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Chicago Welcomes
Electronic Engineers

Edgewater Beach Hotel, Chicago home of the National Electronics Conference. NEC Board of Directors, left to right, W. C. White of G.E. Research Laboratories, Karl Kramer (Sec.) of Jensen Manufacturing Co.; Prof. G. H. Fett, (NEC Pres.) University of Illinois; A. W. Graf (Chairman of Board) Chicago; Prof. L. T. Devore, (Program Chairman) University of Illinois.

NATIONAL ELECTRONICS CONFERENCE

The 1949 annual NATIONAL ELECTRONICS CONFERENCE will be held Monday through Wednesday, September 26-27-28, at the Edgewater Beach Hotel in Chicago, Ill.

Jointly sponsored by the Chicago Section of the Institute of Radio Engineers and the local section of AIEE, together with Illinois Institute of Technology, Northwestern University and the University of Illinois, its program of technical papers is directed toward electronics, particularly the first day, but with some attention to communication subjects, vacuum tubes and circuits on Tuesday and Wednesday.

50 leading firms will exhibit electronic equipment in the East and West Lounges. Admission to exhibits is by invitation. Registration for technical sessions is $3.00. Advance registration may be sent to "National Electronics Conference, Karl Kramer, Secy., 852 East 83rd St., Chicago 19, Ill." A special, advance price of $14.25 will cover registration, the three daily luncheons, and Volume 5 of the "Proceedings of the National Electronics Conference" (separately sold for $4.00). Make check payable to "National Electronics Conference, Inc." and mail before September 12th.

The Institute of Radio Engineers will have a service and information booth in the exhibits.

IRE Regional Meetings Accelerate Electronic Progress!

1949 National Electronics Conference, Chicago, Ill., September 26-28
AIEE Midwest General Meeting, Cincinnati, Ohio, October 17-21
Radio Fall Meeting, Syracuse, N.Y., October 31, November 1-2
1949 Nucleonics Symposium, New York City, October 31, November 1-2
IRE-URSI Fall Meeting, Washington, D.C., October 31, November 1-2
Audio Fair, Hotel New Yorker, New York City, October 27-28-29, 1949, sponsored by the Audio Engineering Society.
Second Annual Nucleonics Symposium October 31 to November 2, 1949 at Hotel Commodore, New York. IRE/AIEE
Southwestern IRE Conference, Baker Hotel, Dallas, Texas, December 9-10, 1949.
1950 IRE National Convention, New York, N.Y., March 6-9

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<table>
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<th>INSTRUMENT</th>
<th>FREQ. RANGE</th>
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<th>ACCURACY</th>
<th>INPUT IMPEDANCE</th>
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<td>-hp- 400A</td>
<td>10 cps to 1 mc</td>
<td>.005 to 300 v 9 ranges</td>
<td>Within 3%</td>
<td>1 meg., 16 µfd shunt</td>
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<tr>
<td>-hp- 400B</td>
<td>2 cps to 100 kc</td>
<td>.005 to 300 v 9 ranges</td>
<td>Within 3%</td>
<td>10 meg., 20 µfd shunt</td>
<td>195.00</td>
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<tr>
<td>-hp- 400C</td>
<td>20 cps to 2 mc</td>
<td>.0001 v to 300 v 12 ranges</td>
<td>Within 3%</td>
<td>10 meg., 15 µfd shunt</td>
<td>200.00</td>
</tr>
<tr>
<td>-hp- 404A</td>
<td>2 cps to 30 kc</td>
<td>.0005 v to 300 v 11 ranges</td>
<td>Within 5%</td>
<td>10 meg., 20 µfd shunt</td>
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<td>0.1 v to 300 v 7 ranges</td>
<td>Within 3%</td>
<td>10 meg., 1.3 µfd shunt</td>
<td>245.00</td>
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For complete data on any -hp- instrument, write direct to factory or contact the nearest -hp- technical representative.

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PROCEEDINGS OF THE I.R.E. September, 1949
**Alarm Indicator Record**

**He can see a thousand miles**

Carrying hundreds of telephone calls, coaxial cable runs through many lonely miles. Far from towns and people, master amplifying stations stand guard with a new automatic alarm system developed by Bell Telephone Laboratories.

At a city terminal, the man on duty makes a check by laying a transparent log sheet over a glass window, and dialing a master station hundreds of miles away. At once the station begins to give an account of itself, lighting lamps under the log sheet to report any abnormal operating condition before it becomes an emergency.

But when something happens that threatens serious trouble, the apparatus acts at once—maybe by switching in a spare coaxial—and calls a distant test board by ringing a bell. Sometimes he can take further steps by remote control; if not, he knows exactly how to brief the nearest repair crew.

With this new alarm system, maintenance men need not be stationed at isolated points, just waiting for something to happen. Instead, they live in their home communities. This makes for better work—and better telephone service.

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<table>
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<th>Time Rec'd</th>
<th>AP BY</th>
<th>No.</th>
<th>Sync. Start</th>
<th>Sync. Stop</th>
<th>Fuses</th>
<th>24-Volts</th>
<th>Abs</th>
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<td>203-204W</td>
<td>205-206W</td>
<td>207-208W</td>
<td>201-202E</td>
<td>203-204E</td>
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<td>202</td>
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<td>204</td>
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<tr>
<td>Rectifier-Inverter Fail</td>
<td>Rectifier-Inverter Fail</td>
<td>64 KC Pilot Alarm at Non-SW. Main</td>
<td>3096 (WKG Line) Pilot at SW. Main</td>
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<td>SP Line Fail at SW. Main</td>
<td>Tot. Line Fail at SW. Main</td>
<td>Auto. Switch at SW. Main</td>
<td>Auto. SW. Locked at SW. Main</td>
<td>201</td>
</tr>
</tbody>
</table>

---

**Bell Telephone Laboratories** Exploring and inventing, devising and perfecting, for continued improvements and economies in telephone service.
QUANTITY production of "GP" Ceramic Condensers is achieved by limiting them to definite capacity values — with a consequent saving in cost without affecting quality. For by-passing and coupling applications which are not frequency determining, "GP" Ceramicons are unexcelled in performance.

General Purpose Ceramicons are sturdy, compact. They are easy to install in small spaces and their use increases production on the assembly line. This feature is proving especially valuable in assembling TV sets.

Erie "GP" Ceramicons are made in insulated and non-insulated styles in popular capacity values up to 10,000 MMF. Write for detailed information and samples.
A transient signal...imagined at 4 KV...but...IMAGED AT 14 KV!

Du Mont Oscillography shows the difference...

At low operating voltages the cathode-ray tube will respond to a high-speed transient signal, but—only at high voltages is the light output sufficient to see and record it.

Du Mont high-voltage Oscillography shows you the difference with these actual (unretouched) oscillograms, and here's how it's done:

...with DU MONT HIGH-VOLTAGE CATHODE-RAY TUBES

Type 280-A is an intensifier-type, high-voltage cathode-ray tube featuring multiple accelerating electrodes for use with accelerating potentials up to 25,000 volts, without serious loss in deflection sensitivity. Writing rates in excess of 280 inches per microsecond have been recorded with this tube.

Type 281-A is an intensifier-type, high-voltage cathode-ray tube featuring multiple accelerating electrodes for use with accelerating potentials up to 25,000 volts, without serious loss in deflection sensitivity. Writing rates in excess of 280 inches per microsecond have been recorded with this tube.

...with these HIGH-VOLTAGE CATHODE-RAY INSTRUMENTS

Type 281-A is a basic cathode-ray indicator utilizing Type SRP-A tube. Provision made for either capacitive or direct-coupling to all deflection plates. Displays single transient writing speeds up to 210 inches per microsecond. Internal power supply provides overall accelerating potential of 8,000 volts; external power supply can be used for higher voltages. The Type 286-A Power Supply is especially designed for use with the Type 286-A indicator, supplying overall accelerating potential of 29,000 volts.

Type 285-AH is a high-voltage version of the versatile Type 250-A. High-voltage Type SRP-A tube replaces Type SCP-A. Provision is made for external high-voltage power supply. Type 250-AH is capable of recording writing speeds ten times those recorded by the Type 250-A. Using Type 263-B Power Supply, accelerating potentials as high as 13,000 volts may be applied. Sufficient light output to project oscillograms up to 30 feet with Type 2542 Projection Lens.

Type 248-A oscillograph is a favorite for high-frequency research. Self-contained, it offers a medium-voltage oscillograph for investigating pulses containing high-frequency components. Vertical amplifiers uniform in response within 30% from 20 cycles to 5 megacycles per second.

With addition of Type 263-B Power Supply, the Type 248-A becomes a high-voltage oscillograph for observation and photography of transients of short duration and extremely low repetition rates. Accelerating potentials up to 14,000 volts may be applied to a Type SRP-A tube.

...with these HIGH-VOLTAGE POWER SUPPLIES

Type 262-A is a regulated rectified R.F. type high-voltage power supply with adjustable output from 18,000 to 25,000 volts. Designed for use with Type 281-A indicator or other component. Especially suited for use with wide-band amplifiers. Types 5RP-A and 5XP- alike are capable of sufficient light output to allow projected oscillograms. Type SXP- is interchangeable with Type 5RP-A except for slightly greater overall length.

For further details and prices, just address...

Du Mont for Oscillography

Allen B. Du Mont Laboratories, Inc., Instrument Division, 1000 Main Avenue, Clifton, New Jersey

Proceedings of the I.R.E. September, 1949
Hi-Q general purpose ceramic capacitors have proven superior to mica and paper condensers of corresponding values. They are available in ratings of 5 mmf to 33,000 mmf. Hi-Q disc capacitors are high dielectric by-pass, blocking or coupling capacitors designed for use where their physical shape is more adaptable than tubular units.

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Our engineers are always available to work with you in the development and production of capacitors, trimmers, resistors and choke coils to meet your specific needs. Write, wire or phone whenever you have a question concerning them.

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

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Satisfaction... year after year of trouble free performance.

**Miniaturization**
The smallest big value components in the business make possible space saving factors which reduce your production costs and increase your profits.

**Hi-Q Electrical Reactance Corp.**

Franklinville, N.Y.

Plants: Franklinville, N.Y. — Jessup, Pa. — Myrtle Beach, S. C.

Sales Offices: New York, Philadelphia, Detroit, Chicago, Los Angeles
New process for depositing selenium gives rectifier stacks greater uniformity, higher efficiency and longer useful life.

Here's real news for rectifier users. G.E.'s new 18-volt selenium cells, made by a special evaporation process which deposits selenium on the aluminum base with greater uniformity than otherwise possible, give you these advantages:

GREATER OUTPUT—With 50% more output than the standard 12-volt cells, the new design can be used for any application except those few which demand 24-hour, year-around service.

HIGHER EFFICIENCY—Not only is the initial efficiency higher, but more uniform coating keeps it high during the life of the stack.

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Selenium stacks are available in several standard sizes. Output in d-c voltage ranges from 18 to 126; applied a-c voltage, from 26 to 161. Bulletin GEA-5258 will give you detailed information. Send for it today.
TIMELY HIGHLIGHTS ON G-E COMPONENTS

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When PRD developed the Type 801 it needed two extremely accurate, matched indicating instruments for measuring Beam Current and Beam Voltage. Because PRD had previous experience with Marion and because Marion is so well known for foolproof, troublefree instruments of this kind... at reasonable cost, it was natural that PRD should turn to Marion. Marion's Standard Type 53SN was selected. This is a 3½" instrument, noted for accuracy and readability.

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PROCEEDINGS OF THE I.R.E. September, 1949
and 7 reasons why they are the criteria of good design in any electronic equipment.

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This new CLARE Type "JMS" Relay is a sensitive relay for switching heavy a-c loads with small d-c controlling currents... as high as 1250 watts can be switched with a 1-watt input.

It combines the outstanding features of the larger CLARE Type "CMS" Relay with the small size and light weight of the CLARE Type "J" Relay and employs a new-type Micro precision switch of unusual efficiency and compact design.

The CLARE Type "JMS" Relay is especially suitable to locations subject to sudden jolts, constant vibration or tilting. It may be provided with either one or two Micro snap-action switches, or with one switch and a pileup of twin-contact springs. For installations where quick removal or replacement may be desirable, it may be fitted and wired to a standard radio type plug.

This new relay is a development of CLARE's unceasing effort to keep pace with every industrial relay requirement. Our engineers and sales representatives are constantly at your service to provide just the relay to meet your specific need.

For full information on the CLARE Type "JMS" Relay, look up the CLARE office in your classified telephone directory... or write for Bulletin 102 to C. P. Clare, 4719 West Sunnyside Avenue, Chicago 30, Illinois. In Canada: Canadian Line Materials Ltd., Toronto 13. Cable Address CLARELAY.

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*First in the Industrial Field*
On Land, Sea, and Air

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Are Relied Upon For Complete Dependability

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Mallory engineering skill, long experience, and complete adherence to quality ideals mean that Mallory Vibrators will give you longer life and less trouble.

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Send complete technical details of your problem. Mallory will supply the answer.

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To Meet Every Need

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These Stackpole Specialties
SIMPLIFY DESIGN and CONSTRUCTION

TINY "GA" CAPACITORS
...that cost no more than "gimmicks"

These sturdy little capacitors cost no more than flimsy, twisted wire "gimmicks," are non-inductive and assure greater stability, higher Q, better insulation resistance and higher breakdown voltage. Standard capacities include .5—.68—1.0—1.5—2.2—3.3 and 4.7 mmfd. types.

INEXPENSIVE SUPPORTS
FOR WINDINGS

Handy Stackpole molded Bakelite coil forms take less space and require one-third fewer soldered connections. Standard forms are available for universal, solenoid, tapped universal and multiple windings. Molded iron center sections can also be provided.

DEPENDABLE, LOW COST
SLIDE SWITCHES

for 3 ampere, 125V. A.C. use

The new Stackpole Type SS-26 Switch is just the thing for electrical appliances and equipment of all sorts! Construction is exceptionally durable and the switches are readily adaptable to various mounting arrangements. Underwriters' approved and conservatively rated for 3 amperes at 125 volts A.C. (or 1 ampere at 125 volts D.C.). Single-pole single-throw and single-pole double-throw types available.

WRITE FOR CATALOG RC7

Stackpole fixed and variable resistors—iron cores for practically any need—inexpensive line and slide switches.

STACKPOLE
Electronic Components Division
STACKPOLE CARBON COMPANY • ST. MARYS, PENNA.

PROCEEDINGS OF THE I.R.E. September, 1949
Kilovolts for Klystrons

New PRD

Power Supply for Microwave Oscillators

Price

$1250.00

F.O.B. Brooklyn, N.Y.

Direct Reading Controls

- Coarse and fine - allow repeller voltage to be accurately adjusted from -20 to +750 volts.

Grid Voltage

May be varied continuously from -300 volts to a positive value limited by the klystron grid current.

Panel Meters

Indicate beam voltage and current directly.

The Type 801 Universal Klystron Power Supply has been developed by PRD to meet the increasing need of research and production engineers for a unit to operate the many high voltage klystrons now in use throughout the microwave spectrum.

Provision is made for c-w, square-wave, sawtooth, and external modulation of conventional internal cavity klystrons, external cavity oscillators, and the new millimeter tubes now coming into use in microwave spectroscopy.

Excellent voltage stabilization guarantees a maximum of oscillator frequency stability. Carefully controlled modulation wave-shapes insure accuracy of standing wave measurements and spectrum analysis when made with oscillators powered by this supply.

Polytechnic Research & Development Company, Inc.

202 Tillary St., Brooklyn 2, New York
COSMALITE* gives
STAR performance in the new ZENITH

This internally threaded Cosmalite coil form of cloverleaf design in the very heart of the Zenith Television Transformer, permits quick tuning of both primary and secondary frequencies through the upper end. The hexagon shaft of the frequency setter easily passes through the upper core and engages in the lower core . . . adjusting the frequencies of both coils with the greatest ease.

Consult us on the many uses of Cosmalite (low cost phenolic tubing) in television and radio receivers.

Cosmalite coil forms are also used in transformers of Zenith's table radios, such as the new Super-Sensitive "Major" FM receiver, above.
NO. 1010 COMPARISON BRIDGE
For precision laboratory adjustment and incoming inspection of resistors, capacitors and inductors. Entirely self-contained, A.C. operated and includes a three frequency oscillator, an A.C. bridge and a null detector.

NO. 1030 "Q" INDICATOR
Frequency range from 20 cycles to 50 kilocycles. "Q" range from 5 to 600. "Q" of inductors can be measured with up to 50 volts across the coil.

NO. 1040 NULL DETECTOR
High gain Null Detector for Bridge measurements. Contains selective circuits for 60-400-1000 cycles. Frequency range 30-30,000 cycles.

NO. 1060 VACUUM TUBE VOLTMETER
For use at audio, supersonic and radio frequencies. Frequency range from 10 cycles to 1.6 megacycles. Input impedance 50 megohms, input capacity 15 MMF. Voltage range of .001 to 100 volts. Frequency range from 10 cycles to 2 megacycles.

NO. 1080 A.C. POWER SUPPLY
A valuable laboratory instrument with continuous variable output from 1 volt to 100 volts @ 60 cycles.

NO. 1120 NULL DETECTOR & VACUUM TUBE VOLTMETER
Vacuum Tube Voltmeter: Sensitivity .1, 1, 10, 100 volts. Input impedance 50 megohms shunted by 20 mmfd. Frequency range 20 cycles - 20,000 cycles.

NO. 1140 NULL DETECTOR
High gain Null Detector for Bridge measurements. Contains selective circuits for 60-400-1000 cycles. Frequency range 30-30,000 cycles.

NO. 1150 UNIVERSAL BRIDGE
For measurement of inductors, capacitors, and determination of resistive and reactive components of impedances. Frequency range 20 cycles - 20,000 cycles. 1% accuracy.

NO. 1160 HIGH FIDELITY OUTPUT TRANSFORMER
High quality output transformer combines unusually wide frequency range together with very low phase shift and harmonic distortion. Frequency range 1/2 DB 20-30,000 cycles.

STEPDOWN TRANSFORMERS
High efficiency Auto Transformer can be used as either Step Up or Step Down transformer equipped with standard receptacle and line cord.

FILTERS
Narrow band pass filters for remote control and tele-metering applications. High pass, low pass, band pass and band elimination filters for communication and carrier systems.

VACUUM TUBE VOLTMETERS

HI-Q MINIATURE TOROIDAL INDUCTORS
Size 3/4" D. x 3/4" - weight 1/4 ounce. Type TI-1. Frequency range 100 cycles - 15,000 cycles. Type TI-7 Frequency range 10,000 cycles - 100,000 cycles. Wound on molybdenum permalloy dust cores. Available in hermetically sealed cans, potted and cased or open units.

HERMETICALLY SEALED COMPONENTS
Class A Grade I components to meet JAN-T-27 specifications. Made to customer requirements where temperature and humidity are factors.
IMPORTANT, EXCLUSIVE FEATURE...

ONLY
BENDIX-SCINTILLA
ELECTRICAL CONNECTORS

offer you this

IMPORTANT, EXCLUSIVE FEATURE...

PRESSURE TIGHT

SOCKET CONTACT ARRANGEMENTS!
Outstanding design and fine workmanship,
combined with materials that will meet every
requirement, make possible the "pressureized
arrangements for ALL sizes of contacts.

PLUS ALL THESE OTHER FEATURES
• Moisture-proof
• Radio Quiet
• Single-piece inserts
• Vibration-proof
• Light Weight

• Easy Assembly and Disassembly
• Power Parts than any other Connector
• No additional solder required
• High Insulation Resistance

Write our Sales Department for detailed information.

SCINTILLA MAGNETO DIVISION of
BENDIX SCINTILLA
SIDNEY, NEW YORK

News—New Products

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

New Plug-In Audio
Amplifiers

The design of new AM, FM and Tele-
vision audio amplifiers has been an-
nounced by the Transmitter Div., General
Electric Co., Electronics Park, Syracuse,
N. Y.

The amplifiers plug into Cannon recep-
tacles mounted at the rear of the trays.
The trays fit into a shelf which can be
mounted in any standard 19 inch cabinet
or relay rack.

Included are the Type BA-1-C Pre-
Amplifier with its' Type FA-22-A Tray,
the Type BA-12-A Program/Monitor
Amplifier with its' Type FA-22-B Tray,
and the Type FA-23-A Shelf which will
accommodate up to six of the pre-amplifiers,
and up to four of the program-monitor
amplifiers.

The Type BA-1-C Pre-Amplifier may
be used as a microphone or transcription
pre-amplifier, booster amplifier, medium-
level line amplifier, or as an isolation
amplifier.

The Type BA-12-A Program/Monitor
Amplifier may be used as a program or line
amplifier, a monitoring amplifier, or an
isolation amplifier.

New Diffusion Pump

A new oil-diffusion type vacuum pump,
the HV-1, is in mass production by Eitel-
McCullough, Inc., 189 San Mateo Ave.,
San Bruno, Calif.

The HV-1 was

(Continued on page 26A)
One of the best known manufacturers of air circuit breakers in the country is the I-T-E Circuit Breaker Company, located at 19th and Hamilton Streets in Philadelphia. From its inception the company has displayed unusual receptiveness to new ideas, whether from within or without; hence it has done its share of pioneering, and perhaps more. Revere is proud to play a part in its progress, through close collaboration with I-T-E engineers, production men, and the purchasing department. The extensive use of Revere Extruded Shapes is but one result of our mutual attack upon I-T-E problems, which the company is good enough to say has saved a great deal of money, as well as made possible a better product. Perhaps similar results would be obtained if you gave us the opportunity to place our knowledge as well as our metals at your disposal. Why not inquire?

REVERE
COPPER AND BRASS INCORPORATED
Founded by Paul Revere in 1801
230 Park Avenue, New York 17, N. Y.

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

This is but a part of the I-T-E Stock of Revere Extruded Shapes in copper, brass, manganese bronze, and aluminum. I-T-E is a great advocate of extruded shapes, from long experience finding them markedly superior, in uniformity, strength, and economy due to the fact that a great deal of machining is avoided.

(Left) I-T-E Contact Block made from an extruded shape. This was formerly extruded in electrolytic copper; changing to Revere Free-Cutting Copper resulted in a saving of 30% in machining time. (Right) I-T-E "K" Breaker, Main Contact Assembly in open position. This is an especially interesting assembly, since it shows no less than eight extruded shapes in copper and bronze. Use of these shapes makes the assembly more compact, stronger, lighter, and considerably more economical to produce. The contacts are silver alloy, and the unit is silver plated. In addition to supplying I-T-E with extruded shapes, and strip, Revere furnishes rolls, bar, rod, sheet, in a wide range of non-ferrous alloys, and seamless brass tube.

(Left) Main movable Contacts and Flexible Connectors in an I-T-E "K" Type Circuit Breaker. The two contacts are made from Revere Extruded Shapes. Revere and I-T-E collaborated closely on the specifications for the thin-gauge copper strip for the pigtails, working out the correct gauge and temper to avoid notch effects and cracking of the connection at the brace. (Right) Main Separable Contacts from an I-T-E Type "LG" Circuit Breaker. These are stamped from Revere Copper Strip with the temper specially controlled to eliminate a de-burring operation previously found necessary to obtain edge surface suitable for electrical contacts. (Inset) Back view of "K" type Breaker showing a similar type of contact.
NEXT BEST THING
TO A

"SKY-HOOK!"

"Sky-hooks" being expensive and somewhat impractical, why not start from the ground up with a Blaw-Knox tower to obtain support for your high-riding FM and TV antennas?

Blaw-Knox, having built towers since spark-gap days, makes available to electronic engineers a degree of practical experience unequalled in this field. So, when you want the next best thing to a sky-hook, call Blaw-Knox.

Shown here is a Blaw-Knox special 417 ft. Type 1140 Heavy Duty tower for Station WHIO, Dayton, Ohio. This tower was designed to support an RCA combination 4-section pylon, plus a 6-section TV antenna and station call letters.

BLAW-KNOX DIVISION
OF BLAW-KNOX COMPANY
2037 Farmers Bank Building • Pittsburgh 12, Pa.

Distributed by
Graybar
OFFICES IN 100 PRINCIPAL CITIES
A 100% INCREASE IN THE GL-502-A's RATED CURRENT CAPACITY AT REDUCED VOLTAGES

0.2 amp average with 180 v on the anode!

"THIS COMPACT METAL THYRATRON WILL REPLACE GLASS TYPE 2050 IN YOUR CIRCUIT, YET IT'S ONLY HALF THE SIZE... AND SELF-SHIELDING!"

CONTINUOUS G-E improvement in design and production makes it possible to rate the GL-502-A thyatron, for low-voltage operation, at twice its former average current capacity, or .2 amp maximum.

Here is performance sure to be welcomed by the electronic designer. No change in size is involved; the GL-502-A (only 2 1/16 inches high when seated) continues to take up minimum space. Also, the tube's self-shielding characteristic, a feature of metal-envelope types, remains an important aid in simplifying circuit and panel design.

Much electronic control equipment is being built to operate at voltages at or near low power-supply potentials. The new, higher-rated current capacity of the GL-502-A under these conditions, gives the designer "more tube to work with." Glass Type 2050—twice the size of the GL-502-A—can be replaced by the smaller thyatron with no loss in tube performance, yet with a pronounced saving in space occupied.

INVESTIGATE THIS GREAT LITTLE METAL THYRATRON NOW... while your new control circuit is in the planning stage! You'll save in space, gain in economy and efficiency. Get the complete story from your nearby G-E electronics office. Or wire or write Electronics Department, General Electric Co., Schenectady 5, New York.

CHARACTERISTICS, TYPE GL-502-A

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Max over-all height</td>
<td>2½ inches</td>
</tr>
<tr>
<td>Max over-all diameter</td>
<td>1 5/16 inches</td>
</tr>
<tr>
<td>No. of electrodes</td>
<td>4</td>
</tr>
<tr>
<td>Cathode voltage, current, approx</td>
<td>6.3 v</td>
</tr>
<tr>
<td>Heating time, typical</td>
<td>10 seconds min</td>
</tr>
<tr>
<td>Voltage drop, typical</td>
<td>8 v</td>
</tr>
<tr>
<td>Arg anode to control-grid</td>
<td></td>
</tr>
<tr>
<td>Capacitance</td>
<td>0.2 mmfd</td>
</tr>
<tr>
<td>Ambient temperature limits</td>
<td>-55 to +90 c</td>
</tr>
</tbody>
</table>

MAXIMUM RATINGS

<table>
<thead>
<tr>
<th>Voltage Type</th>
<th>High-voltage operation</th>
<th>Low-voltage operation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak anode voltage, forward</td>
<td>1,300 v</td>
<td>360 v</td>
</tr>
<tr>
<td>Anode current, instantaneous</td>
<td>1 amp</td>
<td>1 amp</td>
</tr>
<tr>
<td>Average</td>
<td>0.1 amp</td>
<td>0.2 amp</td>
</tr>
<tr>
<td>Time of averaging current</td>
<td>30 seconds</td>
<td>30 seconds</td>
</tr>
</tbody>
</table>

GENERAL ELECTRIC

FIRST AND GREATEST NAME IN ELECTRONICS

PROCEEDINGS OF THE I.R.E. September, 1949
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 27 A)

Precision Asbestos Tubing

Originally designed by Precision Paper Tube Co., 2045 W. Charleston St., Chicago 47, Ill., as a heat resistant base for coils and bobbin, this new asbestos tubing has other potentialities in the industry.

Because it is unaffected by high temperatures, this tubing could be used as insulation and as a dielectric, in such units as heaters, thermal heating devices, for insulating rods, etc.

The tube is made by spirally winding prepared asbestos tape to predetermined sizes around a mandrel, and then deformed into shape. This tubing can be supplied in any length, with wall thicknesses from 0.10 inch.

New Power Supply

The Model B, a recently designed dc power supply that employs new type heavy duty selenium rectifiers, and has a wide range variable voltage control, and damped volt and ammeter, is now available from Electro Products Labs., Inc., 549 W. Randolph St., Chicago 6, Ill.

This source will deliver, from 3 to 9 volts with a rating of 6 volts at 20 amperes continuous, and 35 amperes instantaneous, from 50 to 60 cps 115 volt supply.

The Model B was primarily designed for testing or operating automobile radio receivers, but it will also test faulty vibrators, push button solenoids, 6 volt battery type receivers, and will provide over and under voltage operating conditions for all auto radio receivers.

(Continued on page 30A)
AEREOVOX TYPE '87

AEROCON
SELF-MOLDED PLASTIC TUBULAR

• Brand new! Looks like a paper tubular yet is entirely different. This plastic tubular is molded in its own paper tube. That means a tubular approaching the performance of the molded-plastic capacitor yet available at a price closer to the conventional paper tubular.

For example: In a typical TV receiver using some 30 molded-plastic capacitors, the Type '87 Aerocon scores a saving of 50 cents! And without sacrificing top performance!

It's all due to another exclusive Aerovox development—Aerolene—the combination impregnating-sealing material already featured in Aerovox Duranite tubulars in general use.

So here's real performance insurance for those TV, auto-radio, oscillograph and other severe-service requirements. And at irresistible price, too.

Samples, ratings, quotations available on request

AEROSCON CHECK LIST

✓ Paper-tube tubular but with ends sealed with rock-hard Aerolene.
✓ Aerolene impregnant eliminates stocking and using of both wax and oil capacitors. One impregnant does work of both.
✓ Absence of impregnating oils and waxes eliminates dripping or cracking of wax coating which interaction might cause.
✓ Equal to or even smaller than molded units.
✓ Heat- and humidity-resistant qualities of the order of the best plastic tubulars.
✓ Can be used without drips at 212°F.
✓ Dielectric strength maintained at elevated temperatures. Rated voltages based on 212°F operation.
✓ No softening of dip wax to become gummy, tacky, dirty or dark.
✓ Unimpaired by sub-zero operation. Capacitance increases slightly with temperature rise.
✓ Extremely high initial insulation resistance. Units recover insulation resistance upon heating.

FOR RADIO-ELECTRONIC AND INDUSTRIAL APPLICATIONS

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

SALES OFFICES IN ALL PRINCIPAL CITIES • Export: 13 E. 40th St., New York 16, N.Y.

Cable: 'ARLAB' • In Canada: AEROVOX CANADA LTD., HAMILTON, ONT.
... you draw the curve  
WE'LL BUILD THE CARTRIDGE  

Your specifications . . . your special requirements in phono pickup cartridges . . . are ideally "custom-solved" through E-V creative engineering . . . unusual manufacturing facilities . . . and inherent advantages of exclusive TORQUE DRIVE.*

CUSTOM RESPONSE  
Smooth upper response with roll off frequency to your specifications or wide range, peak-free response to 10 kc. You draw the curve, we'll build the cartridge.

VOLTAGE  
E-V TORQUE DRIVE cartridges provide the highest compliance per volt output. For example, the E-V 14 cartridge tracks at 5 grams with excellent wave form down through 50 c.p.s. on the RCA 12-5-31V record at 1 volt at 1,000 c.p.s.

TRACKING FORCE  
With the high compliance and low mass of the driving system, needle forces at 5 grams for both one and three mil records are used in everyday production by leading manufacturers. Cartridges with even lower needle force with slight reduction in voltage are thoroughly practical. 3 gram tracking pressures are definitely in sight.

COMBINATION One and Three Mil  
E-V TORQUE DRIVE again leads in twin needle cartridge design. Tracking force of 5 grams on both one and three mil records precludes weight changing. Straight line needle position assures accurate set down when used with changers. Approximately the same output is obtained on both stylii. The E-V Twin-Tilt cartridge mounts in any arm with 1/4" mounting holes with no modification except adjustment for correct needle force.

MOISTURE PROOFING  
The cartridge is entirely filled with DC4 Silicone jelly—the material that is used for inhibiting moisture on aircraft wiring. Tests indicate that it increases the life of an ordinary crystal some 20 times. This is a plus feature, found in all E-V crystal cartridges.

Our engineering staff and full facilities are at your service. Contact us today.
FILTER SPECIALISTS
PRODUCERS OF PERMALLOY DUST TOROID COILS AND FILTERS FOR OVER A DECADE

FOR FILTERS

BROAD BAND SHARP CUTOFF FILTER

LOW FREQUENCY — LOW PASS FILTER

SUB-OUNCER TOROID FILTERS
 Filters employing SUB-OUNCER toroids and special condensers represent the optimum in miniaturized filter performance. The band pass filter shown weighs 6 ounces.

UNCASED TOROIDS
ATTENUATES 10KC TO 30 MEGACYCLES

FOR HIGH Q COILS

HQA, C, D
1 1/4" Dia. x 1 1/4" High.

HQB
2 1/8" L. x 1 3/4" W. x 2 1/8" H.

HQC
HOD

United Transformer Co.
150 Varick Street
New York 13, N. Y.
Export Division: 13 East 40th Street, New York 16, N. Y.
Cables: "Arlab"
Look in the IRE Yearbook—
3347 Electronic Supply Firms Listed:

Every firm in the alphabetical directory has address and complete line of products of interest to radio engineers is given in code numbers so you can understand what related products the firm supplies. This kind of listing gives you the most comprehensive and up-to-date picture of the company you are looking up, both as to manufactured products and services. A cross-index provides direct reference to the page on which advertisers provide fuller information. Easy-reading type and spacing!

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 briefer listings.

Complete

Fast Reference Understandable

Published for IRE Members from Associate grade—up.
by THE INSTITUTE OF RADIO ENGINEERS

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 264)

New Pre-Set Type Industrial Counter

A new series of presettable high-speed electronic counters for industrial applications, with counting rates of 10,000 per minute, and operating from photoelectric or magnetic pickup sources, may be purchased from the manufacturer Airlectron, Inc., P.O. Box 151, Caldwell, N. J.

Various models are designed for preset ranges of from 1 to 999 gross with decade selection and direct reading count indication (shown in illustration); fixed preset selection of any four quantities, and for decade selection of any number up to 10,000.

The output circuits of these counters may be connected as to start and stop machine operation, switch loading chutes, or perform functions associated with the counting.

Special timers are furnished if desired for use in rate determination or the precise measurement of rotational speeds.

Dimensions: 14" wide, 10" high, and 8" deep.

Turnover Type Pickup

A new turnover type phonograph pickup with double needle cartridge, which plays 33 1/2, 45, and 78 rpm records at the same pressure, has been manufactured by Astatic Corp., Conneaut, Ohio.

The pickup is the Model CLD, which employs Model LQD cartridge. All three recordings are played at 8 grams pressure. The needles may be removed by light prying with a knife point or small screw driver; this may be performed without removing the cartridge.

When using a 78 rpm Audio-Tone test record, output voltages are 1.2 volts at 1,000 cps, and with a 33 1/3 rpm Columbia 281 record, .9 volt at 1,000 cps.

(Continued on page 58A)
A Few of the Added Features that make the 3rd Edition of this Handbook essential to you

- Radar Fundamentals
- Microwave Links and Propagation
- Pulse-modulation Methods
- Wideband Intersstage Circuit Design
- Filter-Network Design
- Transformers and Other Components
- Expanded Antenna Data
- Multi-vibrators and Special Oscillators
- Electroacoustics Theory and Practice
- Bridges and Impedance Measurements
- Microwave Tubes and Circuits
- Servo-Mechanism Fundamentals
- AM, FM, and TV Broadcasting
- Transmission-Line Formulas Greatly Expanded
- Spurious Frequency Responses
- Expanded Mathematical Formulas
- Laplace Transforms
- Summary of Maxwell's Equations

Over 100,000 satisfied users attest to the real worth of this indispensable data book. Now it has been revised and enlarged from 322 to 640 pages...jammed packed with the kind of reference data you need to have on hand.

Over 653 charts and diagrams and 207 tables give quick answers to the problems that come up in practical radio, television and electronic work. The handy subject index makes it easy to find the exact information you require.

Federal Telephone and Radio Corporation

Publication Department—67 Broad Street, New York 4, N. Y.
THE FOUNTAINHEAD OF MODERN TUBE DEVELOPMENT IS RCA

...the five most popular kinescopes ___from one dependable source

RCA now has a popular type of kinescope to accommodate television receiver designs in practically every class and price range.

Concentrated production on these five accepted types results in longer production runs, which in turn, make possible lower cost, more uniform, and better quality tubes for our customers.

All five types are currently being mass-produced at the famed RCA tube plant in Lancaster, Pennsylvania. In addition, a large new plant is under construction at Marion, Indiana, where the production will be centered on the RCA-16-inch metal-cone kinescope.

RCA Application Engineers are ready to cooperate with you in applying these kinescopes and their associated components to your specific designs. For further information write RCA, Commercial Engineering, Section 471R, Harrison, N.J.

RCA, LANCASTER, PA.

THE WORLD'S MOST MODERN TUBE PLANT...
PROCEEDINGS OF THE I.R.E.

(Including the WAVES AND ELECTRONS Section)

Published Monthly by
The Institute of Radio Engineers, Inc.

Volume 37

September, 1949

Number 9

PROCEEDINGS OF THE I.R.E.

Julius A. Stratton, Director, 1948–1950

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3413. Part II—Investigations of High-Frequency Echoes... H. A. Hess

3416. Microwave Filter Theory and Design

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3420. An Analysis of Magnetic Amplifier Networks

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3424. "Table for Use in the Addition of Complex Numbers" by Jorgen Rybner and K. S. Sorensen

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3429. "Automatic Record Changer Service Manual, Volume Two" (1948)

3430. "Advances in Electronics, Vol. 1" edited by L. Marson

3431. "Fundamentals of 1-Literic Waves" by Hugh Liljefors, Skilling

3432. "A Textbook of Radar" by the Staff of the Radiophysics Laboratory

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Julius A. Stratton

DIRECTOR, 1948–1950

Julius Adams Stratton was born on May 18, 1901, in Seattle, Washington. In 1919 he entered the University of Washington, but transferred the following year to the Massachusetts Institute of Technology. After receiving the bachelor of science degree in electrical engineering from MIT in 1923, he spent the next year studying at the Universities of Grenoble and Toulouse in France.

Dr. Stratton returned to MIT in 1924 to act as research assistant in the Communications Laboratory for two years. During this period he received the master's degree in electrical engineering, in 1925. Returning to Europe in 1926, he was granted the doctor of science degree from the Technische Hochschule, in Zurich, Switzerland, in 1927. The following year he continued his studies in physics at the University of Munich.

In 1928 Dr. Stratton returned to the United States and was appointed assistant professor of engineering at MIT. He was transferred to the physics department four years later. When the Radiation Laboratory was established at MIT in 1940, he became a staff member, with the rank of full professor.

From 1942 to 1947, Dr. Stratton acted as expert consultant in the office of the Secretary of War. Meanwhile, in 1945, MIT established the Research Laboratory of Electronics, and appointed Dr. Stratton its head. He was also Chairman of the Committee on Electronics, Research, and Development from 1946 to 1947; and, for his services rendered during the war, was awarded the Medal for Merit by the Secretary of War.

A Director of the Armed Forces Communications Association, Dr. Stratton is a Fellow of the American Physical Society and the American Academy of Arts and Sciences. He became a Member of the Institute in 1942, a Senior Member in 1943, and a Fellow in 1945.
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—The Editor
Double-Stream Amplifiers*

J. R. PIERCE†, FELLOW, IRE

Summary—This paper presents expressions useful in evaluating the gain of a double-stream amplifier having thin concentric electron streams of different velocity and input and output gaps across which both streams pass.

1. INTRODUCTION

A NEW HIGH-FREQUENCY AMPLIFIER has been described recently.1-4 In this device there are two closely coupled streams of electrons with slightly different velocities. These two electron streams support a space-charge wave which travels with a velocity lying between the two electron velocities, and which increases in amplitude with distance as it travels. Use can be made of this increasing wave in obtaining amplification. The increasing wave can be set up by means of a helix or resonator which forms the input circuit. After the wave has increased as much as is desired, an amplified output can be obtained by means of a helix or resonator which acts as an output circuit.

The double-stream amplifier has many attractive features. For instance, the two electron streams need be close to one another, but need not be close to a long metallic circuit. Also, a high gain can be attained in a relatively short distance. In order to evaluate the new device, however, it is necessary to know how close the electron streams must be to one another, and what overall gain may be expected in a given physical structure. It is the purpose of this paper to carry the theory far enough so that the overall performance of a particular structure can be calculated and so that the importance of various parameters such as separation of the electron streams can be evaluated.

In the structure to be analyzed, which is shown in Fig. 1, the two streams are velocity modulated in passing across the gap between the grids of input resonator $R_1$ which is fed by input line $L_1$. This velocity modulation sets up an increasing space-charge wave. The wave grows in the space between input resonator $R_1$ and output resonator $R_2$. The convection current associated with the wave excites resonator $R_2$ and so transfers power to the output line $L_2$. The electron streams are collected on an anode $A$.

* Decimal classification: R132X339.2. Original manuscript received by the Institute, February 9, 1949; revised manuscript received. June 10, 1949.
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2. MOTION OF THE ELECTRONS

One problem in the analysis of double-stream amplifiers is to express the ac charge in the electron stream in terms of the fields acting on the electrons. In order to save space, expressions derived elsewhere will be used.

Fig. 1—A double-stream amplifier using concentric tubular electron streams and resonators as input and output circuits.

These are linearized (small signal) expressions derived assuming 1. nonrelativistic equations of motion, 2. a static electric field derivable from a potential, and 3. the same potential for all electrons at a given cross section of the beam. This would be essentially true for a thin tubular beam. 4. no electron motion perpendicular to the direction of electron flow. MKS units are used. All quantities are assumed to vary with time and distance as exp $j(\omega t - \beta z)$.

The following additional nomenclature will be used:

- $\varepsilon_0 = $ dielectric constant of vacuum $\varepsilon_0 = 8.85 \times 10^{-12}$ farad/meter
- $\eta = $ charge-to-mass ratio of the electron $\eta = 1.76 \times 10^9$ coulomb/kilogram
- $I_{01}, I_{02} = $ dc currents
- $u_1, u_2 = $ dc velocities
- $p_{01}, p_{02} = $ dc linear charge densities $p_{01} = -I_{01}/u_1, p_{02} = -I_{02}/u_2$
- $\rho_1, \rho_2 = $ ac linear charge densities
- $V_{01}, V_{02} = $ dc voltages with respect to the cathodes
- $\beta_1 = \sqrt{\eta V_{01}}, \beta_2 = \sqrt{2\eta V_{02}}$

By use of the equations of motion and continuity $\rho_1$ and $\rho_2$, the linear charge densities in the two streams, have been shown to be

$$\rho_1 = \frac{I_{01}\beta_1}{2u_1V_{01} \left[ \beta \left( 1 + \frac{b}{2} \right) - \beta \right]^2},$$

$$\rho_2 = \frac{I_{02}\beta_2}{2u_2V_{02} \left[ \beta \left( 1 + \frac{b}{2} \right) - \beta \right]^2},$$

where $b = \frac{e}{2m}$, $e$ is the electron charge, $m$ is the electron mass.

* * *
\[ \rho_2 = \frac{I_2 \beta^2}{2u_1V_1 \left[ \beta \left( 1 - \frac{b}{2} \right) - \beta \right]^2} \]  

Here \( b \) is the fractional velocity separation

\[ b = \frac{2(u_1 - u_2)}{u_1 + u_2} \]

and \( u_0 \) is a sort of mean velocity specified by a mean potential \( V_0 \)

\[ u_0 = \sqrt{2\pi V_0} = \frac{2u_1u_2}{u_1 + u_2} \]

and \( \beta_0 \) is a phase constant related to \( u_0 \)

\[ \beta = \frac{\omega}{u_0} \]

We shall treat only a special case, that in which

\[ \frac{I_0_1}{u_1 V_0} = \frac{I_0_2}{u_2 V_2} = \frac{I_0}{u_0 V_0} \]  

Here \( I_0 \) is a sort of mean current which, together with \( u_0 \), specifies the ratios \( I_0_1/u_1 V_0 \) and \( I_0_2/u_2 V_2 \), which appear in (1) and (2).

In terms of these new quantities, the expression for the total ac charge density \( \rho \) is, from (1) and (2)

\[ \rho = \rho_1 + \rho_2 = \frac{I_0 \beta^2}{2u_0 V_0} \left[ \frac{1}{\beta_0 \left( 1 - \frac{b}{2} \right) - \beta} \left[ \beta_0 \left( 1 + \frac{b}{2} \right) - \beta \right]^2 \right] \]

Equation (7) is a ballistical equation telling what charge density \( \rho \) is produced when the flow is bunched by a voltage \( V \). To solve our problem, that is, to solve for the phase constant \( \beta \), we must associate (7) with a circuit equation which tells us what voltage \( V \) the charge density produces.

3. Circuit Equation

Here it will simply be assumed that the ac voltage is proportional to the ac charge, as in a capacitance. The factor of proportionality \( p \) may be called a coefficient of induction

\[ V = pp. \]

For a tubular beam, \( p \) will be a function of beam radius and of \( \beta \). As we are interested in values of \( \beta \) lying in a small range about \( \beta_0 \), we will make a further approximation, and assume \( p \) to have the value it would have for \( \beta = \beta_0 \). This makes \( p \) a real constant.

The evaluation of \( p \) is merely a problem in electrostatics; this is a simple problem in certain cases. For instance, for a thin tubular beam in free space, one finds

\[ p = F(\beta_0 a)/\varepsilon_0 \]  

\[ F(\beta_0 a) = \frac{I_0(\beta_0 a)K_0(\beta_0 a)}{2\pi} \]

Here \( I_0 \) and \( K_0 \) are modified Bessel functions. In Fig. 2, \( F(\beta_0 a) \) is plotted versus \( \beta_0 a \).

In actually using a tubular beam, a tubular outer conductor must be used. The presence of such an outer conductor will reduce \( p \) somewhat. We can get an idea of the seriousness of this reduction by considering the case of a plane electron beam a distance \( d \) from a parallel plane conductor. In this case, we find the ratio \( R \) of the field produced by a given charge to the field which

\[ R = \frac{2}{1 + \cosh \beta_0 d} \]

Fig. 3—Factor \( R \), by which the coefficient of induction for a plane electron stream is reduced by the presence of a conducting plane a distance \( d \) away, plotted versus \( \beta_0 d \), the separation of the stream and the plane in radians. This factor \( R \) may also be used in connection with sufficiently large tubular beams with a tubular shield.
would be produced if the conducting plane were removed to be

\[ R = \frac{2}{1 + \coth \beta_0 l} \]  

(11)

In Fig. 3, \( R \) is plotted versus \( \beta_0 l \). We see that for \( \beta_0 l \) larger than 1.5, the conducting plane has little effect. Presumably, an outer tubular conductor more than 1.5 radians away from a tubular beam would have little effect. For reasonably large values of \( \beta_0 l \) (perhaps \( \beta_0 l = 1.5 \) or larger), \( R \) can be used as a correction factor in connection with (10) in obtaining \( p \) for a tubular beam with a tubular outer conductor.

4. Combined Equation and Solution

Let us combine (7) with (8). We obtain

\[
\frac{1}{[(1 - b/2) - \beta/\beta_0]^2} + \frac{1}{[(1 + b/2) - \beta/\beta_0]^2} = \frac{1}{U^2}
\]

(12)

\[
U^2 = \frac{(\beta/\beta_0)^2 I_0 \beta}{2\eta V_0} = \frac{(\beta/\beta_0)^2 I_0 \beta}{2^{1/2} \rho^{1/2} \eta \epsilon_0 V_0^{3/2}}
\]

(13)

In assuming \( \beta \) to be a constant over the range considered, we assumed \( \beta \) to be nearly equal to \( \beta_0 \). Hence, in (13) we will replace \( (\beta/\beta_0)^2 \) by unity, giving the approximate relation

\[
U^2 = \frac{(\rho \epsilon_0 I_0 \beta)}{2^{1/2} \rho^{1/2} \epsilon_0 V_0^{3/2}} = 9.52 \times 10^4 \rho \epsilon_0 I_0 \beta / V_0^{3/2}
\]

(14)

In solving (12) it is convenient to let

\[
\beta = \beta_0 (1 + \delta)
\]

(15)

\[
U_M^2 = b^2/8
\]

(16)

We then obtain

\[
\delta = \pm j \frac{b}{2} \left[ \pm (U/2U_M) \sqrt{(U^2/U_M^2)} + 8 - (U^2/2U_M^2) - 1 \right]^{1/2}
\]

(17)

We see that \( \delta/b \) is a function of \( U/U_M \) only. In terms of \( \delta \), the amplitude varies with distance as

\[
\exp(-j\beta_0 - j\beta\delta)
\]

We see that the increasing wave, if there is one, is given by the plus signs. The rate of increase in db per wavelength per unit \( b \), which will be called \( A \), is

\[
A = 20 \log_{10} e (2\pi) \left( \frac{-j\delta}{b} \right)
\]

\[
= 27.3 [(U/2U_M) \sqrt{(U^2/U_M^2)} + 8 - (U^2/2U_M^2) - 1]^{1/2}
\]

(18)

In Fig. 4, \( A \) is plotted versus \( (U/U_M)^2 \). From (14) and (16) the abscissa is

\[
(U/U_M)^2 = 7.61 \times 10^4 \rho \epsilon_0 I_0 / b^2 V_0^{3/2}
\]

(19)

For a given geometry, that is, for a given value of \( \rho \epsilon_0 \), and for given values of \( b \) and \( V_0 \), \( (U/U_M)^2 \) varies directly as the current. Hence, increasing the abscissa is equivalent to increasing the current. For currents below some critical current \( I_M \) given by

\[
(U/U_M)^2 = 1
\]

\[
I_M = 1.313 \times 10^{-6} b^2 V_0^{3/2} / \rho \epsilon_0
\]

(20)

there is no increasing wave. As the current is raised above this value, the gain gradually rises and approaches asymptotically 27.3b db/wavelength.

5. Boundary Conditions

So far the rate of increase of the increasing wave has been evaluated in terms of \( I_0 \), \( V_0 \), \( \rho \epsilon_0 \), and \( b \). In this section, we will consider the initial amplitude of the increasing wave which is set up by velocity modulating the electron streams, and the convection current associated with the increasing wave. In considering these matters, expressions for velocity and convection current will be written in terms of quantities already defined. For instance,

\[
V_1 = \eta \beta / \rho \epsilon_0 (1 - b/2)
\]

(21)

We have already replaced \( \beta/\beta_0 \) by unity in some of our expressions. We may as well do so here. For the increasing wave we can have \( \delta \ll 1 \) only when \( b/2 \) is not very small; for the unattenuated waves \( \delta \ll 1 \) only when \( b/2 < 1 \). Hence, we will assume \( b/2 < 1 \). We can write (21) approximately, with some rearrangement, as

\[
(b/2)V_1 = \frac{\eta}{\rho \epsilon_0 (1 + 2\delta/b)} V
\]

(22)

and similarly

\[
(b/2)V_2 = \frac{-\eta}{\rho \epsilon_0 (1 - 2\delta/b)} V
\]

(23)
Let $q_1$ and $q_2$ be the convection currents of the two streams. We have the general relation

$$\frac{\partial q_1}{\partial z} = -\frac{\partial \rho_1}{\partial t}$$

$$q_1 = \frac{\omega}{\beta} \rho_1.$$  \hspace{1cm} (24)

We have,\(^1\) using the same approximations as before,

$$(b/2)^2 q_1 = \frac{-I_0}{2V_d(1+2\delta/b)^2} V,$$  \hspace{1cm} (25)

$$(b/2)^2 q_2 = \frac{-I_0}{2V_d(1-2\delta/b)^2} V.$$  \hspace{1cm} (26)

If the first stream is velocity-modulated by a voltage $V'$, the initial velocity will be

$$v_1 = \frac{u_1}{2} \frac{V_i}{V}.$$  \hspace{1cm} (27)

In accordance with the approximation we have made $(b/2 \ll 1)$ we will replace $u_1$ by $u_0$. Then we obtain as the initial velocities

$$v_1 = v_2 = \frac{u_0}{2} \frac{V_i}{V} = \frac{\eta V_i}{u_0}.$$  \hspace{1cm} (28)

The initial convection currents will be zero.

Let us denote the $V'$s for the four waves and the four values of $\delta$ for the four waves (which we obtain from (18)) by the subscripts $I$, $II$, $III$, and $IV$. We will use the subscript $I$ for the increasing wave (given by the plus signs in (18)). We see from (22), (23), and (28) that

$$\sum_{N=1}^{IV} \frac{V_N}{(1+2\delta_N/b)} = bV_i/2$$  \hspace{1cm} (29)

$$\sum_{N=1}^{IV} \frac{V_N}{(1-2\delta_N/b)} = -bV_i/2$$  \hspace{1cm} (30)

and from (25), (26), and the fact that the initial convection currents are zero

$$\sum_{N=1}^{IV} \frac{V_N}{(1+2\delta_N/b)^2} = 0$$  \hspace{1cm} (31)

$$\sum_{N=1}^{IV} \frac{V_N}{(1-2\delta_N/b)^2} = 0.$$  \hspace{1cm} (32)

We can solve these equations, obtaining $2V_i/bV_i$, where $V_i$ is the voltage associated with the increasing wave, as a function of $(U/U_M)^2$.

We are more interested in $|q|$, the magnitude of the total convection current, $(q_1+q_2)$, associated with the increasing wave. From (25) and (26) we obtain

$$|q| = \frac{I_0}{bV_0} G$$  \hspace{1cm} (33)

We note that both $\delta_0/b$ and $2V_i/bV_i$ are functions of $(U/U_M)^2$ only. The function $G$ is plotted versus $(U/U_M)^2$ in Fig. 5.

![Figure 5](image-url)

**Fig. 5**—A conductance factor $G$ versus $(U/U_M)^2$. The ratio of initial convection current in the increasing wave to the voltage velocity-modulating both streams is $(I_0/bV_0)G$.

The quantity $|q|/V_i$ is a conductance. If the tube is long enough so that at the output the increasing wave is large compared with the three other waves, the trans-conductance of the tube $g_m$ is, neglecting the effect transit time across the gaps (gap factor),

$$g_m = \frac{I_0}{bV_0} G 10^{bA N/10}.$$  \hspace{1cm} (35)

Here $N$ is the number of beam wavelengths between the gaps. $G$ and $A$ are functions of $(U/U_M)^2$. The number of beam wavelengths is the number of free-space wavelengths times $c/u_0$, where $c$ is the velocity of light, and

$$c/u_0 = 505/\sqrt{V_0}.$$  \hspace{1cm} (36)

6. **Separation of the Electron Streams**

So far we have assumed that the same field acts on both electron streams. We will now consider a case in which all the electrons in stream 1 are acted on by an ac potential $V_1$ and all electrons in stream 2 are acted on by a potential $V_2$. Such would be the case, for instance; for thin concentric electron streams.

In this case we have three coefficients of induction, $p_1$, $p_2$, and $p_3$, which appear in the following relation between charge densities and potentials

$$V_1 = p_2 p_2 + p_3 p_3$$  \hspace{1cm} (37)

$$V_2 = p_3 p_3 + p_1 p_1.$$  \hspace{1cm} (38)
The \( p \)'s will be taken as real constants.

We will now have in place of (1) and (2)

\[
\rho_1 = \frac{\eta l_{01}}{u_1^3} \frac{\beta^2}{(\beta_1 - \beta)^2} V_1 = a_1 V_1 \quad (39)
\]

\[
\rho_2 = \frac{\eta l_{02}}{u_2^3} \frac{\beta^2}{(\beta_2 - \beta)^2} V_2 = a_2 V_2. \quad (40)
\]

As written, (39) and (40) define the parameters \( a_1 \) and \( a_2 \), which relate charge densities to the voltages producing them. We can now write (37) and (38)

\[
V_1 = \rho_1 a_1 V_1 + \rho_m a_2 V_2 \quad (41)
\]

\[
V_2 = \rho_2 a_2 V_2 + \rho_m a_1 V_1. \quad (42)
\]

By eliminating the \( V \)'s we obtain

\[
(1 - \rho_1 a_1)(1 - \rho_2 a_2) = \rho_m a_1 a_2. \quad (43)
\]

We will consider the special case in which

\[
\frac{\eta l_{01}}{u_1^3} = \frac{\eta l_{02}}{u_2^3} = \frac{\rho m l_0}{2\eta s V_0} = K^2 \quad (44)
\]

The quantities \( u_0 \) and \( V_0 \) will have the same meaning as before. The product \( \rho m l_0 \) can be regarded as a single parameter.

By introducing \( \delta \) as before, we obtain

\[
\frac{1}{(\delta - b/2)^2} + \frac{1}{(\delta + b/2)^2} = \frac{1}{K^2} \left[ \frac{K^4(1 + \delta)^4(1 - \frac{\rho m^2}{\rho_1 \rho_2})}{(\delta - b)^2(\delta + b)^2} \right]. \quad (45)
\]

As before, we will neglect \( \delta \) with respect to unity. We will now let

\[
\frac{1}{U^2} = \frac{1}{K^2} \left[ \frac{K^4(1 - \frac{\rho m^2}{\rho_1 \rho_2})}{(\delta^2 - (b/2)^2)} \right]. \quad (46)
\]

In terms of \( U \), (45) can be rewritten

\[
\frac{1}{(\delta - b/2)^2} + \frac{1}{(\delta + b/2)^2} = \frac{1}{U^2}. \quad (47)
\]

This is the same as (12), and the solutions are given by (17).

If we substitute the value of \( \delta \) for the increasing wave (using the + signs in (17)) into (46), we obtain

\[
\frac{1}{K^4} \frac{1}{K^2 U^2} + \frac{(1 - \rho m^2/\rho_1 \rho_2)}{U^4(1 - \sqrt{1 + 8(U_i/U)^2})} = 0. \quad (48)
\]

Let

\[
\alpha = (U/U_m)^2 \quad (49)
\]

and

\[
1 - \rho m^2/\rho_1 \rho_2 = S. \quad (50)
\]

We obtain on solving (48)

\[
\frac{U_i^2}{K^2} = \frac{1}{2a} \left( 1 + \sqrt{1 - \frac{4S}{(1 - \sqrt{1 + 8/\alpha})^2}} \right). \quad (51)
\]

In (50), \( S \) is a measure of the separation of the electron streams. For coincident electron streams, \( \rho_m = \rho_1 = \rho_2 \) and \( S = 0 \). For infinitely remote (uncoupled) streams, \( \rho_m = 0 \) and \( S = 1 \). If the streams are coincident, so that \( S = 0 \),

\[
\frac{U_i^2}{U^2} = \frac{U_m^2}{K^2}. \quad (52)
\]

That is, \( U \) and \( K \) are equal and the solution is equivalent to that for coincident streams.

Suppose we use, for comparison, a case in which \( S = 0 \) (coincident streams). The ratio of the actual current to current which would just give gain if \( S \) were zero is \( K^2/U_m^2 \). The ratio of effective current to critical current is \( \alpha \). We can use (51) to plot \( \alpha \) (which must be entered as the abscissa in Fig. 4 to obtain \( A \)) versus \( K^2/U_m^2 \), the ratio of actual current to a critical current calculated setting \( S = 0 \). Such a plot is shown in Fig. 6.

It is helpful in considering Fig. 6 to express the abscissa, \( K^2/U_m^2 \), by means of (49) and (44) as

\[
\frac{K^2}{U_m^2} = \frac{2\rho_m l_0}{(b/2)^2 u_0 V_0}. \quad (53)
\]

Thus, the abscissa is proportional to current.

The gain per wavelength is proportional to \( \delta \), the value for \( \delta \) for the increasing wave, and \( \delta \) can be obtained by means of (18)

\[
\delta = \frac{b\sqrt{\alpha}}{2\sqrt{2}} (\sqrt{1 + 8/\alpha} - 2/\alpha - 1)^{1/2}. \quad (54)
\]

Actually, in obtaining the gain per wavelength per unit \( b \), \( \alpha \) may be used as the abscissa in Fig. 4.

Let us consider the curves of Fig. 6. We see that for \( S = 0 \) the plot is a \( 45^\circ \) line; in this case \( U = K \). For finite values of the parameter \( S \), the effective current at first rises as current is increased, and then falls. There is thus a maximum value of effective current.

If we examine (51) we see that \( \alpha \) has its maximum value at

\[
1 - \frac{4S^2}{(1 - \sqrt{1 + 8/\alpha})^2} = 0. \quad (55)
\]

A larger value of \( \alpha \) would, from (51), imply a complex current. The maximum value of \( \alpha \), obtained from (55) is

\[
\alpha_{\text{max}} = \frac{8}{[(1 + 2S^2) - 1]}. \quad (56)
\]
We see from (51) and (55) that this value occurs at a current ratio
\[ \frac{K^2}{U_M^2} = 2a. \] (57)

The currents for this maximum can be obtained using (53) and (44). As \( S \) approaches unity, \( \alpha_{\text{max}} \) approaches unity (no gain) and \( \frac{K^2}{U_M^2} \) approaches 2.

We are now in a position to obtain the gain versus current for separated electron streams, and we can obtain the maximum gain in terms of \( S \) from (56) and (54). It remains to evaluate this parameter in some physical case.

Sometimes two thin hollow cylindrical beams are used. The writer has treated this case and found that for such beams
\[ S = 1 - \frac{I_0(\beta_d)}{I_0(\beta_{ob})} \frac{K_0(\beta_{ob})}{K_0(\beta_{oa})} \] (58)

Here \( a \) and \( b \) are the radii of the inner and outer beams. A much simpler expression is adequate when \( \beta_{oa} \) and \( \beta_{ob} \) are sufficiently large so that the beams may be regarded as essentially plane. For a plane sheet of impressed charge of phase constant \( \beta_0 \), the potential falls off normal to the sheet as \( \exp(-\beta_d d) \) where \( d \) is distance from the sheet. Hence, we see at once that
\[ \frac{p_m}{p_1} = \frac{p_m}{p_2} = e^{-\beta_0 d} \]
\[ S = 1 - e^{-2\beta_0 d} \] (59)

where \( \beta_{od} \) is the separation between the current sheets in radians. The parameter \( S \) is plotted versus \( \beta_{od} \) in Fig. 7. The maximum gain per wavelength per unit \( b \) is plotted versus \( \beta_{od} \) in Fig. 8.

To give an idea of the separations involved, it may be pointed out that for 1,000 volts and 4,000 Mc, 1 radian corresponds to 0.030 inches.

![Fig. 6](image1)

Fig. 6—A curve for separated streams, giving \( \alpha = (U/U_M)^2 \), the effective current ratio, versus \( (K/U_M)^2 \), a current ratio for various values of beam separation parameter \( S \).

![Fig. 7](image2)

Fig. 7—Beam separation parameter \( S \) versus beam separation in radians \( \beta_{od} \) for parallel plane beams separated by a distance \( d \). This curve is also applicable to large concentric tubular beams separated radially by a distance \( d \).

In computing gain for separated tubular streams, it is necessary to know not only \( S \), but \( p_1 \) and \( p_2 \) as well. These coefficients of induction can be computed by applying the formulas of section 3 to each stream separately.

The boundary conditions for separated streams have not been worked out.

![Fig. 8](image3)

Fig. 8—A maximum gain per wavelength per unit \( b \), plotted versus \( \beta_{od} \), the beam separation in radians, for parallel plane beams or for large concentric tubular beams.
Part II—Investigations of High-Frequency Echoes

H. A. HESS†

Summary—Oscillographic high-speed records of high-frequency telegraph signals (10 to 20 Mc) which show periodic variations in the field strength are investigated in this paper. A movement of the ionospheric reflector is considered as the cause of Doppler shifts, which are measured within the amplitudes of interfering radio waves between two consecutive minima. The long-distance propagation which usually occurs over multiple paths in the transmission, indicates shifts of 0.1 to 0.5 cps; 2.4 cps were measured at the interference of direct and indirect signal of VIS, Sydney, Australia, on 16,450 kc. Routes passing the auroral zones are characterized by the occurrence of "eject signals" with shifts between 5 and 30 cps.

I. INTRODUCTION

The author's previous paper contains investigations of time-interval measurements between long-distance signals and their echoes. It was intended to find an explanation for the highly constant values of a complete high-frequency circuit around the globe, which vary between 0.13760 and 0.13805 second. The complete circulating signal travels on paths between 41,280 and 41,420 km; about 1,300 km greater than the earth's circuit. Interferences of waves coming from multiple paths in the transmission, and the uniform variations of the phases are considered as the cause for divergences within the measured time intervals. The shape of the signals is influenced within fractions of 1 second, and this fact seems to be an occasion for errors in the measurements. An analysis of the signals in their individual wave components gives a new possibility to explore the kind of the ionospheric propagation.

Schmidt's theory of a sliding-wave propagation along an ionospheric limit layer has been discussed by Dieminger, Hamberger, and Rawer, as well as by Lassen. These scientists explain the high-frequency propagation over long distances, and the occurrence of signals which travel repeatedly around the globe, by multiple reflections between the F layer and the earth's surface.

According to Griffiths, Doppler shifts of about 0.5 cps were perceived at the normal frequency of WWV, 15,000 kc, National Bureau of Standards, Washington, D. C. during observations at Tatsfield, England, in December, 1945. The author found, by his recent studies, many measurable shifts on the interference of high-frequency signals, which were recorded on film rolls at Frederikshavn, 57°26'N, 10°29'E, and at Randers, 56°31'N, 10°02'E, during the period 1942 to 1945. The apparatus used were described in the previous paper. The accuracy for time-interval measurements, surveyed by a 500 cps normal frequency, was in the order of $5 \times 10^{-5}$ seconds, if the velocity of the films was greater than two meters per second.

II. INVESTIGATION OF RADIO-WAVE INTERFERENCES

An interference of two waves with the same frequency either effects an intensification, or a diminution of the resulting amplitude, according to the existing phase conditions. Morse signals are interrupted coherent waves. Their beginning and closing makes possible the analysis of multiple paths in the transmission, by measuring the time intervals between the successively arriving waves.

Fig. 1 shows an oscillographic record of the unmodulated signal of LSA3, Monte Grande, Argentina, 17,600 kc, on October 10, 1944, 11°54 Central European time (CET), at Randers, as an exquisite case of an involved multipath phenomenon. The direct signal arrived probably on four paths within time intervals of 0.4, 0.9, and 1.6 milliseconds between the component which arrives first, or detours of about 120, 270, and 480 km. The end of the direct signal is overlapped by the beginning indirect signal, which travels along the reverse great-circle path, and reaches the receiver after the time difference of $t_0 = 0.0552$ second, measured between the first component of the direct signal. Employing the equation:

$$d = u / 2(1 - t_0 / t_a),$$

where $u = 40,024$ km (earth's circuit), and $t_a = 0.13778$ second (time interval

† Decimal classification: R112.4. Original manuscript received by the Institute, July 13, 1948; revised manuscript received, March 18, 1949.
‡ Schad-Str. 24, 14a-ULM (Donau), Germany.

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of the complete high-frequency circuit), a distance of 11,995 km is obtained. The true distance between Monte Grande and Randers is 12,010 km.

Fig. 2—Direct signal of WK0, Rocky Point, N. Y., 15,970 kc. (Two paths with different phase conditions).

Fig. 2 is a record of the unmodulated direct signal of WK0 Rocky Point, N. Y., 15,970 kc on November 8, 1944, 1408 CET at Randers with two transmission paths. The first wave is characterized by a rather low field strength, while the second retarded component arrived after 0.35 millisecond, and represents a detour of 105 km between both paths. Since 15,970 kc is considered as a rather high transmission frequency with regard to the conditions of 1944 during the minimum of the sun-spot cycle, the propagation occurred under flat angles of incidence toward the horizon; 3° at two hops, and 9° at three hops between earth's surface and F layer coincide well with the measured detour of 105 km. The recorded case indicates opposite phases of the interfering waves, since the end of the second component is characterized by an increase of the amplitude. Variations of the phase conditions are generally observed within a few seconds, and minima alternating with maxima make possible the study of Doppler shifts on the film rolls which had a length of 10 meters.

These investigations demonstrate that measurements between the components, which arrive first, of the direct signals and the indirect circulating signals should only permit exact distance determinations. The weak first components frequently did not appear on all recorded signals of North American stations, probably because of the various and fortuitous sensitivity of the receiver. Their absence principally must be considered as the cause of the errors in the measurements which were found to diverge within a range of about 60 km.

![Fig. 3—Doppler-shift studies at the interference of direct and indirect signal of VIS, Sydney, Australia, 16,450 kc.](image)

Fig. 3 shows a record of two consecutive slightly modulate signals of VIS, Sydney, Australia, 16,450 kc on November 8, 1944, 1005 CET at Randers. Both signals consisting of two interfering waves represent the direct signals A-C and the indirect signal B-D with a measurable time difference of \( t_i = 0.0269 \) second. The distance between Sydney and Randers is 16,110 km. The not overlapped parts of the direct and the indirect signals have approximately equal field strengths, while previously and subsequently the amplitudes are different. Thus, a sharp minimum occurs during the interference period B-C on both recorded signals. The time interval of 0.42 second between these two consecutive minima means a Doppler shift of 2.4 cps in the frequency between direct and indirect signals.

Since a moved reflector is considered to effect the sinusoidal variations in the amplitudes of the interfering sky waves, the Doppler shifts are given by the equation:

\[
 f_d = \frac{v}{c} \cdot f,
\]

where \( f = 16,450 \) kc is the rf used, \( v \) generally expresses a velocity (in this case: the path difference within 1 second), and \( c = 299,776 \) km is the velocity of the electromagnetic waves. The measured frequency shift \( f_s = 2.4 \) cps is equivalent with a path difference of 43 meters per second. For a long distance propagation, however, the angles of incidence are oblique toward the vertically moved F layer, and the Doppler shift is given by the equation

\[
 f_d = \frac{2 \cdot v}{c} \cdot f \cdot \sin (\alpha + \gamma),
\]

where \( \alpha \) is the angle of incidence toward the horizon, \( \gamma \) the half angular width of a single hop between the earth's surface and a F layer reflection, and \( v \) the vertical velocity of the F layer.

The geographical distance between Sydney and Randers is about 16,000 km on the direct path, and about 24,000 km on the reverse path. Direct signal and indirect signals travels on ionospheric paths of 16,548 and 24,822 km, if the assumption of six and nine ionospheric reflections is made. The sum is 41,370 km, or a time of 0.1379 second for the completely circulating signal is well in accordance with the average value \( t_w = 0.13778 \) second.

For the case that the F layer should ascend along the direct path, and descend along the reverse path, 15 effective reflections may be regarded. The equation:

\[
 v = \frac{1}{15 \cdot c} \cdot \frac{f_d}{2 \cdot f \sin (\alpha + \gamma)}
\]

may approximately give an average value for the velocity of an individual point of the reflecting F layer. \( /\alpha = 4^°20' \), and \( /\gamma = 12^° \) were calculated for a direct propagation in six hops, like as for a reverse propagation in nine hops with regard to a F layer height of 250 km. A velocity of 5 meters per second is obtained for
the $F$ layer. This average value may be corrected by the experience that the velocity of the $F$ layer can not be homogeneous at every point along the direct and reverse path.

The ionospheric conditions actually are more involved than a simple calculation would sufficiently explain them. The direct and reverse great-circle path is illustrated in Fig. 4, together with the illumination of the globe on November 8, 10:00 CET. The points of probable ionospheric reflections which are marked with numbers along the direct and reverse path are distributed on different times of the day and seasons. According to the general experiences with the vertical incidence sounding, an ascending layer seems possible at the points 1, 2, 3, 4, and 5 along the direct path, while the movement at point 6 is probably a descending one. Along the indirect path, a slight ascending occurred at points 1, 2, 3, 4, and 5 (summer-night conditions) while certainly an effective descending occurred at points 6, 7, 8, and 9 because of sunrise.

In the case that the great-circle line between transmitter and receiver passes the auroral zone, strongly marked fluctuations within the signal amplitudes were observed. Fig. 5 represents the routes on the earth's globe which are characterized by the occurrence of "cleft signals." Filled-out points are positions on the northern hemisphere, and circles on the southern hemisphere. Both magnetic poles and the auroral zone on the northern hemisphere are marked. The auroral center has been found by many observers on polar lights to be in northwestern Greenland (78°N, 68°W), in the middle between the magnetic and geographical pole. Similar conditions are supposed on the southern hemisphere. Cleft signals are observed in Europe from stations in Hawaii, Alaska, and the western North America. They also occur on routes between eastern North America and East Asia, as well as between South Africa and New Zealand on the southern hemisphere.

