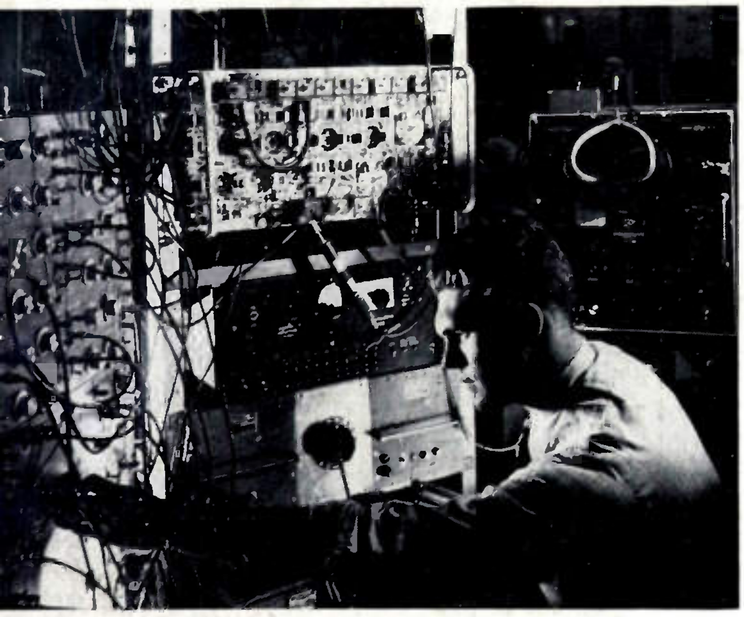
Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering
(Including the WAVES AND ELECTRONS Section)

December, 1949
Volume 37 Number 12



Sylvania Electric Products Inc.

ELECTRONIC-COMPUTER DEVELOPMENT

One section of a computer (in the upper middle of the picture) undergoes systematic tests for necessary precision of its component elements.

The following IRE Standards appear in this issue: Radio Aids to Navigation, Definition of Terms; Railroad and Vehicular Communications, Methods of Testing; Tests for Effects of Mistuning and for Downward Modulation; Piezoelectric Crystals.

PROCEEDINGS OF THE I.R.E.

Waves and Electrons Section

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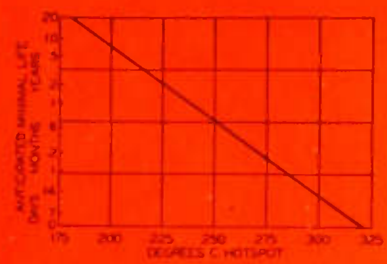
The Institute of Radio Engineers

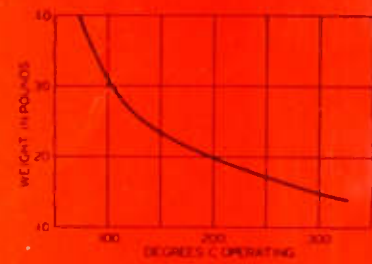
FOR SPECIALTIES

The fields of frequency control, Servomechanisms, etc., are developing rapidly with increasing complexity. UTC is playing a principal role in the development of special components for these and allied fields. A few typical special products are illustrated below:

SMALLER POWER COMPONENTS

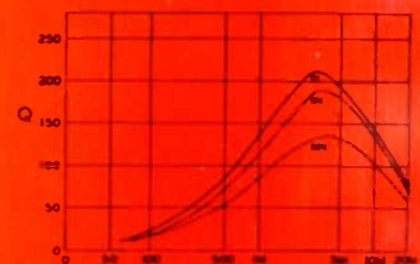
High temperature (class H) insulation, and, in many instances, short life requirements, can effect considerable weight and size reduction where these are important. The curve at the left indicates anticipated life versus temperature rise, using Class "H" insulating materials. The curve at the right illustrates on one typical type the variation in weight with permissible continuous operating temperature.





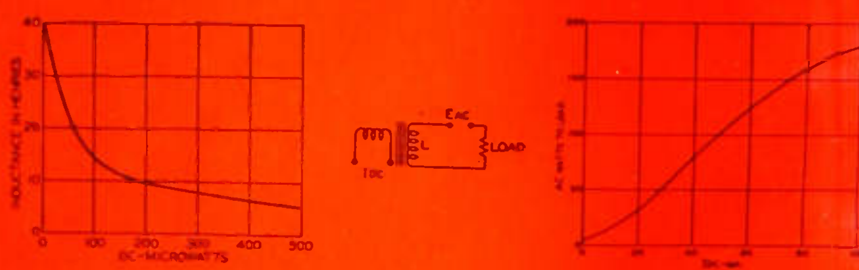
TOROID DUST HIGH Q COILS

UTC type HQ (permalloy dust) coils have found wide application because of their high Q, stable inductance, and dependability. Four standardized groups of stock coils cover virtually any high Q coil application from 300 cycles to 300 Kc.



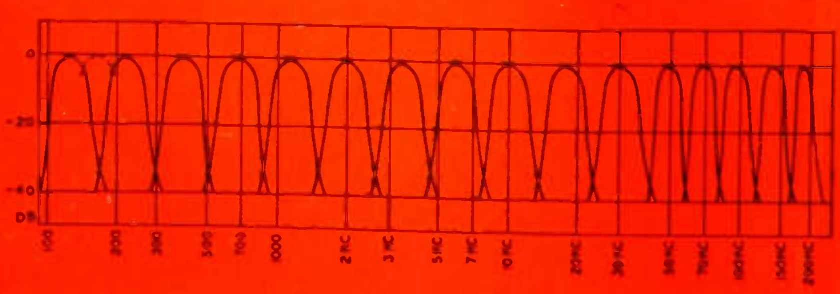
MAGNETIC AMPLIFIERS

Magnetic amplifiers are used extensively for both power control and phase control. The left curve shown is that of a sensitive saturable reactor structure controllable with powers below .5 milliwatt. The right curve is that of a moderate size power control reactor indicating power to the load with saturating DC.



AUDIO FILTERS

The curve illustrated shows a group of filters affording sixteen separate bands in the audio and supersonic region with 35 DB attenuation at the cross-over points. These have also been supplied spaced further apart (40 DB cross-over), with intermediate bands, permitting flat top band pass action for any selected range from 100 cycles to 200 KC.



May we design a unit for your application problem.

United Transformer Co.

150 VARICK STREET • NEW YORK 13, N. Y.

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

COME  AGAIN

to the 1950 IRE National Convention and Radio Engineering Show

March 6-9, 1950

at the Hotel Commodore and Grand Central Palace, New York City

• Technical Sessions

Every field of radio-and-electronics will be explored and progress reported in some 36 three hour sessions and symposiums. More than 150 engineering papers will be presented, skillfully, organized by subjects ranging from Broadcasting to Nucleonics; Circuits to Electronics in Industry.

• Social Events

IRE members renew friendships and make new ones at the Annual Meeting, the Get-Together Cocktail Party, Monday; The Presidents Luncheon, Tuesday; and at the Annual Banquet, Wednesday. It is good to know your fellow members—worth the trip!

• Exhibits

A bigger and better than ever Radio Engineering Show will "Spotlight the New" in 311 exhibits and 12 Theatres featuring the products of 230 manufacturers and U. S. government services. Every exhibit is educational and worth seeing.

• Attendance

Nearly one-third of the total IRE Membership attended the 1949 Convention and Show. Other thousands paid the $3.00 non-member registration to attend. This important national event scored a record 15,710 attendance this year. (Complete analysis on request.)

Calendar of
COMING EVENTS

1949 Annual Meeting, National Society of Professional Engineers, Houston, Texas, December 8-10

Southwestern IRE Conference, Baker Hotel, Dallas, Texas, December 9-10

1950 IRE National Convention, New York, N. Y., March 6-9

Fourth Annual Spring Television Conference of the Cincinnati Section of I.R.E., April 29, 1950, Engineering Societies Bldg., Cincinnati, Ohio

1950 IRE Technical Conference, Dayton, Ohio, May 3-5

IRE Regional Meetings Accelerate Electronic Progress!

• Visitors tell us!

A buying survey of 2373 engineers attending the Radio Engineering Show proved that 20.3% buy equipment, 47.4% specify for purchases, and 12.4% have other authority or influence—80.1% of these engineers in all have buying authority and interest. (We will supply full text of this study upon request.)

Here are 36 of their specific buying interests, with percentage seeking information on these products:

17.8% Aviation Radio, 46.7% Amplifiers
36.5% Antennas, 13.8% Batteries
23.6% Cables, wire, 37.8% Capacitors
13.9% Cabinets, Chassis, 24.7% Coils
9.8% Consoles, 22.8% Crystals
38.8% Electronic Controls, 12.8% facsimile equipment, 19.1% Hardware
20.7% Ceramics, 10.9% Induction equipment
33.8% Loudspeakers, 39.7% Meters
13.7% Motor generators, 14.8% Plastics
31.2% Oscillators, 45.5% Oscillographs
25.2% Power supplies, 25.4% Radar
43.4% Receivers (all), 37.1% Recorders
32.7% Rectifiers, 27.0% Relays
32.4% Resistors, 68.0% Test Equipment
30.0% Transformers, 26.8% Transmitters
26.2% Turntables, pickups, 36.0% U.H.F. Equipment
54.0% Vacuum Tubes, 23.1% Voltage reg.
22.8% Tools for radio manufacturing.

We invite you to come again,
March 6-9, 1950.

TO SELL THE RADIO INDUSTRY

THE INSTITUTE OF RADIO ENGINEERS

Established 1913

A Balanced Promotion Package

"Proceedings of the I.R.E." The IRE Yearbook
The Radio Engineering Show

303 WEST 42nd STREET, NEW YORK 18, N. Y.
Circle 6-6357

TELL THE RADIO ENGINEERS

PROCEEDINGS OF THE I.R.E. December, 1949, Vol. 37, No. 12, Published monthly in two sections by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price $2.25 per copy. Subscriptions: United States and Canada, $18.00 a year; foreign countries $19.00 a year. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P.L. and R., authorized October 26, 1927.

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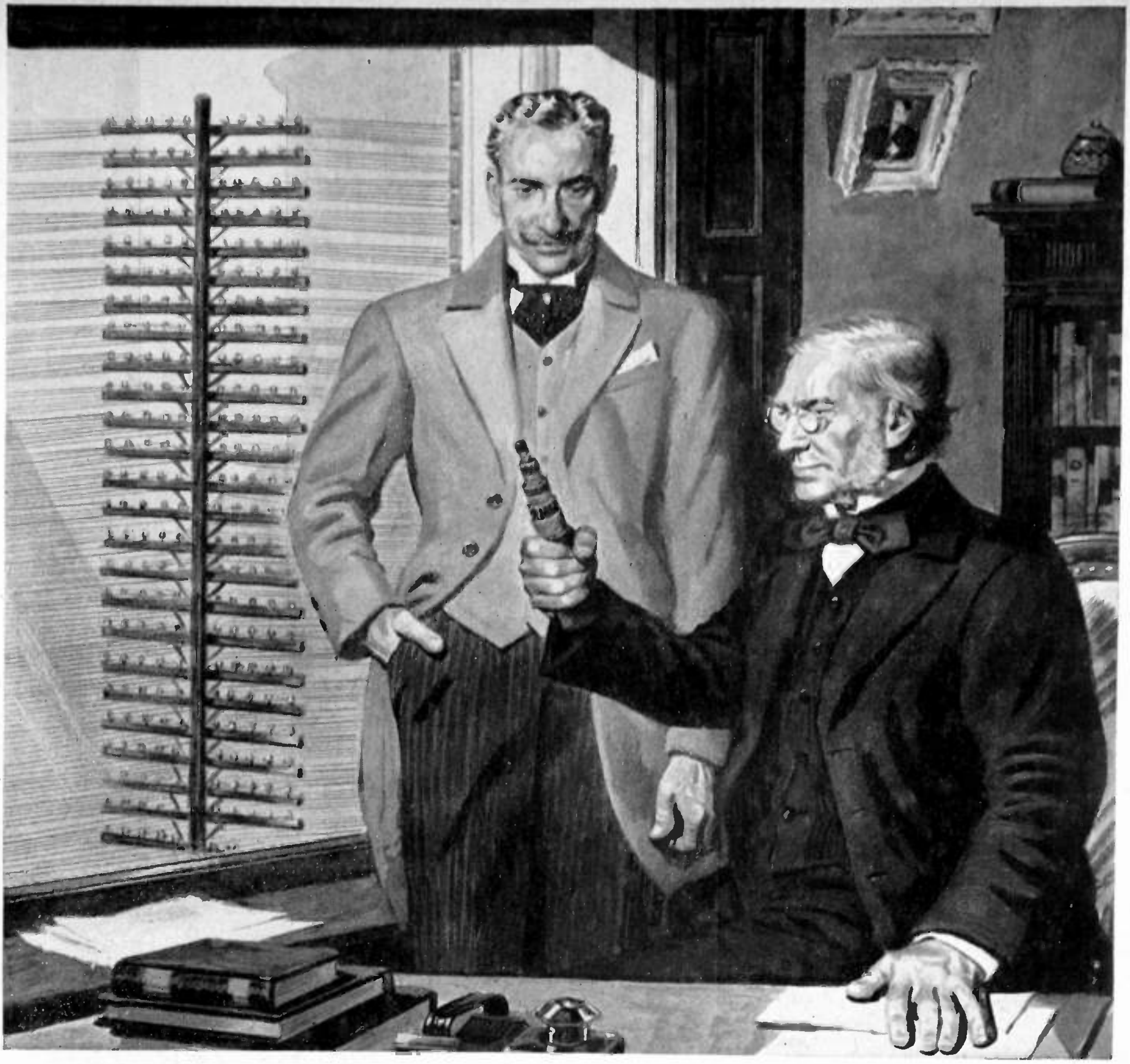


They Packed a Pole Line Into a Pipe

Back in the eighties, telephone executives faced a dilemma. The public demanded more telephone service. But too often, overloaded telephone poles just couldn't carry the extra wires needed, and in cities there was no room for extra poles. Could wires be packed away in cables underground?

Yes, but in those days wires in cables were only fair conductors of voice vibrations, good only for very short distances. Gradually cables were improved; soon every city call could travel underground; by the early 1900's even cities far apart could be linked by cable.

Then Bell scientists went on to devise ways to get more service out of the wires. They evolved carrier systems which transmit 3, 12, or even 15 voices over a pair of long distance wires. A co-axial cable can carry 1800 conversations or six television pictures. This is another product of the centralized research that means still better service for you in the future.

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

SWIFT, SURE FREQUENCY COMPARISON

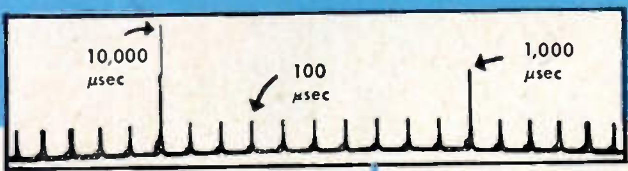


FIG. 1. Timing Comb, -hp- Model 100D

NEW -hp- SECONDARY FREQUENCY STANDARDS

MODELS 100C AND 100D

- Sine or rectangular waves
- 100 μsec time markers
- Built-in oscilloscope
- Stability 1/1,000,000
- Low output impedance
- New, improved circuits
- Audio, supersonic, rf measurements

SPECIFICATIONS

-hp- 100D Secondary Frequency Standard

Accuracy:
: About 2 parts per million per week, normal room temperature.

Stability:
: About 1 part per million over short intervals.

Output:
: Controlled frequencies: 100 kc, 10 kc, 1 kc, 100 cps, 10 cps. Sine or rectangular waves; marker pips. Internal impedance approx. 200 ohms.

Wave Shape:
: Sine wave: less than 4% distortion into 5,000 ohms or higher load.

Marker Pips:
: 10,000, 1,000 and 100 μsec intervals.

Oscilloscope:
: Integral with circuit. Establishes 10:1 Lissajous figures to show division ratio. May be used independently of standard.

-hp- 100C Secondary Frequency Standard

Accuracy:
: Within ± .001% normal room temperature.

Output:
: Controlled frequencies of 100 kc, 10 kc, 1 kc, and 100 cps. Internal impedance approx. 200 ohms.

Wave Shape:
: Sinusoidal only. 4% distortion into 5,000 ohm load.

Power Supply:
: (100C and 100D) 115 v, 50/60 cps, regulated to minimize line voltage fluctuations. Power drawn approx. 150 watts.

Mounting:
: (100C and 100D) Cabinet or relay rack. Panel 19" x 10½". 12" deep.

Data Subject to Change Without Notice

The new -*hp*- 100C and 100D Secondary Frequency Standards incorporate all the features of the time-tested -*hp*- models 100A and 100B, plus important new advantages including rectangular wave output, timing pips, and an internal oscilloscope for convenient frequency comparison. The -*hp*- 100D may be conveniently standardized against station WWV with a minimum of external equipment, and thus provide most of the advantages of an expensive primary standard.

Crystal Controlled Frequencies

The new -*hp*- Models 100D and 100C employ a crystal-controlled oscillator and divider circuits offering a new high in stability and simplicity of operation. Standard frequencies are available through a panel selector switch, and may be employed simultaneously. Internal impedance is low (about 200 ohms), so that standard frequencies can be delivered at some distance from the instrument.

The -*hp*- 100D Secondary Frequency Standard offers sine waves at 5 frequencies and rectangular waves at 4 frequencies, plus a built-in oscilloscope. The instrument also provides a timing comb with markers 100, 1,000 and 10,000 microsecond intervals. Rectangular wave output has a rise time of approximately 5 microseconds. Accuracy is 2 parts per million.

5 v. at all Frequencies

The more moderately priced -*hp*- 100C Standard offers sinusoidal frequencies at 4 crystal-controlled frequencies and, like the -*hp*- 100D, provides 5 volts of output at all frequencies. Accuracy .001%.

Both models operate from a 115 v. ac power supply, and power is regulated to minimize power line voltage fluctuations.

Get full details . . . see your -hp- representative or write direct . . . today!

HEWLETT-PACKARD CO.
1977-D Page Mill Road • Palo Alto, Calif.
Export: FRAZAR & HANSEN, LTD.
301 Clay Street, San Francisco, Calif., U. S. A.
Offices: New York, N. Y.; Los Angeles, Calif.

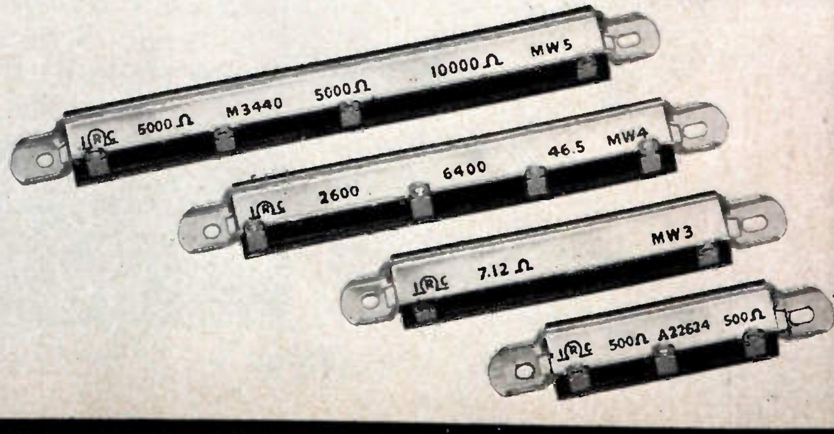
-hp- laboratory instruments
FOR SPEED AND ACCURACY

Power problems can

for resistors too!

... But the most stubborn problems of power and resistance find an answer in IRC resistors. Where crowded chassis demand miniaturization—where portable units call for maximum lightness—where critical applications require complete insulation, unusual heat dissipation, or ability to handle exceptionally high voltages—there's an IRC power resistor for the specific job. Manufacturing the widest line of resistance products in the industry, IRC can specify *without bias!* This is your warranty of efficient, economical resistor performance in virtually any application.

UNSURPASSED FOR ADAPTABILITY to an extremely wide variety of design requirements, IRC Type MW Wire Wound Resistors also give more watts per dollar than conventional types. Initial cost and mounting cost are low. Flexibility in providing taps and savings in space suit flat Type MW's to a host of difficult applications. Mounting bracket of unique design actually transfers heat from inside resistor to mounting surface. MW's may be operated at full "on plate" rating, whether enclosed or not, without exceeding their rated temperature of 100° C. (Thus, an MW 5 can be used at 20 watts where a 60 watt tubular resistor would normally be required.) Use the convenient coupon to obtain full characteristic and specification data.

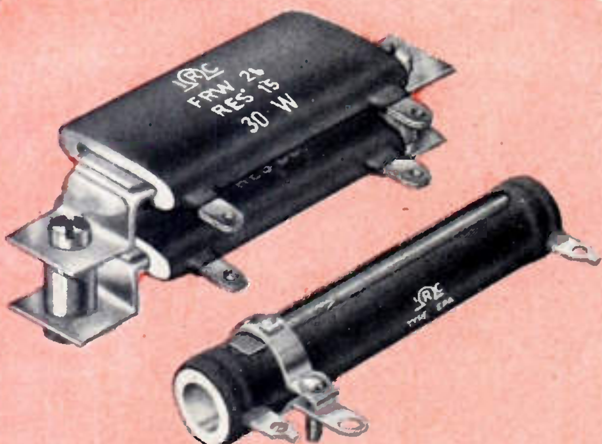


be tough

EXCEPTIONAL STABILITY

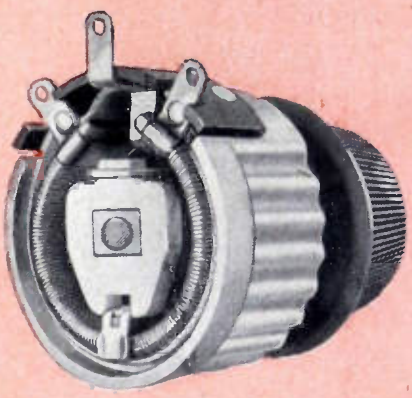
even in very high resistance values assures the dependability of IRC Type MV Resistors in high voltage applications. Unique application of IRC's famous filament coating in helical turns on a ceramic tube provides a conducting path of long effective length, and permits the use of high voltage on the resistor while keeping the voltage per unit length of path comparatively low. Bulletin G-1 gives complete characteristics; use handy coupon.

LENGTH OF RESISTANCE PATH IN INCHES—FOR MV TYPES			
MVF	10	MVS	20
MVG	25	MVT	30
MVJ	50	MVB	70
MVP	80	MVD	95
MVA	190	MVZ	185
MVO	330	MVE	265
MVR	1,100		



HIGHER SPACE-POWER RATIO than tubular power resistors makes IRC Type FRW Flat Wire Wounds ideal for voltage dropping applications in limited space. FRW's can be mounted vertically or horizontally, singly or in stacks—and are available in fixed or adjustable types. Bulletin C-1 gives all the performance facts.

UNEXCELLED IN EXACTING, HEAVY-DUTY REQUIREMENTS, IRC Power Wire Wounds are designed to give balanced performance in every characteristic. Special dark, rough coating assures ability to combat humidity and moisture corrosion, dissipate high heat rapidly, withstand reasonable overloads without opens or breakdowns. Available in a broad range of ratings, sizes and terminal styles in fixed and adjustable types. Send for Bulletin C-2.



VARIABLE POWER requirements within 25 and 50 watt ratings are well handled by IRC Type PR Power Rheostats. All-metal corrugated construction gives maximum heat dissipation. PR Rheostats can be operated at full power in as little as 25% of rotation without appreciable temperature rise. Direct contact between housing and mounting panel increases ability to disperse heat. Bulletin E-2 details specifications; send for your copy.

When you're being "powered" for fast service on small order resistor requirements for experimental work, pilot runs, or maintenance, call your nearest IRC Distributor. IRC's Industrial Service Plan enables him to save you time and worry by giving you 'round-the-corner service on standard types right from his local stocks. He's a handy man to know. May we send you his name and address?

IRC

Wherever the Circuit Says

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

INTERNATIONAL RESISTANCE COMPANY
401 N. Broad Street, Philadelphia 8, Pa.
In Canada: International Resistance Co., Ltd., Toronto, Licensee

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

Send me additional data on the items checked below:

- ☐ MW Insulated Wire Wounds
- ☐ MV High Voltage Resistors
- ☐ PR Power Rheostats
- ☐ FRW Flat Wire Wounds
- ☐ Power Wire Wounds
- ☐ Name and address of local IRC Distributor

NAME ....

TITLE ....

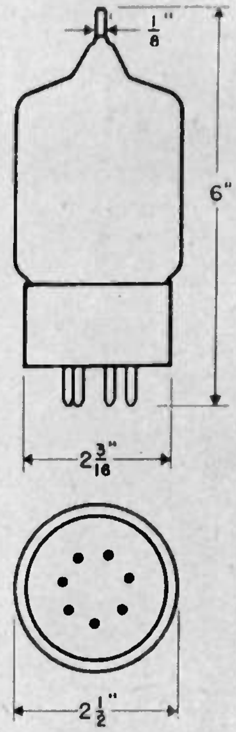
COMPANY ....

ADDRESS ....

....

A NEW, VASTLY IMPROVED 4E27

- MORE RUGGED PLATE-LEAD
- PYROVAC PLATE
- OVERSIZE PLATE
- NON-EMITTING GRIDS
- MECHANICALLY RUGGED
- MOULDED-GLASS HEADER
- LOW-LOSS LEADS
- EASILY COOLED STEM



EIMAC PENTODE TYPE 4E27A/5-75A

Encompassed in the structure of this new version of the 4E27 are many outstanding improvements that now will guarantee performance-dependability to users of this tube type.

The plate-lead of this new Eimac 4E27A/5-75A pentode is of larger diameter than the protype* providing a low-loss, low inductance, more rugged lead. The plate itself is larger assuring a good reserve dissipation capacity above its 75 watt rating. It is made of Eimac Pyrovac plate material, which lengthens the life of the tube and enables it to withstand high momentary overloads.

Primary grid emission has been eliminated and secondary characteristics stabilized through the use of Eimac processed grids. Perfected beam-action and permanent alignment are assured through well engineered internal-element mounts.

The unique moulded-glass header eliminates a base on the 4E27A/5-75A. This simplifies lead cooling, minimizes lead losses, and provides precision alignment of base-pins.

The stability and high power-gain characteristics of this new Eimac pentode make it an excellent VHF or video power amplifier. It is equally well suited for conventional power amplifier service.

Further information and detailed characteristics concerning this latest product of Eimac engineering research may be had by writing the Application Engineering Department of Eitel-McCullough, Inc.

* Lead connector is supplied to make this new tube directly interchangeable with 4E27.

236

EITEL-McCULLOUGH, INC.
San Bruno, California

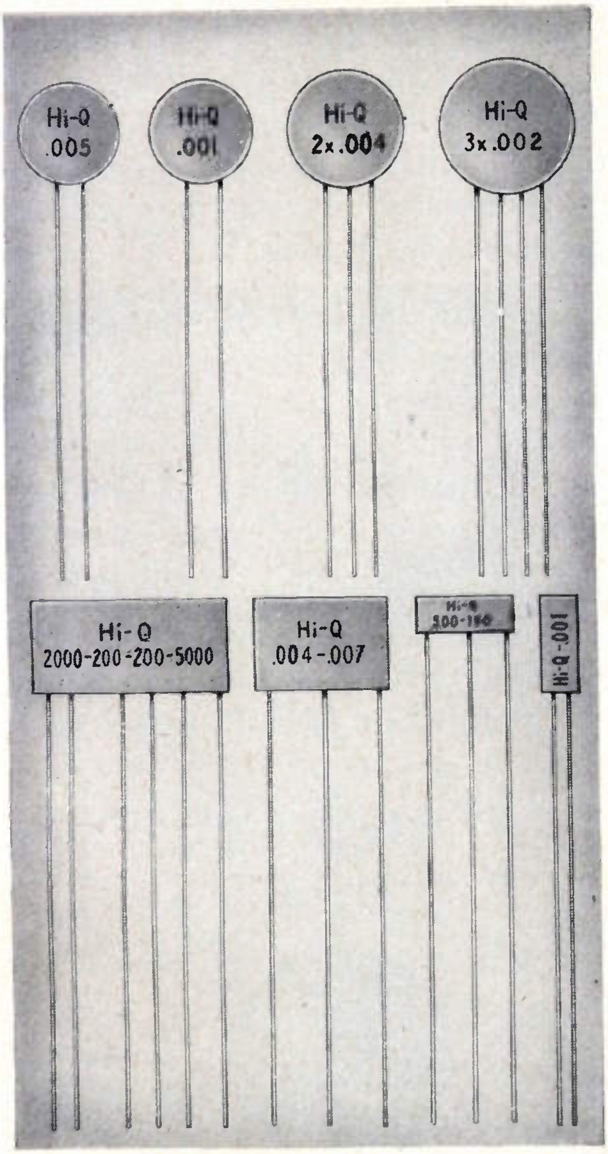
EXPORT AGENTS: FRAZAR & HANSEN, 301 CLAY ST., SAN FRANCISCO, CALIFORNIA

Follow the Leaders to Eimac REG. U.S. PAT. OFF. TUBES

Specify HI-Q COMPONENTS-

BPD's (Disks) and BPF's (Flats)

for SPACE SAVING and ECONOMY



ILLUSTRATIONS APPROXIMATELY ACTUAL SIZE

HI-Q Disk and **HI-Q** Flat Ceramic Capacitors frequently save space simply because their physical shape is more adaptable than tubular units . . . *and even more frequently* because one of them serves in place of two, three or more individual capacitors. The multiple units also simplify soldering and wiring operations and thus effect substantial production economies.

These are just a few of the many types of **HI-Q** Components which are setting the highest possible standards for *Precision, Quality, Uniformity* and *Miniaturization.* Our engineers are always available to work with you in developing capacitors or combinations of capacitors to best meet your specific needs. Please feel free to call on us at any time.

- **HI-Q** BPD's (Disks) are available in capacities of from .001 mf. to .01 mf. Dual units range from 2x.001 mf. to 2x.005 mf. Triple units are supplied in standard rating of 3x.0015 mf. and 3x.002 mf. All are guaranteed minimum values.
- **HI-Q** BPF's (Flats) can be produced in an unlimited range of capacities. The number of capacities on a plate is limited only by the "K" of the material and the physical size of the unit. They do *not* necessarily have to have a common ground as is the case with the disk type.

HI-Q COMPONENTS

BETTER 4 WAYS

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

UNIFORMITY Constancy of quality is maintained over entire production through continuous manufacturing controls.

DEPENDABILITY Interpret this factor in terms of your customers' satisfaction . . . Year after year of trouble free performance. Our Hi-Q makes your product better.

MINIATURIZATION The smallest BIG VALUE components in the business make possible space saving factors which reduce your production costs . . . increase your profits.

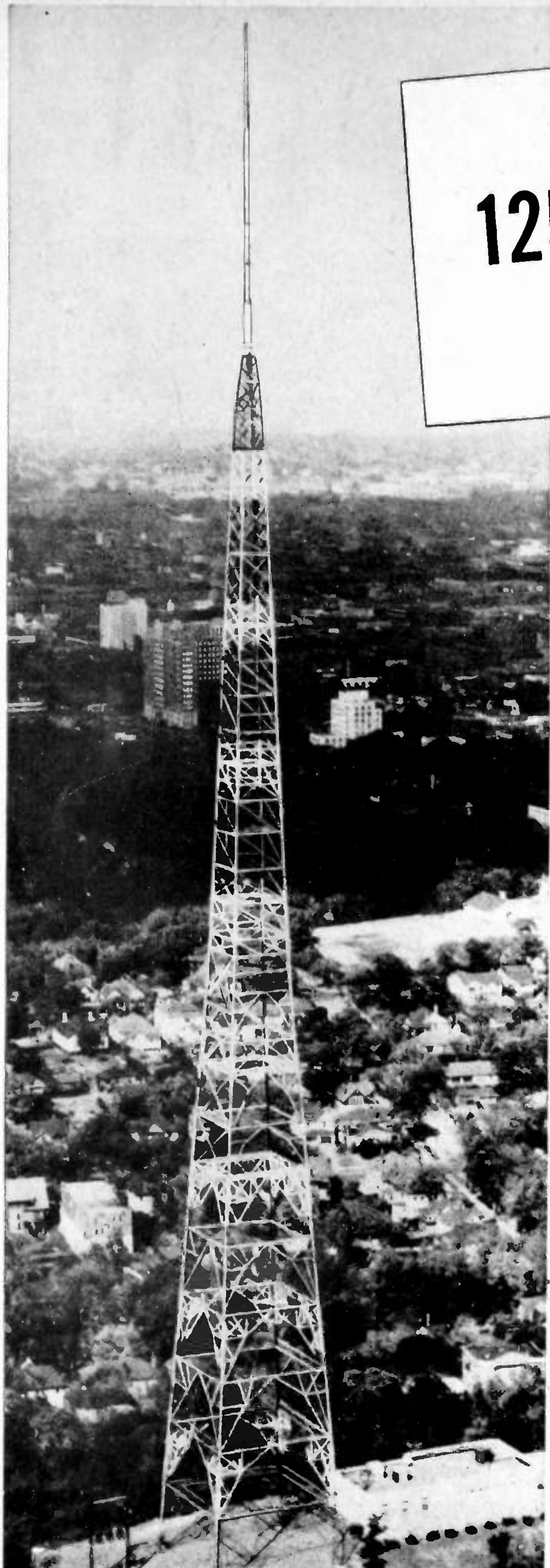
Jobbers—Address Room 1332
101 Park Ave. New York, N.Y.

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



A 125,000 SQ. MILE BLANKET!

The most powerful FM installation in the world recently completed on Red Mountain near Birmingham, Alabama for Station WBRC-FM brings static-free entertainment to residents in a transmission radius of 200 miles.

Important to this installation is the 450 ft. Blaw-Knox type N-28 heavy-duty tower supporting the 8-section Pylon FM antenna. Sturdy, safe and backed by the many years of Blaw-Knox design and engineering in the radio field, it will enable this great new FM Voice of the South to utilize the full capacity of its modern facilities.

BLAW-KNOX DIVISION of Blaw-Knox Company
2037 Farmers Bank Building, Pittsburgh 22, Pa.

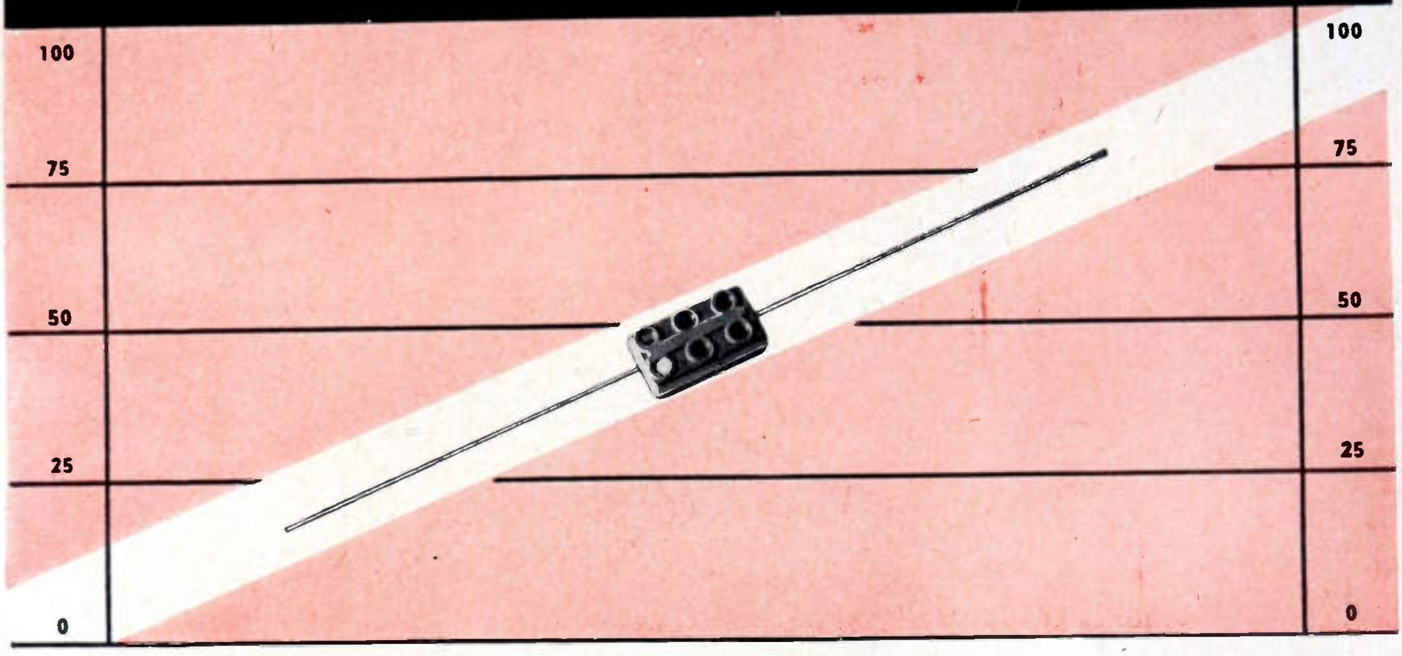
BLAW KNOX

ANTENNA TOWERS

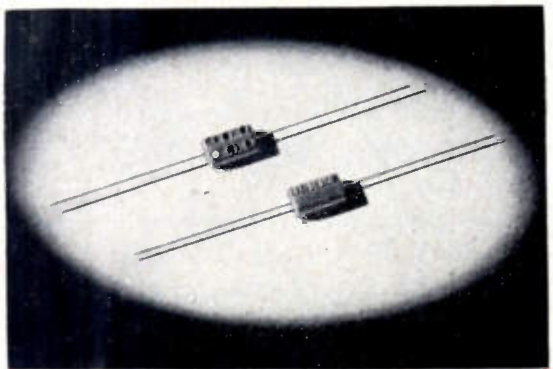
Distributed by Graybar

BLAW-KNOX
ANTENNA TOWERS

For Peak Performance...



EL-MENCO CAPACITORS



CM 15
Actual Size 9/32" x 1/2" x 3/16".
For Television, Radio and other Electronic Applications.
2 — 420 mmf. cap. at 500v DCw.
2 — 525 mmf. cap. at 300v DCw.
Temp. Co-efficient ±50 parts per million per degree C for most capacity values.
6-dot color coded.

You can always depend on these tiny but tried and trusted El-Menco capacitors to give peak performance for long periods of time under the most exacting conditions. Rigid test during and after manufacture insures uniformity and assures quality.

Performance proved, these fixed mica dielectric capacitors are specified by nationally-known manufacturers.

When you need peak performance in capacitors, get the best — get **El-Menco.**

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

Write on your firm letterhead for Catalog and Samples

MOLDED MICA El-Menco MICA TRIMMER CAPACITORS

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

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

ELECTRONICS

Designers

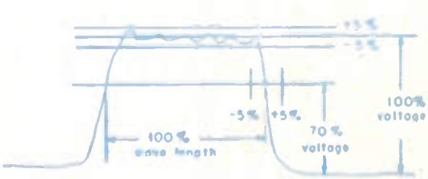
A LINE-VOLTAGE STABILIZER

SO SMALL . . .
. . . it mounts on a radio chassis

These 15-, 25-, and 50-va G-E voltage-stabilizer units are only a little over 2 inches high and about 9 inches long. They'll mount easily on a medium-sized radio or electronic instrument chassis and will give you an even, non-fluctuating 115 volts for your equipment whether your line voltage is 95 or 130. A special transformer circuit provides a stabilized output voltage within 1% of 115 volts for fixed, unity-power-factor loads.

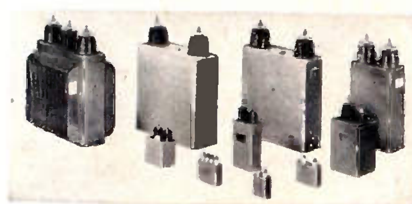
Continuous operation under conditions of short or open circuits will not damage the stabilizer in any way. Since there are no moving parts, there is little maintenance to worry about. For complete information on voltage-stabilizer units of all sizes from 15-va to 5000-va, write for Bulletin GEA-3634.

AN EASY WAY TO PRODUCE SQUARE WAVES



Specially designed G-E Type-E networks will produce impulses which have definite, known energy contents and durations, and thus are ideal for converting a-c or d-c charging voltages into approximately rectangular square waves. These networks consist of capacitor and coil sections adjusted to close tolerances and hermetically sealed in single metal containers.

G.E. helped meet wartime radar demands with thousands of these units and now offers them for commercial use. They are available in a wide range of designs, impedances, ratings, and sizes for pulse lengths of 0.1 to 40 microseconds. See Bulletin GEA-4996.

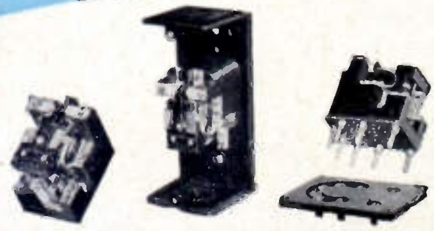


GENERAL ELECTRIC

667-3

Digest

TIMELY HIGHLIGHTS ON G-E COMPONENTS



HEAVY-DUTY RELAYS THAT MOUNT 3 WAYS

This versatile, general-purpose, heavy-duty, a-c relay unit is available in three mounting arrangements: front connected, back connected, or plug-in connected. All three mounting types are available in open or enclosed models and are furnished in spst, dpst, or dpdt circuits. Heavy, long-lasting silver contacts carry 10 amps continuous. Normally-open forms make or break 45 amps; normally-closed forms make or break 20 amps. Relay coils come in 12-, 24-, 115-, or 230-volt, 60-cycle a-c sizes. D-c units are available in similar models. For full details see GEC-257.

ACCURATE BUT RUGGED

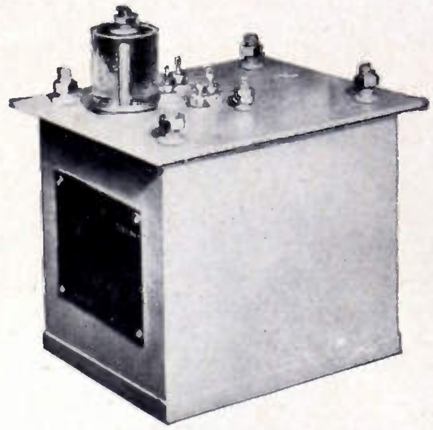
The new, modern-looking, easy-to-read 2½ inch G-E instrument line is improved inside as well as outside. A single, self-contained mechanism supported on an extremely strong Alnico magnet assures permanent alignment even under the most adverse operating conditions. This high-gauss Alnico magnet permits the use of a large air gap with a consequent smoother, non-sticking action. The greater torque-to-weight ratio means better damping and allows the use of heavier vibration-resisting pivots. Accuracy is 5% of full scale on rectifier types, 2% on all others. For complete details, send for Bulletin GEC-368.

SNAP-SWITCH INSTALLATION TIME CUT TO SECONDS

You'll have a firm electrical connection without the use of solder a few seconds after you begin to install this small but rugged Switchette. Only 1½ inches long and weighing only 9 grams, this 230-vac, 10-amp unit has solderless knife-contact terminals made of pure, tinned copper.

G-E Switchettes are available in a variety of forms and circuits, all of which have double-break contact structures. They're particularly well suited for electronic applications because of their low RF noise output (short contact-bounce).

For your convenience there are screw-terminal and soldering-lug types as well as this special quick-connect unit. Send for Bulletin GEA-4888.



A SMALL PACKAGE OF WELL-REGULATED HIGH VOLTAGE

You get both high voltage and good regulation with small lightweight G-E precision rectifiers. This may interest you if you need compact, well-regulated, high d-c voltage sources for cathode-ray tubes, television camera tubes, radar indicator scopes, electron microscopes, Geiger-Mueller counters, or similar jobs.

These supplies are hermetically sealed and oil-filled. Typical units have outputs of 7 kv at 0.1 ma.—have only 3.5% deviation for every 0.1 ma load and output ripple of less than 1%. Size—only 6″ x 6″ x 7″. Weight—8 lbs. For further data, write: General Electric Company, Section 667-3, Schenectady 5, N. Y., giving complete information on the proposed application with specifications required.

General Electric Company, Section H667-3
Apparatus Department, Schenectady, N. Y.

Please send me the following bulletins:

- ☐ GEA-3634 Voltage stabilizers
- ☐ GEA-4888 Switchettes
- ☐ GEA-4996 Capacitor networks
- ☐ GEC-257 Heavy-duty relays
- ☐ GEC-368 Instruments

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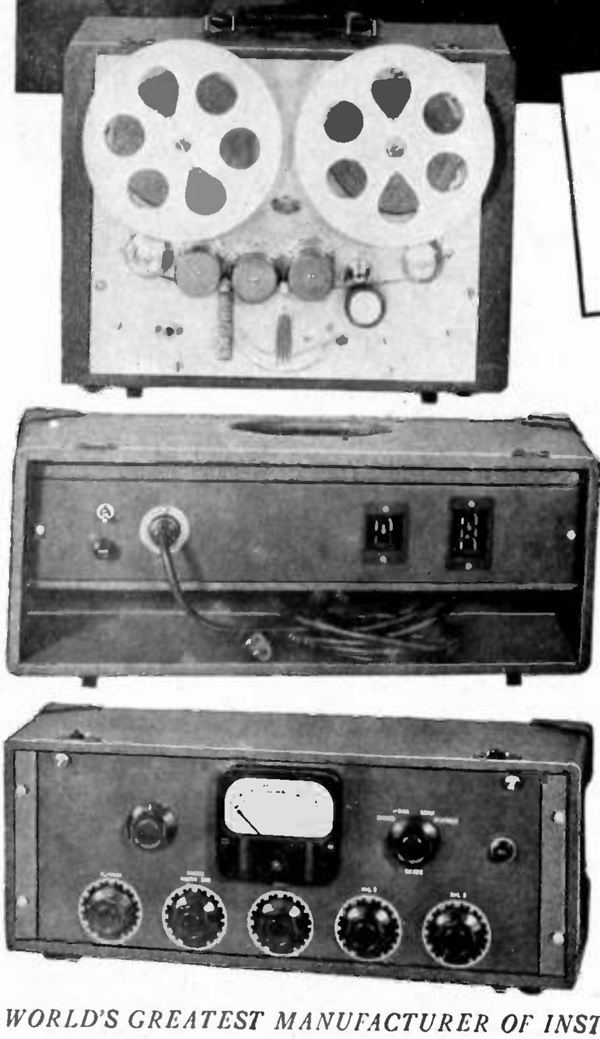
COMPANY__________

ADDRESS__________

CITY__________ STATE__________



NEW PRESTO
PORTABLE TAPE RECORDER

PRESTO'S PT-900 is the answer for delayed sports broadcasts—field recording—wherever there is a need for a portable recorder of complete broadcast quality. Look at these outstanding engineering features:

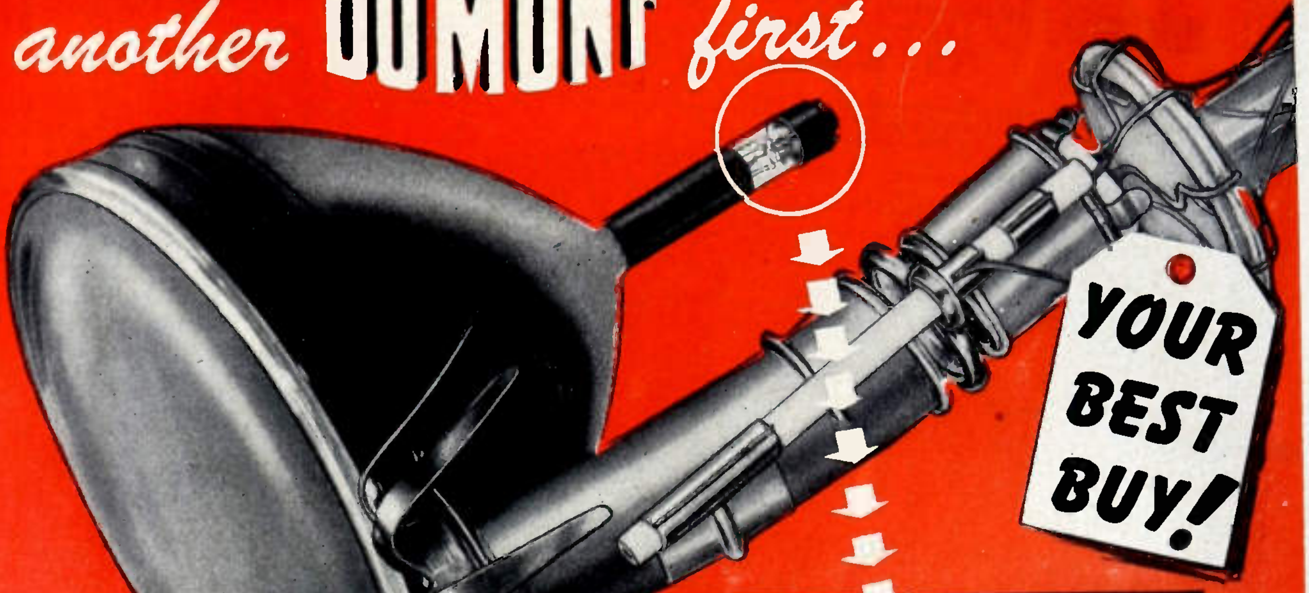
• Three separate heads for superior performance (and for monitoring direct from tape). One head each to erase, record and play back. • 3 microphone channels with master gain control in recording amplifier. • Large V.U. meter with illuminated dial to indicate recording level, playback output level, bias current and erase current, and level for telephone line. • 2-speed, single motor drive system. Toggle switch to change tape speeds from 7½" to 15" per second.

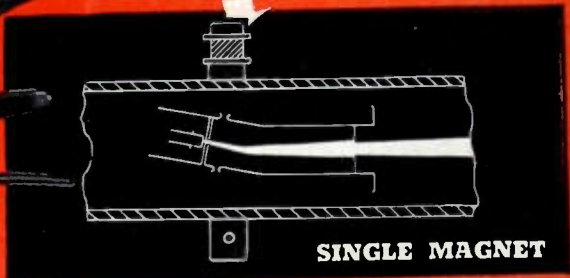
Don't choose your tape recorder until you see the *new* Presto Portable Tape Recorder. Write for complete details today.

PRESTO
RECORDING CORPORATION
Paramus, New Jersey
Mailing Address: P. O. Box 500, Hackensack, N. J.
In Canada: WALTER P. DOWNS, Ltd., Dominion Sq. Bldg., Montreal

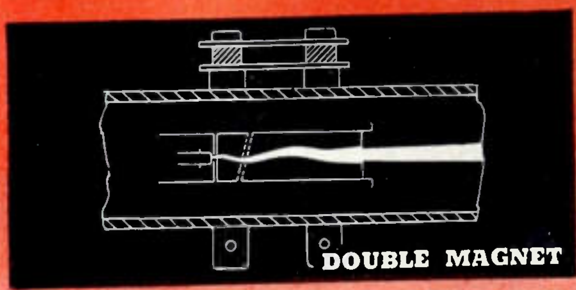
WORLD'S GREATEST MANUFACTURER OF INSTANTANEOUS SOUND RECORDING EQUIPMENT AND DISCS

another DUMONT first...





BENT-GUN *Teletrons**



The new Du Mont Types 12RP4 and 15DP4 (replacing respectively Types 12JP4 and 15AP4) feature the exclusive Du Mont bent-gun. This ion-trap design eliminates ion-spot blemishes while maintaining an undistorted spot for maximum pictorial resolution. Meanwhile, lead-free glass reduces tube weight considerably. Five-pin duodecal base permits using the new half-socket for a significant saving, although old-type full-socket also accommodates these new tubes without modification.

Definitely "Your best buy!" For initial-equipment or replacement purposes — for superlative performance and longest service — insist on Du Mont Teletrons!

Above: Du Mont bent-gun principle, utilizing single ion-trap magnet. Space saved by eliminating double beam-bending magnet results in shorter neck length. Focussed-spot distortion eliminated by use of electrode parts designed to form symmetrical electrostatic fields in G_1 space. Lower-cost magnet.

Below: Conventional straight-gun design. Ion and electron beam is twisted by slanting electrostatic field between second grid and anode, requiring TWO bending magnetic fields. More costly beam-bender. Longer neck. Focussed-spot distortion.

Write for latest literature.

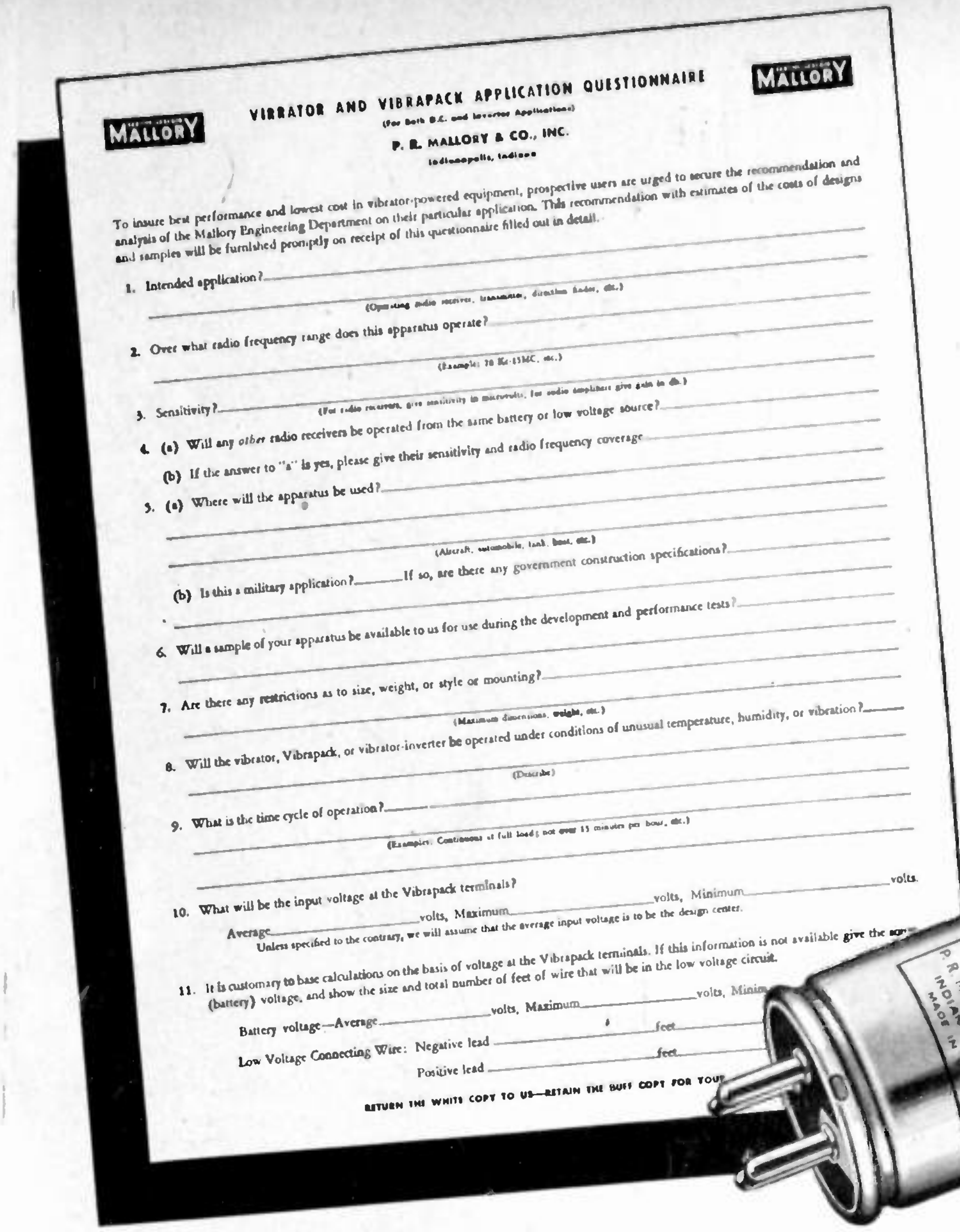
*TRADE MARK

© ALLEN B. DU MONT LABORATORIES, INC.

FIRST WITH THE FINEST IN T-V TUBES

DUMONT *Teletrons*

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



They're Right...

because you write the ticket

More Mallory Vibrators are used in original equipment than all other makes combined.

Creative research is no empty slogan at Mallory. Mallory Vibrators are the world's most popular simply because engineering skill, long experience and adherence to quality ideals have made them better.

But Mallory engineers know that the finest vibrators can fail because of a power transformer design . . . or a wrong value buffer capacitor. That's why they want to know the whole story of your problem.

That's the reason for the inquisitive questionnaire shown above. It's the reason why so many Mallory Vibrators are right for the job. With this information, Mallory engineers can make *intelligent* and *profitable* recommendations.

Do you have a supply of these "tell-all" questionnaires in your engineering files? If not, we earnestly suggest you give your Mallory representative a call—or write to Mallory direct. Do it now. And remember, too, that standard Mallory Vibrators are quickly available from authorized Mallory distributors.

Vibrators and Vibrapack* Power Supplies

P.R. MALLORY & CO., Inc.

MALLORY

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

SERVING INDUSTRY WITH

Capacitors
Contacts
Controls
Rectifiers
Switches
Vibrators
Power Supplies
Resistance Welding Materials

*Reg. U. S. Pat. Off.

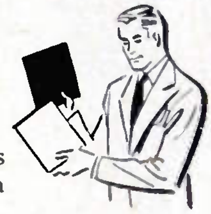
ANNOUNCING...Television's greatest picture tube improvement!

The New Rauland

1. Greatly diminishes tube face halation

2. Reduces reflection of ambient light

New Tube Vastly Increases Contrast!



Here is a scientific advance in picture tube technique which offers a marked, visible picture improvement already widely acclaimed by Television buyers.

In this newest Rauland development, the tube face is a new, practically colorless glass containing a metallic oxide which produces uniform light attenuation throughout the visible range. This light-absorbing characteristic acts in two ways to increase picture contrast, clarity and detail.

In ordinary tubes, light from bright picture areas of the screen striking the exterior tube face surface in angles greater than 48 degrees is completely reflected onto dark screen areas, reducing the apparent blackness. This halation is greatly reduced with the new Rauland "Luxide" screen, because such reflected light is attenuated by passing three times through the glass before it can reach the eye.

Similarly, under normal conditions, clear glass picture tubes can have their maximum contrast only when operated in an otherwise dark rocm. Ambient light passing through the tube face to the phosphor causes the dark picture area to appear light in tone and causes the picture to "wash out." With the "Luxide" screen, such ambient light must pass twice through the attenuating glass while light originating in the phosphor passes through the glass only once. The result is a picture with far greater contrast when viewed in lighted rooms, and since several more steps in the grey scale are available in forming the picture, better detail as well as greater contrast results.

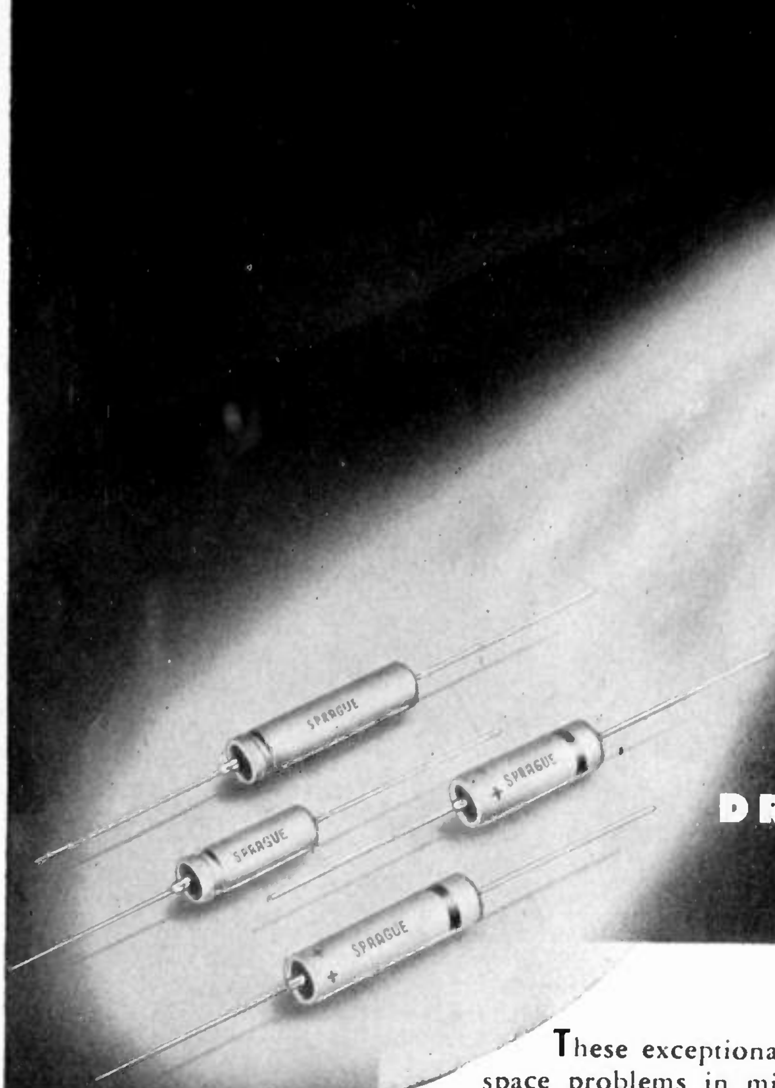
Write for Technical Bulletins. The Luxide screen is available in metal-cone types 16AP4-A, 16EP4-A and 12UP4-A, and in the all-glass 12LP4-A.

THE RAULAND CORPORATION

Perfection Through Research

4245 N. KNOX AVENUE · CHICAGO 41, ILLINOIS

SPACE SAVERS *de luxe*



SPRAGUE
MINIATURE DRY ELECTROLYTICS
Types 16D and 18D

These exceptionally small capacitors really solve space problems in miniaturized electronic and radio equipment. And their performance characteristics actually surpass those of ordinary metal encased tubular dry electrolytic capacitors!

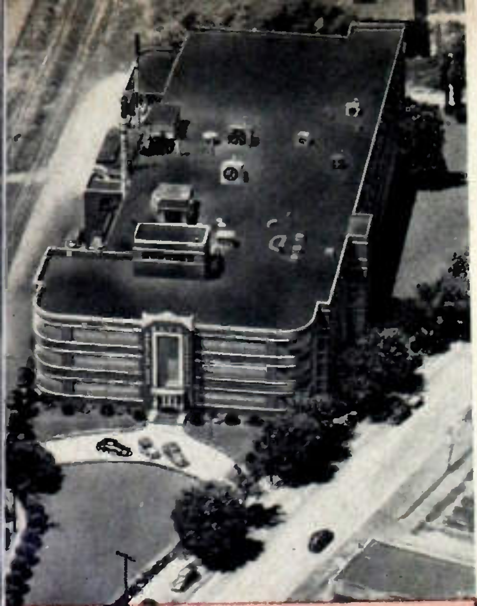
Sealed against moisture, Types 16D and 18D electrolytics are normally furnished for operation at 85° C. to meet the high operating temperatures common in crowded assemblies. Type 18D has an outer insulating tube over the metal case, whereas Type 16D does not have this extra covering.

Write for Sprague Engineering Bulletin No. 303 for complete details.

SPRAGUE

SPRAGUE ELECTRIC COMPANY
North Adams, Mass.

PIONEERS IN ELECTRIC AND ELECTRONIC DEVELOPMENT

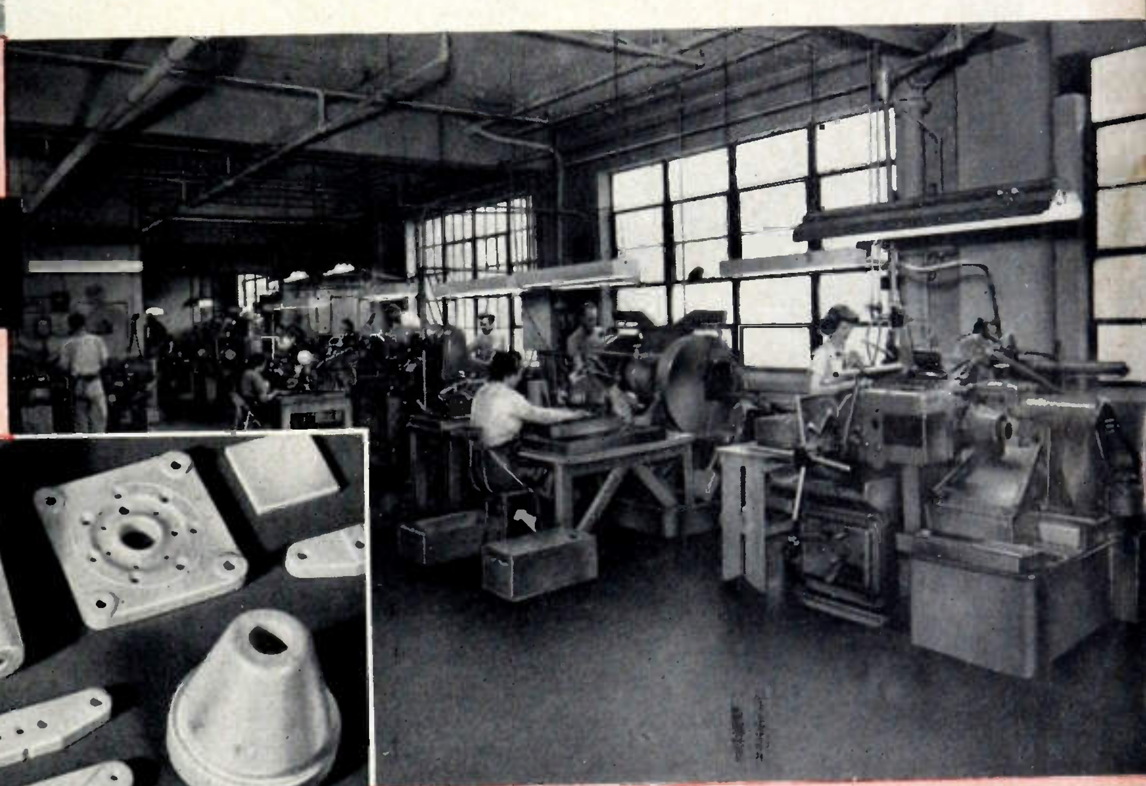


grinding

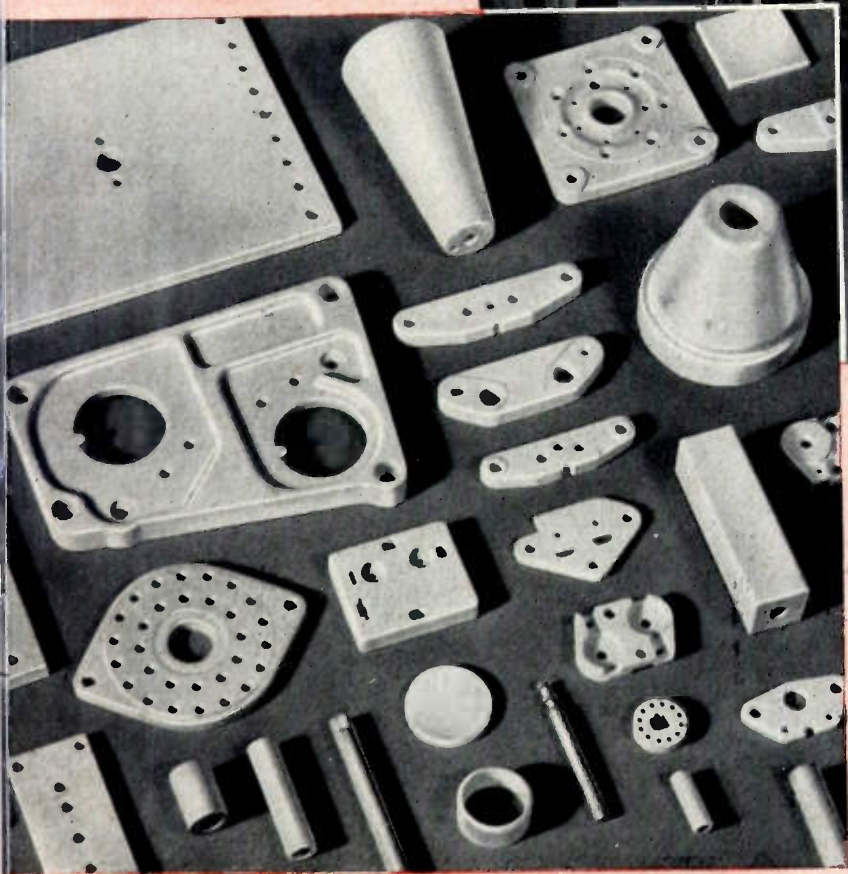
by skilled operators

enables AlSiMag to meet

unusual dimensional tolerances



ALSIMAG
Reg. U. S. Pat. Off.



After firing, AlSiMag is extremely hard. Further finishing requires special tools, great skill. We have the tools and the skill and can meet almost any tolerance required. The closer tolerances involve commensurate cost. Even if you think your requirements are impossible, ask us. It is probable that we can solve your problem . . . well within practical cost limits. Ability to consistently comply with dimensional and physical requirements is another reason why American Lava Corporation is known as Headquarters for Custom Made Technical Ceramics.

AMERICAN LAVA CORPORATION

48TH YEAR OF CERAMIC LEADERSHIP

CHATTANOOGA 5, TENNESSEE

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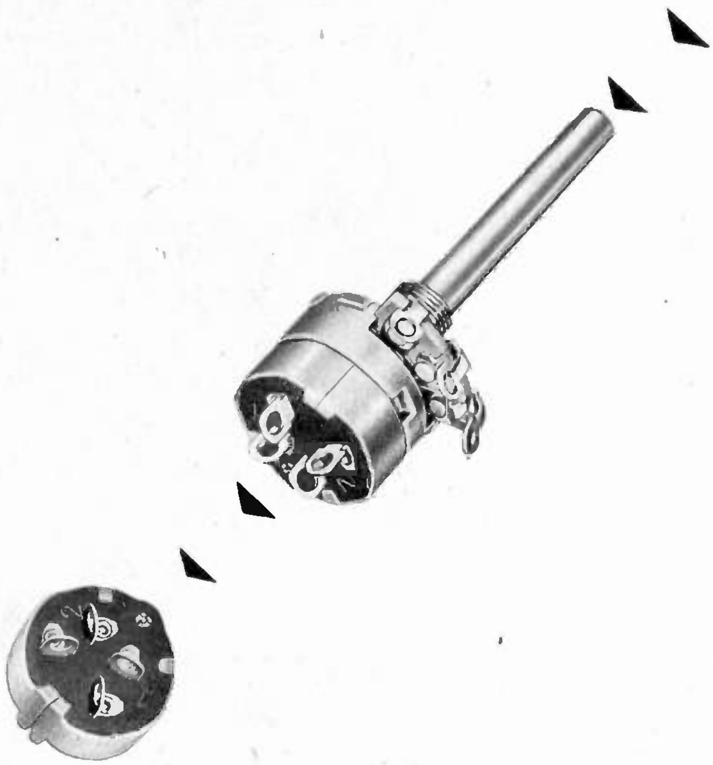
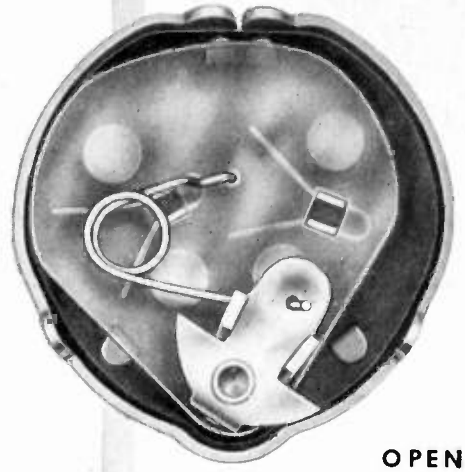
NEW ENGLAND, 38-B Brattle St., Cambridge, Mass., Kirkland 7-4498 • ST. LOUIS, 1123 Washington Ave., Garfield 4959

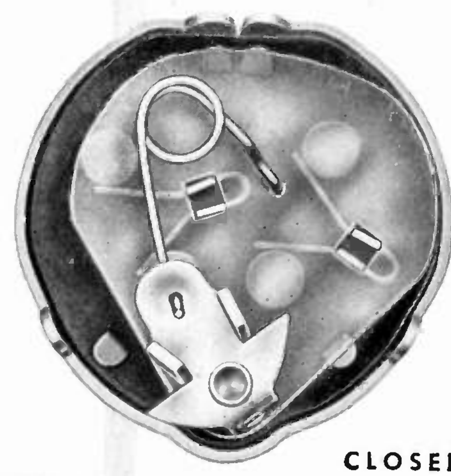
a simplified, outstandingly dependable LINE SWITCH for Stackpole Controls



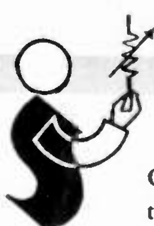


OPEN

(Interior views approximately 2½ times actual size of switch)



CLOSED



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

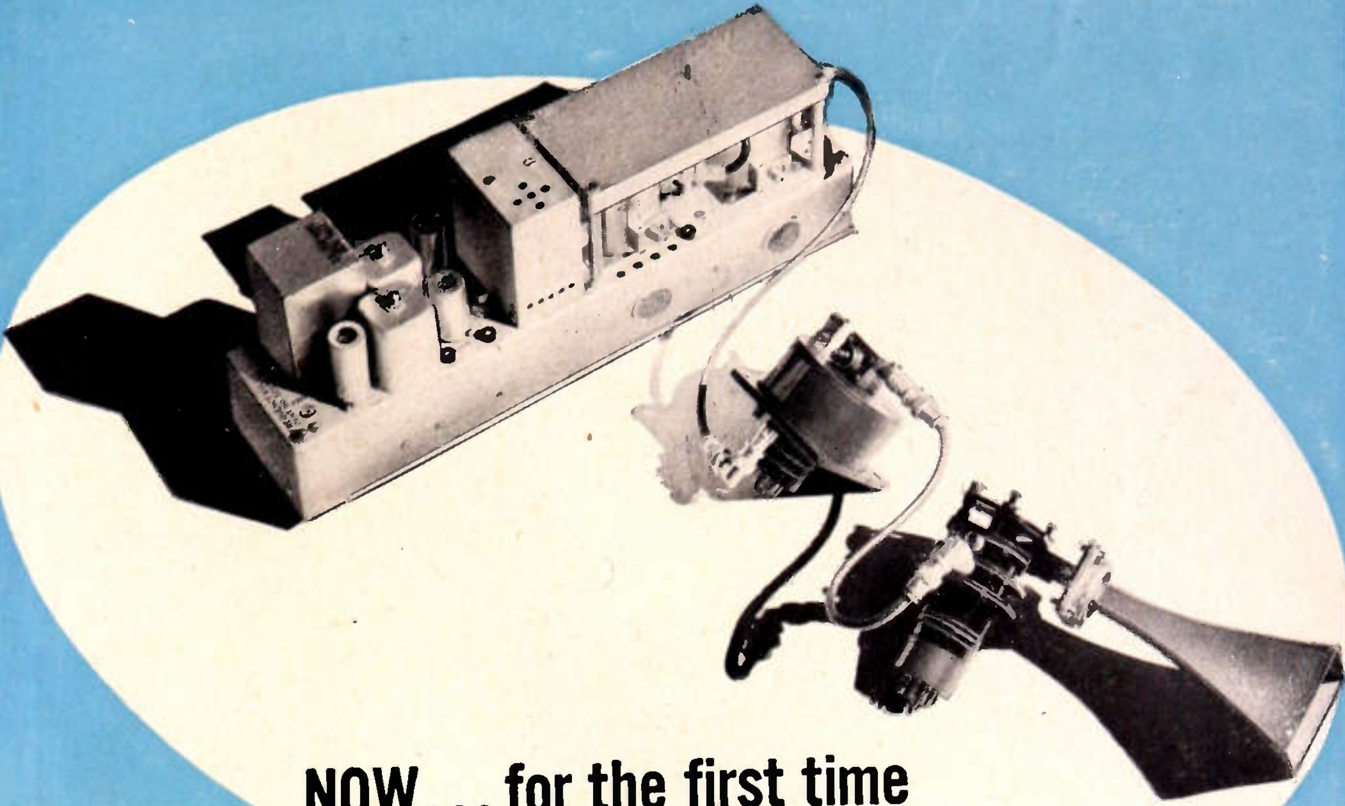
Write for Stackpole Bulletin RC-7

ELECTRONIC COMPONENTS DIVISION

STACKPOLE CARBON COMPANY, ST. MARYS, PA.

STACKPOLE

VARIABLE RESISTORS FOR MODERN RADIO AND TELEVISION NEEDS



NOW... for the first time

SUBSTANTIAL POWER

At Microwave Frequencies with Direct Crystal Control

Now, with two new Sperry Klystron tubes, stabilized frequency control is possible at 10,000 mc. with 1 watt continuous wave power output. These multiplier tubes, the SMC-11 and the SMX-32, permit direct crystal control at microwave frequencies with this power level.

Starting with a 5 mc. crystal, the frequency is multiplied to 830 mc. by use of an *Exciter*. The SMC-11 Klystron multiplies the 830 mc. to a frequency of 5,000 mc. The SMX-32 then multiplies this frequency to 10,000 mc. with the same accuracy which exists in the control crystal (±0.0005%).

This practical achievement of 1 watt power output with continuous accuracy of frequency control at 10,000 mc. exists only through the use of these two Sperry Klystrons.

Write our Industrial Department for further information.

SPERRY

GYROSCOPE COMPANY

DIVISION OF THE SPERRY CORPORATION
GREAT NECK, NEW YORK

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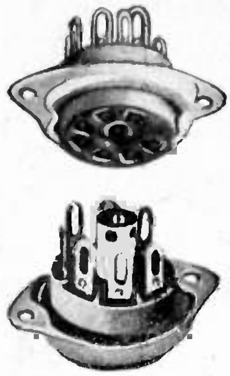
NEWS and NEW PRODUCTS

DECEMBER, 1949

Miniature Sockets in Mycalex

A new organization, **Mycalex Tube Socket Corporation,** operating under exclusive license of Mycalex Corporation of America have started the manufacture of 7 pin miniature tube sockets, utilizing precision molded Mycalex as an insulator. The sockets are obtainable in Mycalex 410 which was developed for applications requiring close dimensional tolerances not possible in ceramics and at much lower loss factor than mica filled phenolic with advantage in economy, the manufacturer claims; and in Mycalex 410X which has been developed to compare favorably with general purpose bakelite in economy but with a loss factor of only about one-fourth of that material.



Illustrated are two views in actual size of the 7 prong socket. Sockets with different numbers of prongs are in the development stage.

These sockets are manufactured to precise specifications and fully meet RMA standards. Further information is obtainable from **Mycalex Tube Socket Corporation,** 30 Rockefeller Plaza, New York 20, N. Y.

New Low-Priced Oscillograph

A new low-priced, lightweight oscillograph especially designed for use in schools, colleges, and industrial laboratories has been announced by **General Electric Co., Meter and Instrument Div.,** Schenectady 5, N. Y.

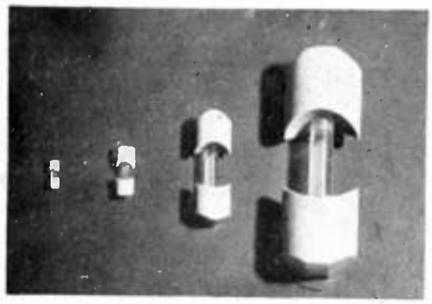
Compact and simple in design, the Type PM-18 instrument is said to be easily operable by inexperienced personnel, in the laboratory, or in the field. It can be used either for visual indications, or for taking oscillograms of current and voltage phenomena.

Additional information is contained in Bulletin GEC-580 which is available by writing to the company.

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

Prodelin Transmission Line

A new 51.5-ohm air-dielectric coaxial transmission line (rigid) with center conductor impedance equivalent to that of air for use in any given position of the spectrum up through the microwave frequencies, is available from the manufacturer, **Product Development Co. Inc.,** 526 Elm St., Arlington, N. J.



Physical properties of a new press-fit construction provide a line with accurate physical tolerances that will withstand high-impact shock or exposure to rapid temperature changes. A threaded self-sealing coupling has also been developed for use at frequencies where a minimum VSWR is required.

New 75-Watt Lamp Ballast

A new lamp ballast has been developed by the engineering department of **Acme Electric Corp.,** 44 Water St., Cuba, N. Y. for use with the new 75-watt, 96-inch T-12 slim-line lamp.

The manufacturer claims that this ballast provides an output of 75 watts, 0.425 amperes, a minimum of 700 volts starting and 195 volts operating. With these electrical characteristics the lamp can be expected to operate with a 4,800 lumen output after 100 hours, and an average brightness of 1,600 foot lamberts. This represents an average of 64 lumens per watt.

New Standard Signal Generator

Measurements Corp., Boonton, N. J., have announced the production of a new standard signal generator covering the wide frequency range of 20 cps to 50 Mc.

This instrument, the Model 82, was designed to provide in one signal generator a continuously variable signal source for most measurements at audio, supersonic, and radio frequencies. Two oscillators are employed to cover the frequency range. The low frequency oscillator, continuously variable from 20 cps to 200 kc, has a metered output from 0 to 50 volts across a resistance of 7,500 ohms. A radio-frequency oscillator covering the range from 80 kc to 50 Mc provides output from 0.1-microvolt to 1 volt, and may be modulated with the low-frequency oscillator.

An improved mutual-inductance type attenuator is said to insure a higher degree of accuracy than may be obtained with the resistor or mutual-inductance type attenuator of earlier design.

Recent Catalogs

• • • An illustrated brochure describing the greatly expanded facilities of The Franklin Institute Research and Development Laboratories may be obtained by writing to **Administration Div., The Franklin Institute Laboratories,** Benjamin Franklin Pkwy., at 20th St., Philadelphia 3, Pa.

• • • The sixth edition of "Johnson Antenna Handbook" was recently published by the **E. F. Johnson Co.,** Waseca, Minn., and is now available from Johnson Jobbers at 60 cents a copy.