Fig. 6 shows that three selected film records with cleft signals of KQF, Kailua, Hawaii, on 13,495 kc, made at Randers, Denmark. The signal amplitudes are characterized by a sinusoidal course between minima and maxima. (a) is the record of August 31, 1944, 08'07 CET. A sharp minimum and maximum of two interfering waves occurred within 80 milliseconds, indicating a Doppler shift of 6 cps. The signal itself is slightly modulated with 360 cps. (b) represents two deep minima in a record of August 24, 1944, 08'35 CET within 43 milliseconds, and indicates a Doppler shift of 23 cps.

Since the minima do not occur within the same time intervals, the signal might be shaped by more than two interfering waves. The distorted 360-cps signal modulation is also a reason to suppose an interference of multiple waves. (c) The record of August 15, 1944, 08'40 CET shows a pulse structure at the commencement of the signal, which lasts 0.4 millisecond, followed by a deep minimum. The maximum occurring after 75 milliseconds indicates a Doppler shift of 7 cps. The strong short pulse is actually the wave which arrives first, overlapped after 0.4 millisecond by the retarded component which has an opposite phase. With regard to

---

Fig. 4—Direct and reverse great-circle path between Sydney and Randers, and illumination of the globe on November 8, 10'00, Central European time.

Fig. 5—Routes across the auroral zone.

Fig. 6—Cleft signals of KQF, Kailua, Hawaii, 13,495 kc.

---

the frequencies used of 15 Mc, the measured Doppler shifts between 5 and 30 cps indicate velocities of 100 to 500 meters per second.

Indirect signals did not occur at KQF, Hawaii. This kind of echoes, however, was found on records of stations in California and Madagascar. The recording place, Frederikshavn, and these two positions are on the same great-circle line, which passes the proximity of the northern and the southern magnetic pole, as shown in Fig. 5. Figs. 7(a) and (b) are records of the direct and the indirect signals of KPH, Bolinas, Calif., 12,735 kc on January 20, 1942, 15°20 CET, and FZT, Tananarive, Madagascar, 17,890 kc on April 27, 1942, 09°30 CET. The cleft direct signal from California passed the northern auroral zone, and the cleft indirect signal from Madagascar passed the southern and the northern auroral zone. Doppler shifts of 15 and 12 cps were measured. It is striking that the indirect signal from California is normal, though it has passed the auroral zone on the southern hemisphere. This probably is an evidence that no interference of multiple waves occurred, as a case of a single-path propagation.

![Fig. 7](image)

**Fig. 7**—(a) Cleft direct signal and normal indirect signal from California; and (b) normal direct signal and cleft indirect signal from Madagascar.

A profile shown in Fig. 8 enables an investigation of the ionospheric transmission between Bolinas and Frederikshavn. The true distance is 8,571 km. With regard to the conditions in 1944, 12,735 kc was a rather high frequency, reflected from the $F$ layer under angles between 0° and 10° toward the horizon. The propagation is considered to occur along two paths, in three and four hops with angles of 3° and 9°. The detour between both paths is 100 km. One ionospheric reflector of the first path is situated within the auroral zone, while two reflectors of the second path are more distant from the turbulence zone. Both waves evidently are affected by shifts, and decline from the transmitter frequency.

### III. Conclusions

On the basis of these investigations, the long distance propagation is explained by repeatedly reflected waves between ionosphere and earth’s surface. An absorbing, refracting, and reflecting influence of the $E$ layer is possible on the long-distance propagation. It can be neglected for a transmission near the critical $F$ frequency, insomuch as no abnormal $E$ ionization occurs. The sun’s radiation effects the daily course of the $F$ layer ionization, connected with the variations of the virtual height. An $F$ layer velocity between 0 and 10 meters per second, concluded from Doppler shifts found in the long-distance propagation, coincides well with the experiences of the vertical incidence sounding. Signals unmodulated or slightly modulated only permit exact Doppler-shift measurements, while strongly modulated signals and pulses are unsuit. Cleft signals, characterized by shifts between 5 and 30 cps, indicate a multiple-path propagation and a strong turbulence of the ionosphere within the auroral zone, since a velocity of 100 to 500 meters per second cannot be explained with the regular uniformly ascending or descending movement of the $F$ layer in lower geographical latitudes.

A radiation of electrically charged particles, probably from the sun, which is deflected by the earth’s magnetic field to the zones of the magnetic poles, is regarded to cause this turbulence. Waves reflected from the ionosphere are split in ordinary and extraordinary components by the earth’s magnetic field; circularly polarized at the poles, and linearly at the equator. Probably the frequency shifts of both components reflected within the auroral zone are also different, and it seems, that they are much stronger absorbed after repeated ionospheric reflections. The formation of ionized layers within the auroral zone is much involved, and depends on many unknown factors, as shown by the occurrence of the sporadic $F$ layer.

Since the occurrence of Doppler shifts indicates a vertically moved ionospheric reflector, no definite limit exists between the $F$ layer and the vacuum. Therefore, there are no arguments which should confirm the “sliding-wave” conclusions, to explain the highly constant time intervals of repeatedly circulating signals.


Microwave Filter Theory and Design

J. HESSEL†, SENIOR MEMBER, IRE, G. GOUBAU†, AND L. R. BATTERSBY†

Summary—The first part of this paper gives the theory of waveguide filters with arbitrary identical links, and their matching to a line. The treatment differs from that previously presented in the literature, in that the electromagnetic state of the impedors and transducers is described by relations between the incident and reflected waves. This simplifies the analysis of all waveguide systems, because each transformation by a line section results only in a phase shift of these waves. Each filter stage is characterized by two angles which can be determined by simple measurements. The formulas which describe the insertion properties of the filters are given in terms of these angles.

The second part gives the application of the theory to direct and quarter-wave coupled band-pass iris filters. The design data are given, including correction factors for irises of finite thickness. Measured insertion loss curves of direct couples filters constructed, show good agreement with the theoretical.

TABLE OF SYMBOLS

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>a</td>
<td>wide dimension of a rectangular waveguide</td>
</tr>
<tr>
<td>a_{11}, a_{12}</td>
<td>coefficients of a transducer matrix</td>
</tr>
<tr>
<td>b, b_o</td>
<td>relative 3-dB bandwidth without and with dissipation loss; also narrow dimension of a rectangular waveguide</td>
</tr>
<tr>
<td>d_o</td>
<td>diameter of a circular iris</td>
</tr>
<tr>
<td>d_{11}</td>
<td>diameter of a slit iris</td>
</tr>
<tr>
<td>D</td>
<td>distance between two irises</td>
</tr>
<tr>
<td>f</td>
<td>frequency</td>
</tr>
<tr>
<td>f_o</td>
<td>center frequency</td>
</tr>
<tr>
<td>f_c</td>
<td>cutoff frequency of the waveguide</td>
</tr>
<tr>
<td>F_o, F_{11}</td>
<td>factor for calculation of circular irises</td>
</tr>
<tr>
<td>F</td>
<td>factor for calculation of slit irises</td>
</tr>
<tr>
<td>g</td>
<td>transmission constant</td>
</tr>
<tr>
<td>j</td>
<td>j = \sqrt{-1}</td>
</tr>
<tr>
<td>L</td>
<td>insertion loss</td>
</tr>
<tr>
<td>m</td>
<td>integer</td>
</tr>
<tr>
<td>n</td>
<td>number of stages</td>
</tr>
<tr>
<td>ρ</td>
<td>\frac{\cos \psi}{\cos \phi}</td>
</tr>
<tr>
<td>q</td>
<td>\frac{\sin \psi}{\sin \phi}</td>
</tr>
<tr>
<td>Q</td>
<td>quality of an unloaded cavity</td>
</tr>
<tr>
<td>\vec{u}, \vec{w}</td>
<td>amplitudes of guide waves</td>
</tr>
<tr>
<td>\vec{z}_0, \vec{z}_c</td>
<td>characteristic reflection factor of a transducer</td>
</tr>
<tr>
<td>\vec{z}_c</td>
<td>normalized impedance</td>
</tr>
<tr>
<td>\vec{z}_0</td>
<td>characteristic impedance of a waveguide</td>
</tr>
<tr>
<td>α, β, γ</td>
<td>factors characterizing the frequency response</td>
</tr>
<tr>
<td>δ</td>
<td>iris thickness</td>
</tr>
<tr>
<td>ε</td>
<td>factor characterizing the dissipation loss</td>
</tr>
<tr>
<td>θ</td>
<td>transmission phaseshift</td>
</tr>
</tbody>
</table>

\* Decimal classification: R143.2X R310. Original manuscript received by the Institute, July 22, 1948: revised manuscript received, March 31, 1949.

† Coles Signal Laboratory, Signal Corps Engineering Laboratories, Red Bank, N. J.

\begin{align}
\lambda &= \text{free-space wavelength} \\
\lambda_o &= \text{wavelength in a guide} \\
\lambda_c &= \text{cutoff wavelength} \\
\nu &= \text{half relative 3-dB bandwidth} \\
\sigma &= \text{factor characterizing the dissipation loss} \\
\tau &= \text{ratio between 3-dB bandwidth of n stage filter to 3-dB bandwidth of one single stage} \\
\phi &= \text{aperture of an iris, characteristic of a transducer} \\
\psi &= \text{phase length of a guide section} \\
\end{align}

step, the determination of equivalent circuits. This intermediate step is eliminated by the use of wave matrices. Some information about wave matrices may be obtained from the Radiation Laboratory Series. A more elaborate treatment of the wave matrix theory will be contained in a monograph now being published.

In the following the wave matrices are used for the analysis of transmission line filters with identical stages, including their matching transformers.

2. Characteristics of Symmetrical Transducers

There are several possible characterizations for a transducer. It is obvious that a characterization is desirable which is based on measurable quantities, such as reflection and transmission coefficients, and which is at the same time convenient for analysis. Thus we write the wave matrix for an arbitrary symmetrical transducer in the form:

\[
(a) = \frac{j}{\sin \phi} \begin{pmatrix} -e^{i\phi} & -\cos \phi \\ \cos \phi & e^{-i\phi} \end{pmatrix} - \frac{\pi}{2} < Re \phi < +\frac{\pi}{2}.
\]  (2)

The angles \(\phi\) and \(\psi\) are the quantities which may be introduced as characteristics of the transducer. They are real if the dissipation losses are negligible; otherwise they contain an imaginary part. In terms of \(\phi\) and \(\psi\), the reflection and transmission coefficients become:

\[
\begin{pmatrix} H_1 \\ H_1 \end{pmatrix}_{\text{for } \psi = 0} = -\cos \phi e^{-i\psi},
\]

\[
\begin{pmatrix} H_2 \\ H_1 \end{pmatrix}_{\text{for } \psi = 0} = j \sin \phi e^{-i\psi}.
\]  (3)

For negligible losses we see \(\phi\) is a measure of the absolute value of the reflection coefficient \(\cos \phi\) and of the transmission coefficient \(|\sin \phi|\). \(\psi\) is a measure of their arguments.

If a transducer consists of a plane thin iris only, \(\psi\) is equal to \(\phi\). Therefore, \(\phi\) can be used as the only characteristic of an iris, and it will be called in the following the "aperture" of the iris. \(\phi\) is connected with the normalized impedance \(z\) by the relation:

\[
z = \frac{j}{2} \tan \phi.
\]  (4)

It can be shown that every symmetrical transducer is equivalent to an iris with an aperture \(\phi\) (defined by (3)) and line sections on both sides, the length of which is \(\frac{1}{4}(\psi - \phi)\). This is also true if dissipation is considered, then the iris and the line sections must be dissipative, i.e., \(\phi\) and \(\psi\) complex.

If a transducer is inserted in a system, and the reflection factor measured on the output side of the transducer is \(w_z = \frac{H_2}{H_1}\), the transformed reflection factor \(w_1 = \frac{H_1}{H_1}\), which appears on the input side is, with regard to eq. (1) and (2):

\[
w_1 = \frac{\cos \phi + w_ee^{-i\psi}}{e^{i\psi} + w_1 \cos \phi}.
\]  (5)

There is a certain value of \(w_e = w_e\) which remains unchanged by the transducer; this means \(w_1 = w_2 = w_e\), \(w_1\), which may be called the "characteristic reflection factor" of the transducer, is found from (5) to be:

\[
w_e = - \rho \pm \sqrt{\rho^2 - 1} \text{ where } \rho = \frac{\cos \psi}{\cos \phi}.
\]  (6)

The transmission constant \(g\) of the transducer is defined by: \(\varepsilon^2 = \left(\frac{H_2}{H_1}\right)_{\text{for } w_1 = w_2 = w_e}\) becomes:

\[
\varepsilon^2 = q + j \sqrt{1 - q^2} \text{ where } q = \frac{\sin \psi}{\sin \phi}.
\]  (7)

Regarding the signs of the roots in (6) and (7), the following statements can be made: The upper sign of (6) corresponds to the upper sign in (7) if we state that \(\sqrt{\sin^2 \phi} = +\sin \phi, \sqrt{\cos^2 \phi} = +\cos \phi\). Furthermore, the law of conservation of energy stipulates that \(|w_e|\) can never be greater than 1, and the real part of \(g\) never positive.

If dissipation is negligible (\(\phi\) and \(\psi\) real) we see from (7) that \(g\) is purely imaginary for \(|q| < 1\) or \(\psi\) is between the limits:

\[
m\pi - |\phi| < \psi < m\pi + |\phi| \text{ for } m = 0, 1, 2, 3, \ldots .
\]  (8)

At the same time \(w_e\) is real (see (6)). If the transducer is loaded by a system the reflection factor of which is \(w_2 = w_1\), the passing wave undergoes only a shift in phase; its amplitude remains unchanged. In the range:

\[
(m + 1)\pi - |\phi| > \psi > m\pi + |\phi|,
\]  (9)

\(g\) contains a real part and \(w_e\) becomes complex with the absolute value 1. In this case the passing wave is attenuated, if the transducer is loaded with \(w_2 = w_e\). Equations (8) and (9) determine the passing and rejection ranges of the transducer.

The characteristic reflection factor of a transducer has a meaning similar to characteristic impedance in the usual four-terminal theory.

3. Filter with Identical Links

Consider a filter of \(n\) identical symmetrical transducers (Fig. 2). The equations for such a ladder may be

\[
\text{Fig. 2—Block diagram of a filter with identical stages,}
\]
written in the form:

\[
\begin{pmatrix}
\mathbf{u}_1 \\
\mathbf{u}_n+1
\end{pmatrix} =
\begin{pmatrix}
\mathbf{a}_{11} & \mathbf{a}_{12} \\
\mathbf{a}_{21} & \mathbf{a}_{22}
\end{pmatrix}
\begin{pmatrix}
\mathbf{w}_n \\
\mathbf{w}_{n+1}
\end{pmatrix},
\]

(10)

where \( \mathbf{u}_1, \mathbf{u}_n+1 \) are the waves on the input side and \( \mathbf{w}_n, \mathbf{w}_{n+1} \) those on the output side. Using the characteristics \( w, g \) introduced in the previous section, it can be easily shown that the \( a \) matrix of a \( n \)-stage ladder becomes:

\[
(a)_n = \frac{1}{w_n - \frac{1}{w_n}}
\]

If we introduce Tschebyscheff’s polynomials of the first kind, (see Table 1) we may write:

\[
\frac{1}{2} (e^{-n\phi} + e^{n\phi}) = \frac{1}{2} \left[ (q \pm j\sqrt{1 - q^2})^n \mp (q + j\sqrt{1 - q^2})^n \right],
\]

\[
= T_n(q),
\]

\[
\frac{1}{2j} (e^{-n\phi} - e^{n\phi}) = \frac{1}{2j} \left[ (q \pm j\sqrt{1 - q^2})^n - (q \mp j\sqrt{1 - q^2})^n \right],
\]

\[
= \pm U_n(q).
\]

<table>
<thead>
<tr>
<th>( n )</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
</tr>
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<tbody>
<tr>
<td>( T_n(q) )</td>
<td>( q )</td>
<td>( 2q )</td>
<td>( 4q^2 - 3q )</td>
<td>( 8q^4 - 8q^2 + 1 )</td>
</tr>
<tr>
<td>( U_n(q) )</td>
<td>( \frac{1}{\sqrt{1-q^2}} )</td>
<td>2q</td>
<td>( 4q^2 - 1 )</td>
<td>( 8q^4 - 4q^2 )</td>
</tr>
</tbody>
</table>

The matrix (3.2) becomes:

\[
(a)_n = \frac{1}{w_n - \frac{1}{w_n}}
\]

with

\[
q = \frac{\sin \psi}{\sin \phi}, \quad \rho = \frac{\cos \psi}{\cos \phi}
\]

If the filter is inserted in a line which is terminated by its characteristic impedance, the transmission constant \( g_n \) of the filter is given by:

\[
e^{\pm n\phi} = T_n(q) - j \frac{\rho}{\sqrt{\rho^2 - 1}} U_n(q)
\]

\[
= T_n(q) - j \frac{\cos \phi}{\sin \phi} \frac{U_n(q)}{\sqrt{1 - q^2}}.
\]

(14)

As long as the dissipation loss is negligible (\( \phi \) and \( \psi \) are real), (14) leads to the following expression for the insertion loss:

\[
L = 1 + \left( \frac{\cot \phi}{\frac{U_n(q)}{\sqrt{1 - q^2}}} \right)^2.
\]

(15)

The phase shift of the passing wave is:

\[
\tan \theta = -\frac{\rho}{\sqrt{\rho^2 - 1}} \frac{U_n(q)}{T_n(q)}.
\]

(16)

The reflection coefficient on the input side

\[
(w_1)_{n+1=0} = -\frac{1}{\rho + j \frac{T_n(q)}{U_n(q) \sqrt{\rho^2 - 1}}}
\]

(17)

If dissipation is to be taken in account, the formulas (15) to (17) become more complicated, because \( \rho \) and \( q \) are no longer real and the Tschebyscheff polynomials must be developed in order to separate the real and imaginary parts.

As may be seen from (15), the insertion loss vanishes (neglecting the dissipation losses) if \( \cos \phi = 0 \) or \( U_n(q)/\sqrt{1 - q^2} = 0 \). If each stage is nonreflecting for a certain frequency, \( \cos \phi \) is zero (see 3). This provides resonating filter links, which act like parallel LC circuits shunted to a two-wire line.

The formulas (13) to (17) describe the behavior of all transmission line filters with identical links, provided

\[
\frac{\rho}{\sqrt{\rho^2 - 1}} \frac{T_n(q) + j \frac{\rho}{\sqrt{\rho^2 - 1}} U_n(q)}{T_n(q) - j \frac{\rho}{\sqrt{\rho^2 - 1}} U_n(q)}
\]

(13)

that the frequency response of the characteristics \( \phi \) and \( \psi \) is known.

**Example 1**

We apply the formulas first to a band-pass iris filter with two identical nonresonating irises (Fig. 3) with
aperture $\phi$. Each filter stage consists of an iris with a guide section on each side, whose length is equal to half the distance between the irises. The angle $\psi$ in our formulas is equal to the transmission angle between the two irises $\psi$ plus $\phi$ (see Section 2).

![Diagram of a two-stage filter](image)

**Fig. 3—Two-stage filter, each stage consisting of an iris with the aperture $\phi$ and guide sections with the transmission angle $\psi/2$.**

Within a certain frequency range, the frequency response of $\phi$ and $\psi$ can be considered linear. Hence, we may write:

$$\phi = \phi_0 + \beta y, \quad \psi = \psi_0 + \gamma y, \quad \text{with} \quad y = \frac{f - f_0}{f_0}. \tag{18}$$

where $\beta$ and $\gamma$ are constants; the subscript 0 refers to the center frequency. $\psi_0$ has according to Section 2 (see (8)) the value $m\pi$. If dissipation losses are considered, $\phi_0$, $\beta$, and $\gamma$ are complex, and $\psi_0$ is $m\pi$ plus a small imaginary part. We calculate only the insertion loss for the dissipationless case when $m = 1$. Using (15) and Table I we get:

$$L = 1 + 4 \left( \frac{\cos \phi \sin \psi}{\sin^2 \phi} \right)^2 \tag{19}$$

This formula is not identical to that usually given for a 2 iris or one cavity filter, because it is not restricted to small bandwidths. For small bandwidths $\phi_0$ becomes small, and within a certain frequency range in the neighborhood of the center frequency, the term containing $\beta \gamma^2$ can be neglected. Thus, the insertion loss becomes quadratic in $y$.

**Example 2**

The second example which we will consider is a band rejection filter, consisting of cavities coupled by small holes in the side of a waveguide, separated by a distance of a quarter wavelength (Fig. 4(a)). If the coupling is effected by the electric field or the longitudinal component of the magnetic field of a TE-wave mode, and the field distortion is compensated by plungers (pl in Fig. 4(a)), each of these cavities behaves like a series tuned circuit in shunt to a line (Fig. 4(b)). According to (4) the aperture of an equivalent iris is given by the relation:

$$\frac{j}{2} \tan \phi = \frac{R}{Z_0} \left( 1 + j2Q \frac{f - f_0}{f_0} \right), \tag{20}$$

where $R$ is the resonance impedance of the equivalent circuit, $Q$ the quality of the cavity, $f_0$ its resonance frequency, and $Z_0$ the characteristic impedance of the waveguide.

![Diagram of a band rejection filter](image)

**Fig. 4—(a) Band rejection filter, each stage consisting of a cavity coupled to a guide section having a length $\psi$, pl indicates the compensating elements. (b) Single filter stage and equivalent circuit.**

Each filter link consists of a cavity and a guide section on both sides, the length of which is half the distance $\psi$ between two cavities. As $\psi = \psi + \phi$, $p$ and $q$ become:

$$p = \frac{\cos \psi}{\cos \phi} = \frac{\cos (\psi + \phi)}{\cos \phi} = \cos \psi - \sin \psi \tan \phi \tag{21}$$

$$q = \frac{\sin \psi}{\sin \phi} = \frac{\sin (\psi + \phi)}{\sin \phi} = \sin \psi \cot \phi + \cos \psi. \tag{21}$$

Because of (20) $\tan \phi$ can be written in the form:

$$\tan \phi = x - j\sigma \text{ with } x = \frac{4R}{Z_0} Q \frac{f - f_0}{f_0} \tag{22}$$

and $\sigma = \frac{2R}{Z_0}$.

As the distance between two cavities is a quarter wavelength at the center frequency, $\psi$ becomes:

$$\psi = \frac{\pi}{2} + \beta \frac{f - f_0}{f_0} = \frac{\pi}{2} + \alpha x \text{ with } \alpha = \beta \frac{Z_0}{4R}. \tag{23}$$
\( \beta \) denotes the frequency response of \( \psi \) and has the value
\[
\beta = \frac{1}{\sqrt{1 - \left( \frac{f}{f_0} \right)^2}} \quad (f_0 = \text{cutoff frequency of the (24) guide}).
\]

Equations (21) and (22) furnish all information necessary for the description of the filter properties.

Within the rejection range, the frequency response of \( \psi \) can be neglected in general. Under this condition \( \rho \) and \( q \) become:
\[
\rho = -\tan \phi = -(x - js), \quad q = \cot \phi = \frac{1}{x - js}. \quad (25)
\]
The insertion loss for an \( n \) stage filter is then
\[
L = |e|^2 = |T_n(q) + j \left( \frac{U_n(q)}{\sqrt{1 - q^2}} \right)|^2 \quad (\text{see (14)}).
\]

For \( n = 2 \), for instance, \( L \) becomes:
\[
L = |e|^2 = \frac{(x^2 + \sigma(\sigma + 2))^2 + 4(1 + \sigma)^2}{(x^2 + \sigma^2)^2}. \quad (27)
\]

4. Filters With Matching Transformers

In general, a filter with identical links requires matching transformers on both ends. These are unsymmetrical transducers of the same type, but applied in opposite directions. (See Fig. 5.) The characteristics of these transducers influence the insertion properties of the filter, and they can be designed for various requirements. In the following, we restrict our consideration to a type of transformer with the following properties:

1. No reflection in the total filter at the center frequency of the pass band.
2. Both transformers in series connected as shown in Fig. 6 shall have the same frequency response as one filter stage.

\[
\text{Fig. 5—Block diagram of a filter with matching transformers.}
\]

\[
\text{Fig. 6—Block diagram for the replacement of a filter stage by two matching transformers.}
\]

The center frequency of the pass band is defined as the frequency for which \( \psi = m\pi \) (see Section 2 (8)). If dissipation losses are taken in account, only the real part of \( \psi \) can be \( m\pi \). Transformers which satisfy these conditions have the matrices:
\[
(a) = \frac{j}{\sin \phi} \left( \begin{array}{cc} -e^{i(\pi/2 + \phi)} & e^{i(\pi/2 - \phi)} \\ \cos \phi e^{i(\pi/2 + \phi)} & e^{-i(\pi/2 + \phi)} \end{array} \right) \quad (28)
\]

if \( \phi' \) is related to \( \phi \) by:
\[
\cos \phi' = \frac{1}{\cos \phi} - \sqrt{\left( \frac{1}{\cos \phi} \right)^2 - 1} \quad (29)
\]
The upper subscript and the upper signs in (28) refer to the input transformer, and the lower subscript and signs to the output transformer. In order to prove that these transformers satisfy the two conditions above, we consider first the matrix \( (a)_n \) for the output transformer.

If the line connected with it is terminated by its characteristic impedance, the reflection factor on the filter side of the transformer is:
\[
w(\psi = \pi) = -(-1)^n \cos \phi' = -(-1)^n \left( \frac{1}{\cos \phi} - \sqrt{\left( \frac{1}{\cos \phi} \right)^2 - 1} \right). \quad (30)
\]

This is the characteristic reflection factor \( w_1 \) of the filter links for \( \psi = m\pi \). (See (6)). The matching transformer on the input side transforms \( w \) for \( \psi = m\pi \) into 0. Thus, condition 1 is satisfied. It may be mentioned that our consideration is not quite exact because of the imaginary part of \( \psi \) due to the dissipation losses. Its effect could be compensated by a small discontinuity in the guide on both sides of the filter which would give the required additional transformation.

To prove the second condition, we have to form the matrix product \( (a)_n \times (a)_n \). Considering the relations (29), the resulting matrix becomes identical with that for one filter link (2). Only the sign becomes opposite for \( Re[\phi] < 0 \); however, this is of no consequence, because it means only a phase shift of 180° for the waves on one side of the transformer.

The matrix for the filter including its matching transformers is found by the following: Because of property (2) each filter link can be replaced by a pair of matching transformers as indicated in Fig. 6. Therefore, the total filter can be considered as a chain of identical symmetrical links, each of which consists of a pair of these transformers, but in the inverse arrangement (Fig. 7).

\[
\text{Fig. 7—The filter of Fig. 5, each stage of which has been replaced by a pair of matching transformers.}
\]

The number of these stages is one greater than the number of stages of the unmatched filter. This means the two matching transformers together form an additional stage. The matrix for one of the new stages is:
The corresponding characteristic reflection factor $w'_r$ becomes, with regard to (29),

$$w'_r = -\tan \phi \left( \cot \psi + \frac{1}{\sin \psi} \sqrt{1 - q^2} \right);$$

$$q = \frac{\sin \psi}{\sin \phi}. \quad (32)$$

The transmission constant $g'$ is the same as for the original stages, because they differ only in the succession of the matching transducers of which they may be thought to consist.

Because the total filter is transformed into a ladder of identical stages, its matrix is of the type (11), $w_r$ is replaced by $w'_r$, and $n$ is one more than the number of original stages. With (34) and (35), the matrix can be written in the form:

$$(a)_n = \left[ \begin{array}{c}
T_n(q) - j \cos \psi \frac{U_n(q)}{\sqrt{1 - q^2}} \\
jq \cos \phi \frac{U_n(q)}{\sqrt{1 - q^2}}
\end{array} \right].$$

The transmission constant is given by:

$$\sigma_n = T_n(q) - j \cos \psi \frac{U_n(q)}{\sqrt{1 - q^2}}; \quad (34)$$

the insertion loss in the dissipationless case:

$$L = 1 + \cos \phi \left( q \frac{U_n(q)}{\sqrt{1 - q^2}} \right)^2; \quad (35)$$

the phase shift $\theta$ of the passing wave and the reflection coefficient $w_1$ on the input side

$$\tan \theta = -\cos \psi \frac{U_n(q)}{\sqrt{1 - q^2}} \frac{1}{T_n(q)}$$

$$w_1 = -\frac{q \cos \phi}{\cos \psi + \frac{T_n(q)}{U_n(q)} \sqrt{1 - q^2}} \quad (37)$$

**Example:**

We apply these general formulas to a special type of bandpass filter (shown in Fig. 8) for which a simplified

![Fig. 8—Band-pass filter (direct coupled type).](image)

design procedure is given in part II. Each filter stage consists of a nonresonating iris with guide sections of the length $\psi/2$ on both sides. The apertures of the irises may

be small, so $\sin \phi$ can be replaced by $\phi$ and the frequency response neglected. The matching transformers also consist of irises with guide sections on both sides, but these are of different length ($\psi_1/2$ and $\psi_2/2$).

The required aperture $\phi'_r$ of the matching iris is given by (29). The wave matrices of this iris $(a)_r'$ and the matrices of the line sections $(a)_{\gamma}$ and $(a)_{\gamma}$ are:

$$(a)_{\gamma} = \frac{j}{\sin \phi} \left( -e^{i\phi'} - \cos \phi \right)$$

$$(a)_{\gamma} = e^{-i\phi} \begin{pmatrix} 0 \\ 0 \end{pmatrix}. \quad (38)$$

If we form the product $(a)_{\gamma} \times (a)_{\gamma} \times (a)_{\gamma}$ and compare the resulting matrix with the matrix (28) of the input transformer we find $\psi_2$ and $\psi_1$ to be:

$$\psi_2 = \frac{\pi}{2} - \phi', \quad \psi_1 = \psi - \phi'. \quad (39)$$

These relations should be satisfied over the entire frequency range. Actually they are only valid for one frequency. However, this is of no serious consequence for the following reasons: For small relative bandwidths the frequency response of $\phi$ and $\phi'$ can be neglected. (See example I, Section 3.) The frequency response of the line section $\psi_2/2$ which is continued by the homogeneous guide, does not affect the insertion loss. $\psi_1$ differs from $\psi$ by only a few per cent if $\phi'$ is small. Therefore, the frequency response of $\psi_1$ is approximately the same as that of $\psi$. Hence, if (39) is satisfied for the center frequency, it holds over a relatively wide frequency range.

For sin $\phi$ cos $\phi$, sin $\psi$, and cos $\psi$, we can use the approximations:

$$\sin \phi \approx \phi \quad \cos \phi \approx 1$$

$$\sin \psi \approx (-1)^m \frac{\beta}{\sin \beta} f \quad (\text{for } \beta \text{ see } (24)). \quad (40)$$

In the following we assume $m = 1$; this means that the distance between the irises should be about one-half wavelength.

For the moment we will neglect the dissipation losses. Then the insertion loss given by (35):

$$L = 1 + \left( x \frac{U_n(x)}{\sqrt{1 - x^2}} \right)^2 \quad \text{with}$$

$$x = -q \frac{\beta}{\phi} \frac{f}{f_0}. \quad (41)$$
\( n \) is the number of inner irises plus one, or in other words it is the number of resonating cavities.

The transmission angle \( \theta \) and the reflection factor \( w_1 \) are obtained from (36) and (37):

\[
\tan \theta = \frac{U_n(x)}{\sqrt{1 - x^2} T_n(x)}
\]

\[
w_1 = \frac{x}{1 - j T_n(x) \sqrt{1 - x^2}}
\]

These formulas are identical with those for the so-called quarter-wave coupled type, which is fully discussed in the literature.\(^3\) The close relationship between the two different filter types is founded in the fact that the quarter-wave coupled type gives the iris performance of the general filter type shown in Fig. 7, in which each original filter stage is replaced by a pair of the matching transducers. Each stage of the original iris filter (Fig. 8), consisting of an iris with the aperture \( \phi \) and the line sections \( \psi \), is replaced by two matching irises in the distance \( \psi \), with guide sections of \( \psi \) on the ends (Fig. 9).

\[\text{Fig. 9—Band-pass filter (quarter-wave coupled type).}\]

We require in Part II the relative 3-db bandwidth \( b \) for \( n = 1, 2, 3 \) or 4 stages. It is given by the solution of the algebraic equation

\[\left( \frac{x}{\sqrt{1 - x^2}} \right)^3 = 1. \quad (44)\]

For \( n = 1 \), the solution is: \( x = 1 \). Hence with (41) and (24)

\[b_1 = 2 \phi = 2 \phi \left( 1 - \left( \frac{f}{f_0} \right)^n \right).\]

The ratio \( \tau \) between the relative 3-db bandwidth for \( n \) through 4 inclusive and \( b_1 \) is given in the following Table II.

<table>
<thead>
<tr>
<th>( n )</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
</tr>
</thead>
<tbody>
<tr>
<td>( b )</td>
<td>0.707</td>
<td>0.762</td>
<td>0.827</td>
<td></td>
</tr>
</tbody>
</table>

Because the dissipation losses are small, they can be considered by assuming that only \( \psi \) has a small imaginary part—\( j \epsilon \). Then, with (40)

\[y = \frac{\sin \psi}{\sin \phi} = -x + j \epsilon = -x + j \sigma. \quad \rho \simeq 1. \quad (45)\]

Hence the transmission constant given by (34) becomes

\[n = 1 \quad e^{\epsilon n} = -x + j(1 + \sigma)\]

\[n = 2 \quad e^{\epsilon n} = [2x^2 - (1 + 2a + 2\sigma)] - j2x(1 + 2\sigma)\]

\[n = 3 \quad e^{\epsilon n} = - \left[ (4x^2 - x(3 + 8\sigma + 12\sigma^2) + j4x^2(1 + 3\sigma) - (1 + 3\sigma + 4\sigma^2 + 4\sigma^4) \right]\]

\[n = 4 \quad e^{\epsilon n} = [8x^4 - 8x^2(1 + 3\sigma + 6\sigma^2) + (1 + 4\sigma + 8\sigma^2 + 8\sigma^3 + 8\sigma^4)] - j8x^2(1 + 4\sigma) - x(4 + 16\sigma + 24\sigma^2 + 32\sigma^3)\]

\[\sigma \text{ can be expressed by the } Q \text{ of a single cavity and the relative 3-db bandwidth } b_1 \text{ (without dissipation):}\]

\[\sigma = \frac{1}{b_1 Q}. \quad (47)\]

From the relations (46), the corresponding insertion losses can be calculated by taking the sum of the squares of the real part and the imaginary part of \( e^{\epsilon n} \).

**PART II. APPLICATION TO FILTER DESIGN**

1. **Design Procedure**

   **A. Introduction**

   On the basis of the theory developed, the following presents design data for band-pass iris filters which give in simplified form, the information necessary for the design of band-pass filters of the types shown in Fig. 10,

   ![Diagram showing the relation between iris distances \( \psi \) and iris aperture \( \phi \), for direct and quarter-wave coupled filters. \( \phi' \) of the type (b) is identical with \( \phi' \) of the type (a).](image.png)

   Where \( n \) is the number of resonating cavities. For \( n \geq 2 \) there are two types of filters which have approximately the same insertion properties. Type (b), is the well-
known quarter-wave coupled type. The type (a), the direct coupled type, has seen little application to date, primarily because of more rigid design and manufacturing requirements. However, this type has the advantages of compactness and somewhat lower dissipation loss. Therefore, the design data presented have been examined especially for this type of filter, and the results are gratifying. The agreement between calculated and measured values is within 5 per cent.

The characteristic variables for the filter are the angles \( \phi \) and \( \psi \), introduced in the theoretical discussion, Part I. \( \phi \) is a measure for the aperture of an iris. The cos \( \phi \) and \( |\sin \phi| \) are the absolute values of the reflection and passing coefficients of the iris, in a guide terminated by its characteristic impedance. It is assumed that \( \phi \) is constant within the passing range of the filter and in the neighborhood of it. For the direct coupled type, the apertures of the irises on the ends, denoted by \( \phi' \), differ from the inner irises \( \phi \). The irises of quarter-wave coupled filters are identical, and have the same apertures as the end irises of the equivalent direct coupled filters. The irises of the one-stage filter are also denoted with \( \phi' \), because the design procedure is identical to that of the end irises. \( \psi \) and \( \phi \) determine the transmission angle \( \varphi \) between two irises. The values for \( \varphi \) indicated in Fig. 1 are referred to the center frequency.

**B. Number of Stages**

The first step in designing a filter is the choice of the proper number of stages. This is done with the aid of

The frequency scale in Fig. 11 is given in steps of \( \nu \). The curves are calculated with (46) for \( \sigma = 0 \). If, for instance, an attenuation of 20 db is desired for a frequency distance \( 3 \nu \) from the mean passing frequency, the number of stages must be at least 2.

**C. Apertures**

The apertures of the inner irises of the filter type (a) are found with the aid of the unbroken curves of Fig. 12. \( \phi \) indicates the aperture in degrees, \( b \) the relative 3-

**D. Dimension of the Windows**

The first steps require no special knowledge as to the kind of waveguide, wave mode, and form of irises. The following data are restricted to rectangular waveguides with \( TE_{10} \) mode excitation, slit-type irises with the slit
parallel to the electric field and circular irises (see Table III).

For infinitely thin irises of these types, the relation between \( \phi \) or \( \phi' \) and the iris dimensions can be calculated with the assumption of a quasi-stationary field distribution adjacent to the openings. This assumption holds with great accuracy as long as \( \sin \phi \) can be replaced by \( \phi \). This is also the supposition made for the formulas above. The width \( d_1 \) of the slit in the case of slit-iris corresponding to aperture in degrees is given by the formula:

\[
d_1 = \frac{\phi}{F_{11}} \frac{\lambda}{\lambda_e}, \quad \text{where} \quad F_{11} = \frac{a}{45\pi}.
\]  

The hole diameter \( d_0 \) for a circular iris is given by the formula:

\[
d_0^2 = \frac{\phi}{F_0} \frac{\lambda}{\lambda_e}, \quad \text{where} \quad F_0 = \frac{a^2b}{120}.
\]

The values of \( F_{11} \) and \( F_0 \) have been calculated for all approved guide types which are shown in Table III. Because the inside guide dimensions \( a \) and \( b \) are given in inches, \( d_1 \) and \( d_0 \) in the formulas above result also in inches. \( \lambda_e/\lambda \) is plotted in Fig. 12 as a function of \( d_0/f_0 \).

It is necessary to make a correction for the finite thickness of the irises. Because the irises with circular holes can be manufactured simply and with high accuracy, this type is preferable to the slit type for most applications; therefore, the correction for finite thickness has been measured only for this type. Fig. 13 shows a curve for this correction; \( d_6 \) is the hole diameter of an infinitely thin iris which is calculated with the design formula above. \( d_6' \) is the diameter of an iris with the thickness \( \delta \) which has the same coupling effect as \( d_6 \). The curve in Fig. 13 gives \( d_6'/d_6 \) as a function of \( \delta/d_6 \). If the diameter \( d_6 \) for the infinitely thin iris is determined and the thickness of the material to be used for the iris is known, \( \delta/d_6 \) can be calculated. Fig. 13 gives the correction factor \( d_6'/d_6 \) with which \( d_6 \) must be multiplied to obtain the real diameter \( d_6' \). It should be noted here that allowance for the decrease of hole diameter and increase of thickness by plating must be made when the iris dimensions are determined.

E. Distance Between the Irises

The transit angles \( \varphi \) between the irises for the center frequencies are given in Fig. 10. The real distances \( D \) in cm or inches are given by the relation

\[
D = \frac{\lambda_e}{360} \frac{\varphi}{\lambda_e} = \frac{\lambda_e}{360} \left( \frac{\lambda_e}{\lambda} \right) \varphi.
\]

\( \lambda_e/\lambda \) is plotted in Fig. 12. The values of \( \lambda_e \) are found in Table III.

### Table III

<table>
<thead>
<tr>
<th>Outside Dimension (&quot;</th>
<th>Inside Dimension (&quot;)</th>
<th>Cut Off Frequency Me</th>
<th>Cut Off Wavelength (cm in)</th>
<th>( F_0 ) (d_e in &quot;)</th>
<th>( F_\pi ) (d_0 in &quot;)</th>
</tr>
</thead>
<tbody>
<tr>
<td>a b</td>
<td>a b</td>
<td></td>
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</tr>
<tr>
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<td>0.420</td>
<td>0.170</td>
<td>14080</td>
<td>2.13</td>
</tr>
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</table>
As the distances are critical (except those corresponding to \( \psi_q \) in the quarter-wave coupled type), and the guide cross section may vary somewhat, it is better to make them smaller by approximately 1 per cent, and to tune the cavities to resonance by means of a capacitive plunger.

F. Dissipation Loss

The dissipation affects the insertion loss of the filter, and also to some extent the bandwidth. The unbroken curves in Fig. 14 show the insertion loss at the center frequency for 1, 2, and 3 stages. The dissipation losses are expressed by the quantity \( 1/bQ \), where \( b \) is the relative 3-db bandwidth for the dissipationless case, and \( Q \) the quality of a single cavity. The broken curves in Fig. 14 show the ratio between the modified relative bandwidth \( b_d \) and \( b \) as a function of \( 1/bQ \). The change of bandwidth because of the dissipation losses can be considered by designing the filter for a bandwidth which is the desired bandwidth divided by \( b_d/b \).

2. Example

The following example is given to illustrate the design procedure:

Three-stage filter of the type (a): A center frequency \( f_0 = 9,350 \text{ Mc} \), and 3-db bandwidth of 13 Mc is chosen. (The measured bandwidth will be a few per cent less than 13 Mc because of the dissipation losses previously discussed.) The type of guide used will be (1 inch \( \times \frac{\lambda}{4} \) inch) for which Table III gives the cutoff frequency \( f_s = 6,560 \text{ Mc} \). First, the values of \( f_s/f_o \) and the relative bandwidth \( b \) must be calculated. They become \( f_s/f_o = 1.43; b = 1.39 \times 10^{-4} \). \( \phi \) may now be determined from Fig. 12. With \( f_s/f_o = 1.43 \) and \( n = 3 \), \( \phi^{(0)}/b = 231^\circ \) or \( \phi^{(0)} = 0.325 \). By (49) we find \( \phi^{(0)} \) to be \( 6.1^\circ \).

Table III gives the value of \( F_o \) for 1 inch \( \times \frac{\lambda}{4} \) inch guide as \( 2.70 \times 10^{-4} \), and \( (d^o)^2 = 16.3 \times 10^{-4} \). Fig. 12 shows \( \lambda_0/\lambda_e \) to be 0.98. With (51) we find \( (d^o)^2 \) to be \( 0.860 \times 10^{-4} \) and \( (d^o)^2 = 16.3 \times 10^{-4} \). Hence: \( d_o = 0.095 \text{ inch} \) and \( d^o = 0.252 \text{ inch} \). For an iris thickness \( \delta \) of 0.009 inch (including plating) \( \delta/d_o = 0.09 \) and \( \delta/d^o = 0.32 \). Fig. 13 gives the correction factors for the iris diameters: \( d_o'/d_o = 1.10 \) and 1.04, respectively. The real diameters are therefore \( d_o' = 0.105 \text{ inch} \) and \( d^o' = 0.262 \text{ inch} \). To find the distance \( D \) between the irises, we use (52), first determining \( \psi_1 \) and \( \psi_2 \) from Fig. 10; \( \psi_1 = 176.8^\circ \), \( \psi_2 = 179.7^\circ \). With \( \lambda_0/\lambda_e = 0.98 \) and \( \lambda_e = 1.80 \text{ inches} \) (see Table III) we get \( D_i = 0.864 \), \( D_k = 0.882 \). To allow for tuning of the filter, we subtract approximately 1 per cent from these values to obtain the real distances.

3. Measurements

A two-stage and three-stage filter were constructed to verify the theory. The two-stage filter was designed for a bandwidth of 14 Mc \( (b = 1.50 \times 10^{-4}) \), centered at 9,350 Mc. The three-stage filter was constructed using the dimensions obtained in the foregoing example, 13 Mc bandwidth \( (b = 1.39 \times 10^{-4}) \), and center frequency of 9,350 Mc. Fig. 15 is a plot of the measured insertion loss of these filters. The curves shown were calculated with (46) for best fit to the measured points. The corresponding values of \( b \) and \( \sigma \) for the two-stage filter are \( b = 1.47 \times 10^{-4}, \sigma = 0.112 \) \( (Q = 4.300) \); and for the three-stage filter \( b = 1.25 \times 10^{-4}, \sigma = 0.097 \) \( (Q = 6.300) \). It should be noted here that the bandwidth of the three-stage filter is about 10 per cent less than 14 Mc. Approximately one-half of this deviation is due to dissipation losses which are considered in Section 1-F (for the two-stage filter the effect of dissipation is negligible). Thickness of plating and the accuracy of measurement account for the remainder.

The measured dissipation loss at the center frequency for the three-stage filter is relatively low, compared with that of the two-stage filter. This may be due to variation in the plating of the cavities. Another reason can be the following: The dissipation loss of an iris becomes greater for larger apertures. In the two-stage filter, each cavity has one large and one small aperture. The three-stage
filter, however, has one cavity with only small apertures, and therefore, the mean Q of the three-stage filter should be higher than that of the two-stage filter. The

tuning of a three-stage filter can be accomplished easily in the following manner: First, one outside cavity is approximately tuned to resonance by noting the reaction on the generator. This is done to obtain some power flow through the filter. The inner cavity is then tuned far off resonance. Following this, the last cavity is tuned for maximum power transfer, and the first cavity readjusted to maximize the output. Finally, the inner cavity is tuned, and no further adjustments are required.

ACKNOWLEDGMENT

Acknowledgment is due to R. E. Lacy, Chief, Radio Relay and Microwave Section, without whose encouragement this paper would not be possible; to C. E. Sharp and A. Meyerhoff for many valuable suggestions which facilitated preparation of the material; and to A. J. Colaguori, who performed much of the technical work.

Stabilization of Simultaneous Equation Solvers*

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Summary—A new stability criterion for multiple-loop feedback systems is developed and applied to the problem of stabilizing electronic simultaneous equation solvers.

THE SOLUTION of a vast number of scientific and engineering problems involves the numerical computation of the unknowns of a system of linear simultaneous equations. Systems with as many as fifteen or twenty unknowns are frequent in many applications. The numerical solution of such systems, even by approximation methods, constitutes a formidable task. On the other hand, many engineering problems demand solutions less accurate than about 1 per cent. Within this limit of accuracy, the problems in question have been found to lend themselves well to solutions by analogue computers of the type described.1-4 In these devices, a physical quantity, such as a voltage, is made to correspond to each unknown by the introduction of convenient scale factors. The physical quantities representing the unknowns are then made to satisfy a set of equations analogous to the given problem. This is achieved by means of "computing elements" used to establish the desired relations between the various voltages. Once the computer is thus set up for a given problem, these voltages must then satisfy the relations established between them and must, therefore, be proportional to the unknowns. Such analogue computers are vastly cheaper than digital computers, and their accuracy is sufficient for many engineering applications.

The purpose of the computers in question is to find the set of numbers (unknowns), \( x_1, x_2, \ldots, x_n \), satisfying the set of linear simultaneous equations

\[
\begin{align*}
\alpha_1 x_1 + \alpha_{12} x_2 + \cdots + \alpha_{1n} x_n + b_1 &= 0 \\
\alpha_{n1} x_1 + \alpha_{n2} x_2 + \cdots + \alpha_{nn} x_n + b_n &= 0
\end{align*}
\]

whose determinant must be different from zero. In the following, we shall consider the case of real coefficients, \( a_k \) and \( b_t \), so that the unknowns will be real as well. It should be mentioned in passing that it will not prove difficult to extend our considerations to the case of complex coefficients.

In the computers to be discussed, the \( x_i \)’s are considered as variables represented by ac or dc voltages which are proportional to the respective numerical values. We shall denote these voltages by the same symbols \( x \), as the respective variables.
Fig. 1 shows the essential features of an analog computer using feedback to establish the relations (1) between voltages $x_i$ for the case $n = 2$. An automatic feedback computer of this type is described in footnote reference 2. It is seen that voltages corresponding to the terms $b_i$ and $a_{ik}x_k$ are obtained by means of potentiometers and summed by means of summing amplifiers. If $e_i$ be the output voltage of the $i$th summing amplifier, the device simulates the relations

$$ A \left\{ \sum_{k=1}^{n} a_{ik} x_k + b_i \right\} = e_i; \quad i = 1, 2, \ldots, n \quad (2) $$

which differ from the desired relations (1) only by the amount of the error or residual $e_i$ in each equation. The computer then attempts automatically to minimize these residuals by means of $n^2$ feedback connections such that $e_i = x_i$, as shown in Fig. 1. The relations (2) established by the computer now become

$$ \sum_{k=1}^{n} \left( a_{ik} - \frac{\delta_{ik}}{A} \right) x_k + b_i = 0 \quad i = 1, 2, \ldots, n \quad (3) $$

Equation (3) approximates (1) for high amplifier gain $A$. The error due to finite gain is easily seen to be inversely proportional to the gain $A$ and can be made very small. The relations (3) are satisfied by approximately the same values of the $x_i$ as the original equations (1). The voltages $x_i$ appearing in the arrangement of Fig. 1 will, therefore, be proportional to the desired unknowns if and only if, the feedback system for the given set of equations is stable at all frequencies.

Some conditions determining the stability of feedback systems of the type in question now will be discussed. If it were possible to construct the summing amplifiers so that the gain $A$ is a negative constant for all frequencies of the input voltages, the stability of the system would depend only on the nature of the feedback networks. The net effect of the latter should be degenerative. That is to say, it should actually minimize the sum of the squares of all the residuals $e_i$. It is easily seen that this need not be true in every case by considering the arrangement of Fig. 1. Regeneration would surely result if $a_{11} = a_{21} = -1$ and $a_{12} = a_{22} = 0$. The nature of the feedback networks depends on the given values of the coefficients $a_{ik}$. Specifically, it was shown in footnote reference 2 that, for constant negative amplifier gain $A$, the system will be stable if, and only if, the matrix of the coefficients $a_{ik}$ is positive definite. This is the case if, and only if, the equation

$$ \text{determinant } | a_{ik} - \delta_{ik} \lambda | = 0 \quad \delta_{ik} = \begin{cases} 0 & \text{for } i \neq k \\ 1 & \text{for } i = k \end{cases} \quad (4) $$

has only roots $\lambda (j=1, 2, \ldots, n)$ with positive real parts. This condition of positive definiteness is not as stringent a limitation as it might appear. Many systems of equations related to engineering problems have positive definite matrices for physical reasons. Again, the matrix of a given set of equations can often be made positive definite by simply rearranging the equations so that the diagonal elements are large and the post-diagonal coefficients are smaller. The diagonal coefficients are then all made positive by appropriate multiplications by $-1$. The reference cited also mentions that it is useful to arrange the equations so that, for each pair of indices $i$ and $k$,

$$ |a_{ik} a_{sk}| > |a_{ik} a_{si}|. \quad (5) $$

In any case, for any given system of linear equations (1) there exists an equivalent linear system

$$ \sum_{i=1}^{n} \sum_{k=1}^{n} a_{ij} a_{ik} x_k + \sum_{k=1}^{n} a_{ij} b_k = 0, \quad j = 1, 2, \ldots, n. \quad (6) $$

This set of equations is satisfied by the same unknowns and its matrix is necessarily always positive definite. Adcock has made the relations (6) the basis of a feedback computer which is stable for all values of the coefficients $a_{ik}$.

In the discussion which follows, it will be assumed that the coefficients $a_{ik}$ satisfy condition (4). It will then be found that almost all problems involved in the design of feedback computers of the type just described center about the design of the summing amplifiers. In practice, the amplifier gain $A$ is not a negative constant, as assumed above, but a complex function $A(j\omega)$ of the angular frequency $\omega$. Both the absolute voltage gain and the phase shift of each amplifier vary with frequency, and one must be sure that they vary in such a manner that the computer is stable at all frequencies. In other words, the amplifiers must be designed so that the feed-

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back cannot become regenerative at any frequency, since this would result in uncontrolled oscillations.

With the performance equations of the computer given in (3), the general stability criterion is that all the roots \( p = j\omega \) of the so-called characteristic equation

\[
\text{determinant} \left| a_{ik} \right| = 0
\]

must have negative real parts.

The necessity of maintaining stability imposes severe restrictions on the amplifiers to be used in feedback computers. The only known practical means of analyzing the stability of a complicated multiple-loop feedback system on the basis of (7) are Routh's rule and Nyquist analysis. Both necessitate cumbersome computations for values of \( n \) larger than 2.

The writer was, however, able to develop a simple condition for stability which is useful in the case of positive definite matrices, and, therefore, applies to the present computer. An inspection of (4) and (7) shows that (7) can hold true only for such values of \( p \) which satisfy one or more of the equations

\[
\frac{1}{A(p)} = \lambda_i
\]

where the \( \lambda_i \) are the roots in (4), often called the "eigenvalues" of the matrix of the \( a_{ik} \). Equation (7) is, then, equivalent to the \( n \) simpler equations (8).

The general stability criterion, therefore, reduces to the much more easily tested requirement that for a stable computer, all roots \( p \) of the \( n \) equations (8) must have negative real parts. This criterion holds whether or not (4) is satisfied by the coefficients \( a_{ik} \).

Assuming now that the matrix of the \( a_{ik} \) is positive definite, so that (4) holds true, it is not necessary to apply the above stability criterion to each new combination of coefficients \( a_{ik} \), or every given problem. If (4) is to hold, the \( \lambda_i \) must have positive real parts, which may vary in value depending upon the nature of the problem. They will, as a matter of fact, never exceed the value 1 since that is the maximum feedback factor possible with the potentiometer arrangement shown. Thus, the computer will be stable for all values of the coefficients \( a_{ik} \) satisfying condition (4) if, and only if, real part of \( p < 0 \) for all \( p \) such that \( 0 < \text{real part of } 1/A(p) < 1 \).

It is notable that this condition is independent of the number of amplifiers or equations. Therefore, in order to find out whether a given type of amplifier will yield a stable computer of the type considered, it is only necessary to perform analytical or experimental tests on one amplifier with simple feedback, which will solve the equation

\[
\text{ax} + b = 0 \text{ by the approximation } A(ax + b) = x
\]

or

\[
x = b \frac{A}{1 - aA} \approx \frac{b}{a}
\]

where \( a \) is now a complex number with positive real part (passive feedback networks with less than 90° phase shift).

If this simple system, shown in Fig. 2, proves stable for all values of \( a \) between 0 and 1, the same will be the case with a computer for \( n \) equations using the same type of amplifier if the matrix of the coefficient \( a_{ik} \) is positive (except for the possible effects of stray coupling, etc.). The fundamental importance of the last-mentioned theorem lies in the fact that it reduces the design of amplifiers for the complicated multiple-loop feedback system to the design of one simple feedback amplifier, so that stability criteria known from experience or a simple Nyquist analysis may be applied. The analysis described can be extended to cases in which the various amplifiers are not exactly identical as has been assumed above. Other more general theorems useful for the design of multiple-loop feedback amplifiers and servomechanisms may be derived from (8).

**Acknowledgment**

The author wishes to thank W. Prager, director of the Graduate Division of Applied Mathematics, Brown University, Providence, R. I., for making available the funds under which the subject research was conducted.
Novel Multiplying Circuits with Application to Electronic Wattmeters

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Summary—The primary relationship underlying this investigation is the exponential law relating plate current to plate voltage in a diode when operated in the so-called "retarding-field" region. By extending this classical mode of operation to multigrill tubes, the tube yielded further operating conditions in which plate current is accurately proportional to the product of a linear function of plate voltage and an exponential function of grid voltage over wide ranges.

An analysis is given of various multiplying circuits using a single multigrill tube under this mode of operation. Two methods are described for compensating accurately the inherent plate rectification due to the exponential grid curve.

The circuits are most suited for electronic wattmeters having exceptionally good features and predictable performance over the frequency range from 20 cycles up to the neighborhood of 50 Mc.

I. INTRODUCTION

For direct measurement of power at the common frequencies of power circuits, the electro-dynamometer type of wattmeter is satisfactory, but for direct measurement of small amounts of power, or for the measurement of power at the high audio frequencies or at radio frequencies, such an instrument cannot be used because of its appreciable insertion loss, inductive and capacitive qualities. For this reason, electronic wattmeters have been developed and used. With one exception, these depend upon the use of two accurately matched tubes operating in the square-law regime. These circuits have the disadvantage that it is difficult to match the tubes sufficiently accurately and that, even when matched, the square-law relation holds over such a small range that voltage dividers are unavoidable for voltages exceeding one or two volts. It is difficult to correct the angles of these dividers at high frequencies, and the power absorbed by the divider may constitute a serious loss. The insertion unit in these instruments is a T or π configuration, and it is difficult to provide for a satisfactory ground system with these configurations.

Another type of electronic wattmeter is described by J. R. Pierce, in which a multigrill tube of the hexode type is used. In these instruments, the multigrill tube is arranged to operate in a regime so that its plate current versus the voltage of the first grid for various values of the voltage on the third grid is a family of straight lines, which when extended meet at one point; while at the same time the plate current varies linearly against the voltage of the third grid. Existing multigrill tubes have such a small range of linearity that the error introduced by rectification is serious enough to preclude the use of one tube in a wattmeter. When two tubes are used in a push-pull arrangement, although the error due to voltage deflection is considerably reduced, the deflections due to current alone are still appreciable. It is difficult to match the tubes under these operating conditions, and even when matched, the useful voltage range is exceedingly small.

This paper describes two types of electronic wattmeters for the measurement of ac power. The first type is mostly designed for use in circuits in which there are almost sinusoidal current and voltage relationships. Its basic multiplying circuit is of novel design and simple construction using a single vacuum tube. The second type is essentially the first, but with an additional circuit to make it generally applicable to voltages and currents having complex wave forms. The primary relationship leading to the main design equations of the incorporated multiplying circuit is the exponential law relating plate current to plate voltage in a diode when operated in the so-called "retarding-field" region. By extending this mode of operation to multigrill tubes, it was found that the tube yields further current-voltage relationships that are of importance in obtaining a predictable nonlinear performance when more than one signal is applied. These relationships are useful directly for a mathematical application to a specific type of a multiplying circuit.

II. "RETARDING-FIELD" CHARACTERISTIC OF DIODES

It is known that in a diode the plate current \( I_b \) is related to the temperature-limited current \( I_e \) by the expression:

\[ I_b = I_e e^{bE_m}, \]

where \( E_m \) is the potential at the virtual cathode, \( b = e/kT = 11600/T \) volts\(^{-1} \), where \( e \) is the charge of the electron \((1.602 \times 10^{-19} \text{ coulomb})\), \( k \) is Boltzman's constant \((1.380 \times 10^{-23} \text{ watt sec}.)\), and \( T \) is the absolute temperature in degrees Kelvin. For a given diode at a fixed cathode temperature, the magnitude of \( E_m \) is a function of the plate voltage \( E_p \). As \( E_p \) becomes sufficiently negative, \( E_m \) approaches \( E_b \) until at a certain
plate voltage $E_b'$, the potential minimum is equal to $E_s'$ and its position is at the plate. For all values of $E_b$ more negative than $E_b'$ a region known as "retarding-field region," $E_b = E_w$, and the $I_b - E_b$ characteristic is exponential and obeys the law:

$$I_b = I_{b_0} e^{b E_b}.$$  (2)

The exponential portion of the diode characteristic occurs at plate voltages more negative than those for which the three-halves power law holds, and only at plate currents determined by the Maxwellian theory of initial electron velocity distribution. The theoretical value of $b$ corresponding to a cathode temperature of 1000° K is 11.6 volts$^{-1}$. This exponential relationship is classical and has found practical applications; for instance, in analyzing the behavior of the so-called peak type of vacuum-tube voltmeter with small applied voltages. It is important both theoretically and practically, because it is essentially independent of variations in tube construction and processing, and therefore gives accurately reproducible results that do not demand careful selection of tubes. Fig. 1 shows this exponential characteristic for the Type 954 connected as a diode with all the grids strapped to the plate. The slope $b$ of the ln $-E_b$ curve is 9.64 volts$^{-1}$.

### III. "Retarding-Field" Characteristics of Multigrid Tubes

The development of the exponential mode of operation of multigrid tubes will be presented here briefly:


but some of the clarifying details are given in Appendix 1. When the screen grid voltage of a tetrode or a pentode is reduced to a value near cathode potential and the grid voltage is sufficiently negative so that a virtual cathode is formed at the grid, and both grid and plate currents will vary exponentially against grid voltage. These low currents are determined by initial velocity distribution and occur at grid voltages more negative than those for which the three-halves power law holds.

Fig. 2 shows these grid and plate currents for the Type 954 plotted against grid voltage on a semilog chart. The curves are taken for plate voltages between 20 and 400 volts with the screen grid voltage held constant at +3 volts. It will be seen that the curves are accurately exponential up to about 10 microamperes and that the grid current $I_g$ is substantially independent of plate voltage. The slope $b_g$ of the ln $I_g - E_g$ curves is constant, being practically 9.13 volts$^{-1}$. The slope $b_p$ of the ln $I_p - E_p$ curve is 10.86 volts$^{-1}$. The curves show that, over the full plate voltage range, the grid and plate currents, respectively, obey the relationships:

$$I_g = I_{g0} e^{b_g E_g},$$  (3)

and

$$I_p = I_{p0} e^{b_p E_p},$$  (4)

where $I_r$ and $I_b$ are, respectively, the extrapolated values of grid and plate currents at $E_s = 0$. Furthermore, the plate current curves of Fig. 2 show that since $b_p$ is fixed, the value of $I_{b0}$ is determined by the plate voltage only.
Fig. 3 shows the same experimental data of Fig. 2, plate current versus plate voltage plotted on a linear chart for various constant values of grid voltage. These curves show that, over the full plate voltage range from 20 to 500 volts, the plate current varies accurately linearly with plate voltage. Also, the lines determine a common intersection on the plate voltage axis at \( E_b = -660 \) volts. Under these conditions, the value of \( I_b \) varies linearly with plate voltage, and (4) for plate current can therefore be rewritten as:

\[
I_b = (mE_b + n)ebEs,
\]

(5) where \( E_b \) and \( E_s \) are, respectively, the plate and grid voltages, \( m \) is the slope of the extrapolated plate current-plate voltage characteristic for \( E_s = 0 \), and \( n \) is the extrapolated plate current for \( E_b = E_s = 0 \).

Referring to Fig. 2 and 3 it is evident that a multi-grid tube operated under initial velocity conditions yields a plate current accurately proportional to the product of a linear function of plate voltage and an exponential function of grid voltage over wide ranges. The grid current is an exponential function of grid voltage, and substantially independent of plate voltage.

IV. MULTIPLYING CIRCUITS

Consider the simplified circuit of Fig. 4 in which a pentode \( V-1 \) is shown with its electrodes suitably polarized to operate in the exponential regime with (3) and (5) valid. Two ac input voltages \( v_p \) and \( v_g \) are applied in the plate and grid circuits, respectively. The source impedances of \( v_p \) and \( v_g \) are not shown, since these can be neglected compared to the corresponding high tube impedances in the operating region of Fig. 3. For the same reason, grid rectification is neglected, and only (5) will be considered. With no alternating voltages applied, let the dc plate and grid voltages be so adjusted that the direct current through the plate is given by:

\[
I_b = (mE_b + n)ebEs,
\]

(6) where \( E_b \) and \( E_s \) are the dc plate and grid voltages, respectively.

Fig. 4—Simplified multiplying circuit incorporating a pentode under initial velocity conditions.

When \( v_p \) and \( v_g \) are applied, the plate voltage becomes \( (E_p + v_p) \) while the grid voltage becomes \( (E_s + v_g) \), and the plate current becomes \( (I_b + i_b) \). Therefore:

\[
i_b = I_b(e^{bE_p} - 1) + \frac{m_0v_pe^{bE_s}}{E_s}.
\]

(7) where \( m_0 = m_e^{bE_s} \) and equals the plate conductance at the quiescent operating point. Over an infinitesimal grid voltage excursion, (7) reduces to:

\[
i_b = I_b(v_p) + m_0(v_p) + m_0b(v_p)v_p.
\]

(8)

Equation (8) indicates that with infinitesimal grid voltage excursion, the change in plate current, consequent upon application of alternating plate and grid voltages, comprises a term proportional to the product of the two voltages in addition to terms proportional to each voltage. The average change in plate current equals \( (i_b)_{av} \) proportional to \( (v_pv_g)_{av} \) provided that \( v_p \) and \( v_g \) have no dc components.

In order to obtain a useful amount of output by increasing the grid voltage excursion, plate rectification occurs due to the nonlinearity inherent in the exponential curvature. The multiplying circuit of Fig. 4 would be ideal, had the plate current been proportional to the product of linear functions of plate and grid voltages. However, two methods will be described here for compensating accurately the effect of plate rectification. One method compensates for the average value of plate rectification, whereas the other compensates for the instantaneous value.

A. Compensation by Average Grid Rectification

Since plate current is exponential against grid voltage, plate rectification occurs and the average change in plate current consequent upon application of alternating plate and grid voltages contains a component determined by grid voltage only. This component is independent of plate voltage and can therefore be compensated by changing the average potential to the grid. Since, as previously indicated, the grid current is substantially independent of plate voltage, it is therefore reasonable to see if the tube can bias itself automatically
to the proper operating point to compensate for average plate rectification by using the grid current to produce average grid rectification.

Consider the circuit shown in Fig. 5 (a) which is similar to that shown in Fig. 4, except that a grid leak resistor $R_p$ by-passed by a capacitor $C_p$ is connected in series with the grid. Also, a plate load resistance $R_p$ by-passed by a capacitor $C_p$ is connected in series with the plate supply $E_b$. Let the source impedances of the alternating voltages $v_p$ and $v_g$ be negligibly small compared to the corresponding tube impedances in the operating region shown in Fig. 3. Let, also, the values of $C_p$ and $C_p$ be such that all the alternating voltages $v_p$ and $v_g$ are effectively applied respectively across the grid-cathode and plate-cathode spaces of the tube.

With no alternating voltages applied to the system, let the dc potential at the plate be $E_b$, and the direct currents flowing in the plate and grid be $I_b$ and $I_g$ respectively. The grid current $I_g$ will flow through $R_p$, thus biasing the tube negatively; the bias depending upon $R_p$ and its magnitude is $E_g = I_g R_p$. Also, the plate current $I_b$ will flow through $R_p$, thus dropping the battery voltage to $E_b$ at the plate.

When $v_p$ is applied, grid rectification occurs and the current through $R_p$ increases. The total grid voltage becomes $(-E_b + v_p - \Delta V_g)$ where $\Delta V_g$ is the average rectified grid voltage depending upon the magnitude of $R_p$ and the applied alternating grid voltage (see Appendix II).

When the alternating voltages $v_p$ and $v_g$ are applied simultaneously, the plate current will be $(i_b + i_b)$; the average value changing from $I_b$ to $(i_b + (i_b)_{av})$. The total plate voltage will be $(E_b + v_p - (i_b)_{av} R_p)$. Therefore, substituting these values in (5), the plate current consequent upon the application of alternating voltages to the system is given by:

$$I_b + I_b = [mE_b + mv_p - m(i_b)_{av} R_p + n] e^{-b_p R_p + b_p R_p - b_p \Delta V_g}.$$
value of the product of two alternating voltages. Under these conditions, when the peak ac grid voltage is limited to less than 1/2 \( b_p \), the accuracy of measurement is better than \( \pm 1.5 \) per cent of full output. Meanwhile, for \( b_p V_p = 0.5 \), the full output is 12.5 per cent of the peak ac plate voltage at unity power-factor and with \( m_0 R_p = 1 \); being only doubled for \( m_0 R_p > 1 \).

As to grid bucking, it is possible (see Appendix II) to adjust the grid rectification efficiency so that the tube biases itself automatically to such an operating point that almost perfect grid bucking is achieved. The condition for optimum grid bucking requires that the magnitude of \( R_g \) be given by:

\[
R_g \equiv \frac{1}{\delta b_p I_c} \quad \text{or:} \quad \delta b_p I_c R_g = 1
\]  

(14)

where

\[
\delta = \frac{I_c}{I_p}
\]

(see Appendix II).

However, by examining (9), it will be apparent that when \( V_p = 0 \), the condition for perfect grid bucking demands that \( (i_b)_{av} \) should be zero independent of the magnitude of \( v_g \). Therefore, practically, the multiplying circuit of Fig. 5 can be adjusted for optimum grid bucking by applying appropriate ac voltages to the grid with the plate input circuit short circuited, and adjusting \( R_g \) for no change in the dc component of plate current. The same result can be achieved by fixing the value of \( R_g \) and introducing a small adjustable grid bias to control the magnitude of \( I_c \) so as to fulfill the requisite grid rectification efficiency for optimum grid bucking. These practical methods of adjustment require only a fair knowledge of the parameters \( \delta, b_p \), and \( I_c \) in (14) for the particular type of tube used. It will be necessary to readjust the screen grid voltage in order to restore the plate current to its initial value before and during the adjustment.

The circuit of Fig. 5(b) is similar to that in Fig. 5(a), except that the alternating voltages \( v_p \) and \( v_s \) are fed in parallel with the plate and grid, respectively. In this arrangement, since capacitive feed is used, any dc components in the applied voltages will not reach the tube. A resistance-capacitance filter \( R_o - C_o \) is connected across the plate and cathode; the dc potential at the plate appearing across \( C_o \).

Fig. 6—A plot of the quotient \( \frac{j I_c (j b_p V_p)}{f(I_c j b_p V_p)} \) and its percent deviation from linearity versus \( b_p V_p \).

Fig. 7 shows results of measurements of the dc output voltage due to imperfect grid bucking versus rms grid voltage for various values of \( R_g \) in the circuits of Fig. 5.
imperfect grid bucking is less than 4 millivolts over an rms grid voltage range up to 40 millivolts. For smaller values of $R_e$ the output is positive, whereas for larger values the output displays a negative loop. The negative loop curves show that the ac grid voltage range can be increased to about 70 millivolts without having an appreciable output voltage due to imperfect grid bucking, as shown by the curve for $R_e = 0.434$ megohm.

Fig. 8—Observed values of dc output voltage due to optimum grid bucking versus rms grid voltage for various values of the heater voltage in the circuit of Fig. 5.

Fig. 9 shows results of similar measurements with the circuit initially adjusted for optimum grid bucking at a heater voltage about 5.7 volts. The curves show that the circuit adjustment for optimum grid bucking is not appreciably affected by a reasonable change in heater voltage.

The pentode Type 954 is particularly suited for a large plate voltage swing without showing a serious error, due to imperfect linearity of plate current-plate voltage characteristics. For this Type, 4 microamperes is a suitable value for $I_p$ and 1.0 microampere for $I_r$. The plate conductance at 4 microamperes is of the order of 0.005 microampere per volt, corresponding to an internal resistance of 200 megohms. The requisite value of $R_e$ for optimum grid bucking is of the order of 0.5 megohm, and a suitable value for $R_e$ is 100 megohms. Since the dc voltage drop across $R_e$ is 400 volts; the resistor should therefore be selected for minimum voltage coefficient. This is necessary in order to minimize resistor rectification due to dc polarization. Also, for the same reason, capacitors $C_r$ and $C_a$ should preferably be of the polystyrene dielectric type. With these circuit components, and with $R_e = 250$ volts, the full output at unity power factor is about 20 volts dc for 200 volts peak ac plate voltage, and 70 millivolts peak ac grid voltage.

Circuit Performance as a Wattmeter: The circuit of Fig. 5(b) is particularly suited for a simple electronic wattmeter. Fig. 9 shows a diagram of the basic circuit of the type in which 1'-1 is the basic measuring tube, and 1'-2 is a twin triode connected as a conventional degenerative dc voltmeter. The indicator meter $m_2$ is connected across a reversing switch $S$. The voltage component of power to be measured is applied to the plate of 1'-1, while all the current component practically passes through the series resistor $r$ and develops a voltage drop which is proportional to current and is applied to the grid of 1'-1.