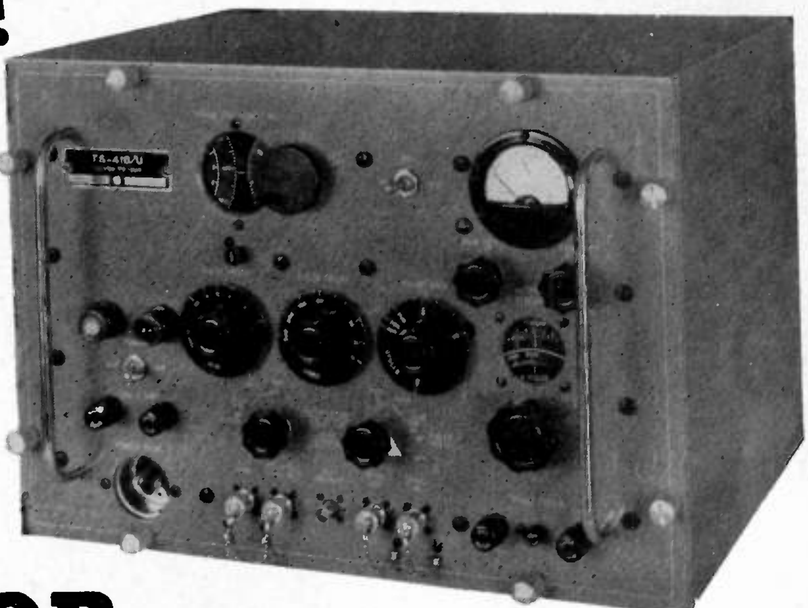
• • • Just released by **Triad Transformer Mfg. Co.,** 2254 Sepulveda Blvd., Los Angeles 64, Calif., a 16-page catalog, TR-49, describing and pricing the entire line of Triad transformers for original equipment, radio and television, replacement, and amateur applications.

(Continued on page 61A)

New!

ACCURATE! CONVENIENT!

PORTABLE!



UHF SIGNAL GENERATOR

Covers the Range of 400 – 1000 MC.

* * * The LAVOIE LA-418 Signal Generator, newest addition to the LAVOIE LABORATORIES' line of precision electronic equipment...

Provides:

- **DIRECT READING Frequency Dial.**
- **DIRECT READING Attenuator calibrated in D B** (0 TO - 120 DBM) **U Volts.**
- **INTERNAL and EXTERNAL Pulse Modulation sine wave modulation external.**

A complete descriptive folder is available promptly on request.
WRITE FOR TECHNICAL BULLETIN LA-418

Lavoie Laboratories
RADIO ENGINEERS AND MANUFACTURERS
MORGANVILLE, N. J.

***Specialists* in the Development and Manufacture of UHF Equipment**

WHY STAND ON YOUR HEAD?

1 FACTORY = 1 SOURCE

1 NEDA DISTRIBUTOR = 100 FACTORIES

WHY chase around the countryside, from factory to factory, looking for this or that odd electronics part, when your NEDA Distributor carries the *diversified* supplies you need?

Why lose valuable time, scheduling a pilot run, when your NEDA Distributor has the quantities you need in stock?

Why stand on your head, trying to explain what you want, when your NEDA Distributor understands your specifications and gives you intelligent cooperation, on all your electronics needs?

On the industrial front, more and more engineers and purchasing agents are doing business with the NEDA Distributor . . . because he expedites your purchase of radio and electronics parts with *greater selection . . . faster delivery* . . . and *on-the-spot consultation and service.*

FOR EXAMPLE:
NEDA Distributors Carry

Square-wave Generators
Master Music Systems
Cathode Ray Tubes
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Voltohmmeters
Potentiometers
Oscillographs
Transformers
Microphones
Condensers
Thyratrons
Amplifiers
Photocells
Recorders
Rheostats
Cabinets
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NATIONAL ELECTRONIC DISTRIBUTORS ASS'N — NEDA

NATIONAL ELECTRONIC DISTRIBUTORS ASSOCIATION
INCORPORATED
2214 LaSALLE-WACKER BUILDING, 221 N. LaSALLE ST., CHICAGO 1, ILLINOIS

Consult the list of NEDA members on the opposite page.

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Thomas H. Brown Co.
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Hatry & Young
STAMFORD
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WATERBURY
Hatry & Young
The Bond Radio Supply

DISTRICT OF COLUMBIA
WASHINGTON
Electronic Wholesalers, Inc.
Kenyon Radio Supply Co.
Rucker Radio Wholesalers
Silberne Radio & Elec. Co.

FLORIDA
MIAMI
Herman Radio Supply Co.
Walder Radio & Appliance

GEORGIA
AUGUSTA
Prestwood Electronics Co.
COLUMBUS
Radio Sales & Service Co.

IDAHO
BOISE
Craddock's Radio Supply
Kopke Electronics Co.

ILLINOIS
BELLEVILLE
Lurtz Electric Co.
BLOOMINGTON
Alverson Sales Co.
CHICAGO
Allied Radio Corp.
J. G. Bowman & Co.
Chauncey's, Inc.
The Lukko Sales Corp.
Pilgrim Distributing Co.
Radio Parts Co.
Walker-Jimieson, Inc.
MOLINE
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BALTIMORE
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MASSACHUSETTS
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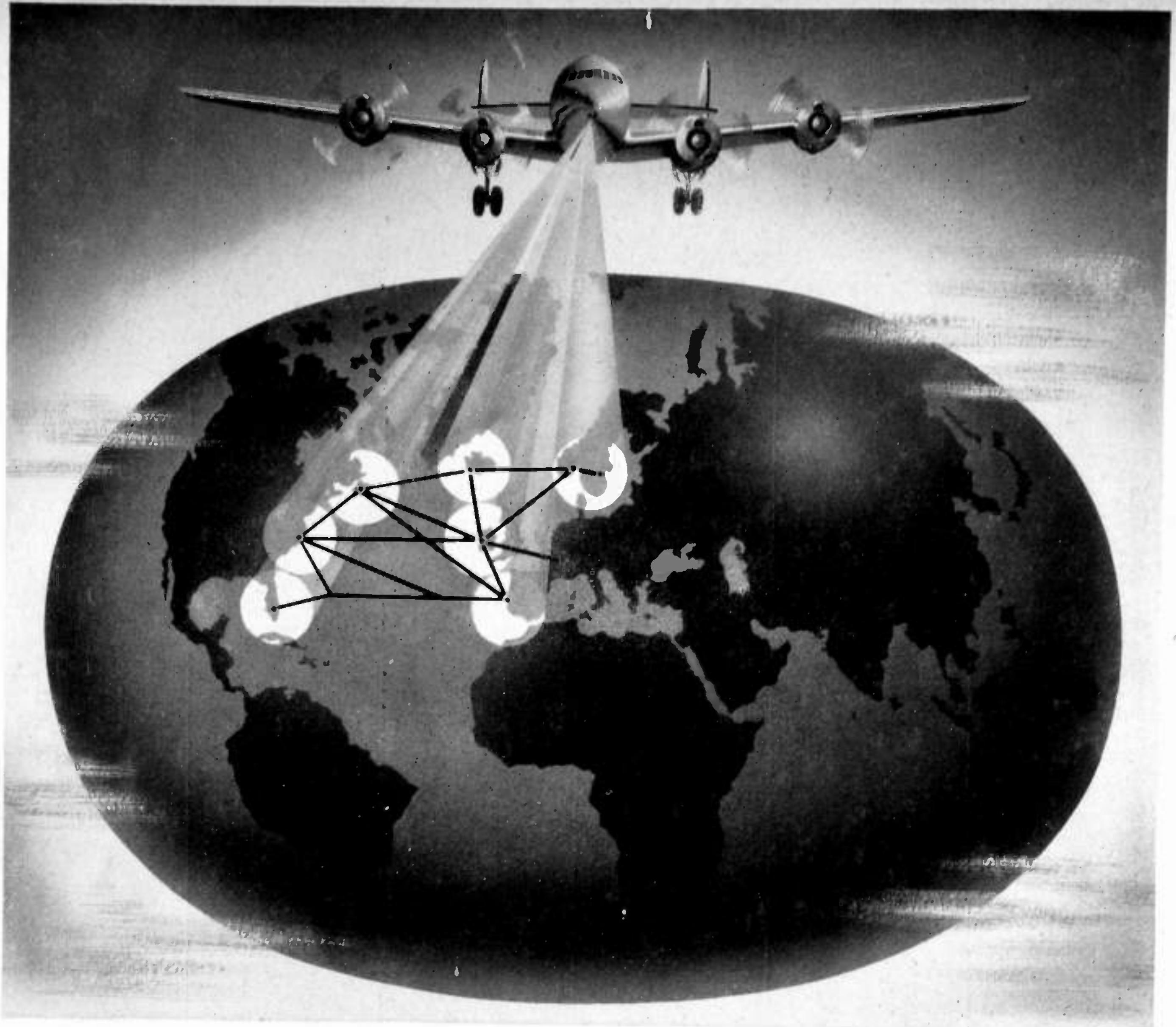


WILCOX ...First Choice for Transatlantic Airline Communication

The whirling propellers of the international air lines make daily mockery of the vast space of the Atlantic Ocean. Intercontinental passengers and cargo come and go hourly at New York, Miami, Gander; Shannon, Ireland, and Lisbon, Portugal. These European and American airports are equipped with modern long-range, multichannel WILCOX Transmitters.

Oslo, Norway, and Stockholm, Sweden, use WILCOX Transmitters as basic communications equipment, and radio beacon service is provided at Reykjavik, Iceland, by WILCOX Type 96-200 Transmitters.

Thus, the giant airliners of the world's major airways are protected in flight and guided safely to the runways of Europe's and America's principal ports of entry.

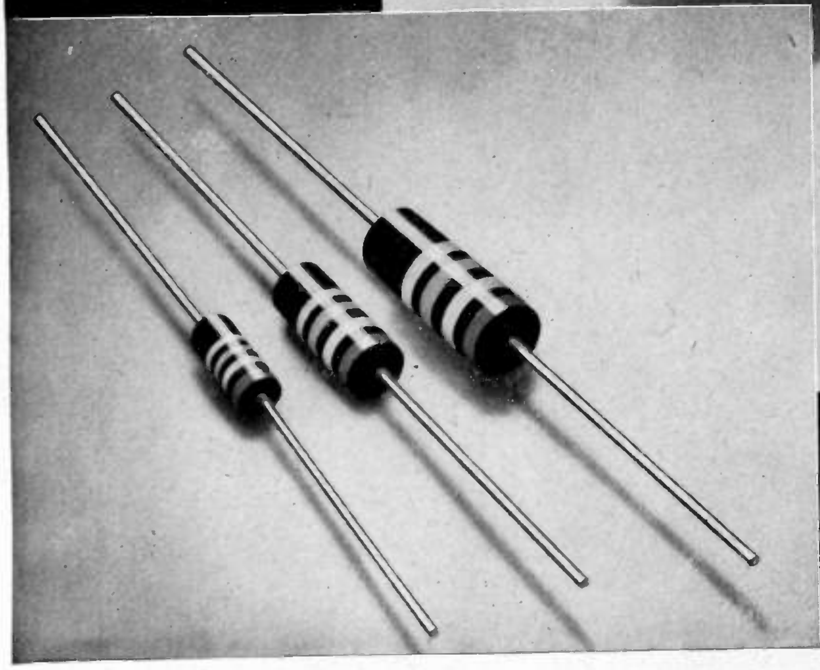
WRITE TODAY...for complete information on air-borne, ground station, point-to-point, or shore-to-ship communications equipment.

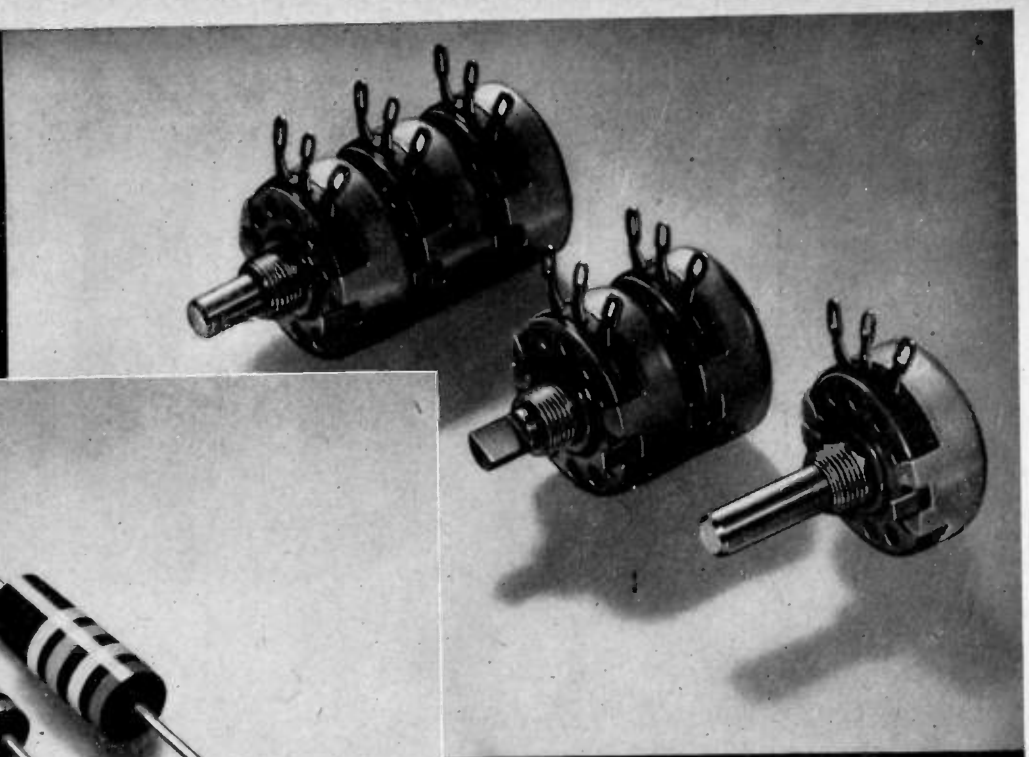
WILCOX ELECTRIC COMPANY

KANSAS CITY MISSOURI

FIXED RESISTORS

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





ADJUSTABLE RESISTORS

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

For circuits that require resistors of unsurpassed quality ...*Specify* Allen-Bradley

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

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

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

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

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

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

A-B Quality

ALLEN-BRADLEY

FIXED & ADJUSTABLE RADIO RESISTORS

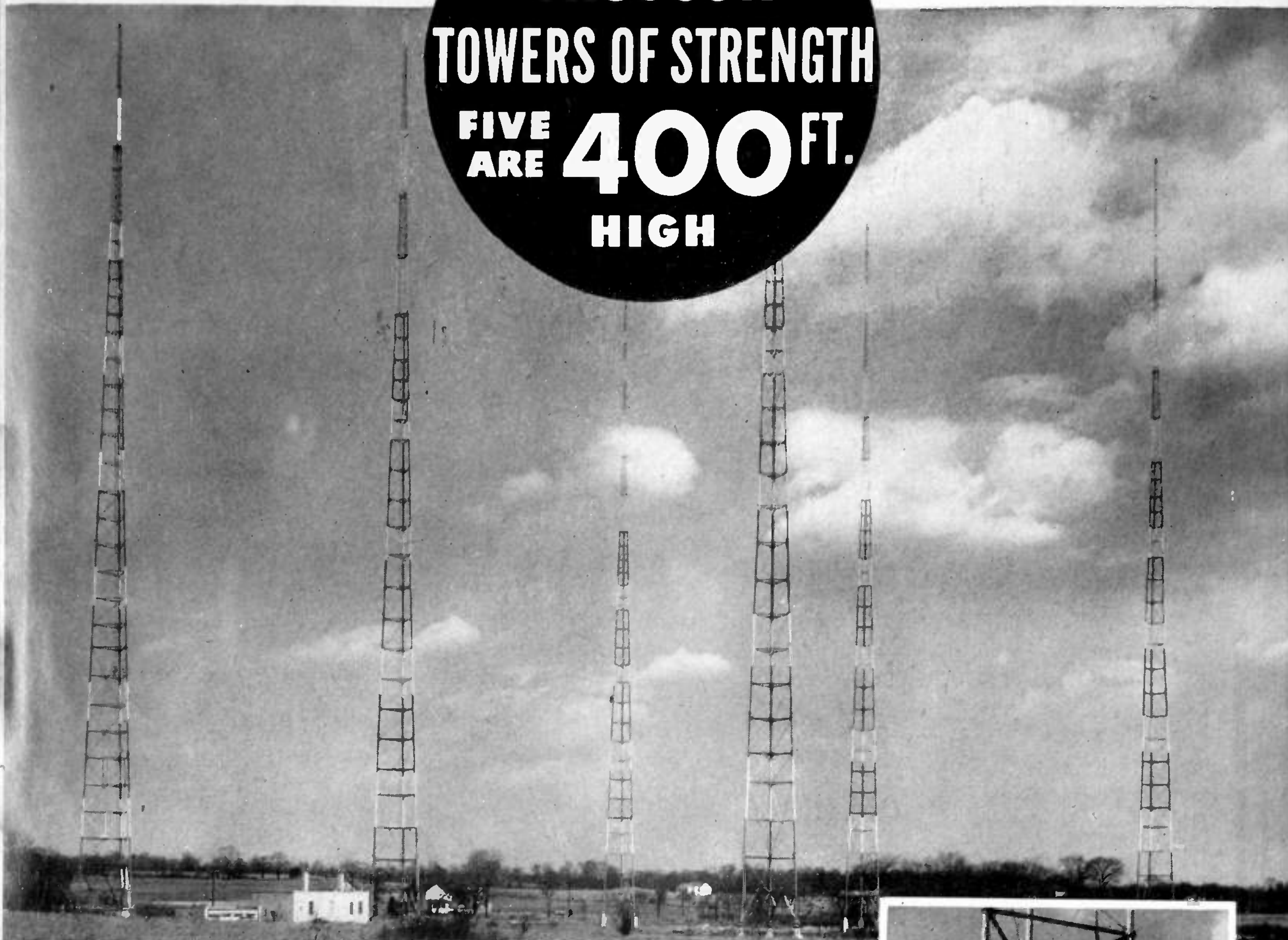
Sold exclusively to manufacturers QUALITY of radio and electronic equipment

Six More

TRUSCON

TOWERS OF STRENGTH

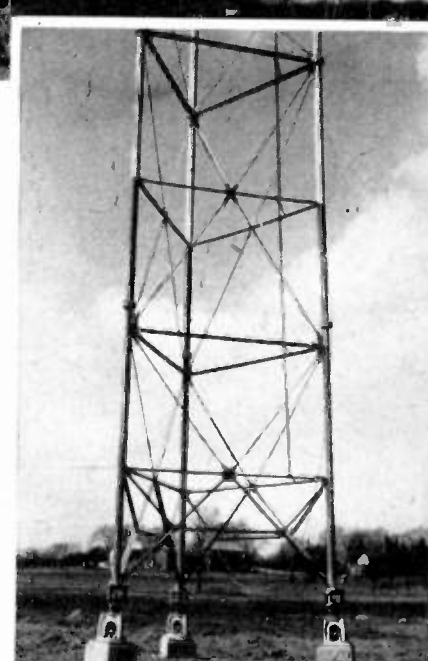
FIVE ARE 400 FT. HIGH



YOU'RE LOOKING AT 2,400 soaring feet of self-supporting radio towers—Truscon-engineered and erected for WFMJ Broadcasting Company in Youngstown, Ohio. These sturdy steel structures climb 400 feet above the Mahoning Valley. One tower carries an RCA 4-section Pylon FM antenna. Together, they give 5,000-watt WFMJ top coverage of the bustling eastern Ohio-western Pennsylvania industrial area.

Competition for Youngstown dialers is intense, with nearby Cleveland and Pittsburgh broadcasters pouring 50,000-watt signals into the market. Facing this problem, Truscon furnished a tower set-up that was exactly right for WFMJ's needs—and then erected the towers for best operating efficiency.

It's one more example of the way in which Truscon engineers tackle *purely local problems* — operational or geographical — in any part of the world. Truscon can engineer and erect exactly the towers you need . . . tall or small . . . guyed or self-supporting . . . for AM, FM or TV. Your phone call or letter to our home office, or to any close-by Truscon District Office, will bring you helpful assistance without obligation.



WFMJ Broadcasting Station, Poland—Boardman Road, Youngstown, Ohio. 6 Truscon Self-Supporting Towers. One Tower is 346 ft. high with RCA 4-Section Light or Heavy Duty Pylon. Five Towers each 400 ft. high. Shows base of one tower.

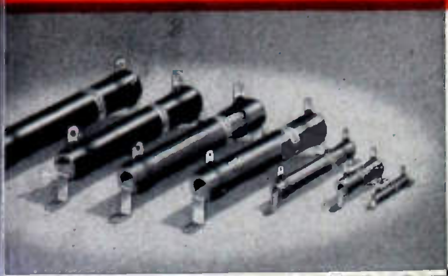
TRUSCON
STEEL COMPANY
YOUNGSTOWN 1, OHIO
Subsidiary of Republic Steel Corporation

TRUSCON
SELF-SUPPORTING AND UNIFORM CROSS SECTION GUYED
TOWERS

TRUSCON COPPER MESH GROUND SCREEN

OHMITE Resistors

Sizes and Types for Every Service

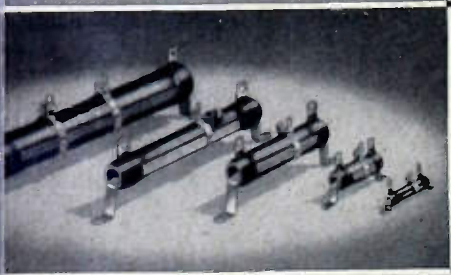


LUG TYPE

Most popular type for general purpose applications. Connected by soldering or bolting to lugs. Protected by vitreous enamel coating.

FERRULE TYPE

Winding terminated on metal bands for mounting in standard fuse clips. Provides easy interchangeability without tools.

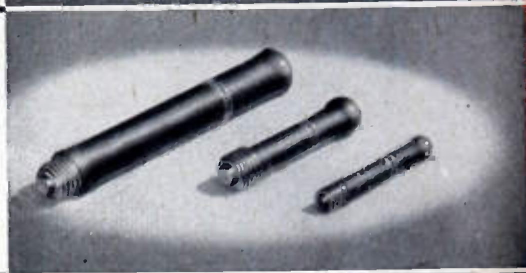


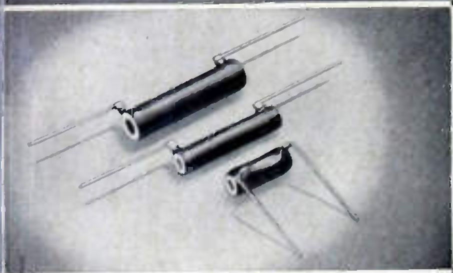
"DIVIDOHM" ADJUSTABLE TYPE

Provided with adjustable lugs for securing odd values of resistance quickly and easily.

EDISON BASE TYPE

Mounted in ordinary lamp type screw sockets for easy interchangeability without the use of tools.



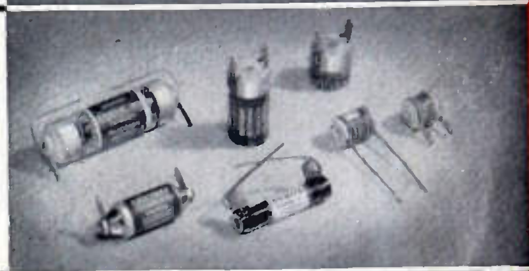


WIRE LEAD TYPE

Small vitreous enameled resistors which can be connected and supported by their own wire terminals. Maximum size approx. 20 watts.

PRECISION TYPE

Low wattage resistors of ± 1% or closer tolerance. Made in vacuum impregnated, glass sealed, or vitreous enameled type units.

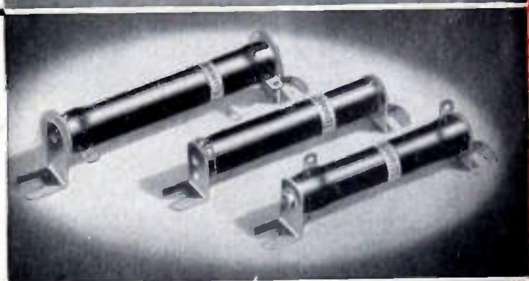


FLEXIBLE LEAD TYPE

Winding is connected to stranded bare or insulated leads. Used where it is desired to have connecting wires a part of the resistor.

BRACKET TYPE

Have metal end brackets. Live bracket type is connected by bolting brackets to panel terminals. Dead bracket type has separate lugs.

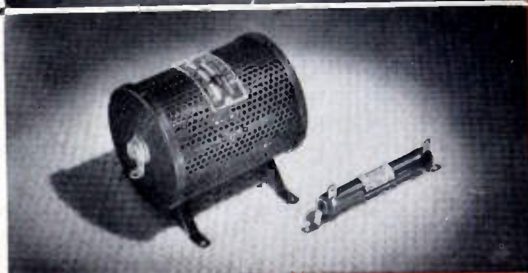


"CORRIB" TYPE

Has edge-wound, exposed corrugated ribbon winding. For low resistances where 100 watts or more must be dissipated in small space.

NONINDUCTIVE TYPE

For radio frequency circuits where constant resistance and impedance are required. Made in rugged, vitreous-enameled type construction.



In addition to the many types of resistors shown above, Ohmite offers resistors in more than sixty different core sizes, and a wide range of wattages and resistance values. Ohmite engineers will be pleased to help you in selecting the right resistors for your needs.

OHMITE MANUFACTURING CO.

4862 Flournoy Street Chicago 44, Illinois

Write on Company Letterhead for Catalog and Engineering Manual No. 40.

Contains 96 pages of useful data on the selection and application of rheostats, resistors, tap switches, and other equipment.

Be Right with...

OHMITE

RHEOSTATS · RESISTORS
TAP SWITCHES

Industry's First Choice

How CTC worked with an Equipment Manufacturer

NEW COINCIDENCE ANALYZER USES CTC TERMINAL BOARDS

Atomic Instrument Company's Coincidence Analyzer, used with Geiger Counters, steps up accuracy of coincidence experiments in cosmic ray and nuclear physics research and general radioactivity measurements.

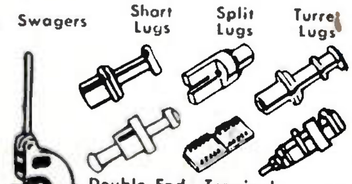
Teamwork between the engineering departments of Atomic Research Company and CTC licked a serious time problem in the manufacture of the Coincidence Analyzer. Atomic Research Company specified what they needed in terminal boards and CTC saw to it that they got what they wanted in a hurry. Five Terminal Boards made of laminated phenolic and equipped with standard CTC feed-through and single-ended lugs comprise CTC's contribution to this excellent piece of equipment.

This is but one of many cases where CTC has cooperated with electronic and radio manufacturers with gratifying results. We are set up to create and produce assembled terminal boards to meet just about any specifications. Special terminal lugs, coils and chokes to fit particular requirements are also part of our custom engineering service.

WHAT IS YOUR PROBLEM?

When you need terminal boards, check with us before your designs are too far advanced. Our engineers' long experience with laminated insulating materials offer you expert analysis and satisfactory recommendations, promptly. CTC's broad line of terminals usually fulfills any requirement.

Custom or Standard
The Guaranteed Components



CAMBRIDGE THERMIONIC CORP.
456 Concord Ave., Cambridge 38, Mass.

Save Space and Weight with TRIAD "HS" Transformers

1939 Transformer
Dims: $3\frac{3}{8}'' \times 3\frac{3}{8}'' \times 3\frac{7}{8}''$
Weight: $2\frac{3}{4}$ lbs.

TRIAD "HS" Audio Input Transformer
Dims: $1\frac{7}{16}'' \times 1\frac{3}{4}'' \times 2\frac{1}{2}''$
Weight: 12 oz.

Both transformers shown above are high fidelity input transformers, frequency response from 20-20,000 cycles and 95db. shielding.

Yet the Triad transformer is only **one-seventh as large by volume,** occupies **one-fourth the space** and is **one-fourth as heavy.** In the production of today's high fidelity equipment, where space is at a premium, that's important.

Triad "HS" (hermetically sealed) transformers, built to meet JAN specifications, are providing new standards of performance for quality electronic equipment — yet they cost little more than ordinary cased types.

Triad builds a complete line of transformers for original equipment, replacement, geophysical and amateur applications.

Write for Catalog TR-49

TRIAD TRANSFORMER MFG. CO.
2254 Sepulveda Blvd.
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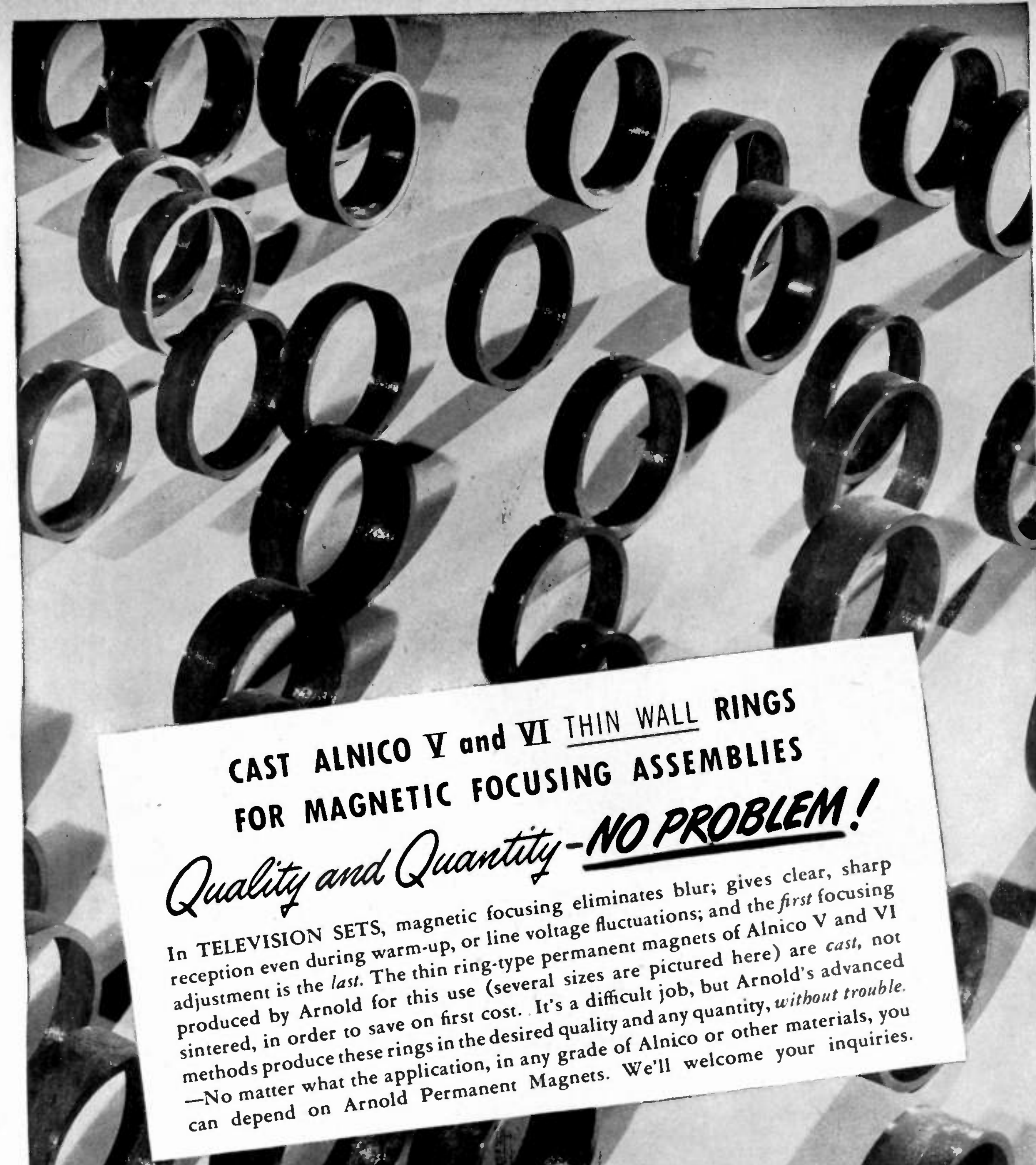

THE ARNOLD ENGINEERING COMPANY

Subsidiary of
ALLEGHENY LUDLUM STEEL CORPORATION
147 East Ontario Street, Chicago 11, Illinois

Specialists and Leaders in the Design, Engineering and Manufacture of PERMANENT MAGNETS

. . . the economy of thoriated-tungsten filaments and improved cooling in high-power tubes

Here is unparalleled tube value . . .

Five new RCA tubes, ranging in power input from 1.5- to 150-kw, and successfully utilizing economical thoriated-tungsten filaments which offer marked savings in filament power and the cost of associated power equipment.

Five tubes with proved features of previous similar types. Two—the 5762 and 5786—have efficient newly designed radiators that permit the use of less expensive blowers.

Five tubes with improved internal constructions that contribute to their more efficient operation and longer service life.

These five new RCA tube types are "musts" for designers of broadcast, communications and industrial electronic equipment where design and operating economies alike are important considerations.

Forced-air-cooled assemblies and water-jacket assemblies are available for most RCA power tubes.

RCA Application Engineers are ready to consult with you on the application of these improved tubes and accessories to your specific designs. For complete technical information covering the types in which you are interested, write RCA, Commercial Engineering, Section 47LR, Harrison, New Jersey.

TUBE DEPARTMENT

RADIO CORPORATION of AMERICA

HARRISON, N. J.

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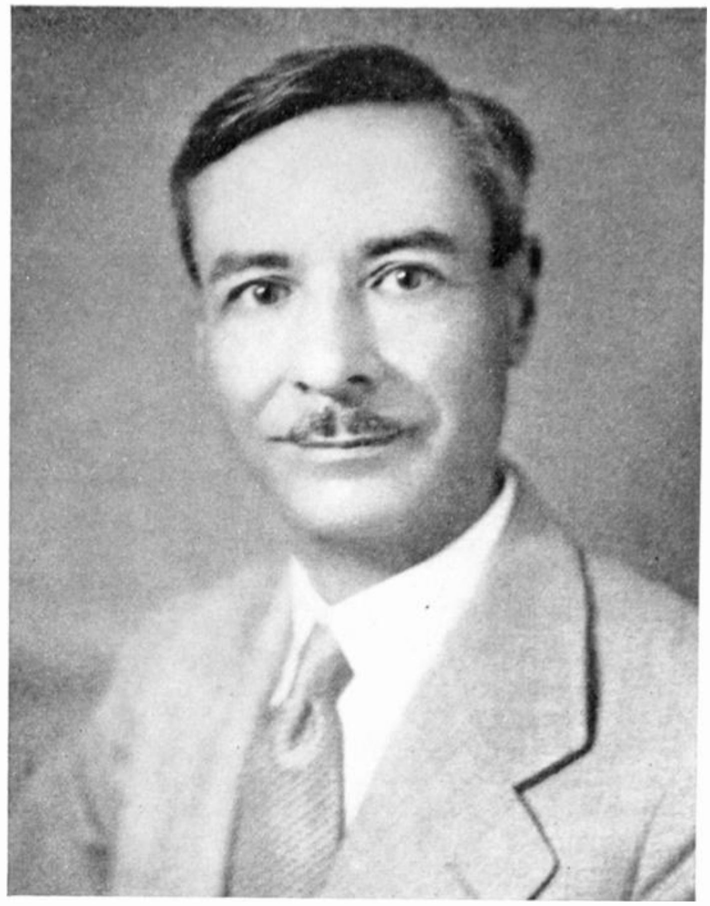

Haraden Pratt

SECRETARY, 1943–1949

Haraden Pratt, Vice-President and Chief Engineer of the American Cable and Radio Corporation, was born in San Francisco, Calif., on July 18, 1891. He commenced his career in radio as an amateur in 1906, and became a wireless telegraph operator and installer of equipment for the United Wireless Telegraph Company and Marconi Wireless Telegraph Company of America during the years 1910–1914.

Mr. Pratt, who received the B.S degree in electrical engineering from the University of California in 1914, became a construction and operating engineer for the Marconi Company's 300-kilowatt spark-type Trans-Pacific radio stations in California.

From 1915 to 1920 he was an Expert Radio Aide for the Navy Department and was primarily occupied with the construction and maintenance of its high-powered radio stations. Commencing in 1920 he established the public service radiotelegraph system of the Federal Telegraph Company on the Pacific Coast. In 1925 he constructed and operated a radiotelegraph system between Salt Lake City and Los Angeles for the Western Air Express, which was followed in 1927 by development work on radio aids for air navigation of which he was in charge at the Bureau of Standards, Department of Commerce, Washington, D. C. In 1928 he became chief engineer and later vice-president of Mackay Radio and Telegraph Company. He constructed its world-wide communication plant.

Mr. Pratt has attended most international radio and telegraph conferences since 1926 as either a technical or industry adviser. He was a director of the American Standards Association from 1939 to 1942, chairman of the Radio Technical Planning Board from 1945 to 1948, and is now a member of the Joint Technical Advisory Committee.

During World War II he was chief of the National Defense Research Committee's Division 13 on Communications, and in 1948 was awarded a Presidential Certificate of Merit.

Mr. Pratt is a member of Sigma Xi, Fellow of the American Institute of Electrical Engineers, Fellow of the Radio Club of America, and an Associate Fellow of the Institute of the Aeronautical Sciences.

He joined the Institute as an Associate in 1914, became a Member in 1917, and a Fellow in 1929. He has been a Director since 1935, Treasurer in 1941–1942, Secretary since 1943, and President in 1938. Mr. Pratt received the Institute's Medal of Honor in 1944.

By decision of the Board of Directors of the Institute, a major new service to the IRE membership has been established. Its genesis and purpose are described in the following statement from the Technical Editor.—*The Editor.*

IRE Standards Publication Policy

Standardization is essential to the orderly expansion and progress of a technological field.

The Institute of Radio Engineers, sensible of its responsibilities in this regard, appointed a Committee on Standardization immediately following its foundation in 1912, and the next year published a report dealing with definitions of terms, letter and graphical symbols, and methods of testing and rating equipment. Since that time, standardization activities have played a permanent and prominent role in Institute affairs, resulting in the eventual formation of more than twenty technical committees. These activities have contributed appreciably to the advancement of the electronic and communication art, bringing conformity and clarity to all fields of endeavor.

The Standards publication program of the Institute has been modified from time to time in the past to meet the needs of a rapidly growing field. The initial Standards report of 1913 was succeeded by revised reports appearing in 1915, 1922, 1926, 1928, 1931, and 1933. Each report contained, in a single document, data on all branches of the art. Due to the rapid strides made in certain fields, the Committee on Standardization, during 1924 and 1925, was supplemented with a number of subcommittees, each one of which concerned itself with one specialized branch. As a result of this change in committee structure, the 1926 and subsequent Standards reports each were separated into several sections. Eventually it was found desirable to give these subcommittees full standing-committee status. Accordingly there was initiated in 1938 a new series of Standards in which each Standard dealt with a separate field. More recently, this subdivision has been carried further by the publication of separate Standards within each field on definition of terms, on symbols, and on measuring and testing methods.

The rapid growth of the electronic and communications field has occasioned a correspondingly large increase in Institute membership. As a result, the method of distributing Standards heretofore has become increasingly inadequate. **Therefore, the Board of Directors plans as a continuing procedure, subject to unforeseen contingencies, to publish all Standards, prepared by the Technical Committees of the Institute, in the PROCEEDINGS OF THE I.R.E.** This is regarded as a valuable new service to the membership as it will make available to all members each Standard that is published, thereby ensuring the widest practicable distribution of Standards without additional cost to the members. As a matter of convenience, all issues of the PROCEEDINGS containing Standards will be identified by a red border on the spine and a corresponding notice in red on the front cover. In addition, those individuals who wish to maintain a separate file of IRE Standards may purchase reprints, while available, of those Standards published in the PROCEEDINGS from Institute headquarters.

The Institute is therefore pleased to place before the membership in this issue of the PROCEEDINGS OF THE I.R.E. the first Standards to be published in accordance with the new Institute Standards publication policy.—*The Technical Editor.*

Standards on
RADIO AIDS TO NAVIGATION: DEFINITIONS OF TERMS, 1949*

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1. INTRODUCTION

1.1. Definition

"Navigation" is the process of finding the position of a vehicle, and directing it to reach a desired destination.

Note: Navigation is inherently three-dimensional. It is often reduced to two dimensions by projecting all positions, courses, and speeds upon the surface of the earth. The measurement of a navigational co-ordinate defines a surface of position. The intersection of this surface of position with the surface of the earth is the conventional line of position. The position of the vehicle is at the intersection of three surfaces of position; it may be so defined or it may be defined as the intersection of two lines of position (at the surface of the earth). Thus altitude is ordinarily dealt with independently as one co-ordinate, while the

* Reprints of this Standard, 49 IRE 12. S1, may be purchased while available from The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.60 per copy. A 20% discount will be allowed for 100 or more copies mailed to one address.

other two are converted into horizontal distance and direction or into latitude and longitude. When the vertical component of a course is of comparable importance to the horizontal components it is often advantageous to treat the navigational process in terms of three appropriate surfaces of position.

Note: Navigation is ordinarily carried out continuously throughout a journey. All observed quantities are functions of time.

1.2. Conditions

Navigation must be carried on under three sets of conditions that apply at various distances from the starting point, way points, or destination; that is:

1.2.1. At distances such that the knowledge of position alone is adequate to determine the proper course to the objective.

1.2.2. At distances such that the operation of other vehicles in the vicinity becomes a vital factor in the choice of procedure.

1.2.3. At distances such that the relation between the course of the moving vehicle and the positions of fixed bodies, such as docks or landing strips, assume paramount importance.

1.3. Operations

Navigation consists of four basic operations: dead-reckoning, fixing, pilotage, and homing.

1.3.1. "Dead-reckoning" is the procedure of advancing a known position to give a position at a later time by addition of one or more vectors representing known courses and distances.

Note: One or more of the vectors may represent currents in the sea or air. The distances are ordinarily found by measurements of speeds and time intervals.

1.3.2. "Fixing" is the determination of position without reference to any former position.

Note: In case the various elements of a "fix" are not obtained simultaneously, they may be converted to a common time. Having obtained two or more "fixes," at known time intervals, the navigator may determine or verify certain of the vectors which he uses in dead-reckoning.

1.3.3. "Pilotage" is navigation without computation of position, by directing a vehicle to its destination through observation of landmarks in the vicinity of the vehicle.

Note: Pilotage may be performed either by direct visual, aural, mechanical, or electronic observation.

1.3.4. "Homing" is approaching a desired point by following a route such that some navigational co-ordinate (other than altitude) is held constant.

1.4. Radio Fixing Aids

The radio fixing aids to navigation may be classified in terms of the geometrical relation between the vehicle and the known points, lines, or surfaces as follows:

1.4.1. Single Fixed Vertical Line in Space. (In two-dimensional navigation, the identification of a single point on the surface of the earth.)

Example: Zone (Z) marker.

1.4.2. Radial Lines-of-Position

a. Directional aids—the frame of reference is attached to the vehicle.

Example: Radio direction finding from the vehicle.

b. Azimuthal aids—the frame of reference is fixed with respect to the earth.

Example: Omnidirectional range.

1.4.3. Distance Measurement

a. Distance from one or more discrete points. (Circular lines of position.)

Example: Shoran, oboe, or distance measuring equipment (DME).

Note: The measurement involves transmission in both directions over the path. The measurement may be initiated from either the fixed points or the vehicle.

b. Distance from a line. (Cylindrical surface of position.)

Example: Maintenance of a signal of constant amplitude by traveling at a constant distance from a long wire radiating electromagnetic waves with uniform cylindrical symmetry.

c. Distance from a surface. (The surface of position dependent upon the reference surface.)

Example: Radio altimeter (reference is surface of earth).

1.4.4. Distance-Difference Measurement (hyperbolic lines of position).

The difference between the distances from two fixed points is measured without knowledge of either distance.

Example: Loran, gee.

1.4.5. Distance-Sum Measurement (elliptical lines of position).

The sum of the distances of two fixed points is measured without knowledge of either distance. This process is similar to 1.4.3 a, except that the transmitter and receiver of the transponder are separated by a fixed known distance.

1.4.6. Composite Aids to Fixing

The basic processes and co-ordinate systems, outlined above, are not mutually exclusive. They may be combined in a wide variety of ways (including the use of computers) to form numerous specific navigational systems. The various "elements of a fix" (i.e., co-ordinates of position) may be determined by different basic methods, of which the following are typical examples:

a. Polar co-ordinate methods

Examples: Radar-PPI (combining 1.4.2 a and 1.4.3 a), omnidirectional range plus distance-measuring equipment (combining 1.4.2 b and 1.4.3 a).

b. Point identification (combining 1.4.1 and 1.4.3 c). Example: Zone (Z) marker plus altimeter.

c. Hyperbolic-elliptical co-ordinate methods (combining 1.4.4 and 1.4.5).

d. Intersection of two identifiable surfaces.

Examples: Beam-type instrument approach system (ILS), Ground control of approach radar (GCA).

2. DEFINITIONS OF STANDARD NAVIGATION NOMENCLATURE

The following table presents a summary of suggested standard nomenclature (defined elsewhere) as the relation of navigation coordinates to the plan, enroute indication, and result.

Co-ordinate	Navigational Plan	Instrument Indication	Result
Horizontal Component of Direction	Course	Heading	Course made good
Vertical Component of Direction	Slope Angle	Pitch Attitude	Slope angle made good
Horizontal Component of Path	Course Line	Position	Track (or Horizontal Track)
Vertical Component of Path	Altitudes	Position	Vertical Track
Path (3 Dimensions)	Path (or Flight Path)	Position	Track (or Flight Track)
Horizontal Component of Speed	Estimated Ground Speed	Speed (any method)	Ground Speed
Vertical Component of Speed	Estimated Rate of Climb	Rate of Climb (or Dive)	Vertical Speed
Time Schedule	Estimated Time at Each Point	Time (at any point)	Elapsed Time (Between any Points)

Three terms used in navigation, "bearing," "heading," and "course," are defined on later pages. When these words are used without modifier, the reference direction from which measurements are made is indefinite. Thus, the definitions of these three terms involve measurement of angles from references which are unstated.

A number of modifiers are used with these words to define the reference. For example, we may have "true bearing," "true heading," and "true course." Each of these is defined hereafter. The modifier "true" gives the reference direction as true north. The angles for bearing, heading, and course in their "true" sense can be measured by any desired means.

The modifiers and their reference directions are as follows:

True. The reference direction is true north. Azimuth is the same as true bearing. It is suggested, however, that the word "azimuth" be reserved for celestial angles used in navigation and for other purposes, such as surveying. Thus, for purely terrestrial navigational use, the term "true bearing" is preferred.

Magnetic. The reference direction is magnetic north.

Compass. The reference direction is the north mark on the compass card of a magnetic compass (or repeater). Deviation has an effect on the angle of the north point of the card with reference to magnetic north. Further, the calibration of the compass card may not be correct. However, the reference as stated is the north mark on the compass card and the reference reading is obtained at the lubber line on the compass case as indicated on the marked compass scale. "Corrected compass" means "magnetic," which is given above.

Relative. The reference direction is the vehicle's heading, which is the forward direction along the vehicle's longitudinal center line. "Relative heading" and "relative course" should not be used. The proper terms are, respectively, "heading" and "drift angle."

Grid. The reference direction is the top of a grid which, for polar navigation, is a grid of rectangular coordinates superimposed over the polar regions. One line on this grid coincides with the Greenwich meridian. North on this grid is the direction upward on the chart, usually the direction of the North Pole from Greenwich.

3. DEFINITIONS OF TERMS

ADF. Automatic direction finder, specifically as used in aircraft.

Aided Tracking. A system of tracking a signal in bearing, elevation, or range, or any combination of these variables, in which a constant rate of motion of the tracking mechanism is maintained such that the motion of the target can be followed. The operator adjusts the rate by controlling an error parameter.

Airborne Radar. A radar set providing information in an aircraft about the relative position of fixed identification points or other aircraft.

Air Position Indicator (API). A dead reckoning computer which integrates headings and speeds to give a continuous indication of position with respect to the air mass in which the vehicle is moving.

Altitude. Vertical distance above sea level.

Ambiguity. The condition obtaining when navigational co-ordinates define more than one point, direction, line of position, or surface of position.

Angle of Elevation. The angle measured at the observer between the horizontal plane and the line to the object.

A-N Radio Range. A navigational aid that establishes four radial equisignal zones, a deviation from the zones being indicated by the audible Morse Code letters A or N and the on-course indication being a continuous tone.

Approach Navigation. Navigation under such conditions that the approach to a dock or runway becomes of major importance.

Approach Path. That portion of the flight path in the immediate vicinity of the landing area where the flight path is ordinarily defined in three dimensions.

A-Scope. A cathode-ray indicator with a horizontal or vertical sweep, giving signal amplitude and distance. Signals appear as vertical or horizontal deflections on the time scale.

Aural Radio Range. A radio range facility whose courses are normally followed by interpretation of the transmitted aural signal.

Automatic Tracking. Tracking in which a servomechanism follows the signal automatically.

Auto-Radar Plot. See *Chart Comparison Unit.*

Azimuth. Same as true bearing and usually associated with celestial navigation.

Azimuth-Stabilized PPI. A PPI on which indicated north is fixed with respect to the heading of the vehicle, usually at the top of the screen.

Baseline. The geodesic between two stations that operate in conjunction for the determination of navigational co-ordinates.

Bearing. The angle in the horizontal plane between a reference direction and the line jointing the observer with an object, usually measured clockwise from the reference direction.

B-Scope. A cathode-ray indicator in which a signal appears as a spot with bearing as the horizontal co-ordinate and distance as the vertical co-ordinate.

Boundary Marker. A marker facility, in an instrument landing system, which is installed near the approach end of the landing runway and approximately on the localizer course line.

Carrier Controlled Approach (CCA). A radar system for aiding landing on aircraft carriers.

Center Line. The locus of the points equidistant from two reference points or lines.

Chain. A network of similar stations operating as a group for the determination of position.

Challenge. See *Interrogation.*

Challenger. See *Interrogator.*

Chart Comparison Unit. A device for positioning a radar map on a navigational chart.

Coder. See *Pulse Coder.*

Coding Delay. An arbitrary time delay in the transmission of pulse signals, usually inserted at a transmitting station.

Compass Bearing. The angle in the horizontal plane between the direction of magnetic north on the compass card and the line joining the observer and the object, usually measured clockwise.

Compass Course. The directng in the horizontal plane of intended travel with respect to the direction of magnetic north on the compass card, usually measured clockwise.

Compass Heading. The angle in the horizontal plane between the direction of magnetic north on the compass card and the line along which the vehicle is pointing, usually measured clockwise.

Corner Reflector. A reflecting object consisting of two or three mutually intersecting conducting surfaces. (Definition of Antennas Committee.)

Note: Corner reflectors may be dihedral or trihedral. Trihedral reflectors may be used as radar targets.

Corrected Compass Course. Same as *Magnetic Course.*

Corrected Compass Heading. Same as *Magnetic Heading.*

Count Down. The ratio of the number of interrogation pulses not answered to the total number of interrogation pulses in a transponder beacon.

Course. The direction of intended travel projected in the horizontal plane expressed as an angle from a reference direction, usually measured clockwise.

Course Line. The horizontal component of path of proposed travel for the vehicle. It comprises course and the element of distance.

Course Error. The angular difference between the planned course and the course made good by an aircraft.

Course (Line) Computer. The equipment which provides the means by which any arbitrary course line may be set up and flown, such as that used in connection with ODR and DME equipment.

Course (Line) Deviation Indicator. A cross-pointer instrument indicating deviation from a course line.

Course Made Good. The resultant direction the vehicle bears from a point of departure or waypoint, usually measured clockwise from true north.

Course (Line) Selector. An instrument providing means to select the course to be flown.

Crossing Angle. The angle at which two lines of position, or courses lines, intersect.

C-Scope. A cathode-ray indicator in which a signal appears as a spot with bearing as the horizontal co-ordinate and elevation angle as the vertical co-ordinate.

Dead Reckoning. Determination of position by advancing a known position through the addition of one or more vectors representing known courses, times, and speeds.

Decoder. A circuit which responds to a particular coded signal and rejects others.

Desired Track. See *Path.*

Destination. The point of intended arrival.

Differential Gain Control. (Also called *Gain Time Control* or *Sensitivity Time Control.*) A device for altering the gain of a radio receiver at the times when various signals are expected, in order to reduce the discrepancy in amplitude between the signals at the output of the receiver.

Directional Homing. The procedure of following a path such that the objective is maintained at a constant relative bearing.

Direction Finder (DF). A radio aid to navigation that determines the direction of arrival of a radio signal by measuring the orientation of the wave front or of the magnetic or electric vector of a radio wave.

Distance Measuring Equipment (DME). A radio aid to navigation that determines the distance from a transponder beacon by measuring the total time of transmission to and from the beacon.

Distance Mark. A mark on a cathode-ray indicator which indicates the distance from a radar set to a target.

Drift Angle. The angular difference between the heading and the course made good.

Electrical Distance. The distance traveled by radio waves in a unit of time. A convenient unit of electrical distance is the light microsecond, or about 983 feet (300 meters). In this unit the electrical distance is numerically equal to the transmission time in microseconds.

Elements of a Fix. The specific values of the navigational co-ordinates necessary to define a position.

Equiphase Zone. The region in space within which the difference in phase of two radio signals is indistinguishable.

Equisignal Zone. The region in space within which the difference in amplitude of two radio signals (usually emitted by a single station) is indistinguishable.

Fan Marker. A vhf radio beacon, in an instrument landing system, having a fan-shaped radiation pattern and located along an airway radio range leg to provide a position fix.

Fix. Position determined without reference to any former position.

Flag Alarm. A semaphore-type flag provided in a navigational indicator to warn the pilot that a system failure has occurred.

Flight-Path. The path in space planned for an aircraft flight.

Flight-Path Computer. A computer including all of the functions of a course line computer and in addition providing means for controlling the altitude of the aircraft in accordance with any desired plan of flight.

Flight-Path Deviation-Indicator. An instrument providing an indication of deviation from flight path.

Flight Track. The track in space actually traced by an aircraft.

Gain Time Control. See *Differential Gain Control.*

Geodesic. The shortest line on the surface of the earth between two points.

Glide Path. See *Glide Slope.*

Glide Slope. (Previously called *Glide Path.*) An inclined surface of signal extending upward at an angle to the horizontal from the point of desired ground contact.

Glide Slope Facility. The means of providing a glide slope.

Grid Bearing. The angle, usually measured clockwise, between grid north and the initial direction of the arc of a great circle through an observer and a point.

Grid Course. A direction of intended travel projected in the horizontal plane expressed as an angle from grid north, usually measured clockwise.

Grid Heading. A direction in the horizontal plane expressed as an angle from Grid North to a line along which a vehicle is pointed, usually measured clockwise.

Grid North. An arbitrary reference direction used in connection with the Grid System of navigation. The reference direction is the top of a grid which, for polar navigation, is a grid of rectangular co-ordinates superimposed over the polar regions. One line on this grid coincides with the Greenwich Meridan. North of this grid is the direction upward on the chart, usually the direction of the North Pole from Greenwich.

Ground Controlled Approach (GCA). A ground radar system providing information by which aircraft approaches may be directed via radio communications.

Ground Distance. The horizontal component of distance from one object to another.

Ground-Position Indicator (GPI). A dead-reckoning tracer, similar to an air position indicator with provision for taking account of drift.

Ground Surveillance Radar. A radar set operated at a fixed point for observation and control of the position of aircraft or other vehicles in the vicinity.

Heading. The angle in the horizontal plane between a reference direction and the line along which the vehicle is pointing, usually measured clockwise.

Homing. Following a course directed toward a point by maintaining constant some navigational coordinates (other than altitude).

Instrument Landing System (ILS). A radio system which provides in the aircraft the directional, longitudinal, and vertical guidance necessary for landing.

Interrogation. Transmission of a radio-frequency pulse or combination of pulses intended to trigger a transponder or group of transponders.

Note: Sometimes called *Challenge.*

Interrogator. See *Interrogator-Responsor.*

Interrogator-Responsor (IR). A radio transmitter and receiver combined to interrogate a transponder and display the resulting replies.

Note: Sometimes called *Challenger.*

Lattice. A grid of identifiable lines of position laid down in fixed positions with respect to the transmitters that establish it.

Leader Cable. A navigational aid consisting of a cable around which a magnetic field is established, marking the path to be followed.

Line of Position. A line along which two navigational co-ordinates are constant; known values of these co-ordinates establish the navigator's position as somewhere along this line.

Localizer. A radio facility which provides signals for lateral guidance of aircraft with respect to a runway center line.

Long-Range Navigation. Navigation at distances such that knowledge of position and objective alone are sufficient to permit determination of the proper course.

Magnetic Bearing. The angle in the horizontal plane between the direction of magnetic north and a line joining the observer and the object, usually measured clockwise.

Magnetic Course. The direction in the horizontal plane expressed as the angle of intended travel with respect to the direction of magnetic north.

Magnetic Deviation. Angular difference between compass reading and magnetic heading.

Magnetic Heading. The angle in the horizontal plane between the direction of magnetic north and the line along which the vehicle is pointing, usually measured clockwise.

Marker. In an instrument landing system, a radio facility providing a signal to designate a small area above it.

Master Station. The station of a synchronized group of stations that governs the emissions of the group.

Matching. In navigation, the process of bringing two quantities into suitable positions for measurement of their relative value.

Middle Marker. A marker facility which is installed approximately 3,500 feet from the approach end of the runway on the localizer course-line.

Minimum Distance. The shortest distance at which a navigational system will function.

Moder. See *Pulse Coder.*

Moving Target Indicator (MTI). A device which limits the display of radar information primarily to moving targets.

Navigation. The process of finding the position of a vehicle and directing it to reach a desired destination.

Navigational Co-ordinate. A quantity whose measurement serves to define a surface of position (or a line of position if one surface is already known) containing the vehicle.

North-Stabilized PPI. See *Azimuth-Stabilized PPI.*

Offset-Course Computer. An automatic computer which translates reference navigational co-ordinates into those required for a predetermined course.

Omnibearing. The bearing, usually magnetic, of an omnidirectional range station from a vehicle.

Omnibearing Indicator. An instrument providing automatic and continuous indication of the omnibearing.

Omnibearing Converter. An electromechanical device which combines the omnibearing signal with vehicle heading information to furnish electrical signals for the operation of the pointer of a radio magnetic indicator.

Omnibearing-Distance Facility. A radio facility, having an omnidirectional range in combination with distance-measuring equipment.

Omnibearing-Distance Navigation (R-THETA). Radio navigation utilizing a polar-coordinate system as a reference, making use of omnibearing-distance facilities.

Omnibearing Selector. An instrument capable of being set manually to any desired omnibearing, or reciprocal thereof, which controls a course line deviation indicator.

Omnidistance. The distance between the vehicle and an omnibearing-distance facility.

Omnirange (or Omnidirectional Range). A facility providing navigators with direct indication of the bearing of the omnirange facility from the vehicle.

Outer Marker. In an instrument landing system, a marker facility installed at approximately 5 miles from the approach end of the runway on the localizer course line.

Path. The proposed route of a vehicle in space. In surface navigation, the proposed route on the surface.

Phase Localizer. A localizer in which two signals are compared in phase to obtain lateral guidance.

Pilotage. Navigation without explicit determination of position, by directing a vehicle to its destination by observation of landmarks.

Pitch Attitude. The angle between the longitudinal axis of the vehicle and the horizontal.

Polar Grid. See *Grid.*

Position. The location of a vehicle as determined by specific values of three or more navigational co-ordinates.

Plan Position Indicator (PPI). A cathode-ray indicator in which a signal appears on a radial line. Distance is indicated radially and bearing as an angle.

Pulse Coder. A circuit which sets up a plurality of pulses disposed in an identifiable pattern.

Pulse Interval. The interval between the leading edges of successive pulses in a sequence characterized by uniform spacing; recurrence interval.

Pulse Spacing. The interval between the leading edges of successive pulses.

Pulse Train. A group of pulses of similar characteristics.

Quadrantal Error. Angular error of a measured bearing caused by disturbances due to the characteristics of the vehicle or station.

Racon. An abbreviation of "radar beacon"; a responder beacon for use with a radar set.

Note: See *Transponder.*

Radar. A device that measures the distance and direction of objects by reflection of radio waves.

Radio-Autopilot Coupler. Equipment providing means by which an electrical navigational signal will operate the automatic pilot to allow automatic flight.

Radio Beacon. A radio facility, usually a non-directional radio transmitter, providing signals for radio direction-finding observations.

Radio Direction Finding. Radiolocation in which only the direction of a station is determined by means of its emissions.

Radiolocation. Determination of a position or of a direction by means of the constant velocity or rectilinear propagation properties of the Hertzian waves.

Radio Magnetic Indicator (RMI). An instrument which presents a combined display of vehicle heading, relative and magnetic bearings, and omnibearings of the radio station being utilized for navigation purposes.

Radio Navigation. Navigation by means of radio signals.

Radio-Range. A radio facility providing radial equisignal zones.

Range. See *Operating Distance*, and *Radio Range.*

Range Mark. See *Distance Mark.*

Recurrence Rate. See *Repetition Rate.*

Reference Direction. The direction used as a reference for angular measurements.

Reference Line. A line passing through a reference point and an observer.

Relative Bearing. The angle, usually measured clockwise, between the heading of a vehicle and the initial direction of the arc of a great circle through an observer and a point.

Relative Course. See *Drift Angle.*

Relative Heading. The heading itself. "Relative" is superfluous.

Repetition Rate. The rate at which recurrent signals are transmitted.

Reply. A radio-frequency pulse or a combination of pulses transmitted by a transponder as a result of an interrogation.

Resolution. The degree to which nearly equal values of a quantity can be discriminated.

Responder Beacon. See *Transponder.*

Responder. A receiver in a transponder whose function is to receive the signals transmitted from a beacon.

Scanning. A periodic motion given to the major lobe of an antenna.

Note: Definition of Antennas Committee.

Searchlighting. Projecting the radar beam continuously at an object instead of scanning.

Sensing. The relative direction of motion of a deviation indicator needle resulting from departure of a vehicle from the desired flight path.

Service Area. The area within which a navigation aid is of use.

Short Distance Navigation. Navigation which is predicated upon aids spaced less than 200 miles apart.

Slant Distance. The distance from an object to another object not at its own elevation. Used in contrast to ground range.

Slave Station. A station of a synchronized group whose emissions are controlled by a master station.

Slope. The projection of a flight path in the vertical plane.

Slope Angle. The direction of a flight path expressed as an angle projected on the vertical plane.

Slope Deviation. The difference between the projecion in the vertical plane of the actual path of movement of a vehicle and the planned slope for the vehicle expressed in terms of either angular or linear measurement.

Sonar. The general name for sonic and/or ultrasonic underwater echo-ranging and echo-sounding systems.

Stable Element. A gyroscopic instrument which maintains a true vertical and shows angles of deviation of the ship's deck or aircraft's axis from the true horizontal.

Stabilization. A system for maintaining a device in a desired direction in space despite the motions of the ship or aircraft.

Star Chain. A group of navigational radio transmitting stations in Y form with the master facility in the center and three (or more) slave facilities around the circumference of a rough circle.

Surface of Position. A surface that is defined by a constant value of some navigational coordinate.

Terrain Clearance Indicator (Sometimes called *Absolute Altimeter*). A device measuring the distance from an aircraft to the surface of the sea or ground.

Tilt. The angle which an antenna axis forms with the horizontal.

Time Gain Control. See *Differential Gain Control.*

To-From Indicator. An instrument to show whether the numerical reading of an Omnibearing Selector for an "on course" indication of the omnibearing indicator represents the bearing toward or away from a vhf omnirange.

Tone Localizer. A localizer which transmits two modulation frequencies for amplitude comparison.

Track. The projection on the earth's surface of the actual path followed by a vehicle.

Tracking. The process of keeping radar beams or the cross hairs of an optical system set on a target.

Track Homing. The process of following a line of position known to pass through an objective.

Transponder. A transmitter-receiver facility whose function is automatically to transmit signals when the proper interrogation is received.

Transponder-Beacon. See *Transponder.*

Triplet. Three radio facilities operated as a group for the determination of positions.

True Bearing. The angle measured clockwise between true north and the initial direction of an arc of a great circle through an observer and a point.

True Course. A direction of intended travel projected in the horizontal plane expressed as an angle measured clockwise from true north.

True Heading. A direction in the horizontal plane expressed as an angle measured clockwise from true north to a line along which a vehicle is pointed.

True Homing. The process of following a course such that the true bearing of a vehicle as seen from an objective is held constant.

Variation. The angular difference between true and magnetic bearings.

VHF Omnirange. A vhf radio navigation station providing direct indication of the magnetic bearing (omnibearing) of that station from a vehicle.

Visual-Aural Radio Range. A radio range which provides aural sector identification and visual course indication. The visual course of this range defines the primary navigation course and is flown by visual means. The aural sector identification results in an aural course at 90° to the visual course.

Visual Radio Range. Any range facility the course of which is flown by visual instrumentation not associated with aural reception.

Way-Point. A course line point the co-ordinates of which are defined in relation to established radio navigation aids.

Zone (Z) Marker. A vhf radio facility located at airways radio range stations to indicate position above such ranges.

4. SUPPLEMENTARY DEFINITIONS

The following supplementary list offers brief descriptions of a number of navigation aids known chiefly by recently adopted names that are not yet universally recognized. The definitions are recorded for information but not for standardization.

Babs. Blind Approach Beacon System. A pulse-type ground-based navigation beacon used for runway approach at airfields.

Benito. A continuous wave navigation system measuring range and azimuth from one or more ground stations. The range is determined by a phase comparison method.

Consol. British code word for *Sonne.*

Decca. A continuous wave differential distance hyperbolic radio aid to navigation in which the receiver measures and integrates the relative phase difference between the signals received from two or more synchronized ground stations.

Electra. A German continuous wave navigation system using radio beacons providing multilobe equisignal patterns.

Eureka. The ground radar beacon of the Rebecca-Eureka navigation system.

GCI. Ground Controlled Interception. A radar system by means of which a controller at the radar may direct an aircraft to make an interception of another aircraft.

Gee. A medium range hyperbolic radio aid to navigation whose position lines are determined by measuring the difference in time of arrival of synchronized pulsed signals.

Gee H. A combination of the *Gee* and *H* systems of navigation.

H. A radar air navigation system using an airborne interrogator to measure distance from two ground responder beacons.

Note: See *Shoran.*

Lanac. Laminar Navigation Anti-Collision. This system consists of air and ground radar and beacon equipments with height coding of the aircraft transmitter pulses.

Loran. A long range pulsed hyperbolic radio aid to navigation whose position lines are determined by the measurement of the difference in the time of arrival of synchronized pulses.

MEW. Microwave Early Warning. A high power long-range, 10-centimeter early warning radar with a number of indicators, giving high resolution and large handling capacity.

Navar. A co-ordinated series of radar air navigation and traffic control aids including both air and ground equipments.

Navaglobe. A long range continuous wave lf navigation system of the amplitude comparison type.

Oboe. A radar navigation system consisting of two ground stations measuring distance to an airborne responder beacon and relaying information to the aircraft.

POPI. Post Office Position Indicator. A continuous wave lf navigation system of the phase comparison type.

Rebecca. The airborne interrogator responder of Rebecca-Eureka, a radar responder beacon system.

SBA. Standard Beam Approach. A continuous wave approach system using a localizer and markers.

Shoran. Short Range Navigation. A precision position fixing system using a pulse transmitter and receiver and two transponder beacons at fixed points.

Sonne. A radio aid to navigation that provides a number of equisignal zones which rotate in a time sequence so that a navigator may determine his true bearing from the transmitter by observation of the instant at which he hears the equisignal.

Teleran. Television and radar navigation system. Television image of ground PPI and map and weather data are presented in the aircraft.

Tricon. A radar system in which the receiver records the coincidence of received pulses from a group of three ground stations pulsed in variable time sequence.

VOR. Vhf phase comparison omnidirectional radio range system developed by the Civil Aeronautics Administration.

Standards on
RAILROAD AND VEHICULAR COMMUNICATIONS: METHODS OF TESTING, 1949*

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1. INTRODUCTION

These Standards cover definitions of terms and methods of testing communication receivers designed to receive frequency-modulated (FM) waves in the frequency range from 25 to 225 megacycles. In view of the several bands presently allocated for communications services in this range and the possibility that these bands may be modified from time to time, carrier-frequency limits for the individual sub-bands are not specified.

2. DEFINITIONS OF TERMS

2.1. Test Frequencies

In general, the test frequencies for the various designated sub-bands consist of one frequency near the lower edge of the sub-band, one near the upper edge, and one near the center. When measurements are to be made at a single frequency only, that frequency should be chosen near the center of the sub-band unless the equipment is designated for a specific frequency, in which case its specified operating frequency should be used.

2.2. Test Input Signals

Input-signal intensities may be expressed in either of two ways:

(a) In terms of available power, in which case the input is expressed in decibels below 1 watt.

(b) In terms of input voltage, in which case the input is expressed in microvolts with the input impedance specified.

2.3. Available Power

The available power is the power delivered by a generator to a matched load. It is equal to $E^2/4R$ where E is the open-circuit voltage of the generator and R is the internal resistance of the generator plus the dummy-antenna resistance. It is expressed in decibels below 1 watt. A signal generator may be calibrated in terms of the available signal power and used on that basis though not matched exactly by the load impedance. If a signal generator is to be used with various values of dummy-antenna resistance, it should be calibrated in terms of the open-circuit voltage, and the available power should be calculated from the above formula. In this report, when values of power input are spoken of, it should be understood that the available power is meant.

2.4. Standard Input Values

Five values of standard input are specified for the purpose of certain tests. The values of standard input voltage are equivalent to the corresponding values of standard input power for receivers designed for input impedances of 50 and 70 ohms.

2.4.1. Standard Mean-Signal Input

The standard mean-signal input is either 123 decibels below 1 watt or the equivalent signal measured in microvolts (10 microvolts at 50 ohms, 12 microvolts at 70 ohms).

TABLE I

(a) Standard Input Powers, decibels below 1 watt	(b) Standard Input Voltages in Microvolts (Open Circuit)	
	50 ohms input impedance	70 ohms input impedance
143	1	1.2
123	10	12
83	1,000	1,200
43	100,000	120,000
23	1 volt	1.2 volts

2.5. Standard Test Modulation

Standard test modulation in tests on frequency-modulation communication receivers refers to a signal that is frequency-modulated at 1,000 cycles with a deviation of 70 per cent of maximum rated system deviation. In this report, maximum rated system deviation is taken as 15 kilocycles, so that the deviation due to standard test modulation is 10.5 kilocycles.

2.6. Usable-Sensitivity Test Input

The usable-sensitivity test input is the least input signal of a specified carrier frequency having standard test modulation which, when applied to the receiver through the standard dummy antenna, results in standard test output with the ratio of the root sum square, signal+noise+distortion, to the root sum square, noise+distortion, equal to at least 12 decibels. This test discloses the influence of the selective circuit of the receiver and of internal receiver noise on the usable sensitivity of the receiver.

2.7. Deviation-Sensitivity Test Input

The deviation-sensitivity test input is the minimum deviation at 1,000 cycles of a carrier wave of standard mean-signal input value (Section 2.4.1) required to give standard test output (Section 2.9) when all controls are adjusted for greatest sensitivity. The deviation sensitivity is expressed in kilocycles or as a percentage of maximum rated system deviation.

2.8. Quieting-Signal-Sensitivity Test Input

The quieting-signal-sensitivity test input is the least unmodulated signal input which, when applied to the receiver through the standard dummy antenna, reduces the receiver noise output by a factor of 20 decibels. It is expressed in dedibels below 1 watt or in microvolts with input impedance specified.

Note. This is not the same characteristic as that described in Section 2.09 of *Standards on Radio Receivers: Methods of Testing Frequency-Modulation Broadcast Receivers, 1947*, which has the same title. It is believed that the above test is more applicable to communication receivers and more easily made with available field test equipment.

2.9. Standard Test Output

Standard Test output is equal to one-half of the rated power output of the receiver.

2.10. Rated Power Output

The rated power output is as specified by the manufacturer.

2.11. Standard Dummy Antenna

The standard dummy antenna comprises a resistor connected in series with the high terminal of the signal generator, of such value that the total impedance presented to the receiver is equal to the rated receiver input impedance. It is intended to simulate the mean value of the impedance of a typical transmission line connected to an antenna. (See Fig. 1.)

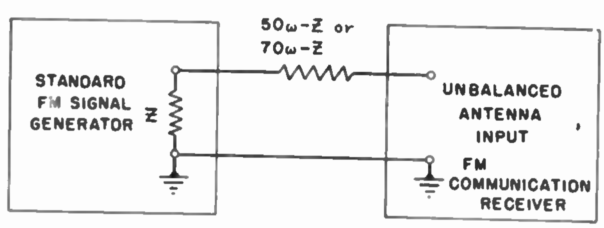

Fig. 1—Standard communication dummy antenna and method of connection.

2.12. Standard Pre-Emphasis Characteristic

The standard pre-emphasis characteristic has a rising response, with modulated frequency, of 6 decibels per octave. This is the characteristic obtained by linear phase modulation.

2.13. Standard De-Emphasis Characteristic

The standard de-emphasis characteristic has a falling response with modulating frequency, the inverse of the standard pre-emphasis characteristic or 6 decibels per octave. This characteristic may be approached over the usable audio-frequency range by taking the voltage across the capacitor when a capacitor and a resistor are connected in series and fed from a constant-voltage source. Deemphasis is usually incorporated in the audio circuits of the receiver.

2.14. Frequency Deviation

Frequency deviation, in frequency modulation, is the peak difference between the instantaneous frequency of the modulated wave and the carrier frequency.

2.15. Maximum System Deviation

Maximum system deviation is the greatest deviation specified in the operation of the system. The maximum system deviation is usually 15 kilocycles.

2.16. Selectivity

Selectivity is the extent to which a receiver is capable of differentiating between the desired signal and disturbances of other frequencies.

2.17. Spurious Response

Spurious response is receiver response which exists because of other than all desired normal frequency translations in the receiver.

2.18. Squelch

Squelch is a means whereby a receiver is prevented from producing audio frequency output in the absence of a signal having predetermined characteristics. A squelch circuit may be operated by signal energy in the receiver pass band, by noise quieting, or by a combination of the two (ratio squelch). It may also be operated by a signal having special modulation characteristics (selective squelch).

3. REQUIREMENTS AND CHARACTERISTICS OF TESTING APPARATUS

3.1. Signal Generator

A frequency-modulated signal generator is required for testing frequency-modulation radio receivers.

The signal generator should cover at least the carrier-frequency ranges of the various sub-bands being considered. It preferably should also cover the intermediate-frequency ranges and frequency ranges required for spurious-response tests.

Single-ended output terminals should be provided. These terminals should be on the end of a flexible cable.

The output meter of the signal generator should indicate the open-circuit voltage at the terminals, and the internal impedance should be stated.

The generator should be capable of being frequency modulated at rates from 300 to at least 3,000 cycles per second, and at deviations from zero to at least rated system deviation and preferable to twice that value. It should be provided with a deviation indicator reading from zero up to the maximum deviation.

The modulation circuit of the generator should be provided with a pre-emphasis network providing 6 decibels per octave pre-emphasis over the audio-frequency range of 300 to 3,000 cycles. A switch should be provided for cutting this pre-emphasis network in or out of the generator circuit at will.

The generator syould provide a frequency-modulated signal at 400 and at 1,000 cycles up to maximum rated system deviation with less than 2 per cent root-sum-square distortion. Amplitude modulation resulting from the frequency modulation should be kept to a minimum.

The frequency and amplitude modulation of the output voltage due to power-supply ripple should be negligible, in comparison with the effects under observation.

3.2. Audio-Output and Distortion-Measuring Devices

Apparatus for the measurement of distortion, required for Sections 4.1 and 4.2, should consist of distortion meter of the type which integrates the total noise and distortion, while balancing or filtering out the fundamental frequency of the audio signal.

3.3. Standard-Signal Generator for Two-Signal Test

For certain tests of radio communications receivers, two radio-frequency input signals are required simultaneously, and consequently two standard-signal generators are employed. The recommended method is to use a dummy antenna on each signal generator of twice the standard dummy antenna resistance. The terminals of the two-dummy antennas are then connected in parallel and to the input terminals of the receiver. By this arrangement, the impedance connected across the receiver input terminals is the normal value, and the open-circuit signal voltages are half the values indicated by each generator.

4. TEST PROCEDURES

4.1. Usable-Sensitivity

A calibrated signal generator is connected to the input of the receiver under test through a standard dummy antenna (Section 2.11). The signal generator is adjusted to the receiver frequency, and standard test modulation (Section 2.5) is applied. A distortion meter of the type which integrates the total noise and distortion while balancing or filtering out the 1,000-cycle fundamental frequency of the signal is connected to the output of the audio-amplifier.

The signal generator output is adjusted to the minimum signal which will provide standard test output from the receiver. The signal+noise+distortion to noise +distortion ratio is then measured, and, unless found to be 12 decibels or more, the signal is increased, holding the receiver audio output constant at standard test output by means of the receiver audio gain control, until this value is obtained. The signal required to produce this result is the usable sensitivity. Measurements should be made at each of the test frequencies (Section 2.1).

4.2. Two-Signal Selectivity

Two signal generators of similar characteristics shall be equally coupled to the input of the receiver in such a fashion that they do not react upon one another, and in combination present an impedance match to the input circuit. Both signal generators shall be modulated equally at 70 per cent of maximum system deviation (Section 2.15), with signal generator No. 1 at 1,000 cycles and signal generator No. 2 at 400 cycles.

With the output of No. 2 at zero, No. 1 shall be set to the receiver frequency and its output adjusted until the receiver input equals that impressed on its terminals when one of the standard input values (Section 2.4) is applied through the standard dummy antenna (Section 2.11).

Signal generator No. 2 shall then be set at a frequency differing from No. 1 and its output increased until the signal+noise+distortion to noise+distortion ratio decreases to 6 decibels. During this measurement the modulation products from signal generator No. 2 are to be considered as noise.

The selectivity of the receiver at the frequency between the two signal generators is then specified by the ratio of their radio-frequency output amplitudes.

If the total audio output of the receiver drops by 6 decibels at a smaller radio-frequency ratio than that referred to above, the selectivity is specified by the ratio of radio-frequency amplitudes when this occurs.

The selectivity characteristic of a receiver may be displayed by a plotted curve showing the variation of radio-frequency amplitude ratio, plotted in decibels as the ordinate, with frequency differences as the abscissa. Both scales should be linear.

These same data should be obtained with desired signal values corresponding to each of the standard input values (Section 2.4) within the limitations of the measuring equipment.

4.3. Spurious Response

The desired-signal input values are the same as used in the selectivity measurements, and the interfering input voltage is referred to the desired-signal input. Proceed as in the measurement of two-signal selectivity, except that the interfering signal is tuned to produce peak response for the spurious mode under test. The spurious response of the receiver at the given interference and test frequencies is specified by the ratio of the radio-frequency output amplitudes of the two-signal generators.

TESTS FOR EFFECTS OF MISTUNING AND FOR DOWNWARD MODULATION, 1949*

Supplement to Standards on

RADIO RECEIVERS: METHODS OF TESTING FREQUENCY-MODULATION BROADCAST RECEIVERS, 1947

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1. MISTUNING

1.1. The degree of mistuning is represented by the total signal output distortion resulting when the receiver is adjusted to a frequency other than the desired signal frequency. The measurement is made by setting the signal generator to standard input voltages successively, modulating the signal generator to 75 kc deviation at standard test output. The signal generator is then adjusted off tune by successive increments, the volume control is adjusted for standard test output, and the total distortion in per cent (or db) is measured. For each value of input signal a curve is plotted, having as abscissa the frequency difference of detuning, and as ordinate the distortion expressed in per cent, or db. Distortion components will comprise all frequencies present except the fundamental frequency of the modulating tone. In these tests the signal generator is adjusted off tune on each side of the signal frequency.

1.2. The standard measurement will comprise setting the signal generator on each side of the signal frequency and noting the amount of mistuning that will produce 10 per cent distortion, expressing the degree of mistuning as the average of the measured plus and minus frequency excursions. The signal input for this test shall be the standard mean signal input (1,100 microvolts).

This mistuning test should be correlated with the frequency drift test, Sec. 4.05.20.

2. DOWNWARD MODULATION

2.1. This test will define the ability of the receiver to withstand the effects of downward amplitude modulation. In this test it is assumed that the principal forms of distortion are caused by the downward component of modulation.

2.2. The test is made at the standard mean-carrier frequency (98 megacycles). Frequency modulation is impressed at a 400-cycle modulation rate at 30 per cent of maximum rated system deviation and the volume control is adjusted for standard output. The input signal is then simultaneously amplitude modulated at a 100-cycle rate. By means of a band cut-off filter, the 100-cycle modulation is eliminated in the receiver output. The amplitude modulation is then increased until the total distortion reaches 10 per cent. The percentage modulation at this point is the downward modulation capability of the receiver. The test is made at all values of standard input signal voltages.

Standards on PIEZOELECTRIC CRYSTALS, 1949*

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INTRODUCTION

In 1945 a report entitled "Standards on Piezoelectric Crystals: Recommended Terminology" was prepared by the Committee on Piezoelectric Crystals and issued by the Institute. The present report involves not only much new material, but also a revision of certain portions of the earlier report.

Part I of the 1945 report (corresponding to Part 1 of this Standard) remains unchanged, with the following exceptions: The use of left-handed axes for left crystals is now abandoned (see Sections 1.9, and 1.11 to 1.14); and the 1945 rules for rotated plates are now supplanted by the rules in the present report.