Assuming that $v = V \cos (\omega t + \phi)$ and $i = I \cos \omega t$, the change in the dc voltage across $R_e$ consequent upon the application of $v$ and $i$ is obtained by substitution in (13). hence:
\[ \Delta E_b = \frac{m_b R_p}{1 + m_b R_p} \frac{1}{2} b_p r (VI \cos \phi) \]

which is proportional to the mean ac power dissipated in the load.

The circuit was tested for wattmetric indication by comparing its reading with that indicated by a dynamometer wattmeter at 60 cycles. Fig. 10 shows relative results obtained at unity power factor and Fig. 11 at variable power factors. The values of power factor indicated in Fig. 11 are calculated from the reading of the dynamometer wattmeter, a voltmeter, and an ammeter. The curves show that both wattmeters agree closely. Fig. 12 shows that the deflection of the electronic wattmeter with a constant ac plate voltage is very closely proportional to the magnitude of the ac grid voltage, as has been theoretically predicted and shown in Fig. 6.

Preliminary trials for determining the frequency characteristic of the circuit shown in Fig. 9 indicated a reasonably flat frequency response up to about 10 mega-
cycles. Some errors were observed above this frequency indicating the presence of feedback. The circuit was then tested with the grid input circuit short circuited and a variable frequency voltage of about 150 volts rms applied to the plate. In this test, a dc output of one per cent of full scale was observed at about 5 Mc increasing approximately as the square of frequency. Since, in this type of operation, the grid-plate capacitance is insufficient to cause appreciable error at this frequency, it was found that the observed errors are due to feedback through the cathode and screen-grid lead inductances.

The effects of these two types of feedback, upon the performance of the tube under such operating conditions, are in opposite directions. One type of feedback can therefore be neutralized by means of adjusting the other. However, in the Type 954, an examination of the electrode structure showed that there are some auxiliary metallic parts, such as a top cap, a bottom ring, and a getter support, all connected to the bottom ring together with the cathode, rather than with the suppressor, as is usually the modern policy of tube construction. The resulting direct capacitance between plate and cathode is of the order of 1.5 \( \mu \)f. In the megacycle region, and with 150 volts rms on the plate, an appreciable radio-frequency current flows through the cathode lead inductance, thus developing an in-phase voltage in the grid circuit of the order of few millivolts. In the Type 954, the amount of feedback through the screen-grid circuit is insufficient to compensate for this cathode feedback. However, with the aid of a small adjustable neutralizing capacitor of the order of 1 \( \mu \)f connected between the plate and the screen grid it was possible to increase the feedback through the latter to an extent sufficient to compensate for the excessive cathode feedback. This arrangement extended the frequency range for the Type 954 to 20 Mc. There is a good possibility of exceeding this range by using tubes which do not have such an appreciable plate to cathode capacitance as in Type 954.

The multiplying circuit of this wattmeter is highly sensitive to supply voltage changes the power supply should be well regulated. A good degree of stability is achieved by the use of a bucking tube of the same type as the wattmeter tube and connected to the other arm of the bridged degenerative voltmeter circuit of Fig. 9. Nevertheless regulation, within 0.2 per cent is necessary at the low alternating plate voltage ranges. The circuit stability may also be improved by using tubes having tungsten or tantalum filaments since these have a value of \( \beta = e/kT \) of the order of 3 to 4 volts\(^{-1}\). The use of such tubes require a maximum ac grid voltage of the order of 100 millivolts rms for \( b_p V_g = 0.5 \). This has the further advantage of reducing effects of stray pickup and feedback at high frequencies.

In some fields of application of this wattmeter, as for instance in the measurement of power at low power-factors, the use of a two-tube push-pull arrangement may be necessary, in order to reduce the errors caused by
imperfect grid bucking and imperfect linearity of plate current-plate voltage characteristics. Fortunately, such errors in this type of operation can be determined and controlled.

This wattmeter has the advantages of a wide frequency range and simplicity, in addition to the fact that it absorbs only a minute fraction of the measured power. It also enables the measurement of very small amounts of power.

B. Compensation by Instantaneous Grid Rectification

In this method of compensation, the instantaneous values of rectified plate current due to the exponential grid curvature are compensated by varying the instantaneous grid voltage on a proper logarithmic curvature. The system, in this case, obtains the product of the applied alternating voltages as if the plate current were proportional to the product of linear functions of plate and grid voltages.

Consider the circuit shown in Fig. 13 which is similar to that in Fig. 5 except that the by-pass capacitor $C_e$ is removed and the alternating voltage $v_p$ is applied to the grid through the grid leak resistor $R_p$. An external bias $E_{re}$ is also shown.

With no alternating voltages applied, the direct current through the grid will bias the tube negatively; the bias will be $E_x = I_x R_e - E_{re}$, where $I_x = I_x e^{-b_p E_x}$. Therefore: $E_x = E + E_{re} = I_x R_e e^{-b_p E_x}$. When the external bias $E_x$ is much larger than $E_p$, the value of $R_p$ will be practically $(E_{re}/I_x)$ and the relation between $E_x$ and $E_p$ will be given by:

$$E_{re} = I_x R_p e^{-b_p E_x}, \quad \text{i.e.,} \quad E_x = \frac{1}{b_p} \ln \left( \frac{E_{re}}{E_x R_p} \right)$$

indicating that the grid maintains a bias proportional to the logarithm of the applied external bias.

When the alternating voltage $v_p$ is applied, grid rectification occurs, and the instantaneous voltage developed at the grid will be $(-E_x + e_p)$, where $e_p$ is given by

$$-E_x + e_p = \frac{1}{b_p} \ln \left( \frac{E_x + v_p}{I_x R_p} \right)$$

i.e.,

$$e_p = \frac{1}{b_p} \ln \left( \frac{v_p}{E_{re}} \right) \quad (15)$$

Let the direct current through the plate be given by (6). The rectified plate current consequent upon application of alternating voltage $v_p$ will be given by:

$$I_b + \Delta I_b = (m b_p + n) e^{-b_p E_x} \ln \left( \frac{v_p}{E_{re}} \right)$$

Let in the ideal case $b_p = b_q$, then $\Delta I_b = I_b(v_p/b_q)$ indicating that the output plate current is an exact duplicate of the input voltage $v_p$ and that plate rectification is perfectly compensated. The system, in this case, behaves as if the plate current were accurately proportional to the product of linear functions of plate and grid voltages.

When the alternating voltages $v_p$ and $v_q$ are applied to the system, the plate voltage will be $(E_x + v_p - (b_p)_{av} R_p)$ and the grid voltage will be $(-E_x + e_p)$. The plate current consequent upon application of alternating voltages will be:

$$I_b + i_b = [m b_p + m b_q - m(b_p)_{av} R_p + n] e^{-b_p E_x} b_p b_q \ln \left( \frac{v_p}{E_{re}} \right)$$

and if $b_p = b_q$, then

$$i_b = m b_q + \frac{m}{F_{re}} \left( I_b - m(b_p)_{av} R_p \right)$$

$$+ \frac{m}{F_{re}} v_p v_q - m(b_p)_{av} R_p \quad (16)$$

Equation (16) holds for all wave forms of the voltage $v_p$ and $v_q$. It will be seen from (15) and (16) that, when the instantaneous voltage across the grid-cathode space of the tube is a proper logarithmic function of the applied voltage $v_p$, the compensation of the exponential curvature is accurate. In this case, the circuit obtains the product of the two voltages $v_p$ and $v_q$ without producing errors due to increasing the grid voltage excursion.

Taking average values of both sides of (16) and assuming that $v_p$ and $v_q$ have no dc components, the change in the dc voltage across $R_p$ is given by:

$$\Delta E_b = (b_p)_{av} R_p = \frac{m b_p R_p}{1 + m b_p R_p} \left( \frac{1}{F_{re}} \right) (v_p v_q)_{av} \quad (17)$$

Although the circuit is extremely simple, yet it has the disadvantage that the accuracy of obtaining the product $(v_p v_q)$ depends upon the discrepancy in the ratio $b_p/b_q$ from unity, in addition to a working frequency range limited by the frequency characteristics of $R_p$ and its associated capacities. The circuit can be arranged to give satisfactory performance up to high audio frequencies.

However, the system is greatly improved by the addition of a radio-frequency logarithmic circuit. This additional circuit depends upon the fact that, with proper
adjustments, the plate current of the pentode tube is accurately logarithmic against plate voltage; the relationship being:

\[ I_b = A + K \ln E_b, \]

where \( A \) is a constant and \( K \) is the slope of the \( I_b - \ln E_b \) curve. Fig. 14 shows the circuit diagram and operating characteristics of the Type 6AU6 for a logarithmic relation over the plate voltage range from 0.5 to 2.0 volts. In this circuit the value of \( K \) is mainly controlled by the screen grid voltage and the self-bias resistor. If the direct current voltage on the plate is adjusted to the middle of the logarithmic range and a limited alternating voltage applied to the plate through a low impedance, the consequent change in plate current will be:

\[ i_b = K \ln \left( 1 + \frac{v_p}{E_b} \right). \]

When the alternating voltages \( v_p \) and \( v_a \) are applied simultaneously, the plate voltage of V-1 will be \( [E_b + v_p - (i_b)_{av} R_p] \), and the plate current will be:

\[ I_b + i_b = \left[ mE_b + mv_p \right] - m(i_b)_{av} R_p + n_{v_b} \left( -E_c + KR \ln \left( 1 + \frac{v_p}{E_b} \right) \right) \]

\[ = \left[ I_b + m_0v_p - m_0(i_b)_{av} R_p \right] + m_0 v_a \left( i_b \right)_{av} R_p, \]

which is similar to (16). The change in the dc plate voltage consequent upon application of \( v_p \) and \( v_a \) is given by:

\[ \Delta E_b = (i_b)_{av} R_p = \frac{m_0 R_p}{1 + m_0 R_p E_b} \left( v_p v_a \right)_{av}. \]

The requisite value of \( R \) can be calculated from \( 1/b_p K \). The practical method of adjusting the system is carried out experimentally by applying the alternating voltage \( v_p \) with the plate input circuit of V-1 short circuited and adjusting \( R \) so that the change in the dc output voltage is zero. The same result can be achieved by fixing the value of \( R \) and adjusting the value of \( K \) by changing the screen grid voltage of V-3. Should \( b_p K R = (1 \pm \Delta) \) where \( \Delta \) is a small fraction, the dc output voltage due to imperfect compensation by the logarithmic circuit is approximately \( \pm (\Delta/4)(V_p/E_b) \). \( I_b R_p \) where \( V_p \) is the peak value of \( v_p \). This amounts to about 30 millivolts for a value of \( \Delta = 0.001 \), \( (V_p/E_b) = 0.6 \) and \( I_b R_p = 400 \) volts. The full output is 30 volts at unity power factor and 200 volts peak on the plate for \( m_0 R_p = 1 \). However, provision can be easily made in the circuit for a fine adjustment of the condition \( b_p K R = 1 \).
In practice, values of $K$ of the order of 1.0 milliamperes are obtained from ordinary receiving tubes. The required value of $R$ is about 100 ohms assuming $b_p = 10$ volts$^{-1}$. With these values the frequency range of the circuit is extended to the neighborhood of 30 Mc. However, it is necessary to compensate for the unavoidable shunting capacities across $R$. These usually amount to about 15 microfarads and a small inductive component of $R$ is desirable.

Appendix I

"Retarding-Field" Characteristics of Multigrid Tubes

The extension of the exponential mode of operation of diodes to multigrid tubes was accomplished by investigating—at first—the retarding-field characteristics of triodes.

In a triode, when the plate voltage is reasonably low, and the grid voltage is sufficiently negative so that a virtual cathode is formed at the grid, the grid current-grid voltage characteristic becomes exponential with a slope $b_p$ closely equal to $b$. Since the plate current constitutes electrons which pass through the grid mesh, the plate current-grid voltage characteristic is also exponential, but with a slope $b_p$ considerably less than $b$. Fig. 16 shows these exponential curves of the Type 954 connected as a triode with grids 2 and 3 strapped to the plate. The curves are taken for plate voltages between 5 and 40 volts. The slope $b_p$ of the $\ln I_p - E_p$ curves is, in all cases, fairly close to that observed for the same tube connected as a diode, i.e., 9.64 volts$^{-1}$; but the slope $b_p$ of the $I_p - E_p$ curves is only 6.81 at $E_p = 5$ volts, decreasing to 3.62 at $E_p = 40$ volts. This considerable decrease in the observed value of $b_p$ is due to the fact that the field distribution between grid and cathode is appreciably affected by the plate voltage.

It was inferred that the dependence of $b_p$ upon plate voltage can be avoided by shielding the plate from the grid-cathode region. This was proved by taking measurements of $b_p$ for tetrodes and pentodes. Fig. 17 shows results of the measured values of $b_p$ against $E_p$ for the Type 954 connected progressively as a triode, tetrode, and pentode. In these measurements the screen grid potential was held constant at a value just convenient to permit a few microamperes to flow in the plate circuit. From these curves it is evident that as the plate is progressively more and more shielded from the cathode, the value of $b_p$ increases and becomes substantially independent of plate voltage.

With a fixed value of $b_p$, the general shape of the retarding-field characteristics of multigrid tubes are as those shown in Figs. 2 and 3 for the Type 954. These characteristics differ from those of triodes in that the slopes $b_p$ and $b_p$, and the grid current are substantially independent of plate voltage. Furthermore, the plate current varies linearly with plate voltage. In general, different types of multigrid tubes have different values of the constants $m$, $n$, and $I_m$ in (3) and (5). The values of $b_p$ and $b_p$ mainly depend upon the operating temperature of the cathode. For tungsten filaments, the value of $b$ is of the order of 3 to 4 volts$^{-1}$. In all types, cathode stabilization is important in order to obtain consistent performance, and tube ageing for at least 100 hours may therefore be necessary.

Perhaps the most interesting feature of these characteristics is the wide range over which the plate current increases with plate voltage.

varies linearly with plate voltage. If, however, we consider the effect of \( E_s \) upon \( E_w \) when all other electrodes are at a constant potential, it is conceivable that the plate current varies in accordance with the function 
\[ e^{bE_s/\mu} \]
where \( \mu \) is the amplification factor. When \( \mu \) is high, the value of the exponent \( bE_s/\mu \) is a small fraction of unity and the function becomes practically linear with \( E_s \). Furthermore, as the plate voltage is increased, the position of the potential minimum moves towards the cathode causing \( \mu \) to increase. This increase in \( \mu \) with \( E_s \) tends to make the mode of variation of \( I_b \) with \( E_b \) more closely linear over a wider range of \( E_s \). The curves of Fig. 3 show practically good linearity over the full plate voltage range from 20 up to 500 volts. In tubes having a higher screening factor, this linear range extends to 1,000 volts, but the internal resistance becomes excessively high.

Fig. 18 shows the values of \( \mu \) and \( r_p \) for the Type 954 as calculated from the data in Fig. 2. The curves show that \( \mu \) increases accurately linearly with plate voltage, but is substantially independent of grid voltage. On the other hand, \( r_p \) is independent of plate voltage, but varies exponentially with the grid voltage. The transconductance can be obtained from Fig. 2 by multiplying the value of \( I_b \) by the slope \( b_p \). All these relations are apparent from an examination of (5).

![Graph](image)

**Fig. 18**—Values of amplification factor and plate resistance versus plate voltage for the Type 954 under initial velocity conditions.

**Appendix II**

**Comparison of Perfect and Practical Grid Bucking**

In the description of the operation of the multiplying circuit of Fig. 5, it was assumed that the tube can bias itself automatically to the proper operating point to compensate for average plate rectification by using the grid current to produce average grid rectification. Under this assumption, the average rectified grid bias should be given by (10) for perfect grid bucking. It will be proved here that, at least for the case where \( v_s = V_s \cos \omega t \), perfect grid bucking can almost be achieved by a particular adjustment of the grid leak resistor. In this case the value of \( \Delta E_e \) from (10) should be given by:

\[ \Delta E_e = \frac{1}{b_p} \ln J_0(jb_pV_s). \]  

(21)

Considering the grid circuit, and with no alternating voltage applied, let the direct current through the grid be given by:

\[ I_e = I_c \cos \theta. \]

When \( v_s \) is applied, the grid voltage becomes \(-E_s + v_s - \Delta E_e\), and the grid current will be:

\[ I_e + i_e = I_c \cos(-\theta + \Delta E_e), \]

giving

\[ i_e = I_c [e^{v_s/\mu} \cos \Delta E_e - 1]. \]

Solving for \( \Delta E_e = (i_e/v_s)R_s \) by taking average values of both sides and multiplying by \( R_s \) we get:

\[ \Delta E_e = I_cR_s [e^{v_s/\mu} \cos \Delta E_e - 1], \]

and when \( v_s = V_s \cos \omega t \), then:

\[ \Delta E_e = I_cR_s [J_0(jb_pV_s) \cos \Delta E_e - 1]. \]  

Equation (22) relates the average rectified grid voltage to the applied ac grid voltage. This expression is similar to that obtained by Aiken\(^8\) on the theory of the diode voltmeter, and has no explicit solution.

If, however, it is possible to adjust the magnitude of the grid leak resistance \( R_s \) so that the average rectified grid voltage is equal to that required for perfect grid bucking, then \( R_s \) should be determined by substituting from (21) into (22) and solving for \( R_s \). Hence:

\[ R_s = \frac{1}{b_p} \frac{(J_0(jb_pV_s))^2 \ln [J_0(jb_pV_s)]}{J_0(jb_pV_s) - [J_0(jb_pV_s)]^2}. \]  

Equation (23) shows that for the case \( \beta = 0 \), i.e., \( b_p = b_s \), the denominator vanishes, and \( R \) should be infinite. But, for the case \( \beta > 0 \), i.e., \( b_s < b_p \) which is usually true, the value of \( R_s \) becomes finite. In this latter case, it is clear from (23) that the magnitude of \( R_s \) for perfect grid bucking should depend upon the magnitude of \( V_s \). This indicates that it is not possible with a simple linear grid-leak resistor to achieve perfect grid bucking at all values of \( V_s \).

Let therefore the adjustment of the grid leak resistance be such that (21) and (22) are satisfied only at small values of \( V_s \). We have then:

right-hand side of (21)

\[ \approx \frac{1}{b_p} \ln (1 + \frac{1}{2}b_p^2V_s^2) \approx \frac{1}{2}b_p^2V_s^2, \]

and right-hand side of (22)

\[ \frac{1}{b_p} \ln (1 + \frac{1}{2}b_p^2V_s^2) \approx \frac{1}{2}b_pV_s^2. \]

\[
Z = \frac{1}{1 + b_{b}I_{e}R_{e}} \approx \frac{1}{1 + b_{b}I_{e}R_{e} + 1}
\]

By equating both results, we get:

\[
R_{e} = \frac{1}{bb_{b}I_{e}} \quad \text{or} \quad bb_{b}I_{e}R_{e} = 1. \quad (14)
\]

With the adjustment of the grid leak resistor as given by (14), it is possible to calculate the dc output voltage of the circuit of Fig. 5 due to imperfect grid bucking. Rearranging (22) we get:

\[
\Delta E_{c} = \frac{1}{b_{o}} \ln \left(\frac{v_{o}}{v_{o}}\right)_{av.} - \frac{1}{b_{o}} \ln \left(1 + \frac{\Delta E_{c}}{I_{e}R_{o}}\right)
\]

For small values of \( v_{o} \) the rectification efficiency is low, and \( \Delta E_{c}/I_{e}R_{o} \) is small compared to unity, thus:

\[
\ln \left(1 + \frac{\Delta E_{c}}{I_{e}R_{o}}\right) \approx \frac{\Delta E_{c}}{I_{e}R_{o}}.
\]

Therefore,

\[
\Delta E_{c} \approx I_{e}R_{o} \frac{\ln \left(\frac{v_{o}}{v_{o}}\right)_{av.}}{1 + b_{b}I_{e}R_{o}}
\]

for a first approximation. Using successive approximation, and substituting from (14) into (24), we get:

\[
e^{b_{b}\Delta E_{c}} \approx \left[\left(\frac{v_{o}}{v_{o}}\right)_{av.}\right]^{1/X}
\]

where

\[
X = 1 + 2\delta + \delta^{2} - \frac{1}{2} \delta^{2} \ln \left(\frac{v_{o}}{v_{o}}\right)_{av.}
\]

Let

\[
Z = \left(\frac{v_{o}}{v_{o}}\right)_{av.} e^{-b_{b}\Delta E_{c}} = \left(\frac{v_{o}}{v_{o}}\right)_{av.} \left[\left(\frac{v_{o}}{v_{o}}\right)_{av.}\right]^{-1/X},
\]

then (9) can generally be rewritten as:

\[
(i_{b})_{av.}R_{p} = I_{b}R_{p} \frac{Z - 1}{1 + m_{0}R_{p}Z} + m_{o}R_{p}Z \frac{(v_{p}b_{b}v_{o})_{av.}}{1 + m_{0}R_{p}Z} \frac{(v_{p}b_{b}v_{o})_{av.}}{1 + m_{0}R_{p}Z}.
\]

from which the absolute value of the dc output voltage due to imperfect grid bucking is given by:

\[
I_{b}R_{p} \frac{Z - 1}{1 + m_{0}R_{p}Z}.
\]

and the per cent error in the dc output due to imperfect grid bucking is given by:

\[
\frac{I_{b}R_{p} - \frac{Z - 1}{1 + m_{0}R_{p}Z}}{m_{0}Z} \times 100.
\]

Equations (25) and (26) are general and apply for any wave form of the ac grid voltage. For the case \( v_{o} = V_{o} \cos \omega t \), Fig. 19 shows a plot of the dc output voltage due to imperfect grid bucking as calculated from (25) for various values of \( b_{b}I_{e}R_{o} \) close to unity. These curves show that for the condition \( b_{b}I_{e}R_{o} = 1 \), the output is positive, increasing as \( b_{b}V_{o} \) is increased; but for values slightly greater than unity, the output shows a negative loop. It is interesting to see that the curves in Fig. 19 agree in shape and order of magnitude with those obtained experimentally in Fig. 7.

Fig. 20 shows the per cent error in the dc output voltage due to imperfect grid bucking as calculated from (26) for unity power factor. These curves show that it is possible to adjust the grid leak resistor so that for rms grid voltages less than 40 millivols, the order of the error is less than one per cent at unity power factor and with only 15 volts rms on the plate.

It should therefore be concluded from the above that if a linear-grid-leak resistor is used, the adjustment for optimum grid bucking corresponds to a value of \( b_{b}I_{e}R_{o} \) slightly greater than unity. The requisite grid rectification efficiency for optimum grid bucking equals \( \frac{1}{b_{b}V_{o}} \) for values of \( b_{b}V_{o} \) less than 0.5, decreasing slightly for higher values.

**Appendix III**

*Accuracy of Measurement by the Multiplying Circuit of Fig. 5 under Complex Wave Forms*

It was proved in the text that \((i_{b})_{av.}\) of (9) is proportional to \((V_{p}V_{o} \cos \phi)\) in the simple case where \(v_{p} = V_{p} \cos (\omega t + \phi), v_{o} = V_{o} \cos \omega t\) and \(b_{b}V_{o} \leq 0.5\). If the alter-
nating voltages \( v_a \) and \( v_p \) have complex wave forms, the accuracy of measurement of \((v_p v_a)_{av}\) is subject to an error caused by the fact that only average plate rectification is compensated. This type of error does not exist in the multiplying circuit of Fig. 15.

Equation (9) gives the quantities generally involved in determining the value of \((v_p v_a)_{av}\) under any wave form of the alternating voltages \( v_a \) and \( v_p \). Also, (26) gives the per cent error in the indicated reading. Estimation of this error under standard wave forms—expressed in Fourier Series—indicates that second harmonic terms contribute almost to the full value of the discrepancy in the indicated reading from that of a perfect multiplying circuit. The error will therefore be considered due to second harmonic. Two cases will be illustrated here.

**Case a:** \( v_a = V_0 \cos (wt + \phi) \), and \( v_p = V_p (\cos wt + k_p \cos 2 wt) \) where \( k_p \) is the amplitude ratio of the second harmonic to the fundamental of the ac grid voltage. The following expressions are used in the analysis:

\[
(v_p v_a)_{av} = J(a(jbV_o) + [J(a(jbV_o) - 1] + \frac{3}{2}k_p(bV_o)^2[(1 + \frac{3}{2}k_p bV_o) - 1] + \frac{1}{2}k_p(bV_o)^2[1 + \frac{1}{2}k_p bV_o + \frac{1}{2}(bV_o)^2]].
\]

**Fig. 21** shows the calculated per cent error versus rms value of the fundamental grid voltage for various values of \( k_p \). These curves show that the error is of the order of 20 per cent for 100 per cent second harmonic at \( b_p V_o = 0.5 \); decreasing to about 3.5 per cent at \( b_p V_o = 0.125 \).

**Case b:** \( v_p = V_p \cos wt \), and \( v_p = V_p \{\cos (wt + \phi) + k_p \cos 2 (wt + \phi)\} \), where \( k_p \) is the amplitude ratio of second harmonic to fundamental of the ac plate voltage. The following expression is used:

\[
(v_p v_a)_{av} = V_p \cos \phi [ -jJ_1(jbV_o) + k_p j^2 J_2(jbV_o)].
\]

**Fig. 22** shows the calculated per cent error versus the rms value of the ac grid voltage for various values of \( k_p \). The curves show that the error is of the order of 14 per cent for 100 per cent second harmonic at \( b_p V_o = 0.5 \); decreasing to about 2 per cent at \( b_p V_o = 0.125 \).

It will therefore be concluded that though the measurement of ac power by the wattmeter circuit of Fig. 9 is subject to a reasonably serious error when the current and voltage are complex, advantage can be taken by reducing the voltage drop across the series resistor \( r \). The accuracy is greatly improved by using the appropriate series resistor which gives a value of \( b_p V_o \) about 0.125 (quarter full scale). Under this condition, the error in the measurement of the total ac power is of the order of +5 per cent when both current and voltage contain 100 per cent second harmonic; decreasing to less than 3 per cent for 50 per cent.

**Acknowledgments**

The author wishes to extend his appreciation to General Radio Company, Cambridge, Mass., whose cooperation made this development possible, with especial indebtedness to D. B. Sinclair, assistant chief engineer, whose constant interest has carried the investigation to a stage of development; and to R. A. Soderman, development engineer, for the able assistance in performing the experimental tests and many helpful calculations.

The author also wishes to express appreciation to Fouad University, Cairo, Egypt, for Fellowship facilities for circuit development.
Graphical Analysis of Tuned Coupled Circuits*

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Summary—A new basis for normalizing the transfer admittance of two coupled tuned circuits permits the representation of this admittance by a single universal curve which is a parabolic locus on the complex plane. Within the limitations of the assumptions of high Q and small frequency deviations, data can be obtained from this curve for different Q ratios, as well as the usual values of coupling and relative tuning. The method also simplifies the calculation of the input admittance of coupled circuits. Extension of the method to triple tuned circuits is possible, but the applicability of a single universal curve is lost.

Although the representation of the transfer admittance of two coupled tuned circuits by a parabolic locus was published by Smith,1 the idea has not received the attention it merited. Subsequently and independently, the same representation was worked out by Johnson,2 Hamilton,3 Spangenberg,4 Chang,5 and probably others.

This method of representing an admittance by a locus on the complex plane with the conductance as the real co-ordinate and the susceptance as the imaginary co-ordinate is quite general. A straight line parallel to the imaginary axis is the locus of the admittance of a single tuned circuit, two coupled tuned circuits have a parabolic locus, and a cubic curve represents a triple tuned circuit. The parabolic locus is the most useful, however. It is applicable only for high-Q circuits when the frequency deviation is small, but the error introduced when the Q is very low will be illustrated.

The parallel circuit of Fig. 1 has been chosen for illustration, because these circuit constants are most convenient for an admittance analysis. The results apply equally well to a circuit with a resistance in series with the inductance if the Q of the circuits is high. In low-Q circuits where the source of the losses becomes important, the circuit is usually loaded by a shunt resistance, and the circuit of Fig. 1 is generally applicable.

An analysis for the case when both circuits are tuned to the same resonant frequency \( \omega_0 \) gives the following relation for the transfer admittance \( Y_r \):

\[
Y_r = \frac{k}{k_x} (1 - k^2) \sqrt{G_1 G_2} \left\{ \frac{\omega^2}{\omega_0^2} - \frac{k^2}{(1 - k^2)} \right\} - Q (Q_1 + Q_2) \left\{ \frac{\omega^2}{\omega_0^2} - \frac{1}{1 - k^2} \right\}
\]

where

\[
k = \frac{M}{\sqrt{L_1 L_2}}
\]

\[
k_x = \frac{1}{\sqrt{Q Q_2}}
\]

and

\[
\omega_0 = \frac{1}{\sqrt{L_1 C_1}} = \frac{1}{\sqrt{L_2 C_2}}
\]

\[
Q_1 = \frac{\omega_0 C_1}{G_1} = \frac{1}{\omega_0 L_1 G_1}
\]

\[
Q_2 = \frac{\omega_0 C_2}{G_2} = \frac{1}{\omega_0 L_2 G_2}
\]

In order to convert (1) into the parabolic form, \( Q \) must be high. A \( Q \) of 100 is sufficiently high to make the parabolic locus accurate within 1 per cent for a frequency deviation corresponding to the 3 db points on the curve. The error will be larger for larger frequency deviations, but in general the parabolic locus is reasonably accurate within the values of frequency deviation of greatest interest if the \( Q \) of the circuits is 100 or more.

Under these conditions, \( \omega/\omega_0 \) and \( \omega/\omega \) can be replaced by unity. For small values of the coupling coefficient \( k \), the term \((1 - k^2)\) may also be considered unity and (1) becomes quadratic in terms of the quantity \((\omega^2/\omega_0^2) - 1\). It will be more convenient to express the relationship in terms of the frequency deviation ratio \( \Delta \omega/\omega_0 \) or \( \delta \). The approximate relation is

\[
\frac{\omega^2}{\omega_0^2} - 1 \approx 2 \delta
\]

When these approximations are substituted in (1), the result is

\[
Y_r \approx \frac{k}{k_x} \sqrt{G_1 G_2} \left\{ 1 + \frac{k^2}{k_x^2} - 4Q_1 Q_2 \delta^2 \right\} - 2(Q_1 + Q_2) \delta
\]

Fig. 1—Tuned coupled circuit considered in the analysis.

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* Decimal classification: R142. Original manuscript received by the Institute, January 13, 1949; revised manuscript received, May 13, 1949.
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‡ Princeton University, Princeton, N. J.
2 W. C. Johnson, Unpublished Lecture Notes, February, 1942.
In this form, the equation for the transfer admittance is a parabola, but a different parabola will be required for every value of the \( Q \) ratio \( (Q_1/Q_2) \). This difficulty is easily overcome by a change of variable. Consider the basic equation for a parabola with its vertex at the origin.

\[
A x^2 = By.
\]

If both terms are multiplied by \( A/B^2 \), the result is of the form

\[
X^2 = Y^2,
\]

indicating that a change of variable can eliminate the coefficients.

A variety of possibilities exist for the new variable in the expression for the transfer admittance. If we choose \( 4Q_1Q_2\delta/(Q_1+Q_2) \), this term reduces to the familiar universal frequency deviation parameter \( 2\delta \) when \( Q_1 \) is equal to \( Q_2 \). This term may also be rewritten

\[
\frac{4Q_1Q_2\delta}{Q_1+Q_2} = \frac{4\delta}{\frac{1}{Q_1} + \frac{1}{Q_2}} = \frac{2\delta}{\frac{1}{Q_1}+\frac{1}{Q_2}},
\]

where \( D \) represents the dissipation factor and is the reciprocal of the circuit \( Q \). An extension of this form of the variable to single- and triple-tuned circuits would give

\[
\alpha_1 = \frac{2\delta}{Q_1},
\]

\[
\alpha_2 = \frac{2\delta}{\frac{1}{Q_1} + \frac{1}{Q_2}},
\]

\[
\alpha_3 = \frac{2\delta}{\frac{1}{Q_1} + \frac{1}{Q_2} + \frac{1}{Q_3}}.
\]

In the limiting case with extremely small coupling and equal \( Q \) for all circuits, a unity value of this parameter would correspond to the 3-db, 6-db, and 9-db points for single-, double- and triple-tuned circuits, respectively. The corresponding phase shifts from the midfrequency would be 45, 90 and 135 degrees.

Another variable can be chosen which relates a unit value of the variable with the 3-db points when the coupling coefficient corresponds to transitional coupling, (maximal flatness). This parameter can also be extended to single- and triple-tuned circuits, and the relations are

\[
\gamma_1 = \frac{2\delta}{Q_1},
\]

\[
\gamma_2 = \frac{2\sqrt{2}\delta}{Q_1 + Q_2},
\]

\[
\gamma_3 = \frac{4\delta}{Q_1 + Q_2 + Q_3}.
\]

\(*\) Suggested in a Sperry Gyroscope Company report by W. W. Hansen.

In this case a unit value of \( \gamma \) corresponds to the 3-db point in each case, and the phase shift from the midfrequency value will be 45, 90, and 135 degrees, respectively.

It is probable that this frequency deviation parameter will be the most useful one in the analysis of coupled circuits, and it has been selected as preferable to the more familiar form. Equation (8) can be converted by introducing the frequency deviation parameter in (13b) and the result is

\[
Y_\tau \equiv \frac{k_e}{k} \sqrt{G_1G_2} \left( \frac{D_1 + D_2}{2k_\tau^2} \right)^2 \left\{ \frac{2k^2 + k_\tau^2}{(D_1 + D_2)^2} \right\} \left[ j \left[ \frac{1 + k^2}{k_\tau^2} \right] \frac{k^2}{1 + k_\tau^2} \right].
\]

The construction of the parabola will be simplified if it is defined by a focus and a directrix. While this could be done without modifying (14), the symbols will be simpler if the dissipation factors are replaced by coupling coefficients. This can be done by relating \( (D_1+D_2) \) to the critical coupling \( k_e \) and the transitional coupling coefficient \( k_\tau \), which is the largest value of \( k \) which allows a single minimum in the magnitude of the transfer admittance. The relations are:

\[
(D_1 + D_2)^2 = 2(k_e^2 + k_\tau^2)
\]

\[
k_\tau^2 = \frac{1}{2} (D_1^2 + D_2^2)
\]

\[
\frac{k_\tau^2}{k^2} = \frac{1}{2} (\frac{D_1 + D_2}{D_1}) = \frac{1}{2} \left( \frac{Q_2 + Q_1}{Q_1} \right).
\]

Substituting these relationships in (14) gives

\[
Y_\tau \equiv \frac{k_e}{k} \sqrt{G_1G_2} \left( 1 + \frac{k_\tau^2}{k_e^2} \right) \left\{ j \left[ \frac{1 + k^2}{k_\tau^2} \right] \frac{k^2}{1 + k_\tau^2} \right] \left[ \left( \frac{2\delta}{\sqrt{k_e^2 + k_\tau^2}} \right)^2 \right] - \sqrt{2} \left( \frac{2\delta}{\sqrt{k_e^2 + k_\tau^2}} \right) \right].
\]

Several important relations can be obtained from (18). First, the parabola is determined uniquely by the term within the brace. If the transfer admittance is normalized, the distance from the vertex to the focus and the directrix is \( \frac{1}{2} \), as indicated on Fig. 2. Second, the position of the origin of co-ordinates is determined by the coupling coefficients and is located a distance \( (1+k^2/k_e^2)/(1+k_\tau^2/k_e^2) \) below the vertex. Third, the frequency scale is proportional to the real component of the admittance, and a unit value of the frequency deviation parameter \( 2\delta/\sqrt{k_e^2+k_\tau^2} \) corresponds to a distance of \( \sqrt{2} \) along the horizontal axis. (Note that
negative values of frequency deviation correspond to

When the frequency deviation parameter has the
value which makes \( Y_T \) a minimum, i.e.,

\[
\frac{2\delta}{k_x} + k_x = 1
\]

the vertical coordinate of the parabola has a value of
unity. These two points on the curve correspond to the
maximum response of the circuit. As the coupling is
increased, the position of the origin of co-ordinates
moves downward, but the distance between the real
axis and the horizontal line determining the admittance
minima remains constant and the minima occur at
more widely spaced frequencies. All of these relations
are illustrated by Fig. 2.

Low Q Circuits

It is often necessary or desirable to use low \( Q \) circuits,
therefore, the deviation from the simple case when the
gain is a constant \( Q \) of the two circuits is high or more is
of considerable interest. A comparison of the high \( Q \)
case with the low for \( Q \) equal to 10 and 5 is shown in
Fig. 3 for two circuits of equal \( Q \) and identical tuning,
when the coupling constant \( k \) is equal to the critical
coupling \( k' \). Unit values of the frequency deviation
parameter \( 2\delta/k \) or \( k' \) are shown as points on the
curves. Note that the frequency deviation is no longer
linearly related to the real component of the
admittance.

**Effect of Detuning**

All of the previous discussion has been limited to the
case when the two circuits have been tuned to the same
frequency. It has been shown that the effect of
detuning is to shift the vertex of the parabola. This
problem is handled by defining a mid-frequency as the average of the two resonant frequencies by the relation

\[
\omega_m = \frac{\omega_1 + \omega_2}{2}
\]

where \( \omega_1 \) and \( \omega_2 \) are the resonant frequencies of the two
circuits, respectively. A fractional detuning ratio \( \delta_0 \) is
defined by

\[
\delta_0 = \frac{\omega_1 - \omega_0}{\omega_0} = \frac{\omega_2 - \omega_0}{\omega_0}
\]

These relations may be introduced to obtain as a final
result for the detuned case:

\[
\begin{align*}
1 + \frac{2h_0}{k_x + k_f} & = \left[ 1 + \frac{k_x}{k_f} \right] \left[ 1 + \frac{k_f}{k_x} \right] \\
- \frac{2\delta_0}{k_x + k_f} & = \left[ \frac{2\delta}{k_x + k_f} \right] \left[ \frac{2\delta}{k_x + k_f} \right] \\
\end{align*}
\]

Comparison of (22) and (18) shows that the effect of
symmetrically detuning the two circuits adds an
imaginary term which is equivalent to increasing the
coupling coefficient \( k \). In addition, if the dissipation

---

*Fig. 2—Locus of the normalized transfer admittance

\( Y_T \) on the complex admittance plane

\( (k_x/k)\times G_{20}(1+(k_f/k)) \)

*Fig. 3—Effect of \( Q \) on the normalized transfer admittance for two
circuits of equal \( Q \) and identical tuning when the coupling is
equal to the critical coupling. The coordinate scale represents
unit values of normalized susceptance and conductance.*
factor $D$ of the two circuits is not the same, there is a translation of the parabolic locus along the real axis which depends upon the ratio $D_2/D_1$ and the fractional detuning. The shape and size of the parabola are not affected, but the vertex follows a path in the complex plane determined by a family of curves which are also parabolas with a width depending on the ratio $D_2/D_1$. These relations are illustrated in Fig. 4. The location of the focus and the directrix with respect to the vertex are as given in Fig. 2, but have been deleted in Fig. 4 in order to simplify the illustration.

**Input Admittance**

Although the transfer admittance of two tuned coupled circuits is probably used most frequently, the input admittance is sometimes desired. It is a considerably more complicated function than the transfer admittance because the frequency deviation ratio appears in both numerator and denominator. The numerators are the same, therefore it is simple to obtain an expression for the ratio of the two admittances. This ratio permits calculation of the input admittance from the transfer admittance, which can be obtained easily from the parabolic locus, using the relation

$$Y_{\text{input}} = -j \frac{k}{k_c} \sqrt{\frac{G_1}{G_2}} \left(1 + j \frac{Q_2}{Q_1} \frac{2(\delta + \delta_0)}{\sqrt{k_c^2 + k_T^2}} \right).$$

**Triple-Tuned Circuits**

The graphical method as outlined in the preceding sections can be extended, with certain modifications, to triple-tuned circuits, although it is no longer possible to represent the admittance loci by a single universal curve. The analysis will not be given in detail, since it is quite involved, but the results will be stated and illustrated.

If the discussion is limited to the high-Q case with zero coupling between the first and third tuned circuits, and all three circuits are tuned to the same resonant frequency $\omega_0$, the expression for the transfer admittance $Y_{73}$ may be written

$$Y_{73} = \frac{\sqrt{G_1 G_2 (D_1 + D_2 + D_3)^3}}{8k_{12}k_{23}\sqrt{D_1D_3}} \left[4g^4 - (f^2 + 1)^2 + 2x^2 - j(2f^2 + 1)x - x^2\right].$$

The symbols $f^2$, $g^4$, and $x$ are used to simplify the writing of (24) and are defined below. The parameter $x$ is a frequency deviation parameter equivalent to $\gamma_3$ in (13c). The symbols $f^2$ and $g^4$ are functions of the coupling and the dissipation factors of the circuits and replace the coupling coefficients used in the analysis of two tuned circuits.

$$x = \frac{4\delta}{D_1 + D_2 + D_3}$$

$$f^2 = \frac{2}{(D_1 + D_2 + D_3)^2} \left(k_{12}^2 + k_{23}^2 + D_1D_3 + D_2D_3 + D_3D_3 - 1\right)$$

$$g^4 = \frac{(f^2 + 1)^2}{4} - \frac{2}{(D_1 + D_2 + D_3)^2} (D_1D_2D_3 + D_1k_{13}^2 + D_3k_{13}^2)$$

$$k_{12} = \frac{M_{12}}{\sqrt{L_1L_2}}$$

$$k_{23} = \frac{M_{23}}{\sqrt{L_2L_2}}$$

Inspection of (24) indicates that the form of the curve obtained by plotting (24) is fixed once the choice of the value for $f^2$ is made. Typical curves for the portion of (24) within the braces are shown in Fig. 5 to illustrate this point. A value of unity for $f^2$ has been chosen. The effect of varying $g^4$ only shifts the curve along the real axis and does not change the shape of the curve, provided $f^2$ is held constant.

1 See forthcoming paper by N. W. Mather, "An analysis of triple-tuned coupled circuits." The notation has been changed somewhat for this paper.
There is no simple geometrical construction for the locus of the transfer admittance \( Y_{tr} \) similar to the directrix for a parabola. However, there are several points on the locus which are easily located, and these points are shown in Fig. 6. If one or more curves are accurately known, it is not difficult to interpolate a curve for a different value of \( f^2 \) after the crossover point and limits to the curve are located with the aid of Fig. 6. Also, since the real component of the transfer admittance varies as the square of the frequency deviation parameter, the point on the locus corresponding to a given value of \( x \) can be located by means of a parabola with an imaginary component equal to \( x \) and the real component equal to \( (2x^2) \). The vertex will be at the point on the locus corresponding to \( x = 0 \). This construction is illustrated by the dotted curve on Fig. 6. The same parabola may be used regardless of the value of \( f \), provided the vertex of the parabola is located as indicated above.

The most interesting case occurs when the coupling factors are related by the equation

\[
4g^4 = f^4. \tag{30}
\]

This relation produces three equal magnitude minima in the magnitude of the transfer admittance \( Y_{tr} \), provided \( f^2 \) is greater than zero.\(^7\) The special case with

\[
4g^4 = f^4 = 0 \quad \tag{31}
\]

is the transitional case between a one-minimum case and the three-minima case, i.e., the case giving "maximal flatness." When \( f^2 \) is negative, there is only one minimum.

A family of curves similar to Fig. 5 can be plotted for various values of \( f^2 \) as shown in Fig. 7. For convenience, the relation for the three equal minima case defined by (30) is illustrated, although the curves will be applicable to other values of circuit loss and choice of coupling by shifting the \( x = 0 \) intercept along the real axis to correspond to the actual value of \( g^4 \).

Although the factors \( f^2 \) and \( g^4 \) are quite useful for obtaining the transfer admittance when the coupling coefficients and dissipation factors are known, they are not very helpful in solving the more difficult problem of determining the proper coupling to use with circuits of known losses in order to obtain the desired response characteristic. The relation between coupling and circuit losses for the case with three equal minima may be obtained by substituting (30) in (27) and solving (26) and (27) simultaneously for \( k_{22}^2 \) and \( k_{32}^2 \):

\[
k_{12}^2 = \frac{1}{8} \left[ 1 - \frac{D_2}{D_1 - D} \right] \left[ (D_2 + D_0)^2 + 3D_1^2 \right.
\]

\[
+ 2f^2(D_1 + D_2 + D_0)^2] \tag{32}
\]

\[
k_{23}^2 = \frac{1}{8} \left[ 1 + \frac{D_2}{D_1 - D} \right] \left[ (D_1 + D_0)^2 + 3D_2^2 \right.
\]

\[
+ 2f^2(D_1 + D_2 + D_0)^2]. \tag{33}
\]

Since \( f^2 \) is positive for the case with three equal minima, it is necessary to restrict the value of \( D_2 \) in these equations in order to obtain real values for the coupling coefficients. Therefore the relation

\[
D_2 \leq |(D_1 - D_2)| \tag{34}
\]

must be satisfied in order to obtain three equal minima in the magnitude of the transfer admittance. It is obvious that \( D_2 \) can be greater than the value indicated by (34), but the coupling cannot be adjusted to give three equal-minima.\(^7\)

The value of \( f^2 \) used in (32) and (33) determines the type of response. If \( f^2 = 0 \), the transitional case will be obtained, while the choice of \( f^2 = 4 \) will give two maxima in the transfer admittance with a magnitude of approximately \( \sqrt{2} Y_{min} \). These points would correspond to 3 db points in the response characteristic of the circuit.
An Analysis of Interlinked Electric and Magnetic Networks with Application to Magnetic Amplifiers*  

D. W. VER PLANCK† AND M. FISHMAN‡, ASSOCIATE, IRE  

Summary—A general system of equations is developed for the analysis of interlinked electric and magnetic networks. These equations are applied to the study of the steady-state behavior of six basic types of magnetic amplifier without feedback. The equations, which are non linear, are solved to give the currents as functions of time for certain given applied voltages and given circuit and core parameters. An experimental check is included to confirm the correctness of the analysis. A comparison of the results for the six basic types shows that the use of two separate magnetic cores has important advantages over the single three-legged core so commonly used in present practice.

**INTRODUCTION**

MAGNETIC AMPLIFIERS, also called dc-controlled reactors, saturable reactors, and transductors, consist of electric and magnetic elements so interlinked that a direct current controls the reactance of an ac circuit.1–6 One form appears in Fig. 1(a). Here there are two identical iron core transformers with one pair of similar windings in series with a dc control voltage, and the other pair in series with an ac source and a load. The winding polarities are such that fundamental-frequency voltages induced in the two control windings are in opposition. The direct current determines the average saturation of the cores which, in turn, affects the inductive reactance of the ac circuit. With proper design, amplification occurs, since but a small change of control power results in a large change of load power.

Besides this simple arrangement, there are more complex schemes involving cores with contricted sections or several materials and circuits with additional coils to provide bias and feedback. Numerous variations of the elementary magnetic amplifiers, such as shown in Figs. 1 through 4, are possible, all having very similar characteristics. While some of these forms have found favor,7–10 no general discussion of their relative merits could be found.

The purpose of this paper is to develop a more precise analysis than those published heretofore.11,12 While the analysis applies in general to all magnetic amplifiers in both transient and steady states, it is carried through here only for a steady-state solution of certain of the simpler circuits. Circuits employing feedback are treated in a companion paper.

**DESCRIPTION OF SOME BASIC ARRANGEMENTS**

Besides the arrangement in Fig. 1(a), already mentioned, the two others in this figure behave exactly as the first, except that in (b) the use of a single control coil...
through both cores eliminates voltage components at power frequency from all parts of the control circuit, which may be advantageous from the standpoint of insulation.

Instead of two separate cores as in Fig. 1, a single core having two magnetic loops with a common branch may be used as in Figs. 2 and 3. In Fig. 2 the center leg carries a direct component of flux but no power-frequency component, while in Fig. 3 the reverse is true. The form in Fig. 2(b) is very common, while the forms in Fig. 3 are rare, although (a) has been used. Because of magnetic saturation, the behavior of a core with a common portion is quite different from that when the core is split into two separate parts as in (b) and (c) of Fig. 1, even though the cross section of the common part is twice that of each separate core.

Wherever, in Figs. 1, 2, and 3, the load circuit has two separate coils in series, additional arrangements may be derived by connecting these two coils in parallel. For example, Fig. 4 is thus derived from Fig. 1. Parallel connection of control coils is not permissible.

Altogether, there are twelve possible arrangements employing a magnetic network of two loops as in the

Fig. 2—Type 2 magnetic amplifiers. These have two magnetic circuits with a common branch in which there is no fundamental-frequency component of flux.

Fig. 3—Type 3 magnetic amplifiers. These have two magnetic circuits with a common branch in which there is no dc component of flux. This type is derived from Fig. 2 by interchanging the control and load windings.

Fig. 4—Type 4 magnetic amplifiers. These are like Fig. 1, except that the load coils are connected in parallel.

### TABLE I

**Classification of Types Having Magnetic Networks of Two Loops**

<table>
<thead>
<tr>
<th>Type</th>
<th>Form</th>
<th>Magnetic Loops Have Common Branch</th>
<th>Fundamental Power-Frequency Components of Flux Aid or Cancel in Common Branch</th>
<th>Number of Control Windings</th>
<th>Number of Load Windings</th>
<th>Load-Winding Connection</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>a</td>
<td>No</td>
<td>—</td>
<td>2</td>
<td>2</td>
<td>Series</td>
</tr>
<tr>
<td>2</td>
<td>a</td>
<td>Yes</td>
<td>Cancel</td>
<td>2</td>
<td>2</td>
<td>Series</td>
</tr>
<tr>
<td>3</td>
<td>a</td>
<td>Yes</td>
<td>Aid</td>
<td>2</td>
<td>2</td>
<td>Series</td>
</tr>
<tr>
<td>4</td>
<td>a</td>
<td>No</td>
<td>—</td>
<td>2</td>
<td>2</td>
<td>Parallel</td>
</tr>
<tr>
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<td>a</td>
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<td>Yes</td>
<td>Aid</td>
<td>2</td>
<td>2</td>
<td>Parallel</td>
</tr>
</tbody>
</table>
following considerations: These twelve are of six basic types, as shown in Table I. The classification is based on the following considerations:

1. Whether or not the two magnetic circuit loops have a common branch;
2. If there is a common branch, whether the fundamental power-frequency fluxes aid or cancel in this branch; and
3. Whether the load windings (if there are two) are connected in series or in parallel.

As will be shown, these points of difference are important in the analysis. The types are subclassified into forms according to the number of coils in the load and control circuits. The forms of each type differ as to leakage reactances and winding resistances, but these differences are less important than those between types, and are neglected here, so that all forms of a given type are assumed to be equivalent.

**Plan of Analysis**

The object of the analysis is to predict the instantaneous currents as functions of time in terms of magnetic material characteristics, core dimensions, numbers of turns, and impressed voltages for magnetic amplifiers in general. The approach is to treat the magnetic cores as simple lumped magnetic circuits neglecting leakage fluxes, the varying cross sections at corners, and the effects of eddy currents. Thus the magnetic amplifier becomes a system of interlinked electric and magnetic networks. The procedure is to introduce the following relationships:

(a) Kirchhoff's voltage law for each loop of the electric network;
(b) Law of continuity connecting flux density, loop fluxes, and cross sectional area;
(c) The $B$-$H$ characteristic of the magnetic material; and
(d) Ampere's line-integral law for each loop of the magnetic network.

While these determine the problem, a general solution would be extremely laborious, if not impossible, and useful results are obtainable only by further simplification, as will be shown for the specific group of magnetic amplifiers classified in Table I.

**Voltage Equations**

The voltage equations are written in terms of loop currents identified by numerals, the general indices being $n$ and $p$. Similarly, in the magnetic network use is made of loop fluxes identified by Greek letters with the general index $v$.

The positive directions for currents are chosen arbitrarily. The positive directions for fluxes are chosen so that they all traverse a common branch in the same direction; which is apparently always possible if the magnetic network is planar. With this convention, the sign of the number of turns of a coil may be plus or minus, and is determined after the positive directions for currents and fluxes have been chosen.

The system of equations resulting from the application of Kirchhoff's voltage law to the electric network may be written compactly in the matrix form:

$$
\begin{vmatrix}
|e_n| &=& |V_{n}| + |R_{np}| \cdot |i_p| \\
\end{vmatrix}
$$

where $e_n$ = the instantaneous emf applied to the $n$th loop of the electric network (volts)
$N_n$ = the number of turns of the coil in the $n$th electric loop linking the $n$th magnetic loop (dimensionless)
$\phi_v$ = the instantaneous time rate of change of flux in the $v$th magnetic loop (webers/second)
$R_{np}$ = the resistance common to loops $n$ and $p$ of the electric network ($R_{nn}$ is the total resistance of loop $n$) (ohms)
$i_p$ = the instantaneous value of the $p$th loop current (amperes).

**Magnetic Equations**

The magnetic network is assumed to be divisible into a finite number of parts $j$ in each of which the flux density is sensibly uniform. Thus, to express flux density in terms of loop fluxes and cross-sectional area, one finds the total flux in a given part of the magnetic network and divides by the cross-sectional area. To express this relation in a general set of equations which will account systematically for branches common to more than one loop flux, it is convenient to introduce a matrix $|S_{pj}|$, an element of which is the reciprocal of the cross-sectional area of the $j$th part of the magnetic network associated with flux loop $v$. As a consequence of the sign convention for loop fluxes, the elements $|S_{pj}|$ are always positive. Thus, one can write

$$
|B_j| = |S_{pj}| \cdot |\phi_v|
$$

where $B_j$ = the magnetic flux density in the $j$th part of the magnetic network (webers/meter$^2$)
$S_{pj}$ = the reciprocal of the area of the $j$th part of the magnetic network if $\phi_v$ threads that area; otherwise, zero (meter$^{-2}$).

In this analysis, the relation of flux density $B$ to field intensity $H$ need be known only in curve form. Since in the solution $H$ will be found from $B$, $H$ is said to be some known function of $B$, or for the $j$th part of the network

$$
H_j = H_j(B_j).
$$

In the application which follows it is assumed that one magnetic material is used throughout; hence, the subscript $j$ after $H$ on the right of (3) may be omitted. Moreover, hysteresis is neglected, so that (3) becomes the normal magnetization curve.

Because of the assumed uniformity of field in each part $j$ of the core, the line integral of $H$ can be ex-
pressed as the summation of a finite number of products of magnetic field intensity and path length. Thus Ampere's law gives the system of equations:

\[ [N_{r}] \| i_{r} \| \| L_{r} \| \| H_{r} \]  \\
\text{(4)}

where

\( N_{r} = \) the number of turns in the \( r \)th electric loop linking the \( r \)th magnetic loop (dimensionless)

\( L_{rj} = \) the length of the \( j \)th part of the magnetic network if \( \phi_{r} \) traverses it; otherwise, zero (meters)

\( [N_{r}] \) in (4) can be obtained from \( [N_{a}] \) in (1) by interchanging rows and columns.

Application of the Analysis

For illustration, the foregoing analysis is applied to the magnetic amplifiers of Table I. In this application further simplifications are made, as follows:

(a) The voltage impressed on the coils in the load circuit is sinusoidal (load replaced by a short-circuit);

(b) The dc voltage impressed on the control circuit is constant and its source has no impedance to harmonic currents;

(c) Steady-state conditions exist; and

(d) Coil and circuit resistances are very small, although, as will be seen, their effects are not neglected entirely.

While these ideal conditions sometimes may be approximated only roughly in practice, results based on them are, nevertheless, useful for gaining a better understanding of magnetic-amplifier behavior and for drawing comparisons between types.

Solving for the Currents

As the first step in solving for the currents, (1) is applied, resulting in the matrices shown in Table II.

<table>
<thead>
<tr>
<th>TABLE II</th>
</tr>
</thead>
<tbody>
<tr>
<td>Matrices of (1) for the Basic Types</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Type</th>
<th>( [E_{a}] )</th>
<th>( [N_{a}] )</th>
<th>( [\phi_{a}] )</th>
<th>( [R_{a}] )</th>
<th>( [i_{a}] )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 and 2</td>
<td>( E_{a} )</td>
<td>( N_{1} N_{2} )</td>
<td>( \phi_{a} )</td>
<td>( R_{1} 0 )</td>
<td>( i_{1} )</td>
</tr>
<tr>
<td>3</td>
<td>( E_{a} )</td>
<td>( N_{1} - N_{2} )</td>
<td>( \phi_{a} )</td>
<td>( 0 R_{2} )</td>
<td>( i_{2} )</td>
</tr>
<tr>
<td>4 and 5</td>
<td>( E_{a} )</td>
<td>( N_{3} 0 )</td>
<td>( \phi_{a} )</td>
<td>( 0 R_{3} )</td>
<td>( i_{3} )</td>
</tr>
<tr>
<td>6</td>
<td>( E_{a} )</td>
<td>( N_{4} 0 )</td>
<td>( \phi_{a} )</td>
<td>( 0 R_{4} )</td>
<td>( i_{4} )</td>
</tr>
</tbody>
</table>

Here the general values of voltages \( e_{a} \) and of turns \( N_{a} \) are replaced by specific values indicated in Figs. 1 through 4, and the general values of total circuit resistance \( R_{aa} \) are shown as \( R_{11} \) or \( R_{22} \), as the case may be. (In types 4, 5, and 6, \( R_{33} = R_{22} \).) Then, neglecting the resistance drops for the time being and noting that as a consequence \( E_{a} \) must also be a negligibly small quantity, one can solve for \( \phi_{a} \) and \( \phi_{b} \). The result for each type is that

\[ \phi_{a} = \Phi_{m} \sin \omega t + \Phi_{0} \]  \\
\text{(5)}

where \( \Phi_{m} = E_{m}/2N_{2} \) for types 1, 2, and 3, and \( E_{m}/\omega N_{2} \) for types 4, 5, and 6. \( \Phi_{0} \) is a constant of integration, as yet unknown. The other loop flux \( \Phi_{b} \) in types 1, 2, 4, and 5 is \( -\Phi_{m} \sin \omega t + \Phi_{0} \), and in types 3 and 6, \( \Phi_{m} \sin \omega t - \Phi_{0} \). Considerations of symmetry show that the time averages of \( \phi_{a} \) and \( \phi_{b} \) must be the same in magnitude, and thus there is but one constant of integration to be found. For types 4, 5, and 6, only two of the three equations of (1) are needed to determine the fluxes, a fact which proves useful later on.

The next step is to find the flux densities in the different parts of the core, using (2). In doing this, symmetrical cores with parts of uniform area, as indicated in Figs. 1 through 4, are assumed. In the three-legged cores, Figs. 2 and 3, the center leg is assumed to have twice the area of either of the other legs.

Having the flux densities \( B \), the field intensities \( H \) are found using the magnetic characteristic of the material,

<table>
<thead>
<tr>
<th>TABLE III</th>
</tr>
</thead>
<tbody>
<tr>
<td>Matrices of (4) for the Basic Types</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Type</th>
<th>( [N_{r}] )</th>
<th>( [i_{r}] )</th>
<th>( [L_{r}] )</th>
<th>( [H_{r}] )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>( N_{1} )</td>
<td>( N_{2} )</td>
<td>( i_{1} )</td>
<td>( L )</td>
</tr>
<tr>
<td>2</td>
<td>( N_{1} )</td>
<td>( -N_{2} )</td>
<td>( i_{2} )</td>
<td>( 0 )</td>
</tr>
</tbody>
</table>
assumed here to be the normal magnetization curve. The
time functions representing values of \( II \) which occur are
indicated symbolically in the last column of Table III
where the order of matrix elements is first the part of the
core carrying \( \phi_a \) alone, then the part with \( \phi_b \) alone, and
finally, if present, the part carrying their sum. The
meaning of the symbols, using the notation of (3), is
\[
\begin{align*}
II_+ &= II(B_m \sin \omega t + B_0) \\
II_- &= II(-B_m \sin \omega t + B_0) \\
II_0 &= II(B_0) \\
II_m &= II(B_m)
\end{align*}
\] (6)

where \( B_m \) is \( \Phi_m/A \) and \( B_0 \) given by \( \Phi_0/A \), replaces the
unknown constant of integration.

The actual time variation of \( II \) is determined graphically
for any assumed value of \( B_m \), as illustrated in Fig.
5 by projecting the displaced sinusoids of \( B \) at the upper
left across to the magnetization curve, and then down
onto a new time scale to give the curves of \( II_+ \) and \( II_- 
\)
shown at the lower right.

Fig. 5—Graphical construction using the normal magnetization
curve of the core material to get the time variation of the field
intensities.

The currents are related to the field intensities by
(4), which is expanded in Table III. Here \( L \) is the total
length of either flux loop, while \( L' \) represents the length of
a part carrying either \( \phi_a \) or \( \phi_b \) alone, and \( L'' \) a part
carrying their sum. For the first three types, these pairs of
equations are solved for the two currents with the
results shown in Table IV. For types 4, 5, and 6, how-
ever, there are three unknown currents, and so another
equation is needed. It is found by manipulating the
voltage equations so as to eliminate the time derivatives of flux, but now not neglecting the resistances. The
resulting equation and the pair in Table III are then
solved simultaneously for the three currents, with the
results shown in Table IV. Here two new quantities ap-
pear: \( I_1 \), which is the average (dc) value of control cur-
rent given by
\[
I_1 = E_{dc}/R_{11},
\]
and \( k \), given by
\[
k = (2R_{12}/R_{11})(N_1/N_2)^2.
\] (8)

The physical significance of \( k \) is that in types 4, 5, and
6 there are two low-resistance circuits linking the two
flux loops similarly; the control circuit and the series
path through the load coils. Then, even though the
resistances otherwise may be negligibly small, their ratio
is significant in fixing the division of induced currents
between the paths.

**Determining the Integration Constant**

The expressions for the currents in Table IV depend through (6) on the average value of flux density \( B_m \) so
far unknown. To determine \( B_m \), one finds \( I_1 \), the dc value
of control current, by taking the time average of \( i_1 \); thus,
\[
I_1 = \frac{1}{2\pi} \int_0^{2\pi} i_1 d(\omega t).
\] (9)

Evaluating this, one finds for types 1 and 4, 2 and 5,
and 3 and 6, respectively, the following values of \( N_1 I_1 \):
$LH_{av}, L''H_{av} + L'''H_0,$ and $L''H_{av}$ where

$$H_{av} = \frac{1}{2\pi} \int_0^{2\pi} H(B_m \sin \omega t + B_0) d(\omega t).$$

(10)

This integral is evaluated graphically as the net area between the curve of $H_{av}$, illustrated in Fig. 5, and its time axis. Since one wishes to find $B_0$ for a given $I_1$, a process of trial and error is necessary, or a family of curves may be constructed as in Fig. 6. This family represents a property of the material independently of the frequency, the core configuration, and circuits assumed.

To obtain $B_0$, then, for given $E_m$ (which determines $B_m$) and given $I_1$ one uses the curves in Fig. 6 directly, or, in the case of types 2 and 3, a trial and error procedure involving Fig. 6.

**Fig. 6—Effect of superimposed alternating flux density on the dc magnetization curve. The upper curve is the measured normal magnetization curve; the others are calculated from it. This family represents a property of the core material independently of the core geometry and circuit conditions.**

**Experimental Check**

The method of analysis has been checked by comparing the calculated currents with oscillograms for a variety of conditions. An example is given in Fig. 7 which applies to type 2, form (b), shown in Fig. 2, using the core material described by Fig. 6. The check is good, even though the conditions are such as to make the flux density reach very high values, which accounts for the large harmonics in the currents. With this material the effect of neglecting hysteresis is imperceptible in a case like this, but it is observable, though never large, in cases where the maximum flux density is low.

**Fig. 7—Oscillograms of currents and applied ac voltage compared with calculated currents for circuit (b) in Fig. 2. The calculated curves are plotted to scales determined by the oscillograph sensitivity.**

- $N_1 = 2,000$ turns
- $R_1 = 135$ ohms
- $L_1 = 0.34$ amperes
- $E_m = 1.2 V$ volts
- $L'' = 0.056$ meters
- $A = 4.6 \times 10^{-4}$ meters
- $\omega = 2\pi 60$ (sec.)

**Series and Parallel Connections Compared**

To compare series and parallel connection of load coils, types 1 and 4, the ones with separate cores, are considered first. All conditions are assumed identical except that $N_2$ for type 4 is twice that for type 1, so that the ac volts per turn is the same in both. Then the expressions for total load current ($i_2$ for type 1 and $i_2 + i_3$ for type 4) shown in Table IV are identical. Moreover, the relation between $B_0$ and $I_1$ is the same for both, and thus, under these conditions, the flux densities in the respective cores of the two types are equal at every instant, and the two have identical external characteristics except for differences in the instantaneous values of $i_3$. The instantaneous values of mmf for each magnetic loop are the same in the two types, but because of the additional electric loop in type 4 the distribution of ampere-turns among the electric circuits differs in the two types. Analyzed in another way, it can be shown that even harmonic components of ampere-turns are necessary to produce the sinusoidal variations in the two fluxes. In type 1 these harmonics are carried entirely by the control circuit, while in type 4 they are carried in part by the series path through the two load coils. It should be noted that, had the control circuit not been assumed to have zero impedance, the series and parallel connections would not be equivalent.

Comparisons of type 2 with type 5 and of 3 with 6 also lead to the conclusion that the series and parallel connections are equivalent. Thus types 4, 5, and 6 may be eliminated from further comparisons.
CORE ARRANGEMENTS COMPARED

To study the relative merits of types 1, 2, and 3, it is assumed that the purpose of a magnetic amplifier is to cause the greatest change in reactance—that is, the greatest change in load current with constant &volage across the load coils—with the least change in control current. For a given type and given applied voltage, the load current $I_2$ is some function of the average control current $I_1$, and of time $t$. The intention is to study in the three types the effect of $I_1$ on $N_2b(I_1, t)$. Values of this function with the control current switched on and off are shown in columns 3 and 4, respectively, of Table V.

Here a new quantity is introduced:

$$H_d = \frac{H_+ - H_-}{2} = \frac{H(B_m \sin \omega t + B_0) - H(B_m \sin \omega t - B_0)}{2}, \quad (11)$$

which varies in time phase with $H_m$. In studying Table V, it is to be noted that, because of saturation,

$$H_{sv} > H_0 \quad (12)$$

$$H_d > H_m. \quad (13)$$

Also, the various $H$'s all are functions of the flux densities, and therefore direct comparisons of items in the table must be made only with $B_m$ and $B_0$ the same in all three types. Furthermore, core total lengths and areas must be the same throughout. Thus comparisons of the currents, which in general will differ from type to type, are made with applied &volage, $B_0, N_1, N_2$, and weight of core held constant.

As a first criterion, take the difference of load current produced, Table V, fifth column. This difference is identical for types 2 and 3, but requires more control excitation in type 2 than in type 3. In this respect, then, type 3 is superior to type 2. Now compare type 1 with type 3. The difference in $I_2$ produced in type 1 is greater than in type 3 in the ratio $L/L'$, but the control current required by type 1 is greater in the same ratio. But type 1 is superior to type 3 because with the same amount of magnetic material a greater change in load current can be obtained, although to realize this superiority it might be necessary to increase the copper in type 1 over that in type 3.

The conclusion is different if one takes as the criterion the ratio of the load current resulting from $I_1$ to that with $I_1=0$, as shown in the last column of Table V. This ratio is identical for types 1 and 2, but is produced by a smaller $I_1$ in type 2; type 2 is thus better than type 1 in this respect. The relative standing of type 3 is not so apparent, but it is probably inferior to type 2.

These comparisons are based on theoretical considerations for one assumed kind of operation; practical considerations, however, might rule. For instance, difficulty of matching two separate cores might outweigh their theoretical advantage and dictate the use of the three-legged core where every lamination is in both magnetic circuits.

CONCLUSION

A general system of equations for interlinked electric and saturable magnetic networks has been developed. As an example of their application, the equations have been applied to the steady-state solution of certain common magnetic-amplifier circuits without feedback, and in this connection a systematic classification of these magnetic amplifiers into basic types was developed. Conclusions of practical interest were reached as to the relative behavior of the several types of magnetic amplifier studied.

ACKNOWLEDGMENTS

The authors gratefully acknowledge the helpful criticisms of B. R. Teare, Jr., and L. A. Finzi of the Department of Electrical Engineering, Carnegie Institute of Technology, and the contributions of the many graduate students assigned to the project in various capacities.

### TABLE V

<table>
<thead>
<tr>
<th>Type</th>
<th>$N_1 I_1$</th>
<th>$N_2b(I_1, t)$</th>
<th>$N_2b(0, t)$</th>
<th>$N_2b(I_1, t) - N_2b(0, t)$</th>
<th>$N_2b(I_1, t)$</th>
<th>$N_2b(0, t)$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$LH_{sv}$</td>
<td>$LH_d$</td>
<td>$LH_m$</td>
<td>$L(H_d - H_m)$</td>
<td>$L_d/H_m$</td>
<td>$L_d/H_m$</td>
</tr>
<tr>
<td>2</td>
<td>$L' H_{sv} + L'' H_0$</td>
<td>$L' H_d$</td>
<td>$L' H_m$</td>
<td>$L'(H_d - H_m)$</td>
<td>$L/H_m$</td>
<td>$L/H_m$</td>
</tr>
<tr>
<td>3</td>
<td>$L' H_{sv}$</td>
<td>$L' H_d + L'' H_m$</td>
<td>$L' H_m$</td>
<td>$L'(H_d - H_m)$</td>
<td>$L/H_m$</td>
<td>$L/H_m$</td>
</tr>
</tbody>
</table>
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Dr. El-Said was engaged in an engineering mission to England and the United States for the Egyptian Government through a fellowship during the period 1945 to 1948. This project was made possible through the co-operation of Marconi's Wireless Telegraph Company in England, and General Radio Company in the United States. Dr. El-Said has invented various types of multiplying circuits, one of which has been developed by General Electric for a general purpose electronic wattmeter. Dr. El-Said has returned to Fouad University where he is a senior lecturer in engineering.

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For a photograph and biography of M. Fishman, see page 901 of the August, 1949, issue of the PROCEEDINGS OF THE I.R.E.
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For a photograph and biography of J. Hessel, see page 1434 of the November, 1948, issue of the PROCEEDINGS OF THE I.R.E.

For a photograph and biography of D. W. Ver Planck, see page 902 of the August, 1949, issue of the PROCEEDINGS OF THE I.R.E.

Correspondence

The Theory and Design of Progressive and Ordinary Universal Windings

A paper by Kantor contains several definite contributions to the science of universal coils. However, due to rather unfortunate wording, a number of statements made and conclusions reached could easily be misinterpreted. It is proposed to clarify these.

1. Accuracy of the Basic Equation of the Universal Winding

From a reading of the Summary, particularly the statement that "equations are derived which are considerably simpler and at the same time more accurate than those given by Simon," one might infer that the form of the basic equation of the universal winding as given by Kantor is more accurate than that originally derived by the writer, whereas, as Kantor himself points out subsequently in the body of his paper, both equations express absolutely identical relationships, the particular one given by Kantor being merely an alternative form of the other. In fact, his equation can be obtained directly from that originally given by the writer merely by inverting both sides and rationalizing the denominator of the fraction appearing on the right-hand side of the resulting equation. The only new contribution made by Kantor in this connection is to show that the expression for the reciprocal of the gear ratio, as defined by the writer, can be put into somewhat simpler form than that for this gear ratio itself. Kantor's derivation of the basic formula of the universal winding is, of course, exactly parallel to that originally given by the writer, except that the gear ratio $r$ is replaced throughout the derivation by its reciprocal $1/R$.