In Part II (corresponding to Part 2 of this Standard), Sections 11 and 13 have been revised. Otherwise, Part II remains in effect. The introduction of new crystals, some belonging to classes for which no satisfactory conventions have existed, makes desirable a self-consistent set of conventions, sufficiently comprehensive to include all piezoelectric crystal classes. This necessitates certain changes in the definitions of the axes of quartz and in the algebraic signs of piezoelectric and elastic constants, as described in Sections 1.12 to 1.14. These changes result not only in agreement with the crystallographers, but, fortunately, in better conformity with shop practice as it has developed.

Part 3 of the new Standard presents basic equations, symbols, and units of piezoelectric theory. It has been prepared in view of the growing tendency to express

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elastic stress and strain as dependent upon the electric displacement, rather than upon the electric field, and also in view of the increasing use of mks units. The formulation in terms of displacement is especially useful in cases where there is a gap between crystal and electrodes, also in dealing with crystals whose constants show a marked dependence on temperature. In the latter case it is found that the "constants" have more nearly constant values when expressed in terms of displacement than in terms of field.

When there is no gap between crystal and electrodes it is often found more convenient to use Voigt's formulation, with the piezoelectric constants expressed in terms of electric field strength rather than displacement.

The equations are interconvertible. The Voigt piezoelectric constants d and e are related to the more recent constants g and h by the dielectric constant. For some purposes it is convenient to use expressions involving combinations of constants from both formulations. In doing so it is necessary, of course, to use the same system of units, esu or mks.

Authorship and Acknowledgment

This Standard is the result of several years of conferences involving all of the members of the Committee. Its form and its completeness, which in effect make it an introduction to the formal treatment of piezoelectricity, come from the fact that it is basically a collection of several papers prepared for the Committee by individual members and their collaborators serving as subcommittees. There would have been entire justification for the separate publication of these papers by their individual authors, but it has been the view of the Committee that the advantages of unification into a system such as the Standard is intended to establish could best be gained by bringing all papers into a common publication with agreement as to symbols, conventions, and equations.

The authorship of the several parts of the Standard is as follows: Part 1 is an elaboration of memoranda prepared for the Committee by W. L. Bond, entitled "Crystal Axis Nomenclature," October 25, 1946, and "Axes For Triclinic Crystals," June 16, 1947. Part 2 stems from a memorandum by W. L. Bond of August 19, 1946, on "Crystal Rotation Systems." Part 3 is an expansion of a memorandum by Paul L. Smith, "The Piezoelectric Relations, Symbols and Units," June 3, 1946, in which he had the collaboration of H. G. Baerwald of the Brush Development Company. The adopted formulation of the theory is based on equations first presented by the latter in a wartime report[1] to the NDRC.

[1] H. G. Baerwald, OSRD Report No. 287, Contract No. OEMsr-120, 1941.

1. Definitions of Axes for Piezoelectric Crystals

1.1 Crystals and Their Classification

For the guidance of those who are not familiar with crystallography, the following summary of those principles to which reference will be made later in this Standard may be helpful.

The term "crystal" is usually applied to solids that possess structural symmetry. In a crystal the atoms may be thought of as occurring in small groups, all groups exactly alike, similarly oriented, and regularly aligned in all three dimensions. If each group is regarded as bounded by a parallelepiped, the latter can be looked upon as one of the ultimate building blocks of the crystal; they are stacked together in all three dimensions without any spaces between. Such a building block is called a *unit cell.* Since the choice of the particular set of atoms to form a unit cell is arbitrary, it is evident that there is a wide range of choice in the shape and dimensions of the unit cell. In practice, that unit cell is selected which is most simply related to the actual crystal faces and X-ray reflections, and which has the symmetry of the crystal itself. Except in a few special cases the unit cell has the smallest possible size.

Depending on their degrees of symmetry, crystals are commonly classified in seven *systems:* triclinic (the least symmetrical), monoclinic, orthorhombic, tetragonal, hexagonal, trigonal, and isometric. Some authorities, however, treat trigonal crystals as a division of the hexagonal system.

The seven systems in turn are divided into point-groups (classes) according to their symmetry with respect to a point. There are thirty-two such classes, of which twelve are of too high a degree of symmetry to show piezoelectric properties. Thus twenty classes can be piezoelectric. Every system contains at least one piezoelectric class.

1.2. General

In determining a suitable nomenclature for the many new crystals that are finding their way into the piezoelectric field, it seems wise to make use of the nomenclature of the highly developed science of crystallography. This facilitates the use of data already recorded by crystallographers. Such data are, for example, atomic cell dimensions and angles, optical properties, and interfacial angles, all of which can be useful for establishing orientations of piezoelectric plates. In crystallography, the properties of a crystal are described in terms of the natural co-ordinate system provided by the crystal itself. The axes of this natural system are the edges of the unit cell. In a cubic crystal these axes are of equal length and are mutually perpendicular; in a triclinic crystal they are of unequal lengths and no two are mutually perpendicular.

The faces of any crystal are all parallel to planes whose intercepts on the natural axes a, b, c are small multiples of unit distances or else infinity, so that their reciprocals, when multiplied by a small common factor, are all small integers or zero. These are the indices of the planes. In this nomenclature we have, for example, faces (100), (010), (001), also called the a, b, c faces, respectively; in the orthorhombic, tetragonal, and isometric systems these faces are normal to the a, b, c axes. Other examples are faces (111) (the unit face), (121), etc. As referred to a set of rectangular axes X, Y, Z, these indices will in general be irrational except for cubic crystals.

On the other hand, the theoretical treatment of electricity and elasticity, which is fundamental in piezoelectric applications, has been developed with rectangular axes. One must therefore adopt some arbitrary relation between the a, b, c axes of crystallography and the rectangular X, Y, Z axes. Unless all workers in the field agree to use the same nomenclature, there will be great confusion. Data expressed in terms of one abc-XYZ relation look very different from the same data in terms of another abc-XYZ relation.

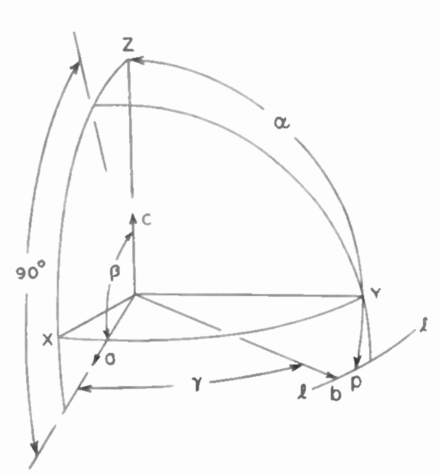

Fig. 1—Axes of triclinic crystals.
The line l–l is on a small circle with pole Z and angle α. For b to left of p, $\gamma<90°$. For b to right of p, $\gamma>90°$

1.3. The Triclinic System

If there are neither symmetry axes nor symmetry planes present in a crystal, it is triclinic. The lengths of the three axes are in general unequal, and the angles α, β, and γ are also unequal. α, β, γ are the angles between b and c, c and a, and a and b, respectively, as shown in Fig. 1. Fig. 2, a triclinic crystal, shows that

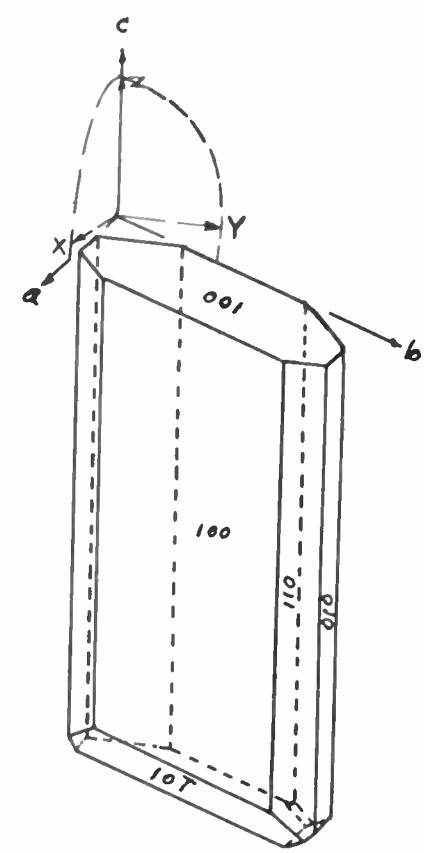

Fig. 2—Aminoethyl ethanolamine hydrogen d-tartrate, triclinic, class 1.
The Z axis is along c, Y normal to (010), X perpendicular to Y and Z.

the a axis has the direction of the intersection of the faces b and c (extend the faces to intersection if necessary), the b axis has the direction of the intersection of faces c and a, the c axis has the direction of the intersection of faces a and b. According to the best current usage the positive directions of a, b, and c are taken so as to make α greater than 90°, and β also greater than 90°. This convention determines uniquely the positive senses of all axes. (Many old works record the complementary angle, i.e., the angle between $-a$ and $+c$ as β.) The a, b, and c axes are in general to be chosen as the three shortest noncoplanar interatomic distances, with c the shortest unit distance, b the longest.

1.3.1. The X, Y, Z axes

The most logical relation is that which associates the X, Y, Z axes most closely with the a, b, c axes, respectively. There are six equally simple ways of completing the specification: X along a and Y in the ab plane; X along a and Z in the ac plane; Y along b and X in the ab plane; Y along b and Z in the bc plane; Z along c and Y in the bc plane; or Z along c and X in the ac plane. In physics and technology Z is commonly pictured as the vertical axis, while in crystallography c is usually so pictured. Hence we take Z along c and have only the choice of Y in the bc plane or X in the ac plane. The choice now accepted is to let Y be normal to the ac plane; this automatically places X in the ac plane.

The rules for the rectangular axes, together with their positive directions, are summarized as follows (see Figs. 1 and 2):

$+Z$ is parallel to $+c$, hence parallel to the (100) and (010) planes.

$+X$ is perpendicular to c in the ac plane, pointing in the general direction of $+a$. X is thus parallel to (010).

$+Y$ is normal to the ac plane (010), pointing in the general direction of $+b$, and forming a right-handed axial system with Z and X.

1.4. The Monoclinic System

If a crystal has but a single axis of twofold symmetry, or but a single plane of reflection symmetry, or both, it belongs to the monoclinic system. Either the twofold axis or the normal to the plane of symmetry (they are the same if both exist, and this direction is called the unique axis in any case) is taken as the b axis. Of the two remaining axes, modern crystallographers always make c the smaller. The angle β between $+a$ and $+c$ is always obtuse (a special convention has to be adopted when this angle is obtuse at some temperatures, acute at others). This convention determines uniquely the positive directions of all axes for classes 2 and $2/m$. In class m there are two alternatives. The choice between them is indicated in Section 1.16. Since the axes chosen by early workers may not give the smallest possible cell, a new cell with a smaller volume can be chosen.

Many physicists seem to prefer to make Z the unique axis of monclinic crystals. This is the convention adopted by Voigt, and continued by Cady in his book "Piezoelectricity."[2] Nevertheless, in most crystallographic literature b is taken as the unique axis in the monoclinic system. Henceforth "Z along c, Y along b" is to be the standard abc-XYZ relation for monoclinic crystals, as illustrated in Fig. 3. According to this convention, with the unique axis b taken as the Y axis, the stresses T and strains S are related through the matrix equations $S=sT$ and $T=cS$, where the compliance constant s takes the form[3]

$$s = \begin{vmatrix} s_{11} & s_{12} & s_{13} & 0 & s_{15} & 0 \\ s_{12} & s_{22} & s_{23} & 0 & s_{25} & 0 \\ s_{13} & s_{23} & s_{33} & 0 & s_{35} & 0 \\ 0 & 0 & 0 & s_{44} & 0 & s_{46} \\ s_{15} & s_{25} & s_{35} & 0 & s_{55} & 0 \\ 0 & 0 & 0 & s_{46} & 0 & s_{66} \end{vmatrix} \qquad (1)$$

The stiffness constant c takes an analogous form. Also, electric displacement and elastic stress are related through the matrix equation $D=dT$, where the piezoelectric strain-constant d takes the form, for class 2 (Y a twofold axis),

$$d = \begin{vmatrix} 0 & 0 & 0 & d_{14} & 0 & d_{16} \\ d_{21} & d_{22} & d_{23} & 0 & d_{25} & 0 \\ 0 & 0 & 0 & d_{34} & 0 & d_{36} \end{vmatrix} \qquad (2).$$

while for class m (a plane of reflection-symmetry perpendicular to Y)

$$d = \begin{vmatrix} d_{11} & d_{12} & d_{13} & 0 & d_{15} & 0 \\ 0 & 0 & 0 & d_{24} & 0 & d_{26} \\ d_{31} & d_{32} & d_{33} & 0 & d_{35} & 0 \end{vmatrix}. \qquad (3)$$

The matrices (1), (2), and (3) replace those that have hitherto been in common use according to Voigt's convention. To each elastic or piezoelectric constant on the new convention there corresponds one on the old, with the same numerical value.

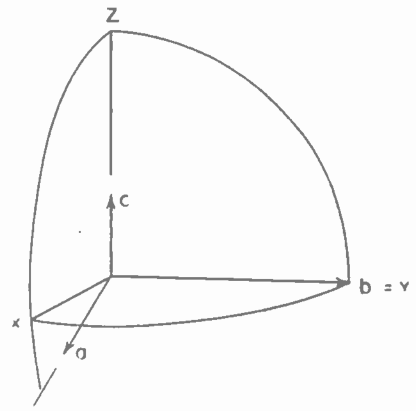

Fig. 3—Axes of monoclinic crystals.
a and c are perpendicular to b but not to each other.

For comparison with the new matrices (1), (2), and (3), the corresponding expressions according to Voigt's

[2] W. G. Cady, "Piezoelectricity," McGraw-Hill Book Co., New York, N. Y., 1946

[3] See part 3 of this Standard

convention will now be given. The Voigt matrix for elastic compliance, now abandoned, is

$$s = \begin{vmatrix} s_{11} & s_{12} & s_{13} & 0 & 0 & s_{16} \\ s_{12} & s_{22} & s_{23} & 0 & 0 & s_{26} \\ s_{13} & s_{23} & s_{33} & 0 & 0 & s_{36} \\ 0 & 0 & 0 & s_{44} & s_{45} & 0 \\ 0 & 0 & 0 & s_{45} & s_{55} & 0 \\ s_{16} & s_{26} & s_{36} & 0 & 0 & s_{66} \end{vmatrix} \qquad (4)$$

with an analogous form for c.

For class 2, the Voigt matrix (in which Z is a twofold axis), now abandoned, is

$$d = \begin{vmatrix} 0 & 0 & 0 & d_{14} & d_{15} & 0 \\ 0 & 0 & 0 & d_{24} & d_{25} & 0 \\ d_{31} & d_{32} & d_{33} & 0 & 0 & d_{36} \end{vmatrix}, \qquad (5)$$

while for class m (reflection-plane perpendicular to Z) the Voigt matrix, now abandoned, is

$$d = \begin{vmatrix} d_{11} & d_{12} & d_{13} & 0 & 0 & d_{16} \\ d_{21} & d_{22} & d_{23} & 0 & 0 & d_{26} \\ 0 & 0 & 0 & d_{34} & d_{35} & 0 \end{vmatrix}. \qquad (6)$$

If we know the s and d matrices for Voigt's convention, we can put them in standard form by means of the following matrices:

$$s = \begin{vmatrix} s_{22}' & s_{23}' & s_{12}' & 0 & s_{26}' & 0 \\ s_{23}' & s_{33}' & s_{13}' & 0 & s_{36}' & 0 \\ s_{12}' & s_{13}' & s_{11}' & 0 & s_{16}' & 0 \\ 0 & 0 & 0 & s_{55}' & 0 & s_{45}' \\ s_{26}' & s_{36}' & s_{16}' & 0 & s_{66}' & 0 \\ 0 & 0 & 0 & s_{45}' & 0 & s_{44}' \end{vmatrix} \qquad (7)$$

where the subscripts of the primed values are those according to Voigt.

Monoclinic, class 2

$$d = \begin{vmatrix} 0 & 0 & 0 & d_{25}' & 0 & d_{24}' \\ d_{32}' & d_{33}' & d_{31}' & 0 & d_{36}' & 0 \\ 0 & 0 & 0 & d_{24}' & 0 & d_{14}' \end{vmatrix} \qquad (8)$$

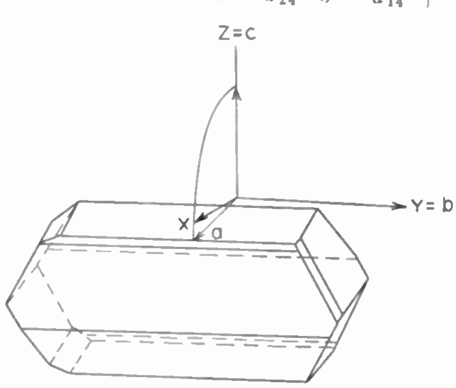

Fig. 4—Lithium sulfate monohydrate, a monoclinic crystal, class 2. The specimen here shown is left-handed. On extension along the b axis, the $+b$ end becomes negatively charged.

Monoclinic, class m

$$d = \begin{vmatrix} d_{22}' & d_{23}' & d_{21}' & 0 & d_{26}' & 0 \\ 0 & 0 & 0 & d_{35}' & 0 & d_{34}' \\ d_{12}' & d_{13}' & d_{11}' & 0 & d_{16}' & 0 \end{vmatrix}. \qquad (9)$$

Examples of monoclinic crystals are shown in Figs. 4, and 5, and 6.

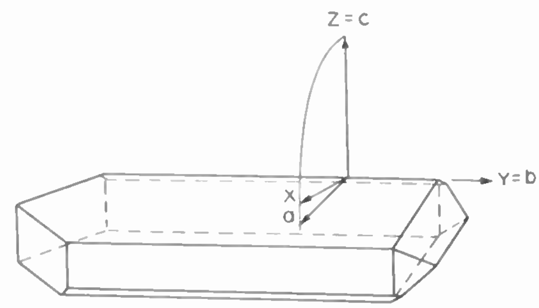

Fig. 5—Ethylene diamene d-tartrate (EDT), a monoclinic crystal, class 2. On extension along the b axis, the $+b$ end becomes positively charged.

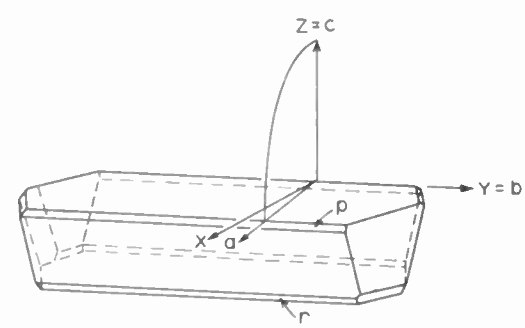

Fig. 6—Dipotassium d-tartrate (DKT), a monoclinic crystal, class 2. On extension along the b axis, the $+b$ end becomes positively charged.

1.5. The Orthorhombic System

Crystals having three mutually perpendicular twofold axes or two mutually perpendicular planes of reflection symmetry, or both, belong to the orthorhombic

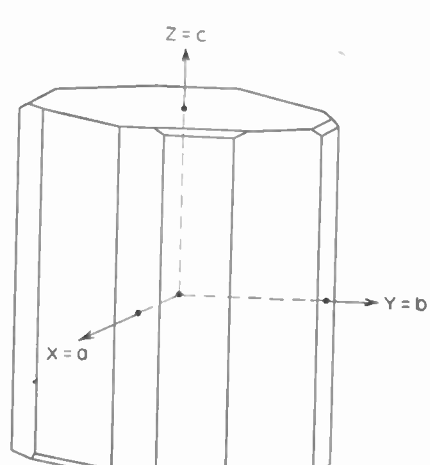

Fig. 7—Rochelle salt, an orthorhombic crystal, class 2 2 2.

system. The unit cell is a right-angled parallelepiped with the a, b, c axes of unequal length, and with unit distances $c_0 < a_0 < b_0$. Either end of any two axes may be taken as positive. The third axis is then given the proper sense to form a right-handed system. The following convention has been universally adopted: "*X* along *a*, *Y* along *b*, *Z* along *c*."

An example of an orthorhombic crystal is Rochelle salt, Fig. 7.

1.6. The Tetragonal System

Crystals having a single fourfold axis (or a fourfold inversion-axis) belong to the tetragonal system. The c axis is always taken along this fourfold axis, and the Z axis lies along c. The a and b axes are equivalent and are usually called a_1 and a_2. The X axis may be parallel to either a_1 or a_2; that is, to either a or b in Fig. 8. There are thus two possible sets of X and Y axes.

The orientation of a and hence of X is not so easily settled. There are seven classes of tetragonal crystals, five of which can be piezoelectric; these are classes $\bar{4}$, 4, $\bar{4}$ 2 m, 4 2 2, and 4 m m. Three of these have no twofold axes to guide in a choice of an a axis; however, for all of them except class $\bar{4}$ 2 m there is no alternative to the choice of an a axis in such a way as to make the unit cell of smallest volume. In class $\bar{4}$ 2 m, which has a twofold axis, the smallest cell may not have its a axis parallel to this axis. There are twelve possible arrangements of matter (space-groups) that have symmetry $\bar{4}$ 2 m. Of these twelve, six have the smallest cell when the a axis is an axis of twofold symmetry, and six have the smallest cell when a is chosen at 45 degrees to twofold axes (while still perpendicular to the c axis). The "International Tables for the Determination of Crystal Structure" give preference to the choice: "a along a twofold axis," and hence do not use the smallest possible cell. For piezoelectric studies this choice is more convenient than the smallest-cell choice. Since there is such good precedent for letting a lie along a twofold axis, there seems little likelihood of a conflict here.

In summary, it may be stated that for all tetragonal crystals having axes of twofold symmetry, one of these axes is taken as the a axis. When there is no twofold axis, the a axis is parallel to one of the two equal dimensions of the smallest unit cell. Arbitrarily take one end of the c axis as positive. Then use section 1.16 as a guide for the sense of the a axis as well as, in class $\bar{4}$ 2m, for the choice of the a axis.

The $+Z$ and $+X$ axes coincide with the $+c$ and $+a_1$ (or $+a_2$) axes respectively. The $+Y$ axis is such as to complete the right-handed rectangular axial system.

1.7. Application to Crystals of the ADP Type

Ammonium dihydrogen phosphate (ADP, Fig. 8), potassium dihydrogen phosphate (KDP), and the dihydrogen arsenates of ammonium and potassium, are all in class $\bar{4}$ 2 m. Since with these crystals the particular faces that would be needed to determine the positive senses of the a axes are usually absent, the following empirical rule is adopted. It is based on the fact that compression along a line 45 degrees from the a and b axes polarizes the crystal in the c direction, causing opposite charges to appear at the ends of the c axis. Stretching along the same line reverses the signs of the charges.

The rule is that the direction of a stretch (extension) that causes a positive charge to appear at the end of the Z axis which is chosen as the positive end, should lie in the quadrant between the positive directions of the X and Y axes. There are obviously two choices of positive directions, since the rule still holds if both axes are reversed. The rule is in accord with Section 1.16, (4) and (9).

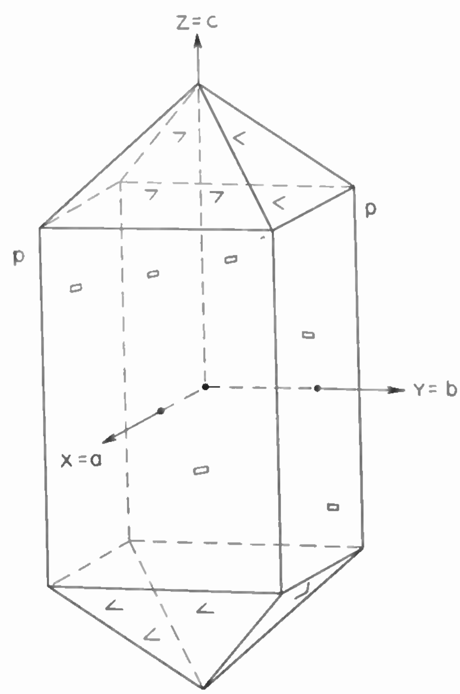

Fig. 8—Ammonium dihydrogen phosphate (ADP), a tetragonal crystal, class $\bar{4}$ 2 m. Potassium dihydrogen phosphate (KDP) is similar.
On extension along the diagonal pp, the $+z$ end becomes negatively charged.
Typical etch figures are shown on various faces.

1.8. The Hexagonal and Trigonal Systems

These systems are distinguished by an axis of sixfold (or threefold) symmetry. This axis is always called the c axis. According to the Bravais-Miller axial system, which is most commonly used, there are three equivalent secondary axes, a_1, a_2, and a_3, lying 120 degrees apart in a plane normal to c. These axes are chosen as being either parallel to a twofold axis or perpendicular to a plane of symmetry, or if there are neither twofold axes perpendicular to c nor planes of symmetry parallel to c, the a's are chosen so as to give the smallest unit cell.

According to the present convention, the Z axis is parallel to c. The X axis coincides in direction and sense with any one of the a axes. The Y axis is perpendicular to Z and X, so oriented as to form a right-handed system. This rule applies to both right- and left-handed crystals.

The axes of tourmaline are shown in Fig. 9.

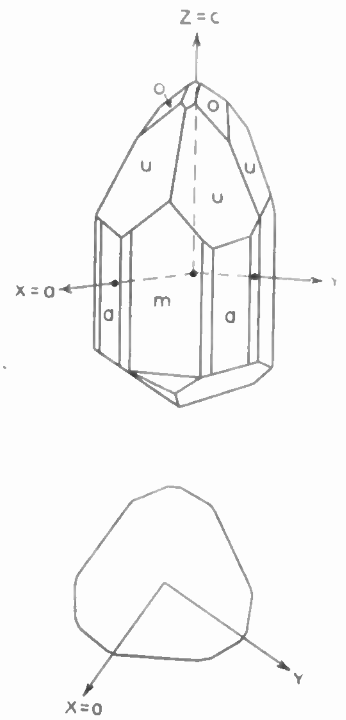

Fig. 9—Tourmaline, a trigonal crystal, class 3 m.
The Y axis lies in the plane of symmetry. On extension along Y the $+Y$ end becomes positively charged. On extension along Z the $+Z$ end becomes positively charged.

1.9. Application to Quartz

The axes according to the present convention are shown in Fig. 10. With both right and left quartz the X, Y, Z axes form a right-handed system. The effect of these changes[4] on the signs of elastic and piezoelectric constants and on the formulas for rotated axes is discussed in Section 1.12.

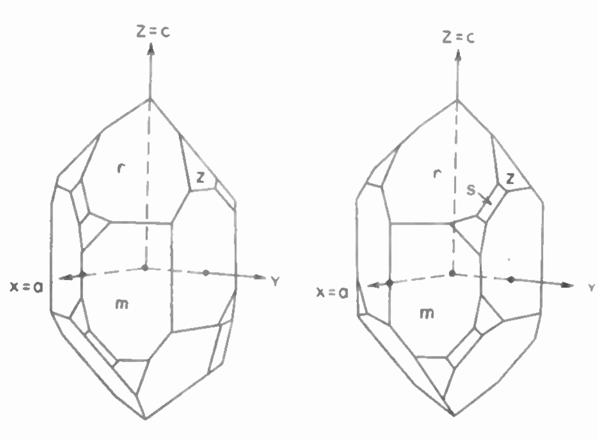

Fig. 10—Left and right quartz crystals, trigonal, class 3 2.

[4] For right quartz both X and Y axes are reversed with respect to the 1945 convention. For left quartz the Y axis is reversed, while the X axis is unchanged.

1.10. The Isometric (Cubic) System

The three equivalent axes are a, b $(=a)$, and c $(=a)$, often called a_1, a_2, and a_3. They are chosen parallel to axes of fourfold symmetry, or, if there is no true fourfold symmetry, then parallel to twofold axes. The X, Y, and Z axes form a right-handed system parallel to the a, b, and c axes.

An example is sodium bromate, Fig. 11.

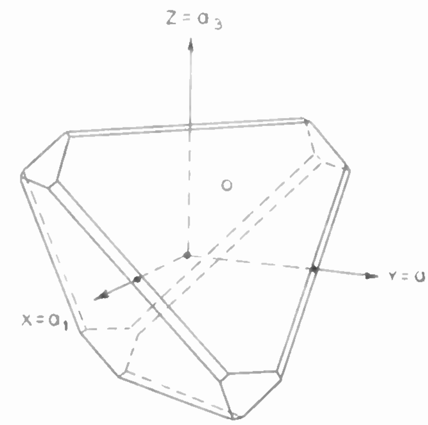

Fig. 11—Right crystal of sodium bromate, an isometric crystal, class 2 3.

1.11. Axes of Enantiomorphous Forms

In the following piezoelectric classes, both right and left forms are possible: triclinic 1, monoclinic 2, orthorhombic 2 2 2, tetragonal 4 and 4 2 2, trigonal 3 and 3 2 hexagonal 6 and 6 2 2, and isometric 2 3.

If a right crystal is placed in any orientation in front of a mirror, the image as seen in the mirror represents the corresponding left crystal. If the right crystal has right-handed rectangular axes, the axes of the left crystal will then appear left-handed. Nevertheless, it is standard crystallographic practice to use ***right-handed axial systems for all crystals, whether right or left.*** This convention is adopted in the present Standard for piezoelectricity. Under this convention the left form should be regarded as the crystallographic ***inversion*** of the right form, rather than as its mirror image.[5]

The signs of all elastic constants are the same for left and right crystals. Piezoelectric constants, however, have ***opposite*** signs for left and right crystals.

1.12. Special Conventions for Quartz

The rules given in the foregoing paragraph will now be applied to ordinary quartz, class 3 2 (alpha-quartz, the form occurring at temperatures below the α-β inversion point at 573° C).

The present convention for axes of right and left quartz has been stated in Section 1.9 (see Fig. 10). This choice of axes involves certain conventions respecting

[5] The inversion of any solid figure is produced by drawing from each point a line through an arbitrary point P and continuing this line for an equal distance beyond P. An inversion is equivalent to a 180 degree rotation about any axis through P followed by reflection in a mirror perpendicular to this axis of rotation at the point P.

the elastic constants c_{14} and s_{14}, as well as the piezoelectric constants $d_{11}=-d_{12}=-d_{26}/2$, $e_{11}=-e_{12}=-e_{26}$, $d_{14}=-d_{25}$, and $e_{14}=-e_{25}$.

Elastic constants: under this convention, *for both right and left quartz, c_{14} is negative, s_{14} is positive.**

As may be verified from the well-known fundamental equations ((11) and (12) in part 3), the *piezoelectric constants* of quartz must have the following algebraic signs in conformity with the present axial system:

Right quartz: d_{11} negative, e_{11} negative, d_{14} negative, e_{14} positive.

Left quartz: d_{11} positive, e_{11} positive, d_{14} positive, e_{14} negative.

When the piezoelectric constants g and h, occurring in (13) to (16) of part 3, are specialized for quartz, they become g_{11}, g_{14}, h_{11}, h_{14}, corresponding to Voigt's d_{11}, d_{14}, e_{11}, e_{14}, respectively. In all cases their signs are the same as those of d_{11}, d_{14}, e_{11}, and e_{14}, respectively.

1.13. Summary of Rules for Determining the Axes and Hand of Quartz Crystals According to the Present Recommendations

(A) The $+X$ axis should coincide with a $+a$ crystallographic axis, and the Z axis should coincide with the c crystallographic axis, as in Fig. 10.

(B) On extension, the positive ends of the a axes, and

TABLE I

Crystal System	Hermann-Mauguin Symbol	Axes: Crystallographic				Axes: Rectangular			+/− Axes	Schoenflies Symbol	Example
		c	a	b		X	Y	Z			
Triclinic $c_0<a_0<b_0$ α & $\beta>90°$	1	.					⊥ (010)			C_1	Aminoethyl ethanolamine hydrogen d-tartrate (AET)
	$\bar{1}$						⊥ (010)	c		S_2	Copper sulfate pentahydrate
Monoclinic $c_0<a_0$ $\beta>90°$; $\alpha=\gamma=90°$	m			/m		⊥ (100)	b	c	X or Z	$C_{1h}=C_s$	Clinohedrite
	2			2		⊥ (100)	b	c		C_2	Ethylene diamine d-tartrate (EDT)
	2/m			2		⊥ (100)	b	c		C_{2h}	Gypsum
Orthorhombic $c_0<a_0<b_0$ (for 2 2 2 & 2/m 2/m 2/m $\alpha=\beta=\gamma=90°$)	2 2 2	2	2	2		a	b	c		$V=D_2$	Rochelle salt (except between Curie points)
	2 m m	2	/m	/m		a	b	c	Z	C_{2v}	Hemimorphite
	2/m 2/m 2/m	2	2	2		a	b	c		$V_h=D_{2h}$	Barite
		c	a_1	a_2							
Tetragonal $a_0=b_0$ $\alpha=\beta=\gamma=90°$	4 m m	4	/m	/m		(a_1)	(a_2)	c	Z	C_{4v}	Silver fluoride monohydrate
	$\bar{4}$ 2 m	$\bar{4}$	2	2		(a_1)	(a_2)	c	Z	$V_d=D_{2d}$	Ammonium dihydrogen phosphate (ADP)
	4 2 2	4	2	2		(a_1)	(a_2)	c	*	D_4	Nickel sulfate hexahydrate
	4/m 2/m 2/m	4	2	2		(a_1)	(a_2)	c	*	D_{4h}	Zircon
	4	4	†			(a_1)	(a_2)	c	Z	C_4	Barium antimonyl tartrate
	$\bar{4}$	$\bar{4}$	†			(a_1)	(a_2)	c	Z	S_4	$Ca_2Al_2SiO_7$
	4/m	4	†			(a_1)	(a_2)	c	*	C_{4h}	Scheelite
Isometric $a_0=b_0=c_0$ $\alpha=\beta=\gamma=90°$	2 3	2	2	2		(a_1)	(a_2)	a_3	Z	T	Sodium chlorate
	$\bar{4}$ 3 m	$\bar{4}$	$\bar{4}$	$\bar{4}$		(a_1)	(a_2)	a_3	Z	T_d	Zinc blende (Sphalerite)
	4 3 2	4	4	4		(a_1)	(a_2)	a_3	*	O	None known
	2/m $\bar{3}$	2	2	2		(a_1)	(a_2)	a_3	*	T_h	Pyrite
	4/m $\bar{3}$ 2/m	4	4	4		(a_1)	(a_2)	a_3	*	O_h	Sodium chloride
		c	a_1	a_2	a_3						
Trigonal $(a_0)_1=(a_0)_2=(a_0)_3$	3	3	†			a_1		c	any two	C_3	Sodium periodate trihydrate
	$\bar{3}$	$\bar{3}$	†			a_1		c		$C_{3i}=S_6$	Dolomite
	3 m	3	/m	/m	/m	a_1		c	Y, Z	C_{3v}	Tourmaline
	3 2	3	2	2	2	a_1		c	X	D_3	α-quartz
	$\bar{3}$ 2/m	$\bar{3}$	2	2	2	a_1		c	X	D_{3d}	Calcite
Hexagonal $(a_0)_1=(a_0)_2=(a_0)_3$	$\bar{6}$	$\bar{6}$	†			a_1		c	X, Y	C_{3h}	None known
	$\bar{6}$ m 2	$\bar{6}$	2	2	2	a_1		c	X	D_{3h}	Benitoite (?)
	6	6	†			a_1		c	Z	C_6	Nephelite
	6 m m	6	/m	/m	/m	a_1		c	Z	C_{6v}	Wurtzite—2H
	6 2 2	6	2	2	2	a_1		c		D_6	β-quartz
	6/m	6				a_1		c		C_{6h}	Apatite
	6/m 2/m 2/m	6	2	2	2	a_1		c		D_{6h}	Beryl

* Equations for the elastic constants of quartz for rotated axes, as, for example, equations (50)–(54) in footnote reference 2, remain unchanged, owing to the reversal in sign of certain direction cosines on passing from the old to the present axial system.

therefore of the X axes, become negatively charged with right quartz, positively charged with left quartz.

(C) For right-handed quartz, the conoscope interference rings expand when the analyzer is turned clockwise by the observer.

1.14. Choice of Axes for Piezoelectric Crystals

Axes are assigned to crystals according to Table I. Crystal classes are listed, using the full Hermann-Mauguin symbols to designate the class. The method of selection of both the crystallographic axes and the rectangular axes of physics and engineering is to be read from the table. a, b, c are the crystallographic axes (Sections 1.3 to 1.10, and 1.13); a_0, b_0, c_0 refer to the dimensions of the unit cell along these axes; X, Y, Z are the rectangular axes, which should always form a right-handed system, whether for a left or a right crystal (Sections 1.9 and 1.11 to 1.13). α, β, γ are the angles between the pairs of crystallographic axes (Section 1.3). Both the Schoenflies and the Hermann-Mauguin symbols are given, although the use of the latter is preferred.

Explanation of Table I

In the column "Hermann-Mauguin Symbols" those classes which are piezoelectric are placed at the left.

Under "Axes," the numerals 2, 3, 4, 6 mean an axis of two-, three-, four- or six-fold symmetry; $\bar{4}$ (read 4 bar) a fourfold, $\bar{6}$ a sixfold axis of inversion; m an axis in a plane of symmetry; $/m$ an axis perpendicular to a plane of symmetry.

a, b, c are the crystallographic axes; X, Y, Z the rectangular axes. In some systems all, or two, of these axes are physically indistinguishable and said to be equivalent; the same symbol is then often repeated for the several equivalent axes, using a different subscript for each, as a_1, a_2, a_3.

a_0, b_0, c_0 are the edge lengths of the unit cell, parallel to the a, b, c axes, respectively.

α, β, γ are the angles between c and b, a and c, b and a, respectively.

The procedure for determining the a, b, c axes of any crystal involves satisfying a series of conventions. The first convention is indicated under the name of each system, and gives general rules for identification of axes in terms of the relative magnitudes of the several unit distances and of the angles between the crystallographic axes. With triclinic crystals, and with orthorhombic classes 2 2 2 and $2/m$ $2/m$ $2/m$, this one rule (Section 1.3) unambiguously prescribes all crystallographic axes and their senses. With monoclinic crystals a further rule is imposed, namely that the b axis is defined in terms of the symmetry according to the third column under "Axes." With the remaining systems the c axis is the first to be identified and is always the axis of high symmetry. Where the symbol † appears there is no special rule beyond that for the choice of the c axis, except that the remaining axes shall be selected in such a way as to give the smallest cell consistent with the specification of c.

Parentheses around a_1 and a_2 in columns 6 and 7 (tetragonal and isometric classes) indicate that the designation is arbitrary as to which of the two crystallographic axes perpendicular to $c(Z)$ shall be X and which shall be Y. Except for class $\bar{4}$ $2m$, either choice of sense may be made for the Z axis, after naming the X axis, and the choice will not affect the signs of the constants. The only restriction is that the axial system shall be right-handed. In three tetragonal classes (4 2 2, $4/m$ $2/m$ $2/m$ and $4/m$) and three isometric classes (4 3 2, $2/m$ $\bar{3}$ and $4/m$ $\bar{3}$ $2/m$) this choice is trivial in the sense that the signs, values and matrix positions of elastic, dielectric or piezoelectric constants are in no way affected thereby. These six classes are designated by an asterisk (*) in column 9.

The column headed "+/− Axes" indicates classes for which the rules given do not uniquely determine the axial system, particularly with respect to the senses of certain axes. A single axial symbol appearing in this column indicates that the sense of the axis named remains to be chosen by the first worker in the field, and that the signs of certain of the piezoelectric constants will depend upon the choice. See Section 1.16 for guidance in making this choice. Two letters in this column indicate that the two axes named may be similarly chosen, in general the first choice affecting the signs of certain piezoelectric constants and the second choice affecting those of some others. In the two situations named, there may also be certain piezoelectric constants whose signs are not affected by the choices. Finally, in class 3, as indicated, the senses of any two axes remain to be chosen and the choice affects the signs of the piezoelectric constants.

1.15. The Hermann-Mauguin Symbols[6]

In this system of notation an axis of rotation is indicated by one of the numbers 1, 2, 3, 4, 6. The number indicates through its reciprocal the part of a full rotation about the axis which is required to bring the crystal into an equivalent position in regard to its internal structural properties. The number 1 indicates no symmetry at all, since any structure must come back into coincidence after a complete rotation, while 2 indicates a twofold axis of rotation. $\bar{1}$, $\bar{2}$, $\bar{3}$, $\bar{4}$, $\bar{6}$ indicate axes of rotary inversion. $\bar{1}$ implies a simple center of inversion. $\bar{2}$ is equivalent to a reflection plane and since reflection planes are so important a feature of the structure the symbol for such a plane, m, is written instead of $\bar{2}$. If an axis has a reflection plane perpendicular to it, this fact is written as part of the symbol for that axis by following the number which describes the symmetry of the axis with the notation $/m$.

[6] Adapted from W. H. Bragg and W. L. Bragg, "The Crystalline State," vol. I, pp. 85–86, G. Bell and Co., London, 1933.

The designation for any class of symmetry is made up of one, two, or three symbols, each indicating an element of the symmetry. *The first symbol* in general refers to the principal axis of the crystal if there is one, indicating the type of symmetry of that axis and the existence of a reflection plane perpendicular to that axis, if any. *The second symbol* refers to secondary axes of the crystal, giving similarly the character of the symmetry involved and including reference to a reflection plane perpendicular to it if such exists, or refers to a reflection plane alone if no secondary axes exist. In the isometric system the secondary axes are the threefold axes inclined to the principal axis. *The third symbol* names tertiary axes if they exist, such as those parallel to $(11\bar{2}0)$ in the hexagonal system or (110) in the tetragonal system, or corresponding planes.