Also from a reading of the Summary, particularly the statement that "the present paper offers a more thorough treatment of the subject... by employing theoretical expressions to replace previously required empirical rules," one might infer that the formulas given originally by the present writer for the gear ratio and the number of crossovers (throws) per turn were largely empirical. There was, however, nothing empirical about either the deduction or the form of these equations: the only empirical thing about them was that the constant $k$ in the crossover formula had to be determined empirically. Kantor's contribution in this connection is to show how the value of this constant can be deduced from a knowledge of the coefficient of friction between the material of the insulation and that of the dowel, thus making the empirical determination of $k$ unnecessary.

2. Selection of the Number of Crossovers (Throws) per Turn

From Kantor's theory on the factors which influence the selection of the number of crossovers per turn, it might be inferred that it is necessary in practice to vary this factor in accordance with the coefficient of friction between the surface of the dowel and that of the insulation; that is to say, with every change of material of the dowel or of the insulation. Actually, this is not the case, for the following reasons: (1) After the first layer is down, the wire is wound not on the surface of the dowel but on the surface of the insulation of the underlying layer; hence, the discussion given by Kantor applies primarily to the problem of producing a stable first layer. (2) The necessity for altering the number of crossovers per turn with every change of material of the dowel can be obviated entirely by the simple expedient of roughening the surface of all dowels (if necessary) prior to winding. In fact, it might be noted parenthetically in this connection, that winding on smooth, i.e., polished or glazed dowels, imposes an unnecessary handicap in manufacture, and should be avoided in any case. Hence, while Kantor's analysis of the factors determining the selection of the number of crossovers per turn, in particular the deduction of the constant $k$ from the coefficient of friction $\mu$, is a definite contribution to the science of universal coils, it is of more academic than practical importance. In the writer's experience the insertion of the empirical constant $k$ in the crossover formula has always given satisfactory results, irrespective of the type of dowel or textile insulation used, provided only that the surface of the dowel was sufficiently rough. In fact, a value of $\frac{1}{2}$ for this constant would correspond, according to Kantor's theory,
Correspondence

to a value of $\mu$ of 0.21, which is in good agreement with the values actually found by him for such materials as cardboard, wood, bakelite, etc.; that is, for materials generally used.

3. Determination of the Progression

With reference to the pitch of the progression, which determines directly the number of turns per unit length of the coil, it should be pointed out that, contrary to the impression one might receive from a reading of Kantor's discussion of the subject, this factor can be arbitrarily selected in general; it is fixed or predetermined only if it is prescribed that the winding is to fulfill also a certain geometric condition; in particular, if it is to exhibit a certain ratio of width of close-packed layer to spacing between close-packed layers. Actually, however, in practice the number of turns per unit length, which will determine, for example, such factors as total band coverage in the case of permeability-tuned coils, will be of greater importance in design than the particular geometric ratio referred to; hence, the pitch will usually be selected independently and the geometric ratio allowed to take what value it may (between limits). It is true, of course, that a value of this ratio of $\frac{b}{a}$, as recommended by Kantor, is desirable but not at all a necessary condition.

4. Accuracy of Various Approximate Formulas for the Gear Ratio (Ordinary Universal Winding)

At the outset of a discussion of the accuracy of a formula for the gear ratio, it should be pointed out that the fundamental factor to be considered in this connection is not the error in the gear ratio itself, but the error in the gear-ratio parameter, since the latter determines directly such fundamental magnitudes as the number of turns per layer, spacing between centers of adjacent wires, etc. The gear-ratio parameter $P$ is the quantity defined by the equation:

$$ r = T_c / T_D = (2/n)(1 \pm 1/P) $$

as explained in a previous paper. Of the various gear-ratio formulas proposed, we have the original (accurate) form given by the author:

$$ r = 2/n (1 \pm \sqrt{a^2 + b^2}) $$

the alternative (accurate) form, which can be derived merely by the inversion of (1), as given by Kantor:

$$ r = T_D / T_c = 2/n (1 \pm \sqrt{a^2 + b^2}) $$

the approximate form, derived from (1), as given by the author:

$$ r = (2/n)(1 \pm \sqrt{a^2 + b^2}) $$

another approximate form derived from (3) by neglecting the $a^2$ term:

$$ r = (2/n)(1 \pm \sqrt{a^2 + b^2}) $$

the first approximate form of Hershey.\footnote{A. W. Simon. \textit{"Winding the universal coil.\textquotedblright} Electronics, vol. 9, pp. 22-24; October, 1936. Errata, p. 52; November, 1936.}

and, finally, a second approximate form, given originally also by Hershey and later by Kantor:

$$ R = (n/2)(1 \mp a) $$

where the upper sign refers in all cases to progressive layering and the lower one to retrogressive layering.

The error introduced in the gear-ratio parameter by the use of the various formulas obviously will depend on the magnitude of the quantities $a$ and $b$; hence, it becomes of interest to determine what range of values these quantities take in practice. They are defined by the relations

$$ a = q_c / q_r = 1/q_r $$

$$ b = m / b_d = a k / m $$

where $q_r$ represents the number of crossover wires per layer, $w$ the number of wires per layer, and $k$ the empirical constant in the crossover formula. The minimum value of $q$ in any case is 2, so that if we let $w$ vary from the rather extreme case of only 5 wires per layer to the more typical case of 50 wires per layer, and take for $k$ the recommended value of 0.1, we have $a$ varying from 0.10 to 0.01 and $b$ from 0.021 to 0.0021. The error introduced into the gear-ratio parameter under these conditions is given in Table I.\footnote{4. L. M. Hershey. \textit{"The design of the universal winding.\textquotedblright} Proc. I.R.E., vol. 29, pp. 442-446; August, 1941.}

\begin{table}[h!]
\centering
\caption{Error in the Gear Ratio Parameter}
\begin{tabular}{|c|c|c|c|c|c|}
\hline
\textbf{a} & (1) & (2) & (3) & (4) & (5) & (6) \\
\hline
0.10 & none & 1.0\% & 11.0\% & 13.4\% & 2.0\% & 2.2\% \\
0.01 & none & 0.0\% & 0.9\% & 3.2\% & 2.0\% & 2.2\% \\
\hline
\end{tabular}
\end{table}

From the Table it is seen that, for the range considered, the approximate formula (3) originally given by the author is more accurate than any of the other approximate formulas, although in the range most likely to occur in practice ($a=0.01$), the difference between the values given by the various formulas is not marked. The Hershey-Kantor form (6) has the advantage over the form (3) originally given by the author in that it is simpler, and over the other approximate formulas (4) and (5) in that it holds over a wider range for a given degree of accuracy. The reason for the error in (6) is, of course, that the effect of diameter (involved in $b^2$) has been neglected. This with the exception, perhaps that in progressive universal-coil design the approximate formulas (10) or (11) can advantageously replace the formulas originally given by the writer; while in ordinary universal-coil design, (6) or (9) can advantageously replace (3). Varying the number of crossovers per turn with a change of material of the dowel, or selecting the pitch of the progression to produce a spacing ratio of 0.5, however, are unnecessary procedures and, accordingly, are not recommended except in special cases.

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\footnote{See footnotes references 2 and 3, and also: A. W. Simon. "Universal coiil design." Radio, vol. 31, pp. 16-17; February-March, 1947.}
The Duo-Mode Exciter*

During the course of an investigation at the Northwestern University Microwave Laboratory of multiplexing systems using two or more modes, a device was required for independently launching the $TE_{01}$ and $TE_{11}$ modes in the same guide. It was necessary that this device have a low voltage-standing-wave-ratio (VSWR) over a wide frequency range with a minimum of cross talk between inputs.

The duo-mode exciter is shown in Fig. 1. The vertical arm we shall designate the $E$ arm, and the horizontal arm the $H$ arm. When energy is transmitted down the $H$ arm in the $TE_{01}$ mode, it widens along the symmetrical 45° nozzle to form a $TE_{01}$ mode in the duo-mode guide. When energy is transmitted down the $E$ arm in the $TE_{11}$ mode, it forms components in each side of the duo-mode guide that are 180° out of phase. This $TE_{01}$ mode is reflected at some point along the $H$ arm nozzle at which the width of the guide is less than cutoff for the $TE_{01}$ mode. The distance of this nozzle from the $E$ arm (see Fig. 1) is such that the $TE_{01}$ mode is reinforced in the direction away from the $H$ arm.

The $TE_{01}$ mode from the $H$ arm cannot propagate up the $E$ arm, since the 0.4-inch dimension of the $E$ arm is less than cutoff width for the $TE_{01}$ mode. Consequently, the $E$ arm and $H$ arm are isolated from each other.

The results of the experimental test of the duo-mode exciter are given in Fig. 2. Without any additional matching, the VSWR remained below 1.6 from a frequency of 8,830 to 9,530 Mc. During these tests the duo-mode guide was terminated with a load having a VSWR of about 1.02 with either mode. With small posts about 0.06 inch in length and positioned in the $E$ arm and $H$ arm as shown in Fig. 1, the VSWR remained below 1.6 from a frequency of 8,960 to 9,740 Mc. Any frequency in this range could have been chosen as center frequency for matching. Our center frequency was 9,375 Mc and the VSWR at this frequency was 1.02 looking into the $E$ arm, and 1.04 looking into the $H$ arm.

Fig. 2—Voltage-standing-wave-ratio of duo-mode exciter.

The separation between inputs was 37 db at 9,375 Mc with the duo-mode guide terminated with a load having a VSWR of 1.02 with either mode. The modes produced were very pure, the null point of the $TE_{01}$ mode being more than 60 db below the peak value of electric intensity of the $TE_{01}$ mode.

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Note on the Theoretical Efficiency of Information Reception with PPM*

For small $P/N$ ratios, the now classical expression for the information reception capacity of a channel

$$C = W \log (1 + P/N)$$

can be written, substituting $kTW$ for $N$,

$$CT_s = W T P/N \log e = \frac{P T_s}{kT} \log e = \frac{E}{kT} \log e$$

where $E$ designates the energy available for the reception of the information packet containing $CT_s$ binary symbols.

If quantized PPM and an ideal low-pass channel are utilized, and if the available energy $E$ is concentrated into one pulse, the ratio of the pulse height to the rms thermal noise will be $2E/kT$, and the voltage gate should be set at a proper fraction $a(2E/kT)$ of the pulse height, so as to minimize the mathematical expectation of a random positive thermal pulse exceeding this gate at any one of $N$ prearranged sampling epochs, or of a random negative thermal pulse preventing the information carrying pulse from exceeding the gate at the proper epoch. This mathematical expectation can be written:

$$N \left( \frac{2}{\sqrt{\pi}} \right) \int_{a \sqrt{2E/kT}}^{\infty} e^{-x^2} dx$$

$$+ \left( \frac{2}{\sqrt{\pi}} \right) \int_{(1-a) \sqrt{2E/kT}}^{\infty} e^{-x^2} dx,$$

and the minimizing operation just indicated determines the optimum value for $a$:

$$a = \frac{1}{2} \pm \frac{kT}{2E} \log e N.$$
IRE-URSI to Meet in Fall

The regular IRE-URSI Fall Meeting, sponsored jointly by the IRE Professional Group on Wave Propagation and Antennas and the U. S. A. National Committee of the International Scientific Radio Union, will be held on Monday, Tuesday, and Wednesday, October 31 and November 1 and 2, in Washington. Four U. S. A. National Commissions of URSI will participate: Commission 2—Tropospheric Radio Propagation; Commission 3—Ionospheric Radio Propagation; Commission 5—Extraterrestrial Radio Noise; and Commission 6—Radio Waves and Circuits, including General Theory and Antennas.

The first two days will be devoted to individual or joint meetings of two or more Commissions, held in the National Academy of Sciences, 2101 Constitution Avenue, N. W., and the Auditorium of the New State Department Building, Twenty-first and Virginia Avenues, N. W. The third day will consist of a general assembly of all Commissions, followed by individual administrative meetings of the several Commissions.

Meetings will be of a modified symposium type, consisting of invited papers, contributed papers, and informal discussions at the discretion of the chairmen.

Industrial Engineering Notes

Electronic Brain Developed

A fundamental advance in the organization, storage, and dissemination of knowledge is foreseen in an "electronic brain" developed jointly by the U. S. Departments of Commerce and Agriculture. The machine stores "vast amounts of scientific information in its system," selects what is desired by its operator, and then hands him copies of the material requested. A report describing the Rapid Selector in detail (PB 97535) is available from the OTS, Department of Commerce, Washington 25, D. C., for $2.50 each copy.

Interference Suppression for Arc Welders Developed

A successful method for effectively suppressing radio interference caused by the operation of high-frequency stabilized arc welders was reported on by the Signal Corps. With the use of a double-screened room and with adequate filtering of the power line at the point of entry into the screened room, it was possible to reduce radio interference outside the room to a point where measurement was almost impossible. The

Notes

The data on which these Notes are based were selected, by permission, from "Industry Reports." Issues of June 17 and 24, and July 1 and 8, published by the Radio Manufacturers' Association, whose helpful attitude is gladly acknowledged.

Institute News and Radio Notes

Technical Committee Notes

The Radio Transmitters Committee met on June 13 at IRE Headquarters, with J. F. McDonald, Chairman, presiding. At present the committee is concerned chiefly with the preparation of Standards on Methods of Testing Transmitters. J. B. Heffelfinger has been appointed Chairman of the Radio-telephone Subcommittee, which will operate in the midwest. Reports on the status of work were submitted by the subcommittees on Telegraph Transmitters, H. R. Butler, Chairman; Pulse Transmitters, Cedro Brunetti, Chairman; Single Side Band Transmitters, A. E. Rerwin, Chairman; and FM Transmitters. A meeting of the Wave Propagation Committee was held on June 20 under the Chairmanship of C. R. Burrows to consider the FCC's Ad Hoc Committee's Report for The Evaluation of the Radio Propagation Factors concerning the Telemetry and Frequency Modulation Broadcasting Services in the Frequency Range between 50 and 250 Mc. The results were summarized in a report submitted to the JTAC. The Committee on Electron Tubes and Solid State Devices met at Headquarters on June 2, with L. S. Neergaard, Chairman, leading the discussion. A proposal to have this Committee sponsor a Professional Group on Electron Tubes was tabled temporarily. The relative merits of operating the Electron Tube Conference with simultaneous sessions and extending the length of the meeting as compared to having two short conferences during the year were discussed and a consensus of opinion favored extending the length of the Conference by one day. On June 8 the Electroacoustics Committee headed by E. S. Seeley met. F. V. Hunt, H. F. Olson, W. F. Meeker, and P. S. Veneklaas were appointed members of the new subcommittee on Loudspeaker Testing, with M. J. Di Toro as Chairman. B. B. Bauer, Chairman, and H. F. Olson make up the subcommittee appointed to study the proposed ASA Standard on Secondary Microphone Calibration for the IRE. P. F. Siling, Chairman of the Joint Technical Advisory Committee, announced the appointment of Donald G. Fink as Chairman and John V. L. Hogan as Vice-Chairman for the coming year, July 1, 1949, to June 30, 1950 at the June 23 meeting of the Committee. It was agreed to bind the official correspondence between the FCC and the JTAC, together with the minutes of all JTAC meetings to date. This volume will be released as the PROCEEDINGS OF THE JTAC, Volume III. The Nucleonics Symposium Planning Group for the Second Annual Joint IRE-URSI Conference on Electronic Instrumentation in Nucleonics and Medicine met on June 15, with Harner Selvidge, Chairman, presiding. The Symposium is scheduled for October 31 and November 1 and 2, and will be held at the Hotel Commodore in New York City. The papers which will be presented at the Symposium will be published in a single volume entitled "Proceedings of the Symposium," which will be made available to those registering or upon order at an extra charge to those who did not attend the meetings. W. A. Geoghegan is Chairman of the Papers Procurement and Program Committee; R. D. Chipps of the Committee on Local Arrangements; Norman Beers of the Publicity Committee. Ward Davidson is Treasurer. The Papers Procurement and Program Committee of the IRE-AIEE Planning Group for the Symposium met on June 27, and the papers to be presented will be announced at an early date.

IRE-URSI MEET

The U. S. A. National Committee of the International Scientific Radio Union (URSI) and the Institute of Radio Engineers held a joint technical meeting in Washington, D. C. on May 2, 3, and 4. General open technical sessions were held on May 2 and 3, and organization meetings of the four U. S. A. National Commissions which sponsored the meeting on May 2 and 4. Twenty-seven fundamental scientific and research papers were presented on radio standards, methods of measurement, terrestrial radio noise (natural and man-made), communication theory, antennas, and circuits.


Abstracts of the papers were prepared in booklet form and can be obtained at $1.00 each, and may be obtained from Newbom Smith, Secretary, U.S. A. National Committee, URSI, National Bureau of Standards, Washington, D. C.
FCC REGULATIONS

The FCC issued an initial decision renewing the license of Sarkes Tarzian for a high-frequency AM experimental broadcasting station at Bloomington, Ind. The experiments in transmission and reception of high-frequency AM broadcasting have been carried on since the spring of 1946. . . . An order revising its rules governing performance measurements of AM and FM broadcast systems was issued recently by the FCC. The Commission rules require all AM and FM stations to make certain performance measurements at yearly intervals, with one such set of measurements being made during the four-month period preceding the date of filing application for renewal of station license. These rules were eased so that applicants for renewal of licenses expiring prior to February 1, 1950, are not required to indicate that these measurements have been made. The full requirements for AM and FM broadcast stations are contained in Part 3 of the “Commission Rules and Regulations” and the “Standards of Good Engineering Practice Concerning Both Standard and FM Broadcast Stations,” which can be purchased from the Superintendent of Documents, Government Printing Office, Washington 25, D. C. . . . The FCC issued a check list of its rules and regulations to enable individuals possessing books of the Commission’s Rules and Regulations to check for completeness. The list (Mimeograph No. 37927), which brings the rules up to date as of June 27, 1949, may be obtained from the Secretary of the FCC, Washington 25 D. C. . . . Printed copies of the new rules governing the mobile and other nonbroadcast services involved in the Report and Order Dockets 8638, 8965, 8972, 8973, 8974, 9001, 9018, 9046, and 9047, issued by the FCC May 3, 1949, are now available from the Superintendent of Documents, U. S. Government Printing Office.

TELEVISION NEWS

E. U. Condon, Director of the National Bureau of Standards, is assembling a committee of independent authorities to study the present status and future of color television. The objective of the study will be to determine the present status of color television and to estimate when color television may be feasible for public service and commercial operation. Among those asked to serve on the committee were Donald G. Fink, JTAC Vice-Chairman and editor of Electronics; Stuart L. Bailey, President of the IRE; William L. Everitt, head of the University of Illinois’ department of electrical engineering; and Newbern Smith, Chief of the Central Radio Propagation Laboratory at the Standards Bureau. Dr. Condon would serve as Committee Chairman. . . . The FCC explained its rule preventing the separate operation of aural and visual transmitters of a television station, stating that it is intended to insure that television channels shall be used only for simultaneous visual and aural television programming and for incidental or test purposes, and not for separate aural broadcasts. To permit a television sound channel to be used either to duplicate AM or FM aural broadcasts, or to originate aural broadcasts only, would not be an economical use of radio frequencies and would not be in the public interest. . . . Precautionary cathode-ray safety rules for tube and set manufacturers, service men and dealers, and television set owners were issued by the RMA Cathode Ray Safety Committee. The Committee laid emphasis on the fact that the cathode-ray tube is not dangerous except when improperly handled, and stated further that rumors concerning the harmful effects of ultra-violet rays reputedly emitted by picture tubes are unfounded. . . . At the end of June there were 69 commercial television stations on the air. There were 49 construction permits outstanding, and 386 applications pending but “frozen.”

RADIO AND TELEVISION NEWS ABROAD

Prime Minister J. B. Chifley of Australia announced that his country will use a television standard of 625 lines when television is inaugurated there as a government monopoly. The use of a greater number of picture lines than either the British (405) or the American (525) should, he stated, ensure a better image than is available under either of the other two standards. Australia plans to erect stations in its six capital cities—Brisbane, Sydney, Melbourne, Adelaide, Perth, and Hobart. . . . Five television manufacturers are now producing television receivers in Canada, with the present Canadian market estimated at 1,500,000 persons residing along the United States border within the range of American television stations at Toledo, Buffalo, Rochester, Detroit, Cleveland, and Seattle. Sales of radio receiving sets by Canadian manufacturers in March totaled $4,050,501, compared with 40,551 sets valued at $3,978,361 during the corresponding month in 1948.

PRODUCTION NOTES

Television receiver production by RMA member-companies in May was slightly under the previous month’s output. May’s production was 163,262 sets, as compared with 166,536 in April. For the first two months of 1949, 752,335 television sets were produced; 383,869 FM-AM and FM sets; and 2,586,135 AM only sets—all sets totaling 3,722,339.

Calendar of COMING EVENTS

1949 National Electronics Conference, Chicago, III., September 26–28
SMIE 66th Semiannual Convention, Hollywood, Calif., October 10–14
AIEE Midwest General Meeting, Cincinnati, Ohio, October 17–21
Radio Fall Meeting, Syracuse, N. Y., October 31, November 1–2
1949 Nucleonics Symposium, New York, N. Y., October 31, November 1–2
1950 IRE National Convention, New York, N. Y., March 6–9

Books

Table for Use in the Addition of Complex Numbers (Table til Brug ved Addition af Komplekse Tal) by Jørgen Rybnér and K. Steenberg Sørenson

Published (1948) by Jul, Gjellerup Forlag, Copenhagen, Denmark; obtainable from Scandinavian Book Service, P. O. Box 99, Audubon Station, New York 32, N. Y., 95 pages $1.50, pages 91–121 $5.50.

Published in parallel Danish and English texts, this pamphlet contains a table which facilitates calculations with complex numbers by rendering possible the addition or subtraction of such numbers in polar form. It is actually a supplement to Professor Rybnér’s “Nomograms of Complex Hyperbolic Functions” (see page 1271 of the October, 1948, issue of the Proceedings for review), which includes the conversion between rectangular and polar co-ordinates and the function $R/a = 1 + r/\phi$ represented in this table.

It was considered appropriate to prepare an extended numerical table over 90 pages long of this function, because it was found impossible to construct the corresponding nomograms with an accuracy sufficient for practical calculations. The function $R/a = 1 + r/\phi$ is represented giving $R$ and $a$ as functions of $r$ and $\phi$ for $0 \leq \phi \leq 180^\circ$ at intervals of $0.01$ and for $0 \leq a \leq 1$ at intervals of $1^\circ$. These quantities are connected by the following relations:

$$R^2 = 1 + r^2 + 2r \cos \phi$$
$$\sin a = \frac{r}{R} \sin \phi$$
$$\tan \left( \frac{\phi}{2} - a \right) = \frac{1 - r \phi}{1 + r \phi}$$

in the following forms:

$$\frac{\phi}{2} = \frac{1 - r \phi}{1 + r \phi}$$

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1949 report (PB 97469) may be obtained from the OTS, U. S. Department of Commerce, Washington 25, D. C.
PROCEEDINGS OF THE I.R.E.

First Call

Authors for National Convention

R. M. Bowie, Chairman of the Technical Program Committee for the 1950 IRE National Convention, requests that prospective authors of papers to be considered for presentation submit the following information to him as soon as possible:

1. Name and address of author.
2. Title of paper.
3. Abstract of sufficient length to permit the Committee to assess the paper's suitability for inclusion in the Technical Program. Since the merit of the prospective paper must necessarily be judged by the abstract, it should be clear and informative.

Materia1s should be mailed to R. M. Bowie, Sylvania Electric Products Inc., Box 6, Bayside, L. I., N. Y. The deadline for acceptance of abstracts is November 21, 1949.

Waveforms, edited by Britton Chance, Frederick J. Williams, Vernon Hughes, and Edward F. Mac Nicol, and David Sayre


This book, which is a notable addition to the Radio Laboratory Series, is written to prove the great value to workers in the communication and electronics fields. Based upon wartime developments in this country and in the United Kingdom, it contains a wealth of material not previously available, and is the first book to cover completely the application of nonlinear circuit elements to the generation and shaping of current and voltage pulses and waves. The purpose of the book is to give a comprehensive survey of basic circuit techniques used in the generation and manipulation of voltages and currents by linear and nonlinear circuit elements.

An introductory discussion of operations on wave forms leads to a treatment of the generation of sinusoidal waves, pulses, and special wave forms, such as triangular waves, rectangular waves, exponentials, hyperbolas, and parabolas. Chapter 7: amplitude selection (clipping), comparison, and discrimination, and on time selection include the subjects of switch circuits and multiple coincidence circuits. The subject of amplitude modification of electrical waves by electrical and mechanical signals is followed by a discussion of time modulation and demodulation circuits that is particularly timely. Chapters on frequency multiplication, frequency division, and counting, and on mathematical operations on wave forms should prove of special value in the field of electronic computers. The final chapters cover oscillographic techniques, storage tubes, electrical delay lines, and super sonic delay devices.

The treatment throughout the book is comprehensive and clear, and the authors and editors are to be complimented upon the excellence of style and freedom from errors. Some readers may be handicapped in places, however, by unfamiliar nomenclature. A small amount of duplication of subject matter is inevitable in a book written in so short a time by a large number of authors.

Although the book contains the first adequate analysis of the effects of tube and circuit capacitance upon the triggering of trigger circuits (multivibrators), the authors have not discussed sufficiently the effects of shunt tube and circuit capacitances upon the upper frequency limit and output wave form of other devices, such as choppers and differentiating circuits.

The reviewer has found the reading of this book very profitable, and will undoubtedly deal with it as an invaluable reference source. He enthusiastically recommends it to teachers and research workers in the fields of electrical engineering, physics, and applied mathematics. Since much of the subject matter is fundamental, the book will not become obsolete rapidly, and it should prove to be a bible in its field.

Herbert J. Rich
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Published (1949) by Howard W. Sams and Co., Inc., 2924 E. Washington St., Indianapolis 7, Ind. 166 pages. 124 figures. $5.75.

This pamphlet is intended primarily for the guidance of service technicians in the selection and installation of proper television antennas and accessories. An introduction with a discussion of television receivers, frequency allocations, and television networks is followed by chapters on receiving antenna principles, antenna construction, commercial antennas, antenna installation, and common installation problems.

Tables of Generalized Sine- and Cosine-Integral Functions

Published (1949) by the Harvard University Press, Cambridge, Mass. 1 volume, 600 pages, 7 in. x 9 in. $1.25.

The tables in these two volumes, XVIII and XIX in the Annals of the Harvard University Computation Laboratory, were computed by the Automatic Sequence Controlled Calculator. Carried to six decimal places, the tables are timesavers in the investigation of such problems as self- and mutual impedances, radiation resistance, and distributions of currents in antennas and antenna rays of various types.

Automatic Record Changer Service Manual, Volume Two (1948)

Published (1949) by Howard W. Sams and Co., 2924 E. Washington St., Indianapolis 7, Ind. 432 pages. 8 X 11. $6.75.

This volume covers forty-five models of automatic record changers manufactured in 1938, including the new LP and dual-speed changers, plus wire and tape recorders. Included are change cycle data, information on adjustments, needle landing data, "hints and kinks," complete parts lists, and copious illustrations and diagrams.

Harold A. Zuhl and George G. Hower
Signal Corps Engineering Laboratories
Fort Monmouth, N. J.
Kenneth A. Norton discusses service range and the effects of factors such as antenna height, terrain, reflection, and interference in "Propagation in the FM Broadcast Band" (44 pages). "Electronic Aids to Navigation," by J. A. Pierce, covers only general aspects of the subject.

With ten contributors, each writing on ten different subjects, it is not surprising that there is no uniformity in the volume; yet all the monographs are written on a completely professional level. Some of the authors, such as Eisenstein, McKay, and Hutter, assume a more specialized preparation and interest in the part of the reader than the others, such as Rostenhouse, such that the authenticity of the information given in all of the papers can seldom be questioned. On the other hand, some formulas in at least one paper are incomplete or there are missing terms. In another paper special symbols are used in mathematical expressions with no explanation of their meaning, while a third contains poorly composed or improperly punctuated sentences which require two readings.

Most of the papers constitute the only recently published surveys of these subjects generally available and in practically all of them the references are extremely up-to-date. This work should, in the opinion of this reviewer, win wide recognition and become an important part of the literature of electronics.

GEORGE D. O'NEILL
Sylvania Electric Products Inc.
Bayside, L. I., N. Y.

Fundamentals of Electric Waves, by Hugh Hildreth Stilling

Published (1948) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 240 pages + 5-page index + vii pages. 86 figures. $4.00.

This is the second edition of an excellent book which first appeared in 1912. Intended for use by individuals who have the equivalent of electrical engineering training at the college senior level, it presupposes a knowledge of static and low-frequency electric and magnetic fields and an introduction to matrices through calculus. For its purpose, that of introducing the student to vector analysis and Maxwell's equations, together with simple applications, the book has been well planned, and should serve, particularly for those who are somewhat timid, to take the chill out of that first Legion in waters often suspected of being rather icy. Some students may not need all the pictorial and pedagogic aids Professor Stilling uses so skillfully—for example, in his exposition of gradient, divergence, and curl—but many others will find them most helpful in their first introduction to vector fields.

The first half of the book treats magnetic and electric fields by vector methods, and the emphasis throughout is on imparting sound physical ideas and concepts. Maxwell's equations are subsequently introduced and discussed in discussing a variety of topics, such as wave propagation, reflection, radiation, antennas, waveguides, and cavity resonators. However, in a book of this size and scope, the treatment must necessarily be kept at a fairly elementary level.

The present edition differs from the first in that the miter system of units is used rather than the "Gaussian" system, and that the material used to illustrate the application of Maxwell's equations has been increased. Problems are given with each chapter, and this fact, taken in conjunction with the lucid style of writing employed, makes the book suitable both for self-study and for classroom use.
why "grid-cathode impedances" and the "electron beam of the indicator" were singled out for special mention in connection with thermal noise is unfortunate.

The book omits any mention of a number of important items, such as servo systems for antenna position or range tracking, radar altimeters, and pulse doppler techniques for moving target indication. A number of other important items, such as circuitry for automatic frequency control circuitry for automatic gain control, and waveguide joints are never discussed very inadequately or are merely alluded to with no discussion at all.

Some of the material in the book is hopelessly out of date. For example, the chapter on receivers (page 340) shows "typical" receivers utilizing tube types 6A6T, 6166, 954, and 951! Many who read this book will be exasperated by the frequent references to unpublished reports. The reviewer is puzzled to know the reason for these references; they are generally unavailable to readers and the material in the book is obviously not original, so there is no necessity for disclosing credit for any ingenious devices or methods described.

Radio Wave Propagation, by the Committee on Propagation of the National Defense Research Committee; C. R. Burrows, Chairman, S. S. Atwood, Editor

This book should be of great interest and value to engineers concerned with radar and communication circuits operating at frequencies above 30 Mc. A vast amount of information on radio wave propagation at these frequencies has been compiled in this voluminous publication. There are excellent theoretical treatments of basic phenomena and also many good discussions of practical application problems. More than 130 charts and 30 monographs greatly facilitate the use of the material.

The book is essentially a consolidation of the three volumes of the Summation Technical Report of the Committee on Propagation of the N.D.R.C. The work under the general supervision of this committee during World War II involved the cooperation of many groups in the United States, England, Canada, New Zealand, and Australia. Numerous wartime reports by experts from these groups are assembled in the book, the names of about forty-five of these authors being found in footnotes. Since the book is a compilation of many reports by many authors, there is understandably, a noticeable lack of co-ordination and considerable duplication, although in some cases this permits an insight into the historical development of a subject. However, the greatest improvement for users will probably be the lack of an index.

There are three sections or volumes into which the book is divided. Among the subjects treated in the first section are standard propagation, nonstandard propagation, diffraction, refraction, ducts, siting, and coverage. Meteorological theory and methods of forecasting propagation characteristics by meteorological measurements are presented in the second section, together with such topics as reflection coefficients, absorption, scattering, and echo. The results of many experiments are included. The final section is a general discussion of propagation through the standard atmosphere, involving such subjects as ground reflection, atmospheric refraction, antenna gain, and the calculation of field strength and coverage diagrams. Results of a number of transmission experiments and an extensive bibliography are included in an appendix.

The chairman, editor, and publishers have done a real service by making such a wealth of information on microwave propagation available in a single book.

Radio Laboratory Handbook, by M. G. Scroggie

The purpose of this handbook is to guide a laboratory worker in setting up and properly utilizing a laboratory. Dedicated to the use of both "home experimenters" and "dull professionals," the book comprises a compilation of data concerning laboratory techniques, instruments, and procedures. Some special instruments not commercially available are described, but a large proportion of the material is devoted to a description of commercially available British laboratory instruments.

Since practically all of the laboratory instruments considered in the book are British made, full use of the volume for the purpose intended is somewhat limited in this country. However, it has value as a study of British laboratory practices and instruments; furthermore, a large proportion of the material is devoted to general methods of measurement which do not concern specific instrument. Hence, for that portion all that is required is the ability to interpret such British terms as valve, II. T. accumulator, etc.

Much of the material presented is cautionary, in that pitfalls are pointed out—"what may be neglected," "alternative formulae," and the use of judgment in plotting results. Preferred practices, such as the use of decals and the importance of handiness in instruments, are also stressed. It is interesting to observe how few test instruments are exported from the United States to Great Britain. An occasional General Radio instrument and the Bonton 0-meter seem to be the only ones used.

The book is presented in a clear and readable manner. Such mathematics as is used consists only of working formulas, such as inductance and capacitance formulas, and the like.

Radio Laboratory Handbook, by M. G. Scroggie

The purpose of this handbook is to guide a laboratory worker in setting up and properly utilizing a laboratory. Dedicated to the use of both "home experimenters" and "dull professionals," the book comprises a compilation of data concerning laboratory techniques, instruments, and procedures. Some special instruments not commercially available are described, but a large proportion of the material is devoted to a description of commercially available British laboratory instruments.

Since practically all of the laboratory instruments considered in the book are British made, full use of the volume for the purpose intended is somewhat limited in this country. However, it has value as a study of British laboratory practices and instruments; furthermore, a large proportion of the material is devoted to general methods of measurement which do not concern specific instrument. Hence, for that portion all that is required is the ability to interpret such British terms as valve, II. T. accumulator, etc.

Much of the material presented is cautionary, in that pitfalls are pointed out—"what may be neglected," "alternative formulae," and the use of judgment in plotting results. Preferred practices, such as the use of decals and the importance of handiness in instruments, are also stressed. It is interesting to observe how few test instruments are exported from the United States to Great Britain. An occasional General Radio instrument and the Bonton 0-meter seem to be the only ones used.

The book is presented in a clear and readable manner. Such mathematics as is used consists only of working formulas, such as inductance and capacitance formulas, and the like.
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For example, the chapter on cavity resonators leads one to believe that the calculation of resonances and loss factors of even simple cylindrical shapes commonly used are subject to large errors. This is not necessarily true. Methods are available—with which the author is evidently not acquainted—which allow the calculation of resonances and loss factors to any desired degree of accuracy, for many cylindrical shapes in common use. The results have been consistently checked by tests, even including some glass as a dielectric. Thus, the author's conclusion that "the calculated loss would bear little relation to the measured loss" is obviously due to poor methods of calculation.

Again, in the chapter on velocity-modulated amplifiers, the impression is given that "delaunched" as a result of space-charge effects may not be very important and cannot be calculated with any accuracy. Both of these conclusions are wrong. A wave theory has been developed which does exactly this. Although a reference to this theory was given, the author apparently did not read his own work.

W. C. Hahn
General Electric Co.
Schenectady, N. Y.

Radio Servicing: Theory and Practice, by Abraham Marcus

Mr. Marcus' book is intended for those "who are not beginners in radio nor yet advanced enough to study the subject on an engineering level." These readers will find a clear, nonmathematical discussion of electron tubes and their use as rectifiers, AM detectors, amplifiers, and oscillators. They will also find good explanations of many practical circuits of power supplies, radio receivers, and amplifiers, such as are encountered in commercial equipment. Also included are chapters on electrical and radio theory, components and parts, special tubes, servicing procedures and techniques, and repair and alignment.

Thorough in its scope, the book discusses volume control (manual, AVC, delayed AVC, amplified delayed AVC, and quiet AVC), tuning indicators, tone compensation and control, selectivity control, band-switching and band-spread, noise suppression circuits, push-button tuning, and automatic frequency control.

One fault might be that the book is written mainly from the AM point of view. The comparison of the relative bandwidth requirements of AM and FM could be more adequately discussed.

"Radio Servicing: Theory and Practice" should be of value to those who want not only a textbook, but also a reference book.

Frank R. Arams
RCA Victor Division
Lancaster, Pa.
1949

Television, How It Works


Designed as an elementary textbook in television, this book, written by various staff members of the Rider Co., explains the operation of the television receiver on today's market.

The opening chapter presents an over-all picture of television, followed by a chapter on the television channels and characteristics of the video signal, operating bandwidth characteristics, and the carrier and intermediate frequencies. The third chapter discusses antennas designed for the reception of television signals, and the next eight chapters cover the receiver proper. The final chapter contains practical discussions of text instruments, signal generators, etc., used in the alignment and maintenance of receivers, alignment procedures, and “trouble-shooting” suggestions.

Auto Radio Manual

Published (1949) by Howard W. Sams and Co., Indianapolis, Ind., over 300 pages, profusely illustrated. 8$ X 11. 4$ 95.

This volume gives uniform service data for most of the automobile radios produced since 1946. It covers 100 different postwar models put out by 24 manufacturers.

IRE People

The Government of the Republic of France has awarded William Dubilier (A'14-M'18-F'29), radio inventor and pioneer, the Diploma of Officer of the Academy and the Order of Academic Palms.

Further, at the annual meeting of the Association des Ingenieurs-Docteurs de France, held at the Sorbonne in Paris on June 25 at 1949, the executive committee (consisting solely of professors of the Sorbonne and members of the French Academy of Sciences) unanimously voted to Mr. Dubilier the honorary medal of the organization. The presentation was made on June 25 at the French Embassy in New York City.

Mr. Dubilier is well-known among his fellow engineers as a long-time contributor to the fields of radio telephony; transmitting and receiving condenser theory, practice, and construction; and a successful and effective inventor in many communications and radio equipment. As a result of his development of the power condenser or capacitor that has made modern broadcasting and commercial radio possible.

New York, which became the Cornell-Dubilier Electric Corp.

Among the 300 patents which have been issued to him in the United States is one for his invention of the first portable electric heater. Currently he is vice-president and technical director of the Cornell-Dubilier Electronic Corp., and president of the Radio Patents Corp., both in New York City.

Donald G. Fink (A35-M5-F45-F47), editor of Electronics and newly appointed Chairman of the Joint Technical Advisory Board, was chosen by the Radio Manufacturers Association as technical advisor to the American delegation which attended an international meeting on television standards in Zurich, Switzerland, from July 1 through 5. The delegation was instructed to support American television standards as adopted by the FCC and to try to prevent hasty adoption by the international body of television standards not compatible with the United States system.

Mr. Fink was born in Englewood, N.J., on November 8, 1911. After receiving the bachelor of science degree from the Massachusetts Institute of Technology in 1933, he spent a year in graduate study there before joining the editorial staff of Electronics in 1934. During the war he took a four-year leave of absence in order to work on the development of the Loran system of long-range navigation at the Radiation Laboratory. The Loran transmitters now in use are based on his designs, and he is the author of “Microwave Radar,” the first radar text-book.

Appointed head of the Loran division of the Laboratory in 1943, Mr. Fink served both in the European and Pacific areas, later being transferred to the Washington, D.C., office of the Secretary of War. In 1945 he returned to Electronics as executive editor. He was awarded the War Department's Medal of Freedom in 1946 and the Presidential Certificate of Merit two years later.

Ross K. Gessford (A38-M44), formerly engineering specialist in cathode-ray tubes, has been appointed chief engineer for the television tube division of Sylvania Electric Products Inc.

Born on June 27, 1906, at Niagara Falls, N.Y., Mr. Gessford was educated at George Washington University and the University of Maryland, receiving the B.S.E.E. degree from the latter in 1929. From 1929 to 1936 he was employed by the Westinghouse Electric Corp., where he progressed from student engineer to research worker in design and development engineering. In 1936 and 1937 he worked for the Ken-Rad Tube and Lamp Corp., as a design and development engineer for radio tubes. He joined Sylvania in the latter year and has been continuously associated with the research, development, and engineering of radio and cathode tubes since that time.

Mr. Gessford is a past chairman of the Emporium Section of the IRE.
A. Daniel Collins (S'47), student at the University of Michigan College of Engineering, died early this year. 

Born in Ohio on March 19, 1917, Mr. Collins was educated at the Celina High School in Celina, Ohio. He expected to be graduated from the University of Michigan with the B.S.E.E. degree this year.

consultant for the city of Los Angeles on radio safety ordinances.

From 1942 until 1944 Mr. Lamm was a staff member of the Massachusetts Institute of Technology, conducting research and development on radar receivers and systems at the Radiation Laboratory. In 1944, as director of the MIT field experimental station at the National Bureau of Standards, Mr. Lamm was responsible for the research, development, and engineering of all electronic equipment used in automatic homing missiles, including the "Hat" and the "Pelecan." He also served as consultant to the Navy Department’s Ordnance Bureau on guided missile work being done by the National Bureau of Standards. He joined the NBS staff in 1947.

In recognition of his contributions to the war effort, Mr. Lamm has been awarded the Presidential Certificate of Merit, Certificates from the Office of Scientific Research and Development, the U.S. Navy Bureau of Ordnance Merit Award, and the Award of Commerce Meritorious Service Award. He is a member of the Electronics Club.

Directing the activities of the committee secretariat will be the task of Fred A. Darwin (S'46), newly appointed executive director of the Committee on Guided Missiles of the Research and Development Board, National Bureau of Standards.

Mr. Darwin was born in Chattanooga, Tenn., on May 28, 1913. He attended the University of Chattanooga from 1929 to 1931 and the U.S. Naval Academy at Annapolis, Md., from 1931 to 1935, receiving the bachelor of science degree in the latter year. In 1946 he was granted the M.S. degree in electrical communications engineering by Harvard University, and the following year he entered the employment of the Western Union Telegraph Company as an apprentice engineer. When he left in 1941, he had risen to the rank of senior engineering supervisor.

Upon the outbreak of war, Mr. Darwin joined the service as a naval reserve command, heading the IFF (radar identification) and radar beacon design section in the radio division of the Navy Department's Bureau of Ships. In 1946 he joined the Hazelton Electronics Corp. at Little Neck, L.I., N.Y.

New associate director of Tele-Tech, television and communications engineering magazine, is John H. Bottsman (S'47-M'49). A prolific writer of scientific articles for leading publications, Mr. Bottsman resigned from the staff of the E. I. du Pont de Nemours Co., where he had been employed as assistant chief allocations engineer since July, 1947, in order to take up his new post.

Mr. Bottsman was born in England on August 11, 1913, and educated at the University of London, from which he received the bachelor of science degree in radio communication in 1936. Meanwhile, in 1941, he had entered the employment of the E. I. du Pont de Nemours Co., where he had been employed as assistant chief allocations engineer since July, 1947, in order to take up his new post.

Victor B. Corey (A'45) has been appointed manager of the engineering physics division of Fredric Flasher, Inc. Associated with the company since 1946, Mr. Corey has supervised research and development on sonic true airspeed, true air temperature, and machine number indicators; radiation physics; servomechanisms for vibration and control; analogue computers; and long-range automatic navigation.

Born in Missouri on February 9, 1915, Dr. Corey received the B.A. degree from Central College in Fayette, Mo., in 1937. Until 1941 he served as a teaching assistant in the physics department of the State University of Iowa; and was a research assistant at the University for a year following. He received the M.A. degree from the University of Iowa in 1939; the Ph.D. in 1942. From 1942 until 1946 he was research physicist for the Sylvania Electric Products Inc., in Framingham, Mass.
The Program for New Aids to Air Navigation*

D. W. RENTZEL†

Air navigation requires adequate communication and desirable and comprehensive guidance means. A summary of present-day practices and a schedule of future plans in this field are presented in some detail by the Administrator of Civil Aeronautics of the United States Government. Engineers working, or planning to work in this field, will be well advised to study the contents of this address.—The Editor.

† Civil Aeronautics Administration, Washington, D. C.

Air operations using the same system during instrument weather conditions. Speed is the essence of modern warfare; in case of sudden attack we must be able to move large numbers of military aircraft quickly and unerringly to the points where they are needed. The enemy will not wait for favorable weather, or give us time to acquaint our pilots with unfamiliar devices.

Fortunately, all significant groups connected with civil and military aviation have agreed on a definite program to modernize our airways and make all-weather flying a universal reality. This program was developed through the Radio Technical Commission for Aeronautics, and the plan itself is commonly referred to in aviation circles as "SC-31," because it was prepared by Special Committee 31 of the RTCA.

The first, or transition, phase of this revolutionary new air navigation program will be completed about 1951. A good start already has been made in developing and installing the new devices needed for this part of the program. The ultimate program, which envisions some devices which a highly imaginative Buck Rogers might envy, is scheduled for completion about 1963.

OMNI RANGE AND DME

Now let us look at some of the old and the new air navigation equipment. Earlier, I mentioned the four-course low-frequency range. This range offers, as the name implies, only four paths to or from the range. In order to stay on one of these courses, the pilot must listen continuously to dots and dashes which blend together when he is in the exact center of the airway. Needless to say, this is exacting, and during thunderstorms and periods of heavy static, the range becomes difficult and even impossible to hear. There is danger, too, of the pilot confusing the courses and flying on a wrong heading.

To replace this kind of range, the CAA has been installing what is known as omni-directional, or omnaranges. These offer the pilot an almost unlimited number of courses which he may fly. And the omnaranges, operating in the very high-frequency part of the radio spectrum, are largely free of static and interference.

Best of all, with the omnirange, the pilot can fly by eye instead of ear. An occasional glance at a vertical needle in his cockpit is all the pilot needs to keep him on the right heading. About 250 of these omnaranges are now operating in the United States, and the CAA program calls for an eventual total of about 400, blanketing most of the country with signals.

The omnirange gives the pilot simple, clear information about the course he is flying. If he is flying northeast, for example, on a course of 45 degrees, the numerals zero four five will be continuously visible. And the words "to" or "from" will tell him clearly whether he is on a course to or from the station. This course indication is entirely independent of the aircraft compass, and shows the track actually being flown, regardless of cross winds and the plane's heading.

The difference between the omnirange course and the indicated magnetic heading continuously shows the pilot the amount of correction necessary for cross winds. But the pilot need not concern himself with this unless he wishes; if he flies by the vertical needle his wind correction is automatic.

DME

Each omnirange eventually will be equipped with a device called "distance-measuring equipment," or DME. With suitable equipment in the aircraft, the pilot always will know his exact distance to the omnirange. This information will be displayed in the cockpit by a simple pointer on a dial. With the omnirange and the DME combined, the pilot continuously will know his exact position in space, without having to work out navigational problems.

In addition to all this, an electronic brain called a course-line computer has been developed. This device solves difficult navigation problems with the speed of light. Using this computer, a pilot will not need to fly directly to or from an omnirange. He can set a course from one selected point to another, and then let the computer, which uses signals from near-by omniranges, guide him accurately to his destination.

These new devices, all of which will come into general use in the next few years, will make possible multiple airways between cities, relieving the traffic congestion which already has passed the saturation point in many parts of the country.

Very-high-frequency voice radio, which is static-free, is coming into general use along the airways. It is making a definite contribution to safer flying. For the ultimate pro-
gram, however, a private-line system will be developed for instantaneous automatic transmission of information between ground and air.

So far, we have discussed the new equipment which will guide aircraft along their routes. Equally important, however, is the problem of getting them safely into the air, and onto the ground, during low visibility. For all-weather flying, this is just as important as safe and reliable navigation on route.

**ILS and GCA**

We have available today two entirely different methods of bringing aircraft safely to a landing through low ceilings. One, mentioned earlier, is called the "instrument landing system," and uses radio beams. The second, using precision radar principles discovered during the war, is called "ground-controlled approach (GCA)." Each system has advantages, and each system has drawbacks. Each can be used separately. But when used together, as recommended under the RTCA program, they provide the pilot with a double check on his position at all times, and achieves the closest to ultimate safety which our present knowledge permits.

With the "Instrument landing system (ILS)" two radio beams are transmitted from the airfield. Received aboard the aircraft, these beams operate a cross-pointer indicator, which is simply a dial with two needles crossing in the center. The vertical needle, which also is used with the omirange, tells the pilot whether he is properly lined up with the center of the runway and, if not, which way he must turn. The horizontal needle tells him whether he is above or below his proper glide path, and how to correct his descent, if necessary.

The ILS system is simple, positive, and in wide use by our scheduled airlines. Already, it has permitted the CAA to lower landing minimums from 400-foot ceilings to as low as 200 feet in many locations, greatly improving schedule reliability. Similar reductions in ceiling minimums have been approved where radar systems are in use.

The radar landing system, called "ground-controlled approach (GCA)," permits a controller on the ground to "talk the pilot down" over ordinary voice radio channel. The ground controller watches two radar screens.

The first, known as the surveillance radar screen, enables the operator to locate aircraft flying within a 40-mile radius of the airport. After positively identifying the aircraft on approach as a particular dot on the screen, the controller guides him safely into and through the holding pattern.

When the plane is ready to head in for a landing, a precision radar screen comes into play. The correct path to the runway is shown by lines on the screen, and if the dot representing the plane gets off the lines, the controller tells the pilot exactly how to correct his course.

This ground-controlled approach radar may be used independently, or to monitor an approach made on the instrument landing system.

At present there are about 80 civilian instrument landing systems in operation. We have improved-type surveillance and precision radar equipment for ground-controlled approaches at LaGuardia Field in New York, at Washington National Airport, and at Chicago. As rapidly as funds and manufacturers' delivery schedules permit, we are installing additional GCA radar sets at the busiest airports.

**RADAR**

At other large airports CAA is planning to install the surveillance radar unit alone. This will permit the traffic controller to watch all the aircraft in his vicinity through radar, even when the weather has closed in. The controller can be certain that each plane is in its reported position, thus reducing collision hazards and specifying the landing and takeoff sequences at the airport.

There has been some misunderstanding by the public of the whole subject of radar. Many people believed that war-developed radar would, in some magic way, instantly transform aviation into an all-weather transportation system, free of hazards and navigation problems. Ultimately, it promises to do just that. But we still have quite a way to go.

For one thing, military ground radar equipment designed for use on the fighting fronts proved to be too efficient and unsatisfactory for everyday civilian use. An extensive program was necessary to design, test, and produce ground radar which is economical and equally useful for civilian and military aircraft.

Airborne radar, as produced during the war, was a heavy item of equipment. Also, it required one or more men to operate it, in addition to other members of the crew.

Overseas, where there were no other navigation aids, it was a necessary piece of military equipment, well worth the extra weight and manpower.

But in a country like the United States, with adequate navigation aids, airborne radar of the wartime type cannot justify itself in commercial operation. A pilot can get far more navigational information from radio ranges, and use it more easily, than from radar equipment in his plane.

However, airborne radar does show promise in two special fields. Numerous experiments have indicated that a satisfactory light-weight radar can be produced which will help pilots to detect and fly around thunderstorms and other turbulent areas. Eventually, also, someone may develop a satisfactory radar collision warning device.

New applications of radar and television really will come into their own in the ultimate RTCA program, which will provide an air traffic system of almost inconceivable magnitude and precision. Some of the equipment needed has not yet been invented, but the specifications have been laid down, and the principles of operation are understood. No one doubts the ability of American electronic engineers to produce the needed air and ground devices.

By 1963

Here, in a general way, is how this ultimate air navigation system will work:

Even before a pilot takes off on a flight, a landing time will be reserved for him at his airport of destination. As he flies along, a dial will tell him in minutes and seconds whether he is ahead or behind his exact schedule, and he will slow down or speed up accordingly.

In the cockpit the pilot will see a pictorial presentation of everything around him. This picture, probably televisual from the ground, will show his own aircraft in relation to others in his vicinity, indicate obstructions or other hazards, and even show the location of storms and turbulent air.

At the same time, radar will be continuously watching him from the ground. By means of a block system something like that used on railroads, the pilot will be assured that he is in safe air space at all times.

The automatic pilot, which will carry equipment which continuously transmits to the ground the readings of the various cockpit instruments. Electronic brains on the ground will check these readings automatically against information derived from radar and other sources. If, for example, the altitude shown by ground radar differs from altitude reading in the cockpit, the pilot will be instantly and automatically notified.

If the pilot wishes to change his altitude or his flight plan, he will be able to query the ground stations by pushing an appropriate button. Approval or disapproval will be flashed back to his cockpit in a fraction of a second, since it will be made by automatic machines on the ground.

This ultimate system, fantastic though it may sound, is designed to meet the everyday needs of civil and military aviation 15 years hence. It will, of course, solve the weather problems which plague aviation today, and it will permit aircraft to fly their schedules with clocklike precision and absolute reliability.

Furthermore, the RTCA system is designed with military as well as civilian requirements in mind. In case of war, the system will give instant warning of unfriendly aircraft, and permit interceptors to be vectored to attack. It will permit quick and heavy concentration of airpower anywhere it is needed within the country, and then will assist in maintaining a continuous flow of supplies and manpower to the area.

This tremendous new program, on which the Army, Navy, Air Force, and CAA are jointly agreed, will open the way for a whole new era of aviation in which the blessings of fast, safe, reliable low-cost transportation will be shared by every American citizen.
Summary — The frequency dependence of the direction from which reflected television signals reach the receiver complicates the antenna design necessary to reduce reception of these reflected signals. A formula is derived for the strength of these reflections, and an analysis is made of some reflecting areas in order to find a reason for this frequency dependence. An antenna arrangement that appears particularly useful for reflection reduction is described.

ONE OF THE most difficult types of television interference to eliminate is that causing multiple images, or "ghosts," in the picture. Although in a great number of locations the transmitting stations are all in the same direction from the receiver, the reflections causing displaced images do not generally come from the same directions for all the stations. This has resulted in the widespread use of double antennas, one for the high and one for the low television bands, which can be individually oriented.

Let us consider a building or other object located at point A in Fig. 1 as causing a reflection such that signals from the transmitter T reach the receiver R by path $d_2 - d_3$ as well as by path $d_1$. For an equivalent isotropic transmitter power $P_t$ in the directions of interest the power per unit area incident on $A$ is $P_t/4\pi d_2^2$. Let $A$ be the equivalent isotropic reflecting area defined for incident and reflecting directions $T$, $A$ and $A$, $R$, which differs from the area much used in radar calculations in which direction $T$, $A$ is coincident with $-(A, R)$, while in this case of television reflection, no such relation exists between the two directions. The reflected energy incident on $R$ will then be

$$P_t \frac{A}{4\pi d_2^2 \frac{A}{4\pi d_3^2}}$$

The energy incident on $R$ per unit area due to direct transmission from $T$ will be $P_t/4\pi d_1^2$. The square root of the ratio of these two energy intensities will be the ratio of "ghost" field strength to signal field strength or

$$\frac{E_o}{E_s} = \frac{1}{\frac{A}{4\pi}} d_1/d_2.$$  (1)

The reflected signal will be relatively strongest when either $d_2$ or $d_3$ is a minimum and the other is nearly equal to $d_1$, which means that the source of reflection is near either the transmitter or the receiver. Due to the symmetry of (1) in $d_2$ and $d_3$, it makes no difference which the reflector is near. The formula can then be written

$$\frac{E_o}{E_s} = \frac{1}{\frac{A}{4\pi}} \frac{1/2d_1}{2d}$$  (2)

where $d$ is the shorter distance to either $T$ or $R$ when $A$ is near one of them.

Reflections can then be divided into two categories: (1) those originating near the transmitter, all of which come from nearly the same direction as the desired signal, and (2) those that originate near the receiver, which can come from any direction. Reflections of the former type are obviously the harder to discriminate against by means of directional receiving antennas.

Since reflections do not all come from the same directions at different frequencies even though the transmitters are all located in the same direction from the receiver, if these reflections are originating in the vicinity of the receiving antenna, this must mean that the reflectors are relatively frequency sensitive. If the reflections originate in the vicinity of the transmitters no such conclusion can be drawn, since the angles at which incident waves from the various transmitters strike a given reflector would be different unless the transmitting antennas were extremely near each other relative to the distance from them to the source of reflection, which is not generally true of transmitting antennas located on tall buildings in a large city.

To see just how frequency sensitive various reflectors are, let us first consider a plane reflecting area large compared to a wavelength and oriented so as to produce the maximum reflection toward the receiver as shown in Fig. 2. The total energy incident on the plane will be equal to $S \cos a_1 P_t/4\pi d_2^2$ where $S$ is the physical area of the plane and $a_1$ the angle of incidence. The plane will then act as a uniformly excited plane array and will re-reflect with a beam width inversely proportional to frequency. Its isotropic gain, considering it as an antenna of area $S$ will be $4\pi S/\lambda^2$ where $\lambda$ is the wave-
length. Its equivalent isotropic reflecting area will then be

$$A = \frac{4\pi S^2 \cos a_1}{\lambda^2}$$  \hspace{1cm} (3)

when the angle of incidence equals the angle of reflection. The ratio of carrier frequencies of the channel 13 and channel 2 television stations is 211.25 Mc/55.25 Mc or 3.82. The relative signal-“ghost” field strengths in the two channels for large flat reflecting surfaces would then be different by the same factor or 3.82. For reflectors several wavelengths across the angle of incidence equals the angle of reflection regardless of frequency. Therefore, if located near the receiver, such a plane should cause “ghosts” in all channels with the relative “ghost”-signal field strength changing by a factor of 3.82 as frequency is shifted from channel 2 to channel 13. If located near the transmitters, it would be coincidental if the angles of incidence and reflection were constant for more than one channel.

Let us now consider the echoing area of a sphere when the incident and reflected wave directions do not coincide. The laws of geometrical optics\(^3\) apply if the sphere is large in wavelengths. Let us assume that a square cylinder of incident rays bounded by the lines 1, 2, 3, 4 of Fig. 3 impinges on the sphere. Assume that the rays are all parallel to the XY plane and that rays 1 and 2 lie in the XY plane. If incident ray 1 makes an angle of \(a_1\) with the normal to the surface, its angle of reflection will also equal \(a_1\). The angles of incidence and reflection of ray 2 are greater than those of ray 1 by \(\Delta a\), so the two rays leave the surface no longer parallel but diverging at an angle \(2\Delta a\).

Rays 3 and 4 are deflected both upward and to the left upon reflection. The angle \(\Delta B\) that ray 4 makes with the XY plane after reflection is given by the formula from trigonometry

$$\sin \Delta B = 2 \tan \Delta C \cos (a_1 + \Delta a),$$

again assuming the angle of incidence equals the angle of reflection. This formula as \(\Delta a, \Delta B, \Delta C\) approach zero approaches

$$B = 2\Delta C \cos a_1.$$  \hspace{1cm} (4)

We see then that all of the energy incident in the square cylinder is reflected in a rectangular beam 2\(\Delta a\) wide and 2\(\Delta C\) \(\cos a_1\) deep. The area of the incident beam is \(R^2\Delta C\Delta a \cos a_1\) where \(R\) is the radius of the sphere. If the beam has an intensity of \(I\) watts per unit area, the incident power is \(IR^2\Delta C\Delta a \cos a_1\) watts. Dividing this power by the solid angle of the reflected beam we get \(IR^2/4\) watts per steveradian. If the sphere of radius \(R\) has an isotropic reflecting area of \(A\), the power reflected by it in the direction of interest will be the same as if the total incident power \(IA\) were reflected uniformly over all 4\(\pi\) steradians. The signal intensity will then be

$$IA/4\pi = IR^2/4\text{ watts per steveradian.}$$

The echoing area of the sphere will then be, solving for \(A\),

$$A = \pi R^2$$  \hspace{1cm} (4)

as is well known to be true when the directions of incidence and reflection coincide as in radar operation. Equation (4) shows that the area is independent of both frequency and the directions of incidence and reflection. Displaced images, then, caused by reflections from a spherical shaped object, should appear on all television channels with equal intensity regardless of where the object is located, provided the television transmitters are close together, compared to the distance between them and the reflector.

A similar analysis to the above for vertical cylinders, such as smoke stacks, where the radiation is always normal to the cylinder axis shows the echoing area of

\(^3\) L. J. Chu, “Microwave Beam-Shaping Antennas,” Research Laboratory of Electronics, MIT, Report 40; June, 1947.
such a cylinder to be

\[ A = \frac{2\pi Rh^2 \cos a}{\lambda} \]  

(5)

where \( h \) is the cylinder height, \( R \) its radius, and \( a \) the angle of incidence with respect to a normal to the surface. This type of echoing area, like that of a flat plane, is dependent on the included angle between directions from the reflector to the transmitter and to the receiver, but is not dependent on a particular reflector orientation as is the plane reflector, and does not vary as rapidly in magnitude with frequency as does the equivalent isotropic area of a plane. The "ghost"-signal field strength ratio due to reflection from such a cylinder would then change from channel 2 to channel 13 by a factor of only \((3.82)^{1/2}\) or 1.96.

In considering smaller reflectors we might see how close an object must be for the receiver to give a visible displaced image. Assuming a "ghost"-signal field strength ratio visibility threshold for the television receiver of approximately 0.01 and an area equivalent to that of a sphere of 1.96 foot radius which, because of resonances occurring with such a small sphere, is 16.3 square feet in both channels 2 and 13, the distance to the reflector would have to be

\[ d = (\frac{A}{\pi})^{1/2}E_s/2E_g = 114 \text{ feet}, \]

which is so small that a displaced image caused by it would barely be visible as a "ghost." For objects smaller than this 1.96-foot sphere, the Raleigh sixth power law soon governs, and reflections quickly become negligible with decreasing reflector size, even though they are much more frequency-dependent than larger sized reflectors.

The corner reflectors so effective in radar work would, of course, be of little interest here because of the sparsity of structures simulating them, and because they would cause "ghosts" only if placed on a line passing through the receiver and transmitter, and not located on the portion of the line between the two.

**Separately Orientable Array**

One antenna type that can be used for countering the difficulty presented by the frequency dependence of the direction from which reflected television signals approach the receiver is one made up of several separately orientable arrays; one for each station in a particular locality. The various array outputs are fed into a filter unit which selects the desired array for a particular channel.

The antenna of Fig. 4 is made up of 7 four-element parasitic arrays. As the maximum number of stations presently to be located in any one service area is seven, and since in no area will two stations occupy adjacent channels, the arrays of Fig. 5 each cover two adjacent channels, except for the low-frequency array for each band which cover only one channel each. Fig. 5 shows a picture of the filter box used with the array. The box is designed to fit around the antenna mast so as to minimize the necessary lengths, for economic reasons, of the seven 92-ohm coaxial cables leading into the box. A single 300-ohm twin lead carries energy from the box to the receiver. The box contains seven doubly tuned filters, one of which is shown removed in the figure. The circuit of the filter unit is shown in Fig. 6. The filter unit serves to pass energy from a particular array to the main transmission line only at the desired frequencies.
of operation of the array, and to transform the 92-ohm impedance of each array up to the 300 ohms of the main transmission line.

![Diagram of a filter unit](image)

**Fig. 6**—Circuit of a filter unit.

Fig. 7 shows radiation patterns of the complete antenna with filter box when the arrays corresponding to the two filters in operation are oriented 90 degrees apart. These patterns were taken for two adjacent channels and at a frequency midway between the channels. The gain of the antenna is the 8 db of each array, minus the losses in the filter box at any particular frequency.

The coaxial cable feeding the folded dipole of each array passes through a hole in the center portion of the grounded part of the folded dipole and is run inside the tube of the dipole, so that the braid can be fastened to one of the dipole terminals and the central conductor of the cable to the other dipole terminal without unbalancing the normally balanced folded dipole. The seven arrays of the antenna show negligible mutual effects when two adjacent arrays are separated by a quarter wavelength at the frequency of the longer array.

![Diagram of reception patterns](image)

**Fig. 7**—Reception patterns of a complete antenna.

**Conclusions**

If the frequency dependence of reflected signal directions were reduced, the television receiving antenna problem would be considerably simplified. One possible solution to the problem would be to place all television transmitting antennas of a given service area on the same structure. Other possible solutions would be, of course, the development of improved receiving antennas, or improved "ghost" suppressors, such as of the delay line type, or the adoption of circularly polarized transmitting and receiving antennas.

**Regenerative Amplifiers**

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*Summary*—This paper describes the principles and applications of regenerative amplifiers, which may be used, for example, to mark the instant when two voltages become equal. A peak voltmeter circuit based upon the switching properties of a regenerative amplifier is introduced to minimize the error encountered in measuring low duty-cycle pulses. The use of a regenerative amplifier in forming a pulse-width discriminator circuit is also described.

**INTRODUCTION**

By employing a large amount of regeneration, the output voltage of an amplifier can be made to change abruptly from one constant value to another when the input voltage is raised to a critical value.

* Decimal classification: R363.23. Original manuscript received by the Institute July 19, 1948; revised manuscript received, November 8, 1948.
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This circuit can be adjusted to trigger on or off for the difference trigger on with the 1949 Regenerative Amplifiers because positive feedback is introduced when the input voltage is increasing or decreasing, respectively. Because positive feedback is introduced by the cathode resistor \(R_1\), plate current can flow in only one tube at a time, and can be caused to transfer abruptly from one tube to the other by varying the input voltage. When the input voltage is zero or small, \(T_1\) is cut off, and \(T_2\) conducts. When the input voltage reaches a certain value, say \(E_{in}\), the plate current of \(T_1\) starts to flow, a switching process takes place and the plate current transfers from \(T_2\) to \(T_1\). Because the voltage across \(C\) cannot change instantly, the voltage on the screen of \(T_1\) remains constant during the switching process. After the switching process, \(T_1\) conducts, capacitor \(C\) discharges, and the screen voltage of \(T_1\) reduces. The amount of reduction is governed by the magnitude of voltage \(E_1\), because at the instant the grid of \(T_4\) reaches \(E_1\), diode \(T_3\) conducts and further discharge of \(C\) is thereby stopped. This reduction lowers the operating path of \(T_1\), i.e., causes \(T_1\) to operate along a lower transfer characteristic. Thus the second switching process can be arranged to occur as soon as input voltage is reduced to \(E_1\) again (or even above \(E_1\)), by adjusting the magnitude of \(E_1\). After the second switching process, \(T_1\) is cut off and \(C\) charges back to its original value. The circuit is then ready to repeat the cycle when the input voltage reaches the same critical value. The choice of time constant \(R_4C\) is governed by the characteristics of the input voltage, such as repetition frequency, rise, and decay times.

A somewhat different form of regenerative amplifier is shown in Fig. 3. This circuit generates a sharp positive pulse as soon as the unknown voltage equals the reference voltage. Either thermionic diodes or germanium crystals can be used for \(T_1\) of this circuit, or for \(T_1\) and \(T_2\) of the circuit of Fig. 4. When the circuit is in its quiescent condition, the unknown voltage is higher than the reference voltage, and diode \(T_1\) conducts. Thus the impedance between the grid of \(T_2\) and ground is very small (approximately equal to the plate resistance of \(T_1\) plus the internal impedance of the reference voltage source). The amount of regeneration is therefore very small and no oscillation can be expected. As soon as \(T_1\) is cut off, the amount of regeneration greatly increases, and the grid voltage of \(T_2\) reduces. Thus a switching process similar to that of other regenerative amplifiers
takes place, and ends with $T_2$ cut off and $T_3$ carrying maximum current. Since the grid-leak resistor $R_1$ is returned to 250 volts, the grid voltage of $T_3$ increases exponentially toward 250 volts as $C_1$ discharges. When the cutoff point is reached, the plate current of $T_2$ begins to flow, a second switching process occurs, and the circuit returns to its normal condition. The sharpness of output pulse depends on the time constant $R_1C_1$.

The circuit of Fig. 4 generates a sharp negative pulse as soon as the unknown voltage equals the reference voltage. At this instant, diode $T_1$ conducts, and the amount of regeneration is hereby greatly increased. Therefore, cumulative processes will take place. The purpose of diode $T_1$ is to prevent $C_3$ from discharging through the unknown voltage source. The coupling transformer should provide a phase shift of 180 degrees. If the duration of the pulse is to be determined by the time constant $R_1C_1$, the mutual inductance of the coupling transformer must be large enough to support the pulse without material decay.

**Peak Voltmeter Circuits**

Regenerative amplifiers are well adapted to the measurement of the peak amplitude of voltage pulses. A peak voltmeter circuit, based upon the switching properties of a regenerative amplifier, is shown in block form in Fig. 5. Before the application of an input pulse, triode $T$ is cut off, and the potential across $C$ is zero. The regenerative amplifier is triggered by the leading edge of the input pulse, and produces a positive voltage to drive the grid voltage of $T$ to zero, or slightly positive. Then the memory capacitor $C$ charges. The voltage across $C$ is coupled through the cathode-follower to the indicating device which consists of a dc milliammeter and a resistor in series. The input resistance of the cathode-follower is made very high in order to prolong the hold-on time of the memory capacitor $C$. The voltage on the indicating device is fed back negatively to the input terminal of the regenerative amplifier for the purpose of reducing the potential developed by the input pulse. Therefore the regenerative amplifier will trigger off as soon as the voltage on the indicating device equals the amplitude of the input pulse. This in turn cuts off the plate current of $T$ and blocks $C$ from further charging. Once triggered off, the regenerative amplifier remains so until the input voltage rises above the voltage on the indicating meter.

The principal advantages of peak voltmeters of this type are: (1) Errors encountered in measuring pulses of low repetition frequency and short duration are greatly reduced. This results because the use of the B-supply instead of the measured pulse or the output of an amplifier to charge the memory capacitor greatly shortens the charging time, and makes it possible to prolong the hold-on time by increasing the capacity of the memory capacitor. Furthermore, this meter employs a regenerative amplifier to trigger the charging current, thus permitting the final voltage of the memory capacitor to reach a value equal to the peak amplitude of the input pulse. (2) The meter reading is practically independent of circuit constants and operating voltages because the instrument is based on a comparison principle.

The circuit diagram of a practical instrument of this type is shown in Fig. 6. $T_1$, $T_2$, and $T_3$ are connected as a regenerative amplifier similar to that of Fig. 1. The function of $T_4$ is identical to that of triode $T$ of Fig. 5. The negative bias voltages of $T_1$ to $T_3$ are all supplied by gas diode $T_5$. $C$ functions as the memory capacitor. $T_5$ serves as a cathode-follower with output meter $M$ and $R_4$ in series as its load. To minimize the leakage due to the grid current of $T_3$, resistor $R_2$ should be so dimensioned that $T_5$ operates just at its "floating-grid" potential, at which the positive grid current becomes equal to the negative (positive-ion) grid current. Switch $S$ is used for reset, and potentiometer $R_4$ is the zero panel control. Before applying a signal to the input, potentiometer $R_1$ is adjusted so that $T_1$ operates just at the critical point of cutoff, a milliammeter may be connected in series with the cathode of $T_1$ as an indicator for this adjustment. The regenerative amplifier is triggered by the application of a positive input pulse, and then $T_4$ conducts because the plate current of $T_4$ suddenly falls to zero. This in turn causes memory capacitor $C$ to charge
and develops a negative voltage across meter $M$ and $R_s$. While this instrument is designed for measuring positive pulses, in conjunction with a phase inverter, it may also be used to measure negative pulses.