1.16. Positive Sense Guide

In lieu of other guiding factors, the following is suggested as a rule for guidance in setting up the axial system for crystals which are piezoelectric. Axial senses (as well as the selection of the Z axis or the choice between X and Y axes when this needs to be made) shall be those which provide a positive sign for the first one of the following constants which does not vanish: d_{33}, d_{11}, d_{22}, d_{36}, d_{31}. This selection of the group of sense-determining constants is somewhat arbitrary, giving emphasis to the crystallographic importance of the Z axis and to providing positive signs for the piezoelectric constants of extensional strain along axial fields. If the first application of the rule is not sufficient to determine uniquely the senses of all axes, then the rule is to be applied again to the second one of the constants which does not vanish; and again to the third if necessary. For crystals which are enantiomorphic, the rule should be applied as stated to the right crystal; for left crystals, the rule should read "negative" instead of "positive" in each reference to a piezoelectric constant. The senses of axes in the left crystal should be such as to reverse the signs of all piezoelectric constants with respect to those of the right crystal.

The effects of the application of the rule in the several classes may be summarized as follows:

(1) A positive d_{33} determines the sense of the Z axis in classes m, 2 m m, 4 m m, 4, 3, 3 m, 6, 6 m m.
(2) A positive d_{11} determines the sense of the X axis in classes 3, 3 2, $\bar{6}$, $\bar{6}$ m 2.
(3) A positive d_{22} determines the sense of the Y axis in classes 3 m, $\bar{6}$.
(4) A positive d_{36} determines the distinction between X and Y axes in class $\bar{4}$ 2 m.
(5) A positive $d_{14}=d_{36}$ determines the distinction between X and Y axes in classes 2 3, $\bar{4}$ 3 m after any one of the three crystallographic axes has been chosen arbitrarily as Z axis.
(6) A positive d_{31} determines the sense of the Z axis in class $\bar{4}$.
(7) The senses of all axes are trivial and reversals such as to maintain a right-axial system do not affect the signs of piezoelectric constants in classes 2 2 2, 4 2 2, 6 2 2.
(8) The senses of X and Z axes are trivial in class 2, and reversals such as to maintain a right-axial system do not affect the signs of piezoelectric constants.
(9) The senses of X and Y axes are trivial in classes $\bar{4}$ 2 m, 2 3, $\bar{4}$ 3 m, 2 m m, 4 m m, 4, 6, 6 m m.
(10) The sense of the Y axis is trivial in classes 3 2, $\bar{6}$ m 2.

2. Standard for Specifying Crystal Plate Orientation

2.1. All crystal plate specifications for orientation are to be determined by a "Rotational Symbol." This symbol is a set of letters and angles that indicate a way in which the orientation of the plate, assumed to be rectangular, can be derived from one of six initial orientations by successive rotations about plate edges.

The initial orientation is that in which the thickness, length and width fall along the X, Y, and Z axes, but not necessarily respectively. The X, Y, and Z axes for the various crystal systems are defined in part 1.

A. The first two letters of the symbol indicate the initial orientation used.
  (a) The *first* letter is x, y, or z and indicates the direction of the plate *thickness* before any rotations have been made.
  (b) The *second* letter is x, y or z and indicates the direction of the plate *length* before any rotations.
  (c) These two letters completely specify unrotated plates. Figs. 12 to 17 show the six possible cuts that require no rotation.

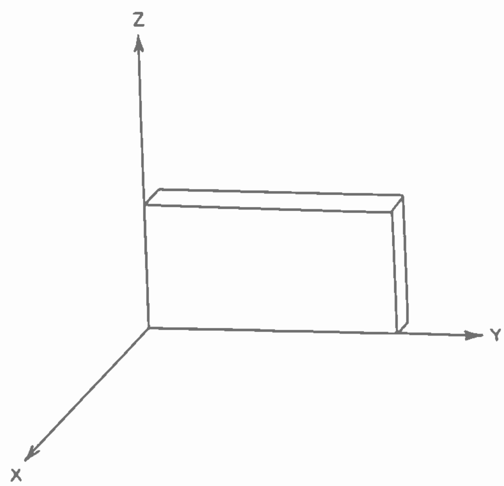

Fig. 12—An (xy) cut.

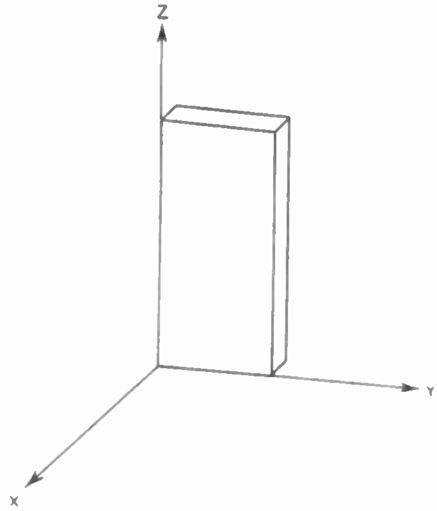

Fig. 13—A (*xz*) cut

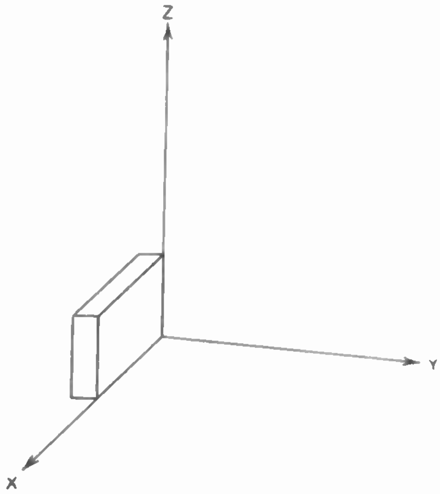

Fig. 14—A (*yx*) cut

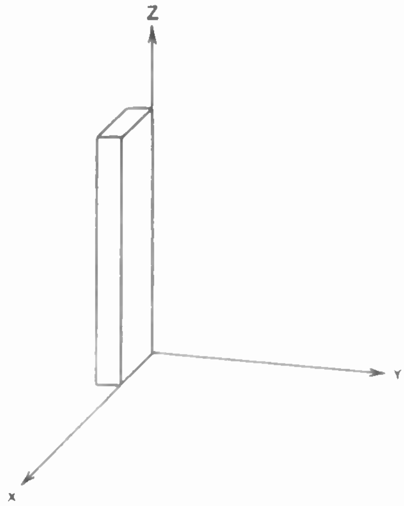

Fig. 15—A (*yz*) cut.

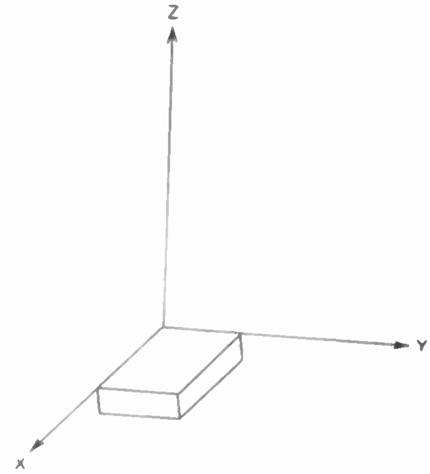

Fig. 16—A (*zx*) cut

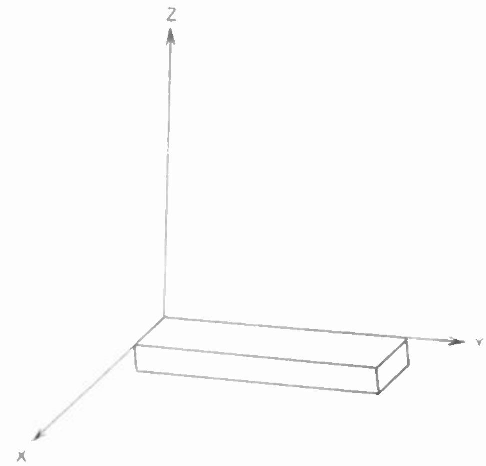

Fig. 17—A (*zy*) cut

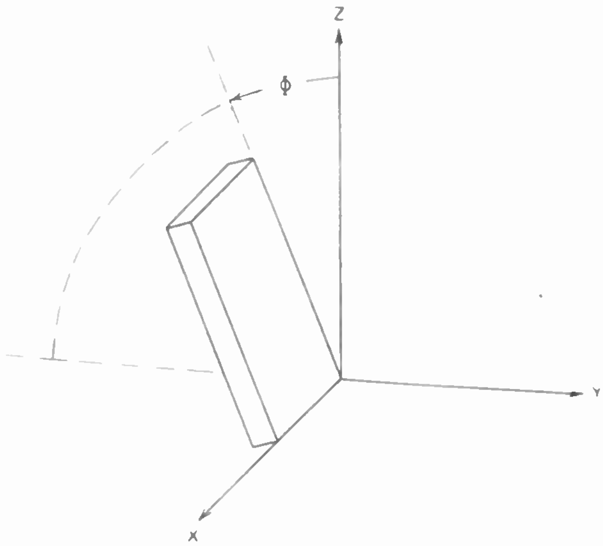

Fig. 18—A (*yzw*) cut, positive single rotation about *X*.

B. The remaining letters of the symbol indicate the plate edges used as axes of rotation.

(a) The *third* letter of the symbol is *t*, *l*, or *w* according to whether the thickness direction, length direction or width direction is the axis of first rotation. If one rotation suffices there are only three letters in the symbol. Several commonly used single rotation cuts of quartz are shown in Figs. 18 to 21.

(b) The *fourth* letter is *t*, *l*, or *w* according to the edge used for the second rotation. If two rotations suffice there are only four letters in the

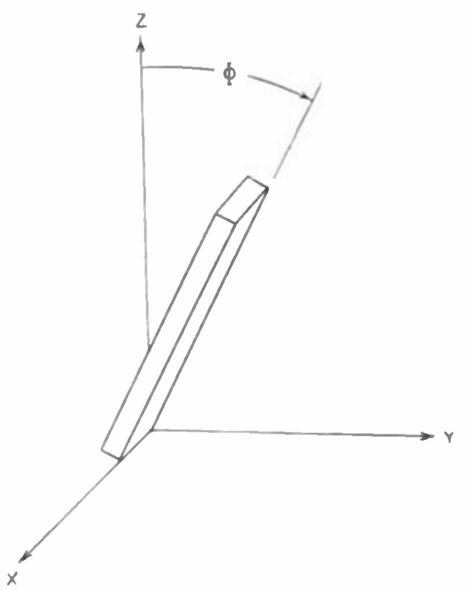

Fig. 19—A (*yzw*) cut, negative single rotation about *X*.

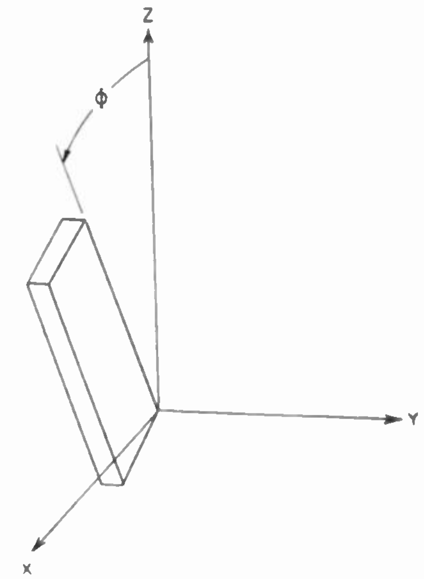

Fig. 21—A (*yzt*) cut, rotation about *Y*.

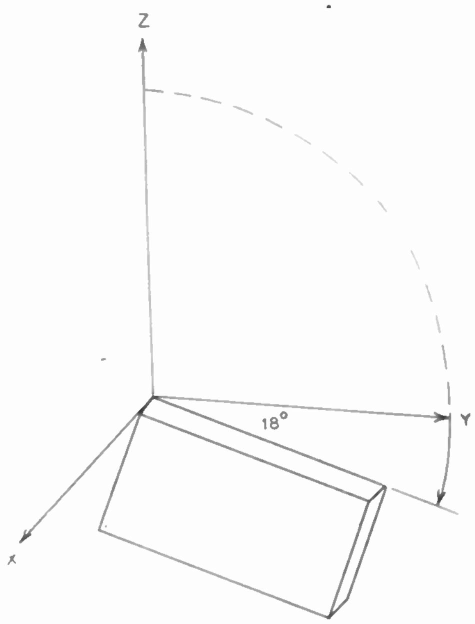

Fig. 20—An (*xyt*) cut, −18° rotation about *X*

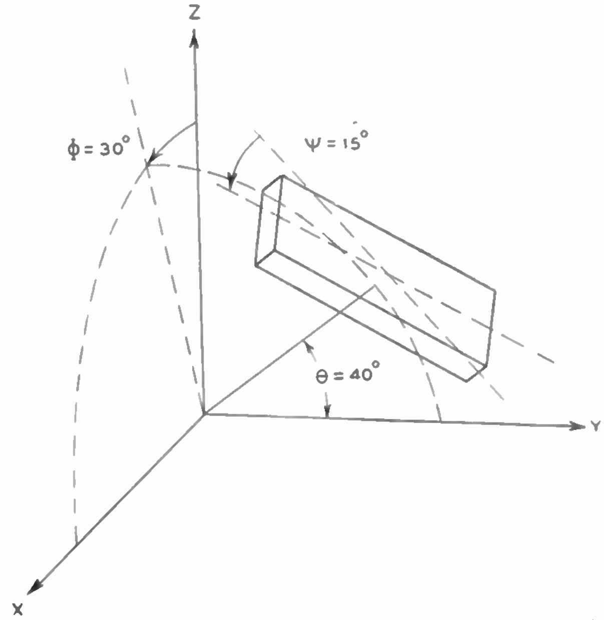

Fig. 23—A triple-rotation cut (*yztwt*), produced from Fig. 22(c) by 15° rotation about *t*.

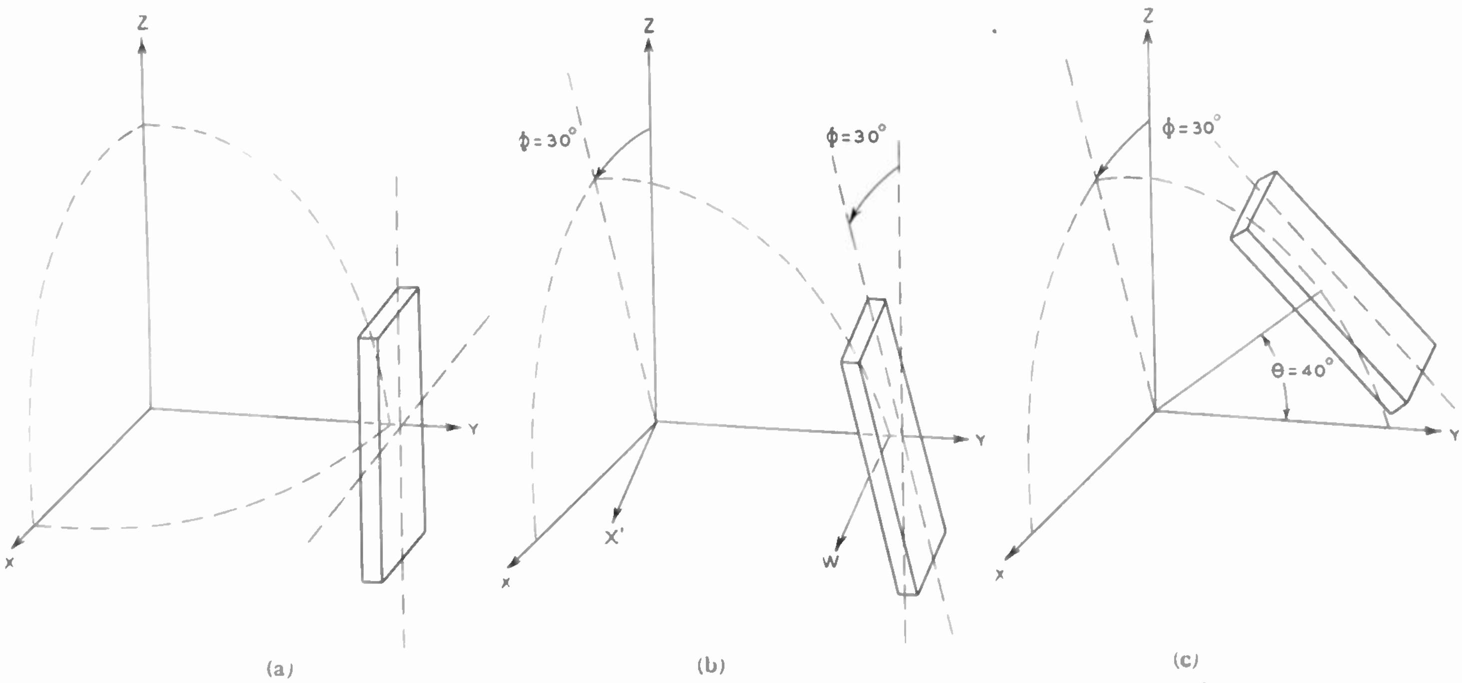

Fig. 22—A double-rotation cut (*yztw*) arrived at by 30° rotation about *t* followed by 40° rotation about *w*

symbol. A double-rotation cut is shown in Fig. 22.

(c) The *fifth* letter is *t*, *l*, or *w* according to the edge used for the third rotation. There need be no more than five letters in the symbol. A triple-rotation cut is shown in Fig. 23.

C. The symbol is to be followed by a list of rotation angles Φ, Θ, Ψ; angles negative in sense will be indicated by a negative sign.

(a) A positive angle means rotation counter-clockwise as seen looking toward the origin from the positive end of the axis of rotation.

(b) The positive ends of the axes *t*, *l*, and *w* are the ends that initially pointed in the positive directions of the three coordinate axes *X*, *Y* and *Z*.

D. Right-handed crystals and left-handed crystals (see Fig. 10) use exactly the same symbol and angle list to give ***elastically*** equivalent plates. As is explained in Section 1.12, the signs of all piezoelectric constants for right crystals are opposite to those for left crystals. Nevertheless, the numerical values of the piezoelectric constants are the same for left as for right crystals (see also Section (b) below).

(a) A left-handed crystal is considered as the crystallographic inversion of a right-handed crystal rather than as its mirror image (see Section 1.11).

(b) The case of crystals that exist in both right-handed and left-handed forms, as is the case for quartz, will now be discussed. Consider a plate *R* made from a right-handed crystal according to some one orientation specification. What orientation specification will produce, from a left-handed crystal, a plate *L* with properties identical to those of plate *R*? The answer is that it is impossible to orient the *L* plate so that *all* its properties are identical with those of the *R* plate. It is, however, possible so to orient *L* as to have its elastic properties identical with those of *R*, and the piezoelectric properties differing only in sign. This "equivalence" is satisfactory in most piezoelectric work. A left crystal can be so positioned with respect to a right crystal that it can be considered as the right crystal inverted through a center of inversion. (Place them in a mirror-image relation first, then rotate one of them 180° in the plane of the mirror.) In this relative position, we can describe the two crystals on identical systems of axes. Each face (*hkl*) of one crystal is now matched by a parallel face, ($\bar{h}\bar{k}\bar{l}$) of the other crystal; the plate *R* of one crystal is now parallel to the plate *L* of the other crystal, *R* and *L* having the "equivalence" mentioned above. Hence ***the rotation-system symbol is the same for both plates.*** If when sawing to produce a given orientation the crystals are supported on two or more faces (reference faces) which rest on surfaces of a sawing fixture, this fixture will serve for both right and left forms if the crystal has a second face parallel to each reference face. This is ordinarily the case for all but crystals of low symmetry, such as triclinic asymmetric crystals.

E. As examples of the association of familiar angles, a quartz *AT* plate, with its length along the *X* axis, is designated as (*yxl*) 35 1/4°. A *BT* cut is (*yxl*) −49°. It is to be noted that most manufacturers are accustomed to the descriptions 35 1/4° rotated *Y* cut and −49° rotated *Y* cut respectively for these two crystals. The designation of the familiar 45° *X* cut Rochelle salt plate is *xyt* 45°; of the 45° *Z* cut plate of *ADP*, *zxt* 45°; and of the *Y* cut of *EDT*, *yx*.

It is convenient to add the angles and the dimensions to the rotation symbol to complete the crystal plate specification. For example, as in Fig. 23, (*yztwl*) 30°/15°/40° means (*yztwl*) with Φ=30° about *t*, Θ=15° about *w*, and Ψ=40° about *l*. A statement of the final required dimensions will now complete the specification, so that a full specification might read:

(*yztwl*) 30°/15°/40°

$t = 0.80 \pm 0.01$ mm

$l = 40.0 \pm 0.1$ mm

$w = 9.03 \pm 0.03$ mm.

F. It might be thought that an independent set of rotation systems would result if the rotations were about *X*, *Y*, and *Z* in some order instead of about *t*, *l* and *w*. For example, we might extend our nomenclature to cover such a symbolization as (*yxxyz*). Here the *yx* tells us that the thickness was originally along *Y* and the length originally along *X*. The *xyz* tells us that the first rotation is about the *X* axis, the second about the *Y* axis, and the third about the *Z* axis. For single rotations there is no difference between this and the previous system as, for example, a (*yxx*) is identical with a (*yxl*) since *x* and *l* coincide in this case. For multiple rotations it can be shown that a simple relation connects the two sets, (*yxxyz*)$\alpha/\beta/\gamma$ being in fact equivalent to (*yxwtl*)$\gamma/\beta/\alpha$. (Note that although *x*, *y*, *z* coincide initially with *l*, *t*, *w*, respectively, the *ltw* is reversed with respect to *wtl* and the order of the angles is also reversed in writing the equivalent symbol.) For example, a quartz *GT* cut is either a (*yxlt*) 51°/45° or alternatively a (*yxyx*) 45°/51°. The (*yxwtl*) form is to be preferred for the reason that most designers think in these terms, not in the (*yxxyz*) terms. This is because the (*yxwtl*) type preserves the direction of one crystal edge for each rotation and hence preserves certain plate properties but allows other properties to vary, thus permitting their improvement. The (*yxxyz*) type preserves no edge direction of a plate except in the case of singly rotated cuts.

3. The Piezoelectric Relations, Symbols, and Units

3.1. General

Piezoelectric investigations usually involve the elastic and dielectric constants of the material as well as the piezoelectric. As in other fields, confusion has arisen through the use of different symbols for the same quantity and of the same symbol for different quantities. Additional difficulties have come from Voigt's selection of a compressive stress as positive. This is not in accord with accepted usage by elasticians, nor is it followed by all writers on piezoelectricity.

A suitable notation for the quantities of interest in piezoelectricity should provide a single symbol for each quantity with the various components designated by subscripts to permit the use of either the matrix or tensor methods of writing the equations. This requirement precludes the adoption of either of the two most widely used notations for stress and strain. Piezoelectric notation is further complicated by the fact that, in general, the electrical, mechanical and sometimes the thermal conditions of measurement must be specified before a unique meaning can be given to the constants of the material. It is therefore highly desirable to provide a notation where the boundary conditions can be specified in the symbol.

In order to promote uniformity and accuracy of presentation, a set of standard symbols is here presented. Wherever possible these symbols conform to accepted usage in the field concerned. A table relating them to those used by various authors is also given. The piezoelectric relations that serve to define the constants of the material, together with the various relations among these constants, are given in both matrix and tensor form. These relations are written in rationalized mks units, and a set of factors is given for converting from cgs electrostatic units to rationalized mks units. For comparison with Voigt's notation, the relations are also given, in matrix form, with electrostatic units and Voigt's symbols, but calling the stress positive when tensile.

3.2. Specification of the Variables

3.2.1. The electrical state of a medium is known when two vector quantities such as electric field and electric displacement are specified. The elastic state is known when two second-order tensor quantities, stress and strain, are specified. Piezoelectricity is concerned with the interaction between the electrical and elastic behavior of a crystal and therefore with relations involving the two electrical and the two elastic variables.

All quantities are referred to rectangular axes designated by X, Y, and Z, or by X_1, X_2, and X_3. These axes are related to the crystallographic axes as explained in part 1 of this Standard.

3.2.2. The Electrical Variables

The electrical variables are chosen as the electric field (E) and the electric displacement (D). The electric displacement is chosen as the second electrical variable in preference to other possible variables (*e.g.*, polarization) as being more useful from an engineering and experimental point of view.

The components of the electric field and electric displacement are designated by E_i and D_i, respectively. The subscript i takes the values 1, 2, 3 and denotes the axis along which the component is directed.

3.2.3. The Elastic Variables

The elastic variables are stress (T) and strain (S). The tensor components of stress T are designated by two subscripts ($i=1, 2, 3$ and $j=1, 2, 3$). Since $T_{ij}=T_{ji}$, only 6 of the 9 components are independent, and the stress components can be written with a single subscript $T_p(p=1 \cdots 6)$. T_p can be related to T_{ij} in different ways. The usual convention is represented below.

The relations of the T_{ij} to T_p, and to the components of stress as written in certain texts, are as follows (the abbreviated form is the basis of the matrix notation used below).

Tensor Form	Abbrev. Form	Voigt	Other Texts	
T_{11}	T_1	X_x	σ_{xx}	Normal stresses
T_{22}	T_2	Y_y	σ_{yy}	
T_{33}	T_3	Z_z	σ_{zz}	
T_{23}	T_4	Y_z	τ_{yz}	Shearing stresses.
T_{13}	T_5	Z_x	τ_{zx}	
T_{12}	T_6	X_y	τ_{xy}	

The tensor strain components are designated by S_{ij}. As with the stress components, $S_{ij}=S_{ji}$, so that only six of the strain components are independent and the strain components can also be written with a single subscript $S_p(p=1 \cdots 6)$.

The relations between the strain components in different notations are as follows:

Tensor Form	Abbrev. Form	Voigt	Other Texts	
S_{11}	S_1	x_x	ϵ_{xx}	Normal strains
S_{22}	S_2	y_y	ϵ_{yy}	
S_{33}	S_3	z_z	ϵ_{zz}	
$2S_{23}$	S_4	y_z	γ_{yz}	Shearing strains.
$2S_{13}$	S_5	z_x	γ_{zx}	
$2S_{12}$	S_6	x_y	γ_{xy}	

Components of elastic displacement are designated by $u_j(j=1, 2, 3)$ and are related to the strain components as follows:

$$S_1 = S_{11} = \frac{du_1}{dx_1}$$

$$S_2 = S_{22} = \frac{du_2}{dx_2}$$

$$S_3 = S_{33} = \frac{du_3}{dx_3}$$

$$S_4 = 2S_{23} = \frac{du_3}{dx_2} + \frac{du_2}{dx_3}$$

$$S_5 = 2S_{13} = \frac{du_1}{dx_3} + \frac{du_3}{dx_1}$$

$$S_6 = 2S_{12} \quad \frac{du_2}{dx_1} + \frac{du_1}{dx_2}.$$

A normal strain is positive when extensional, negative when compressional. A shearing stress, and also a shearing strain, is positive when a rectangle becomes deformed so that an acute angle lies in the quadrant between the positive directions of the axes that form the sides of the rectangle.

3.3. The Piezoelectric Relations

3.3.1. General

When T_p and E_i are chosen as the independent variables, the piezoelectric equations for the most general case (that of a triclinic crystal) are:

$$\begin{aligned}
S_1 &= s^E_{11}T_1+s^E_{12}T_2+s^E_{13}T_3+s^E_{14}T_4+s^E_{15}T_5+s^E_{16}T_6 +d_{11}E_1+d_{21}E_2+d_{31}E_3\\
S_2 &= s^E_{12}T_1+s^E_{22}T_2+s^E_{23}T_3+s^E_{24}T_4+s^E_{25}T_5+s^E_{26}T_6 +d_{12}E_1+d_{22}E_2+d_{32}E_3\\
S_3 &= s^E_{13}T_1+s^E_{23}T_2+s^E_{33}T_3+s^E_{34}T_4+s^E_{35}T_5+s^E_{36}T_6 +d_{13}E_1+d_{23}E_2+d_{33}E_3\\
S_4 &= s^E_{14}T_1+s^E_{24}T_2+s^E_{34}T_3+s^E_{44}T_4+s^E_{45}T_5+s^E_{46}T_6 +d_{14}E_1+d_{24}E_2+d_{34}E_3\\
S_5 &= s^E_{15}T_1+s^E_{25}T_2+s^E_{35}T_3+s^E_{45}T_4+s^E_{55}T_5+s^E_{56}T_6 +d_{15}E_1+d_{25}E_2+d_{35}E_3\\
S_6 &= s^E_{16}T_1+s^E_{26}T_2+s^E_{36}T_3+s^E_{46}T_4+s^E_{56}T_5+s^E_{66}T_6 +d_{16}E_1+d_{26}E_2+d_{36}E_3\\
D_1 &= d_{11}T_1+d_{12}T_2+d_{13}T_3+d_{14}T_4+d_{15}T_5+d_{16}T_6 +\epsilon^T_{11}E_1+\epsilon^T_{12}E_2+\epsilon^T_{13}E_3\\
D_2 &= d_{21}T_1+d_{22}T_2+d_{23}T_3+d_{24}T_4+d_{25}T_5+d_{26}T_6 +\epsilon^T_{12}E_1+\epsilon^T_{22}E_2+\epsilon^T_{23}E_3\\
D_3 &= d_{31}T_1+d_{32}T_2+d_{33}T_3+d_{34}T_4+d_{35}T_5+d_{36}T_6 +\epsilon^T_{13}E_1+\epsilon^T_{23}E_2+\epsilon^T_{33}E_3
\end{aligned} \qquad (10)$$

where s^E is the elastic compliance at constant field-strength, in matrix notation (3.3.8), ϵ the dielectric permittivity,[7] and d the piezoelectric strain-constant. Rationalized mks units are implied in the equations, and throughout this report, unless otherwise stated. Since, according to the definitions given above, a positive stress is associated with a positive strain, the signs of all terms containing stress differ from those used by Voigt.

[7] The term "permittivity" is synonymous with "capacitivity" as defined in *Standards on Abbreviations, Graphic Symbols, Letter Symbols, and Mathematical Signs, 1948*. The word "permittivity" is adopted for this *Standard* because of its almost universal use in piezoelectric and dielectric theory.

In order to minimize the number of subscripts the well-established symbols ϵ_0 and K are used in this Standard instead of the symbols ϵ_v and ϵ_r given in *Standards on Abbreviations*, etc., 1948, referred to above.

The 36 compliance constants s form a matrix relating strain S to stress T. In matrix notation this relation may be written $S=s^ET$. Similarly the 18 piezoelectric strain-constants d form a matrix relating S to E or D to T. In matrix notation the piezoelectric contributions to S and D are written as $S=d_tE$, $D=dT$, where d_t signifies a transposed matrix.

It will be observed that the d-matrix appearing in the first three equations (10) has six rows and three columns, while in the last three equations it has three rows and six columns; in the latter case the rows and columns are interchanged, and the d-matrix in the first six equations is a *transposed matrix* with respect to the one having three rows and six columns. The transposed d-matrix has the special designation d_t. It should be noted that the transposed matrix (three columns) is to be used wherever a piezoelectric constant is associated with an electric vector, the latter being regarded as a matrix with one column and three rows.

3.3.2. Equations in Voigt's Notation

Before passing to the formulation that is of chief concern in this report, we give for comparison the four familiar equations of state according to Voigt, in cgs electrostatic units, but with the stress X positive when the strain x is positive, using matrix notation.

$$\begin{aligned} x &= s^EX + d_tE \qquad & X &= c^Ex - e_tE \\ P &= dX + \eta^XE & P &= ex + \eta^xE \end{aligned} \qquad \text{(not standard)} \qquad (11)$$

where P is the polarization, and η^X and η^x are the dielectric susceptibilities at constant stress and constant strain. The equations for P can be changed to equations for D by multiplying each side by 4π and then adding E. Then, since in esu $1+4\pi\eta=K$, it follows that $D=4\pi dX+K^TE$, and $D=4\pi ex+K^SE$. When these equations, together with those for x and X above, are converted to mks units, with T and S in place of X and x, the result is

$$\begin{aligned} S &= s^ET + d_tE \qquad & T &= c^ES - e_tE \\ D &= dT + \epsilon^TE & D &= eS + \epsilon^SE. \end{aligned} \qquad (12)$$

Here d and e are in rationalized mks units, obtained by dividing the quantities in esu by $3(10^4)$ and $3(10^5)$, respectively. The two equations at the left are the same as (10) in matrix notation. Those at the right are equivalent to another set of equations similar to (10), but with S_p and E_i as independent variables.

3.3.3. *Recommended Equations of State in Matrix Notation*

In many piezoelectric investigations it is desirable to use as independent variables S_p and D_i, or T_p and D_i. In these cases the c's and s's should have the superscript D instead of E, indicating that they give the relations between stress and strain when D is held constant. Two more sets of equations similar to (10) are needed. In one set, S and E are expressed in terms of T and D; in the other set, T and E are expressed in terms of S and D.

It is unnecessary to write all these equations in full, since their content can be expressed at once in matrix form, as follows:[8]

$$S = s^D T + g_t D \qquad (13)$$

$$E = -gT + \beta^T D \qquad (14)$$

$$T = c^D S - h_t D \qquad (15)$$

$$E = -hS + \beta^S D \qquad (16)$$

For comparison with these expressions, (12) are here repeated:

$$S = s^E T + d_t E \qquad (12a)$$

$$D = dT + \epsilon^T E \qquad (12b)$$

$$T = c^E S - e_t E \qquad (12c)$$

$$D = eS + \epsilon^S E. \qquad (12d)$$

In the foregoing equations ϵ is the dielectric permittivity, and β is the dielectric impermeability, related to ϵ as indicated in Tables III and IV. g and h are the piezoelectric constants corresponding to d and e. When d and e are in rationalized mks units, as in (12a) to (12d), the relations are:

$$d = \epsilon^T g \qquad e = \epsilon^S h. \qquad (17)$$

The superscripts designate the boundary conditions. s^E is the elastic compliance when the electric field is constant or zero (short circuit); ϵ^T is the permittivity when the stress is constant or zero (free crystal).

Constants without a superscript to designate the thermal conditions, such as s, d, ϵ, are considered to be those measured under adiabatic conditions. Constants measured isothermally should be designated by a superscript θ.

For greater generality, (12) to (16) could include thermal terms as well as elastic and electric; in that case each coefficient belonging to any one of the three effects (elastic, electric, and thermal) would have to be provided with two superscripts, indicating the boundary conditions with respect to the other two effects. For example, a compliance constant measured adiabatically and at constant displacement would be $s_{ij}^{\sigma D}$. In all expressions involving thermal effects the symbol σ is to be used for entropy per unit volume, and θ for absolute temperature. Further consideration of the treatment of thermal effects lies outside the scope of this Standard.

[8] The form of the piezoelectric relations in (12) to (16) was first proposed by H. G. Baerwald, in OSRD Report No. 287, Contract No. OEMsr 120, 1941.

TABLE II

Quantity	Units[5] (mks)	Symbols						
		Adopted	(a) Cady	(b) Mason	(c) Wooster	(d) Voigt	(e) Mueller	(f) Baerwald
Stress[1]	N/m^2	T	X_x	X_x	t	X_x	X_x	σ
Strain[1]		S	x_x	x_x	r	x_x	x_x	η
Elastic displacement	m	u						
Sign of tensile stress		+	−	−	+	−	−	+
Electric field strength	V/m	E	E	E	E	E	E	E
Electric displacement	C/m^2	D	D	D	D	D	D	D
Absolute temperature		θ						
Entropy per unit volume		σ						
Elastic compliance[2] (E = constant)	m^2/N	s^E	s^E	s^E	s	s	s	s
Elastic compliance (D = constant)	m^2/N	s^D	s^D	s^Q				s
Elastic stiffness (E = constant)	N/m^2	c^E	c^E	c^E	c	c		c
Elastic stiffness (D = constant)	N/m^2	c^D	c^D	c^Q			c	c
Permittivity (T = constant)	F/m	ϵ^T	k'	K^F			ϵ_F	ϵ
Permittivity (S = constant)	F/m	ϵ^S	k''	K^C				ϵ
Dielectric constant, relative[3]		K						
Dielectric impermeability (T = constant)	m/F	β^T	θ'					β
Dielectric impermeability (S = constant)	m/F	β^S	θ''					β
Dielectric susceptibility (T = constant)		η^T	η'	κ	k	η	κ	
Dielectric susceptibility (S = constant)		η^S	η''	κ^C		η'	κ_{cl}	
Reciprocal susceptibility (T = constant)		χ^T	χ'	χ^F				
Reciprocal susceptibility (S = constant)		χ^S	χ''	χ			χ	
Dielectric polarization	C/m^2	P	P	P	p	P	P	P
Piezoelectric strain-constant[4] Eq. (12)	$\frac{C}{N}$ or $\frac{m}{V}$	d	d	d	q	d	d	d
Piezoelectric stress-constant[4] Eq. (12)	C/m^2	e	e	e	e	e	e	e
Piezoelectric strain-constant[4] Eq. (14)	m^2/C	g		g				h
Piezoelectric stress-constant[4] Eq. (16)	$\frac{N}{C}$ or $\frac{V}{m}$	h		f				g
Coupling coefficient		k		k			k	κ

(a) As given in "Piezoelectricity," McGraw-Hill, 1946.
(b) As given in several papers in the *Physical Review*, 1939–1945.
(c) As given "Crystal Physics," Cambridge Univ. Press, 1938.
(d) As given in "Lehrbuch der Kristallphysik," Teubner, 1928.
(e) As given in several papers in the *Physical Review*, 1935–1940.
(f) As given in OSRD Report No. 287, Contract No. OEMsr-120 (1941).

For numbered footnotes see *Explanation of Table II* on page 1394.

This superscript notation can be extended to mixed boundary conditions. For example, in the important practical case of thickness vibrations, the elastic constant governing the vibration would properly be designated as $c^{E_t D_n}$, indicating that the transverse components of the field E are constant and that the normal component of the displacement D is constant. This is the meaning to attach to c in thickness vibrations, although the superscript may be omitted.

In equations (12) to (16) all symbols denote matrices. S and T are column matrices of 6 terms; E and D column matrices of 3 terms; c and s symmetrical matrices of 6 columns and 6 rows; and d, e, g, h, 6 columns and 3 rows; ϵ and β, symmetrical matrices of 3 columns and 3 rows.

3.3.4. The four piezoelectric constants (d, e, g, and h)

are related as indicated above, but each presents a different aspect of the piezoelectric relationship and is useful for a particular set of conditions. For example, d measures the strain in a free crystal for a given applied field, e the stress developed by a given field when the crystal is clamped, g the open circuit voltage for a given stress, and h the open circuit voltage for a given strain.

3.3.5. In Table II are listed the symbols adopted for this Standard and also those used by various authors for elastic, electric, and piezoelectric quantities.

Explanation of Table II. The numbered paragraphs relate to the numerical superscripts in the table.

1. X_x for stress and x_x for strain indicate that the notation used was X_x, Y_y, Z_z, Y_z, Z_x, X_y for stress and x_x, y_y, z_z, y_z, z_x and x_y for strain; or, alternatively, $X_1 \cdots X_6$ and $x_1 \cdots x_6$.
2. Other superscripts are to be used for other boundary conditions as necessary; *e.g.* for zero polarization or infinite air gap.
3. ϵ_0 is the permittivity of free space. In the electrostatic system ϵ_0 is equal to 1 and $\epsilon = K$, numerically. In the mks system $\epsilon_0 = 8.85 \times 10^{-12}$ farad/meter $= 1/(36\pi \times 10^9)$ farad/meter. See footnote 7.
4. The numbers in parentheses refer to equations (13) to (16) which define the different piezoelectric constants. The piezoelectric constant f was introduced by Mason[9] to designate the ratio of elastic stress to charge-density on the electrodes, in cgs esu. As can be seen from (15), the piezoelectric stress-constant h corresponds to f, but represents the ratio of stress to displacement rather than to charge-density. The constants d and e have long been used, following Voigt; g was introduced by Baerwald.[8]
5. The letters referring to the units of the mks system have the following meaning:
   - N = newton, the unit of force = 10^5 dynes.
   - m = meter, the unit of length = 100 centimeters
   - V = volt, the unit of potential = 1/300 statvolt
   - C = coulomb, the unit of charge = 3×10^9 statcoulombs
   - F = farad, the unit of capacitance = 9×10^{11} statfarads.

3.3.6. The relations among the constants in (13) to (16) are given in matrix notation in Table III. Units are rationalized mks.

TABLE III

$$c^E = (s^E)^{-1} \qquad d = es^E = \epsilon^T g$$
$$c^D = (s^D)^{-1} \qquad e = dc^E = \epsilon^S h$$
$$\beta^T = (\epsilon^T)^{-1} \qquad g = hs^D = \beta^T d$$
$$\beta^S = (\epsilon^S)^{-1} \qquad h = gc^D = \beta^S e$$
$$\epsilon^T - \epsilon^S = dc^E d_t = es^E e_t = de_t$$
$$\beta^S - \beta^T = hs^D h_t = gc^D g_t = hg_t$$
$$c^D - c^E = e_t \beta^S e = h_t \epsilon^S h = h_t e$$
$$s^E - s^D = g_t \epsilon^T g = d_t \beta^T d = d_t g$$

[9] W. P. Mason, "A dynamic measurement of the elastic, electric, and piezoelectric constants of Rochelle Salt," *Phys. Rev.*, vol. 55, pp. 775–789; June, 1939.

Matrices of the c, s, d, e, and ϵ constants applicable to each of the crystal classes except the monoclinic are given by Cady.[10] For the revised forms of the monoclinic matrices see Section 1.4. From the matrices and the relations in Table III the other constants may be obtained.

3.3.7. Tensor Notation

In full tensor notation (12) to (16) are written:

$$S_{ij} = s^E{}_{ijkl} T_{kl} + d_{ijm} E_m \qquad S_{ij} = s^D{}_{ijkl} T_{kl} + g_{ijm} D_m$$
$$D_n = d_{nkl} T_{kl} + \epsilon^T{}_{nm} E_m \qquad E_m = -g_{mkl} T_{kl} + \beta^T{}_{mn} D_n$$
$$T_{kl} = c^E{}_{ijkl} S_{ij} - e_{klm} E_m \qquad T_{kl} = c^D{}_{ijkl} S_{ij} - h_{kln} D_n$$
$$D_n = e_{nij} S_{ij} + \epsilon^S{}_{nm} E_m \qquad E_m = -h_{mij} S_{ij} + \beta^S{}_{mn} D_n$$

where subscripts repeated in the same term indicate summation. All subscripts take the values 1, 2, 3.

The relations among the constants in tensor notation are given in Table IV. In this Table, I is the idemfactor relating a tensor with its reciprocal.

TABLE IV

$$c^E{}_{ijpq} s^E{}_{pqkl} = I_{ijkl} \qquad d_{nkl} = \epsilon^T{}_{nm} g_{mkl} = e_{nij} s^E{}_{ijkl}$$
$$c^D{}_{ijpq} s^D{}_{pqkl} = I_{ijkl} \qquad e_{nkl} = \epsilon^S{}_{nm} h_{mkl} = d_{nij} c^E{}_{ijkl}$$
$$\beta^T{}_{mp} \epsilon^T{}_{pn} = I_{mn} \qquad g_{nkl} = \beta^T{}_{nm} d_{mkl} = h_{nij} s^D{}_{ijkl}$$
$$\beta^S{}_{mp} \epsilon^S{}_{pn} = I_{mn} \qquad h_{nkl} = \beta^S{}_{nm} e_{mkl} = g_{nij} c^D{}_{ijkl}$$
$$\epsilon^T{}_{mn} - \epsilon^S{}_{mn} = d_{nkl} e_{mkl}$$
$$\beta^S{}_{mn} - \beta^T{}_{mn} = h_{nkl} g_{mkl}$$
$$c^D{}_{ijkl} - c^E{}_{ijkl} = e_{mij} h_{mkl}$$
$$s^E{}_{ijkl} - s^D{}_{ijkl} = d_{mij} g_{mkl}.$$

To obtain explicit relations between the c_{ijkl} and the s_{ijkl} or the ϵ_{mn} and β_{mn} it is necessary to use the matrix relations given in Table III, Section 3.3.6.

3.3.8. Since, according to Section 3.2.3, the shearing strain components in the matrix notation (Section 3.3.1) are set equal to twice the tensor strain components (*e.g.*, $S_4 = 2S_{23}$) the numerical values of the tensor components of some of the constants are not equal to the corresponding components in the matrix notation of Section 3.3.3. The relations are given in Table V. The subscripts i, j, k, and l take the values 1, 2 or 3 in all the terms in the Table; p and q take the values given in the first column.

TABLE V

when $p, q = 1, 2, 3$			$s_{iijj} = s_{pq}$
$p = 1, 2, 3$, and $q = 4, 5, 6$			$s_{iijk} = s_{pq}/2$
$p, q = 4, 5, 6$			$s_{ijkl} = s_{pq}/4$
when	$p = 1, 2, 3$	$d_{ijj} = d_{ip}$	$g_{ijj} = g_{ip}$
	$p = 4, 5, 6$	$d_{ijk} = d_{ip}/2$	$g_{ijk} = g_{ip}/2$

The tensor components of c, e, and h are equal to the corresponding components in the matrix notation.

3.3.9. Units

It is proposed that numerical values of all the quantities involved in the piezoelectric relations be given in

[10] See footnote reference 2.

rationalized mks units. To aid in using these units, factors for converting from cgs electrostatic units to rationalized mks units are given in Table VI.

TABLE VI

CONVERSION FACTORS: *cgs* ELECTROSTATIC TO *mks* (RATIONALIZED) UNITS

The conversion factors are given as dimensionless expressions each equal to unity, except that for β and η the value is 4π, and for E and D, $1/4\pi$. They may be inserted as factors into an equation without changing its validity. Insertion of the appropriate conversion factors into an equation in which cgs electrostatic units are explicit converts the equation to a form which is explicit in mks rationalized units.

Quantity	Symbol	Conversion Factor	
Mechanical force	F	10^{-5}	newton per dyne
Elastic strain	S	1	numeric = relative deformation
Elastic stress	T	10^{-1}	newton/meter² per dyne/cm²
Elastic displacement	u	10^{-2}	meter per centimeter
Elastic compliance	s	10	meter²/newton per cm²/dyne
Elastic stiffness	c	10^{-1}	newton/meter² per dyne/cm²
Electric potential	V	300	volt per statvolt
Electric field	E	3×10^{4}	volt/meter per statvolt/cm
Electric charge	q	$\frac{1}{3}\times10^{-9}$	coulomb per statcoulomb
Electric displacement	D	$\dfrac{1}{12\pi\times10^{5}}$	coulomb/meter² per statcoulomb/cm²
Dielectric permittivity	ϵ	$\dfrac{1}{36\pi\times10^{9}}$	farad/meter per statfarad/cm
Dielectric impermeability	β	$36\pi\times10^{9}$	meter/farad per cm/statfarad
Dielectric constant, relative	K	1	numeric $=\epsilon/\epsilon_0$
Dielectric polarization	P	$\frac{1}{3}\times10^{-5}$	coulomb/meter² per statcoulomb/cm²
Dielectric susceptibility	η	4π (numeric)	units (mks) per unit (cgs)
Piezoelectric constant	d	$\frac{1}{3}\times10^{-4}$	coulomb/newton per statcoulomb/dyne
Piezoelectric constant	e	$\frac{1}{3}\times10^{-5}$	coulomb/meter² per statcoulomb/cm²
Piezoelectric constant	g	3×10^{5}	meter²/coulomb per cm²/statcoulomb
Piezoelectric constant	h	3×10^{4}	newton/coulomb per dyne/statcoulomb

Explanation of Table VI

The conversion factors for D, ϵ, β and η are expressions for either 4π or $1/4\pi$ instead of unity owing to the differences in definitions of these quantities in rationalized and unrationalized systems, and to the compensating appearance of 4π explicitly in the equations using these quantities in one system or the other.

The term "permittivity" is synonymous with "capacitivity" as defined in *Standards on Abbreviations, Graphic Symbols, Letter Symbols, and Mathematical Signs, 1948*. The word "permittivity" is adopted for this Standard because of its almost universal use in piezoelectric and dielectric theory.

In order to minimize the number of subscripts the well established symbols ϵ_0 and K are used in this Standard instead of the symbols ϵ_v and ϵ_r given in *Standards on Abbreviations*, etc., 1948 referred to above.

In rationalized mks units, polarization and susceptibility, as here defined, are related by the equation $P=\eta\epsilon_0 E$. This convention, though not universally adopted, has the advantage of assigning to P and η the same dimensions on both systems of units. It also conforms to good modern usage.[11]

[11] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., New York, N. Y., 1941.

3.3.10. Examples of the Use of Conversion Factors

The following equation is equally valid whether cgs or mks units are regarded as implicit along with the numerics in the several symbols:

$$E = -gT + \beta^T D.$$

Upon bracketing the symbols when they are intended to represent only the numerics, the equation becomes in cgs units (now shown explicitly)

$$[E]\,\frac{\text{statvolt}}{\text{cm}} = -[g]\,\frac{\text{cm}^2}{\text{statcoulomb}} \times [T]\,\frac{\text{dyne}}{\text{cm}^2} + [\beta^T]\,\frac{\text{cm}}{\text{statfarad}} \times [D]\,\frac{\text{statcoulomb}}{\text{cm}^2}.$$

Insert as unit factors, in parentheses, the several conversion factors given in Table VI:

$$[E]\,\frac{\text{statvolt}}{\text{cm}}\left(\frac{3\times10^4\ \text{volt/meter}}{\text{statvolt/cm}}\right) = -[g]\,\frac{\text{cm}^2}{\text{statcoulomb}}\left(\frac{3\times10^5\ \text{meter}^2/\text{coulomb}}{\text{cm}^2/\text{statcoulomb}}\right)$$
$$\times [T]\,\frac{\text{dyne}}{\text{cm}^2}\left(\frac{10^{-1}\ \text{newton/meter}^2}{\text{dyne/cm}^2}\right) + [\beta^T]\,\frac{\text{cm}}{\text{statfarad}}\left(\frac{36\pi\times10^9\ \text{meter/farad}}{\text{cm/statfarad}}\right)$$
$$\times [D]\,\frac{\text{statcoulomb}}{\text{cm}^2}\left(\frac{\text{coulomb/meter}^2}{12\pi\times10^5\ \text{statcoulomb/cm}^2}\right).$$

After cancelling, and dividing out the factor 3×10^4, the equation becomes

$$[E]\,\frac{\text{volt}}{\text{meter}} = -[g]\,\frac{\text{meter}^2}{\text{coulomb}} \times [T]\,\frac{\text{newton}}{\text{meter}^2} + [\beta^T]\,\frac{\text{meter}}{\text{farad}} \times [D]\,\frac{\text{coulomb}}{\text{meter}^2}.$$

and thus the original equation

$$E = -gT + \beta^T D$$

is seen to be equally valid whether the symbols imply cgs or mks units along with the numerics.

Following are a few numerical examples of conversion of units. They are for ammonium dihydrogen phosphate, based on Mason's measurements.[12]

	Electrostatic cgs units		Rationalized mks units	
d_{36}	1.55×10^{-6}	statcoulomb/dyne	5.17×10^{-11}	coulomb/newton
e_{36}	0.096×10^{6}	statcoulomb/cm²	0.319	coulomb/meter²
g_{36}	0.012×10^{-4}	cm²/statcoulomb	0.375	meter²/coulomb
h_{36}	0.088×10^{6}	dyne/statcoulomb	0.0263×10^{11}	newton/coulomb

[12] W. P. Mason, "The elastic, piezoelectric, and dielectric constants of potassium dihydrogen phosphate and ammonium dihydrogen phosphate," *Phys. Rev.*, vol. 69, pp. 173–194; March, 1946.

Response of Circuits to Steady-State Pulses*

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Summary—A method of calculating the steady-state response of circuits to repeated pulses is given using the method of the steady-state operational calculus. A short table of transforms which have been found useful in these calculations is also presented. The response of several basic circuits to these pulses is obtained and shown as calculated curves, and the calculated curves are then compared with curves obtained experimentally. These curves have been found to be very useful in adjusting circuits to be used with pulses. Several other possible applications are discussed.

Introduction

Much work has been done lately on the manner in which electrical circuits affect the shape of pulses.[1,2,3] MacLachlan[4] and others[5,6,7] have studied the response of circuits to the impulsive type of pulse, i.e., a pulse in which the energy has been transferred to the circuit before the circuit has had time to respond. Very little work on the response of circuits to steady-state pulses of the impulsive type has been done, however, and it is the purpose of this paper to examine this type of response.

Analysis

When considering the impulsive type of pulse, the main requirement to be satisfied is that the duration of the pulse be much shorter than any of the time constants or natural periods of oscillation of the circuits used. For example, a one-microsecond pulse width should be satisfactory for a circuit whose smallest time constant is of the order of ten microseconds, and whose highest natural frequency is of the order of 100 kc. If the above is satisfied, the shape of the particular pulse used should have very little effect on the results. The area E is a satisfactory measure of the strength of a pulse, and a unit pulse will be defined as one with unit area.

The pulse voltage wave form of Fig. 1(a) has the steady-state direct transform[8]

$$S(e) = \frac{E}{ap}(1 - \epsilon^{-pa}). \tag{1}$$

For very small values of a, the transform approaches the impulsive form

$$S(e) = E. \tag{2}$$

In applying the repeated pulses to circuits, a table of steady-state transforms was found very useful, and a short table of this kind is presented in Table I. The symbol Im_1 indicates an impulse of the first order and would be that of Fig. 1(a) with $E=1$ and a approaching zero. Similarly Im_2 indicates an impulse of the second order and is that of Fig. 1(b) with $E=1$ and a approaching zero. Figs. 1(c) and (d) show impulses of the third and fourth order, i.e., Im_3 and Im_4.

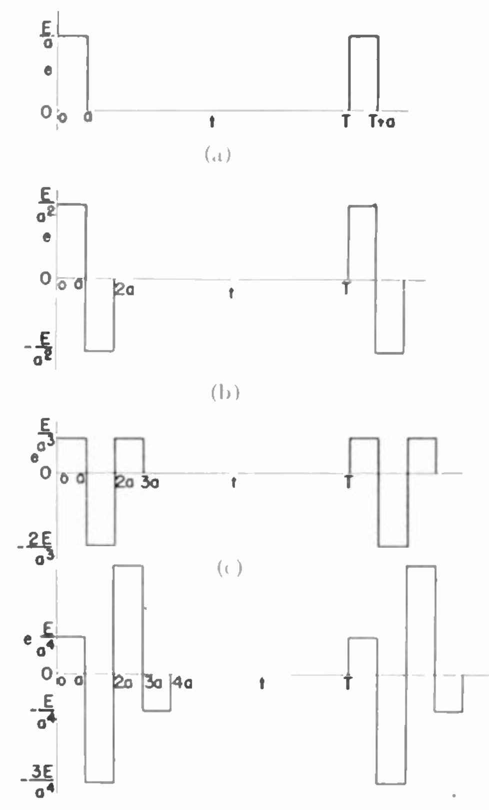

Fig. 1—Wave forms of the voltage pulses of orders (a) one, (b) two, (c) three, (d) four.

Response of Circuits

In determining the response of circuits to repeated pulses, the first circuit to be considered is that of a resistance R and a capacitance C in series. If the repeated pulses have the direct transform of (2), the direct transform of the current i flowing is

$$S(i) = \frac{E}{R}\frac{p}{p + \frac{1}{CR}}, \tag{3}$$

and that of the voltages across the resistance R and the capacitance C are

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[1] E. Frank, "Pulsed Linear Networks," McGraw-Hill Book Company, New York, N. Y., 1945.

[2] E. C. Cherry, "Pulse response," *Jour. I.E.E.*, pt. III, vol. 92, pp. 183–196; September, 1945.

[3] G. N. Glasoe, and J. V. Lebacqz, "Pulse Generators," McGraw-Hill Book Company, New York, N. Y., 1948.

[4] McLachlan, "Operational form of $f(t)$ for a finite interval with application to impulses," *Phil. Mag.*, vol. 26, Ser. 7/ pp. 695–704; November, 1938.

[5] M. F. Gardner, and J. L. Barnes, "Transients in Linear Systems," John Wiley and Sons, Inc., New York, N. Y., pp. 252–263; 1942.

[6] G. A. Campbell and R. M. Foster, "Fourier Integrals for Practical Applications," D. Van Nostrand Co., New York, N. Y., pp. 15–19; 1942.

[7] J. C. Jaeger, "Switching problems and instantaneous impulses," *Phil. Mag.*, vol. 36, Ser. 7, pp. 644–651; September, 1945.

[8] D. L. Waidelich, "Steady-state operational calculus," Proc. I.R.E., vol. 34, pp. 78–83; February, 1946.

TABLE I

STEADY-STATE TRANSFORM TABLE

Direct Transform $S[f(t)] = F(p)$	*Inverse Transform* $S^{-1}[F(p)] = f(t)$
1. 1	Im_1
2. p	Im_2
3. p^n	$Im_{(n+1)}$
4. $\dfrac{1}{p+a}$	$\dfrac{\epsilon^{-a}}{1-\epsilon^{-aT}}$
5. $\dfrac{p}{p+a}$	$-a\dfrac{\epsilon^{-at}}{1-\epsilon^{-aT}} + Im_1$
6. $\dfrac{1}{p^2+a^2}$	$\dfrac{T}{2m}\,\dfrac{\cos m\left(\dfrac{t}{T}-\dfrac{1}{2}\right)}{\sin(m/2)}$ where $m = aT$
7. $\dfrac{p}{p^2+a^2}$	$-\dfrac{\sin m\left(\dfrac{t}{T}-\dfrac{1}{2}\right)}{2\sin(m/2)}$ where $m = aT$
8. $\dfrac{p^2}{p^2+a^2}$	$-\dfrac{m}{2T}\,\dfrac{\cos m\left(\dfrac{t}{T}-\dfrac{1}{2}\right)}{\sin(m/2)} + Im_1$ where $m = aT$
9. $\dfrac{1}{(p+a)^2}$	$\dfrac{\epsilon^{-a(t-T)}[t-\epsilon^{-aT}(t-T)]}{2(\cosh aT - 1)}$
10. $\dfrac{p}{(p+a)^2}$	$\dfrac{\epsilon^{-a(t-T)}}{2(\cosh aT - 1)}[(1-at)(1-\epsilon^{-aT}) - aT\epsilon^{-aT}]$
11. $\dfrac{p^2}{(p+a)^2}$	$\dfrac{\epsilon^{-a(t-T)}}{2(\cosh aT - 1)}[a(at-2)(1-\epsilon^{-aT}) + a^2T\epsilon^{-aT}] + Im_1$
12. $\dfrac{1}{(p+\alpha_1)(p+\alpha_2)} = \dfrac{1}{(p+a)^2-b^2} = \dfrac{1}{(p+a)^2+\beta^2}$ $\alpha_1 = a+b = a+j\beta$ $\alpha_2 = a-b = a-j\beta$ $b = j\beta \neq 0$	$\dfrac{1}{(\alpha_2-\alpha_1)}\left[\dfrac{\epsilon^{-\alpha_1 t}}{1-\epsilon^{-\alpha_1 T}} - \dfrac{\epsilon^{-\alpha_2 t}}{1-\epsilon^{-\alpha_2 T}}\right]$ $= \dfrac{\epsilon^{-a(t-T)}}{2b}\left[\dfrac{\sinh bt - \epsilon^{-aT}\sinh b(t-T)}{\cosh aT - \cosh bT}\right]$ $= \dfrac{\epsilon^{-a(t-T)}}{2\beta}\left[\dfrac{\sin\beta t - \epsilon^{-aT}\sin\beta(t-T)}{\cosh aT - \cos\beta T}\right]$
13. $\dfrac{p}{(p+\alpha_1)(p+\alpha_2)} = \dfrac{p}{(p+a)^2-b^2} = \dfrac{p}{(p+a)^2+\beta^2}$ $\alpha_1 = a+b = a+j\beta$ $\alpha_2 = a-b = a-j\beta$ $b = j\beta \neq 0$ when $b > a$, $\phi = \tanh^{-1}\dfrac{a}{b}$ when $a > b$, $\psi = \tanh^{-1}\dfrac{b}{a}$ $\sigma = \tan^{-1}\dfrac{a}{\beta}$	$\dfrac{1}{(\alpha_2-\alpha_1)}\left[\dfrac{-\alpha_1\epsilon^{-\alpha_1 t}}{1-\epsilon^{-\alpha_1 T}} + \dfrac{\alpha_2\epsilon^{-\alpha_2 t}}{1-\epsilon^{-\alpha_2 T}}\right]$ $= \dfrac{\epsilon^{-a(t-T)}\sqrt{b^2-a^2}}{2b}\left\{\dfrac{\cosh(bt-\phi) - \epsilon^{-aT}\cosh[b(t-T)-\phi]}{\cosh aT - \cosh bT}\right\}$ $= \dfrac{\epsilon^{-a(t-T)}\sqrt{a^2-b^2}}{2b}\left\{\dfrac{-\sinh(bt-\psi) + \epsilon^{-aT}\sinh[b(t-T)-\psi]}{\cosh aT - \cosh bT}\right\}$ $= \dfrac{\epsilon^{-a(t-T)}\sqrt{a^2+\beta^2}}{2\beta}\left\{\dfrac{\cos(\beta t+\sigma) - \epsilon^{-aT}\cos[\beta(t-T)+\sigma]}{\cosh aT - \cosh bT}\right\}$

TABLE I—(*continued*)

	Direct Transform $S[f(t)] = F(p)$	*Inverse Transform* $S^{-1}[F(p)] = f(t)$
14.	$\frac{p^2}{(p+\alpha_1)(p+\alpha_2)} = \frac{p^2}{(p+a)-b^2} = \frac{p^2}{(p+a)^2+\beta^2}$ $\alpha_1 = a + b = a + j\beta$ $\alpha_2 = a - b = a - j\beta$ $b = j\beta \neq 0$ $\rho = \tanh^{-1}\left(\frac{2ab}{a^2+b^2}\right)$ $\delta = \tan^{-1}\left(\frac{2a\beta}{a^2-\beta^2}\right)$	$\frac{1}{(\alpha_2-\alpha_1)}\left[\frac{\alpha_1^2\epsilon^{-\alpha_1 t}}{1-\epsilon^{-\alpha_1 T}} - \frac{\alpha_2^2\epsilon^{-\alpha_2 t}}{1-\epsilon^{-\alpha_2 T}}\right] + Im_1$ $= \frac{\epsilon^{-a(t-T)}\lvert a^2-b^2\rvert}{2b}\left\{\frac{\sinh(bt-\rho) - \epsilon^{-aT}\sinh[b(t-T)-\rho]}{\cosh aT - \cosh bT}\right\} + Im_1$ $= \frac{\epsilon^{-a(t-T)}(a^2+\beta^2)}{2\beta}\left\{\frac{\sin(\beta t-\delta) - \epsilon^{-aT}\sin[\beta(t-T)-\delta]}{\cosh aT - \cos\beta T}\right\} + Im_1$

$$S(e_R) = E\frac{p}{p + \frac{1}{CR}} \tag{4}$$

and

$$S(e_C) = \frac{E}{RC}\,\frac{1}{p + \frac{1}{CR}}. \tag{5}$$

By the use of transforms 4 and 5 of Table I and putting $\theta = (RC/T)$ and $\tau = (t/T)$, where $0 < \tau < 1$,

$$i = \frac{E}{R}\left[-\frac{\epsilon^{-\tau/\theta}}{\theta T(1-\epsilon^{-1/\theta})} + Im_1\right]. \tag{6}$$

$$e_R = E\left[-\frac{\epsilon^{-\tau/\theta}}{\theta T(1-\epsilon^{-1/\theta})} + Im_1\right]. \tag{7}$$

and

$$e_C = \frac{E}{RC}\,\frac{\epsilon^{-\tau/\theta}}{1-\epsilon^{-1/\theta}}. \tag{8}$$

The current i and the voltage across the resistance e_R have the same shape, and this shape is shown in Fig. 2 for various values of θ. The parameter θ is equal to the time constant of the circuit divided by the period of the applied pulses, so a larger time constant in the circuit is indicated by a larger value of θ for a given pulse period. The pulses of Fig. 2 are shown with a finite height which is as they would appear in a practical case. For low values of θ, the circuit approaches a differentiating circuit, and the wave form of Fig. 2 approaches the derivative of the impulse wave form which is an impulse of the second order. Experimental confirmation of the wave form for $\theta = 1$ is shown in Fig. 3, but only the tops of the pulses are visible in the photograph. The wave forms of the voltage e_C across the capacitance are shown in Fig. 4, and it should be noted that no pulses appear in this case. For large values of θ, the circuit approaches an integrating circuit, and the wave form of Fig. 4 ap-

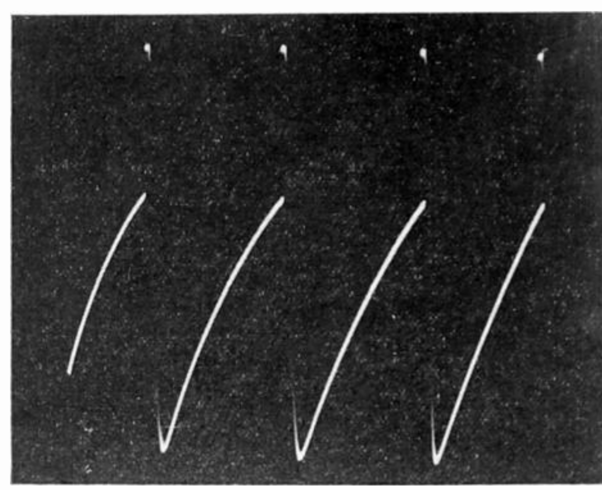

Fig. 3—Experimental wave form for $e_R(\theta = 1.0)$ in a series RC circuit.

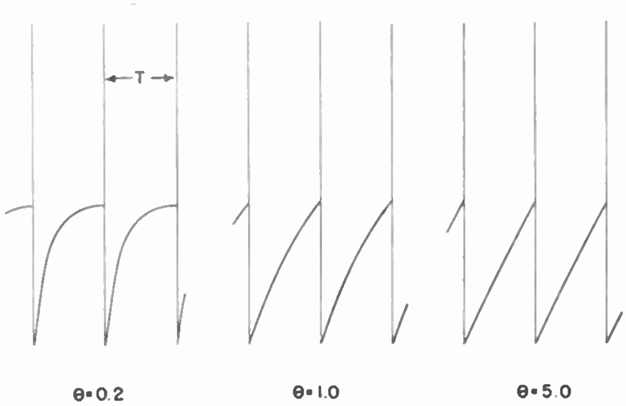

Fig. 2—Calculated wave forms for e_R in a series RC circuit.

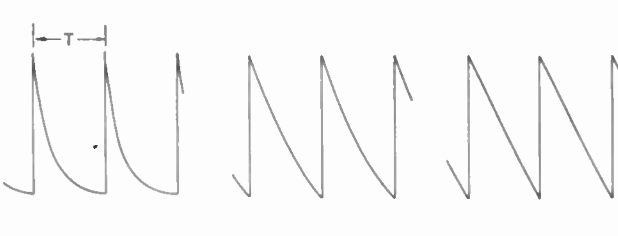

Fig. 4—Calculated wave forms for e_C in a series RC circuit.

proaches the integral of the pulse wave minus its average value, resulting in the saw-tooth wave shown. Again, the experimental wave form for $\theta=1.0$ is shown in Fig. 5.

A number of other circuits have responses which are the same as those of a series resistance-capacitance circuit. Among these are series resistance-inductance, parallel resistance-capacitance, and parallel resistance-inductance circuits. The applied pulses for the parallel circuits are current pulses.

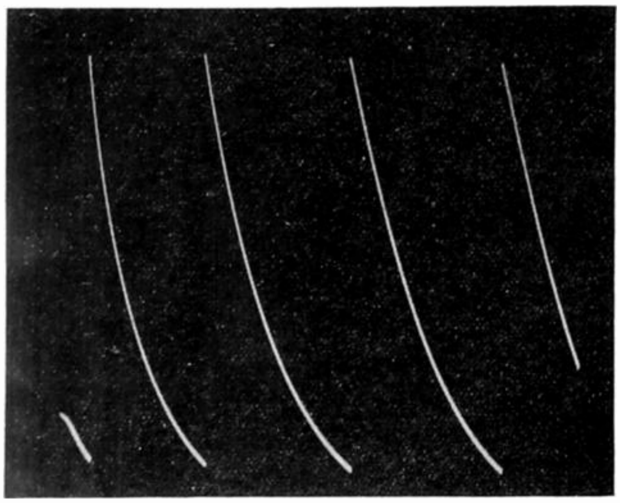

Fig. 5—Experimental wave form for $e_C(\theta=1.0)$ in a series RC circuit.

Another circuit that is very commonly used is that of an inductance in series with a capacitance. The resistance is assumed zero, although it will be considered different from zero later on. The direct transforms of the current i, the voltage e_L across the inductance, and the voltage e_C across the capacitance are

$$S(i) = \frac{E}{L}\,\frac{p}{p^2+\dfrac{1}{LC}} \tag{9}$$

$$S(e_L) = E\,\frac{p^2}{p^2+\dfrac{1}{LC}} \tag{10}$$

and

$$S(e_C) = \frac{E}{LC}\,\frac{1}{p^2+\dfrac{1}{LC}}\,. \tag{11}$$

By the use of transforms 6, 7, and 8 of Table I and putting $m=(T/\sqrt{LC})$ and $\tau=(t/T)$ where $0<\tau<1$,

$$i = -\frac{E}{2L}\,\frac{\sin m(\tau-1/2)}{\sin(m/2)}, \tag{12}$$

$$e_L = E\left[-\frac{m\cos m(\tau-1/2)}{2T\sin(m/2)} + Im_1\right], \tag{13}$$

and

$$e_C = \frac{ET\cos m(\tau-1/2)}{2m\sin(m/2)}\,. \tag{14}$$

The current i is shown in Fig. 6 for various values of m, and it should be noticed that $(m/2\pi)$ is the number of cycles of the natural frequency of oscillation which occur during one period of the applied pulses. When $m=2n\pi$ where n is a positive integer, the circuit is resonant at the nth harmonic of the applied pulse wave, and since the resistance in the circuit is assumed zero,

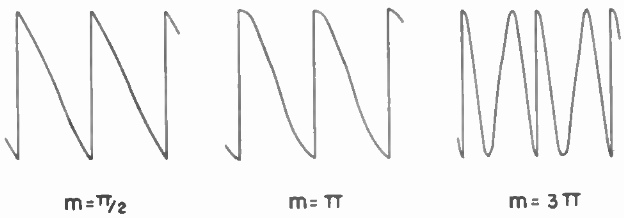

Fig. 6—Calculated wave forms for the current in a series LC circuit.

the response becomes infinitely large. When m becomes very small, the current wave form is the integral of the applied pulse wave form and is similar to that of Fig. 4 for the case of θ very large. The corresponding wave forms for the inductance voltage e_L are shown in Fig. 7,

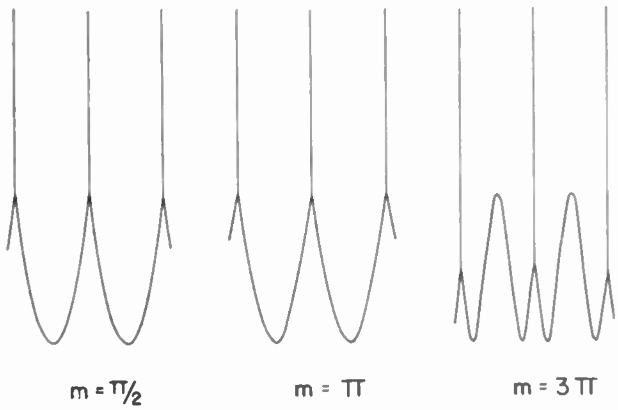

Fig. 7—Calculated wave forms for e_L in a series LC circuit.

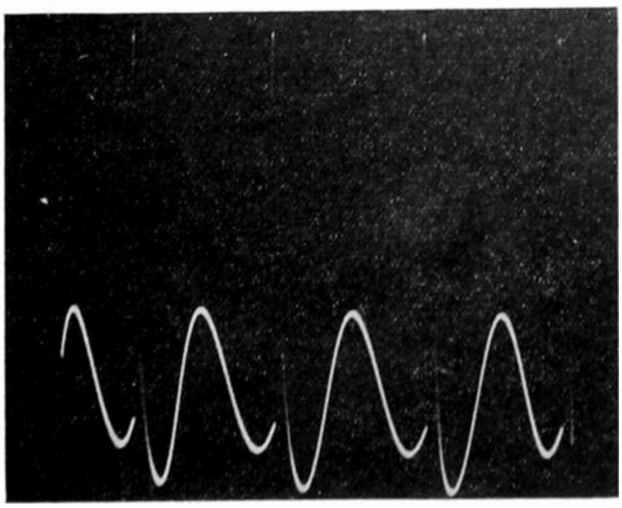

Fig. 8—Experimental wave form for $e_L(m=3\pi)$ in a series LC circuit.

and the experimental wave for $m=3\pi$ is given in Fig. 8. The effect of resistance in the circuit may be noticed in that the peaks of the natural oscillation of the circuit voltage decrease in height with time. Similar results for the capacitive voltage e_C are shown in Fig. 9. For very low values of m the capacitive voltage approaches the second integral of the applied pulse voltage, which is a parabola. The paralleled inductance and capacitance circuit with a current-pulse wave form applied to it has wave forms exactly similar to those discussed above for the series inductance-capacitance circuit.

Fig. 9—Calculated wave forms for e_C in a series LC circuit.

When the resistance is not negligibly small in a series resistance-inductance-capacitance circuit, the response differs largely, depending upon whether the resistance is small or large. From transforms 12, 13, and 14 of Table I and with

$$\tau = (t/T), \qquad 0 < \tau < 1,$$

$$\alpha_1 = \left[\frac{R}{2L} + \sqrt{\left(\frac{R}{2L}\right)^2 - \frac{1}{LC}}\right]T, \quad \text{and}$$

$$\alpha_2 = \left[\frac{R}{2L} - \sqrt{\left(\frac{R}{2L}\right)^2 - \frac{1}{LC}}\right]T,$$

$$e_R = Ri = \frac{RE}{L(\alpha_2 - \alpha_1)}\left[\frac{-\alpha_1\epsilon^{-\alpha_1\tau}}{1-\epsilon^{\alpha_1}} + \frac{\alpha_2\epsilon^{-\alpha_2\tau}}{1-\epsilon^{\alpha_2}}\right], \tag{15}$$

$$e_L = \frac{E}{T(\alpha_2 - \alpha_1)}\left[\frac{\alpha_1^2\epsilon^{-\alpha_1\tau}}{1-\epsilon^{-\alpha_1}} - \frac{\alpha_2^2\epsilon^{-\alpha_2\tau}}{1-\epsilon^{-\alpha_2}}\right] + \frac{EIm_1}{T}, \tag{16}$$

$$e_C = \frac{ET}{LC(\alpha_2 - \alpha_1)}\left[\frac{\epsilon^{-\alpha_1\tau}}{1-\epsilon^{-\alpha_1}} - \frac{\epsilon^{-\alpha_2\tau}}{1-\epsilon^{-\alpha_2}}\right]. \tag{17}$$

This is the nonoscillatory case $(R>2\sqrt{L/C})$, and a typical set of response curves are shown in Fig. 10 for the parameters $\alpha_1=2.0$ and $\alpha_2=1.0$. The shapes of the curves for e_R and e_L resemble those of a series resistance-inductance circuit, as evidenced in Figs. 2 and 4. The voltage e_C across the capacitance, on the other hand, resembles the capacitive voltage of a series inductance-capacitance circuit as given in Fig. 9.

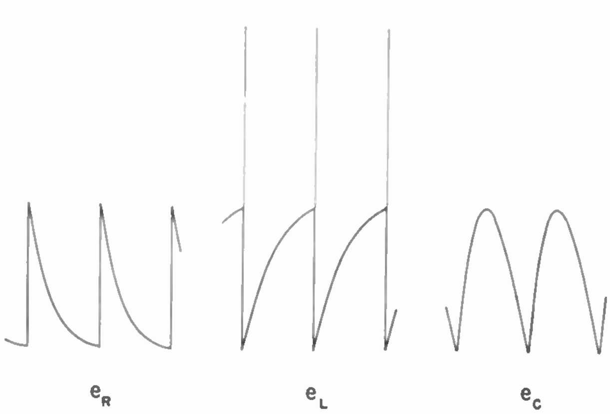

Fig. 10—Calculated wave forms for e_R, e_L, and e_C in a series RLC circuit with $R>2\sqrt{L/C}$.

The oscillatory case $(R<2\sqrt{L/C})$ is obtained also from transforms 12, 13, and 14 of Table I. If $\tau=(t/T)$, $0<\tau<1$,

$$m = T\sqrt{\frac{1}{LC} - \left(\frac{R}{2L}\right)^2}, \qquad n = (TR/2L),$$

$$\sigma = \tan^{-1}(n/m), \quad \text{and} \quad \delta = \tan^{-1}\left(\frac{2mn}{n^2 - m^2}\right),$$

$$e_R = Ri = \frac{RE\epsilon^{-n(\tau-1)}\sqrt{m^2+n^2}}{L2m(\cosh n - \cos m)}\{\cos(m\tau + \sigma) - \epsilon^{-n}\sin[m(\tau-1)+\sigma]\}, \tag{18}$$

$$e_L = \frac{E\epsilon^{-n(\tau-1)}(m^2+n^2)}{2mT(\cosh n - \cos m)}\{\sin(m\tau - \delta) - \epsilon^{-n}\sin[m(\tau-1)-\delta]\} + \frac{EIm_1}{T}, \tag{19}$$

$$e_C = \frac{ET\epsilon^{-n(\tau-1)}}{2LCm(\cosh n - \cos m)}\{\sin m\tau - \epsilon^{-n}\sin m(\tau-1)\}. \tag{20}$$

A typical set of response curves for the case $m=3\pi$, $n=1.0$ is shown in Fig. 11. It should be noticed that the curves are very similar to those of Figs. 6, 7, and 9, except that the effect of the additional resistance is to damp out the oscillations. The responses of many other circuits may be obtained by using linear combinations of the transforms given in Table I.

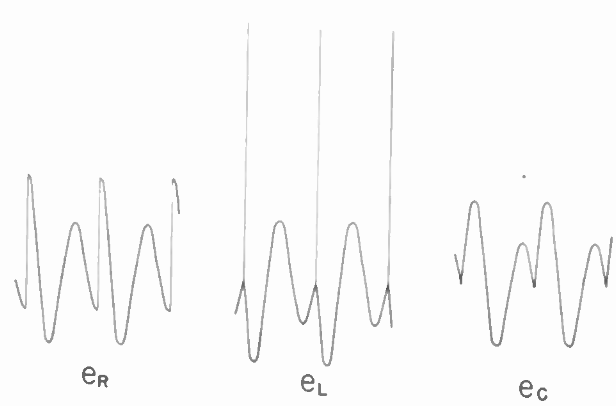

Fig. 11—Calculated wave forms for e_R, e_L, and e_C in a series RLC circuit with $R<2\sqrt{L/C}$.

In applying the above wave forms to the testing of amplifiers,[9,10] the pulse-generator output is applied to

[9] L. B. Arguimbau, "Network testing with square waves," *Gen. Rad. Exp.*, vol. 14, pp. 1–6; December, 1939.
[10] D. L. Waidelich, "Steady-state testing with saw-tooth waves," Proc. I.R.E., vol. 32, pp. 339–348; June, 1944.

the amplifier, and the output of the amplifier is viewed on an oscilloscope screen. The frequency of the pulse generator may be varied to determine the width of the pass band of the amplifier and the type of distortion present at the edges of the pass band. For example, a simple resistance-capacitance coupled amplifier at low frequencies acts much like a series resistance-capacitance circuit with the output voltage appearing across the resistance. The resulting wave forms are similar then to those of Fig. 2, and the approximate lower half-power frequency would be that when the wave form for $\theta=(1/2\pi)$ of Fig. 12(a) would appear. The upper half-power frequency may be obtained in a similar manner, and the wave form to be used is that of Fig. 12(b) for $\theta=2\pi$. Resonance occurring in the amplifier may be found by noticing for what frequencies of the pulse generator the output voltage becomes very large. The output voltage wave forms should be similar to those of Fig. 9. When the sinusoidal or nearly sinusoidal wave form is that for $m=2\pi$, the frequency of the pulse generator is the same as the resonant frequency of the amplifier.