**Pulse-width Discriminator**

The name applied to circuits whose output is proportional to the width of the input pulse is pulse-width discriminator. A regenerative amplifier may be used to augment the special circuit to be described to form a pulse-width discriminator. In Fig. 7, the input to the grids of $T_1$ and $T_2$ is taken from the plate of the normally off tube of the regenerative amplifier, which is adjusted to trigger on and off when the input voltage reaches levels $a$ and $b$ respectively. Thus, the regenerative amplifier acts as a pulse-slicer. When the circuit is in its quiescent condition, the grid-to-cathode potential of both $T_1$ and $T_2$ is zero, and $T_b$ (a sharp cutoff triode) is operating just at the critical point of cutoff. $T_1$ and $T_2$ are chosen to be identical in characteristics, and $R_s$ is made equal to $R_4$. Thus $e_{t_1}$ and $e_{t_2}$, the potentials across capacitors $C_1$ and $C_2$ respectively, are equal to $E_0$ (see Fig. 8). The application of an input pulse drives the grid of $T_1$ at $t_1$ and $T_2$ far below cutoff. Therefore $C_1$ (with initial voltage $E_0$) charges, $e_{t_1}$ rises exponentially toward a final value of $E_{bb}$ with a time constant of $R_4C_1$. We have

$$E_{11} = (E_{bb} - E_0)[1 - e^{-t/(R_4C_1)}] + E_0.$$  

$C_2$ (with initial voltage $E_0$) also charges, but $e_{t_2}$ rises with a different time constant, namely, $(R_2+R_4)C_2$, assuming the plate resistance of diode $T_2$ equal to zero. We can write

$$E_{2} = (E_{bb} - E_0)[1 - e^{-t/(R_2+R_4)C_2}].$$

After the pulse, $T_1$ and $T_2$ again conduct. $C_1$ (with initial voltage $E_{11}$) discharges, $e_{t_1}$ falls exponentially toward a final value of $E_0$ with a time constant equal to $C_1r_{p_2}R_4/(r_{p_2} + R_4)$, where $r_{p_2}$ is the plate resistance of $T_2$. But $C_2$ cannot discharge until $e_{t_1}$ reaches the value $E_{12}$, at which the potential across $C_1$ and $C_2$ are equal. Since $t_1 - t_2$ is the amount of time required for $C_1$ to discharge from $E_{11}$ to $(E_0 + E_2)$, we can find

$$t_2 - t_1 = \frac{C_1R_2r_{p_2}}{R_4 + r_{p_2}} \left[ \ln \left( 1 - e^{-t_1/(R_4C_1)} \right) - \ln \left( 1 - e^{-t_2/(R_2+R_4)C_2} \right) \right].$$

After time $t_2$, both $C_1$ and $C_2$ discharge through triode $T_2$. If the plate resistance of diode $T_2$ is very small, the potentials across both capacitors are approximately equal and decrease exponentially toward $E_0$ with a time constant equal to $(C_1 + C_2)r_{p_2}R_4/(r_{p_2} + R_4)$. During the pulse, from $t_2$ to $t_1$, triode $T_5$ conducts and capacitor $C_3$ charges because the grid of $T_5$ rises with $e_{t_2}$. After the time $t_2$, triode $T_6$ is cut off and capacitor $C_3$ does not discharge. During the next cycle, $T_6$ will not conduct unless the width of the input pulse increases or the voltage across $C_3$ drops below that at the end of the previous cycle. The time between $t_1$ and $t_2$ serves to insure that the output voltage increases by an amount equal to $E_2$. Therefore $(t_2 - t_o)$ must be long enough to allow $C_3$ to charge up to a voltage equal to $E_2$. When the circuit of Fig. 7 is required to follow a series of pulsed signals of decreasing width, a proper resistance must be used to shunt capacitor $C_3$. In general the time constant of $C_3$ and its shunting resistance should be small enough to allow the output voltage to follow the decreasing of the input pulse width, and at the same time must be large enough to permit $C_4$ to hold most of its charge between pulses.

**Acknowledgment**

The author wishes to thank H. S. Dixon for his helpful suggestions in the preparation of this paper.

**Bibliography**

Design of Dissipative Band-Pass Filters Producing Desired Exact Amplitude-Frequency Characteristics*

MILTON DISHAL†, SENIOR MEMBER, IRE

Summary—The purpose of this paper is to present a basic method of obtaining the exact required values for all circuit constants in a band-pass network using \( n \) finite-\( Q \) resonant circuits to obtain either of two types of exact amplitude responses; the so-called critical shape-coupled, Butterworth, or transitional type of response, and the so-called over-coupled or Chebyshev\(^1\) type of response.

The equation giving the gain obtained with the desired response shape is derived. Equations for the exact phase characteristics associated with the above exact amplitude characteristics are also given.

Some comments are made concerning a somewhat similar method of design, which makes use of the so-called "poles" of the network.

Design sheets are presented giving the necessary equations for single-, double-, triple-, and stagger-tuned networks to produce either of the above two amplitude-response shapes.

1. Symbols

\( n \) = total number of resonant circuits in networks of Figs. 3 and 4.
\( N \) = total number of cascaded networks between which there is no coupling, e.g., separated by vacuum tubes.
\( I \) = complex voltage output at any frequency.
\( I' \) = magnitude of the voltage output at any frequency. (See Fig. 7.)
\( I'_n \) = magnitude of the voltage output at the frequency of peak response. (See Fig. 7.)
\( I'_p \) = magnitude of the voltage output at that frequency on the response curve that the designer defines as the edge of the pass band. For response-shape \( C \), this voltage output is identical with the response at the valleys of the response inside the pass band. (See Fig. 7.)

\( \Delta f \) = frequency bandwidth between response points whose voltage output is \( I' \). (See Fig. 7.)

\( \Delta f_p \) = frequency bandwidth between the peaks of response-shape \( C \) at the voltage output of \( I'_p \). (See Fig. 7.)

\( \Delta f_p \) = frequency bandwidth between the response points whose voltage output is \( I'_p \), i.e., the frequency bandwidth between the edges of the defined pass-band width. (See Fig. 7.)

\( f \) = resonant frequency of each resonant circuit. See Section 3.2. \( f \) is also the geometric mean frequency between two frequencies \( f_1 \) and \( f_2 \) having the same voltage response.

\( f_0 \equiv \) resonant radian frequency = \( 2\pi f_0 \).

\( F = \) total percentage bandwidth between two frequencies \( f \equiv (f_0 - f_0/f)/f_0 \), where \( f = (f_0 f_2)^{1/2} \).

\( F_p = \) percentage bandwidth between peaks of response-shape \( C \equiv (\Delta f_p/f_0) \).

\( F_p = \) percentage bandwidth between edges of the defined pass-band width \( = (f_0 - f_0)/(f_0 f_2)^{1/2} \).

\( d \) = total decrement of a resonant circuit. See Figs. 3 and 4.

\( G_{n} = \) total conductance across \( n \)th resonant circuit of node network \( = \left( \frac{1}{R_n} + \frac{1}{Q_{L}X_{u,n} + \frac{1}{Q_{L}} X_{w,n}} \right) \).

See Fig. 4(a).

\( R_n = \) total resistance in series with \( n \)th resonant circuit of mesh network \( = \left( R_n + \frac{X_{u,n}}{Q_{L}} + \frac{X_{w,n}}{Q_{L}} \right) \).

See Fig. 4(b).

\( Q = \) inverse of the total resonant-circuit decrement.

\( K = \) resultant coefficient of coupling between resonant circuits. (See Figs. 3 and 4, and Sections 3.1.2 and 3.2.2.)

\( I = \) magnitude of the equivalent constant-current generator that drives the network of Fig. 3. For a pentode generator \( I = e_n I'_p \). For a "transformed" low-resistance generator, see Fig. 6.

\( C_{n} = \) total resonated capacitance in the \( n \)th resonant circuit.

\( L_{n} = \) total resonated inductance in the \( n \)th resonant circuit.

\( U_p = \) general symbol for a coefficient of some power of \( (jF) \) in the complex polynomial form of the circuit-response equations. The subscript of the \( U \) is identical with the power of \( (jF) \) for which \( U \) is the coefficient. (See Section 6.2 and (2.1).

\( L_{p} \) = general symbol for a coefficient in that complex polynomial that produces the desired response-shape \( B \) and \( C \), respectively. The subscript is identical with the power of \( (jF) \) for which \( U \) is the coefficient. \( A_n, B_n, C_n, \ldots \) = specific coefficient of that \( (jF) \) whose power is \( n \), in that specific complex polynomial for a network that has as many resonant circuits as the numerical position of the letter in the alphabet, e.g., \( C_2 \) would be the coefficient of \( (jF)^2 \) in the polynomial

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* Decimal classification: R143.2. Original manuscript received by the Institute, May 11, 1948; revised manuscript received, March 25, 1949. Presented, 1948 IRE National Convention, New York, N.Y., March 22, 1948, under the title, "Application of Tschebyscheff polynomials to the exact design of band-pass filters."

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\( n \) This name is spelled variously in English, commonly as "Tschebyscheff."

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for a triple-tuned network. (See Design Equations — Groups 2 and 3.)

\[ \Delta n_{\text{min}} = \text{Minimum value of the magnitude of the complex polynomial for an } n\text{-resonant-circuit network.} \]

(See (3).) \[ \Delta n_{\text{real}}, \Delta n_{\text{imag}} = \text{same as above, except for the complex polynomial for shapes } B \text{ and } C \text{, respectively.} \]

\( r_n, \gamma_n = \text{same as above for response-shape } C. \) (See (26).)

\[ \theta_n, \phi_n = \text{phase angle, at the percentage bandwidth } F_n \text{ of the response shape } (V_n/V'), \text{i.e., the } \theta \text{ in } (V_n/V') < \theta, \text{ for amplitude-response-shapes } B \text{ and } C \text{ respectively, for an } n\text{-resonant-circuit network.} \]

(See (10).) \[ T_n(F/F_0) = \text{general Chebyshev}^4 \text{ polynomial in terms of the variable } (\Delta f/\Delta f_0) \text{ of highest power } n. \] (See (17), (18), and (19).)

\( s_n, \epsilon_n = \text{defined by (26).} \)

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2. Introduction

As the need increases for more and more channels in a given frequency band and as the voltage ratios between desired and undesired signals continue to increase, there will also be an increasing need for design information for band-pass filter networks that is more exact than that supplied by classical filter theory.

It is well known that when a continuously increasing attenuation is required outside of a given pass band, a straightforward method of producing a very high rate of increase in attenuation is to use a large number of correctly coupled and correctly damped resonant circuits.

For many designers, a further very important practical requirement arises at this point. For the exact design to be of practical value, no lossless elements (no infinite Q's) should be required. For multiple-resonant-circuit filters, this last requirement has apparently not received much attention. In general, it has been stated that any dissipation in the reactive elements of a filter degrades its performance at the edges of the pass band.

With correctly designed networks, this last statement is not true. With correct circuit element values using finite Q's, the amplitude response can be made identical to that obtained with infinite-Q elements.

To show the increase in sharpness of cutoff as the number of resonant circuits is increased, the graphs of

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Fig. 1—Selectivity characteristic of 5 cascaded stages, each stage containing a correctly resonated n-resonant-circuit network having a proper Q distribution and being critical-shape-coupled. When \( f_0 \) and \( \Delta f \) are given, the two frequencies between which \( \Delta f \) occurs are \( f_{1b} = f_0 + \Delta f/2 \) and \( f_{1a} = f_0 - \Delta f/2 \). The number of tuned circuits per stage \( n \) is indicated on each curve. See (4) and (4a).

Fig. 2—Selectivity characteristic of 5 cascaded stages, each stage containing a correctly resonated n-resonant-circuit network having a proper Q distribution and overcoupled for a 1-dB peak-to-valley ratio. When \( f_0 \) and \( \Delta f \) are given, the two frequencies between which \( \Delta f \) occurs are \( f_{2b} = (f_0 + (\Delta f/2))^{12} + (\Delta f/2) \) and \( f_{2a} = (f_0 + (\Delta f/2))^{12} - (\Delta f/2) \). When \( f_{1a} \) and \( f_{1b} \) are given, then \( f_{2} = (f_{1b})^{12} \). The number of tuned circuits per stage \( n \) is indicated on each curve. See (14) and (15).
takes place, and ends with $T_1$ cut off and $T_3$ carrying maximum current. Since the grid-leak resistor $R_3$ is returned to 250 volts, the grid voltage of $T_3$ increases exponentially toward 250 volts as $C_3$ discharges. When the cutoff point is reached, the plate current of $T_3$ begins to flow, a second switching process occurs, and the circuit returns to its normal condition. The sharpness of output pulse depends on the time constant $R_3C_3$.

The circuit of Fig. 4 generates a sharp negative pulse as soon as the unknown voltage equals the reference voltage. At this instant, diode $T_3$ conducts, and the amount of regeneration is hereby greatly increased. Therefore, cumulative processes will take place. The purpose of diode $T_1$ is to prevent $C_3$ from discharging through the unknown voltage source. The coupling transformer should provide a phase shift of 180 degrees. If the duration of the pulse is to be determined by the time constant $R_1C_1$, the mutual inductance of the coupling transformer must be large enough to support the pulse without material decay.

**Peak Voltmeter Circuits**

Regenerative amplifiers are well adapted to the measurement of the peak amplitude of voltage pulses. A peak voltmeter circuit, based upon the switching properties of a regenerative amplifier, is shown in block form in Fig. 5. Before the application of an input pulse, triode $T$ is cut off, and the potential across $C$ is zero. The regenerative amplifier is triggered by the leading edge of the input pulse, and produces a positive voltage to drive the grid voltage of $T$ to zero, or slightly positive. Then the memory capacitor $C$ charges. The voltage across $C$ is coupled through the cathode-follower to the indicating device which consists of a dc milliammeter and a resistor in series. The input resistance of the cathode-follower is made very high in order to prolong the hold-on time of the memory capacitor $C$. The voltage on the indicating device is fed back negatively to the input terminal of the regenerative amplifier for the purpose of reducing the potential developed by the input pulse. Therefore the regenerative amplifier will trigger off as soon as the voltage on the indicating device equals the amplitude of the input pulse. This in turn cuts off the plate current of $T$ and blocks $C$ from further charging. Once triggered off, the regenerative amplifier remains so until the input voltage rises above the voltage on the indicating meter.

The principal advantages of peak voltmeters of this type are: (1) Errors encountered in measuring pulses of low repetition frequency and short duration are greatly reduced. This results because the use of the B-supply instead of the measured pulse or the output of an amplifier to charge the memory capacitor greatly shortens the charging time, and makes it possible to prolong the hold-on time by increasing the capacity of the memory capacitor. Furthermore, this meter employs a regenerative amplifier to trigger the charging current, thus permitting the final voltage of the memory capacitor to reach a value equal to the peak amplitude of the input pulse. (2) The meter reading is practically independent of circuit constants and operating voltages because the instrument is based on a comparison principle.

The circuit diagram of a practical instrument of this type is shown in Fig. 6. $T_1$, $T_2$, and $T_3$ are connected as a regenerative amplifier similar to that of Fig. 1. The function of $T_4$ is identical to that of triode $T$ of Fig. 5. The negative bias voltages of $T_1$ to $T_3$ are all supplied by gas diode $T_6$. $C$ functions as the memory capacitor. $T_8$ serves as a cathode-follower with output meter $M$ and $R_5$ in series as its load. To minimize the leakage due to the grid current of $T_6$, resistor $R_2$ should be so dimensioned that $T_4$ operates at the “floating-grid” potential, at which the positive grid current becomes equal to the negative (positive-ion) grid current. Switch $S$ is used for reset, and potentiometer $R_4$ is the zero panel control. Before applying a signal to the input, potentiometer $R_1$ is adjusted so that $T_1$ operates just at the critical point of cutoff, a milliammeter may be connected in series with the cathode of $T_1$ as an indicator for this adjustment. The regenerative amplifier is triggered by the application of a positive input pulse, and then $T_3$ conducts because the plate current of $T_2$ suddenly falls to zero. This in turn causes memory capacitor $C$ to charge
and develops a negative voltage across meter $M$ and $R_6$. While this instrument is designed for measuring positive pulses, in conjunction with a phase inverter, it may also be used to measure negative pulses.

**Pulse-width Discriminator**

The name applied to circuits whose output is proportional to the width of the input pulse is pulse-width discriminator. A regenerative amplifier may be used to augment the special circuit to be described to form a pulse-width discriminator. In Fig. 7, the input to the grids of $T_1$ and $T_2$ is taken from the plate of the normally off tube of the regenerative amplifier, which is adjusted to trigger on and off when the input voltage reaches levels $a$ and $b$ respectively. Thus, the regenerative amplifier acts as a pulse-slicer. When the circuit is in its quiescent condition, the grid-to-cathode potential of both $T_1$ and $T_2$ is zero, and $T_5$ (a sharp cutoff triode) is operating just at the critical point of cutoff. $T_1$ and $T_2$ are chosen to be identical in characteristics, and $R_3$ is made equal to $R_4$. Thus $e_{c1}$ and $e_{c2}$, the potentials across capacitors $C_1$ and $C_2$ respectively, are equal to $E_0$ (see Fig. 8). The application of an input pulse drives the grid of $T_1$ to $E_0$, causing $T_1$ to conduct. At the same time, the grid voltage of $T_2$ is taken to be zero, which prevents $T_2$ from conducting. During each cycle, the output voltage is equal to $E_0$. After the pulse ends, both $C_1$ and $C_2$ discharge through triode $T_3$. If the plate resistance of diode $T_3$ is very small, the potentials across both capacitors are approximately equal and decrease exponentially toward $E_0$ with a time constant equal to $R_3 C_3$. During the next cycle, $T_3$ conducts and capacitor $C_2$ charges because the grid of $T_2$ rises with $e_{c2}$. After the time $t_3$, triode $T_3$ is cut off and capacitor $C_2$ does not discharge. During the next cycle, $T_3$ will not conduct unless the width of the input pulse increases or the voltage across $C_3$ drops below that at the end of the previous cycle. The time between $t_1$ and $t_2$ serves to insure that the output voltage increases by an amount equal to $E_0$. Therefore $(t_2 - t_0)$ must be long enough to allow $C_2$ to charge up to a voltage equal to $E_0$. When the circuit of Fig. 7 is required to follow a series of pulsed signals of decreasing width, a proper resistance must be used to shunt capacitor $C_3$. In general the time constant of $C_3$ and its shunting resistance should be small enough to allow the output voltage to follow the decreasing of the input pulse width, and at the same time must be large enough to permit $C_2$ to hold most of its charge between pulses.

**Acknowledgment**

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MILTON DISHAL†, SENIOR MEMBER, IRE

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The equation giving the gain obtained with the desired response shape is derived. Equations for the exact phase characteristics associated with the above exact amplitude characteristics are also given.

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Design sheets are presented giving the necessary equations for single-, double-, triple-, and stagger-tuned networks to produce either of the above two amplitude-response shapes.

1. Symbols

\( n \) = total number of resonant circuits in networks of Figs. 3 and 4.

\( N \) = total number of cascaded networks between which there is no coupling, e.g., separated by vacuum tubes.

\( V \) = complex voltage output at any frequency.

\( V' \) = magnitude of the voltage output at any frequency. (See Fig. 7.)

\( V_p \) = magnitude of the voltage output at the frequency of peak response. (See Fig. 7.)

\( V_p' \) = magnitude of the voltage output at that point on the response curve that the designer defines as the edge of the pass band. For response-shape \( C \), this output voltage is identical with the response at the valleys of the response inside the pass band. (See Fig. 7.)

\( \Delta f \) = frequency bandwidth between response points whose voltage output is \( V \). (See Fig. 7.)

\( \Delta f_p \) = frequency bandwidth between the peaks of response-shape \( C \) at the voltage output of \( V_p \). (See Fig. 7.)

\( \Delta f_p' \) = frequency bandwidth between the response points whose voltage output is \( V_p' \), i.e., the frequency bandwidth between the edges of the defined pass-band width. (See Fig. 7.)

\( f_0 \) = resonant frequency of each resonant circuit. See Section 3.2. \( f_0 \) is also the geometric mean frequency between two frequencies \( f_i \) and \( f_2 \) having the same voltage response.

\( \omega_0 \) = resonant radian frequency = \( 2\pi f_0 \)

\( F \) = total percentage bandwidth between two frequencies \( F = \left( \frac{f_2 - f_1}{f_1} \right) / f_1 \), where \( f_n = (f_1 f_2)^{1/2} \).

\( F_p \) = percentage bandwidth between peaks of response-shape \( C \) = \( \Delta f_p / f_1 \).

\( F_p' \) = percentage bandwidth between edges of the defined pass-band width = \( (f_2 - f_1) / f_1 f_n \).

\( d \) = total decrement of a resonant circuit. See Figs. 3 and 4.

\( G_n \) = total conductance across \( n \)th resonant circuit of node network \( \left( \frac{1}{R_n} + \frac{1}{Q_1 X_{atn}} + \frac{1}{Q_2 X_{atn}} \right) \). See Fig. 4(a).

\( R_n \) = total resistance in series with \( n \)th resonant circuit of mesh network \( \left( R_n + \frac{X_{atn}}{Q_1} + \frac{X_{atn}}{Q_2} \right) \). See Fig. 4(b).

\( Q \) = inverse of the total resonant-circuit decrement.

\( K \) = resultant coefficient of coupling between resonant circuits. (See Figs. 3 and 4, and Sections 3.1.2 and 3.2.2.)

\( l \) = magnitude of the equivalent constant-current generator that drives the network of Fig. 3. For a pentode generator \( l = g_m v_0 \). For a "transformed" low-resistance generator, see Fig. 6.

\( C_n \) = total resonated capacitance in the \( n \)th resonant circuit.

\( L_n \) = total resonated inductance in the \( n \)th resonant circuit.

\( u_p \) = general symbol for a coefficient of some power of \( (jF) \) in the complex polynomial form of the circuit-response equations. The subscript of the \( u \) is identical with the power of \( (jF) \) for which \( u \) is the coefficient. (See Section 6.2 and (2).)

\( u_p', u_p'' \) = general symbol for a coefficient in that complex polynomial that produces the desired response-shape \( B \) and \( C \), respectively. The subscript is identical with the power of \( (jF) \) for which \( u \) is the coefficient.

\( A_p, B_p, C_p \) etc. = specific coefficient of that \( (jF) \) whose power is \( p \), in that specific complex polynomial for a network that has as many resonant circuits as the numerical position of the letter in the alphabet, e.g., \( C_2 \) would be the coefficient of \( (jF)^2 \) in the polynomial.
for a triple-tuned network. (See Design Equations—Group 1.)

\[ A^p, B^p, A^p, B^p; \text{etc.} = \text{same as above, except applied to}
\]
\[ \text{that specific polynomial that produces response-shapes } B \text{ and } C, \text{ respectively. (See Design Equations—Groups 2 and 4.)}
\]

\[ |\Delta_{a}|_{\min} = \text{Minimum value of the magnitude of the complex polynomial for an } n\text{-resonant-circuit network.}
\]

(See (3).)

\[ |\Delta_{a}| \text{ and } |\Delta_{a}'|_{\min} = \text{same as above, except for the complex polynomial for shapes } B \text{ and } C, \text{ respectively.}
\]

\[ r_m, r_m' = \text{magnitudes of the real and imaginary parts, respectively, of the general expression for the complex roots of the equation giving response-shape } B. \text{ These}
\]
\[ \text{roots always occur in conjugate pairs and } m \text{ is the}
\]
\[ \text{pair number. (See Section 7.1 and (6).)}
\]

\[ r_m, r_m' = \text{same as above for response-shape } C. \text{ (See}
\]

\[ \theta_n, \theta_n' = \text{phase angle, at the percentage bandwidth } F, \text{ of}
\]
\[ \text{the response shape } (V_p/V), \text{i.e., the } \theta \text{ in } (V_p/V) < \theta,
\]
\[ \text{for amplitude-response-shapes } B \text{ and } C \text{ respectively,}
\]
\[ \text{for an } n\text{-resonant-circuit network. (See (10).)}
\]

\[ T_n(F/F_p) = \text{general Chebyshev}^1 \text{ polynomial in terms of}
\]
\[ \text{the variable } (\Delta f/\Delta f_p) \text{ of highest power } n. \text{ (See (17),}
\]

\[ (18), \text{ and (19).)}
\]

\[ s_n, e_n = \text{defined by } (26).
\]
Figs. 1 and 2 should be examined. For five cascaded stages, and a required adjacent-channel rejection of 100 db, the use of three resonant circuits per stage having an allowable peak-to-valley ratio in the pass band of 1 db more than doubles the available number of channels over the use of two resonant circuits per stage with critical-shape-coupled characteristics.

The major purpose of this paper is to present a method and some of the resulting equations for obtaining the necessary \( n \) simultaneous equations that must be solved to determine the exact circuit constants required for band-pass circuits that use \( n \) finite-\( Q \) resonant circuits \((n=1, 2, 3, 4, 5, \text{etc.})\) to produce either one of two types of exact response shapes; the so-called critical shape-coupled (maximally flat or transitional-shape-coupled) response, and the so-called overcoupled response.

Briefly the method proposed is as follows:

A. Express the determinant of the exact circuit-response equation in the form of a polynomial in \( j(\Delta f/f_o) \) of highest power \( n \), with descending consecutive powers.

B. Solve for the complex roots of the exact amplitude equation that describes the desired-response shape.

C. Use these roots to place the desired-response equation in the complex polynomial form described in A.

D. Equate the corresponding \( n \) coefficients.

3. Circuits and Circuit Constants

3.1 Circuit Producing an Amplitude-Frequency Characteristic Having Exact Geometric Symmetry

Fig. 3 shows the basic unbalanced band-pass "ladder network." As in the well-known constant-\( K \) filter, the reactance structure of the series and shunt arms are in reverse arms. However, it should be noted that in this paper each resonant circuit is considered to be dissipative, each circuit having its own specific \( Q \). When correctly designed, this circuit can produce for any percentage bandwidth an amplitude response exactly described by the Butterworth and Chebyshev equations given in sections 7 and 8. This result can be accomplished using finite \( Q \)'s in all resonant circuits. It is worth repeating that, when this circuit can be used, no small-percentage passband approximation is required.

The chain may start and/or end with either a series or shunt arm. It should be realized that if the network starts with a series arm and a constant-current generator, e.g., pentode tube, drives the network, then the constant-current generator should be connected across the resistor that produces the required resonant-circuit decrement. Similarly, if the network ends in a series arm and output voltage is to be used, it must be obtained across the resistor that produces the correct resonant circuit decrement.

The circuit constants that exactly and conveniently describe the circuit of Fig. 3 are the resonant frequency \( f_o \) of the series and shunt arms, the coefficient of coupling \( K \) between adjacent resonant circuits, and the decrement \( d \) of each resonant circuit. The definitions of these circuit constants are obtained from examination of the exact circuit response equations for a specific network in the form of Fig. 3 (using, for example, five resonant circuits).

3.1.1 Resonant Frequency \( f_o \)

In the circuit of Fig. 3, all series and shunt arms are tuned to exactly the same resonant frequency, \( f_o = 1/2\pi (LC)^{1/2} \). This frequency will also be the geometric mean frequency \((f_{12})^{1/2} \) between any two frequencies \( f_1 \) and \( f_2 \) having the same amplitude response.

3.1.2 Coefficients of Coupling

It is found that the determinant of the network of Fig. 3 can be expanded into the form of a polynomial in descending consecutive powers of \( j(f/f_o - (f/f_o)) \); the coefficients of these various powers are completely independent of frequency and involve only the decrements of the resonant circuits and certain capacitance ratios. The capacitance appearing in the numerator of these ratios is always that of a series resonant circuit, and the denominator capacitance is always that of an adjacent shunt resonant circuit. In this paper, these capacitance ratios will be called coefficients of coupling (squared) because they define a required relation between adjacent resonant circuits, and also because they are exactly equivalent to the well-known coefficients of coupling in the small-percentage pass-band circuits of Figs. 4(a) and 4(b). Thus, between any two adjacent resonant circuits, \( K = C_{\text{series}}/C_{\text{parallel}} \).

3.1.3 Resonant-Circuit Decrement

As they appear in the determinental equation, the resonant-circuit decrements \( d \) are the inverse of the well-known resonant-circuit \( Q \). For any series arm, \( d = R_s/\omega_0 C_s \), where \( R_s \) is the resistance in series with that arm and \( C_s \) is the resonated capacitance of the arm. For any shunt arm, \( d = 1/R_p/\omega_0 C_p \), where \( R_p \) is the resistance in parallel with that arm and \( C_p \) is the resonated capacitance of the arm.
For high- and very-high-frequency band-pass circuits where shunt capacitance of the usual generators cannot be neglected, it should be noted that if the circuit of Fig. 3 is used only odd numbers of resonant circuits can be employed, i.e., the first and last resonant circuits must be parallel arms.

\[ L_0 = \frac{1}{(q_0 - \gamma) \omega_0^2} \]

**Circuit Constants**

**Resonant Frequency**

\[ \frac{g_0}{C_0} \]

**Coefficient of Coupling**

\[ K = \frac{g_0}{C_0} \]

**Decrement**

\[ \frac{d}{q_0} = \frac{1}{Q_0} \]

**Fig. 4(a)**—Basic node network to be analyzed and the resonant-circuit constants that are used. For small-percentage pass bands, this network is exactly equivalent to that of Fig. 3, when equal numbers of resonant circuits are used.

### 3.2 Circuits Producing an Amplitude-Frequency Characteristic Having Geometric Symmetry for Only a Small-Percentage Bandwidth

When the required pass band is small, the circuits of Figs. 4(a) and 4(b) are exactly equivalent to that of Fig. 3, and are usually more practical to build physically. When a small-percentage band pass is needed, the circuit of Fig. 3 is often not the most desirable. For instance, the required values for the coefficients of coupling between adjacent resonant circuits \([C_{\text{series}}/C_{\text{parallel}}]\) for Fig. 3 are approximately equal to the percentage bandwidth; thus we see that, for a 5 per cent bandwidth, the resonating capacitance of a series arm would have to be approximately 5 per cent of that used in the adjacent parallel branch; to satisfy the resonance requirements, the inductance in the series branch must be twenty times that in the adjacent shunt branch; a satisfactory inductance of this size is often undesirable or impractical.

Figs. 4(a) and 4(b) are not exactly equivalent to Fig. 3 because the effective coefficients of coupling between adjacent resonant circuits, which appear in the determinantal equations for the circuits of Figs. 4(a) and 4(b), are functions of frequency; i.e., we find that for Fig. 4(a), e.g., \( K = K_C(f/f_0) - (K_L + K_M)(f_0/f) \). If we make the approximation that \( K_C \approx (K_L + K_M) \) and \( f_0/f = 1 \), then the above \( K \) varies negligibly with frequency, and for the same number of resonant circuits the determinantal equations for Figs. 4(a), and 4(b) are identical.

The above assumption automatically means that the response null, which can be obtained with the circuits of Figs. 4(a) and 4(b), is placed far from the pass band. In the circuits of Fig. 4(a), this response null occurs exactly at \( f_{\text{null}}/f_0 = [(K_L + K_M)/K_C]^{1/2} \). In the circuits of Fig. 4(b), this response null occurs exactly at \( f_{\text{null}}/f_0 = [K_C/(K_L + K_M)]^{1/2} \).

Since the circuit of Fig. 4(a) has \( n \) nodes (where \( n \) is the number of resonant circuits used) and is most simply analyzed by the use of node equations, it will be called the node network, and the \( n \)-mesh circuit of Fig. 4(b) will be called the mesh network. It should be realized that these networks are physically different (thus supplying the designer with a variety of physical configurations) but are electrically related, in that wherever \( G, C, L, I, \) and \( V \) appear in the determinantal equation for Fig. 4(a), then \( R, L, C, E, \) and \( I \) appear in the corresponding determinantal equation for Fig. 4(b) (principle of duality).

**Fig. 4(b)**—Basic mesh network to be analyzed and the resonant-circuit constants that describe it. For small-percentage pass bands, this network is exactly equivalent to that of Fig. 3, when equal numbers of resonant circuits are used.

It may be noted that when we use only the first two nodes of Fig. 4(a) with mutual-inductive coupling between these nodes, the familiar intermediate-frequency transformer of the common broadcast receiver is obtained.

For a small-percentage band pass, the constants that best describe the above circuits are the coefficients of coupling \( K \) between resonant circuits, the resonant frequency \( f_0 \) of the resonant circuits, the decrement \( d \) of the resonant circuits (inverse of the resonant-circuit \( Q \)).

It may be helpful for the reader to realize that the definitions of the above constant circuits are obtained from the exact node equations for a node network and from the exact mesh equations for a mesh network. Thus, if the reader will write the exact node equations for a triple-tuned circuit, then by correct manipulation, the above constants will be recognized. These constants, as they appear in the exact node and mesh equations, will now be described briefly.

#### 3.2.1 Exact Resonant Frequency \( f_0 \)

The resonant frequency \( f_0 \) of each node is that frequency at which the susceptance of the total capacitance (including mutual) attached to the node equals the susceptance of the total inductance (including mutual).
attached to the node. Figs. 4(a) and 4(b) give this exact general resonance frequency.

In line with this definition, it should be noted that a fundamental and practical method of experimentally "tuning up" the resonant circuit attached to any node is effectively to short-circuit the two nodes on either side of the node in question, and then tune this resonant circuit for maximum output. In practice, the effective short-circuiting can be done by completely detuning the two adjacent nodes, thus allowing some signal transfer through the filter so that an output indicator at the end of the filter chain can be used for all the nodes. Mesh networks can be tuned up by effectively open-circuiting the two meshes on either side of the mesh in question, and then tuning the desired mesh for maximum response.

### 3.2.2 Coefficient of Coupling

The coefficient of any one type of coupling, $C$, $L$, or $M$, between any two nodes is the ratio of the susceptance of that type common to the two nodes in question to the geometric mean of the total susceptance of that type connected to each node. For the correct coefficient of coupling between any two meshes of Fig. 4(b), substitute in the above statement the word reactance for susceptance and mesh for node. Figs. 4(a) and 4(b) give these coefficients.2

Either inductive, capacitive, or mutual-inductive coupling, or a combination of them, may be used between the adjacent resonant circuits. The resultant coefficient of coupling is

$$K = \left[K_c(f/f_0)\right] - (K_L + K_M)(f_0/f)$$

and in the analysis to be considered, the frequency at which this quantity equals zero must not occur within or near the pass band.

It should be realized that it is not necessary to use the same type of coupling between all the nodes of the network.

The designer will find that there is less chance of making an error in designing the coefficient of coupling if the following procedure is used: first, decide whether to use a node or a mesh network; second, decide what type of coupling to use; third, if a node network is to be used design the network using the exact circuit configuration of Fig. 3 or if a mesh network is to be used use the exact configuration of Fig. 4; fourth, after the above design is completed then the $T$, $\pi$, or transformer equivalents of Fig. 5 can be used to obtain different circuit configurations that, depending on the specific problem, may be desirable.

### 3.2.3 Resonant-Circuit Decrement $d = 1/Q$

For node networks, the decrement $d$ of the resonant circuit is the ratio of the resultant equivalent conductance across the resonant circuit to the susceptance of the resonant frequency of either the total capacitance or total inductance. The term "equivalent" is used to indicate that any series resistance in the inductance or capacitance of the resonant circuit should be transformed into its equivalent shunt conductance and added to any actual shunt conductance present to obtain the total resonant-circuit conductance. Figs. 3 and 4 give the resonant-circuit decrement. Naturally any shunt conductance due to the generator and the load must be considered, when calculating the decrement of the input and output resonant circuits.

#### 3.2.4 Equivalent Generator

With reference to the equivalent constant-current generator that drives the first node, there are in practice two situations to be considered. If a vacuum tube is attached directly to the input circuit, the equivalent generator is, of course, $g_mV_0$. If a transforming circuit is used to couple the generator to the resonant circuit, then Fig. 6 gives the equivalent "reflected constant-current generator" for use with the node circuits (equivalent constant-voltage generator for use with the mesh circuits) and the "reflected decrement" portion of the total resonant-circuit decrement that results when a resistive generator and/or load are "transformed" into the first and last resonant circuits of the network. The equivalents of Fig. 6 thus allow application of the following analysis to the important practical cases where the actual generator is not a pentode but is, for example, the much-used equivalent 50-ohm generator and the load is, for example, a low-input-resistance crystal mixer. For this example, one could use a transforming circuit as given in Fig. 6 to couple the untuned generator and untuned load to the first and last resonant circuits of the chain.

When a certain response shape is to be produced, the function of the transforming circuit that couples a nonresonant generator and/or load to the first and/or last resonant circuit should be thought of in the following manner: The transforming circuit is used in conjunction with the generator and/or load resistance to couple a certain amount of decrement to the resonant circuit to make the total resulting resonant-circuit decrement equal to that value required to produce the desired response shape. Note that one does not design the transforming circuit to produce a certain desired equivalent constant-current (or constant-voltage) generator. The total resonant-circuit decrement is the sum of the above-considered "reflected decrement" and the "unloaded decrement" of the resonant circuit (which is the inverse of the unloaded $Q$ of the resonant circuit). The inverse of the sum of these two decrements is then, of course, the resultant resonant-circuit $Q$.

### 4. Basic Response Shapes

When the resonant circuits are correctly tuned, there are three basic types of symmetrical band-pass shapes
that can be used to give the same pass band width.

Shape A: This shape corresponds to that obtained with the well-known "under-coupled" condition for the familiar double-tuned circuit. It may be described as a shape having a single maximum in the center of the pass band and the attenuation outside the pass band does not increase as rapidly as possible.

Shape B: This shape corresponds to that obtained with the well-known "critical-shape coupled" condition for the familiar double-tuned circuit, and has also been called the "maximally flat" and the "transition" shape. It may be described as the type having the squarest possible single maximum in the center of the pass band and its attenuation outside the pass band increases as rapidly as possible while still maintaining a single-peaked response.

Fig. 5—Five different coupling methods and the transformer, $T$, and $r$ equivalents. The use of these equivalents in the circuits of Fig. 4 enables the designer to obtain the same electrical performance with a large number of equally different circuits.
Shape C: This shape corresponds to that obtained with the well-known "overcoupled" condition for the familiar double-tuned band-pass circuit. This type of shape has \( n \) maxima of equal height and \( (n - 1) \) minima of equal height inside the pass band, where \( n \) is the number of resonant circuits used. For a given allowable number of decibels down for the edges of the pass band, this shape gives the maximum possible rate of increase of attenuation outside the pass band.

An additional characteristic of response-shape C should be mentioned here: It will be found that no matter what the peak-to-valley ratio, there is a fixed ratio, dependent only on \( n \), between the bandwidth across the outside peaks and the bandwidth of the skirts at that number of decibels down equal to the valley response. With response-shape \( C \), the symbol \( \Delta f_3 \) denotes this particular skirt bandwidth at the response value \( V_3 \) that is equal to the valley response \( V_2 \).

Shape \( C_1 \): In many practical cases, the designer can allow, for example, 1-db dips inside the pass band, but would like to define the edges of the pass band as the 3-db down points. This type of response shape is a simple modification of the basic shape-\( C \) response. For example, suppose four coupled resonant circuits are to be used with an allowable peak-to-valley ratio of 1 db; the ratio between \( \Delta f_{3db} \) and \( \Delta f_{3db} \) is fixed by the above data and, if we know this ratio, we thus know that \( \Delta f_{3db} \) required to give the desired \( \Delta f_{3db} \) bandwidth. One then designs for the basic shape-\( C \) response (i.e., 1-db dips and the "pass-band edges" at 1 db down equal to the above-found \( \Delta f_{3db} \)).

In this paper, only response-shapes \( B \) and \( C \) will be
considered. Fig. 7 gives the above described three response shapes, and the shape constants that describe the responses, for the voltage produced across the last node, when the node network of Fig. 4(a) is driven by a constant-current generator, and for the current produced in the last mesh when the mesh network of Fig. 4(b) is driven by a constant voltage generator.

5. Mathematical Procedure

The analytical procedure to be used in this paper consists of the following steps.

Step A. Express the general circuit response in its simplest possible form. For the networks of Figs. 3, 4(a), and 4(b), this form is that in which the determinantal equation is expressed as a polynomial in descending powers of $j(f/f_0) - (f_0/f)$.

Step B. Express the desired response equation in the same form as the general circuit-response equation.

Step C. Equate the corresponding coefficients in the two equations.

This will produce the necessary number of simultaneous equations.

6. Circuit Response Equations

6.1 Polynomial Coefficients of Complex-Circuit-Response Equation

Consider the design of the shape of the transfer impedance of the networks of Figs. 3 and 4(a), i.e., the resulting output voltage when the networks are correctly driven by a constant-current generator; and for the shape of the transfer admittance of the networks of Figs. 3 and 4(b), i.e., the resulting output current when the networks are correctly driven by a constant-voltage generator.

By the straightforward application of Kirchhoff's laws to any specific network in the form of Figs. 3 and 4, we find that the above-described transfer admittance and transfer impedance can be exactly written as

$$\left( \frac{V_{\text{out}}}{I_{\text{gen}}} \right) \quad \text{or} \quad \left( \frac{I_{\text{out}}}{E_{\text{gen}}} \right) = (jF)^n + U_{n-1}(jF)^{n-1} + U_{n-2}(jF)^{n-2} \cdots U_1(jF)^2 + U_0,$$

For all the node networks of Fig. 4(a) and for all the networks of Fig. 3 that begin and end with shunt arms, the numerator for $(V_{\text{out}}/I_{\text{gen}})$ of (1) is

$$(\text{Numerator})_n = \frac{1}{\omega_0(C_0C_n)^{1/2}} K_1K_2K_3 \cdots K_{(n-1)n}, \quad (1a)$$

where the $K$'s are defined in Figs. 3 and 4(a).

For all the mesh networks of Fig. 4(b) and for all the networks of Fig. 3 that begin and end with series arms, the numerator for $(I_{\text{out}}/E_{\text{gen}})$ of (1) is

$$(\text{Numerator})_n = \frac{1}{\omega_0(L_1L_n)^{1/2}} K_1K_2K_3 \cdots K_{(n-1)n}, \quad (1b)$$

where the $K$'s are defined in Figs. 3 and 4(b).

The denominator of (1) is a polynomial in consecutive powers of $j(f/f_0) - (f_0/f)$, the highest power being $n$ the number of resonant circuits used and, within the limitations of the discussion of Section 3, the coefficients $(U)$ of the polynomial are independent of frequency and are functions of only the coefficients of coupling $K$ and decrements $d$ as they are described in Section 3.

Since the numerator of (1) is independent of frequency within the limitations of the discussion of Section 3, the response shape is fixed entirely by the denominator of (1). For the same number of resonant circuits, the denominators $\Delta_n$ are identical for all the networks of Figs. 3 and 4.

In the general notation of (1), $U$ is used to denote an arbitrary number of resonant circuits. For any specific network, the letter used in the coefficients will be that one whose numerical position in the alphabet corresponds to the number of resonant circuits in the network. Thus, for a 5-resonant-circuit network, we would use $E$ to represent the coefficients. The subscript on any coefficient is exactly the same as the power of the $(jF)$ for which it is the coefficient. Thus, for example,
for a 4-resonant-circuit network, the coefficient of \((jF)^3\) would be represented by \(D_3\).

Since the numerator of (1) is independent of frequency, the peak value of the transfer admittance or transfer impedance is given by

\[
\left( \frac{V_{\text{out}}}{I_{\text{gen}}} \right)_{\text{peak}} \text{ or } \left( \frac{I_{\text{out}}}{E_{\text{gen}}} \right)_{\text{peak}} = \frac{\text{(Numerator)}_n}{\Delta_n|_{\text{min}}},
\]

where \(\Delta_n|_{\text{min}}\) is the minimum magnitude of the polynomial of (1).

Equation (2) will be used in a later section of this paper to find the gain obtained at the amplitude response peaks.

Dividing (2) by (1), we obtain the ratio of the peak response to the response at any frequency; this is the basic response-shape equation:

\[
\frac{V_n}{I_n} \text{ or } \frac{I_n}{I} = \left| \frac{(jF)^n + U_{n-1}(jF)^{n-1} + \cdots + U_2(jF)^2 + U_1(jF) + U_0}{\Delta_n|_{\text{min}}} \right| \quad (3)
\]

In Design Equations—Group 1, the specific circuit-response-shape equations are listed for \(n = 1, 2, 3,\) and 4.

Careful study of the series of response equations given in Design Equations—Group 1 will show the law of formation of the coefficients, and it should now be possible to write the general exact complex response equation for a chain containing any number of resonant circuits.

Insofar as plotting of a response curve is concerned, the above equations enable us to obtain methodically the equations required to plot the amplitude- and phase-response shapes for any number of tuned circuits, once the various circuit coefficient of couplings and \(Qs\) are specified. The complex equation of Design Equations—Group 1 are, of course, expanded into their magnitude and phase-angle form when we plot the amplitude-and phase-response shapes. (All the even powers of \((jF)\) are algebraically added together to give the real part of the determinant, all the odd powers of \((jF)\) are added together algebraically to give the imaginary part of the determinant.)

7. RESPONSE SHAPE B

7.1 Desired-Shape Equation

Our next step is to express the desired-shape equation given below in the form of Design Equations—Group 1, so that we can equate the above coefficients \((U_1, U_2, U_3, \text{ etc.})\) to the corresponding coefficients of the desired-response equation.

When the straightforward procedure mentioned in Section 5 of this paper is applied to single-, double-, and triple-tuned filters, we find that the magnitude equations that result are exactly given by the general equation (4). This equation and the corresponding response shape are also shown in Fig. 8.
where the real \((r_n^b)\) and imaginary \((i_n^b)\) parts of the roots are given by (6a) and (6b).

\[
r_m^b = F_\beta \left| 1/\left( (V_p/V_\beta)^2 - 1 \right)^{1/2} \right| \sin \left( \frac{2m-1}{n} \pi \right), \quad \text{(6a)}
\]

\[
i_n^b = F_\beta \left| 1/\left( (V_p/V_\beta)^2 - 1 \right)^{1/2} \right| \cos \left( \frac{2m-1}{n} \pi \right), \quad \text{(6b)}
\]

\(m = 1, 2, 3, \ldots n/2\) for \(n\) even or \((n+1)/2\) for \(n\) odd, where \(n\) is the total number of tuned circuits used.

The meaning of the letter \(m\) should be made clear by the following discussion. The complex roots of the response equation always occur in conjugate pairs, i.e., \((r+ji)\) and \((r-ji)\), and \(m\) is the pair number of the various pairs of roots. As plotted in the complex plane, these roots fall on a half circle whose center is \(f_0\) as shown in Fig. 8 for 4 resonant circuits. It will be seen that \(m=1\) gives the pair of roots whose real-frequency component is furthest from the midfrequency, \(m=2\) gives the pair of roots whose real-frequency components is next farthest from the midfrequency, etc. The maximum value that \(m\) can reach is that which makes the cos factor equal zero and simultaneously, of course, the sin factor equals unity. Thus \(m_{\text{max}}\) is \(n/2\) for an even number of resonant circuits and \((n+1)/2\) for an odd number of resonant circuits.

By multiplying out the correct number of terms of the above general equation (5), we can prepare a list of general-shape equations for \(n = 1, 2, 3, \ldots\), which are in exactly the form taken by the general-response equations of Design Equations—Group 1. We use exactly as many factors of (5) as there are tuned circuits \(n\). These resulting equations are given in Design Equations—Group 2. We can now compare Design Equations—

\[
V_p^b = \frac{1}{2\pi \Delta f_p (C_i C_n)^{1/2}} \left| (V_p/V_\beta)^2 - 1 \right|^{1/2} \left( \frac{K_{12}}{F_\beta} \right) \left( \frac{K_{22}}{F_\beta} \right) \left( \frac{K_{34}}{F_\beta} \right) \ldots \left( \frac{K_{(n-1)n}}{F_\beta} \right)
\]

Equation (8) alone is not of much use insofar as numerical gain calculations are concerned because the values of the \(K\)'s in (8) must first be determined from the solution of the simultaneous equations given in Design Equations—Group 3. The required values for coefficients of coupling as obtained from Design Equations—Group 3 will be found to be in terms of \(\Delta f_0 f_0\) and \([(V_p/V_\beta)^2 - 1]^{1/2}\) and, when the expressions for the required coefficients substituted in (8), a useful gain equation will result, which is in terms of \(C_i, C_n, \Delta f_\beta, (V_p/V_\beta)\) and the constant-current generator \(I\).

7.3 Resulting Phase Response of Amplitude-Response-Shape B

The exact phase-response shape associated with amplitude-response-shape \(B\) can, of course, be obtained from (5). We are neglecting the actual magnitude of the phase shift at the midfrequency, which from (2) is always plus or minus somemultiple of 90 degrees, depending on the number of inductive and capacitive couplings used. From (5), we see that \(\theta_n\), the phase shift of \((V_p/V_\beta)\) at any percentage bandwidth \(F\) is given by

\[
\theta_n^b = \tan^{-1} \left( \frac{F - i_n^b}{r_n^b} \right) + \tan^{-1} \left( \frac{F + i_n^b}{r_n^b} \right) + \tan^{-1} \left( \frac{F - i_n^b}{r_n^b} \right) + \tan^{-1} \left( \frac{F + i_n^b}{r_n^b} \right) + \ldots
\]

Groups 1 and 2 and equate the corresponding coefficients. Carrying out this procedure, we obtain the sets of simultaneous equations given in Design Equations—Group 3 which have to be solved to find the required circuit constants that will produce the \(B\) type of response shape. The simultaneous equations up to \(n = 4\) are listed, and the procedure for obtaining the simultaneous equations where \(r_n^b\) and \(i_n^b\) are given by (6a) and (6b). There will be exactly as many terms in (9) as there are resonant circuits in the network.

As an example of the use of (9), (6a), and (6b), we see that when a triple-tuned circuit is used to produce amplitude-response-shape \(B\), the resulting phase shift of \((V_p/V_\beta)\) at any \(\Delta f/\Delta f_\beta\) is given by
\[ \theta^b = \tan^{-1} \left\{ 2 \left( \frac{V_p}{V_g} \right)^2 - 1 \right\}^{1/6} \left( \frac{\Delta f}{\Delta f_0} \right) - 1.73 \}
\]
\[ + \tan^{-1} \left\{ 2 \left( \frac{V_p}{V_g} \right)^2 - 1 \right\}^{1/6} \left( \frac{\Delta f}{\Delta f_0} \right) + 1.73 \}
\]
\[ + \tan^{-1} \left\{ \left( \frac{V_p}{V_g} \right)^2 - 1 \right\}^{1/6} \left( \frac{\Delta f}{\Delta f_0} \right) \right\}. \tag{10} \]

In a similar way, the exact phase-shift equation may be written for any number of coupled circuits that are correctly used to obtain response-shape B.

7.4 Exact Design Equations for Response-Shape B
\((n = 1, 2, 3)\)

The design sheets given next in this paper for single-, double-, and triple-tuned circuits used to produce shape B were obtained by solving the first three sets of simultaneous equations given in Design Equations—Group 3 for the required circuit constants; substitution in (8) then gives the gain equations and substitution in (9) gives the phase-shift equation. In each case, the \(Q\) distribution chosen is the one that allows the designer to use the lowest possible \(Q\) value for the high-\(Q\) circuits of the network. Thus for the double-tuned network, the case of \(Q_1 = Q_2\) is considered. For the triple-tuned network, the \(Q\) distribution \(Q_1 = Q_2\) (or \(Q_1 = Q_3\)) is considered, and the coefficient of coupling distribution considered is \(K_{12} = K_{23}\). The reader should realize that the equations of Design Equations—Group 3 are perfectly general and any \(Q\) distribution and coefficient of coupling distributing can be investigated.

The problem of successfully solving the simultaneous equations of Design Equations—Group 3 in the cases where there are more than three resonant circuits per network, will be considered in another paper. It should be clearly realized that the coefficients of Design Equations—Group 2 give exact numerical values to which the general coefficients of Design Equations—Group 1 are equated; thus, even though it may be impossible to obtain closed-form design equations for the required circuit-element values, it may still be possible to obtain the numerical solutions of these simultaneous equations by some form of "try-and-try-again" method. Graphs or alignment charts can then be prepared from these numerical values and thus complete and exact design information can be satisfactorily presented to the engineer.

7.4.1 Exact Design Equations for \(N\)-Cascaded Triple-Tuned Circuits for Response-Shape B. (See Figs. 3, 4, and 7)

\[ Q_1 = Q_2 = \left( \frac{\Delta f_0}{\Delta f_0} \right)_{1.312} \left( \frac{V_p}{V_g} \right)^{2N - 1} \]
\[ Q_3 = \left( \frac{\Delta f_0}{\Delta f_0} \right)_{0.734} \left( \frac{V_p}{V_g} \right)^{2N - 1} \]
\[ K_{12} = K_{23} = \left( \frac{\Delta f_0}{\Delta f_0} \right)_{0.716} \left( \frac{V_p}{V_g} \right)^{2N - 1} \]

7.4.2 Exact Design Equations for \(N\)-Cascaded Double-Tuned Circuits for Response-Shape B (See Figs. 3, 4, and 7)

\[ Q_1 = Q_2 = \left( \frac{\Delta f_0}{\Delta f_0} \right)_{1.314} \left( \frac{V_p}{V_g} \right)^{2N - 1} \]
\[ K_{12} = \left( \frac{\Delta f_0}{\Delta f_0} \right)_{0.707} \frac{1}{\left( \frac{V_p}{V_g} \right)^{2N - 1}} \]

Gain per stage = \( \frac{G_m}{2\pi \Delta f_0 \left( \frac{\Delta f_0}{\Delta f_0} \right)} \) \( \frac{0.707 \left( \frac{V_p}{V_g} \right)^{2N - 1}}{\left( \frac{V_p}{V_g} \right)^{2N - 1}} \)

7.4.3 Exact Design Equations for \(N\)-Cascaded Single-Tuned Circuits for Response-Shape B

\[ Q = \left( \frac{\Delta f_0}{\Delta f_0} \right) \left( \frac{V_p}{V_g} \right)^{2N - 1} \]

Gain per stage = \( \frac{G_m}{2\pi \Delta f_0} \) \( \left( \frac{V_p}{V_g} \right)^{2N - 1} \)

\[ V_p/V = \left[ 1 + \left( \frac{V_p}{V_g} \right)^{2N - 1} \right] \left( \frac{\Delta f_0}{\Delta f_0} \right) \]

or

\[ \frac{\Delta f}{\Delta f_0} = \left( \frac{V_p}{V_g} \right)^{2N - 1} \]

\[ \theta = \tan^{-1} \left\{ \left( \frac{V_p}{V_g} \right)^{2N - 1} \left( \frac{\Delta f}{\Delta f_0} \right) \right\} \]
8. Response-Shape C

8.1 Desired-Shape Equation

When the straightforward procedure given in Section 3.1 is applied to single-, double-, and triple-tuned networks, the three resulting shape equations can be generalized for \( n \) resonant circuits into the following general form

\[
\frac{V_p}{V} = \left[ 1 + \frac{1}{(V_p/V_p)^2 - 1} \right]^{1/2} , \tag{11}
\]

where \( T_n \) is the Chebyshev polynomial of highest power \( n \) as given by

\[
T_n \left( \frac{F}{F_p} \right) = \begin{cases} 
\cos \left[ n \cos^{-1} \left( \frac{F}{F_p} \right) \right] , & \text{for } \left( \frac{F}{F_p} \right) < 1 \\
\cosh \left[ n \cosh^{-1} \left( \frac{F}{F_p} \right) \right] , & \text{for } \left( \frac{F}{F_p} \right) > 1.
\end{cases} \tag{13a}
\]

It is known that the above Chebyshev polynomial is also given by (13).

Thus we can write the shape equation for type C as

\[
\frac{V_p}{V} = \left[ 1 + \frac{1}{(V_p/V_p)^2 - 1} \right] \left[ \cos \left( n \cos^{-1} \left( \frac{F}{F_p} \right) \right) \right]^{1/2} \tag{14a}
\]

and

\[
\frac{V_p}{V} = \left[ 1 + \frac{1}{(V_p/V_p)^2 - 1} \right] \left[ \cosh \left( n \cosh^{-1} \left( \frac{F}{F_p} \right) \right) \right]^{1/2} \tag{14b}
\]

and, solving (14) for \( \Delta f/\Delta f_p \), we obtain the useful equations

\[
\frac{F}{F_p} = \cos \left[ \frac{1}{n} \cos^{-1} \left( \frac{V_p}{V_p} \right) \left( \frac{V_p}{V_p} \right)^2 - 1 \right]^{1/2} , \tag{15a}
\]

and

\[
\frac{F}{F_p} = \cosh \left[ \frac{1}{n} \cosh^{-1} \left( \frac{V_p}{V_p} \right) \left( \frac{V_p}{V_p} \right)^2 - 1 \right]^{1/2} , \tag{15b}
\]

where \( n \) is the total number of resonant circuits used in the filter chain.

Equation (15b) was used to obtain selectivity curves given in Fig. 2. The discussion following (4) and (4a) concerning the quantities \( F \) and \( \Delta f \) should now be re-read. The voltage ratios in (14) and (15) are the ratios for one network so that for \( N \)-cascaded identical networks the voltage ratios to be used in (15) are the \( N \)th root of the desired resultant voltage ratios.

8.2 Location of Peaks and Valleys Inside Pass Band

A little thought will show that we will obtain the location of the peaks of the response if, in (15a) we set \( V = V_p \); and we will obtain the locations of the valleys of the response if in (15a) we set \( V = V_p \).

Making the above substitutions, we obtain (16), which gives the location of the maxima of the response, and (17), which gives the location of the minima of the response,

\[
\frac{\Delta f_{\text{peaks}}}{\Delta f_p} = \cos \left( \frac{2m - 1}{n} \frac{\pi}{2} \right) , \tag{16}
\]

\[
\frac{\Delta f_{\text{valleys}}}{\Delta f_p} = \cos \left( \frac{2m}{n} \frac{\pi}{2} \right) . \tag{17}
\]

As before, \( m = 1, 2, 3 \cdots n/2 \) for \( n \) even, or \( (n+1)/2 \) for \( n \) odd. \( m = 1 \) gives the location of the pair of peaks and pair of valleys that are most distant from the mid-frequency; \( m = 2 \) gives the location of the pair of peaks and pair of valleys that are second most distant from the midfrequency, etc.

From (16), we note the interesting point that the ratio of the bandwidth between outside peaks \( \Delta f_p \) to the bandwidth \( \Delta f_p \) (where the skirt response equals the valley response) is 0.707 for double-tuned circuits; 0.866 for triple-tuned circuits; 0.922 for quadruple-tuned circuits, etc.
where the real \(r_m^e\) and imaginary \(i_m^e\) parts of the roots are given by (19a) and (19b).

\[
\begin{align*}
\mathbf{r}_m^e &= F_\beta [s_n] \sin \left( \frac{2m - 1 \pi}{n} \right), \\
\mathbf{i}_m^e &= F_\beta [c_n] \cos \left( \frac{2m - 1 \pi}{n} \right),
\end{align*}
\]  

(19a)

where

\[
\begin{align*}
s_n &= \sinh \left( \frac{1}{n} \sinh^{-1} \frac{1}{1/\left(1/r_1^p/1/r_2^p\right)^2 - 1} \right)^{1/2}, \\
c_n &= \cosh \left( \frac{1}{n} \sinh^{-1} \frac{1}{1/\left(1/r_1^p/1/r_2^p\right)^2 - 1} \right)^{1/2}
\end{align*}
\]  

(19b)

and \(m = 1, 2, 3, 4 \ldots n/2\) for \(n\) even, or \((n+1)/2\) for \(n\) odd.

Consideration of (19) shows that if plotted in the complex plane, the roots of the Chebyshev or Shape- \(C\) type of response fall on a half ellipse as shown in Fig. 9.

The meaning and use of \(m\) in (19) should be clear from Fig. 8 and from the paragraph in Section 7 that gives the meaning of \(m\) in connection with (6).

By “multiplying out” the correct number of terms of the above general equation (18), we can prepare a list of general shape equations for \(n = 1, 2, 3, \text{etc.},\) which are in exactly the form taken by general response equations of Design Equations—Group 1. These resulting equations are given in Design Equations—Group 4. (We use exactly as many factors of (18) as the number of resonant circuits in the networks.)

We can now compare Design Equations—Groups 4 and 1 and equate the corresponding coefficients. Carrying out this procedure, we obtain the sets of simultaneous equations given in Design Equations—Group 5, which have to be solved to obtain the exact required circuit constants \((Q\) and \(K))\) that will produce the \(C\) type of response shape. The simultaneous equations up to \(n = 4\) are listed, and the procedure for obtaining the simultaneous equations for any \(n\) should now be quite clear.

8.4 Gain Obtained with Response-Shape \(C\).

Comparison of the equation given in Design Equations—Group 1 and those given in Design Equations—Group 4 shows that for response-shape \(C\) the value of \(\Delta_n^e\) is

\[
\Delta_n^e = \frac{F_\beta^m}{2^{n - 1} \left(1/r_1^p/1/r_2^p\right)^2 - 1}^{1/2}
\]  

and using (2) and (1a), we see that the gain obtained at the voltage response peaks with response-shape \(C\) can be written as

\[
1^{1/2} \left( \begin{array}{c}
K_{12} \\
F_\beta \\
K_{23} \\
F_\beta \\
K_{34} \\
F_\beta \\
\end{array} \right) \ldots \left( \begin{array}{c}
K_{(n-1)a} \\
F_\beta \\
\right) 
\]  

(21)

Equation (21) alone does not enable us to make any numerical gain calculations, because the required values for the coefficients of coupling \(K\) must first be found from the solution of the simultaneous equations given in Design Equations—Group 5. It will be found that the required coefficients of coupling will be a function of the desired percentage bandwidth \(F_\beta\) and the ratio \((V_p/V_\beta)\), and when the expression for the required coefficient of coupling is substituted in (21) a useful gain equation results.

8.5 Resulting Phase Response of Amplitude-Response-Shape \(C\).

From (18), we can obtain the exact phase-response shape that is obtained when response-shape \(C\) is used. This will neglect the actual magnitude of the phase shift at the midfrequency, which from (2) is always plus or minus some multiple of 90 degrees, depending on the number of inductive and capacitive couplings used. From (18), we see that the phase shift \(\theta_n\) of \((1/r_1^p/1/r_2^p)\) at any percentage bandwidth \(F\) is given by

\[
\theta_n^e = \tan^{-1} \left( \frac{F - i_1^p}{r_1^p} \right) + \tan^{-1} \left( \frac{F^* + i_1^p}{r_1^{p*}} \right) + \tan^{-1} \left( \frac{F - i_2^p}{r_2^p} \right) + \cdots 
\]  

(22)

where \(r_m^e\) and \(i_m^e\) are given by (19). There will be exactly as many terms in (22) as there are resonant circuits in the network.

As an example of the use of (22) and (19), we see that when a triple-tuned circuit is used to produce amplitude-response-shape \(C\), the resulting phase shift of \((1/r_1^p/1/r_2^p)\) is given by
\[
\theta_d = \tan^{-1} \left[ \frac{2}{s_3} \left( \frac{\Delta f}{\Delta f_\beta} - 1.73 \frac{c_3}{s_3} \right) \right] + \tan^{-1} \left[ \frac{2}{s_3} \left( \frac{\Delta f}{\Delta f_\beta} + 1.73 \frac{c_3}{s_3} \right) \right] + \tan^{-1} \left[ \frac{1}{s_3} \frac{\Delta f}{\Delta f_\beta} \right].
\]

where \(s_3\) and \(c_3\) are given by (19).

In any exactly similar way, the exact phase-shift equation may be written for any number of resonant circuits when they are used to obtain response-shape C.

8.6 Exact Design Equations for Response-Shape C

The design sheets given next in this paper for single-, double-, and triple-tuned circuits used to produce response-shape C were obtained by solving the first three sets of simultaneous equations in Design Equations—Group 5 for the required circuit constants (with the \(Q\) distribution and \(K\) distribution given below). Single-tuned design is, of course, identical for both response-shapes B and C. Substitution in (21) gives the gain equation, and substitution in (22) gives the phase-shift equation.

\[
\theta_{\text{per stage}} = \tan^{-1} \left[ \frac{2}{s_3} \left( \frac{\Delta f}{\Delta f_\beta} - 0.87(1+s_2^2)^{1/2} \right) \right] + \tan^{-1} \left[ \frac{2}{s_3} \left( \frac{\Delta f}{\Delta f_\beta} + 0.87(1+s_2^2)^{1/2} \right) \right] + \tan^{-1} \left[ \frac{1}{s_3} \frac{\Delta f}{\Delta f_\beta} \right].
\]

In the case of the double-tuned circuit, the \(Q\) distribution considered is the one that allows the designer to use the lowest possible \(Q\), i.e., the distribution \(Q_1=Q_2\).

In the case of the triple-tuned circuit, the \(Q\) distribution considered here is the one that produced the simplest mathematical solution, i.e., \(Q_1=Q_3=Q\) and \(Q_2=\infty\); the \(K\) distribution considered is \(K_{12}=K_{23}\). This is not the most practical useful \(Q\) distribution because of the infinite \(Q\) required in the middle resonant circuit.17

The solution of the three simultaneous equations given in Design Equations—Group 5 for the triple-tuned circuit has been accomplished for the much more practical \(Q\) distribution of \(Q_1=Q_2=Q\) (or \(Q_2=Q_3=Q\)), and this solution will be presented in a later paper.

The problem of successfully solving the simultaneous equations of Design Equations—Group 5 for the cases where there are more than three resonant circuits per network will be considered in another paper. It should be clearly realized that the coefficients of Design Equations—Group 4 give us exact numerical values to which we are equating the general coefficients of Design Equations—Group 1. Thus, even though it may be impossible to obtain closed-form design equations for the required circuit-element values, it may still be possible to obtain the numerical solution of these simultaneous equations by some form of "try-and-try-again" method. Graphs or alignment charts can then be prepared from these numerical values, and complete exact design information can thus be satisfactorily presented to the engineer.

8.6.1 Exact Design Equations for N-Cascaded Triple-Tuned Circuits for Response-Shape C (See Figs. 3, 4, and 7)

Let

\[
s_3 = \sinh \left\{ \frac{1}{2} \sinh^{-1} \left[ \left. \frac{1}{(V_p/V_\beta)^{2/N} - 1} \right]^{1/2} \right] \right\}
\]

\[
Q_1=Q_2=\frac{f_0}{\Delta f_\beta} = 1.414 \frac{1}{s_2},
\]

\[
Q_2=\infty,
\]

\[
K_{12}=K_{23}=\frac{\Delta f_\beta}{f_0} = (0.375+0.5s_2^2)^{1/2}.
\]

\[
\text{Gain}_{\text{per stage}} = \frac{G_m}{2\pi\Delta f_\beta(C'C'\Pi)} (1.5+2s_3^2) \left[ \left( V_p/V_\beta \right)^{2/N} - 1 \right]^{1/2},
\]

\[
\frac{\Delta f}{\Delta f_\beta} = \cosh \left\{ \frac{1}{2} \cosh^{-1} \left[ \left. \left( V_p/V_\beta \right)^{2/N} - 1 \right]^{1/2} \right] \right\},
\]

outside pass band,

\[
\frac{\Delta f}{\Delta f_\beta} = \cos \left\{ \frac{1}{2} \cos^{-1} \left[ \left. \left( V_p/V_\beta \right)^{2/N} - 1 \right]^{1/2} \right] \right\},
\]

inside pass band.

8.6.2 Exact Design Equations for N-Cascaded Double-Tuned Stages for Response-Shape C (See Figs. 3, 4, and 7)

Let

\[
s_3 = \sinh \left\{ \frac{1}{2} \sinh^{-1} \left[ \left. \frac{1}{(V_p/V_\beta)^{2/N} - 1} \right]^{1/2} \right] \right\},
\]

\[
Q_1=Q_2=\frac{f_0}{\Delta f_\beta} = 1.414 \frac{1}{s_2},
\]

\[
Q_3 = \frac{f_0}{\Delta f_\beta} = 0.707(1+s_2^2)^{1/2},
\]

\[
\text{Gain}_{\text{per stage}} = \frac{G_m}{2\pi\Delta f_\beta(C'C'\Pi)} (1+s_2^2)^{1/2},
\]

\[
\Delta f = \cosh \left\{ \frac{1}{2} \cosh^{-1} \left[ \left. \left( V_p/V_\beta \right)^{2/N} - 1 \right]^{1/2} \right] \right\},
\]

outside pass band,

\[
\Delta f = \cos \left\{ \frac{1}{2} \cos^{-1} \left[ \left. \left( V_p/V_\beta \right)^{2/N} - 1 \right]^{1/2} \right] \right\},
\]

inside pass band.

\[
\theta_{\text{per stage}} = \tan^{-1} \left[ \frac{1.414}{s_2} \left( \frac{\Delta f}{\Delta f_\beta} - 0.707(1+s_2^2)^{1/2} \right) \right] + \tan^{-1} \left[ \frac{1.414}{s_2} \left( \frac{\Delta f}{\Delta f_\beta} + 0.707(1+s_2^2)^{1/2} \right) \right].
\]
9. Comments on Method of Design that Uses Complex Roots (or "Poles") of Network Response Equation

9.1 Mathematical Procedure

Although perhaps not stated from quite the viewpoint given below, there has recently been presented a method of design that also finds the complex roots of the desired amplitude-response-shape equation, and then finds the complex roots or poles of the circuit—response equations given in Design Equations—Group 1. The complex roots of the desired amplitude-response equation are then equated to the corresponding complex roots of the circuit—response equation to obtain the necessary simultaneous equations required for the solution for the unknown circuit constants.

Now, unfortunately, as we increase the number of resonant circuits in our n resonant-circuit band-pass network, the expressions for the complex roots or poles of the network in terms of the circuit constants $K$ and $Q$ become more and more complicated. To generalize, we can see from (2) that it will be necessary to solve an $n$th order equation to find the roots or poles of an $n$ resonant-circuit band-pass network. It is thus theoretically impossible to obtain general expressions for the poles of band-pass circuits employing more than 5 resonant circuits and, in practice, the general expression for the poles of even a triple-tuned circuit having three finite and different $Q$'s seems almost hopelessly complicated.

To demonstrate the above fact, the pole location $(r + ji)$ for single-, double-, and triple-tuned networks are given below in (23), (24), (25), i.e., these are the roots of the corresponding equations of Design Equations—Group 1.

\[ F_1 = -d \pm j0, \text{ single tuned.} \] (23)

\[ F_{12} = -\left(\frac{d_1 + d_2}{2}\right) \pm j\left[K_{12}^2 - \left(\frac{d_1 - d_2}{2}\right)^2\right]^{1/2} \text{, double tuned.} \] (24)

The roots of a triple-tuned circuit are

\[ P_{1,2} = -\left[\frac{1}{2}C_2 + \frac{1}{2}(\alpha + \beta)\right] \pm j\left[\frac{1}{2}\beta\left(\frac{1}{2}ight)^{1/2}\right], \] (25)

where

\[ \alpha = \left\{ -\left(\frac{q}{2}\right) + \left[\left(\frac{q}{2}\right)^2 + \left(\frac{1}{3}\right)^2\right]^{1/2}\right\}^{1/2}, \]

\[ \beta = \left\{ -\left(\frac{q}{2}\right) - \left[\left(\frac{q}{2}\right)^2 + \left(\frac{1}{3}\right)^2\right]^{1/2}\right\}^{1/2}, \]

where

\[ q = C_0 - \frac{1}{2}C_2 - \frac{1}{2}C_1, \]

\[ p = C_1 - \frac{1}{2}C_2. \]

where $C_0, C_1, C_2$ are given in Design Equations—Group 1.

When the above three roots of the triple-tuned-circuit response equation are simultaneously equated to the corresponding three roots of the desired triple-tuned-response equation, as obtained from (6) or (19), it is readily apparent that the solution of the resulting simultaneous equations for the required values of the circuit constants ($K$'s and $Q$'s) will indeed be a formidable task.

9.2 Stagger Tuning of Simple Interstage Circuits

The great practical importance of the pole type of design method must not be overlooked, however. When we consider the case of an over-all network consisting of many simple band-pass networks (i.e., single and double tuned) that are separated from each other by vacuum tubes, i.e., there is no coupling between the different simple circuits, then this design method is extremely useful. For this case, the expressions for each of the many poles retain the simplicity of (23) and (24), and it is a relatively simple matter to solve the resulting simultaneous equations for the required circuit constants. (It will be noted that the expression for the poles of a double-tuned circuit (24) becomes quite simple for the case of $Q = Q_0$).

9.2.1 Single-Tuned Interstage Circuits As Used to Obtain Response-Shape B (Small-Percentage Pass Band)

For example, let us briefly consider the case of stagger tuning of a single-tuned-interstage circuit. When obtaining the expressions for the poles of the networks, it is always important to consider the question of whether the resonant frequency of the networks used is identical with the midfrequency of the desired response shape. For the case of stagger tuning of single-tuned interstage circuits, it is of course obvious that this cannot be the case, and therefore it is necessary to express the equation for the pole of the network in terms of both the desired midfrequency of the response and the resonant frequency of the circuits.

A little thought will show that for the small-percentage pass-band case, the desired general expression for the pole of a single-tuned network which allows us mathematically to place the pole anywhere on the frequency axis, is given by (26)

\[ P_m = d_m \pm j(\Delta f_m/f_0). \] (26)

where the subscript $m$ has the same significance as in Section 7 and 8, $d$ is the decrement (i.e., reciprocal of $Q$) of the single-tuned circuit being considered, and $(\Delta f_m/f_0)$ is the percentage bandwidth (from the mid-frequency of the desired response) between the required pair of resonant frequencies of the $m$th pair of single-tuned interstages. (Equation (26) can be obtained from the usual simple expression for the transfer impedance of a single-tuned circuit by effectively changing the coordinate-system reference by simply expressing the resonant frequency of the resonant circuit as $f_r = f_0(1 \pm K/2)$, where $K$ is twice the percentage frequency difference between the desired midfrequency $f_0$.
and the circuit resonant frequency \( f_r \).

By equating the above circuit pole expression of (26) to the amplitude-response-shape-\( B \) pole expression of (6), we obtain the first two design equations of the design sheet for stagger tuning to produce response-shape \( B \).

The following equations give the desired-response-shape equation in three different forms. The final equations are the gain equations in two different forms. These gain equations are obtained by realizing that the total output voltage of the stagger-tuned chain at any frequency is given by simply multiplying together the responses of all the networks in the chain, thus obtaining (27).

\[
\sum_{i=1}^{n} \frac{1}{f_i} = \sum_{i=1}^{n} \frac{I_i}{\omega_0 C_i} = \sum_{i=1}^{n} \frac{I_i}{\omega_0 C_i} \prod_{j=1}^{d} \left[ jF \left( -d_1 \pm j \frac{\Delta f_1}{f_0} \right) \right] \prod_{j=1}^{d} \left[ jF \left( -d_2 \pm j \frac{\Delta f_2}{f_0} \right) \right] \cdot \cdot \cdot (27)
\]

The gain at the peak of the response is obtained when the magnitude of the denominator of (27) has its minimum value. We have already seen that for response-shape \( B \) this minimum value is given by (8). Thus, making use of (8), we obtain the total-gain equation; the \( n \)th root of this total-gain equation gives the equation for gain per stage as included in the design sheet.

9.2.2 Single-Tuned Interstage Circuits Used to Obtain Response-Shape \( C \) (Small-Percentage Pass Band)

By going through exactly the same line of reasoning used in Section 9.2.1, using the pole expressions of (19) for amplitude-response-shape \( C \), and the value of \( |\Delta s|_{\text{min}} \) for response-shape \( C \) given by (20), we obtain the equations given on the design sheet for stagger tuning of single-tuned interstage networks used to produce response-shape \( C \).

It should be realized that the voltage ratios given on the stagger-tuning design sheets are the ratios for a one-staggered \( n \)-tuple design, if \( N \) of these \( n \)-tuples are to be cascaded, then the voltage ratios to be used on the design sheets should be the \( N \)th root of the overall resulting desired voltage ratio.

In the interests of simplicity, a general many-termed phase-shift equation has not been included on the stagger-tuned design sheets. By referring to Sections 7.3 and 8.5, the reader should be able to write the correct phase-shift equation for any specific stagger-tuned design.

9.3 Stagger Tuning of Single-Tuned Interstage Circuits for Response-Shape \( B \) (See Fig. 10)

\[
\frac{1}{Q_m} = \frac{\Delta f_B}{f_0} \sin \left( \frac{2m-1}{n} 90^\circ \right),
\]

\[
(f_u - f_s)_{m} = \Delta f_B c_n \cos \left( \frac{2m-1}{n} 90^\circ \right).
\]

\[
V_p/V = \{1 + [(V_p/V_c)^2 - 1]|\Delta f/\Delta f_b|^{2n}\}^{1/2}
\]

where

\[
G_m = \text{geometric mean of all } G_m \text{'s},
\]

\[
C = \text{geometric mean capacitance}.
\]
For shape outside the pass band,

\[ \frac{V_p}{V_1} = 1 + \left[ \left( \frac{V_p}{V_0} \right)^2 - 1 \right] \]

\[ \cosh^2 \left[ n \cosh^{-1} \left( \frac{\Delta f}{\Delta f_0} \right) \right] \]

or

\[ \Delta f = \cosh \left[ \frac{1}{n} \cosh^{-1} \left[ \left( \frac{V_p}{V_0} \right)^2 - 1 \right] \right] \]

\[ n = \frac{\cosh^{-1} \left( \frac{\Delta f}{\Delta f_0} \right)}{\cosh^{-1} \left[ \left( \frac{V_p}{V_0} \right)^2 - 1 \right]} \]

For shape inside the pass band,

\[ \frac{V_p}{V_1} = 1 + \left[ \left( \frac{V_p}{V_0} \right)^2 - 1 \right] \]

\[ \cos^2 \left[ n \cos^{-1} \left( \frac{\Delta f}{\Delta f_0} \right) \right] \]

\[ \Delta f_{\text{max}} = \cos \left( \frac{2m - 1}{n} \theta_0^2 \right) \]

\[ \Delta f_{\text{min}} = \cos \left( \frac{2m}{n} \theta_0^2 \right) \]

\[ \text{Gain}_{\text{total}} = \frac{G_m}{2^\frac{1}{2} \pi \Delta f / \theta} \left[ \left( \frac{V_p}{V_0} \right)^2 - 1 \right]^{1/2} \]

\[ \log \left( \frac{2 \text{Gain}_{\text{total}}}{\pi \Delta f / \theta} \right) \]

\[ n = \frac{\log \left( \frac{G_m}{2^\frac{1}{2} \pi \Delta f / \theta} \right)}{\left[ \left( \frac{V_p}{V_0} \right)^2 - 1 \right]^{1/2}} \]

where

\[ G_m = \text{geometric mean of all } G_m \text{'s.} \]

\[ C = \text{geometric mean of all } C \text{'s.} \]

BIBLIOGRAPHY


Design Equations—Group 1

Exact Circuit-Response-Shape Equations in the Complex Polynomial Form for an n Resonant Circuit Network.

Single-Tuned Circuit

\[ \frac{V_p}{V_1} = \frac{1}{\left[ \Delta f_{\text{min}} \right]} \left[ \frac{1}{j \Delta f + j \theta} \right] \]

\[ \Delta f = d \theta \]

Double-Tuned Circuit

\[ \frac{V_p}{V_1} = \frac{1}{\left[ \Delta f_{\text{min}} \right]} \left[ \frac{1}{(j \Delta f) + B_1(j \theta + B_2)} \right] \]

\[ B_1 = d_1 + d_2 \]

\[ B_2 = K \theta^2 + d \theta \]
### Triple-Tuned Circuit

\[
V_p^* = \frac{1}{\Delta \min} [(jF)^3 + C_3(jF)^2 + C_4(jF) + C_5,]
\]

\[C_3 = d_1 + d_2 + d_3,\]

\[C_4 = K_{a d}^2 + d_3d_4 + d_3d_5 + d_4d_5.\]

\[C_5 = K_{a d}^2d_6 + K_{a d}d_7 + d_3d_4d_5d_6.\]

### Quadruple-Tuned Circuit

\[
V_p^* = \frac{1}{\Delta \min} [(jF)^4 + D_3(jF)^3 + D_4(jF)^2 + D_5(jF) + D_6].
\]

\[D_3 = d_1 + d_2 + d_3 + d_4,\]

\[D_4 = K_{a d}^2 + K_{a d}^2d_7 + K_{a d}d_8 + d_3d_5 + d_4d_5 + d_3d_6 + d_4d_6.\]

\[D_5 = K_{a d}^2d_6 + K_{a d}d_7 + K_{a d}^2d_9 + K_{a d}d_{10} + d_3d_7d_8 + d_4d_8d_9 + d_3d_9d_10 + d_4d_9d_10.\]

\[D_6 = K_{a d}^2K_{a d}^2 + K_{a d}^2d_7d_8 + K_{a d}^2d_7d_9 + K_{a d}^2d_8d_9 + d_3d_7d_9d_10 + d_4d_8d_9d_10.\]

### Design Equations—Group 2

**Single-Tuned Circuit**

\[
V_p^* = \frac{1}{F_0^2/[(V_p/V_0)^2 - 1]} [jF^* + A_1^*],
\]

\[A_1^* = \frac{1}{[(V_p/V_0)^2 - 1]} F_0.\]

**Double-Tuned Circuit**

\[
V_p^* = \frac{1}{F_0^2/[(V_p/V_0)^2 - 1]} [(jF)^2 + B_2^*(jF) + B_4^*],
\]

\[B_2^* = 1.414 \left[\frac{1}{[(V_p/V_0)^2 - 1]}\right] F_0.\]

\[B_4^* = \frac{1}{[(V_p/V_0)^2 - 1]} F_0^2.\]

**Triple-Tuned Circuit**

\[
V_p^* = \frac{1}{F_0^2/[(V_p/V_0)^2 - 1]} [(jF)^3 + C_3^*(jF)^2 + C_4^*(jF) + C_5^*].
\]

\[C_3^* = 2 \left[\frac{1}{[(V_p/V_0)^2 - 1]}\right] F_0.\]

\[C_4^* = 2 \left[\frac{1}{[(V_p/V_0)^2 - 1]}\right]^2 F_0^2.
\]

\[C_5^* = \frac{1}{[(V_p/V_0)^2 - 1]} F_0^3.\]

**Quadruple-Tuned Circuit**

\[
V_p^* = \frac{1}{F_0^2/[(V_p/V_0)^2 - 1]} [(jF)^4 + D_3^*(jF)^3 + D_4^*(jF)^2 + D_5^*(jF) + D_6^*].
\]

\[D_3^* = 2.64 \frac{1}{[(V_p/V_0)^2 - 1]} F_0^3.
\]

\[D_4^* = 3.41 \frac{1}{[(V_p/V_0)^2 - 1]} F_0^4.
\]

\[D_5^* = 2.64 \frac{1}{[(V_p/V_0)^2 - 1]} F_0^5.
\]

\[D_6^* = 1 \frac{1}{[(V_p/V_0)^2 - 1]} F_0^6.\]
Design Equations—Group 3

Exact Simultaneous Equations to be Solved for Circuit Constants (K's and Q's) of an n-Resonant-Circuit Network to Produce Response-Shape B when the Circuits are Correctly Resonated.

Single-Tuned Circuit

\[ d_1 = \frac{F_B}{\left[(V_p/V_B)^2 - 1\right]^{1/2}}. \]

Double-Tuned Circuit

\[ d_1 + d_2 = 1.414 \frac{F_B}{\left[(V_p/V_B)^2 - 1\right]^{1/4}}. \]

\[ K_{n+1} + d_1d_2 = \frac{F_B^2}{\left[(V_p/V_B)^2 - 1\right]^{1/4}}. \]

Triple-Tuned Circuit

\[ d_1 + d_2 + d_3 = 2 \frac{F_B}{\left[(V_p/V_B)^2 - 1\right]^{1/4}}. \]

\[ K_{n+2} + K_{n+3} + d_1d_2 + d_1d_3 + d_2d_3 = 2 \frac{F_B^2}{\left[(V_p/V_B)^2 - 1\right]^{1/4}}. \]

\[ K_{n+2}d_3 + K_{n+3}d_2 + d_1d_2d_3 = \frac{F_B^3}{\left[(V_p/V_B)^2 - 1\right]^{1/4}}. \]

Quadruple-Tuned Circuit

\[ d_1 + d_2 + d_3 + d_4 = 2.61 \frac{F_B}{\left[(V_p/V_B)^2 - 1\right]^{1/4}}. \]

\[ K_{n+2}d_4 + K_{n+3}d_3 + K_{n+4}d_2 + d_1d_2d_3 + d_1d_3d_4 + d_2d_3d_4 = 2.41 \frac{F_B^2}{\left[(V_p/V_B)^2 - 1\right]^{1/4}}. \]

\[ K_{n+2}K_{n+3}d_4 + K_{n+2}d_3d_4 + K_{n+3}d_2d_4 + K_{n+4}d_1d_2d_3 + d_1d_2d_3d_4 = \frac{F_B^3}{\left[(V_p/V_B)^2 - 1\right]^{1/4}}. \]

Design Equations—Group 4

Exact Complex Polynomials for Response-Shape C.

Single-Tuned Circuit

\[ V_p \quad V_p = \left( \frac{1}{F_B} \right) \left[ (V_p/V_B)^2 - 1 \right]^{1/2} \text{Re} \left( jF + 1 \varepsilon \right). \]

\[ A \varepsilon = \left( \frac{1}{F_B} \right) \left[ (V_p/V_B)^2 - 1 \right]^{1/2} F_B. \]

Double-Tuned Circuit

\[ V_p \quad V_p = \left( \frac{F_B^2}{2} \right) \left[ (V_p/V_B)^2 - 1 \right]^{1/2} \text{Re} \left( (jF)^2 + B \text{Re}(jF) + B \varepsilon \right). \]

\[ B \varepsilon = 1.414s_2F_B. \]

\[ B \varepsilon = 0.5(s_2^2 + c_2^2)F_B. \]

\[ s_2 = \sinh \left( \frac{1}{2} \sinh^{-1} \left( \frac{1}{\left[ (V_p/V_B)^2 - 1 \right]^{1/2}} \right) \right), \]

\[ c_2 = \cosh \left( \frac{1}{2} \sinh^{-1} \left( \frac{1}{\left[ (V_p/V_B)^2 - 1 \right]^{1/2}} \right) \right). \]

Triple-Tuned Circuit

\[ V_p \quad V_p = \left( \frac{F_B^3}{3} \right) \left[ (V_p/V_B)^2 - 1 \right]^{1/2} \text{Re} \left( (jF)^3 + C \text{Re}(jF)^3 + C \text{Re}(jF) + C \varepsilon \right). \]

\[ C \varepsilon = 2s_2F_B. \]

\[ C \varepsilon = (1.25s_2^2 + 0.75c_2^2)F_B. \]

\[ C \varepsilon = s_2(0.25s_2^4 + 0.75c_2^4)F_B. \]
\[ s_5 = \sinh \left( \frac{1}{3} \sinh^{-1} \left( \frac{1}{(V_p/V_d)^3 - 1} \right) \right) \]

\[ c_5 = \cosh \left( \frac{1}{3} \sinh^{-1} \left( \frac{1}{(V_p/V_d)^3 - 1} \right) \right) \]

**Quadruple-Tuned Circuit**

\[ V_p = \frac{1}{F_{sd}^4 \left[ (V_p/V_d)^3 - 1 \right]^{1/2}} \left[ (jF)^4 + D_F(jF)^3 + D_F(jF)^2 + D_F(jF) + D_F \right] \]

\[ D_F = 2.61T_k^2 \]
\[ D_F = (2.41T_k + c)^2F_k^2 \]
\[ D_F = s_5(0.923T_k + 1.06c)^2F_k^2 \]
\[ D_F = 0.125(s_5^4 + 6s_5c^3 + c^4)F_k^2 \]

\[ s_6 = \sinh \left( \frac{1}{4} \sinh^{-1} \left( \frac{1}{(V_p/V_d)^3 - 1} \right) \right) \]
\[ c_6 = \cosh \left( \frac{1}{4} \sinh^{-1} \left( \frac{1}{(V_p/V_d)^3 - 1} \right) \right) \]

**Design Equations—Group 5**

Exact Simultaneous Equations to be solved for Circuit Constants \((K' s \ and \ Q' s)\) of an \(n\)-Resonant-Circuit Network to Produce Response-Shape \(C\) when the Circuits are Correctly Resonated.

**Single-Tuned Circuit**

\[ d_1 = \left( \frac{F_k}{(V_p/V_d)^3 - 1} \right)^{1/2} \]

**Double-Tuned Circuit**

\[ d_1 + d_2 = 1.414T_k F_k \]
\[ K_{12}d_1 + d_3d_4 = (0.5 + c T_k)F_k^2 \]

**Triple-Tuned Circuit**

\[ d_1 + d_2 + d_3 = 2s_5F_k \]
\[ K_{12}d_3 + d_4d_5 + d_5d_6 + d_5d_4 = (0.125 + s_5T_k)^2F_k^3 \]

**Quadruple-Tuned Circuit**

\[ d_1 + d_2 + d_3 + d_4 = 2.61T_k F_k \]
\[ K_{12}d_3 + K_{12}d_4 + d_4d_5 + d_5d_4 + d_5d_5 + d_5d_6 + d_5d_4 = (1 + 3.41T_k)F_k^3 \]
\[ K_{12}d_4 + K_{12}d_4 + K_{12}d_5 + K_{12}d_5 + K_{12}d_5 + d_5d_6 + d_5d_5 + d_5d_6 + d_5d_5 = s_5(1.69 + 2.61T_k)F_k^4 \]
\[ K_{12}d_5 + K_{12}d_5 + K_{12}d_6 + K_{12}d_6 + d_6d_5 + d_6d_6 = (0.125 + s_5T_k)^2F_k^3 \]

In all the above equations, \( s_5 = \sinh \left( \frac{1}{4} \sinh^{-1} \left[ (V_p/V_d)^3 - 1 \right] \right) \).

**CORRECTION**

The authors have brought to the attention of the Editor the following error in the paper "Considerations in the Design of a Radar Intermediate-Frequency Amplifier," by Andrew L. Hopper and Stewart E. Miller, which appeared on pages 1208-1220 of the November, 1947, issue of the PROCEEDINGS OF THE I.R.E.

On page 1215, the expression \( V_o = \sqrt{BKT \Delta f R_s} \) should read \( V_o = \sqrt{4BKT \Delta f R_s} \).1

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\[ \frac{2}{\text{amp}} = 4BT F \Delta f R_s \]

Since

\[ V_o = i_o R_e \quad \text{and} \quad g_e = \frac{1}{R_e} \]

\[ V_o = \sqrt{4BKT \Delta f R_e} \]
Cathode Neutralization of Video Amplifiers

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Summary—The usual cathode bypass capacitors are eliminated, and replaced by a resistor connected from cathode to cathode of succeeding stages. It is shown that no gain need be sacrificed, and a great reduction in low-frequency phase shift is obtained. The addition of a small capacitance across a portion of the intercathode resistance gives an improvement in high frequency response and phase shift. Gain and stability equations are derived, and a circuit diagram of a practical amplifier is given.

In video amplifier stages having tubes of very high transconductance, it is customary to use cathode bias resistors in order to minimize variations in plate current that might be caused by variations in electrode operating potentials, replacement or aging of tubes, etc. A cathode bypass capacitor is normally used in each stage to avoid loss in amplification due to degeneration. However, even with the largest practical values of capacitance, there is usually an appreciable amount of phase shift at very low frequencies.

The cathode bypass is sometimes omitted when a very small value of cathode resistance can be used. This arrangement may be moderately successful in an early stage where the signal is so small that little bias is needed to prevent the flow of grid current. However, the dc stability is not nearly so good as that which is obtained with a large cathode resistance.

In the circuit described here, the cathode bypass capacitors are omitted, and the resulting cathode degeneration is effectively eliminated or greatly reduced by the use of a neutralizing resistor connected between the cathodes of succeeding stages.

The circuit of such an amplifier is shown in Fig. 1. The high-frequency compensation is not shown. The circuit is otherwise normal, except for the resistor $R_{bl}$ connected between the cathodes of the tubes, and the capacitor $C$, which serves merely to improve the performance at high frequencies. The effect of $C$ will be neglected in the following expressions for the gain of the amplifier.

Also, the plate load resistance of each tube will be assumed to be negligible compared with the internal plate resistance of the tube; a condition that normally exists in pentode video amplifiers.

For convenience, $R_L$ has been included in $R_1$ in deriving the gain equations, and $R_2$ has been included in $R_L$. It has also been assumed that there is no voltage drop across the plate and screen grid supply bypass capacitors or across the grid blocking capacitors. While this condition does not usually hold completely true in practice, the gain equations would otherwise become very unwieldy, not only because of the complex circuit meshes that result, but because the equations would include signal frequency as a function. As a practical matter, the value of $R_{bl}$ must be determined for proper operation at medium frequencies, unless a complex network is substituted for $R_{bl}$ that would correct for the attenuation and phase shift in the bypass and grid blocking capacitors.

At medium frequencies, the gain of the amplifier is

$$\text{Gain} = g_m g_m R_2 \left[ R_{bl} + R_{11} + R_{12} + g_m R_{12} R_{bl} (R_{11} + R_{12}) + g_m R_{11} (R_{11} + R_{12}) + g_m g_m R_{12} R_{bl} (R_{bl} - R_1) \right] \quad (1)$$

When $R_{bl}$ and $R_{12}$ equal zero, a situation that is equivalent to a conventional amplifier circuit having perfectly bypassed cathodes, (1) reduces to

$$\text{Gain} = g_m g_m R_2 R_1 \quad (2)$$

In order to obtain the same gain with unbypassed cathode resistors, $R_{bl}$ must have the value

$$R_{bl} = \frac{R_1 R_{12} (1/R_1 - g_m - g_m^2 + g_m g_m R_{12})}{g_m R_{11} + g_m^2 R_{12} + g_m g_m R_{12} R_{12}} \quad (3)$$

The value of $R_{bl}$ for oscillation, or infinite gain, is

* Decimal classification: R363.4. Original manuscript received by the Institute, August 2, 1948; revised manuscript received, April 20, 1949.

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The derivation of equations (1) through (4) is given in the Appendix.

It is of some interest to determine how close together the values of \( R_{40} \) for oscillation and for normal gain may come in a typical case. Let us assume that \( g_{m1} \) and \( g_{m2} \) are 0.009 mho; \( R_{k1} \) and \( R_{k2} \) are 160 ohms, and \( R_1 \) equals 0.000 ohms. These are values that might exist in an amplifier using 6AC7 tubes. From (3), the value of \( R_{40} \) for normal gain is 331 ohms, and from (4), the value for oscillation is 215 ohms. It is apparent, therefore, that a reasonably close tolerance should be placed on \( R_{40} \) in the manufacture of amplifiers of this type. The cathode bias resistors need not be held to such close tolerances, and normal variations in tube transconductances are not troublesome. However, resistor \( R_1 \) must be held to a fairly close tolerance.

Since the cathode of the first tube is at an ac potential that is 180 degrees out of phase with its grid voltage, there will be some increase in the input capacity of the amplifier, accompanied by a decrease in the input capacitance of the second stage. If desired, these effects can be corrected by shunting cathode resistors \( R_{k1} \) and \( R_{k2} \) with small capacitors of such value that the amplifier will function in the manner of conventional amplifiers with bypassed cathodes at high frequencies, but tube internal plate resistance, the output impedance of the amplifier will be reduced, as the system consists essentially of positive current feedback.

If desired, \( R_{k1} \), \( R_{k2} \), and \( R_{40} \) may be rearranged to form a wye rather than a delta, with \( C \) connected in shunt with either of the ungrounded cathode resistors.

Fig. 2 shows the circuit diagram of a portion of a practical amplifier that uses cathode neutralization. \( C_5 \) yields a considerable increase in gain and reduction in phase shift, for frequencies above 3 mc per second.

As the screen grids of the tubes in Fig. 2 are bypassed to the low potential side of the cathode resistors instead of directly to the cathodes, the screen grid alternating currents will flow through the cathode resistors, so that the expressions for gain given previously will apply only approximately

As a practical consideration, amplifiers incorporating the circuit of Fig. 2 that were manufactured in production runs were very uniform in gain and frequency response without any individual selection of parts or tubes. Both low- and high-frequency response were greatly improved over previous designs.

The cathode neutralization principle could also be used in some audio amplifier applications. The value of \( R_{40} \) chosen would be somewhat modified by the fact that

\[
R_{40} = \frac{g_{m1}g_{m2}R_{k1}R_{k2}R_1}{(1 + g_{m1}R_{k1})(1 + g_{m2}R_{k2})} - (R_{k1} + R_{k2} + g_{m2}R_{k1}R_{k2} + g_{m1}R_{k1}R_{k2})
\]

\[
(1 + g_{m1}R_{k1})(1 + g_{m2}R_{k2})
\]

in most audio amplifiers the tube plate load resistance is not negligible, compared with the tube internal plate resistance. Also, heater-cathode leakage might create a hum problem in the early stages of high gain amplifiers.

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Footnotes:

1. Millr: Neutralization of Video Amplifiers


3. See Sec. 5, Par. 11, p. 402 of footnote reference 1.

APPENDIX

The four simultaneous equations for the operation of the circuit of Fig. 1 are:

\[ i_1 = [e_1 + K_{kl}(i_3 - i_1)]g_{m1} \]  
(5)

\[ i_2 = [i_1R_1 - (i_2 - i_3)R_{k2}] g_{m2} \]  
(6)

\[ K_{kl}i_3 + K_{k2}(i_3 - i_1) = K_{k2}(i_2 - i_3). \]  
(7)

Gain = \( i_2R_2/e_1 \)  
(8)

where \( g_{m1} \) and \( g_{m2} \) are the transconductances of 1'-1 and 1'-2.

Solving (5) for \( i_3 \) we obtain

\[ i_3 = \frac{-e_1 + R_{kl}i_1 + i_1}{g_{m1}}. \]  
(9)

Dividing both sides of (6) by \( g_{m2} \), and substituting (9),

\[ \frac{i_2}{g_{m2}} = i_1R_1 - i_2R_{k2} + i_2R_{k2} \]

\[ = i_1R_1 - i_2R_{k2} + \frac{R_{k2}}{R_{k1}} \left( -e_1 + R_{kl}i_1 + \frac{i_1}{g_{m1}} \right), \]

or

\[ i_2 \left( R_{k2} + \frac{1}{g_{m2}} \right) = i_1 \left( R_1 + R_{k2} + \frac{R_{k2}}{R_{k1}g_{m1}} \right) - \frac{R_{k2}}{R_{k1}} e_1. \]

whence

\[ i_1 = \frac{i_2 \left( R_{k2} + \frac{1}{g_{m2}} \right) + \frac{R_{k2}}{R_{k1}} e_1}{R_1 + R_{k2} + \frac{R_{k2}}{R_{k1}g_{m1}}}. \]  
(10)

Solving (7) for \( i_3 \), and substituting (9),

\[ i_2 = \frac{R_{k1}i_1 + R_{k2}i_2}{R_{k0} + R_{k1} + R_{k2}} = \frac{-e_1 + R_{kl}i_1 + \frac{i_1}{g_{m1}}}{R_{k1}} \]

Gain = \( R_2 \)

Multiplying out numerator and denominator, cancelling equal terms of opposite signs, and multiplying numerator and denominator by \( g_{m1} g_{m2} \), we obtain equation (1) of the main text:

Solving (11) for \( i_1 \), and substituting (10),

\[ i_1 = \frac{R_{k1}R_{k2}i_3 + e_1(R_{k0} + R_{k1} + R_{k2})}{\left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2} \]

\[ + \frac{i_2 \left( R_{k2} + \frac{1}{g_{m2}} \right) + \frac{R_{k2}}{R_{k1}} e_1}{\left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2} \]

\[ + \frac{R_{k2}}{R_{k1}} \left( R_{k0} + R_{k1} + R_{k2} \right) \]

\[ \left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2 \]

Transposing, and factoring out \( i_2 \),

\[ i_2 \left[ \left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2 \right] \]

\[ = \frac{R_{k1}R_{k2}i_3 + e_1(R_{k0} + R_{k1} + R_{k2})}{\left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2} \]

\[ + \frac{i_2 \left( R_{k2} + \frac{1}{g_{m2}} \right) + \frac{R_{k2}}{R_{k1}} e_1}{\left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2} \]

\[ + \frac{R_{k2}}{R_{k1}} \left( R_{k0} + R_{k1} + R_{k2} \right) \]

\[ \left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2 \]

Cross-multiplying,

\[ i_2 \left\{ R_{k1}R_{k2} \left( R_1 + R_{k2} + \frac{R_{k2}}{R_{k1}g_{m1}} \right) - \left( R_{k2} + \frac{1}{g_{m1}} \right) \right\} \]

\[ - \left( R_{k0} + R_{k1} + R_{k2} \right) \left( R_1 + R_{k2} + \frac{R_{k2}}{R_{k1}g_{m1}} \right) - \left( R_{k2} + \frac{1}{g_{m1}} \right) \]

\[ \left[ \left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2 \right] \]

\[ = e_1 \left\{ - (R_{k0} + R_{k1} + R_{k2}) \left( R_1 + R_{k2} + \frac{R_{k2}}{R_{k1}g_{m1}} \right) \right. \]

\[ + \frac{R_{k2}}{R_{k1}} \left[ \left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2 \right] \].  
(12)

Substituting (12) into (8), we obtain

\[ \frac{R_{k1}R_{k2} \left( R_1 + R_{k2} + \frac{R_{k2}}{R_{k1}g_{m1}} \right) - \left( R_{k2} + \frac{1}{g_{m1}} \right) \left[ \left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2 \right] \}

\[ + \frac{R_{k2}}{R_{k1}} \left[ \left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}) - R_{k1}i_1^2 \right] \]

Cross-multiplying,

\[ R_{k1}R_{k2}i_1 + R_{k2}i_2 = -e_1(R_{k0} + R_{k1} + R_{k2}) \]

\[ + i_1 \left( R_{k1} + \frac{1}{g_{m1}} \right)(R_{k0} + R_{k1} + R_{k2}). \]  
(11)
Gain = $g_{m1}g_{m2}R_2 \left[ \frac{R_i(R_k0 + R_k1 + R_k2) + R_k1R_k2}{R_k0 + R_k1 + R_k2 + g_m2R_k2(R_k0 + R_k1) + g_m1R_k1(R_k0 + R_k2) + g_m1g_m2R_k1R_k2(R_k0 - R_i)} \right].$ (1)

In order for the gain in (1) to be equal to $g_{m1}g_{m2}R_1R_2$, the fraction inside the brackets must be equal to $R_i$. Dividing both sides of this identity by $R_i$, cross-multiplying, and solving for $R_k0$, we obtain (3) of the main ext:

$$R_k0 = \frac{g_{m1}g_{m2}R_k1R_k2R_i}{(1 + g_m1R_k1)(1 + g_m2R_k2)}$$ (4)

A New Figure of Merit for the Transient Response of Video Amplifiers

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Summary—A figure of merit suitable for comparing the transient response of television video amplifiers is proposed. The parameters are adjusted so that when applied to a shunt-peaked interstage, the figure of merit reaches a maximum for an overshoot of approximately 2 per cent. Application of the suggested form to other types of interstages arranges the various networks in the order of their suitability in television amplifiers as considered from their transient responses.

The study of amplifiers for video applications has, in the past, been considered from the viewpoints of both steady-state and transient response. In the process of developing wide-band amplifiers on a steady-state basis, it has been assumed that the desired goal is to achieve a flat amplitude characteristic for the widest band of frequencies. It has been further assumed that the accompanying phase-shift characteristic be linear within limits which will make the overall amplifier useful for the particular application. Although the experimental verification of a given amplitude characteristic is not difficult, the determination of the accompanying phase characteristic often presents a considerable problem. For minimum phase-shift networks, and these are the types which will be considered here, Bode1 has shown that the phase characteristic is uniquely defined by the amplitude characteristic; however, this determination, while straightforward, is nonetheless involved.

A figure of merit for comparing the relative utility of wide-band amplifiers on a steady-state basis has been generally accepted as the gain-bandwidth product. By considering the interstages alone, normalizing these to a reference midfrequency impedance level, the steady-state figure of merit degenerates to a comparison of the upper frequency limit at which the impedance of the network falls to some specified value. This cutoff frequency may be considered as that at which the interstage network impedance has decreased to 70.7 per cent of its midfrequency value, or alternatively, as that frequency which, when multiplied by the midfrequency impedance, gives the same product as the area contained under the impedance versus frequency curve. This latter value may be more useful for mathematical manipulation, as considered by Hansen2 and DiToro.3

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In video applications, the fidelity of pulse reproduction is, in reality, the final criterion by which we seek to judge the utility of various interstages. Since the steady-state response only indirectly measures the fidelity of pulse reproduction, it is but an intermediate step in the evaluation of the transient response. In particular, since the steady-state figure of merit does not take cognizance of the variation of the time-delay characteristic near the cutoff frequency, it is of relatively little use in transient work.

The purpose of this paper is to analyze video interstages from a transient viewpoint solely, and to propose a figure of merit by which the relative utility of an amplifier for transient applications can be specified. Although the criterion that is proposed does not rest upon a coherent mathematical development, it, nevertheless, does consider for the first time those factors which, from both an experimental and theoretical viewpoint, must be dealt with in order to make a fair appraisal.

Some definitions are in order before proceeding further. Assume the input signal takes the form of a unit step, which is the integral of the sometimes used $\Delta$ function. An ideal amplifier from a transient standpoint will reproduce such a wave form exactly, and a typical amplifier which fails to do this will, in general, reproduce the initial step with a function that has a finite slope and may also have an “overshoot” or “undershoot” which, in a simple case, will take the form of a damped sinusoid. The steepness of the maximum slope will be a measure of the rise time, and this parameter has been previously defined in several different ways. One must decide whether the rise time should be specified in a manner which is of practical value, or in a manner which lends itself to easy mathematical manipulation. Although we penalize ourselves somewhat for doing so, the former method is elected here. Kallman, Spencer, and Singer have suggested that this parameter be defined in terms of the time interval within which the output amplitude wave form traverses the 10 to 90 per cent points. This has been found, in practice, to be desirable inasmuch as one is seldom concerned with the delay which may take place from the origin to the 10 per cent value, and provided the overshoot is not excessive and the wave form follows a normal shape, the 90 per cent value represents a convenient point at which the wave form excursion may be said to have been completed. If the interstage is normalized with respect to impedance level, the absolute magnitude of the maximum variation of the output wave form from its end value of unity will be defined as the overshoot. These two parameters will be referred to as the rise time $\tau$ and the overshoot $\gamma$, and are shown in Fig. 1. The

![Fig. 1—Typical transient response.](image)

Laplace transformation has been used in analyzing the interstage networks to be described here, the method following those of Gardner and Barnes. In carrying out such an analysis, it is found that the rise time varies inversely with both $R$ and $C$, so that it is convenient to conduct one analysis for each network configuration and specify the rise time in $t/RC$ units. The value of shunt capacitance $C$, in the case of two-terminal networks, will consist of the total distributed interstage capacitance; in the case of four-terminal networks, the convention that has been followed has been to consider $C$ as the sum of the input and output capacitances of the network. This then permits the use of recurrent filter networks without penalizing the ultimately derived figure of merit for the midpoint shunt capacitances of the system.

In comparing the performance of any interstage, one is concerned with the quantities absolute rise time, overshoot, and gain (or impedance level). In order to make the figure of merit independent of impedance level, the relative rise time $\tau$ must be introduced as a reciprocal function. It is considered important that the figure of merit also be some inverse function of the overshoot $\gamma$. The amount that the figure of merit should be penalized for increasing values of overshoot rests upon two foundations. If the figure of merit relates only to single interstages, it should be so arranged that the value of the figure of merit increases with decreasing rise time, up to a point where the overshoot becomes so large as to be objectionable from a practical standpoint. For values of overshoot greater than this value, the figure of merit should decrease rapidly. In television applications, values of overshoot greater than 2 per cent may lead to a perceptible and possibly objectionable ringing in the reproduced picture in the case of sharp black and white transitions. Consequently, such a value of overshoot should be given some weight in determining

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1 That is, excluding any delay which merely translates the wave form along the time axis.


what point the figure of merit will begin to become
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value of $\gamma$, however, occurs in relation to the use of
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ted for three different values of $\gamma$ of a single such
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the overshoot has a value equal to or less than ap-
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with the relationship given above, and the overshoot
remains substantially constant. However, for values of
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decrease sharply inasmuch as such a configuration is
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It has seemed expedient, therefore, to derive the figure
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tween rise time and $m$ is given graphically, and we can
see that as $m$ is increased from a value of 0 (which cor-
responds to a simple $RC$ interstage), the value of $r$
decreases initially at approximately a linear rate. Also,
in Fig. 3 is shown the relationship between $\gamma$ and $m$
for the same network. One observes that for a value of
$m < 0.25$ there is no overshoot; and as $m$ increases
beyond this value, the overshoot increases, eventually
becoming approximately a linear variation. It is found
that a value of $m = 0.388$ will give an overshoot of ap-
proximately 2 per cent, and based upon the data of
Bedford and Fredendall for recurrent shunt-peaked
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for the figure of merit, $F$, is as follows

$$F = a \tau^{-1} e^{-b \gamma}$$

where

- $F$ = a numeric specifying the relative quality of a
  shunt-peaked interstage from the transient view-
  point
- $a$ = a constant to give a suitable magnitude to $F$
- $\tau$ = the rise time in units of time normalized by the
  factor $RC$
- $\gamma$ = the fractional overshoot

$W. C. Elmore, "The transient response of damped linear net-
works with particular regard to wideband amplifiers," Jour. Appl.

A. V. Bedford and G. L. Fredendall, "Transient response of
multi-stage video-frequency amplifiers," Proc. I.R.E., vol. 27,
pp. 77-284; April, 1939.
APPENDIX

The four simultaneous equations for the operation of the circuit of Fig. 1 are:

\[ i_1 = \left[ e_1 + R_{i1}(i_3 - i_1) \right] g_{m1} \]  
\[ i_2 = \left[ i_1 R_1 - (i_2 - i_3) R_{k2} \right] g_{m2} \]  
\[ R_{k0} i_3 + R_{k1}(i_3 - i_1) = R_{k2}(i_2 - i_2), \]  
\[ \text{Gain} = i_2 R_2 / e_1 \]

where \( g_{m1} \) and \( g_{m2} \) are the transconductances of \( V - 1 \) and \( V - 2 \).

Solving (5) for \( i_3 \) we obtain

\[ i_3 = \frac{-e_1 + R_{k1} i_1 + \frac{i_1}{g_{m1}}}{R_{k1}}. \]  

Dividing both sides of (6) by \( g_{m2} \) and substituting (9),

\[ \frac{i_2}{g_{m2}} = i_1 R_1 - i_2 R_{k2} + i_2 R_{k2} \]

\[ = i_1 R_1 - i_2 R_{k2} + R_{k2} \left( \frac{-e_1 + R_{k1} i_1 + \frac{i_1}{g_{m1}}}{R_{k1}} \right), \]

or

\[ \frac{i_2}{R_{k1}} \left( R_{k2} + 1 \right) = i_1 \left( R_1 + R_{k2} + \frac{R_{k2}}{R_{k1} g_{m1}} \right) - \frac{R_{k2}}{R_{k1}} e_1, \]

whence

\[ i_2 = \frac{i_1 R_1 - i_2 R_{k2} + R_{k2}}{R_{k1}} \left( e_1 + R_{k1} i_1 + \frac{i_1}{g_{m1}} \right). \]  

Solving (7) for \( i_3 \), and substituting (9),

\[ i_3 = \frac{R_{k0} i_3 + R_{k2} i_2}{R_{k0} + R_{k1} + R_{k2}} = \frac{-e_1 + R_{k1} i_1 + \frac{i_1}{g_{m1}}}{R_{k1}}. \]

Solving (11) for \( i_1 \), and substituting (10),

\[ i_1 = \frac{R_{k1} R_{k2} i_2 + e_1 (R_{k0} + R_{k1} + R_{k2})}{\left( R_{k1} + \frac{1}{g_{m1}} \right) (R_{k0} + R_{k1} + R_{k2}) - R_{k1}^2} \]

\[ i_2 \left( R_{k1} + \frac{1}{g_{m1}} \right) (R_{k0} + R_{k1} + R_{k2}) - R_{k1}^2 \]

Transposing, and factoring out \( i_2 \),

\[ \frac{i_2}{R_{k1}} \left( R_{k2} + 1 \right) = \frac{1}{R_{k1} g_{m1}} \left( R_{k0} + R_{k1} + R_{k2} \right) - R_{k1} \]

\[ + e_1 \left( R_{k2} \right) \frac{g_{m1}}{R_{k1} g_{m1}} \]

whence

\[ i_2 = \frac{i_1 R_1 - i_2 R_{k2} + R_{k2}}{R_{k1}} \left( e_1 + R_{k1} i_1 + \frac{i_1}{g_{m1}} \right). \]

Solving (7) for \( i_3 \), and substituting (9),

\[ i_3 = \frac{R_{k0} i_3 + R_{k2} i_2}{R_{k0} + R_{k1} + R_{k2}} \]

\[ = \frac{-e_1 + R_{k1} i_1 + \frac{i_1}{g_{m1}}}{R_{k1}}. \]

Gain = \( R_2 \)

\[ R_{k1} R_{k2} \left( R_1 + R_{k2} + \frac{R_{k2}}{R_{k1} g_{m1}} \right) - \left( R_{k2} + \frac{1}{g_{m2}} \right) \]

Cross-multiplying,

\[ i_2 \left( R_{k1} R_{k2} \left( R_1 + R_{k2} + \frac{R_{k2}}{R_{k1} g_{m1}} \right) - \left( R_{k2} + \frac{1}{g_{m2}} \right) \right) \]

\[ - \left( R_{k1} + R_{k1} + R_{k2} \right) \left( R_1 + R_{k2} + \frac{R_{k2}}{R_{k1} g_{m1}} \right) + R_{k2} \left[ \left( R_{k1} + \frac{1}{g_{m1}} \right) \left( R_{k0} + R_{k1} + R_{k2} \right) - R_{k1}^2 \right] \]

Multiplying out numerator and denominator, cancelling equal terms of opposite signs, and multiplying numerator and denominator by \( g_{m1} \) \( g_{m2} \), we obtain equation (1) of the main text:
Gain = $g_m g_2 R_2 \left[ \frac{R_1 (R_{k_0} + R_{k_1} + R_{k_2}) + R_{k_1} R_{k_2}}{R_{k_0} + R_{k_1} + R_{k_2} + g_m R_{k_2} (R_{k_0} + R_{k_2}) + g_m R_{k_2} (R_{k_0} + R_{k_2}) + g_m g_2 R_{k_1} R_{k_2} (R_{k_0} - R_{k_2})} \right]. \quad (1)$

In order for the gain in (1) to be equal to $g_m g_2 R_{k_1} R_{k_2}$, the fraction inside the brackets must be equal to $R_1$. Dividing both sides of this identity by $R_1$, cross-multiplying, and solving for $R_{k_0}$, we obtain (3) of the main text:

$$K_{k_0} = \frac{g_m g_2 R_{k_1} R_{k_2} R_1 - (R_{k_1} + R_{k_2} + g_m R_{k_2} (R_{k_0} + R_{k_2}) + g_m R_{k_2} (R_{k_0} + R_{k_2}))}{(1 + g_m R_{k_1}) (1 + g_m R_{k_2})}. \quad (4)$$

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**A New Figure of Merit for the Transient Response of Video Amplifiers**

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Summary—A figure of merit suitable for comparing the transient response of television video amplifiers is proposed. The parameters are adjusted so that when applied to a shunt-peaked interstage, the figure of merit reaches a maximum for an overshoot of approximately 2 per cent. Application of the suggested form to other types of interstages arranges the various networks in the order of their suitability in television amplifiers as considered from their transient responses.

The study of amplifiers for video applications has, in the past, been considered from the viewpoints of both steady-state and transient response. In the process of developing wide-band amplifiers on a steady-state basis, it has been assumed that the desired goal is to achieve a flat amplitude characteristic for the widest band of frequencies. It has been further assumed that the accompanying phase-shift characteristic be linear within limits which will make the over-all amplifier useful for the particular application. Although the experimental verification of a given amplitude characteristic is not difficult, the determination of the accompanying phase characteristic often presents a considerable problem. For minimum phase-shift networks, and these are the types which will be considered here, Bode has shown that the phase characteristic is uniquely defined by the amplitude characteristic; however, this determination, while straightforward, is nonetheless involved.

A figure of merit for comparing the relative utility of wide-band amplifiers on a steady-state basis has been generally accepted as the gain-bandwidth product. By considering the interstages alone, normalizing these to a reference midfrequency impedance level, the steady-state figure of merit degenerates to a comparison of the upper frequency limit at which the impedance of the network falls to some specified value. This cutoff frequency may be considered as that at which the interstage network impedance has decreased to 70.7 per cent of its midfrequency value, or alternatively, as that frequency which, when multiplied by the midfrequency impedance, gives the same product as the area contained under the impedance versus frequency curve. This latter value may be more useful for mathematical manipulation, as considered by Hansen and DiToro.

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In video applications, the fidelity of pulse reproduction is, in reality, the final criterion by which we seek to judge the utility of various interstages. Since the steady-state response only indirectly measures the fidelity of pulse reproduction, it is but an intermediate step in the evaluation of the transient response. In particular, since the steady-state figure of merit does not take cognizance of the variation of the time-delay characteristic near the cutoff frequency, it is of relatively little use in transient work.

The purpose of this paper is to analyze video interstages from a transient viewpoint solely, and to propose a figure of merit by which the relative utility of an amplifier for transient applications can be specified. Although the criterion that is proposed does not rest upon a coherent mathematical development, it, nevertheless, does consider for the first time those factors which, from both an experimental and theoretical viewpoint, must be dealt with in order to make a fair appraisal.

Some definitions are in order before proceeding further. Assume the input signal takes the form of a unit step, which is the integral of the sometimes used $\Delta$ function. An ideal amplifier from a transient standpoint will reproduce such a wave form exactly, and a typical amplifier which fails to do this will, in general, reproduce the initial step with a function that has a finite slope and may also have an "overshoot" or "undershoot" which, in a simple case, will take the form of a damped sinusoid. The steepness of the maximum slope will be a measure of the rise time, and this parameter has been previously defined in several different ways. One must decide whether the rise time should be specified in a manner which is of practical value, or in a manner which lends itself to easy mathematical manipulation. Although we penalize ourselves somewhat for doing so, the former method is elected here. Kallman, Spencer, and Singer have suggested that this parameter be defined in terms of the time interval within which the output amplitude wave form traverses the 10 to 90 per cent points. This has been found, in practice, to be desirable inasmuch as one is seldom concerned with the delay which may take place from the origin to the 10 per cent value, and provided the overshoot is not excessive and the wave form follows a normal shape, the 90 per cent value represents a convenient point at which the wave form excursion may be said to have been completed. If the interstage is normalized with respect to impedance level, the absolute magnitude of the maximum variation of the output wave form from its end value of unity will be defined as the overshoot. These two parameters will be referred to as the rise time $\tau$ and the overshoot $\gamma$, and are shown in Fig. 1. The

Laplace transformation has been used in analyzing the interstage networks to be described here, the methods following those of Gardner and Barnes. In carrying out such an analysis, it is found that the rise time varies inversely with both $R$ and $C$, so that it is convenient to conduct one analysis for each network configuration and specify the rise time in $t/RC$ units. The value of shunt capacitance $C$, in the case of two-terminal networks, will consist of the total distributed interstage capacitance; in the case of four-terminal networks, the convention that has been followed has been to consider $C$ as the sum of the input and output capacitances of the network. This then permits the use of recurrent filter networks without penalizing the ultimately derived figure of merit for the midpoint shunt capacitances of the system.

In comparing the performance of any interstage, one is concerned with the quantities absolute rise time, overshoot, and gain (or impedance level). In order to make the figure of merit independent of impedance level, the relative rise time $\tau$ must be introduced as a reciprocal function. It is considered important that the figure of merit also be some inverse function of the overshoot $\gamma$. The amount that the figure of merit should be penalized for increasing values of overshoot rests upon two foundations. If the figure of merit relates only to single interstages, it should be so arranged that the value of the figure of merit increases with decreasing rise time, up to a point where the overshoot becomes so large as to be objectionable from a practical standpoint. For values of overshoot greater than this value, the figure of merit should decrease rapidly. In television applications, values of overshoot greater than 2 per cent may lead to a perceptible and possibly objectionable ringing in the reproduced picture in the case of sharp black and white transitions. Consequently, such a value of overshoot should be given some weight in determining

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*That is, excluding any delay which merely translates the wave form along the time axis.


at what point the figure of merit will begin to become penalized with respect to increasing values of $\gamma$.

A much more important factor in establishing his value of $\gamma$, however, occurs in relation to the use of the figure of merit for multistage applications. Kallman, et al., and Elmore have shown the relation between the rise time of a single interstage as compared to the overall rise time accruing from the recurrent use of such interstages. If the value of $\gamma$ is small, these results can be summarized by saying that the over-all rise time of $n$ similar interstages is equal to the rise time of a single such interstage multiplied by the square root of $n$. However, if $\gamma$ is not small, this relationship does not hold, and we are led to inquire as to the value of $\gamma$ beyond which the approximation becomes in error. The work of Bedford and Fredendall fortunately gives us some clue to the establishment of the value of $\gamma$ from this standpoint. They have given graphical data showing the performance of 16, 32, and 64 recurrent shunt-peaked interstages, in each case these data being tabulated for three different values of $\gamma$ of a single such interstage. Based upon these results, it is found that if the overshoot has a value equal to or less than approximately 2 per cent, the rise time agrees fairly well with the relationship given above, and the overshoot remains substantially constant. However, for values of $\gamma$ greater than 2 per cent, the overshoot of the over-all amplifier increases enormously, and concurrently the given relationship for the rise time does not hold. From this standpoint, therefore, it seems fair, in the case of a shunt-peaked amplifier, to permit the figure of merit to increase for decreasing rise time, up to a point where the overshoot approximates 2 per cent. For values of parameters in such a network that lead to greater values of $\gamma$, although the rise time may still continue to decrease, it is felt reasonable to have the figure of merit decrease sharply inasmuch as such a configuration is deemed of little use for transient work such as television, where the recurrent use of such a network would cause excessive overshoots.

It has seemed expedient, therefore, to derive the figure of merit based upon this data and to consider the shunt-peaked network first. In Fig. 2 is shown a typical shunt-peaked network, and it will be observed that there are an infinite number of variations possible, depending upon the assigned value of the parameter $m$, where $m$ is given by $m = L/CR$. In Fig. 3 the relationship between rise time and $m$ is given graphically, and we can see that as $m$ is increased from a value of 0 (which corresponds to a simple $RC$ interstage), the value of $\tau$ decreases initially at approximately a linear rate. Also, in Fig. 3 is shown the relationship between $\gamma$ and $m$ for the same network. One observes that for a value of $m < 0.25$ there is no overshoot; and as $m$ increases beyond this value, the overshoot increases, eventually becoming approximately a linear variation. It is found that a value of $m = 0.388$ will give an overshoot of approximately 2 per cent, and based upon the data of Bedford and Fredendall for recurrent shunt-peaked interstages, this value has been selected as representing the point beyond which the figure of merit should be penalized for increasing values of $\gamma$. The expression for the figure of merit which we shall now propose will take cognizance of this fact, leading to a figure of merit for a shunt-peaked interstage which will have a maximum value for this value of $m$. A suitable form for the figure of merit, $F$, is as follows

$$F = a \tau^{-1} e^{-b\gamma}$$

(1)

where

- $F = a$ numeric specifying the relative quality of a shunt-peaked interstage from the transient viewpoint
- $a = $ a constant to give a suitable magnitude to $F$
- $\tau = $ the rise time in units of time normalized by the factor $RC$
- $\gamma = $ the fractional overshoot


From an engineering standpoint, it does provide a useful rule-of-thumb measure of the relative quality of such an interstage for certain video applications.

One is now led to inquire as to the application of such a figure of merit to other types of interstages, both two- and four-terminal varieties. There are two aspects to such networks: (1) their configuration, and (2) the parameter values specified for such a configuration. If we apply the figure of merit derived herein to a typical four-terminal network, for example, one has no assurance that the value thus determined represents the highest figure of merit possible with the particular configuration. However, the derivation of the optimum parameter values for a given four-terminal network represents a vast amount of effort. Since it must be additionally demonstrated that the overshoot resulting from the recurrent use of such an optimum interstage is not objectionable, there seems to be a real value at this time in appraising certain configurations for at least the parameter values which have already been investigated. There appears to be some assurance that for any network, if the overshoot resulting from its use singly is small, the recurrent use of such networks will not give rise to a poor transient response, at least from a standpoint of excessive overshoot. With this thought in mind, and assuming that a 2 per cent overshoot permits of this extrapolation, we have tabulated the figure of merit for a number of typical interstages.

In Fig. 5 is given the transient response for a series-peaked network for two values of the parameter \( k \) indicated in the circuit diagram. For the value \( k = 1.4 \), it is seen that there is a 6.4 per cent "undershoot," which, of course, may be as objectionable as an overshoot. The fact that \( \gamma \) appears squared in the definition of \( F \) means, of course, that an "undershoot" is considered in the same fashion as an overshoot. A value of \( k = 2.0 \) gives a 10 per cent overshoot, further penalizing the resulting figure of merit. In Fig. 6 is given the transient response of a network which will be referred to as "Doba's network" because the particular value of

\[
b = \text{a second constant which determines the amount by which the figure of merit } F \text{ is penalized for different values of } \gamma.
\]

Carrying out computations with such a figure of merit and utilizing the known data for shunt-peaked interstages

\[
F = 1000e^{-100y^2}.
\]

Plotting \( F \) versus the network parameter \( m \) in Fig. 4, it is observed that the maximum figure of merit for the shunt-peaked interstage has a value of 755. For \( m = 0 \), corresponding to a simple \( RC \) interstage, the corresponding figure of merit is 450. The figure of merit is permitted to increase beyond the critically damped case into the

region where overshoot becomes apparent, reaching a peak for a value of \( m \) corresponding to an overshoot which we consider the maximum permissible from two viewpoints:

1. The practical consideration, wherein greater overshoot may result in objectionable distortion in a transmitted picture, in the case of a television system.

2. From the standpoint that a greater overshoot will render a multistage amplifier of limited application because its over-all overshoot is then very objectionable. The logic of this arrangement becomes apparent when one realizes that the figure of merit for \( n \) similar stages of a shunt-peaked amplifier with small overshoot is given simply as the figure of merit for one interstage divided by the square root of \( n \). It should also be clear that this figure of merit is independent of the gain of the amplifier or the corresponding impedance level of the interstage. Of course, in a given case and for a fixed value of shunt capacitance, one is always at liberty to reduce the gain by reducing \( R \); this then makes the absolute rise time in microseconds less, but the value of \( F \), the figure of merit, remains constant.

The figure of merit defined above, while resting on a well-established experimental background, does not have any profound mathematical basis. Nevertheless, from an engineering standpoint, it does provide a useful

![Figure 4](image-url)  
Fig. 4—Figure of merit versus \( m = L/CR^2 \) for shunt-peaked interstage.

![Figure 5](image-url)  
Fig. 5—Transient response, series-peaked interstage.
parameters specified for this configuration is due, it is believed, to S. Doba of the Bell Telephone Laboratories. By way of summary, Table I compares the figures of merit for the interstages discussed.

In view of the decreased rise time that one may achieve with some of the more complicated four-terminal networks, and the concurrent increase in figure of merit, one gains the feeling that if a more complicated network may be justified from the standpoint of improved rise time, such an improvement may be gained at the expense of the more involved adjustment. There has been a tendency in the past with respect to production types of video amplifiers to adhere to the simplicity of the shunt-peaked amplifier, because the complexity of adjustment of the more involved networks has rendered their use ill advised. The justification for this feeling may well be questioned in the future as more profound methods become available for the evaluation of the sensitivity of the figure of merit to change in parameter values. However, in the light of present limited knowledge, there seems to be a definite engineering utility for the application of a simple figure of merit of the type described herein to all of the common types of video interstages. It must be borne in mind, however, that the figure of merit assigned on the basis established above, to a given interstage configuration with presently proposed parameter values, may not in fact represent the ultimate in either performance or figure of merit for such an interstage.

ACKNOWLEDGMENT

The assistance of J. H. Mulligan, Jr., and S. Shamis in the computation of transient responses is gratefully acknowledged.
Design Equations for Reactance-Tube Circuits*

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Summary—Without assuming the usual approximations, design equations are derived for several systems of reactance-tube modulation. Empirical methods are used to derive expressions for the total band swept, in which the effect of each parameter can be directly evaluated. The analysis was completed without the use of the usually accepted simplifying relationships between the impedances of the feedback network.† The critical point where a given-type network changes from an apparent inductance to an apparent capacitance is noted.

INTRODUCTION

Variable-Impedance Circuit

The system under consideration is shown in Fig. 1, where

\[ E_1/E_0 = Z_1/(Z_1 + Z_2) \]  

and assuming a constant current generator

\[ E_0/E_1 = -g_mZ_0/(Z_1 + Z_1 + Z_2). \]

From (1) and (2) it follows that

\[ -g_mZ_0Z_1 = Z_0 + Z_1 + Z_2. \]

The following definitions refer to Fig. 1, where \( Z_1 \) and \( Z_2 \) are parallel CR circuits:

\[ \omega/\omega_0 = \left\{ \left( \frac{g_mR_0R_1 + R_0 + R_1 + R_2}{R_0R_1R_2(C_1C_1 + C_1C_2 + C_2C_2)} \right)^{1/2} \right\}^{1/2} \]

Now, in (5), let \( g_m \rightarrow g_c \), the "Class A" operating point on the "\( g_c - e_c \)" curve, corresponding to the "center" angular frequency \( \omega \), and let the equal positive and negative increments \( \Delta g_m \) correspond to increments \( \Delta \varphi \) and \( \Delta \vartheta \).

Then, defining \( \Delta \varphi = \Delta \omega + \Delta \vartheta \), and rearranging terms, the following will result:
Then, from (5), with \( g_m \equiv g \) and \( \omega \equiv \omega_c \), we can refer the total band swept, \( \Delta \omega \), to the operating angular frequency \( \omega_c \), as follows:

\[
\frac{\Delta \omega}{\omega_c} = \frac{\Delta \omega}{\omega_c} \cdot \frac{\omega_c}{\omega_c}.
\]

Then we see that by a number of elementary operations, the ratio of total band swept to operating center frequency is

\[
\frac{\Delta \omega}{\omega_c} = \left(1 + \frac{1}{1 + \frac{R_2}{R_1} + \frac{R_c' + R_c''}{L_0} + \frac{g_m}{R_2}} \right)^{1/2} - \left(1 - \frac{1}{1 + \frac{R_2}{R_1} + \frac{R_c' + R_c''}{L_0} + \frac{g_m}{R_2}} \right)^{1/2}. \tag{7}
\]

Rewriting (5) will give the necessary explicit expression for \( \omega_0 \)

\[
\omega_0 = \frac{1}{\omega_c^2} \left[1 + \frac{C_2}{C_0(C_1 + C_2)} \right] \left(1 + \frac{R_2}{R_1} + \frac{R_c' + R_c''}{L_0} + \frac{g_m}{R_2} \right)^{1/2} - \left(1 - \frac{1}{1 + \frac{R_2}{R_1} + \frac{R_c' + R_c''}{L_0} + \frac{g_m}{R_2}} \right)^{1/2}. \tag{8}
\]

From (8) the quantity under the radical must be positive and (7) is then defined only for

\[
\omega_0^2 > \frac{g_m R_0 R_1 + R_0 + R_1 + R_2}{R_0 R_1 R_c(C_1 + C_2 + C_1')}, \tag{9}
\]

If the inequality in (9) is reversed, we see that \( \omega_0 \) has no real definition in terms of \( \omega_c \) and we must return to (4). It appears that, for the region where the inequality is reversed, apparently the significance of the imaginaries and reals interchange, and we must use the nominally imaginaries which are now the reals. It can be shown that this is legitimate as it will be found to check in the correct limits with published information.

In this region, then, from the "imaginaries" of (4), \( \Delta \omega/\omega \), and \( \omega_0 \) are defined as follows:

\[
\frac{\Delta \omega}{\omega_c} = \left(1 + \frac{1}{1 + \frac{C_1}{R_0} + \frac{C_1}{R_0} + \frac{C_0}{R_0} + \frac{C_0}{R_0} + \frac{C_1'}{R_0} + \frac{C_1'}{R_0} + \frac{g_m}{R_2}} \right)^{-1/2} - \left(1 + \frac{1}{1 + \frac{C_1}{R_0} + \frac{C_1}{R_0} + \frac{C_0}{R_0} + \frac{C_0}{R_0} + \frac{C_1'}{R_0} + \frac{C_1'}{R_0} + \frac{g_m}{R_2}} \right)^{-1/2}
\]

\[
\omega_0 = \omega_c \left[1 + \frac{g_m R_0 R_1 R_0 C_0 + R_0 C_1 H_0 + C_1' R_0 + C_1' R_0 + C_1'' R_0 + C_1'' R_0 + R_1}{R_0 C_0(R_0 + R_1)} \right]^{1/2}. \tag{10}
\]

The circuits of case I, along with the associated formulas, are listed in Fig. 2.

Fig. 2—Formulas for case I.

**Case II**

Again referring to Fig. 1, the definitions of case II are identical to case I, except that \( Z_1 \) and \( Z_2 \) are now parallel LR circuits, and the corresponding \( t' \)'s are

\[ t_1 = \frac{R_1}{\omega L_1}, \]
\[ t_2 = \frac{R_2}{\omega L_2}. \]
Replacing the \( t \)'s in (4) with quantities defined here the vanishing of the imaginaries this time yields the desired expression for \( \omega / \omega_0 \).

\[
\frac{\omega}{\omega_0} = \left( 1 + \frac{R_2L_2L_1(g_mR_0R_1 + R_0 + R_1) + R_1L_2(R_0 + R_2)}{R_0L_1L_2(R_1 + R_2)} \right)^{1/2}
\]

(12)

Considerations similar to those used in case I, yield the following expressions:

\[
\frac{\Delta \omega}{\omega_c} = \left( 1 + \frac{1}{\omega_c + \frac{g_e}{L_2}} \right)^{1/2}
\]

(13)

\[
\Delta g_m = \left( 1 + \frac{1}{\omega_c + \frac{g_e}{L_2}} \right)^{1/2}
\]

and (1.3) will hold only as long as

\[
\omega_c^2 > \frac{R_2L_1(g_eR_0R_1 + R_0 + R_1) + R_1L_2(R_0 + R_2)}{L_1L_2C_0R_0(R_1 + R_2)}
\]

(15)

In the region where the inequality is reversed, we must return to (4) and obtain

\[
\frac{\Delta \omega}{\omega_c} = \left( 1 + \frac{1}{\omega_c + \frac{g_e}{L_2}} \right)^{-1/2}
\]

(16)

\[
\Delta g_m = \left( 1 + \frac{1}{\omega_c + \frac{g_e}{L_2}} \right)^{-1/2}
\]

and

\[
\omega_0 = \left\{ \begin{array}{l}
\omega_c^2 \left[ 1 + \frac{L_1L_2(g_eR_0R_1 + R_0 + R_1 + R_2)}{R_0R_1C_0(L_1 + L_2)} \right] \\
- \frac{1}{C_0(L_1 + L_2)} \end{array} \right\}^{1/2}
\]

(17)

in which again the quantity under the radical sign must also be positive.

Case II should not be interpreted to cover the conditions wherein the interelectrode capacitances become of primary importance. Some of the more conventional inductive circuits are listed in Fig. 3.

It should be noted that cases I and II are special developments of a more general condition in which \( Z_1 \) and \( Z_2 \) are both parallel LCR circuits.
Rationalizing (19) will give
\[
\frac{E_1}{E_0} = \frac{A}{D} + j \frac{B}{D}
\]
where
\[
A = R_1(R_1 + R_2 + R_{12}(R_{12} + R_{13}))
\]
\[
B = R_1R_2(t_1 - t_2).
\]
Corresponding to the definitions given under case I,
\[
B = R_1R_2(-\omega_1R_1 + \omega_2R_2)
\]
and under case II,
\[
B = R_1R_2\left(\frac{1}{\omega t_1} - \frac{1}{\omega t_2}\right)\]
In both cases,
\[
\tan \phi = \frac{B}{A} = \frac{R_1R_2(t_1 - t_2)}{A};
\]
and, since \(A\) is always positive, if \(t_2 > t_1\) the phase angle will be negative, and if \(t_2 < t_1\) the phase angle will be positive.

This shows that there are distinct regions where a given phase-shift network is capable of either phase delay or of phase advance corresponding to a change-over from effective inductance to effective capacitance for that circuit. These regions are defined by the inequalities (9) and (15).

APPENDIX II

Numerical Example

The parameters for the following problem were chosen from an example given in footnote reference 4, the essential details of which are shown in Fig. 4, to illustrate the present method and to show that its use obviates the necessity for the more cumbersome methods of successive approximations.

The problem will be to accomplish the maximum bandwidth swept at a geometrical center frequency of 23 Mc. The oscillator is a 955 and the reactor a 6AC7. The reactance circuit of Fig. 1 for case I is chosen as the basic circuit, and use is made of the associated formulas.

The following information is available:
\[f_c = 23 \text{ Mc (geometrical center frequency)}\]
\[g_c = 0.0075 \text{ mho (since } g_c = g_{\text{max}}/2 \text{ by definition, and maximum } g_{\text{max}} \text{ of a } 6\text{AC7 can approach } 0.015 \text{ mho)}\]
\[\Delta g_m = 0.0075 \text{ mho (since the change in } g_m \text{ is seen to be one-half of maximum)}\]
\[C_0 = 25 \mu\text{f} \text{ (this is the distributed capacitance of the oscillator coil plus the tube and stray capacitances)}\]
\[C_1 = 20 \mu\text{f} \text{ (this is the input plus stray capacitances of the } 6\text{AC7)}\]
\[C_2 = 1 \mu\text{f} \text{ (this order of capacitance is realizable in practice)}\]

\[R_0 = 20K \Omega\]

A practical value is \(Q_0 = 110\).
\[R_{0'} = 30,000 \text{ ohms. } r_p = 10^4 \text{ for a } 6\text{AC7}. R_e = 50,000 \text{ ohms was used in this example} \]
\[R_1 = 12K \Omega \text{ (this should be as large as possible. The effective } R_1 \text{ is the electronic loading of } 13,000 \text{ ohms in parallel with the } 200,000-\text{ohm grid resistor used)}\]
\[R_2 = 2,000 \Omega \text{ (as shown by (9) should be as small as possible for maximum bandwidth swept. The minimum of } R_2 \text{ is limited by its damping effect on the oscillator)}\]

In solving the problem it is first necessary to find \(\omega_0\), which in turn will give the design value of \(L_0\). From (8)
\[
\omega_0 = \frac{\omega_0}{2\pi} = 19.17 \text{ Mc}
\]
and
\[
L_0 = \frac{1}{\omega_0^2 C_0} = 2.76 \mu\text{h}.
\]

For this problem the bandwidth swept is given by (7)
\[
\Delta f = 7.61 \text{ Mc}
\]

The results of an approximate solution obtained by using formulas (8.4) and (7.4) are given in the Table I. Following is a summary of the results obtained for the problem, as given by the various methods of approach.

<table>
<thead>
<tr>
<th>Method of Solution</th>
<th>(f_c) Mc</th>
<th>(L_0) (\mu\text{h})</th>
<th>(\Delta f) Mc</th>
</tr>
</thead>
<tbody>
<tr>
<td>Successive approximations as given in footnote reference 4</td>
<td>19.7</td>
<td>2.62</td>
<td>7.60</td>
</tr>
<tr>
<td>Present design using (7), (8)</td>
<td>19.17</td>
<td>2.76</td>
<td>7.61</td>
</tr>
<tr>
<td>Present design using less exact equations (8.4), (7.4)</td>
<td>18.4</td>
<td>2.99</td>
<td>8.30</td>
</tr>
</tbody>
</table>
It will be noted that the solution presented here involves considerably less work than usual procedures, and this typical example will sufficiently illustrate the usefulness of the present method.

**Conclusion**

In approaching the solution of any reactance tube problem, it is first necessary to decide which one of the circuits given in Figs. 2 and 3 is most nearly applicable to the problem at hand. The several associated equations readily show the parameters that should be changed and the direction of the change to produce desired results. The degree of accuracy of the design depends on the particular method chosen and its respective formulas. For usual purposes, the approximate methods as shown in Figs. 2 and 3 will give sufficient accuracy. Many times when a particular system is under consideration, it is often useful to know the theoretical maximum bandwidths available; and these limits are easily obtainable as can be seen from the text. It is believed that these particular formulas might be extremely helpful in problems of reactance tube circuits wherein theoretical limits are being approached.

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**Medium-Frequency Crossed-Loop Radio Direction Finder with Instantaneous Unidirectional Visual Presentation**

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**Summary**—A radio direction finder is described which uses a crossed-loop collector system, electronic switch, single superheterodyne receiver, and synchronous rectifier to produce an instantaneous unidirectional visual indication of the direction of arrival of an electromagnetic wave. Design data and operating characteristics are considered, with details given of the new components.

**I. Introduction**

THE RADIO DIRECTION FINDER to be described was a Signal Corps development calling for a direction finder covering the frequency range of 1.5 to 18 Mc. The equipment was to be transportable in a small vehicle and consist of components of such size and weight as to be readily portable by three men. Each set was to be assembled for operation in less than twenty minutes by two men, with operation being possible in the field, in the open, or under canvas. The performance requirements were that it should be as accurate and sensitive as the electronic art permitted without undue complexity of operation or compromise of transportability. It was desirable that the equipment be capable of obtaining bearings on transmissions of short duration.

Development of the radio direction finder was first begun in August, 1943, and pursued actively until August, 1945, when field tests were completed and the performance of the set demonstrated.

The initiation of development first required a decision as to the basic mode of direction finding to be employed.

* Decimal classification: RS01X561. Original manuscript received by the Institute, July 27, 1948; revised manuscript received, November 18, 1948.
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‡ Signal Corps Engineering Laboratories, Fort Monmouth, N. J.

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The requirements of transportability and operation limited the collector system to loop antennas. For direction finding on transmissions of short duration, direction finders requiring a manually or motor-rotated collector system were ruled out; the dual-channel instantaneous direction finder was eliminated because of difficulties of providing equalized phase and gain. This narrowed the field of possible modes of operation to the selective-modulation type of direction finder in which radio-frequency voltages proportional to the direction of arrival of a signal are suitably “tagged” at the input, pass through a common receiver, and are then decoded. Some prior successful work on a left-right, square-wave-switched, cardiod type of direction finder led to a further examination of this method of operation. A brief examination indicated that, provided certain circuit characteristics could be achieved, a crossed-loop square-wave-switched cardiod type of direction finder would work satisfactorily.

**II. Theory of Operation**

The over-all operation of the set can best be understood from the block diagram in Fig. 1. The collector system is a set of crossed balanced loops with a vertical sense antenna. Under ideal conditions, loop 1 receives a radio-frequency signal proportional to $\sin \theta$ ($\theta$ is azimuth angle of arrival of signal). This voltage is fed push-
pull into a radio-frequency switch, switching at a frequency $f_1$ (approximately 253 cps). The output of this switch (voltages $e_1$ and $e_2$ as shown in Fig. 1) is essentially a constant-amplitude radio-frequency voltage with $e_1$ and $e_2$ 180 degrees out of radio-frequency phase, since the switch receives input voltage from either end of the loop to ground. Loop $B$ operates similarly, except that its radio-frequency signal is proportional to $\cos \theta$ and the radio-frequency switching frequency is $f_B$ (approximately 340 cps). Generally, $e_1$ and $e_2$ are not equal in amplitude to $e_1$, $e_2$, although for the case illustrated in Fig. 1 ($\theta = 45^\circ$) the amplitudes are equal. The outputs of the $A$ and $B$ switches are combined with the sense-antenna voltage, which has been shifted through 90 degrees in order that it will be in phase with either $e_1$ or $e_2$. Ideally, the sense antenna voltage should be just equal to the maximum loop-antenna voltage, although practically it is sufficient to ensure that the sense-antenna voltage is always larger than the loop-antenna voltages. When voltages $e_1, e_2$ are added to the sense-antenna voltage, the result is a square-wave modulated signal; similarly, $e_1, e_2$ added to the sense-antenna voltage, gives a square-wave modulated signal. The combination of these two signals produces a signal with a complex envelope which is nonrecurring in shape, since $f_A$ and $f_B$ are purposely chosen to have a nonintegral relationship. The resulting radio-frequency signal passes through a conventional receiver composed of two rf stages, pentagrid converter, two if stages, diode detector with provision for beat-frequency oscillator injection; and an audio amplifier and power amplifier. Automatic-volume-control and pulse noise-suppression circuits are also provided for optional use. The output of the diode detector, consisting of the combination of two square waves of frequencies $f_A$ and $f_B$ also, goes to buffer amplifiers which feed into synchronized rectifiers $A$ and $B$. Each rectifier is switched synchronously with its associated rf input circuit. Consider first the nonsynchronous square-wave signal, which is applied in parallel to both grids of the synchronous rectifier; the net result is that the output is balanced out due to the push-pull action of the circuit. In the case of a square-wave signal which is in synchronism with the switching frequency, a difference voltage is produced at the anodes which is proportional in magnitude to the peak amplitude of the input square wave of like frequency. The polarity of this difference frequency is determined by the relative phase of signal and synchronous switching voltage applied to the grids of the synchronous rectifier. The charging time constant in the anode circuit of the synchronous rectifier is made long compared to the switching frequency, so that the difference voltage is largely direct current with a small superimposed alternating component. The output voltage passes through additional dc filters to remove the remaining alternating components, and is then connected to an appropriate pair of oscilloscope deflection plates.
The oscilloscope electron beam is deflected out in an X and Y direction proportional to \( \cos \theta \) and \( \sin \theta \) and therefore indicates the direction of arrival of the signal. Since dc voltage polarities change for a signal coming from a direction 180 degrees from \( \theta \), an unambiguous direction of arrival is shown on the oscilloscope. For improved presentation, the deflected oscilloscope spot is converted into a line by having a trigger discharge circuit periodically short the oscilloscope deflection plates to ground.

In order to permit net operation of several direction finders for position location, a communication control unit is incorporated in the set; this control unit permits talking and listening over an interconnecting line as well as the connection of the receiver output to the line for remote monitoring. A third position permits signal matching by connecting one earphone to the receiver and the other earphone to the line.

In the above discussion, audio modulation on the radio-frequency signal was not considered. Audio modulation will have no effect on the operation of the direction finder as long as it does not contain a component that is synchronized with either switching frequencies \( A \) or \( B \) or any harmonics thereof; this condition is unlikely to exist in practice. Since the square-wave switching voltages are also present in the audio output, some loss in intelligibility is experienced. In practice, it is found that this is not too serious, particularly for code signals; simultaneous direction finding and monitoring are therefore possible. A switch is provided for disabling the switching circuit; the receiver is then connected directly to the sense antenna and serves as a sensitive intercept receiver. When mounted on a turret-head mounting plate, the set can also be operated as a manually rotated aural null direction finder (one loop only directly connected to the input), or a visual null direction finder (only one rf switch in operation; null indicated by electron beam at center).

A rotating crossed-hair alidade and a reverse illuminated angular scale are provided to facilitate reading bearings. A two-position switch is provided for varying the time constant of the dc filter. The short-time-constant filter allows bearings to be taken on shorter-duration signals, as well as giving a clearer picture of the character of the bearing. The longer-time-constant filter acts to "freeze" a bearing in position; i.e., tends to read the average bearing for a "swinging" reading. In addition, the longer time constant removes more of the noise.

<table>
<thead>
<tr>
<th>Frequency (Mc)</th>
<th>1.5</th>
<th>2.0</th>
<th>3.5</th>
<th>9.0</th>
<th>12.5</th>
<th>16.0</th>
<th>18.0</th>
</tr>
</thead>
<tbody>
<tr>
<td>RF voltage gain</td>
<td>8.0</td>
<td>6.0</td>
<td>2.3</td>
<td>1.5</td>
<td>0.7</td>
<td>1.3</td>
<td>5.0</td>
</tr>
<tr>
<td>Switching ratio</td>
<td>75</td>
<td>70</td>
<td>35</td>
<td>25</td>
<td>13</td>
<td>10</td>
<td>9</td>
</tr>
<tr>
<td>Grid input impedance (ohms)</td>
<td>2,000</td>
<td>2,200</td>
<td>750</td>
<td>360</td>
<td>200</td>
<td>55</td>
<td>—</td>
</tr>
<tr>
<td>Grid rf input (( \mu )V) for receiver ( (S+N/N) = 10 )</td>
<td>3.2</td>
<td>5.0</td>
<td>4.0</td>
<td>4.0</td>
<td>6.5</td>
<td>4.5</td>
<td>4.0</td>
</tr>
</tbody>
</table>

**Note:** For all tests, the multivibrator was operating and measurements were made from one grid to the common output. Signal modulation was 400 cps, 30 per cent.
loops are 22 and 16 inches on a side, respectively. The center of each loop antenna is grounded and the loop terminals are capacitively coupled to the balanced rf switch input. The sense antenna is inductively coupled to the tuned receiver input in order to provide the 90° phase shift required to bring the sense-antenna voltage in phase with the loop-antenna voltage.

In the development of the rf switch, the following characteristics are important:

1. A high switching ratio.*
2. Large balanced rf gain with good stability and minimum noise.
3. Adequate square-wave output voltage of good shape for use in operating synchronous rectifier circuit.

The following four types of rf switches were designed and tested.

1. Grounded Anode (Fig. 3(a)). This circuit utilizes a dual triode connected as a conventional multivibrator* oscillating at a low frequency. Additional resistors are added in series with anode-to-grid coupling capacitor in order to maintain a large rf impedance at the grids. The rf is connected to the triode grids, with the output taken from a common cathode impedance. This circuit exhibited good switching ratios but its rf gain (0.3 for a 6SN7GT) was poor.

2. Grounded Grid (Fig. 3(b)). This circuit is similar to the one above except that small capacitors serve to ground the grid at radio frequency, and the input is to separate cathodes with the output from a common impedance coupled to the anode by means of small rf coupling capacitors. This circuit exhibited excellent switching ratios and moderate gain, but was considered inadequate because of small square-wave output voltage and poor square-wave shape.

3. Grounded Cathode (Fig. 3(c)). This circuit was found to have the best operating characteristics and was incorporated in the final design. Detailed performance data as a function of frequency are tabulated in Table I. Potential instability of the triode necessitated careful choice of circuit constants and placement of parts. Small resistors in series with the control grid together with large resistors in series with anode-grid coupling capacitor serve to reduce regeneration. Several types of dual triodes were tried, including the 7FB, 6SN7GT, 6SU7GT, and 6SL7GT; of these the 7FB was found to be the best. In order to insure equal rf gains both between triode sections and between tubes, it was necessary to test tubes and utilize only those tubes that had transconductances matched within 5 per cent.

4. Dual Tetrode (Fig. 3(d)). This circuit is similar to Fig. 3(c) except for the addition of the two screen grids.

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* H. M. Wagner and J. F. Herrick, “Self-switching amplifier,” Electronics, vol. 20, pp. 128–131; June, 1947. (This article contains an analysis of a switch similar to those described herein except designed for operation at higher radio frequencies.)

* Switching ratio is defined as the ratio of radio-frequency output voltage during on and off half-cycle of the multivibrator.

Development of this circuit was initiated with the object of improving the grounded cathode circuit, both from the standpoint of improved stability and improved operating characteristics. Comparative tests using 7F8 dual triode and 7G8 dual tetrode tubes indicated that (a) the 7F8 had a signal plus noise-to-noise ratio averaging three times better than the 7G8 over the frequency range of 1.5 through 18 Mc; (b) the radio-frequency gain characteristics were about the same for both tubes; (c) switching voltage stability and wave shape were satisfactory for both tubes; and (d) the square-wave output voltage of the 7F8 averaged 30 per cent higher than for the 7G8. The choice of the 7F8 tube was due primarily to (a) above, and, in addition, to the fact that the 7G8 was considered a developmental tube at the time.

The superheterodyne receiver covered the frequency range of 1.5 to 18 Mc in six bands. It was believed at first that, because of the necessity of passing square-wave modulated signal with small envelope distortion, a flat-top if response curve was desired. Tests indicated, however, that a bell-shaped response curve, center frequency at 470 kc with 3 kc bandwidth (at the half-power point), was sufficient for passing the carrier and sidebands and, in addition, was convenient in alignment. A narrower bandwidth of, perhaps, 1 kc would be more desirable for separation of communication channels, but is insufficient for passing the direction finder intelligence. Because of the existence in the if stages of direction-finder intelligence at several amplitude levels, the linearity of these stages is also of considerable importance. For this reason, the automatic volume control can be used only when the receiver is operated for intercept purposes.

Sensitivity versus frequency data at both sense-antenna and switching-tube inputs are tabulated in Table II.

The set was designed to operate with a dynamotor unit driven from a 12-volt storage battery. In addition, a 180-cps resonant-reed vibrator was used in conjunction with a transformer to furnish 6.3 volts at 0.6 ampere, insulated for 3,000 volts from ground, for the oscilloscope heater; a 1,200 volts at 1 ma dc (rectified by means of 110 small disk selenium rectifiers) for oscilloscope acceleration and focus; and +500 volts, 10 ma dc (selenium rectifier) for the anode voltage of the synchronous rectifier tubes. Careful shielding and filtering was used to reduce rf noise.

In order to eliminate radar, ignition, and similar impulse noises, an automatic noise limiter was provided. The circuit was adjusted to provide automatic limiting when impulse noise peaks exceed the average carrier level. Square-wave direction-finder intelligence modulation is, accordingly, passed without distortion.

The function of the synchronous rectifiers is to separate the N-S and E-W direction-finder intelligence and to provide dc voltages proportional thereto. The basic circuit consists of a pair of triodes operating as balanced modulators, as shown in Fig. 4. The detected signal voltage is introduced to the triode grids in parallel, while the square-wave switching voltage is applied in push-pull. For proper operation, the grid bias is chosen so that each tube operates as a class-A amplifier during the "on" portion of the switching cycle and is completely cut off during the "off" portion of the cycle. If this is done, the push-pull dc output voltage is proportional only to the amplitude of an input voltage of frequency synchronous with the switching frequency. The polarity of the dc voltage difference is dependent upon the relative phase (established at the associated input switch) of the square-wave switching voltage and synchronous input signal voltage. Proper "sense" is thus obtained.

Since large output dc voltage differences are required for full oscilloscope deflection, the problem of linear operation is a severe one. Some consideration was given

![Fig. 4—Synchronous rectifier circuit.](image-url)

**TABLE II**

**SENSITIVITY VERSUS FREQUENCY DATA**

<table>
<thead>
<tr>
<th>Frequency (Mc)</th>
<th>Signal Input</th>
</tr>
</thead>
<tbody>
<tr>
<td>A (µv)</td>
<td>B (µv)</td>
</tr>
<tr>
<td>2.2</td>
<td>0.5</td>
</tr>
<tr>
<td>3.0</td>
<td>0.8</td>
</tr>
<tr>
<td>4.0</td>
<td>0.8</td>
</tr>
<tr>
<td>5.0</td>
<td>0.8</td>
</tr>
<tr>
<td>6.5</td>
<td>0.8</td>
</tr>
<tr>
<td>8.0</td>
<td>0.9</td>
</tr>
<tr>
<td>9.5</td>
<td>0.9</td>
</tr>
<tr>
<td>11.0</td>
<td>1.0</td>
</tr>
<tr>
<td>12.5</td>
<td>1.0</td>
</tr>
<tr>
<td>14.0</td>
<td>1.0</td>
</tr>
<tr>
<td>16.0</td>
<td>1.0</td>
</tr>
<tr>
<td>18.0</td>
<td>1.0</td>
</tr>
</tbody>
</table>

*Note: (a) Signal input A is the input at the sense antenna with the switching tubes off for 4:1 signal-plus-noise to noise ratio. Signal modulation is 400 cps, 30 per cent.
(b) Signal input B is the input at the grid of the switching tube with the switching tube on, for 4:1 signal-plus-noise to noise ratio. Signal modulation is 400 cps, 30 per cent.
(c) Signal input C is the same as for signal input B but for full-scale deflection on the oscilloscope tube (approximately 150 volts between plates).
(d) Phantom antenna consists of a 100 µf capacitor in series with the input to the receiver.*
to the use of dc amplifiers, but due to additional complexity and potential instability, this was discarded.

Instead, the problem was resolved by exhaustive circuit and tube studies together with use of a high B+ voltage (500 volts). Of the various tubes tested (7F8, 6SN7GT, 6SL7GT, 7G8, 6N7, 6E6, and 6SC7), the 6SN7GT gave the best performance. As in the case of the rf switching tubes, matched performance of the four triodes is required, and tubes are selected with transconductances matched to 1 per-cent. A screwdriver-adjustable gain control is also provided, mainly to compensate for variation in transconductance due to aging.

A three-section resistance-capacitor filter network couples each synchronous rectifier anode to the corresponding oscilloscope deflection plate and acts to remove any ac variations in the anode voltage. The total time constant of this circuit is 50 milliseconds; by parallelizing additional capacitance using a panel control, this time constant can be increased to 0.1 second.

The method used for forming a radial line from the deflected oscilloscope spot consists in using triodes to short, periodically, the oscilloscope deflector plates to ground, as shown in Fig. 5. Three dual triodes are used; one acting as an unbalanced multivibrator, and the other two as electronic switches. Dissimilar tube characteristics caused the discharge trace to differ from the charge trace, but since the discharge trace was hardly visible, this was not objectionable and the resulting line presentation was considered very acceptable.

Since manufacturing tolerances on the oscilloscope (3BP1) permit perpendicularity of the deflection plates to be off by as much as ±3°, it is necessary to check tubes; no difficulty was experienced in obtaining tubes in which the plates were perpendicular within ±3°. The unequal horizontal and vertical deflection sensitivities were compensated for by using different load resistors and providing screwdriver-adjustable controls in the switching rectifiers.

IV. SYSTEM PERFORMANCE

The over-all performance of a direction finder is probably best summed up by data presenting the field strength required for a given quality of bearing. The data for a bearing readability of ±1° are shown in Table III; calibrated fields were obtained by operating the equipment in a calibrated screen room. Some indication of the character of the instantaneous bearing can be obtained from Fig. 6(a), (b), and (c). By using the long-time-constant integrating circuit, moderately good bearings can be obtained on signals well below the input noise level.

![Fig. 5—Oscilloscope line formation circuit.](image)

![Fig. 6—Bearing indication.](image)

(a) Signal strength, below noise level.
(b) Signal strength, weak.
(c) Signal strength, good.

**TABLE III**

<table>
<thead>
<tr>
<th>Frequency (Mc)</th>
<th>Loop Antenna</th>
<th>Field Strength for ±1° Readability (µv/m)</th>
</tr>
</thead>
<tbody>
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<tr>
<td>18.0</td>
<td>High frequency</td>
<td>1.5</td>
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</table>

The evaluation of the field performance of the set in the lower frequency range (1.5-3.0 Mc) can be ascertained from Fig. 7, in which is plotted the percentage of

ACKNOWLEDGMENTS

The development work described in this paper was conducted under the leadership of the authors and Aldo Scandurra, formerly of the Signal Corps Engineering Laboratories. Others who contributed in a substantial manner to the project are Benjamin Bernstein, Julia Herrick, Josephine Hollingsworth, Gustave Shapiro, and Cosimo Testa.

It is believed that the method of operation described represents a new approach to an instantaneous direction finder. It is the purpose of this paper to record the work that has been accomplished, so that it will be readily available to anyone contemplating similar development.

While this equipment was designed primarily for radio communication direction finding, it is believed that it may also have important application to homing and navigation of ground, sea, and air vehicles, for position finding for rescue or control operations, and for wave propagation studies. With a suitable collector system and receiver, the same method of direction finding could be employed at frequencies other than the 1.5 to 18 Mc range employed in this set. Of particular interest would be the adaptation of the set to an Adcock-type collector system to reduce bearing errors on abnormally polarized radio waves. Further development work could be spent advantageously on the electronic switching circuits; and, in particular, the improvement to be realized in separating the switching and wave-generation functions in the input tubes should be studied. Other methods of "tagging" or modulating the input signal should be investigated; of the alternate methods available, the following were considered but not tried:

(a) The sequential switching (electronic single-pole four-position switch) of the four loop voltages to the phase-shifted sense voltage with synchronously switched rectifiers. With this mode of operation, only a single fundamental switching voltage is employed.

(b) The switching of the loop voltages directly to the quadrature sense voltage to produce phase modulation instead of amplitude modulation. Synchronously switched phase detectors are then employed to decode the data.
Contributors to Waves and Electrons Section

H. M. Beck (A'37-M'48) was born on January 18, 1912, at Jefferson, Wis. After graduation from RCA Institutes, Inc., of Chicago, he worked for broadcasting station WHBL of Sheboygan, Wis., from 1937 to 1939. He was with RCA Victor Distributing Corp. in Chicago and went to Zenith Radio Corp. in 1940 as test equipment engineer. In 1942 Mr. Beck came to the Naval Research Laboratory, where he is now employed as radio engineer in the Radio Countermeasures Branch of the laboratory.

For a biography and photograph of Milton Dishal, see page 530 of the May, 1949, issue of the PROCEEDINGS OF THE I.R.E.

E. Guy Hills (A'41) was born in El Paso, Texas, on February 1, 1918. He received the B.S. degree in electrical engineering from the New Mexico State College in 1939, and the M.S. in mathematics from the University of Michigan, where he had a teaching assistantship in 1940. After five months as a test engineer for the General Electric Company in Schenectady, N. Y., he became a radio instructor for the U. S. Air Corps. During the following two years, Mr. Hills worked at night toward the Ph.D. degree in mechanics, at Washington University.

In 1934, Mr. Hills joined the engineering staff at the Belmont Radio Corporation in Chicago, Ill., and spent the next six years in radar and television development work. At present he is the manager of the Government Engineering Division of the Webster-Chicago Corp.

L. J. Giacolitto (S'37-A'42-M'44-SM'48) was born in Clinton, Ind., on November 14, 1916. He received the B.S. degree in electrical engineering from Rose Polytechnic Institute, Terre Haute, Ind., in 1938, and the M.S. degree in physics from the State University of Iowa in 1939, where he was a research assistant. From 1939 to 1941, while holding an appointment as teaching fellow, he completed graduate course requirements for a doctorate degree at the University of Michigan.

Mr. Giacolitto was associated with Collins Radio Company during the summers of 1937 and 1938, and with the Bell Telephone Laboratories in the summer of 1940. From 1941 to 1945 he was on active military duty with the Signal Corps, engaged in development activities in the field of radio, navigational, and meteorological direction-finding equipment. In 1945 and 1946 he was on duty in Japan, concerned with technical intelligence work, after which he returned to inactive status as a major in the Signal Corps Reserve. In December, 1947, and January, 1948, he returned to temporary active duty with the Thermionics Branch, Evans Laboratory.

Since June, 1946, Mr. Giacolitto has been a research engineer with the Radio Corporation of America, RCA Laboratories Division, Princeton, N. J., engaged in research and development on electron tubes and electronic equipment. He has served on IRE New York Section committees, and helped to organize the Monmouth Subsection, serving as Chairman. He is now Secretary-Treasurer of the Princeton Section.

Leonard Mautner (M'46-SM'47) was born on October 30, 1917, in New York, N. Y. He received the degree of B.S. in electrical engineering from the Massachusetts Institute of Technology in 1939. He did graduate study at the Stevens Institute of Technology from 1940 to 1941, and at the Massachusetts Institute of Technology in 1942.

In 1939, Mr. Mautner was an illuminating engineer at the Macbeth Daylighting Corp., New York, N. Y., later joining the Army Signal Corps, as radio engineer. In 1942 he joined the television department of the National Broadcasting Co., and when broadcasting was curtailed because of the war, he became a staff member of the Radiation Laboratory at the Massachusetts Institute of Technology. Here he was a member of the Indicator Group, which developed a variety of indicator units for radar equipment. In 1944, Mr. Mautner was asked to serve as a Radiation Laboratory member of the Combined Research Group at the Naval Research Laboratory, Washington, D. C., where he took charge of the display section. In this capacity, he supervised the development of all of the display and interconnection equipment for the Mark V IFF/UNB Project. In 1945, he joined the Research Division of the Allen B. DuMont Laboratories, and in 1947 became manager of their television transmitter division. He recently organized a new firm, Television Equipment Corp., where he is president and director.

Mr. Mautner is a member of Eta Kappa Nu and is active on several committees of the Radio Manufacturers' Association, serving as chairman of the Television Studio Facilities Subcommittee of the Television Transmitter Dept. He is author of number of technical papers, as well as a text book entitled "Mathematics for Radio Engineers."

E. Guy Hills

L. J. Giacolitto

Leonard Mautner
Contributors to Waves and Electrons Section

Richard C. Palmer (S'42-A'44) was born on October 9, 1922, in Washington, D.C. After graduation from the University of Virginia in 1943, he entered the student test program of the General Electric Co. From 1944 to 1946, he did developmental work on automatic controls for the Tennessee-Eastman Corp. at Oak Ridge, Tenn., leaving to engage in television camera development with Remington Rand, Inc., at South Norwalk, Conn. Since 1946, Mr. Palmer has been associated with the Allen B. DuMont Laboratories, Inc., doing developmental work on television and oscillographic equipment. Mr. Palmer is an associate of the AIEE, and a member of Sigma Xi and Tau Beta Pi.

Delos W. Rentzel was born in Houston, Tex., on October 20, 1909. He was graduated from the Engineering School of Texas Agricultural and Mechanical College in 1929. He worked on radio station installations for the United States Navy until 1931, when he joined American Airways, Inc., as a radio operator and station manager. He was promoted successively to system chief operator, assistant director of communications, and finally to director of communications.

In 1943 Mr. Rentzel became president of Aeronautical Radio, Inc., where he was active in helping to develop improved aviation equipment and techniques as chairman of the Radio Technical Planning Board's Aeronautical Radio Panel and vice-chairman of the Radio Technical Commission for Aeronautics. During this period he was also a member of the Board of Aeronautical Radio of Mexico, serving as its president in 1948; and as a member of the Board of the Airborne Instrument Laboratory.

Mr. Rentzel was appointed Administrator of Civil Aeronautics on April 8, 1948, and was confirmed by the Senate on May 5. As Administrator, he is continuing his interests in improving airway and airline operations by his committee memberships on the Telecommunications Coordinating Committee, the Radio Technical Commission for Aeronautics, the National Advisory Committee for Aeronautics, and the Committee on Navigation of the National Military Establishment Research and Development Board.

Mr. Rentzel is a member of the Institute of Aeronautical Sciences, Society of Automotive Engineers, Aircraft Owners and Pilots Association, American Ordnance Association, National Press Club, and the Aero Club.

Samuel Stiber (A'45) was born in New York, N.Y., in 1909. He received the B.S. and M.S. degrees from the College of the City of New York, in 1931 and 1933, respectively. From 1935 to 1940 he was an instructor in physics and mathematics in Puerto Rico. Returning to the United States in 1941, he became an instructor in radio engineering at the Air Force Technical Training Command, at Scott Field, Ill. Since 1943, Mr. Stiber has been a radio engineer at the Evans Signal Laboratories, Belmar, N. J.

Mr. Stiber has made numerous inventions pertaining to communications and direction-finding, such as the instantaneous direction finder, resonant structures for panoramic receivers, and new receivers for detecting radio signals over wide frequency bands.

Jack D. Young (S'42-A'45) was born in Council Bluffs, Iowa, on April 17, 1916. He received the B.S. degree in electrical engineering from the University of Iowa in 1942. During the time he was an undergraduate, he undertook certain developmental work in connection with the recording of speech as an aid to students taking courses in public speaking. He was employed by the Nebraska Power Company of Omaha, Neb., during summer vacations.

Shortly after graduation he entered the Naval Research Laboratory in Washington, D.C., as a radio engineer, and was closely associated with various projects involving the development of Naval radar equipment and special electronic devices. During 1945, Mr. Young undertook certain specialized work in the Radio Countermeasure Branch of the Naval Research Laboratory. This work is primarily concerned with research and design of electronic circuits utilizing cathode-ray tubes and their applications in various naval electronic systems.

Yofo Pay Yu (A'48) was born on August 27, 1917, in Canton City, China. He received the M.S. degree in electrical engineering from Lehigh University. From 1942 to 1947 he was engaged in the development of special electronic instruments, FM-AM receivers, and television with industries, as a project engineer. Since 1947, Professor Yu has been associated with the School of Engineering, North Dakota Agricultural College, as an associate professor in charge of the electronics option. Professor Yu has published various articles in the electronics and communication field. He is a member of the American Institute of Electrical Engineers and the Society for American Engineering Education.
Abstracts and References
Prepared by the National Physical Laboratory, Teddington, England, Published by Arrangement with the Department of Scientific and Industrial Research, England, and Wireless Engineer, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications and not to the IRE.

The Institute of Radio Engineers has made arrangements to have these Abstracts and References reprinted on suitable paper, on one side of the sheet only. This makes it possible for subscribers to this special service to cut and mount the individual Abstracts for cataloging or otherwise to file and refer to them. Subscriptions to this special edition will be accepted only from members of the IRE. Subscribers to the Proc. I.R.E. at $15.00 per year. The Annual Index to these Abstracts and References, covering those published from February, 1948, through January, 1949, may be obtained for 4s. 8d. postage included from the Wireless Engineer, Dorset House, Stamford St., London S. E., England.

ACOUSTICS AND AUDIO FREQUENCIES

A 015:334


534:143:530.652 Motion-Impulse Measurements on a Magnetostrictive System—F. P. Finton. (Jour. Acoust. Soc. Amer., vol. 21, pp. 217-182; May, 1949.) Describes a method of mounting a nickel-titanium coil in a block of Permalloy, a nonmagnetic plastic, so that motion due to magnetostriction can be damped out. The motion-impulse can then be obtained as the difference between the clamped and unclamped impedances. Frequency range, to 45 kc.


534:22-13+534.231.3-13:534.321.0 Ultrasonic Velocities and Absorption in Gases at Low Pressures—L. F. Zartman. (Jour. Acoust. Soc. Amer., vol. 21, pp. 171-174, May, 1949.) Improvements in an interferometer of the Hubbard type (1912 Wireless Ray). Absorptions, p. 171, are described. Measurements of velocity and absorption in air, CO₂, Na₂, and H₂ are given for frequencies from 500 kc to 2.16 Mr. temperatures from 4°C to 36.9°C and pressures from 82.17 cm Hg to 0.45 cm Hg.


534:7:611:85 The Structure of the Middle Ear and the Hearing of One’s Own Voice by Bone Conduction—G. V. Bckesy. (Jour. Acoust. Soc. Amer., vol. 21, pp. 217-212; May, 1949.) A detailed discussion of the construction of the animal ear and throat as an acoustic system shows how the ear’s sensitivity is not upset by sounds originating in the throat.

534:7:611:85 The Vibration of the Cochlear Partition in Anatomical Preparations and in Models of the Inner Ear—G. V. Bckesy. (Jour. Acoust. Soc. Amer., vol. 21, pp. 233-245; May, 1949.) Translation of an article published in Akadémiai Z., vol. 7, pp. 173-186; 1942. An account of experimental methods (acoustic and optical) by which the vibrations of the round window (cochlear partition) were studied. A working model of the cochlea is described.


534:833 Absorption by Sound-Absorbent Spheres—R. K. Cook and P. Chretienkov. (Jour. Acoust. Soc. Amer., vol. 21, p. 167-170; May, 1949.) Theory and measurements show that the absorption coefficient of a sphere covered with hair can be greater than unity. The normal impedance assumption does not appear to be valid.


534:861.1+534.862.1 A Demonstration Studio for Sound Recording and Reproduction and for Sound Film Projection—Philips Tech. Rev., vol. 10, pp. 196-294; January, 1949.) Description of the construction and special features of a new studio at Eindhoven. The reverberation time at the higher frequencies (0.9 second at 2,000 cps) is only slightly less than at the lower frequencies (1.3 second at 100 cps); this has a good effect on high-note response.

621:395:61 The Miniature Electrodynamic Microphone of the Société indépendante de T.S.—(Ann. Radiotechn., vol. 4, pp. 161-163; April, 1949.) A short description of Type S.I.F. Mld. 8, which is 30 mm in diameter, 18 mm thick, and weighs 30 gm. Response is linear to within ± 7.5 db in the frequency band 100 to 6,000 cps and signal-to-noise ratio and sensitivity are good. Impedance is 70±10 per cent at 1,000 cps.

621:395:61+681:85:621:315:612.4 Application of Activated Ceramics to Transformers—H. W. Kern. (Jour. Acoust. Soc. Amer., vol. 21, pp. 198-201; May, 1949.) The conditions are examined under which granular ceramics can be made to have pronounced piezoelectric properties. Methods of applying the necessary stress to ceramic strips are described. Various applications are mentioned, including a sensitive phonograph pickup.


621:395:61+621:317:32 The Substitution Method of Measuring the Open Circuit Voltage Generated by a Microphone—M. S. Hawley. (Jour. Acoust. Soc. Amer., vol. 21, pp. 189-199; May, 1949.) Analysis shows that the “normal” substitution voltage equals the open-circuit voltage for all types of acoustic measurement and for any value of the electrical impedance loading the microphone.

A Stereophonic Magnetic Recorder—N. Conran. (Proc. I.R.E., vol. 37, pp. 442-447; April, 1949.) An experimental 3-channel recorder and play-back unit is described and results obtained with it are discussed. Test results for a small room are obtained with a dished mounting of two loudspeakers.

Graphical Analysis of Linear Magnetic Recording Using High-Frequency Excitation—M. Conran. (Proc. I.R.E., vol. 37, pp. 569-573; May, 1949.) A study of the properties of a high-frequency component to an audio signal is to which it is to be recorded magnetically results in a low-distortion, linear recording characteristic under certain conditions. This paper gives a graphical method for constructing the recording characteristic from the Bg versus H curve of the recording material. An analysis accounts for such magnetic-recording characteristics, as variation in sensitivity with bias, linearity at low recording levels, and adjustment for minimum distortion.

Contrast Expansion—Wheeeler. (See 2171.)

Non-Linear Distortion in Dynamic Loudspeakers due to Magnetic Effects—W. J. Cuninghham. (Jour. Acous. Soc. Amer., vol. 21, pp. 202-207; May, 1949.) An analysis of two kinds of distortion; that due to (a) the force between the voice coil and the magnet iron, and (b) the nonuniformity of the magnetic field. Distortion corrections may be several tens of 1 percent. To minimize (b), the voice coil and the magnet gap should have unequal lengths.


High-Impedance Cable—S. Frankel. (Proc. I.R.E., vol. 37, p. 406; April, 1949.) An approximate formula for the impendance of a long solenoid surrounded by a cylindrical shield is derived by consideration of the multiversion-transmission line obtained when coil and shield are cut lengthwise and unwrapped. Results obtained appear to confirm the value of the correction coefficient given by Winkler. The formula obtained for the distributed capacitance of the system agrees with that for a coaxial cable. See also 1278 of June (Hodell).


Geometrical Representation of the Characteristics of an Active Obstacle Inserted in a Waveguide—J. Ortusi and P. Fechner. (Ann. Radioloe., vol. 4, pp. 131-145; April, 1949.) In waveguide problems, it is often necessary to consider the reflections caused by impedances thrown back into the main waveguide high or series matching stubs, which may be variable. A simple geometrical method is described for determining the coefficients of reflection, transmission, and energy loss corresponding to such impedances and, conversely, for determining the coefficient from measured values of the coefficients. The slunt and series cases are considered separately.

The Conditions of Propagation of H0 Waves, and their Radiation Characteristics (Ann. Radioloe., vol. 4, pp. 94 116; April, 1949.) H0 waves are defined as those for which the longitudinal current in the waveguide is everywhere zero. Discussion: (a) mathematical study for such waves in waveguides of circular section, (b) methods of producing them practically free from parasites, (c) filters of various types favoring the propagation of H0 waves, (d) measurement methods, (e) attenuation, (f) effects of waveguide deformation or curvature and of the dielectric filling the waveguide, (g) curves showing the attenuation in circular waveguides comparable with that obtained in the space air by diffraction, and (h) applications to the construction of cavity wave-meters with very high Q and to problems connected with radar scanning.


Antennas for Circular Polarization—J. Maillard. (Elec. Éduc., vol. 29, pp. 110-123; March, 1949.) The general characteristics of antenna radiation in circular and rectangular waveguides and filters of uhf antennas are considered. Various types are described and their use for particular services is discussed.


Use of a Reflecting Mirror and of Simple Electromagnetic Lenses for the Experimental 23-cm Link Between France and Corsica—J. Hucon. (Ann. Radioloe., vol. 4, pp. 157-160; April, 1949.) Communication could be effected by a link with direct visibility between Mont Angol and Calenzana. With Grasse and Calenzana (a) directly, with part of the path beyond optical range, or (b) indirectly, using a reflector installed on Monte Grasso (Corsica), with direct visibility to Grasse and oriented so as to reflect signals from Grasse to the receiving antenna at Calenzana, or vice versa. The reflector was of perforated sheet-iron and had an aperture of 10 meters. It reflected a gain of 20 db compared to the direct link beyond optical range. The electromagnetic lenses used with the Mont Angol transmitter had radius of curvature of the order of 4000 mm. The propagation velocities in these waveguides were so calculated as to correct the phase shifts in the aperature plane and to give an effective plane wave output, resulting in improved directional characteristics. See also 3508 of 1948 (Kivéré).


Aerial Arrays with Horizontal Beams without Side Lobes—O. Schnirch. (Bull. Schweiz Elektrotech. Ver., vol. 38, pp. 15-20; January 11, 1947. In German, with French summary.) Combined array theory developed by considering the resultant obtained by superposition of the radiation distributions from each element. Technical requirements demand a minimum number of such elements with optimum efficiency. Practical design formulas are derived and applied in two numerical examples.

CIRCUITS AND CIRCUIT ELEMENTS

A Simple Regenerative Amplifier Using High-Frequency Excitation—R. M. Symons. (Proc. I.R.E., vol. 37, pp. 569-573; May, 1949.) A simple regenerative amplifier is described which is suitable for use in radio receivers. The circuit is shown to be stable and to have a gain of about 10 db.

Graphical Analysis of Diode Circuits—G. L. Hamburger. (Wireless Eng., vol. 67, pp. 147-153; September, 1948.) The analysis of the basic diode rectifier circuit leads to an interpretation which has no formal explicit solution; it can, however, be solved by a simple graphical method which is applicable to periodic input voltages of arbitrary wave form. The method is applied to a typical AM detector, and to squaring and clipping circuits. Circuits involving reactance are considered briefly.

Design of I.F. Transformers—B. Sandel. (Radioeng., no. 9, pp. 4-9; May and June, 1948.) Design procedures can be reduced to a few routine operations with the aid of charts and tables if certain assumptions are made. Only the 2-winding transformer with mutual inductance coupling is here considered; the added capacitance coupling is taken into
Abstracts and References

a c c o u n t  b y  a d j u s t i n g  t h e  c o e f f i c i e n t  o f  c o u p l i n g
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6 2 1 . 3 9 6 . 6 1 1 . 3 : 6 2 1 . 3 6 5 .
Section 5, by Frey, outlines work on frequency multiplication.

Section 6, by Schumann, discusses plasma oscillations.

Section 7, by Marx, deals briefly with spark transmitters for centered waves.

Section 8, by Heil, discusses the production of waves in the range 0.1 to 0.3 mm with the high-pressure He-cvapotron lamp.

621.360.615 2176 Oscillation Amplitude in Simple Valve Oscillators—A. S. Glaubin (Wireless Eng., vol. 29, pp. 127-129; May and June, 1949.) A method is derived of calculating this amplitude in oscillators of the regenerative type where grid leak is used. The amplification factor is found in terms of parameters which are functions of the tube and circuit constants, and the solution is presented graphically.

Two types of amplitude instability are studied and criteria for their existence are derived. The first type is in dynamic instability of squeezing; the second type gives rise to the effect known as oscillation hysteresis.

The analysis is applicable to all the components of an oscillator circuit, subject to the conditions that the tube should operate always in the space-charge-limited condition, and that the anode voltage should never fall below the point where the negative current is rapidly diverted to the grid or screen.

621.360.615 2177 Wide-Range Deviable Oscillator—M. E. Ames (Electronics, vol. 22, pp. 96-100; May, 1949.) A cathode follower working into a capacitive load will produce a phase-shift dependent on grid voltage, which, in turn, depends on the grid bias. The type of oscillator described uses four such phase-shifting stages following an amplifier. The system oscillates at the frequency of the phase of the fourth stage is 45° and this effect is d by making the capacitive reactance of each stage equal to its equivalent resistance. FM is accomplished by simultaneous variation of the anode current to all the four phase-shifting stages. Each time the effective resistance is altered by changing the grid bias, the oscillation frequency changes to a new value at which the reactance of the fixed capacitors is equal to the new effective resistance. The voltage loss of the phase-shifting stages remains constant because the amplification factor of the grid is changed and because the phase-shift required is constant.

With careful design, a linear frequency versus modulating-voltage characteristic is observed. The circuit constructed has a frequency range from 150 cps to 15 kc, frequencies than are usually possible with this type of oscillator. Circuit diagrams are given.

621.360.619.13 2178 A Simple Method of Producing Wide-Band Frequency Modulation—Rahelk and Sankar (See 2162.)

621.360.617 2179 Blocking Oscillators—W. T. Cocking (Wireless World, vol. 55, pp. 280-283; June, 1949.) A method is derived for improving the linearity of the sawtooth output by eliminating the linearity of the sawteeth by superimposing a wave on the wave at the comman-pointed of each stroke. This enables the circuit to be used as an oscilloscope time base at higher repetition frequencies than can usually be possible with this type of oscillator. Circuit diagrams are given.


621.360.617 2181 Symmetrical Multivibrators—R. Feinberg (Wireless Eng., vol. 26, pp. 154-158; May, 1949.) "Formulae are derived from an equivalent circuit diagram to give the frequency and waveform of oscillation of a symmetrical multivibrator circuit with point-to-point connection of the coil windings such as it is, as well as that of the interelectrode capacitance of the values that correspond to an equivalent resistance of the coil and reactances of the coil. The effect of these elements has no effect on the circuit performance. The waveform of oscillation is rectangular when the time constant of the capacitance of the coil is relatively small, and is triangular when the time constant is relatively large. The frequency is governed by the d.c. supply voltage, the grid and screen voltages, and eventually by the values of the reservoir capacitance and the grid-coupling resistance. Predicted frequencies and waveforms are verified by experiment.

621.360.619.13 2182 Frequency Swing with Variable-Reactance Valves—F. Lojewski (Wireless Eng., vol. 29, pp. 140-146 and 167-17; March and April, 1949.) The positive impedance of a reactance tube and its RC circuit, when R is connected between grid and anode, is shown theoretically to be equal to the value of the reactance as rigorously as possible, for the frequency swing and the damping. An approximate equation for the reactance impedance is given and its general case and for a positive sinusoidal input, with a simplified formula, for the frequency swing and with numerical examples. When R is connected between grid and anode, the equivalent impedance is capacitive. Comparison shows that the capacitive arrangement gives a greater frequency swing for a given damping than the inductive connection. Circuits in which the damper is connected to the grid, or, more generally, an unsymmetrical, is discussed, with particular reference to the values of the resistance and the grid-coupling factor, as described by Helfrich (2489 of 1948.) A circuit insensitive to supply-voltage variations is also given.

621.360.619.23 2183 The Serrasolid F.M. Modulator—Day (See 2162.)

621.360.619.23 2184 Non-Linear Effects in Ring Modulators—V. Bedloiti, (Wireless Eng., vol. 26, pp. 72-77; May, 1949.) The operation of a 4-tetering ring modulator is considered for signal voltages of various characteristic impedances of the demodulation product. A geometric representation of the geometric geometric function. Curves are drawn for the lowest products, illustrating the departure from linearity with increasing signal voltages.

621.360.619.23-621.360.615.17 2185 A Modulator Producing Pulses of 10 µ Second Duration at a 1-Mc Recurrence Frequency—M. G. Morgan (Proc. I.R.E., vol. 37, pp. 505-509; May, 1949.) A modulator for use with a spark transmitter, of pulse duration about 0.1 µ second for radar observing equipment. A large 1-µc voltage is applied to the grid of a triode with low internal resistance and a small mutual 1-µc signal is applied to the anode. Suitable plane adjustment results in the production of steep-fronted positive pulses which are applied to the grid of the modulator. The modulating voltage is built up across a capacitor which charges and discharges rapidly. The occurrence of a spark reduces the lead inductance to a low value and assists in producing a rapid drop of the modulating voltage.

621.360.615+621.360.53 2186 Increase of Sensitivity of Amplifier and Mixer Stages for Merle and Decimetre Waves—Strutt (See 2172.)

621.360.645 2187 On the General Theory of Linear Amplifiers: Part I—S. P. Strokov (Izvestia i Tekhniku, vol. 19, pp. 234-244; May and June, 1944.) In a linear amplifier, the input and output are related by a linear differential equation. A general solution is given of the problem of finding such parameters of this equation as will ensure amplification without excessive distortion. Means of obtaining an approximate estimate of the distortion will be discussed in part 2.

621.360.645 2188 Note on the Sensitivity of an Amplifying Stages—K. J. H. Koster (Proc. Radio Soc., vol. 4, pp. 156-157; April, 1949.) Discussion of the grounded anode circuit shows that, at low frequencies, the sensitivity is the same whether the grid or anode is grounded, though at high frequencies the sensitivity of the grounded anode circuit differs slightly from that of the other two. See also 1458 of June.

621.360.645 2189 Amplification of Pulses of Gating Methods—J. A. Bect (Proc. Nat. A. E., vol. 2, pp. 49-49; February, 1949.) Discussion pp. 49-50.) The limitations of conventional methods of pulse amplification are discussed, with reference to (a) the minimum pulse length for which amplification is possible, (b) the use of flat and for internal modulation of by the application of linear gating methods, whose principles are explained, all eliminating ringing and enabling very short pulses to be amplified. Neglecting fluctuations in the setting of the amplifier, the ratio of the input signal to noise ratio obtained by any of the three methods is equal to the ratio of pulse recurrence frequency to the bandwidth of the amplifier. For a linear amplification of double, the amplifier is described for which the input signal to noise ratio showed an improvement of 51.8 compared with the ratio for an amplifier not using gating technique. The application of gating techniques to an amplifier with a damped circuit is considered and methods of locking to the leading edge of the de-modulating pulses are described.

621.360.645 2190 Stabilized Decade-Gain Isolation Amplifier—J. F. Roberts (Proc. I. R. E., vol. 37, pp. 98-100; April, 1949.) An input impedance of over 20.0 MΩ and less than 60-µA current capacity is obtained by employing the input circuit in a way which at least the same amount is used. The potential of the transformer output circuit is the amount, and low dynamic input impedance is achieved by adjusting the input impedance to be connected to it by means of the observation of various characteristic values in high-level circuits. See also 2281 of 1945 (Dunlop)

621.360.645 2191 A Coaxial 50-kW F.M. Broadcast Amplifier—Baltic. (See 2165.)


621.360.645 2193 Stagger-Tuned Amplifiers—J. L. Robs. (Proc. I. R. E., vol. 29, pp. 124-129; March, 1949.) Formulas are derived which enable the calculation of the number of stages and the value of the various circuit components to be calculated directly, without the use of ab, for an amplifier with (a) the minimum number of stages; with a given gain, and (b) at a flat response curve within the pass band as possible.

621.360.645.012.8 2194 Network Representation of Input and Output Admittances of Amplifiers—J. M. U. Ville. and R. V. Tucker (Proc. I. R. E., vol. 37, pp. 162-163; April, 1949.) A clear picture of circuit performance is obtained if these admittances are represented by networks derived from the equivalent circuit of the amplifier, together with certain series or parallel branches whose elements are functions of a such networks are shown.
Abstracts and References

1949

Theoretical and experimental investigations into various aspects of the photoelectric effect. Section 2, by Knoll, in collaboration with E. Kinder, discusses electron optics, with particular reference to crosstalk deflection systems, electron lenses, and the electron microscope.

1953: 412.285

An Electron Tube for Viewing Magnetic Fields—Lutz and Tetenbaum. (See 2375.)

1953: 2208

Nonlinear Theories of the Electromagnetic Field—F. Bertelín. (Roy. Sci. [Paris], vol. 86, pp. 349-356; April 1, 1948.)

1953: 2209


Some fundamental problems of classical electromagnetic theory are outlined and brief reference is made to the work of Dirac and L. de Broglie.

1953: 2209

Recent Trends in Radio Technique—M. Adam. (Tech. Mol. [Paris], vol. 41, pp. 165-165; May 1 to 15, 1949.)

Discussion of formalism and its application to all types of components, and (b) the use of printed circuits in subminiature assemblies. (c) Methods of ensuring satisfactory performance of equipment under extreme conditions of humidity, temperature, and altitude, and (d) component design to withstand shock, vibration, or rapid acceleration.

1959: 2200


Comment on 2766 of 1948.

1959: 2200

GENERAL PHYSICS

1954: 532.41+538.566: 537.228.1


1959: 535.37


1959: 2202


1959: 2204


Approximate expressions are obtained for the potential distribution, maximum current density, and spread of beams of finite thickness. Results are compared with those of Harri (127 of 1943) and of Kofler and Housh (240 of 1940).

1959: 2205

551.510.4  

551.510.52  

551.510.535  

551.510.535  

LOCATION AND AIDS TO NAVIGATION

621.396.9  

621.396.9  
Doppler Radar—E. J. Barlow (Proc. I. R. E., vol. 37, pp. 120–145; April, 1949.)

621.396.9  
Aids to Training—The Design of Radio Synthetic Training Devices for the R.A.F.—G. W. A. Dummer. (Proc. I. R. E., part 111, vol. 96, pp. 101–112; March, 1949. Discussion, pp. 113–116.) The training devices described are of two main types, a bench trainer of simple design, and a complete crew trainer. The former was required as an aid to the introduction of new radar systems; the latter provided accurate presentation of moving targets, complete operational practice, and also error-recording facilities.

621.396.9:371.3  
A Phenomenological Theory of Radar Echoes from Meteors—McKinley and Millman. (Sci., 1949.)

551.510.535  

551.510.535  


Radio Aids for Approach and Landing: Control of Aerial Traffic—A. Violet. (Onr., vol. 29, pp. 91–109; March, 1949.)

551.788  

Radio Thermometer Gauge of Compensation Type—A. Rostagni and L. Filippini. (Sacro Cim., vol. 4, pp. 74–84; February 1, 1947. In Italian, with English summary.)


535.37  
An experimental investigation of film prepared in air by a simultaneous condensation of KI and metallic Ti on a quartz plate. The quenching process can be followed by Sott's theory. The energy of activation $U$ depends on the wave-length of the exciting light and on the concentration of the activator; the higher the concentration, the lower the value of $U$. The co-efficient indicating the degree of binding of the mixture with the crystalline lattice remains constant, for a given concentration of the ac-
Abstracts and References

64: February, 1949.) Methods of preparing UO2 powder and beads are described and test results are given showing the effect of applied voltage and of temperature on the resistance of the material. On the part of its variability and high resistivity, the use of UO2 is likely to be limited to bolometer detectors for infrared radiation; for such a purpose it is necessary to use a very thin layer of the oxide.

666.3:621.315.612 2251 Supersonic Tinning of Aluminum Wires—(J. Inst. Metals, vol. 74, pp. 546–547; April 28, 1949.) Discussion of apparatus for making soldered joints without using flux. A Ni striker is immersed beneath the surface of molten solder in an electrically heated crucible. The striker is made to oscillate at about 18 kc. The wire to be tinmed is held by hand near or touching the striker, and becomes tinmed over the short time in which it comes in contact. For full details see B.I.O. Report No. 1844 (2253 below).


MATHEMATICS

517.5 2254 Remarks on the Harmonic Analysis of Astley Functions—A. Blanc-Lapierre. (Rev. Sci. (Paris), vol. 85, pp. 1027–1040; November 1 to 15, 1947.) Astley functions of the second kind are defined and the method of filters is applied to their analysis. See also 1666 of 1948 (Blanc-Lapierre and Fortet).


517.9:53 2256 Non-Linear Vibrations—M. L. Cartwright. (Adv. Sci., vol. 6, pp. 64–69; April, 1949. Bibliography, pp. 69–74.) Discussion of (a) general methods of solving nonlinear differential equations, including methods involving differential analyzers, (b) special types of second-order equation, (c) difficulties of formulating the equation for a physical problem, (d) standards of accuracy of approximations, (f) relaxation oscillations, (g) numerical and graphical solutions, (h) topological methods, and (i) miscellaneous recent work. The bibliography is arranged according to the above sections, which are mainly a commentary upon it. See also 156 of 1948 (Minorsky).

518.5 2257 Electronic Digital Counters—W. H. Bliss. (Proc. Phys. Soc., vol. 66, pp. 309–314; April, 1949.) Discussion of the circuit consisting of four multivibrators, with modifications that convert it to a decimal system. The associated switching circuit is also considered. Various applications are mentioned.


MEASUREMENTS AND TEST GEAR

531.763 2259 An Electronic Stopclock—K. J. Brimley. (Electronic Eng. (London), vol. 21, pp. 180–183; May, 1949.) Fundamentally, the arrangement is a high-speed mechanical counter in which a Scothophony torque motor is used together with two thyatron inverters. This combination counts the quarter-cycles of a standard oscillator within the interval defined by the operation of two три� of its variables, and the details are included, and the principle of the motor drive is explained.

621.3.018.4(083.74) 2260 Some Electromechanical Methods for Producing Low Frequencies from a Primary Frequency Standard—W. R. C. McKinley, (Cowan. J. R., vol. 27, sec. F, pp. 49–54; February, 1949.) The primary-standard crystal frequency of 50 kc is divided by the conventional chain of multivibrators to 10 kc, 1 kc, 600 kc and 100 kc, the output of 1-kc phonic clock motor, one shaft of which rotates at 1 rpm and carries sector disks or needle disks whose elements generate pulses in the form of an electromagnetic impulse, as they pass between its poles. These pulses are used as gate pulses to select pulses of higher timing precision at the desired repetition rate.

621.317.32+621.317.34:621.396.11 2261 A Simple Method of Measuring Electrical Earth-Constants—R. B. Westwater. (Proc. Inst. Elec. Eng., vol. 96, pp. 141–144; March, 1949.) An incoming ground-wave is received in success on two short sloping antennas of equal length, one in the vertical plane and the other in a vertical plane at right angles to a. The recency of the ground-wave ellipse, and hence the earth constants, are calculated from the electric field and the vector, as they pass between its poles. These pulses are used as gate signals for selecting the timing pulses at various times within the desired repetition rate.

621.317.323(083.74) 2262 A Primary High-Frequency Voltage Standard—(Tech. Bull. Nat. Bur. Stand., vol. 33, pp. 15–44; April, 1949.) A bridge containing a very small thermistor of diameter only 0.015 inch, has been developed for measurements from 20 mv to 1.5 volts at all frequencies from 1 to 800 Mc. Careful description of a special mount for a 2-thermistor arrangement eliminated the need for frequency corrections and reduced the time required to balance the bridge. Up to 200 Mc measurements by means of a cro, thermoelement, or electrostatic voltmeter agreed with the bridge measurements to within 1 per cent. Above 200 Mc the accuracy of available data is limited, and no special precautions are taken for the case of the slotted transmission lines used for power measurement. Up to 50 Mc the voltage range was extended to 10 volts by use of thermistor beads of considerably larger diameter.

621.317.336:621.314.2 2263 Measurement of Transformer Impedance using a Low-Current Bridge Techniques—K. Goldsmith. (Electro. Eng. (London), vol. 115, pp. 522–526; April 21, 1949.) Bridge methods have definite advantages, but if carried out at the usual frequency of 1,000 cps the results obtained may differ so greatly from those which would be obtained at the 50-cps operating frequency that they would give an entirely wrong idea of the transformer performance. By measurement of a suitable bridge, therefore, it can be carried out at the normal operating frequency, using a vibration galvanometer or other suitable detector.

021.317.34
Transmission-Line Impedance Measurement—R. J. Lec, C. H. Westcott and F. Kay. (Wireless Eng., vol. 26, pp. 78 19, (1921). Balance-against devices are discussed in the range 500 to 600 Mc. In the measuring section was constructed, consisting of a length of wire of similar dimensions to the tube, but with a slot along which the measuring could be made. A connection was used instead of a shorting bar. Experiments results discussed include admittance measurements on various slots.

021.317.35
The Optimum Performance of a Wave Analyser—N. F. Ratner. (Electronic Eng. (London), vol. 21, pp. 175 179, (1949)).

021.317.37

021.317.61:021.317.69
Physical Society's Exhibition—(See 2179).

021.317.71:021.37
Electrometer Tubes for the Measurement of Small Currents—Victoreen. (See 2176).

021.317.73:021.317.67:021.629.135
Measurement of Aerial Impedances in Aircraft—P. Vandam, (Onde Élec., vol. 29, pp. 73 78, February). For the aerial impedance to be measured, determined by the voltage obtained by means of a transformer, to be used in the frequency range 9 to 30 Mc. The reactive and resistive components of the impedance are determined by the output of the test equipment. An example shows the results obtained on an 8-meter aerial.

021.317.73
Pulse Excitation of Impedance Bridges—J. G. Yates. (Nature (London), vol. 163, pp. 132 133, January 22, (1949)). Measurements can be made in any practical case with resultant pulses of duration 1 to 10 microseconds and of suitable recurrence frequency; sinusoidal excitation would require frequencies as low as 150 c.p.s., under similar conditions, to make the quantity component in the bridge output negligible. A number of bridges can be excited in sequence by pulses and their output applied to a common amplifier, which can be used for capacitance as well as inductance bridges.

021.317.761
A Compact Direct-Reading Audio-Frequency Meter—A.A.M.K. (Electronic Eng., vol. 22, pp. 108 109, April, (1949)). A cascade amplifier is followed by a squaring amplifier, the output of which was used to trigger a bi-servo oscillograph, the integrated signal current in the final tube driven by the positive half of the oscillation is recorded by a microammeter. The instrument is calibrated and proportioned to the frequency. Maximum readings at the three scales are 1000, 5000, and 10,000 cps.

021.317.761:021.029.58
The Additive Frequency Meter—O. Grammer. (QST, vol. 45, pp. 32 37, May, (1949)). A suitable harmonic of a 100-kc crystal oscillator is fed to a probe together with the output of a variable-frequency oscillator covering a 50-kc range. The sidemodes seen to be generated supply a series of signals that can be used in the signal from an ordinary heterodyne meter. Dead errors are only of the order of 50 cps and are independent of the frequency being measured. Errors due to instability of the variable-frequency oscillator are measured, as well as due to instability of the variable-frequency oscillator. Small measurements can be made in the chart of the frequency range 100 kc. A similar principle has been used at the National Bureau of Standards for the measurements.

021.317.784.088

021.317.79:021.317.61:021.317.69.111
Low-Distortion M. Signal Generator—F. S. Sampson. (Proceeding, vol. 22, pp. 121 120, April, (1949)).

021.317.79:021.317.61:021.317.822
Atmospheric Noise Measurement—I. Reiche. (Electronic Eng., vol. 22, pp. 110 114, April, (1949)). A description is given of a continuous measurement of noise levels shown to be 60 volts per meter over the frequency range 75 kc to 30 Mc. Three conveniently situated wide-band preamplifiers, each covering about 10 Mc, at the frequency range, have antennas attached and feed six receivers through coaxial cables. Each receiver is sampled in turn and the noise is recorded. The design of the first stage of the preamplifiers to give a low noise figure is discussed.

021.317.79:021.317.90.080
Radar Range Calibrator—R. L. Root. (Electronic Eng., vol. 22, pp. 114 117, April, (1949)). Design of an instrument for production calibration of the measuring distance of a PPI. Range ring pulses generated in the radar are compared on a triggered scope with spaced pulses from a crystal-controlled clock.

021.317.690.014:083.75
Climatic and Durability Tests for Radio Components—Engineer (London). (London), vol. 18, p. 393, April 8, (1949)). Brief details of specification No. 10.11, obtainable (price 6s.) from the Radio Industry Institute, London, W.C.1. It covers approximately the same ground as the Inter Service Specification No. RCS/11.

021.317.755

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.321.9.001.8
Investigation of an Ultrasonic Generator—(Electronic Eng. (London), vol. 21, pp. 154, 151, May, (1949)). Designed as a laboratory tool for research workers in industry. An output of 1 Mw can be obtained directly by a silica tube. Interchangeable collets allow the test operation at frequencies around 1, 2, and 3 Mc. The quartz crystal oscillator can be safely immersed in liquids at temperatures up to 150°C.

538.569.2.047:021.315.011.5
Dielectric Properties of the Human Body in the Microwave Region of the Spectrum—R. W. E. England and H. A. Sharpe. (Wireless World, vol. 164, pp. 887 888, March, (1949)). Homogeneous specimens were inserted in a waveguide cell between a metallic flange and a plug of polystyrene. Measured values of the absorption coefficient and phase constant for various body substances are tabulated and discussed.

538.569.2.047:021.380.118
Investigations on the Biological Effects of Microwaves in view of their Therapeutic Application—L. de Seguini, G. Castellini, and M. Pelletier. (Revue Sci. (Paris), vol. 86, p. 334 April, (1948)). Experiments are described which show that certain waves may be, and even stimulate, certain biological processes. They provide a convenient and precise means for the therapeutic application of heat.

538.569.2.047:021.380.118
Exposure of [animals] to Microwaves—W. E. White and W. A. Clark. (Microwave (London), vol. 8, pp. 66 67, May, (1949)). Experiments to find the effect of high-intensity 12-cm radiation on rabbits are briefly described. An intensity greater than 3 watts per cm2 can cause damage to certain tissues, such as those of the eye, without accompanying pain. The most vulnerable parts of the body are those not abundantly supplied with blood.

538.569.2.047:021.380.118

021.179.16
Location of Internal Defects by Supersonics—J. N. W. Deacon. (Instruments, vol. 19, pp. 725 726, December, (1946)). An account of the Sperry supersonic reflectoscope, which provides (a) the wave-guide of frequency continuously variable from 0 to 10 Mc., (b) a 1000-volt pulses of recurrence frequency variable from 0.5 to 12 Mc. and of duration 400,000 or more, which are applied to the crystal of sound image in a test piece through a film of oil. Illustrations are given of the pulse echoes seen on the screen of the oscilloscope when the case of cracks or similar defects in the material under investigation.
538.506


539.7

A New Solution to the Problem of Vertical Angle-of-Arrival of Radio Waves—E. W. Hamlin, P. A. Seay, and W. E. Gordon. (Jour. Appl. Phys., vol. 20, pp. 248–251; March, 1949.) A practical mathematical solution of the case of a signal consisting of direct and reflected waves arriving from different directions. The measurement of the amplitudes and relative phase of the two waves at different positions in the vertical plane gives sufficient information to enable the angle of arrival and intensity of each wave to be calculated. The formulas are applied to 3-cm transmissions over a 27-mile desert path in Arizona.

539.9

Radar Reflections in the Lower Atmosphere—A. B. Crawford. (Proc. I.R.E., vol. 37, pp. 404–405; April, 1949.) Simultaneous radar and visual observations suggest that "angels" are due to reflections from flying insects and birds. For other views, see 2769 of 1947 (Frisi), 722 of 1948 (Gould), and 1761 of July (Gordon).

551.510.52

Ionospheric Absorption and the Calculation of Fields at a Distance—A. Houbert. (Onze Elec., May 29, pp. 152–159 and 216–226; April and May, 1949.) A review of the theories and experimental results of many authors, including the presentation of the principal conclusions and results as a whole.

551.510.535

Ionospheric Absorption and the Calculation of Fields at a Distance—A. Houbert. (Onze Elec., May 29, pp. 152–159 and 216–226; April and May, 1949.) A review of the theories and experimental results of many authors, including the presentation of the principal conclusions and results as a whole.

551.510.535

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551.510.535

Ionospheric Absorption and the Calculation of Fields at a Distance—A. Houbert. (Onze Elec., May 29, pp. 152–159 and 216–226; April and May, 1949.) A review of the theories and experimental results of many authors, including the presentation of the principal conclusions and results as a whole.

2306

Temperature of Variations of Ground-Wave Signal Intensity at Standard Broadcast Frequencies—F. R. Gracely. (Proc. I.R.E., vol. 37, pp. 360–363; April, 1949.) Measurements over six paths of lengths from 76 to 558 miles at frequencies from 630 to 1,170 kc show that variations of ground-wave signal intensity appear to be more closely related to changes in temperature than to changes in any other commonly observed meteorological quantity. The main conclusions are that there is a marked decrease in the intensity at the higher temperatures and that this decrease is approximately proportional to the path length in wavelengths.

2307

Ground-Wave Propagation Across a Land/Sea Boundary—G. Millington. (Nature (London), vol. 163, p. 128; January 22, 1949.) A 77.5 Mc 10-kw transmitter was situated at sea level about 1.4 km south of the Blackwater, Essex. A similar transmitter-receiver was moved toward the shore, across the Blackwater (p. 2.2 km) and beyond the opposite shore. Field-strength readings taken at intervals confirm the marked rise of field strength expected theoretically under certain conditions at a land-sea boundary. See also 1753 of July and 2304 above.

2308

Tropospheric Propagation on Lower Radio Frequencies—D. W. Hitchman. (Nature (London), vol. 163, pp. 527–528; April 2, 1949.) Tropospheric phenomena observed at signal strengths as low as 1 Mc at signal-strength measurements at 59 Mc, 3.58 Mc, 877 kc, 660 kc, and 804 kc are shown graphically and correlated with the observed change in relative humidity per 50-m step of the corresponding Larkhall balloon soundings and the height of this change. Comparable ionospheric sounding records did not account for the variations noted.

RECEPTION

2309

Superregeneration—An Analysis of the mechanism of the device as a whole. (Proc. I.R.E., vol. 37, pp. 500–504; May, 1949.) The effect of a sinusoidally varying damping factor on the behavior of a tuned circuit is considered. The
amplitude and frequency of this variation are the fundamental parameters distinguishing the superregenerator from the ordinary resonant circuit. Sensitivity and selectivity are considered as functions of these parameters. Multiple super-regenerative and other circuit properties are deduced from the solution of the differential equation. See also 3501 of 1948 (Macfarlane and Whitehead).

621.396.621 G.E.C. Model BR740—( "Wireless World," vol. 55, pp. 171-174; May, 1949.) An 11-tube super-regenerative receiver with an integral ac supply unit which can operate from mains voltages between 95 and 180 volts or 195 and 250 volts at 40 to 80 cps. Frequency coverage is 150 to 350 kc to 34 kc in six switched ranges. Six alternative bandwidths between 0.5 kc and 9 kc are provided.

621.396.621-621.396.619.13 2311 The Response of Frequency Discriminators to Pulses—E. F. Grant. (Proc. I.R.E., vol. 37, pp. 387-392; April, 1949.) The time response of a simple shunt resonant circuit is analyzed, and the results are applied to the behavior of the Round-Tracks and Foster-Seeley frequency discriminators. The condition for the discrimination of any one crossover frequency in the desired frequency band is derived.

621.396.621.53+621.396.645 2312 Increase of Sensitivity of Amplifier and Mixer Stages for Metre and Decimetre Waves—M. J. O. Strutt. (Bull. Schweiz. Elektrotech. V., vol. 28, No. 363-371, June 18, 1949, in German, with French summary.) Formulas for the maximum power amplification are derived for narrow and for wide frequency bands in the decimetre-wave region. Theory relative to interference factors is developed for the case of noise given whose application enables such factors to be reduced considerably and in ideal cases even eliminated. Practical application of these rules to deshielding of transmission systems and multigrid mixers results in a reduction of the noise factor of about 15 db.

621.396.622+621.394.16 2313 Rectification—Müller; Sieber; Sachsue—See 2330.


621.396.622.621.621 2315 Noise Figures for Receiver Input Circuits—P. G. Sulzer. (Tele- Tech, vol. 8, pp. 40-42; May, 1949.) The following six circuits are compared from the noise standpoint and suggestions for the proper application of each are made: (a) single-ended grounded-cathode amplifier, (b) push-pull grounded-cathode amplifier, (c) cathode-follower circuit, (d) grounded-grid amplifier, (e) cathode-coupled amplifier, and (f) d'Allesandro amplifier. Circuits (c), (b), (c), (d), and (f) have essentially the same noise figure with similar gain. The cathode-coupled amplifier is definitely inferior to the other circuits. The choice of the best circuit for a given application depends largely upon whether or not the triode tube is to be used. The pentode type of circuit is satisfactory for low-frequency narrow-band applications, but triode circuits are usually preferable for high-frequency wide-band receivers.

621.396.828 2316 Suppression of Electrical Intercourse to Highly Inductive Apparatus in Naval Vessels—A. Hunter. (Proc. IEE, part 1, vol. 96, pp. 159-165; March, 1949.) Shielding and bonding, and internal and external suppression are considered for the range 10 kc to 150 Mc. Details and performance of x-type filter boxes with air or dust-core chokes rated up to 150 amperes at 220 volts are given. Virgin suppression of the circuit is obtained by using coupling capacitors. The use of x-networks in the internal bus-bar leads of machines is advocated if shunt capacitance is inadequate.


621.398.101.01 2319 Theoretical Limitations on the Rate of Transmission of Information—W. G. Tidale. (Proc. I.R.E., vol. 37, pp. 468-478; May, 1949.) A theory is developed which takes account of code noise effects. The transmission of a quantity of information II over a given circuit is governed by the relation

$$H = \frac{1}{2} \ln (1 + C/N)$$

where B is the transmission-link bandwidth, T the ratio of noise or current and C/N the carrier-to-noise ratio. For large signal-to-noise ratio B, N, this formula leads to

$$S/N \leq (C/N)^{1/2} f$$

where f is the channel bandwidth. Coded transmission is realized with the following capabilities of the general system: B, N, S, f, C, N, and (B/N). The inefficiency of existing communication systems is discussed and the advantages to be gained by the removal of internal message correlations and by analysis of the information content of a message are mentioned. The theory is applied to radar relays, telemeters, voice communication systems, and communication circuits, etc. See also 1907 of 1947 (Gabor) and 1949 of July (Shannon).


621.396.111.01 2321 A Note on the Theory of Communication—J. D. Weston. (Phil. Muc., vol. 40, pp. 449-453; April, 1949.) An approximate quantitative theory is suggested. A coded message is represented as a vector in a space of an infinite number of dimensions; the process of transmission over an ideal signal channel is equivalent to a projection of this vector on to a subspace. A message will be accurately transmitted if, and only if, it is coded so that its associated vector lies entirely in the subspace characterizing the signaling system.


621.395.305.2 2323 Automatic Change-Over to an Emergency Apparatus in a Communication System—Hepp. (Philips Tech. Rev., vol. 8, pp. 310-314, October, 1946.) Two methods for automatic change-over to an emergency oscillator when the output signal falls below a predetermined amplitude are discussed. Where low-frequency amplifiers are involved, a constant auxiliary signal outside the band of the signal to be amplified is added to this auxiliary signal by means of a change-over to the emergency amplifier.

621.396.619.16 2324 Pulse Communication Systems—S. Van Michel. ("IEEE (Brussels), nos. 1 and 2, pp. 16-25 and 45-51; 1949.) In French.) The principal pulse-amplitude, pulse-position, and pulse-code systems are reviewed and discussed with particular reference to bandwidth and signal-to-noise ratio. Different types of distributors, modulators, and demodulators are mentioned; two-pulse-position systems and two commercial equipments are also discussed.

621.396.619.10 2325 Signal-to-Noise-Ratio Improvement in a P.C.M. System—A. G. Clavier, P. F. Pantier, and W. Dite. (Proc. I.R.E., vol. 37, pp. 355-359; April, 1949.) The output signal-to-noise power ratio (expressed in decibels) is approximately twice the square root of the signal-to-noise power ratio, and is independent of the number of code digits provided this large is enough. The distortion due to quantization varies greatly with the number of code digits and is found to be proportional to the number of digits for which the output noise power is equal to the distortion power for a given input signal-to-noise ratio.

621.396.65 2326 Directional Transmission Investigations in the Alps—W. Klein. ("Tele-Tech, vol. 27, pp. 49-69; April 1, 1949.) In German.) A detailed account of experiments carried out between Monte Generoso at the southern end of Lake Lugano, and the Jungfraujoch, using wavelengths of 15 cm and 2 meters, and with 90-watt transmitter power. Repeatability is found to be due to quantization variances greatly with the number of code digits and is found to be proportional to the number of digits for which the output noise power is equal to the distortion power for a given input signal-to-noise ratio.

621.396.69 2327 Radio-Telephony at Whitemoor Marshalling Yard—(Engineer, London), vol. 187, pp. 227-228; March 25, 1949.) A 92-kw two-way 85-45-Mc system having a fixed 12-watt transmitter-receiver station in the control tower and mobile 12-watt receiver units in the engine cars, where 12-volt batteries are fitted.

621.396.97 2328 Broadcasting at the 5th Olympic Winter Sports, St Moritz—F. Dupuis. ("Tech. Mitt. Schweiz. Telegr.-Telephon. Ver.," vol. 26, pp. 258-263; December 1, 1948.) In French.) Details of the arrangements for Switzerland and also of the international connections with many European countries and with the United States. Altogether 359 transmissions were arranged, and 116 in Switzerland and the remainder abroad, the total duration of the transmissions being 273 hours. See also 3243 of 1948 (Wettstein).
SUBSIDIARY APPARATUS

621.526:061.3 2320


621.314.6-621.306.622 2330


Part 1, by Möller, discusses tube rectification and heterodyne reception, with particular reference to Döbler's method of rectification (which is that in which the total voltage is applied to the rectifier by means of Barkhausen oscillations). Part 2, by Seiler, on detectors, outlines the Schottky theory of blocking-layer and point rectifiers and describes the synthesis and measuring of the rear effect, wave-sensitive, low-resistance detector materials and also the construction of the Telefunken detector Type R705.

Part 3, by Sachs, briefly reviews work on tube rectifiers, for high frequency.

621.316.7 2331

Application of the Method of Logarithmic Frequency Characteristics to the Investigation of the Stability of Monitoring and Regulating Systems and to the Estimation of their Efficiency—V. V. Solodovnikov. (Automatica i Telemechanika, vol. 9, pp. 144-147; March and April, 1948, In Russian.)

621.316.72 2332

Carrier Communication Level Regulator—W. S. Chakin. (Electronics, vol. 22, pp. 104-107; April, 1949.)

621.316.722 2333


The operation of the usual type of electronic-ionic voltage regulator (Fig. 1), consisting of an excitation variable feedback coupling and a voltage generator and a measurement element, is discussed. Equations for various circuits are given and it is shown that while the use of oscillographic measurements and the time constant of the regulator increases the speed of the system and reduces the speed of the regulation process, a discussion of the circuit equations shows that it is impossible to use high amplification voltage generators and measurement elements, as discussed. A high voltage feedback coupling and a low amplification in the measuring element, Design formulas and curves for the feedback coupling circuit are given.

621.316.722:621.3.013 2334

Rectifier Voltage Control using Saturable-Core Transformers—F. Butler. (Wireless World, vol. 55, pp. 227-229; June, 1949.) The principle is outlined and different methods of working the reactors are discussed. A circuit diagram and performance figures are given for a full-wave Hg-vapor rectifier with reactor control; the output voltage remains between 970 and 1000 volts for a current range of 0 to 600 mamp, while the voltage change for a current range of 100 to 400 mamp does not exceed 1.5 per cent.

621.316.726 2335

An Electronic Frequency Regulator—L. S. Bruk, S. S. Chugunov, and N. V. Pautin. (Automatica i Telemechanika, vol. 9, pp. 144-147; March and April, 1948, In Russian.)

A description of a regulator employed to control the frequency of a 400-cps oscillator feeding a circuit analyzer. The regulator uses a tuning fork as its frequency standard and its accuracy is within 0.1 per cent. A circuit diagram is given with values of the components, and the operation is discussed in detail. Experimental curves are also included.

621.316.725.078:621.397.6 2330


Circuit diagrams are given and discussed for three types of arc system, namely: (a) a saw-tooth system in which the saw-tooth is formed from the pulse present across the deflection yoke, and the phase of this saw-tooth is compared with that of the synchronizing pulse to control the sweep circuit, (b) a sinusoidal system in which a stable oscillator is controlled in phase and frequency by the synchronizing pulse, and in turn controls the sweep circuit, or (c) a pulse system in which the area of the synchronizing pulse, which is changed by phase variations, is used to develop a control voltage.

621.319.3 2337


Discussion of the energy loss in electronic machines at the commutator and due to gas friction leads to the conclusion that for the production of electric energy it is superior to electromagnetic generators. Although great progress has been made recently in the design of electronic machines, they are not likely to supercede electromagnetic generators for very high power.

TELEVISION AND TELEPHOTOGRAPHY

621.397.21.3 2338


Drawings, printed matter etc., up to 22 cm wide and of any length and electronically "stuck" on an endless belt and scanned by a rapidly rotating optical system. A document of quarto size can be transmitted in 8 seconds. The image signals may be sent over either cable or radio links. At the receiving end, positive or negative reproductions, of one-sixth the size of the original, are "written" on film, which can be rapidly processed to provide enlarged prints.

Resolution and linearity of the reproduction are discussed and applications are suggested and comparative advantages of the system assessed.

621.397.24 2339


Transmission line in closely populated areas, particularly in blocks of flats, could be improved by using a common antenna and a local wire distribution system. In the system described, the complete carrier and sidebands of the transmitted programs are received, amplified, and distributed over concentric cables. The amplifier, of which a circuit diagram is included, has a uniform gain of 55 db between 42 and 48 Mc. Input voltage to receivers is between 7.5 mvols and 0.75 mvols for not more than 30 receivers on each line at distances up to 480 meters.

621.397.26 2340


Reprint of 2055 of August. See also 1203 of May.

621.397.331.2 2341

High-Speed Production of Metal Inseec-scopes—H. P. Steier and R. D. Faukner. (Electronics, vol. 22, pp. 81-83; May, 1949.)

New techniques used in the manufacture of the RCA 16-inch metal-cone cathode-ray tube Type 16 AP4.

621.397.5 2342


A survey of the difficulties of obtaining bandwidths in excess of 40 Mc in the various elements of a television service. Video amplifiers, if amplifiers, discriminators, FM klystrons, and microwave antennas having the required performance are discussed. The limiting factors are considered to be transmitter output-stage bandwidth, propagation irregularities, and the cost of the domestic receiver.

621.397.50(83.74) 2343


A review leading to the conclusion that the logical solution of the problem lies in the adoption of a standard of 945 lines, with a video-frequency bandwidth of 15 Mc.

621.397.50(83.74) 2345


621.397.5083.73 2346


Discussion of the various factors which led to the selection of the 819-line standard for France.

621.397.6:621.396.05 2347

Apparatus to-Schenectady Television Telev—F. M. Deerlake. (Electr. Eng., vol. 68, pp. 419-422; May, 1949.)

For an earlier account see 1792 of 1948.

621.399.6-182.3 2348

The WOW-TV Television Field Car—J. Herold. (Communications, vol. 29, pp. 12-13; April and May, 1949.)

An illustrated description.

621.399.6-182.3 2349

Mobile TV Studio for WDTV—W. M. Crenshaw. (Electronics, vol. 22, pp. 20-21; March, 1949.)

An illustrated description.

621.397.645 2350


Design equations are derived and illustrated for
several types of rf amplifier stage for television receivers, including a cathode-coupled amplifier, and for several types of mixer. Emphasis is placed on the problem of obtaining the optimum signal-to-noise ratio while satisfying gain, bandwidth, and adjacent-channel rejection requirements.

621.397.7

Low-Cost TV Operation — G. W. Ray. (Em-Tv, vol. 9, pp. 24–27; March, 1949.) The video signals of selected programs from New York television relayed by a microwave link to New Haven, the relay station being located on Oxford Hill, 8 miles from the New Haven transmitter. Audio signals are transmitted by telephone. Reception in the New Haven area is quite satisfactory. Equipment is described.

621.397.8(494)

First Practical Tests of Television Reception in Switzerland—J. Dufour. (Tech. Mitt. Schweiz. Telecom. Teleph. Verw., vol. 26, pp. 241–249; December 1, 1948. In French, with German summary.) An account of trials carried out in and near Zürich during the 20th Swiss radio exhibition, August 26 to 31, 1948, in which the first television demonstrations were given, by Phillips-Lampe, and others. Subjective and length measurements of picture quality were made at many points up to a maximum distance of 16 km from the 80-kW transmitting antenna, which was installed on the Feldberg, about 110 meters above the center of town. The results obtained are tabulated and discussed. For field strengths > 4 mv per meter in Zürich, the picture was generally good, but reception was not possible for fields < 0.7 mv per meter. The most common interference was that from car ignition systems, but some industrial high-frequency generators were troublesome; one located several km from the transmission point reportedly received reception in its neighborhood quite impossible, as the video frequency used was 61.4 Mc. An analysis of the results shows that a 26-kW transmitter on the Feldberg should give good reception over the whole of Zürich.

621.397.8(73)


621.397.92

TVI Patterns—C. G. (QST, vol. 33, pp. 430–435; May, 1949.) Photographs are reproduced and discussed which show the interference to television caused by a 28 Mc amateur transmitter. This interference is not detectable by various refined measures.

621.397.823

Ignition Interference—V. V. Calendur. (Wireless Eng., vol. 26, p. 116; March, 1949.) To reduce ignition interference with television sound, it is given as important to reduce the number of pulses in the train of radio-frequency waves applied to the field radiated. See also 3741 of 1946 (Leedsfield) and 1779 of July (Pressey and Ashwall).

621.397.828

Further Advances in T.V.I. Suppression—J. Varnes. (R.G.M.R. Bull., vol. 24, pp. 268–273; May, 1949.) It was found possible to alleviate various commercial television receivers close to a 14-Mc transmitter when suitable harmonic-suppression devices were used in the transmitter. Several devices and methods of suppression are discussed in detail. A harmonic monitor is described.

621.397.5

Television [Book Review]—M. G. S. Roger. Blackie and Son, Glasgow, 2nd edition, 77 pp., 60s. (1949, 55p. 234, June, 1949) " ... a very simple and lucid explanation of how television works ... an excellent introduction to television."
Abstracts and References

548–556; May, 1949.) A general survey of methods used. Typical mount-inspection service, the use of statistical control charts, and sampling procedures are discussed.

621.385: 538.122

2375

An X-ray Tube for Viewing Magnetic Fields—S. G. Lutz and S. J. Tretenbaum. (Elect. Eng., vol. 67, pp. 1143–1146; December, 1948.) The development of the special tubes is discussed. The construction and operation of theoretical constants, and finally to an equation which defines implicitly the propagation constant and thus indicates the waves which can be propagated in the tube in particular cases are considered. The practical importance of the traveling-wave tube is considered. Results are similar to those for plane electrodes. The theory may be useful for the development of cylindrical tetrodes and pentodes, especially transmitting tubes.

621.385.032.29

2386

Nonlinear Distortion and Noise in a Diode acted upon by U.H.F. Signals—Y. I. Kuznetsov. (Bell. Acad. Sci.U.S.S.R., pp. 1173–1189; September, 1949.) Equations are derived for the current in the circuit of a plane diode for the most general initial conditions using a method similar to that proposed by Müller (1933 Abstracts, Wireless Eng., p. 433). In passing over from electron equations to current equations, a different method from that proposed by Benjam (148 of 1939) is used, a clearer physical interpretation of the theory is thus achieved. From the current equations, general equations are derived determining the nonlinear distortion occurring in the multimode signal of signals, intermodulation of signals and noise, and the effect of the transit time of electrons on noise. The cases of weak and strong signals are treated separately. The results obtained can be applied to multi-electrode tubes which can be regarded as consisting of a number of diodes.

621.385.2: 621.396.822

2387

Measured Noise Characteristics at Long Transit Angles—G. A.C. Humphreys. (Wireless Eng., vol. 26, pp. 383–386; April, 1949.) Tests on diodes at 3,000 Mc indicate that the space-charge reduction of noise is of the order of 10 to 1 when the transit time exceeds a given value. The diodes indicate that moving tungsten, thoriated-tungsten, or oxide emitters. The observed magnitude of the noise and its variation with transit time agrees qualitatively with theory.

621.385.3

2388

Carrier Distribution in Triodes Neglected in Space Charge and Initial Velocities—L. C. Hammer. (Appl. Sci. Res., vol. B1, no. 2, pp. 77–104; 1948.) A theory of current distribution, originally due to Davies (Wireless Eng., p. 505) is clarified and developed for positive-grid triodes. A graphical method of checking the applicability of this theory is discussed; the different distribution functions which are involved in the equations can easily be determined from the graphs given. In some cases theory and experiment are in agreement; discrepancies occurring in other cases are discussed. The basic assumptions underlying the theory are examined in the light of the experimental results.

621.385.3: 621.396.619.13.029.64

2389

The grid short circuits the magnitude of the $r$ voltages in such circuits. Mechanical detail, design, and performance are discussed for a particular case in which a frequency deviation of 5 Mc was obtained.

New Microwave Tube.—(Electronics, vol. 22, pp. 177-177; April, 1949.) Description of a coaxial type, close-packed planar triode, Type BTI 1553, for operation at 4,000 Mc.


Generation of Oscillations—Uurt; Gundlach; Frey; Schumann; Marx; Hettner. (See 2175.)

Increasing the Power Output of Vacuum Tubes—B. M. Hudeld. (Radio News, Radio- Electronics Eng., Suppl., vol. 10, pp. 10-11; May, 1948.) The circuit of a three- or four-element triode-tetrode, connected for use as a triode or diode, can be arranged so that a large part of the anode dissipation is transferred to an external resistance in the reduced diode characteristic. The conditions governing the maximum value of $R$ and the reduction in the anode dissipation obtainable are discussed. Application to vacuum-tube stabilizers of the cathode-follower type is considered.

The Choice of Operating Conditions for Resistance-Capacitance-Coupled Pentodes—F. Langford-Smith. (Radiotronics, pp. 63-69; July, 1948.) A discussion of (a) the optimum value of the anode load resistor for minimum distortion, (b) the optimum anode current for minimum distortion under given conditions, and (c) the relative distortion characteristics of a pentode and a triode for given output voltage, and (d) pentode operating conditions.

Variation of the Input Impedance of Television-Amplifier Pentodes—F. Juster. (Thé. Franc., pp. 23-26, 36; April, 1949.) Discussion of the dependence of input impedance variations on frequency and on tube characteristics. Results of measurements performed on specified tubes are graphically shown. Methods of reducing such variations are indicated.

Electron Emission and Electron Currents—Mayer; Knoll. (See 2050.)

Cathode-Ray Tubes with Post-Deflection Acceleration—W. G. White. (Electronic Eng., London, vol. 23, pp. 197-200, 1949.) The outstanding advantage of post-deflection acceleration is the increase in brightness obtainable for a given accelerating voltage. One form of p.d.a. electron is a band of graphite on the inner circumference of the cathode-ray-tube envelope near the fluorescent screen. Such several electrodes can be used. P.d.a. causes a slight reduction in sensitivity, and the electric fields near the p.d.a. electrode lose their radial symmetry. Various quantities are comparatively tabulated for p.d.a. and ordinary types of electron guns with electrostatic, magnetic, and electromagnetic deflection, but no general formula expressing the over-all advantage of a p.d.a. tube can be deduced. The work of Pierce (1965 of 1941) is critically discussed.

Modern Vacuum-Pump Design—Mellen. (See 2241.)


Generation of Oscillations—Urt; Gundlach; Frey; Schumann; Marx; Hettner. (See 2175.)

Transistor-Tube Characteristics—(Electronic, vol. 22, p. 128; March, 1949.) Satisfactory results are obtained when the two points contacts on a Ge crystal plate are placed on opposite faces instead of the same face as in previous designs. The coaxial mounted contacts rest in polished spherical depressions in the disk. Improved mechanical stability, complete electrostatic screening between input and output, and easier construction is claimed.

Transit-Time Determination of Space-Charge Reduction of Shot Effect—D. K. MacDonald. (Phil. Mag., vol. 40, pp. 561-566, May, 1949.) With a Ge point contact in a tube, the emission current from the cathode exhibits less fluctuation than in the absence of space charge. If this current drifts for some distance, as in a v.m. tube, it is to be expected that the fluctuation will increase until, after a sufficient time, the full shot noise is reached again. Analysis of this problem leads to a result which shows the progressive increase of noise with drift time.

Radio Valve Data [Book Notice]—Ilfie and Sons, London, 80 pp., £a.6d. (Wireless Eng., vol. 26, p. 84; March, 1949.) The characteristics of 1,000 British and American receiving tubes are presented. The charts are useful for a v.m. tube. It is to be expected that the noise will increase until, after a sufficient time, the full shot noise is reached again. Analysis of this problem leads to a result which shows the progressive increase of noise with drift time.

MISCELLANEOUS

The Radio Research Board.—(Wireless Eng., vol. 26, pp. 145-146; May, 1949.) A historical review of its work (a minor part of which is the preparation of these abstracts) and of its relationship with the Department of Scientific and Industrial Research.

National Radio Conferences.—R. L. Smith-Rose. (Nature (London), vol. 163, pp. 493-495; March 26, 1949.) A general survey of the main conclusions of various conferences held in the summer of 1948.

Proposed Standard Frequency-Band Designations.—(Proc. I. R. E., vol. 37, p. 467; May, 1949.) Discussion of a system in which "Band n" includes all frequencies from 10 $n$ cps up, but not including, 10$(n+1)$ cps. Standard abbreviations for frequency and length units are also listed. For an alternative system see 2413 below.

Nomenclature of Frequencies—C. E. Booth. (P.O., Elec. Eng. Jour., vol. 42, part 1, pp. 47-49; April, 1949.) A new classification is proposed in which frequencies between 0.3X 10$^8$ cps and 3X10$^8$ cps are designated as the "Band n." This is capable of unlimited extension. See also 2412 above.

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3. Insulator's high dielectric strength permits breakdown test at 1000 volts R. M. S. Dust and dirt can't get in.

4. Stop, of cup design, provides superior switch shielding... gives you excellent torque strength without distortion.

5. High grade laminated phenolic base maintains high insulation resistance (under humidity conditions.)

6. Contact Spring gives you double wiping contacts on both resistor and center terminal ring... it accurately formed to maintain uniform pressures and minimize noise.

7. Electro tin-plated terminals provide soldering ease. Tightly crimped, terminals give you direct contact to resistor... assure constant contact under humidity and soldering conditions.

8. Resistor is made of special resistance material bonded to high quality phenolic for smooth operation, low noise level, outstanding humidity characteristics.

9. Cadmium tipped center terminal provides easy soldering... good shelf life without oxidation. Adequately lubricated for good rotation life, center terminal is finished to give you smooth take-off... minimum noise.

10. Laminated phenolic base maintains high insulation resistance (under humidity conditions.)

11. Cadmium plated steel ground plate assures positive grounded cover.

12. Cadmium plated steel bushing is accurately finished and fit to shaft for smooth rotation.

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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
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ATLANTA

Baltimore
Inspection Tour of Television Station WAAM; June 26, 1949.

BRAHMA-POR-ARTHUR

COLUMBUS

"Television Transmitting and Microwave Pickup," by C. Sloan, Chief Engineer, Radio Station WLWC, Tour of Radio Station WLWC; May 25, 1949.


DENTON-AMES

HOUSTON

LOS ANGELES

MILWAUKEE


"Application of Proximity Effects in Induction Heating," by E. Bennett, Faculty of University of Wisconsin; May 4, 1949.


"Audio-Frequency Transformers and Their Functions in High-Fidelity Amplifiers," by D. Schwennensen, Chief Engineer, Chicago Transformer Corporation; May 31, 1949.

(Continued on page 40A)
Do you know

**HOW THESE 9 FACTORS AFFECT THE QUALITY OF DISC REPRODUCTION?**

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<th>INTERMODULATION DISTORTION?</th>
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<td>STYLUS FORCE?</td>
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<td>4</td>
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<td>UNWANTED OUTPUT?</td>
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<td>9</td>
<td>NOISE PICK-UP?</td>
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The reproducing equipment must provide the correct frequency compensation for the recording characteristics most commonly used. Since different recording companies use widely varying characteristics, a correspondingly wide choice of equalization characteristics must be available.

A choice of scratch equalization is also necessary to meet the surface noise conditions of all records. "Rolloff" of reproducing curves must permit maximum scratch reduction while retaining as much as possible of the original material on the record.

The signal-to-noise ratio must not be impaired by induced noise pick-up in the reproducer or equalizing circuits. Design of the equalizer and repeating coil should minimize hum pick-up from motor fields or other sources.

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**DISTRIBUTORS: IN THE U.S.A.—Graybar Electric Co.**

**IN CANADA—Northern Electric Company, Ltd.**

Graybar Electric Co.
420 Lexington Avenue,
New York 17, N. Y.

_Gentlemen: Please send me a copy of Bulletin T2351, "109 Reproducer Group."

Name__________________________

Title__________________________

Company_______________________

Street Address__________________

City____________________________

State__________________________

PROCEEDINGS OF THE I.R.E. September, 1949
CRYSTAL PICKUP CARTRIDGES

for High Fidelity Phonograph Reproduction...

for those applications where

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new metal "Big Picture" tubes by Rauland!

AVAILABLE WITH NEW HIGH-CONTRAST LUXIDE SCREEN
Gives 60% greater contrast! Reduces glare! Gives sharper, easier-to-view pictures!

50% LIGHTER WEIGHT
Can be safely shipped installed in sets, reducing field installation costs.

LARGER SCREEN
105 square inches useful screen area

LOWER PRICE
Lower bulb cost permits lower prices

BETTER FACE QUALITY
Optical quality of metal tube faces is superior to the pressed faces of all-glass tubes.

SHORTER LENGTH
2 3/8" shorter than the 16AP4

USES AVAILABLE OPERATING COMPONENTS
Same focus and deflection coils as used with the 16AP4

allows improved cabinet design and lower cabinet cost
Because of reduced cabinet depth requirement

Other Available Types—10BP4, 10FP4, 12LP4, 12KP4, 16AP4

WRITE FOR TECHNICAL BULLETINS

"Perfection through Research"

THE Rauland CORPORATION
4245 N. Knox Avenue • Chicago 41, Illinois
“WANT TO KNOW WHY I’M SOLD ON SORENSEN?”

“Voltage regulators — even electronic regulators — are not all equally accurate! And I know ACCURACY is important! The Sorenson electronic voltage regulator gives me the kind of accuracy I want.

“But I don’t want to buy accuracy and costly maintenance at the same time. The Sorenson Regulator is a strong rugged instrument designed to give accuracy at no sacrifice to low-cost-of-maintenance.

“Furthermore I like my instruments simple — well designed — because I know that a complex instrument loaded with added components can mean poor basic design — and inferior performance. The Sorenson Electronic Voltage Regulator is a beauty for simplicity.”

“That’s why I’m sold on Sorenson”

Get highly stabilized DC regulation with the NOBATRON and B-NOBATRON.

STANDARD AC SPECIFICATIONS

<table>
<thead>
<tr>
<th>Model in VA Capacity</th>
<th>150</th>
<th>500</th>
<th>250</th>
<th>1000</th>
<th>2000</th>
<th>3000</th>
<th>5000</th>
<th>10000</th>
<th>15000</th>
</tr>
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<tbody>
<tr>
<td>Regulation Accuracy</td>
<td>± 0.1% against line or load</td>
<td></td>
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<tr>
<td>Harmonic Distortion</td>
<td>5% max.</td>
<td>5% max.</td>
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<tr>
<td>Input Voltage</td>
<td>95-130 VAC; also available for 90-260 VAC single phase 50-60 cycles</td>
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<tr>
<td>Output Voltage</td>
<td>Adjustable between 110-120; 220-240 in 230 VAC models</td>
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<tr>
<td>Load Range</td>
<td>0 to full load</td>
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<tr>
<td>P.F. Range</td>
<td>Down to 0.7 P.F; all 5 models temperature compensated</td>
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NOTE: REGULATORS CAN BE HERMETICALLY SEALED

The ORIGINAL SORENSEN CIRCUIT is easily adapted to meet your special requirements. SORENSEN engineers are always available to solve any unusual problem not handled by the STANDARD SORENSEN LINE. JAN requirements can be met by all models.

Write for detailed “Sorenson Regulator Performance Chart”*

*Copyright 1949

Sorenson and company, inc.
375 Fairfield Ave., Stamford, Conn.
The CYCLO-TROL® Register is the latest addition to the well-known line of Cyclotron Impulse Registers. The same principle of operation which has gained for these registers such wide use and recognition is applied in this new unit to provide accurate control over a wide range of mechanical cycles.

The CYCLO-TROL Register has two calibrated dials which can be instantly set by means of shaft thumbscrews to any number from 0 to 10,000. When pulsed by an external circuit, the CYCLO-TROL continues to register until the preset number of counts is reached. At this point, CYCLO-TROL's output circuit is completed and a contact is made to external circuit, thus actuating, as desired, operation under control.

The CYCLO-TROL can be reset to original setting by merely pressing the button on top of register. By this simple step, repeat cycles of control can be secured as many times as desired.

**APPLICATIONS OF CYCLO-TROL REGISTER**

The CYCLO-TROL Register is made available because of insistent demand from users of other types of Cyclotron Specialties Registers. Here are only a few of the many applications of this new unit—

★ Counting problems involving positive, accurate control over any number of revolutions or cycles up to 10,000.

★ Electrical circuits may be opened or closed at any predetermined number of counts.

★ Ideal for coil winding machines. The exact number of turns can be preset and machine stopped at exact point, making possible any number of identical coils. Operator needn't watch counter...his attention can be concentrated on winding.

**SPECIFICATIONS AND SPECIAL FEATURES**

<table>
<thead>
<tr>
<th>Feature</th>
<th>Specification</th>
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<tbody>
<tr>
<td>Counting Rate</td>
<td>60 impulses per second maximum</td>
</tr>
<tr>
<td>Power Source</td>
<td>115 volts A.C.</td>
</tr>
<tr>
<td>Power Supplied to Impulse Contact</td>
<td>110 volts D.C. — self-contained</td>
</tr>
<tr>
<td>Output Circuit</td>
<td>50 volts D.C. (direct or to auxiliary relay)</td>
</tr>
<tr>
<td>Dimensions</td>
<td>7&quot; x 4&quot; x 4&quot; high</td>
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<tr>
<td>Weight</td>
<td>5 pounds (approx.)</td>
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**CYCLO-TROL REGISTER**

The CYCLO-TROL® Register is made available because of insistent demand from users of other types of Cyclotron Specialties Registers. Here are only a few of the many applications of this new unit—

★ Counting problems involving positive, accurate control over any number of revolutions or cycles up to 10,000.

★ Electrical circuits may be opened or closed at any predetermined number of counts.

★ Ideal for coil winding machines. The exact number of turns can be preset and machine stopped at exact point, making possible any number of identical coils. Operator needn't watch counter...his attention can be concentrated on winding.

**Specifications and Special Features**

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**Cyclotron Specialties Company**

MORAGA 10, CALIFORNIA

**Price**

$195

**Made by the Manufacturers of these Famous Impulse Registers**

**PROCEEDINGS OF THE I.R.E.**

September, 1949
New Yeoman Transformer gives you ADC Quality at LOW COST

Designed to meet the needs of engineers, experimenters and amateurs who demand high quality at low cost, the new ADC Yeoman line provides many of the well-known performance standards of the Quality Plus and Industrial series, also several items not previously offered. This has been accomplished primarily by improved production engineering methods, standardization of parts and a simplified type of construction.

The ADC Yeoman line includes:
- Output Transformers with carefully balanced windings offering unusually low distortion over a wide frequency range.
- Interstage Transformers with balanced humbucking features providing equal push-pull grid voltages at high audio frequencies for inverse feedback circuits.
- Power Transformers limited to 55°C temperature rise and especially quiet in operation.
- Replacement Units for Audio and TV circuits, miniatures, filament transformers, reactors, and many others.

ADC invites your critical appraisal of this new Yeoman line.

Audio Develops the Finest

Send for the new ADC catalog which you will find convenient to use in selecting almost any transformer you may need. Special requirements not covered by the catalog will receive prompt attention.

ADC
Audio DEVELOPMENT CO.
2855 - 13th Avenue SOUTH, MINNEAPOLIS 7, MINN.

(Continued from page 42A)

Ingram, K. R., Director Telecommunications. Box 48, Nassau, Bahamas.
Keyes, J. C., 414A Nimitz Ave., Chula Lake, Calif.
Kubicek, W. G., Dept. of Physical Medicine, Medical School, University of Minnesota, Minneapolis 14, Minn.
Lambert, C. O., 358 B Crane Walk, Akron 11, Ohio
Neelands, L. J., 117 Parkside Ave., Syracuse, N. Y.
Neubauer, J. R., 3118 Dillon Ave., Cheyenne, Wyo.
Oehmke, W. P., Cauca Postal 2726, Rio De Janeiro, Brazil
Shah, K. L., 124 Sri Krishnamurti Road, Basavangudi, Bangalore, South India.
Toufexis, S., The Delman Company, 105 E.
13th St., New York 14.

The following admissions to Associate were approved and are effective as of August 1, 1949:

Arnold, A. Jr., 6026 Madison St., West New York, N. J.
Banning, W. P., 4166 St. Catherine St., Montreal, Que. Canada.
Beiser, P. D., 532 Thatchers Forest, Ill.
Bell, W. L., R. D. 1, Coatesville, Pa.
Bergman, L. C. H., Government Radio and Telephone Administration, Willemsed, Caracas, Netherlands West Indies.
Bod, D. B., General Delivery, Hilliards, Pa.
Bryant, R. M., Jr., Pines Lake Rd., Box 307, Paterson, N. J.
Buck, D. T., 37-41 Mary St., Freehold, N. J.
Dula, J. E., 10217 Ave. M, Chicago 17, Ill.
Edens, R. L., 1015 S. Fifth St., Waco, Tex.
Evans, R. E., Apartado Postal No. 1106, Bugot, Columbia.
Fujimoto, J. J., 5327 W. Winthrop Ave., Chicago 40, Ill.
Gatch, C. E., Box 444, St. George, S. C.
Goddard, C. A., 6517 Kimbark Ave., Chicago 37, Ill.
Gudzin, M. G., 3259 Queenstown Dr., Mt. Rainier, Md.
Hall, P. C., 5331 W. Congress, Chicago 44, Ill.
Hall, M. E., 43 Leon St., Boston 15, Mass.
Harris, R. R., Taylor Instrument Co., 95 Ames St., Rochester 1, N. Y.
Hodkinson, W. S., 58 Adella Ave., West Newton, Mass.
Holt, C. H., Box 1372, Wright Patterson A.F.B., Dayton, Ohio.
Hooper, E., 1120 19 St., S., St. Louis, Mo.
Inanc, A. D., 606 W. South St., Angola, Ind.
Kilpatrick, L. L., 3981 Menlo Ave., Los Angeles 37, Calif.
King, R. L., Box 313, Lockbourne A.F.B., Columbus 17, Ohio.
Knox, R. M., 818 B S. Catalina Ave., Redondo Beach, Calif.
Kofer, E. J., 1409 Franklin St., Racine, Wis.
Lombard, W. A., 2500 Lalla St., Beaumont, Tex.
Marchetta, P., ASW Branch, Aerial Nads, Johnsville, Pa.
Marhine, E. A., 2173 N. 70 St., Wauwatosa 13, Wis.
Marsh, S. V., 1308 W. Rosedale St., Chicago 40, Ill.
Matter, J. T., 1724 Patton Dr., S. Erickson 7, N. Y.
Mentia, O. N., 2 Royal Hotel Jubilpie C. P., India.
Mohan, M. A., Tech. Officer, Civil Aviation Dept., Radio Development Unit, Factory Rd., New Delhi, India.

(Continued on page 47A)
An ideal radio instrument for laboratory frequency measuring

The new Collins 51J-1 communications receiver is a double conversion superheterodyne of such extreme accuracy and stability that it is admirably suited for use in the laboratory as a dependable secondary frequency standard.

The 51J-1 is permeability tuned throughout. It is continuously tunable over a frequency range of 0.5 to 30.5 megacycles. This range is divided into 30 bands of 1,000 kc each. The tuning mechanism is based on a decade system in which the megacycle figure is set by means of a band switch. The 100 kc figure is indicated on the slide rule dial and the kilocycle figure on the circular dial. Under normal operating conditions and with a 10-minute warmup, the dial reading is within 2 kc of the receiver’s exact frequency throughout the frequency range. Dial accuracy is improved by means of a crystal calibrator and dial corrector which are included.

Frequency over the temperature range —20°C to +60°C does not vary from the frequency at 20°C by more than 30 parts per million plus 1 kc; thus the frequency stability is within 2 kc at the highest operating frequency. Frequency does not vary more than 100 cycles from the frequency at 115 line volts when this voltage is varied through the range 105 to 125. Changes in atmospheric pressure from sea level to 10,000 feet altitude, relative humidity from 10 to 90%, and mild shock, do not vary the frequency of the 51J-1 by more than 500 cps.

This new time and labor saving instrument is also an excellent all around communications receiver of advanced design, with outstanding operating characteristics. We will be glad to give you more complete information on request.

Collins 51J-1. 0.5 to 30.5 mc radio receiver. Normally furnished for rack mounting.
FLEXIBLE SECTION.

PHASE SHIFTER.

TUBE.

WAVEGUIDE.

MICROWAVE PLUMBING & ACCESSORIES

MICROWAVE ANTENNAS

MICROWAVE PHASED ARRAY ANTENNA.

SEARCH ANTENNA.

AN/TPS-15 ANTENNA.

AN/FPW-2 ANTENNA.

MARINE RADAR.

SO-1 AND SO-8 RADAR SETS.

TEST EQUIPMENT.

TRANSMIT.

SLOTTED "PEEL" REFLECTOR.

As Shown.

AP-15 ANTENNAS.

AP-43 ANTENNA.

RELAY SYSTEM.

TOY.

yor TELEGRAPHIC ANTENNA.

10 CM. 30 DEGREE, 115 VAC. DRIVE.

SO-13 ANTENNA.

DEM ANTENNA.

AP-12/AFR.

AP-2 CMF RFT HEAD COMPLETE WITH HARD TUBE.

AP-17 ANTENNAS.

CABLE.

RF EQUIPMENT.

LIGHTHOUSE ASSEMBLIES.

Lightwave Cables with Line of Sight and Type N CPLG. To Rev. Uses 2C40, 2C42.

Transmitter: RT-077.

Cable: 2000 to 6000 mc. 115 VAC.

Beacon light house cable 10 cm. with minor 20 volt DC FM motor. Magnetism.$64.50.

AP-11/AFR.

AP-16.

SO-12.

SO-IX.

X.

AP-13.

AP-12.

AP-11.

AP-10.

AP-6.

AP-4.

AP-3.

AP-2.

AP-1.

Cable: Cmprov. Ph. Diggby 4124.

WRITE FOR LATEST FLYER.

ALL MERCHANDISE GUARANTEED.

PROCEEDINGS OF THE I.R.E.

September, 1949

131 "I" Liberty St., New York, N.Y.
COMMUNICATIONS EQUIPMENT CO.

131 "I" 8th Liberty St., New York, N.Y.
General Electric announces a new line of Permafil d-c paper-dielectric capacitors designed especially for operation in high ambient temperatures. They require no derating for temperatures up to 100°C and can be used up to 125°C.

Hermetically sealed in metallic containers, these new units are available in case styles 61, 63, 65 and 70, as covered by Joint Army-Navy Specifications JAN-25-C, in ratings of 0.10 to 4.0 muf 600-, 1000- and 1500-volts. Permanently sealed silicone bushings are provided on all types.

Permafil capacitors were developed to provide suitable components for the many new applications involving continual operation at ambient temperatures above 85°C—another example of capacitors “designed for the job” by G-E engineers. For further information on these or on capacitors for other applications, write Capacitor Sales Division, General Electric Company, Pittsfield, Mass.
Some transfers to the Associate grade were approved to be effective as of June 1, 1949:

- Albin, F. L., 44 Totten Pl., Babylon, L. I., N. Y.
- Babitz, H. B., Philco TECH. R.E., BOQ Bldg.
- Bakker, T. J., 3417 Tibbet Ave., New York 63, N. Y.
- Babbitt, H. B., University of Illinois, Champaign, Ill.
- Berman, S. J., 50 Academy St., Poughkeepsie, N. Y.
- Bixler, W. V., 3417 Tibbet Ave., New York 63, N. Y.
- Blum, L. J., 2115 Avenue B, Schenectady, N. Y.
- Bollinger, J. R., 3243 Sepulveda Blvd., Los Angeles 34, Calif.
- Bonham, R. L., 368 S. 21st St., St. Louis, Mo.
- Brabson, R. L., 718 First St., Temple, Tex.

The reliability, convenience, dependability . . . those are the "unseen" qualities built into Amphenol RF Cables and Connectors.

And it's those "extras" . . . combined with a tough, highly resistant vinyl outer jacket and a continuously solid and uniformly high dielectric . . . that give lasting performance with minimum loss and interference.

Whatever the requirements—whether for one small part, or a million complete assemblies—nowhere are results more apparent . . . more economical . . . more important than in the use of Amphenol RF Cables and Connectors.

Components and COMPLETE ASSEMBLIES

Amphenol Connectors and Cables are designed in many types of construction. Thus, where connectors are to be permanent, Amphenol is ready to supply complete light-weight assemblies or harnesses with molded connections and RF Connectors, eliminating separate costly parts.

Rugged, compact, providing unsurpassed performance, each component in the assembly gives uninterrupted service and positive protection against all weather.

Catalog D-1 is a ready reference to the regular line of Amphenol RF cables and Connectors. Write on business letterhead to Department H for your copy.
WESTINGHOUSE RESEARCH
in TELEVISION
PHYSICISTS-ENGINEERS

Specialists in OPTICS, ELECTRON-OPTICS, PHOSPHORS, PHOTO-SURFACES, SYSTEMS and CIRCUITS needed for an expanding program at the Westinghouse Research Laboratories, East Pittsburgh, Pa.

For information write:
Manager, Technical Employ.
Westinghouse Elec. Corp.

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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.

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THE SPERRY CORP.
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16th and PARK ROAD N.W.
WASHINGTON 10, D.C.
Approved for Veteran Training

Radio and Radar Development and Design Engineers

Openings for experienced men at HAZELTINE ELECTRONICS CORPORATION
Little Neck, L.I., N.Y.

Please mail complete résumé of experience with salary desired to:
Director of Engineering Personnel

(All inquiries treated confidentially)

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.

ELECTRONICS TEACHER

Electronics Teacher to take charge of electronics option at accredited state land grant college in Northwest. Salary to $4800.00 for nine months. Write giving picture, education, experience, references, and complete personal data to Box 372.

ELECTRICAL ENGINEER—PHYSICIST

Graduate electrical engineer or physicist experienced on microwaves and servo-mechanism for design and layout of electronic circuits. Write Chance Vought Aircraft, Division United Aircraft Corp., Box 5907, Dallas, Texas.

INSTRUCTOR

There will be an opening in September for an instructor to teach electronics, transmission line and waveguide theory. Salary depends on qualifications and is up to $5000.00 for nine months. Possibility of later appointment at lower rank and salary.

ELECTRONIC ENGINEERS—RESEARCH ANALYST

Research and development positions open at Army Security Agency, Washington 25, D.C., in the field of electronics and mathematics. B.S. degree in appropriate field from accredited university necessary. Salaries for electronics engineers—P-1, $2074.80 to P-3, $4779.60 per annum. Add $500.00 per annum for those qualified in mathematics. Positions are permanent in so far as the Agency is concerned. Those qualified communicate immediately with Chief, Army Security Agency, CSGAS-61, The Pentagon, Washington 25, D.C.

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.

TECHNICAL WRITER


ENGINEER

Excellent opportunity for a man with experience in servo mechanisms and cir-
Positions Open For

PHYSICISTS

SR. ELECTRONIC ENGINEERS

Familiar with ultra high frequency and micro wave technique.

Experience with electronic digital and/or analog computer research and development program.

Salaries commensurate with experience and ability. Excellent opportunities for qualified personnel.

CONTACT:

C. G. Jones, Personnel Department

GOODYEAR AIRCRAFT CORPORATION

Akron 13, Ohio

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

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 of Project Engineer caliber with experience in the design of electronic circuits (either pulse or c.w.), computers, or precision mechanical instruments. 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

POSITIONS NOW OPEN

PHD—Physicist—Infrared, Optics

MS—Electronic Engineer—Servos

BS—Electronic Engineer—Circuits

ME—Designer—Small Mechanisms

Have you considered the advantages of working for a smaller, growing Company?

1. Closer contact with management. YOUR abilities quickly recognized!
2. Quicker delegation of responsibility. YOU are in position to develop!
3. Diversity of problems. YOUR job has greater interest!
4. Faster advancement. YOU have opportunities!
5. No complex wage structure. YOUR salary determined by ability!

If you are alert to these advantages, please send résumé of qualifications to

Personnel Director

SERVO CORPORATION OF AMERICA

2020 Jericho Turnpike

New Hyde Park, New York

(Continued from page 50A)
THESE 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 C" needle—with either one or three-mil 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 CQ1 fits standard 1/2" mounting and RCA 45 RPM record changers. Model CQ-1J fits RCA No. 2 Specifications for top mounting 453" mounting centers. (Needle pressure five grams. Output 0.7 volts at 1,000 c.p.s. Employs one-mil tip radius, Q-33 needle. Cast aluminum housing.

3 LQD Double-Needle Crystal Cartridge

The LQD Cartridge—for 45, 78 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 pry 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.

Positions Wanted

(Continued from page 51A)

BROADCAST ENGINEER

Specializing in broadcast station construction 20 years experience in broadcasting and aviation radio. History and picture sent on request. Go anywhere. T. L. Kild, P. O. Box 860, Fredericksburg, Texas.

SENIOR ENGINEER

B.S.E.E. 8 years experience Army electronics training at Harvard, and M.I.T. Research at Radiation Laboratory, M.I.T. Radar project officer, Aircraft Radio Laboratory, Wright Field. Postwar electronic research engineer. Also broadcast installation, operation, and maintenance experience. Married, children, require family housing Prefer west, southwest, midwest. Box 287 W.

JUNIOR ENGINEER


ELECTRONIC ENGINEER

M.E. expected August 1949 Polytechnic Institute of Brooklyn. B.S. Cooper Union. Age 24. 2 years as electronic technician U.S.N. 11/2 years design and development of radar receivers and microwave components. Prefer position in vicinity of New York City. Box 290 W.

 Lockheed Aircraft Corporation

(Continued on page 56A)

LOCKHEED AIRCRAFT CORPORATION has positions available for RESEARCH ENGINEERs in PHYSICS and ELECTRONICS.

These openings are for men who have achieved advanced status in researchers in either of these fields and who possess the following qualifications:

1. Advanced degree (preferably Ph.D.);
2. 7-10 years industrial or graduate research experience;
3. Ability to perform mathematical analysis in field of specialization.

Lockheed's research and development facilities are among the most advanced in the aircraft field. The accomplishments of Lockheed's research staff have resulted in the acquisition of new far-reaching scientific studies.

Immediate openings available. Salary commensurate with ability.

If you are interested, and meet the qualifications listed above, send a resume of your training and experience to—

LOCKHEED AIRCRAFT CORPORATION Employment Manager Post Office Box 551 BURBANK, CALIFORNIA
Fully in keeping with the trend towards larger, direct-viewing tubes originally pioneered by Dr. Allen B. Du Mont—and also the lower price range for higher grade TV offerings.

Type 16FP4 is a 16-inch magnetic focus and deflection television picture tube designed to give high brilliance and sharp definition. Electron gun design utilizes a bent electrode structure to be used with a single external magnet for the elimination of ion spot blemishes. The exclusive Du Mont screen depositing technique assures the longest pleasurable usage.

Detailed Specifications on request. Let us quote on quantity requirements.

CHECK LIST OF 16FP4 ADVANTAGES...

✓ All glass! No mounting problems.
✓ A mass-produced standard TV tube for maximum value at minimum cost.
✓ Overall length of only 20¼ inches.
✓ Deflection angle: 62°.
✓ Maximum diameter: 16¼ inch ± ¼ inch.
✓ Bent-gun ion trap requiring but a single magnet.
✓ Accelerating potential: Maximum 16 KV; (Design Center Value).
✓ New type small shell duodecal 5-pin instead of 7-pin base, for use with economical half-socket.
✓ Ideal compromise between large picture size and moderate tube cost.

ALLEN B. DU MONT LABORATORIES, INC.
PRECISION Transformers
FOR TODAY'S MORE EXACTING REQUIREMENTS
POWER -- AUDIO CHOKESS -- FILTERS

For Television and all other applications where specifications are precise and the emphasis is on quality and performance, famous FERRANTI transformers offer superior value.

Into each unit goes long years of specialized experience, plus up-to-the-minute knowledge of today's improved practices and latest materials. Our large and varied stock of patterns, tools, and dies often permits us to supply "custom" requirements from standard parts, effecting worthwhile savings. We invite your inquiries.

FERRANTI ELECTRIC, INC.
30 ROCKEFELLER PLAZA
New York 20, N.Y.
No matter what your panel instrument problem is, Simpson Electric Company engineers will be glad to help you solve it. Every day they are confronted with individual design problems.

Behind every Simpson instrument is a world-wide reputation for quality. Simpson movements have greater ruggedness and accuracy, because of the full bridge-type construction and soft iron pole pieces.

When Simpson helps you with your problem, you benefit from this world-wide reputation and the years of experience of Simpson engineers.

Let Simpson engineers help you with your next instrument problem and for your standard instrument requirements take advantage of our large stock, available for immediate delivery.
Positions Wanted

(Continued from page 54A)

TECHNICAL WRITER—COMMUNICATIONS ENGINEER


ELECTRONIC PHYSICIST


ELECTRONIC ENGINEER

Stanford University, B.S.E.E., communications major. Age 24. 1 year broadcasting, 2 years navy electronics experience. 1st class license. Prefer position in broadcasting or electronic development and production. Box 296 W.

SALES ENGINEER

Desires selling or distributing products of electronic industries. Graduate of R.E.A. Institute, education at Bowdoin College, Harvard and M.I.T. in radio engineering. Holder of all FCC commercial radio licenses. 7 years experience in industry. Presently on staff of a state college. Box 297 W.

SALES ENGINEER

B.S. in G.E., Naval Academy. One year P. G. in applied communications. Five years Naval experience as radar, radio and communication officer. Telephone, electrical design and production experience. Extensive background in electrical and electronic equipment. Widely traveled with positive aggressiveness and firmly settled on a sales career. Age 32. Married. Box 311 W.

TELEVISION ENGINEER

B.S.T.E., May 1949. American Television Institute of Technology, age 29, married. 3 years communications maintenance U.S. Marine Corps. 2 years U.S. Border Patrol radio laboratories. 7 years bench and field electronics. 1st class FCC license. At present engaged in post graduate work on micro-waves. Desires position in electronic laboratory or TV station engineer. Locate anywhere. Resume upon request. Box 312 W.

TECHNICAL WRITER

B.S.E.E. September 1949. Lawrence Institute of Technology. Honor graduate with aptitude for technical writing. Managing editor college newspaper, instructor in radio electronics in Air Corps. Box 313 W.

RADIO (BROADCAST) ENGINEER

B.S.E.E. 1939. Over 5 years' experience in broadcast work includes AM, FM, Television, 50kw transmitters. Last position chief engineer construction large station with complex directional antenna. Signal Corps radio staff officer during war. Box 314 W.

(Continued on page 54A)
DC AMPLIFIER

This new moderately-priced DC Amplifier combines a great number of desirable features. Designed for minimum overshoot, rather than maximum band width, it is, nevertheless, useful for steady-state operation at frequencies up to 1.5 megacycles. The response to steep wave fronts is excellent, as indicated by the above photographs. Balanced or unbalanced outputs may be used as desired. There is no need to compensate for a fixed dc voltage between output terminals and chassis, since the output is at ground potential. An undistorted voltage swing of 220 volts peak-to-peak makes it possible to obtain large patterns on a cathode ray tube connected directly to the amplifier output. On the other hand, a peak power of a watt into 6000 ohms, together with a low effective output impedance, make the amplifier useful for driving a variety of devices. An exceptionally well-regulated power supply furnishes plate power to all stages except the cathode-follower output, and heater power to the first three stages. This, together with careful circuit design, reduces the drift to the order of 0.5 millivolts per hour, after a 30-minute warm up. A low capacity input cable and probe make connection to driving circuits convenient, and reduce the danger of feedback between external elements connected to input and output. This amplifier offers the circuit engineer a new tool of unusual scope and versatility.

Voltage Gain: 6000.
Rise Time: 0.4 microseconds.

Stability: Drift is usually less than 0.5 millivolts per hour referred to the input, after 30-minute warm up.

Voltmeter and Calibrating Signal: A zero center voltmeter of ±15, 50 and 150 volts is provided for measuring dc output. Associated with this meter is a calibrating signal circuit for checking the gain.

Output Balanced: Internal impedance of 450 ohms. 220 volts peak-to-peak into a high impedance. Maximum undistorted output into 6000 ohms, 500 milliwatts ac, or 1 watt dc.

Output Unbalanced: Internal impedance of 450 ohms. 110 volts peak-to-peak into a high impedance. Maximum undistorted output into 3000 ohms, 250 milliwatts ac, or 500 milliwatts dc.

Input: Unbalanced, 1 megohm shunted by 40 μf maximum. This includes capacitance of the flexible input cable.

Gain Control: Continuously variable over a range of 11 db, plus 10 db steps.

Shock Proofing: The first two stages are mounted on a spring suspended chassis, offering excellent protection against microphonics.

Cabinet Dimensions: 11" high x 17½" wide x 13½" deep.

Power Supply: Approximately 300 watts of 60-cycle power is required.
Positions Wanted

ENGINEER

MS Communications Engineering Harvard University, February 1948. BS Physics, Harvard 1944. Married, 2 years communications officer U.S. Navy. 1 year experience electronic design. Desires position in New York area, electronic controls, computers or microwaves. Box 315 W.

ELECTRONIC ENGINEER


ELECTRICAL ENGINEER

B.E.E. January 1949, City College of New York. Several months experience in psycho-acoustics and medical electronics in leading medical institutions. Will consider any offer in other fields. Box 317 W.

ELECTRONIC ENGINEER

Communications, B.E.E., recent graduate, ranked ninth, former Navy electronic technician; design, development, and research preferred. Anywhere in the U. S. Salary secondary. Box 318 W.

COMMUNICATIONS ENGINEER

B.S.E.E., December 1948, Michigan State College; age 24, married 1 year electrical test equipment maintenance in Navy; experienced in audio and mobile high frequency communication. Desire position in development in west or southwest United States. Box 319 W.

ADMINISTRATIVE ENGINEER

Schools: Harvard, NCE University of Dayton, Columbia and M.I.T. Ex-flying officer, industrial engineer with wide background in electronic research, development and production. Substantial experience in military UHF and aircraft development. Presently assistant to President of medium size concern. Desires industrial management or government contractual management position. Box 320 W.

JUNIOR ENGINEER

Graduate, B.S.E.E., June 1949. Age 26. University of Illinois 4 years varied electronics experience with emphasis on radar and audio. Prefers electronic design and development position in New York City area. Box 324 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 58A)

RECENT CATALOGS

- The current issue of Traceolog, which always contains articles of general interest to those engaged in utilizing radioactivity for research or industrial control problems, with a product listing and various other office addresses of Tracerlab, Inc., 55 Oliver St., Boston 10, Mass.

(Continued on page 60A)
The homing pigeon goes to sea

Now science gives the navigator an improved "homing pigeon instinct," a way which—without checking the sun or the stars—he can head his ship directly home.

Already thoroughly proved, Loran equipment has been simplified through RCA research and engineering, so that almost anyone can learn to use it in a few minutes. Free of human error, readings appear directly on the instrument. A quick check gives position.

Brain of this Loran system is a circuit developed at RCA Laboratories which splits seconds into millions of parts—and accurately measures the difference in the time it takes a pair of radio signals to travel from shore to ship.

Given this information, the navigator, hundreds of miles from shore, can determine his position quickly and accurately. Loran's simplicity adapts it to every type of vessel from merchant ship to yacht. Manufactured by Radiomarine Corporation of America, a service of RCA, it is already being installed in U. S. Coast Guard rescue ships.

The meaning of RCA research

RCA's contribution to the development of this new direct-reading Loran is another example of the continued leadership in science and engineering which adds value beyond price to any product or service of RCA.

The newest advances in television, radio, and electronics can be seen in action at RCA Exhibition Hall, 36 West 49th St., N. Y. Admission is free. Radio Corporation of America, RCA Building, Radio City, N. Y. 20.
New FM and TV Sweep Signal Generator

A new Model 8G, a sweep signal generator with a range from 0-227 Mc with no band switching, has been designed and developed by Transvision, Inc., 385 North Ave., New Rochelle, N. Y.

The Model 8G has a sweep width from 0-12 Mc completely variable, and has a calibrated built-in marker generator.

The dial is calibrated in frequency. There is sufficient voltage output to permit stage-by-stage alignment. The crystal controlled output makes possible any crystal controlled frequency from 5-230 Mc. Unmodulated rf signal provides marker pips simultaneously with the main variable oscillator; these markers can be controlled as to output strength in the pin oscillator.

Magnetic Tape Recorder

A new magnetic tape recorder for studio and broadcast station use is being offered by Presto Recording Corp., P.O. Box 500, Hackensack, N. J., plant, Paramus, N. J.

A desirable feature of this recorder is the accessibility to all working parts. The entire top panel hinges upward, so that the lower side may be easily serviced. Both (Continued on page 61A)
WHO SAID
IT'S SPECIAL? Not BUD!
THAT'S WHY YOU GET MORE...SAVE MORE WITH BUD.

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)

recording and playback amplifiers, as well as a separate regulator power supply, are vertically mounted on the front door of the cabinet. Access to both front and back of the amplifiers can be had without removal.

Keyboard-Type Oscillator
The Model 150-AO-1/100K Keyboard oscillator, improved and redesigned, is at present being marketed by Weinschel Engineering Co., Dept. 1, 123 William St., New York 7, N. Y.

This instrument covers a range from 0.3 cps to 100 kc with decades of push buttons. In both lower ranges frequency is selected with four decades of push buttons, and in the top range with three decades together with a continuous control which covers a deviation of approximately ± 1 per cent.

Between 100 and 100,000 cps, frequency may be varied in steps smaller than 0.1 per cent. Between 10 and 100 kc in steps smaller than 1 per cent, however adjustment to 1 cps is obtainable.

Description: 19 inches long, 17½ inches high, and 13 inches deep. Weight, 29 lbs.; power supply, 105–125 volts, 50–60 cps.

Improved AF Measurement Equipment
The Model GA-1002A, sound pressure measurement equipment with major improvements which include a specially isolated socket tip which effectively separates the microphone clamping structure from the preamplifier extension tube, is being manufactured by Massa Laboratories, Inc., 3868 Carnegie Ave., Cleveland 15, Ohio.

The M-101 standard microphone which is part of the equipment has also been redesigned to provide increased polar symmetry of the mechanical structure, resulting in exact acoustic symmetry about the normal axis of the microphone even at frequencies above 20 kc. The range of the

(Continued on page 62A)
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 61A)

instrument is such that sound pressures from less than 1 dyn/cm² to 20,000 dynes/cm² (160 db level) may be directly measured. A multiplier is available for extending the upper range to 200,000 dynes/cm² (180 db level). A built-in calibrating circuit permits setting the system gain so that 1 millivolt/dyne/cm² is delivered across the 10,000 ohm output circuit.

New Interlocking Relay

A new interlocking relay, series 30500, consisting of two Type 2 Phil-trol relays, is the newest addition to the relays offered by Phillips Control Corp. 612 N. Michigan Ave., Chicago 11, Ill.

(Continued on page 61A)
Kenyon Fits Your Production To A "T"

Kenyon "K"s—high quality, uniform transformers, are your best bet for development, production and experimental work. For over 20 years, the KENYON "K" has been a sign of skillful engineering, progressive design and sound construction.

Now—reduce inventory problems, improve deliveries, maintain your quality—specify KENYON "T's," the finest transformer line for all high quality equipment applications.

New Catalog Edition! Write Today! Kenyon new modified edition tells the complete story about specific ratings on all transformers. Our standard line saves you time and expense. Send for your copy of our latest catalog edition now!

KENYON TRANSFORMER CO., Inc.
840 Barry Street • New York 59, N.Y.

MEASUREMENTS CORPORATION

Peak-to-Peak VOLTMETER

.0005—300 VOLTS

MODEL 67

Designed for accurate indication of the peak-to-peak values of symmetrical and asymmetrical waveforms, varying from low frequency square waves to pulses of less than five microseconds duration.

.0005-300 volts peak-to-peak, .0002-100 volts r.m.s. in five ranges. Semi-logarithmic, hand calibrated scales.

Provision for connection to 1500 ohm, 1 milliampere graphic recorder or milliammeter.

INPUT IMPEDANCE: 1 megohm shunted by 30 mmfd.
DIMENSIONS: Height 7½", width 7", depth 8½". Weight 8 lbs.
POWER SUPPLY: 117 volts, 50-60 cycles, 35 watts.

MEASUREMENTS CORPORATION

BOONTON ★ NEW JERSEY

Sigmund Cohn Corp.
44 Gold St. • New York
Since 1901

PROCEEDINGS OF THE I.R.E. September, 1949
The SKL type 202 Wide-Band Chain Amplifier employs a new and unusual principle of amplification: the traveling wave circuit. This new design makes possible a bandwidth of from 100 KC to 200 MC with a gain of 20 db. Because of its low impedance — 200 ohms — existing coaxial cables can be used. Because of its flat response — ± 1.5 db — and low standing wave ratio — less than 1.5 db — the Model 202 offers new advantages in general laboratory measurement. The very fast rise time of the Model 202 is invaluable in oscillography, nuclear instrumentation and television applications.

The SKL Spencer Kennedy Laboratories, Inc.
185 Massachusetts Ave., Cambridge 39, Mass.

The Tektronix Type 511-AD Oscilloscope has been the choice of the leaders in television research as well as in TV transmitter and receiver design, production and maintenance.

CHECK ITS FEATURES
We are confident that you, too, will find utility and usefulness in the Type 511-AD normally expected only from instruments considerably higher in price.

**TELEVISION?**

The Model 80 is an automatic slide-back type, and provides direct peak voltage readings of positive or negative pulses of widths from 0.25 microsecond to 20 milliseconds at repetitive rates from 10 to 50,000 pps, provided the duty cycle is under 50 per cent. Ranges of 10, 50, 100, 500, 1,000, and 5,000 volts are provided. Accuracy is 3 per cent full scale on any scale.

For detailed information consult H. Langstroth at C. G. S.

**RECENT CATALOGS**

• • • A new "Commercial Radio Operators Q & A Manual," by Milton Kaufman, containing the questions used by the FCC in the April 1949 examination, with their answers, is to be released in August, by John F. Rider, Publisher, Inc., 480 Canal St., New York 18, N. Y.

• • • A comprehensive catalog, #300, containing the most up-to-date information on the electronic and electrical components and their costs of Cambridge Thermionic Corp., 445 Concord Ave., Cambridge 38, Mass.

(Continued from page 65A)

Interlocking relays may be furnished to operate on either ac or dc. When the lock-up relay is energized, it automatically locks in mechanically by means of a tension spring catch, which holds the armature in the energized position even though the circuit has been opened.

When the second relay, known as the release relay, is energized, the lock-up relay is automatically released from its mechanically held position. Other features are: relays may be equipped with as many as 12 springs; dc relays may be provided with copper slug coils, making them slow either to operate or release; contacts may be supplied either single or twin, and with contacts of various precious metals having ratings as high as 6 amperes, 110 volts ac, noninductive; relays may operate over a wide pressure range. A new descriptive catalog in color gives coil characteristics, contact assembly, and weight.

**Peak Pulse Voltmeter**

The new Model 80, a Peak Pulse Voltmeter that is capable of readings at pulse widths under 1 microsecond and at 10 pps, has recently been designed and developed by C. G. S. Laboratories, Inc., 36 Ludlow St., Stamford, Conn.
ANALYZE
COMPLEX
IMPEDANCES
WITH THE
Z-ANGLE
METER

APPLICATIONS...
- Loudspeakers
- Microphones
- Transmission Line Filters
- Amplifier Inputs and Outputs
- Resonant Circuits—series or parallel
- Transformers
- General Laboratory Measurements

FOR MEASUREMENTS OF...
- Impedance (Z) 0.5 to 100,000 ohms
- Phase Angle (θ) 0° (X) to 90° (R)

D = 0.1 to 8.1
Q = 10 to 0.1

FREQUENCY 0.1 to 200,000 c.p.s.

PRICE $425.00

Write today for bulletins on other T.I.C. products: R.F. Z-Angle Meter ... R.F Power Oscillator ... Translatory Variable Resistors ... Slide Wire Resistance Boxes ... Phase Angle Meter.

CO-AX AIR-SPACED ARTICULATED R.F. CABLES

LOWEST EVER CAPACITANCE & ATTENUATION

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We are specially organized to handle direct enquiries from overseas and can give IMMEDIATE DELIVERIES. U.S.A.

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VARIAN MICROWAVE ENGINEERING

WAVEGUIDE TEST equipment for use between 2600 and 3950 mc; 1½ by 3 by 0.080 in.; RG-48/U waveguide with UG-53/U flanges. These and special units for early delivery.

A Standing-wave detector. Precision ground for continuing accuracy better than 1 per cent.
B Variable attenuator. Attenuation 05 to 10 db; vswr less than 1.1, 2600 to 3400 mc; average power 1 watt, peak 1 kw.
C Termination. Average power 1 watt, peak 1 kw; vswr less than 1.05, 2600 to 3400 mc.

VARIAN ASSOCIATES

99 Washington st.
San Carlos, Calif.
Of particular interest to all who need resistors with inherent low noise level and good stability in all climates

HIGH VALUE RANGE

10 to 10,000,000 MEGOHMS

This unusual range of high value resistors was developed to meet the needs of scientific and industrial control, measuring and laboratory equipment—and of high voltage applications.

SEND FOR BULLETIN 4906

It gives details of both the Standard and High Value resistors, including construction, characteristics, dimensions, etc. Copy with Price List mailed on request.

Cramer Type SX Synchronous Motors are highly efficient permanent magnet type motors that produce an exceptionally high torque. Self-starting... quick start and stop... operate at synchronous speed only. Close tolerance of magnetic field construction and precision alignment of gear train assure long, uninterrupted service.

APPLICATION

Designed for applications requiring a constant speed at a given frequency, Cramer Type SX Motors are widely used throughout the instrument and control fields to fill the gap between the low torque clock motors and the fractional horsepower group.

CHARACTERISTICS

High Torque: 30 in. oz. at 1 RPM, 60 cy.

Quick Response: Reach synchronous speed within 1/2 to 2 cy. Stop within 1 pole of motion on 240 RPM rotor shaft (1/60 sec.). Speeds: Standard gear trains from 60 RPM through 1/24 RPH. Cycles: Available for 25, 50 and 60 cy. operation.

Coils: Easily replaceable. Lubrication: Sealed within housing containing rotor and gear train.

SEND FOR BULLETIN 10B
THE R. W. CRAMER COMPANY, INC.
Box #12, Centerbrook, Conn.
News—New Products

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

Low Power Transmitter
Mica Capacitors

A new line of transmitter mica capacitors with universal mounting, which are intended for use in low power transmitters for plate or grid coupling, filament, and plate by-pass applications, is announced by Cornell-Dubilier Electric Corp., South Plainfield, N. J.

This line, the Faradon NF series, is similar to Type 9 except for different case style and mounting arrangement. The dimensions of these capacitors are 1 1/4"x1 1/2"x1 1/8" over all, with a choice of a vertical mounting with insulated mounting holes tapped for 3/32" screw, or a horizontal mounting with brass terminal bushings tapped for a 3/16" screw. The latter type has insulated slotted mounting holes and insulated spacer feet to permit assembly against a chassis. Solder lug terminals and brass terminal bushings tapped for 5/64" screws provide an optional method of making connections. They are rated in a range from 0.00005 μF with 2,500 volts dc to 0.03 μF with 600 volts dc.

New Pin Riveter For Light Operations

Keller Tool Co., Grand Haven, Mich., announces production of a Pin Riveter, for very light riveting operations, with soft-metal tubular and standard rivets.

The new Pin Riveter is said to be suited to a variety of special jobs, such as setting small drive screws, driving brads in the assembly of wood, light peening and scalping operations on thin sheet-metal sections and bakelite.

The net weight of the Pin Riveter is 13 ounces; length, 6 1/16 inches. Piston diameter is 19/32 inch; stroke, 3/8 inch. Tool has a speed of 9,000 blows per minute. Standard equipment consists of one blank rivet set, with special rivet sets available on order.

WRITE FOR DESCRIPTIVE BULLETIN

WELLER
MANUFACTURING COMPANY
821 PACKER STREET • EASTON, PA.

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 5/4" length and new RIGID-TIP mean real soldering efficiency.

Chisel-shaped RIGID-TIP offers more soldering area for faster heat transfer, and new design gives bracing action for heavy jobs. Here you 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. Order from your distributor or write for bulletin direct.

SOLDERING TIPS—get your copy of the new Weller guide to easier, faster soldering—20 pages fully illustrated. Price 10c. at your distributor, or order direct.
News—New Products

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

(Continued from page 67A)

Vertical Lead Shield for Radiation Protection

A new vertical lead shield, Model 800, that incorporates new designed principles, has been developed by Atomic Instrument Co., 160 Charles St., Boston 14, Mass.

The Model 800 is steel-cased, aluminum-lined, and, the manufacturer claims, offers at least 1/2 inch equivalent of lead shielding in all directions. The top is of aluminum and lead construction and, while removable, can be securely locked. The connector opening, located near the top of the cylinder, is aluminum-lined.

The interior is equipped with a lucite holder and is specially designed to be completely light-proof and background-proof.

Dimensions are as follows: inside, 4 inches diameter by 8 inches high; outside, 9 inches by 14 inches; door opening, 3 1/2 inches by 4 inches; weight, 250 lbs.

New Microvolt Signal Generator

The new Model 292X, signal generator, a microvolt generator designed to cover both upper channel TV and mobile band frequencies on fundamentals, is announced by Hickok Electrical Instrument Co., 10551 Dupont Ave., Cleveland 8, Ohio.

The manufacturer states that its major use will be in the coverage of mobile band frequencies for taxicabs, police departments, railways, ships, etc., for which no expanded scale instrument with accuracy to 0.05 per cent was previously available.

Model 292X covers all AM, FM, TV, and mobile frequencies; measures both input and output of units under test; has modulated and unmodulated output from 1 to 100,000 microvolts; may be externally modulated from 15 to 10,000 cps; employs a decimal meter for faster servicing—over 100 inches of scale; and a self-contained crystal oscillator circuit.

Dimensions are 14 x 16 x 8 inches, weight 20 lbs.

For information write to H. D. Johnson at Hickok.

New 2-Watt Fixed Resistor


From voice through UHF COMMUNICATION CIRCUITS

Third Edition

By Lawrence A. Ware and Henry R. Reed

The third edition of Communication Circuits gives all the basic principles of communication transmission lines and their associated networks . . . all the frequencies, including the ultrahigh . . . impedance matching . . . attenuation in wave guides . . . microwave transmission by means of rectangular and cylindrical wave guides and coaxial cables . . . and much more.

Many chapters have been enlarged; all are brought up-to-date.

1949 490 pages 202 illus.
$5.00

10 DAYS' EXAMINATION

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John Wiley & Sons, Inc., 460 Fourth Avenue, New York 16, N.Y.
Please send me, on 10 days examination, a copy of WARE & REED'S COMMUNICATION CIRCUITS if I decide to keep the book. I will remit $5.00 plus postage; otherwise, I will return the book postpaid.

Name
City
State
Address
Employed by
(If other than college, indicate)

NOTE: This offer not valid outside U.S.A.

1949 IRE-9-49
News—New Products

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

(Continued from page 68A)

These resistors are available in a complete range from 10 to 100,000 ohms and in standard tolerances of ±5, 10 or 20 per cent as required. The new resistors have been designed to meet JAN specifications. They are fully insulated and moisture resistant. A new anchoring method assures that lead strength will exceed the standard 10-lb. pull test. Standard RMA color coding is employed. The new resistors are 11/16 inches long by 0.312 inch in diameter.

TV Antenna With Four Driven Elements

A new antenna designed to decrease co-channel interference, is announced by Technical Appliance Corp., Sherburne, N. Y.

The Type 900 TV antenna has four driven elements, two in the vertical and two in the horizontal plane.

The manufacturer claims that with this new antenna it is now possible by means of a diplexer network to eliminate the co-channel interference present in many locations where two stations are on the same channel or adjacent channels and are located about 180° apart.

Type 900 is supplied with diplexer, which is mounted at the receiver. The diplexer serves as a matching transformer between the line and the receiver, eliminating any standing waves due to mismatch. It also serves as a reversing switch forwitching directivity lobes.

Precision Micrometer Head

The announcement of production of a new type precision micrometer head, designed for the electronics industry is made by the manufacturer, Frequency Standards Corp., P. O. Box 68, Easton, N. J.

(Continued on page 70A)

Federal Telecommunication Laboratories, Inc.

500 Washington Ave. NUTLEY 10, N. J.

Design for making impedance, standing wave ratio, and wavelength measurements in the range of 60 to 1000 megacycles per second. Careful design and precision manufacture enable highly accurate measurements to be made with the line.

High sensitivity and selectivity due to efficient probe tuning.

End connectors adapted to use of Type N or similar fittings for solid dielectric cables as well as for 5/8, 1/8 and 3/8 inch air lines.

Write for complete FTL-30A Brochure
News—New Products

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(Continued from page 69A)

The size of the thinfilm in all models is 2½ inches in diameter. Readability is assisted by contrasting colors on the scale. The antireflection screw thread, which compensates for thread wear, provides a positive spindle reading in either direction or rotation. The head is available in either English or Metric scales with either 1 inch or ½ inch thread offered in either scale.

New Four-Beam Oscillograph

A new four-beam cathode-ray indicator, capable of simultaneously displaying four related or unrelated independent phenomena on a single screen, has been made commercially available by the designer, Special Products Section, Allen B. DuMont Labs., Inc. 1000 Main Ave., Clifton, N. J.

This oscillograph is equipped with Type K1027*11 C-R tube, which contains 4 independent electron guns. It also contains its own power supply and horizontal amplifier for each channel. Sweep circuits and vertical amplifiers are not provided because the instrument was not intended for use with a record camera of the moving-film type.

The horizontal amplifiers are directly coupled, with conductive or capacitive inputs to the four channels. Each channel may be individually calibrated by means of an internal voltage calibrator. Selector switches permit either conductive or capacitive input coupling.

Spectrum Analysis

from AF to UHF

Faster and Simpler with these

Panoramic Instruments

Whether your problem is investigation of noise, vibrations, harmonics, characteristics of A.M. P.M., or pulsed signals, or transmission characteristics of lines and filters, a new group of instruments are now available which will help you in your investigation. These instruments will help you in your investigation in a way which is at least easier and more economically by automatically recognizing frequencies in a broad spectrum.

Ponoramic Sonic Analyzer AP-1
Complete Audio Waveform Analysis In
One Second

Recognized as the practical answer for analyzing waves of random or static character, the AP-1 automatically separates and measures complex wave components in only one second. It reads frequency, carrier, modulation, or noise levels.

Frequency Range: 40-20,000 c.p.s., log scale.
Input Voltage Range: 500 µV-500V.
Rating: 50 volts. Overall range.
Input Voltage Range: 500 µV-500V.
Rating: 50 volts. Overall range.

Ponarmonic Ultrasonic Analyser
Entirely New for Ultrasonic Studies

An invaluable new direct reading instrument, the SB-7 enables overall obtainment of the ultrasonic spectrum or very highly detailed examination of any selected narrow spectrum segment.

Frequency Range: 2KC-300 KC linear scale.
Input Voltage Range: 0-90 volts, 0-300 volts.
Dimensions: 10 x 10 x 15 inches.
Weight: 30 lbs.

Ponaralyzer Panadaptor for
RF Spectrum Analysis

Long accepted as the simplest and fastest means of observing segments of the RF spectrum. Panadaptors are used to adapt to any type of receiver, voltage controlled oscillator, or any other type of signal generator for this purpose and have a flat amplitude response for determining relative levels of signals. Both types are available in any color desired.

Write Now for Complete Technical Data

See these instruments demonstrated at the National Electronics Convention, Booth No. 5

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For Every Application...

Outstanding in every respect, JOHNSON Variable and Fixed Inductors are available in a wide variety of types to meet every electronic application. JOHNSON has available a wide range of standard models — or can build special types in production quantities, on short notice.

Among the different types are:

222 SERIES
(Illustrated above)
For low power electronic heating and medium power transmission, internal sliding contact type. Mica-resistance, conductor 1 ½铜 strip nickel plated. Inductors of this standard type are wound to specific requirements.

224 SERIES
For high power application, Roller contact type. Approximate inductance 75 to 150 µh with 3/8" tubing, 50 µh with 1 ½" tubing. Cast aluminum end frames.

226 SERIES
For high frequencies, Rotating coil type. Optional variable pitch windings for wide frequency band coverage. Edge-wise copper strip, silver plated. Wound to customer specifications.

227 SERIES
A high current inductor especially adapted to Electronic Heating Equipment. Rotating coil type. Available in single or dual models, with or without coupling links. End frames and support bars, Mica-resistance, Conductor 3/4" flat wound silver plated copper.

229 SERIES

TYPE M
Inductance: built to any specified inductance from 10 µh up. Basic M design permits any length and diameter.

TYPE VM
Same as M except supplied with variable coupling rotors, flippers, or as varimeters. Fixed coils may be incorporated to reduce eddycurrent losses.

TYPE N
Fixed Inductors wound with either copper strip, ribbon, tubing or wire. Inductance: built to any specified inductance from 10 µh up. May be supplied with either internal or external coupling winding.

TYPE VN
Same as N except variable. Main winding stationary with rotating winding connected as varimeter or coupling inductor.

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PROCEEDINGS OF THE I.R.E., September, 1949
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Ruggedly designed, with unique mechanical construction and simplified circuits, this generator has exceptionally low leakage. Through the elimination of a number of circuit frills it has been possible to keep its cost to a moderate figure.

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**CARRIER-FREQUENCY RANGE:** 5 kc to 50 Mc in eight logarithmic, direct-reading ranges.

**FREQUENCY CALIBRATION:** accurate to ±1%.

**INCREMENTAL-FREQUENCY DIAL:** indicates increments of 0.1 per cent of frequency per division up to 15 Mc.

**OUTPUT VOLTAGE:** open-circuit voltage adjustable from less than 0.1 microvolt to 200 millivolts. Two-volts output available from a second jack.

**AMPLITUDE MODULATION:** adjustable from 0 to 80% either with 400-cycle built-in source, or over 20 to 15,000 cycles from an external source.

**LEAKAGE:** stray fields are substantially less than 1 microvolt per meter two feet from the generator.

**INCIDENTAL FREQUENCY MODULATION:** varies from 10 to 100 parts per million, at 80% a.m., over each range except 15-50 Mc where it may be three times this amount.

**ENVELOPE DISTORTION:** about 6% at 80% modulation.

**NOISE LEVEL:** carrier noise level corresponds to about 0.1% modulation.

Fully 90 per cent of the needs for a standard-signal generator for general laboratory use are adequately met in this carefully designed G-R instrument, where the ultimate in accuracy and stability is not required.

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