It has been shown[10] that if the steady-state response of an amplifier is known to a saw-tooth wave of period T, the steady-state response to any nonsinusoidal wave of the same period T may be calculated. The same is true if a pulse wave is used. For example, if $e_0(t)$ is the steady-state output response voltage of a pulse of period T applied to an amplifier, and if $e(t)$ is any other non-

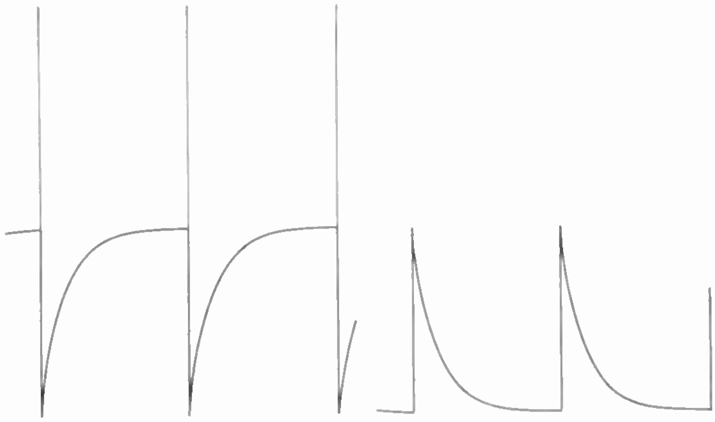

Fig. 12—Calculated wave forms for (a) the lower, (b) the upper half-power frequencies.

sinusoidal wave of period T, then the steady-state response e_s of the amplifier to $e(t)$ is

$$e_s = \int_0^T e(t-\tau)e_0(\tau)d\tau, \tag{21}$$

or another equivalent form is

$$e_s = \int_{t-T}^{t} e(\tau)e_0(t-\tau)d\tau. \tag{22}$$

If the equations for e and e_0 are known, it is possible to integrate (21) or (22). If equations are not known for either e or e_0 or both, a numerical solution is still possible as outlined in the previous reference.[10]

Diode Phase-Discriminators*

R. H. DISHINGTON†

Summary—**Two sinusoidal phase-discriminators are analyzed and it is found that universal curves of their general phase characteristics can be plotted as a function of two parameters. From these curves it is concluded that the resistances in series with the tubes and also the tube resistances themselves are the most important factors in determining optimum performance.**

Introduction

The phase-discriminator, otherwise known as phase-comparator or phase-detector gives a measurement of the phase difference between two waves. Diode discriminators, having the advantage of simplicity, indicate the phase angle by a voltage at the output terminals. At present, the principles of operation are well known, but there is a noticeable lack of an accurate analysis of the circuits.[1,2] The present paper deals with the problem by applying a recently introduced general method of diode circuit analysis.[3] For all practical purposes, this gives an exact solution. The circuits' general characteristics are given graphically, and only a simple calculation must be made to obtain the complete phase-characteristic for any practical values of the circuit parameters.

The Basic Method

In footnote reference 3 it was shown that a tube and any series resistance R_s have a *combination characteristic*

$$i_b = Ke_d^{\alpha_c} \tag{1}$$

where i_b is the plate current, e_d is the voltage across both the tube and R_s, and K and α_c are constants. Mathematically,

$$e_d = e_b + i_bR_s \tag{2}$$

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[1] W. L. Emery, "Ultra-High-Frequency Radio Engineering," Macmillan Co., New York, N. Y., 1944; p. 41.

[2] L. I. Farren, "Phase detectors, some theoretical and practical aspects," *Wireless Eng.*, vol. 23, pp. 330–340; December, 1946.

[3] R. H. Dishington, "Diode circuit analysis," *Elec. Eng.*, vol. 67, pp. 1043–1049; November, 1948.

where e_b is the plate voltage of the tube. To use the solutions presented further on, two quantities must be computed. First, referring to Figs. 1 and 5,

$$i_2 = \frac{e_{1\max}}{R_{bb}} \tag{3}$$

and, from (2),

$$E_{21} = e_b\rbrack_{i_2} + i_2 R_s \tag{4}$$

where $e_b\rbrack_{i_2}$ is the plate voltage of the tube at i_2, taken directly off the static plate characteristic. Second, the exponent α_c can be found very simply, as explained in footnote reference 3.

The Simple Sinusoidal Phase-Discriminator

Phase difference between two sinusoidal waves can be measured by the circuit in Fig. 1. The magnitudes of the *open circuit* input voltages e_x and e_y are assumed

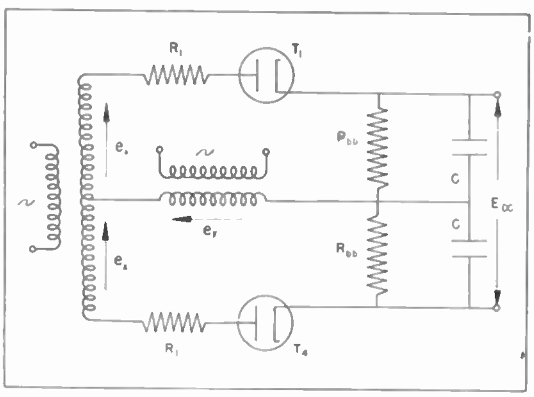

Fig. 1—The simple phase-discriminator.

equal. Both sides of the circuit are identical except for the net voltage applied to each. The driving voltages for T_1 and T_4 respectively are

$$\left.\begin{aligned} e_1 &= e_y + e_x \\ e_4 &= e_y - e_x \end{aligned}\right\}. \tag{5}$$

Adding a fixed lead of $\pi/2$ to e_y, to resolve the ambiguity in ϕ for positive and negative angles,

$$\left.\begin{aligned} e_1 &= E \sin\left(\omega t + \phi + \frac{\pi}{2}\right) + E \sin \omega t \\ e_4 &= E \sin\left(\omega t + \phi + \frac{\pi}{2}\right) - E \sin \omega t \end{aligned}\right\} \tag{6}$$

Transforming (6),

$$\left.\begin{aligned} e_1 &= (e_{1\max}) \sin\left(\omega t + \phi + \frac{\pi}{4}\right) \\ e_4 &= (e_{4\max}) \cos\left(\omega t + \phi + \frac{\pi}{4}\right) \end{aligned}\right\} \tag{7}$$

where

$$\left.\begin{aligned} e_{1\max} &= 2E \cos\left(\frac{\phi}{2} + \frac{\pi}{4}\right) \\ e_{4\max} &= 2E \sin\left(\frac{\phi}{2} + \frac{\pi}{4}\right) \end{aligned}\right\}. \tag{8}$$

The peak values of e_1 and e_4 are functions of ϕ, but not of time.

Equation (7) reveals that the voltages applied to the two opposite sides of the circuit are always 90° out of phase. This means that, except for a short period of overlap, one tube conducts while the other does not. Little error is introduced if both halves are assumed completely independent. Once this assumption is made, reduction to the equivalent circuit is simple, each half of the circuit being reduced separately. The calculated output of T_4 is then subtracted from that of T_1 to give E_{DC} (Fig. 1). Completely general curves for the solution are shown in Figs. 2, 3, and 4. To use the curves, it is necessary to evaluate R_s. In the present circuit, R_s is the sum of the internal resistances of e_x and e_y plus R_1. The curves are plotted for various values of the ratio $(E_{21}/e_{1\max})_{-90°}$ at $\phi = -90°$. Actually $E_{21}/e_{1\max}$ changes with ϕ. A correction for this is used to obtain the solution. The results give E_{DC}/E for negative values of ϕ, but the positive angles give the same shape of characteristic with negative voltage.

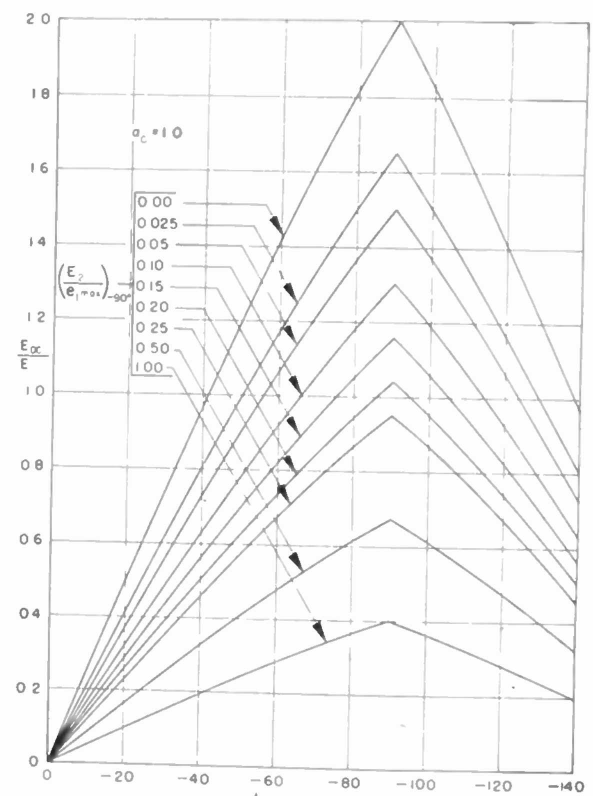

Fig. 2—General phase characteristics of the simple discriminator with sinusoidal input when $\alpha_c = 1.0$. To find $(E_{21}/e_{1\max})$ use the one value $e_{1\max} = 2E$.

The sensitivity of phase measurement for any given value of α_c is a function of $E_{21}/e_{1\max}$, which can be expressed

$$\frac{E_{21}}{e_{1\max}} = \frac{1}{R_{bb}} \frac{e_b}{i_2}\bigg\rbrack_{i_2} + \frac{R_s}{R_{bb}}. \tag{9}$$

Equation (9) makes it apparent that large values of R_{bb} and small values of R_s tend to lower $E_{21}/e_{1\max}$ and thereby increase the sensitivity. The quantity $e_b/i_2\rbrack_{i_2}$ is of the order of magnitude of R_{T_1}, so a low R_{T_1} also in-

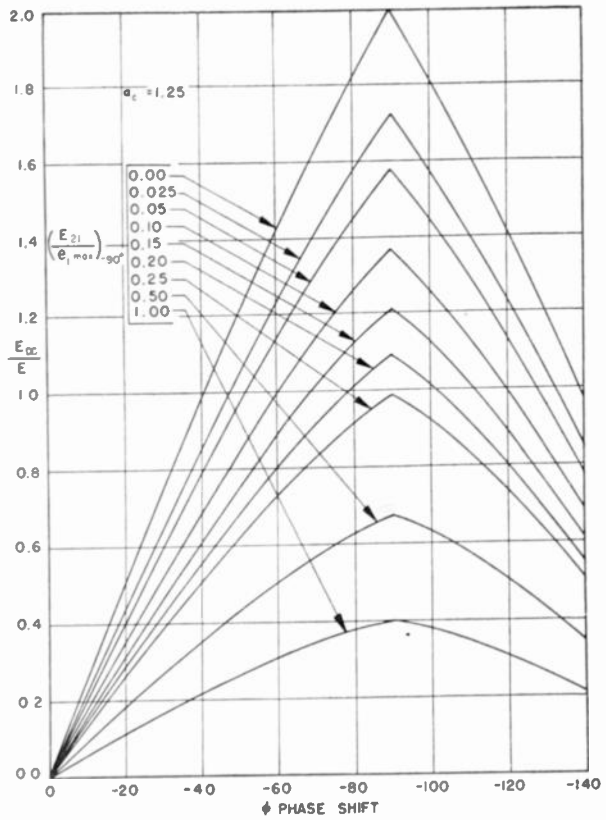

Fig. 3—General phase characteristics of the simple discriminator with sinusoidal input when $\alpha_c = 1.25$. To find $(E_{21}/e_{1\,max})$ use the one value $e_{1\,max} = 2E$.

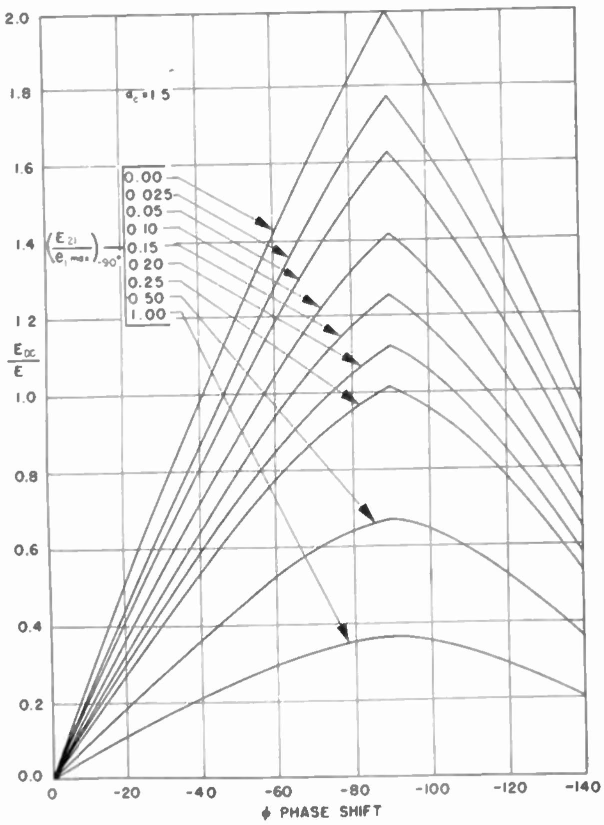

Fig. 4—General phase characteristics of the simple discriminator with sinusoidal input when $\alpha_c = 1.5$. To find $(E_{21}/e_{1\,max})$ use the one value $e_{1\,max} = 2E$.

creases the sensitivity. (For values of R_{T_2} see footnote reference 3.)

The linearity is better for values of α_c near 1.0. However, unless the tube α_s is originally near unity, α_c can only be made linear by adding R_1. From the foregoing, this increases R_s and decreases the sensitivity. For high sensitivity, the difference in nonlinearity of the output between $\alpha_c = 1.0$ and $\alpha_c = 1.5$ is very small. Therefore, an optimum design will have no R_1, making R_s as small as possible.

The Balanced Sinusoidal Phase Discriminator

Another well-known comparator is the balanced circuit shown in Fig. 5. The tubes and resistors R_1 are the same for each branch. Both RC loads are also similar. Given the same conditions for e_x and e_y, the driving

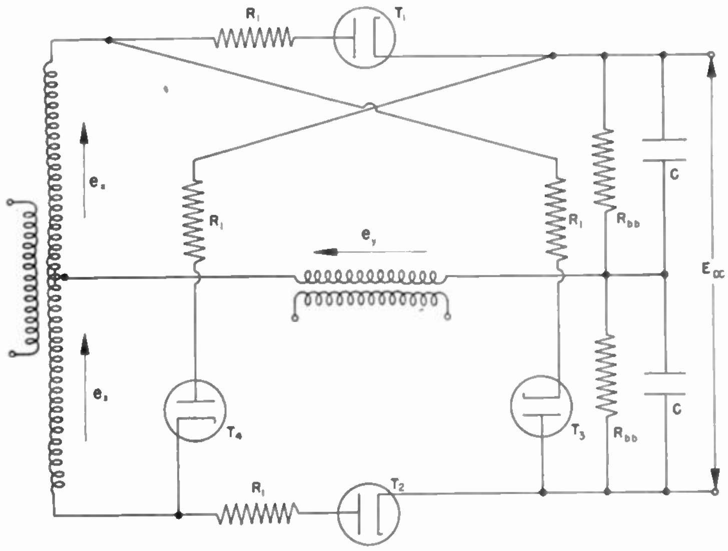

Fig. 5—The balanced phase-discriminator.

voltages for tubes T_1, T_2, T_3, and T_4 are e_1, e_4, $-e_1$, and $-e_4$ respectively. Again, except for a slight overlap, each tube conducts when the other three do not. Consequently, it is assumed that the two halves of the circuit are separable. It is conventional to show E_{bb} as positive with respect to the reference diode plate. Tube T_1 is chosen as the reference for the top half, and inasmuch as the constant output voltage is actually produced across an RC load, E_{bb} will be negative for negative phase angles. For the same conditions, E_{bb} tends to make the plate of T_4 positive. For this reason, it is important in the derivation to remember that for negative ϕ, T_1 operates class C and T_4 operates class AB or A. The second half of the circuit produces an output indentical to the first and in series with it. Thus, the two output voltages are added to give the total E_{DC}. The final solution for sinusoidal input voltages is given in Figs. 6, 7, and 8. Remarks on how to calculate $(E_{21}/e_{1max})_{-90°}$ are exactly the same for this circuit as for the simple comparator. Also the effects of the various resistors on the sensitivity are the same as before. Examining the curves, it appears that unless the flat-topped phase characteristic of Fig. 6 is desired for some particular reason, better sensitivity with more over-all performance is obtained with operation as near to $\alpha_c = 1.5$ as possible. This means less R_1; but a precaution is necessary here. Originally, it was assumed that E_{bb} had a constant value for each ϕ. This is made

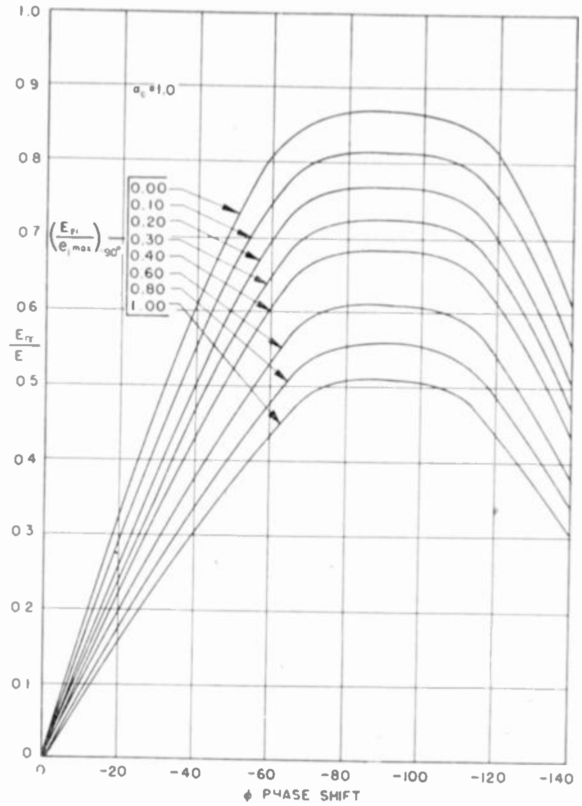

Fig. 6—General phase characteristics of the balanced discriminator with sinusoidal input when $\alpha_c = 1.0$. To find $(E_{21}/e_{1\,max})$ use the one value $e_{1\,max} = 2E$.

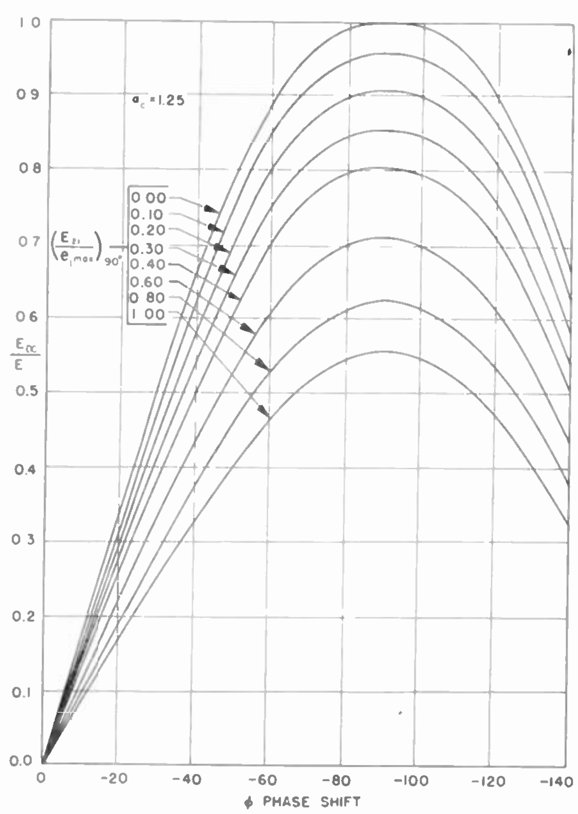

Fig. 7—General phase characteristics of the balanced discriminator with sinusoidal input when $\alpha_c = 1.25$. To find $(E_{21}/e_{1\,max})$ use the one value $e_{1\,max} = 2E$.

possible by a large enough time constant $R_{bb}C$. Now, however, the capacitor can discharge through T_4 for example, and unless $(R_s + R_t)C$ is large, E_{bb} may not remain constant. This generally means that some R_1 must be added.

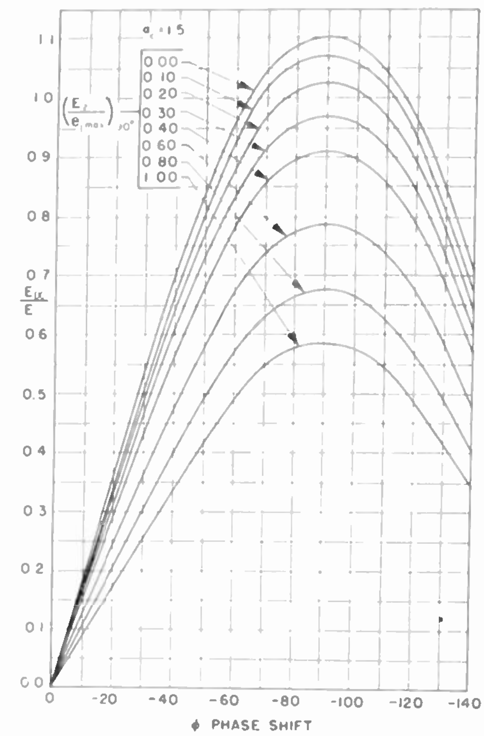

Fig. 8—General phase characteristics of the balanced discriminator with sinusoidal input when $\alpha_c = 1.5$. To find $(E_{21}/e_{1\,max})$ use the one value $e_{1\,max} = 2E$.

Conclusions

The total phase characteristics of two basic types of phase-discriminators are given in a form which enables quick calculation of the proper performance curve. Only two assumptions are made; one, that the ripple across the load is negligibly small; and two, that each tube conducts when the others are nonconducting. The first assumption is easily justified, and the second introduces only a minute error in practical cases. It appears that, in both circuits, the sensitivity is increased by large R_{bb}, and small R_1 and tube resistance. However, the value of R_1 must be large enough in the balanced circuit to ensure the constancy of E_{bb} by giving a large time constant $(R_s + R_t)C$. The slight increase in linearity, over only a part of the range, which is gained by adding R_1 is more than offset by the undesirable loss of sensitivity.

The balanced circuit seems to be less desirable than the simple one, but there is one important feature to consider. The output of the simple circuit is the difference between two large voltages. This gives inaccurate operation for small phase angles in a practical circuit where tubes and resistors are not perfectly matched. To its advantage, the balanced circuit output is the sum of two large voltages and this tends to reduce the effect of an error in either.

Acknowledgment

The author wishes to acknowledge the aid of R. C. Dishington and the Rand Corporation in producing the illustrations, and of F. Forbath in checking the manuscript.

Loading and Coupling Effects of Standing-Wave Detectors*

K. TOMIYASU†, MEMBER, IRE

Summary—**When measuring impedances on transmission lines, insensitive standing-wave detectors have the effect of yielding lower standing-wave ratios than the true values. Double-hump distribution curves are shown to be the result of very tight coupling of the detector. Detectors than can be represented by a susceptance component may indicate unsymmetrical distribution curves. Sensitive detectors used on transmission lines having low power levels can introduce tight coupling effects. Conditions are given for a loosely coupled detector.**

Introduction

Impedances at ultra-high frequencies, and higher, are usually measured indirectly either from the shape of a standing-wave distribution or from the resonance-curve distribution[1] using a sensitive detector. The presence of a detector disturbs the electromagnetic field it measures and the pointer readings are subject to corrections. It is desirable to know how much a detector disturbs the field and how to eliminate or correct for the errors involved.

Altar, Marshall, and Hunter,[2] state that, for deep probe penetrations in waveguides, "... the appearance of more than one maximum per half-cycle ... is clearly unaccounted for by the shunt-admittance theory. ..." Dowker and Redheffer[3] also report unusual effects when deep probe penetrations are used. Since both investigators made use of voltage probes, the effects are somewhat obscured because greater coupling, and larger susceptance effects occur simultaneously. These effects can be separated by using a current probe.

In the literature an interesting discussion of detectors is found,[4] together with a partial treatment of their effects on a line having a matched generator. However, the important case of a loosely coupled generator and the unusual effects of tightly coupled detectors have not been considered heretofore.

Equivalent Detector Admittance

A voltage detector that is tuned to resonance can be represented by a pure conductance, whereas if it is significantly detuned or has a large shunting capacitance, it can be represented by a pure susceptance.

The loading and coupling effects of standing-wave detectors depend upon the combination of three admittances: generator, detector, and load. When the generator is loosely coupled to the line, the effects of tightly coupled detectors are much more conspicuous than when the generator is matched. Usually detectors are tuned to resonance in order to obtain the maximum sensitivity, but occasionally detuned detectors are used—for instance, in broad-band applications. In order to analyze the most useful combinations of generator and detector admittances, the loosely coupled generator and tuned detector are considered in Part I, the matched generator and tuned detector in Part II, and the detector represented by a pure susceptance in Part III.

I. Loosely Coupled Generator and Tuned Detector

The circuit under consideration is shown in Fig. 1. A

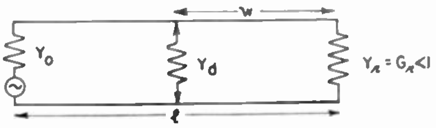

Fig. 1—Schematic circuit diagram of the transmission line with a detector in shunt.

loosely coupled generator would require the generator admittance, $y_0 \to 0$, and a tuned detector would mean that $y_d = g_d$.

A. Apparent Standing-Wave Ratio

The relationship between the apparent and true VSWR (voltage-standing-wave-ratio) can be computed from the line equations. Referring to Fig. 1 and letting $y_d = 0$ for the moment, the voltage at w is

$$V_w = V_g \frac{\cosh \gamma w + y_r \sinh \gamma w}{(1 + y_r/y_0) \cosh \gamma l + (y_r + 1/y_0) \sinh \gamma l}, \quad (1)$$

where

V_g = generator voltage
$\gamma = \alpha + j\beta$, complex propagation constant
w = distance from load
l = total length of line
y_0 = normalized generator admittance
y_r = normalized load admittance
y_d = normalized detector admittance.

Without loss in generality, the terminating impedance can be specialized to a pure conductance having a normalized value less than unity since any phase angle may be added by a section of line. For an arbitrary gene-

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† Formerly, Cruft Laboratory, Harvard University, Cambridge, Mass.; now, Sperry Gyroscope Co., Great Neck, L. I., N. Y.

[1] D. D. King, "Impedance measurement on transmission lines," Proc. I.R.E., vol. 35, pp. 509–514; May, 1947.

[2] W. Altar, F. B. Marshall, and L. P. Hunter, "Probe error in standing-wave detectors," Proc. I.R.E., vol. 34, pp. 33P–45P; January, 1946.

[3] Y. N. Dowker and R. M. Redheffer, "An Investigation of RF Probes," Radiation Laboratory Report 483–14, NDRC Div. 14, OEMsr-262, February 6, 1946.

[4] C. G. Montgomery, "Technique of Microwave Measurements," vol. 11, chap. 8, Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., 1947.

rator admittance, a tuned detector ($y_d=g_d<1$), and a lossless line ($\alpha=0$) the apparent VSWR, which is less than the true VSWR, $1/g_r$, can be calculated by combining the apparent load admittance at the expected distribution extremes with the detector admittance. This yields

$$VSWR_a = \frac{1}{g_r}\left|\frac{y_0+g_r+y_0g_rg_d}{y_0+g_r+g_d}\right|. \quad (2)$$

For a loosely coupled generator, $y_0\to 0$, and

$$VSWR_a = \frac{1}{g_r+g_d}. \quad (3)$$

This is the ratio of the voltage at $\lambda/4$ from the voltage minimum to the voltage at the minimum. If the detector coupling is less than critical, this ratio is simply that of voltage maximum to voltage minimum. When the loading effect of the detector at a voltage minimum is not severe, $g_d \ll 1/g_r$, the true VSWR, $1/g_r$, and hence g_d, can be obtained from either the width of the resonance curve with the detector near the voltage minimum or the width of the distribution minimum.[1,5]

B. Detector Deflection as a Function of Detector Coupling

Ideally, a detector must give maximum response from a given weak signal. Detectors of this type must be tuned to resonance and thus be tuned to the same frequency as the signal on the resonant transmission line. From low-frequency circuit theory,[6] if two resonant circuits are coupled more tightly than a critical value, the secondary current will decrease from a maximum. At the critical value, the reflected resistance is equal to the primary resistance.

Let us consider a resonant transmission line to be the primary and a resonant detector to be the secondary of two coupled circuits. The detector deflection, D, which is proportional to the secondary current, is given by:

$$|D| \propto \frac{k}{k^2+k_c^2}, \quad (4)$$

where

k = coefficient of coupling,

$k_c = \frac{1}{\sqrt{Q_1Q_2}}$, critical coupling.

Equation (4) is plotted in Fig. 2 and clearly shows the linear relationship of deflection to coupling when the latter is much less than critical. Such a curve can be obtained if the location of the standing-wave detector is fixed and the coupling varied.

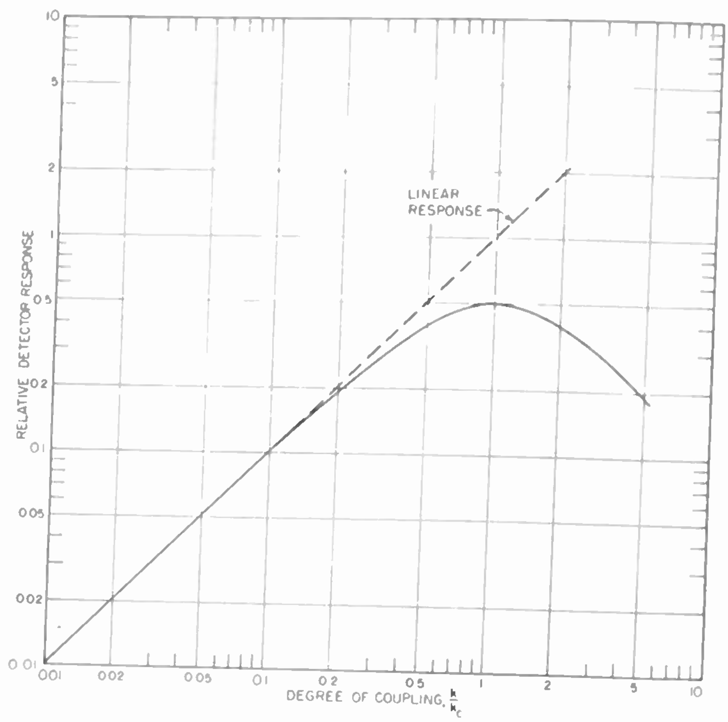

Fig. 2—For loose couplings, $k/k_c \leq 0.1$, the detector response is linear. Maximum deflection is obtained at critical coupling.

C. Distribution Curves as a Function of Detector Coupling

The distribution curves on a transmission line depend not only on the coupling of the detector but on the coupling of the generator as well. If the generator is loosely coupled to an unloaded line and the tuned detector is displaced longitudinally, the deflections obtained are exactly the same as the currents in the secondary of a conventional lumped-constant circuit whose coefficient of coupling is varied cosinusoidally. This can be shown readily by equating the power absorbed by the detector on the transmission line to the power absorbed by the secondary of the lumped-constant circuit. For a low-damped line, the effect is the same as varying cosinusoidally the coefficient of

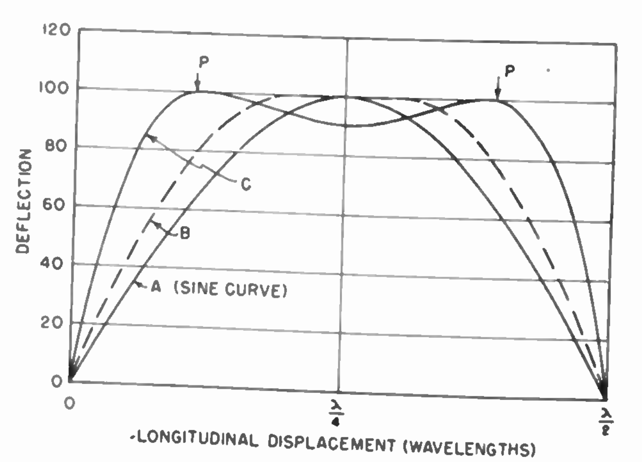

Fig. 3—Standing-wave distributions on an unloaded line for three detector couplings. The normalized curve A is for loose coupling, curve B for critical coupling, and curve C for a coupling greater than critical.

[5] K. Tomiyasu, "Problems of Measurement on Two-Wire Lines with Application to Antenna Impedance," Cruft Laboratory Technical Report No. 48, Harvard University, June 15, 1948.

[6] Cruft Electronics Staff, "Electronic Circuits and Tubes," chap. 7, sec. 8, McGraw-Hill Book Co., New York, N. Y., 1947.

coupling in the lumped-constant circuit in all regions not near minimum deflections.

For a sinusoidal variation in coupling on an unloaded line, a sinusoidal distribution will be found only if the detector is coupled far below critical at the distribution maximum. Such a distribution is shown as curve A in Fig. 3. If the detector is tightly coupled, the maximum deflection will not occur at the expected position, but rather at two symmetrically located peaks, one on each side of the expected position. This is shown by curve C. It is quite apparent that if a detector is over-coupled at the normal distribution maximum, the deflection will be smaller at this location than at either of the side peaks P, where the coupling between the two circuits, line and detector, has been reduced to the critical value. In the transition, i.e., critical coupling, the distribution curve will assume a flat top as shown by curve B. Actually, if the plots are of observed deflections, curve A would not reach the same maximum value as curve B or curve C, since it is coupled less than critical. In the graph, curve A has thus been normalized to the increased maximum value for clarity. All curves are symmetrical since the detector is tuned.

From ordinary transmission-line equations, the positions of distribution extremes for a tuned detector and a loosely coupled generator can be shown to satisfy the equation

$$CSg_r^2(g_r^2 + 2g_rg_d - 1) + CSg_d^2(C^2 + g_r^2S^2)^2 = 0 \quad (5)$$

where

$$C = \cos \beta w$$
$$S = \sin \beta w.$$

The slope of the distribution curve is zero when $\cos \beta w = 0$ and when $\sin \beta w = 0$. This is true for all values of g_d. However, when g_d equals a critical value,

$$g_d \doteq g_r(1 - \tfrac{3}{2}g_r^2), \text{ (for small values of } g_r), \quad (6)$$

not only is the slope of the distribution curve equal to zero when $\cos \beta w = 1$, but the second space derivative as well. This means that the distribution curve becomes "flat-topped" at the expected maximum. Hence for critical coupling the effective shunting conductance of a tuned detector is approximately equal to the terminating conductance. This assumes a lossless line.

For couplings greater than critical, the locations of maximum deflections are obtained from the equation

$$\cos\beta w = \pm(1 - g_rg_d)\sqrt{g_r/g_d} \text{ (neglecting } g_r^2 \text{ compared to 1)} \quad (7)$$

and the distribution curve is characterized by a double hump.

If the detector admittance has a susceptance component as well, the distribution curves will become unsymmetrical within each half-wavelength. For couplings greater than critical, the higher peak will be indicated by the shift of the distribution maximum. (The shifts due to a susceptance detector are considered in Part III.)

D. Resonance-Curve Width as a Function of Secondary Coupling

A further investigation of conventional lumped-constant circuits reveals some interesting results concerning the over-all damping of the coupled system. The over-all damping effect can be measured by the width of the resonance curve.[1] From the secondary current equation[7] the resonance-curve width ΔS is given by

$$\Delta S = \frac{\Delta\lambda_1}{\lambda_0} = \frac{1}{\sqrt{1 - Q_2(k_c^2 + k^2)}} - \frac{1}{\sqrt{1 + Q_2(k_c^2 + k^2)}}, \quad (8)$$

where $\Delta\lambda_1$ = change in primary resonance wavelength for the half-power points, and
λ_0 = operating wavelength.

Applying (8) to special cases, a loosely coupled secondary ($k^2 \ll k_c^2$) yields

$$\frac{\Delta\lambda_1}{\lambda_0} \doteq \frac{1}{Q_1}\left(1 + \frac{k^2}{k_c^2}\right) \doteq \frac{1}{Q_1} \equiv \Delta S_0$$
$$\text{for 1 per cent maximum error, } Q_1 > 10. \quad (9)$$

For a critically coupled secondary ($k = k_c$) the resonance-curve width is

$$\Delta S = 2\Delta S_0. \quad (10)$$

An infinite half-power width is obtained for a coupling coefficient of:

$$k \doteq \frac{1}{\sqrt{Q_2}} \quad \text{for 1 per cent maximum error } Q_1 \geqq 100$$
$$= k_c\sqrt{Q_1}. \quad (11)$$

In Fig. 4 the resonance-curve width from (8) as a function of secondary coupling is plotted. For couplings much smaller than critical the resonance-curve width ΔS is a constant and the effect of the tuned detector is negligible.

E. Condition for a Loosely Coupled Detector

For negligible effects a tuned detector must be loosely coupled to the line. If loose coupling is arbitrarily defined as

$$\frac{k}{k_c} \leqq 0.1, \quad (12)$$

this is equivalent to the following relationship:

$$g_d \leqq \frac{1}{98} g_r. \quad (13)$$

When loosely coupled, the apparent VSWR differs by less than one per cent from the true value and the deflections are proportional to the coupling.

[7] Equation (8.10), footnote reference 6.

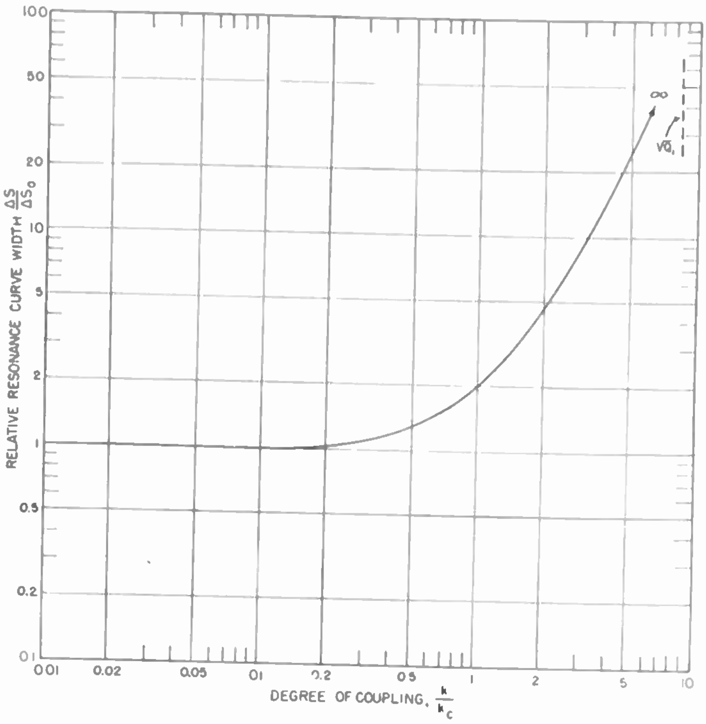

Fig. 4—Relative resonance-curve width of a low-damped line as a function of detector coupling. ΔS_0 is the width when the detector is very loosely coupled.

Since the condition for loose coupling depends upon critical coupling, it also depends upon the primary Q_L. Increased dissipation in the primary decreases Q_1 and increases k_c. A larger k_c reduces the possibility of over-coupled effects. This further implies that if over-coupled effects are not observed on an unloaded line, there certainly will not be any over-coupled effects when the line is terminated in a dissipative load.

II. Matched Generator and Tuned Detector

When the generator is matched to the line ($y_0=1$) all readings and effects are independent of the distance between the generator and the detector. This means that resonance-curve widths cannot be measured.

A. Apparent VSWR

For a matched generator and a tuned detector, the apparent VSWR from (2) is given by:

$$VSWR_a = \frac{1}{g_r} \frac{1 + g_r + g_r g_d}{1 + g_r + g_d}. \tag{14}$$

B. Distribution Curves as a Function of Dector Coupling

For a matched generator and a tuned detector the zero slopes of the distribution curve occur only at cos $\beta w=0$ and sin $\beta w=0$. This means that double-hump curves cannot be found, and critical coupling of the detector does not exist. If the loading effect is significant, the distribution curves will be altered to yield the apparent VSWR given by (14). The distribution curves will be symmetrical.

C. Condition for a Loosely Coupled Detector

For a line with a low-damped termination, the condition for loose coupling of the tuned detector is given by:

$$g_d < \frac{1 + g_r}{99 - 100 g_r} \doteq \frac{1}{99} \quad \text{for} \quad g_r^2 \ll 1. \tag{15}$$

The error in the apparent VSWR for this condition will be less than one per cent.

III. Detector Represented by a Pure Susceptance

The most apparent effect of a susceptance detector regardless of the coupling of the generator is an unsymmetrical distribution curve. The analytical solutions for the effects of this detector are readily obtainable only when the generator is matched to the line.

A. Shift of the Distribution Extremes

From conventional transmission-line equations, and assuming a lossless line, the distribution extremes are obtained by taking the space derivative of the absolute magnitude of the voltage at the detector and setting it equal to zero. The shifts δ of the voltage maximum and minimum are found to be:

$$\text{for } V_{\max}: \quad \frac{\delta_{\max}}{\lambda} \doteq \frac{-b_d}{2\pi(1 + g_r)^2} \tag{16}$$

$$\text{for } V_{\min}: \quad \frac{\delta_{\min}}{\lambda} \doteq \frac{-b_d g_r^2}{2\pi(1 + g_r)^2} \tag{17}$$

$$\text{subject to the condition } |b_d| \leqq \frac{(1 + g_r)^2}{4 g_r}.$$

A negative shift means that the observed location is closer to the load than the normal location. In order to clarify any ambiguity which may arise from the sign of the susceptance, $b \equiv -x/|z|^2$.

Both shifts are in the same direction and increase with increasing susceptance. Since $\delta_{\min}$ is less than $\delta_{\max}$ by a factor of g_r^2, $\delta_{\min}$ is indeed very small for small values of g_r.

From (16) the magnitude of the susceptance can be determined if the load conductance is known. A load of $g_r=0$ would be convenient for its determination. It can be readily shown that the combination of load and detector admittances at the distribution extremes does not yield a pure conductance.

B. Apparent VSWR

Assuming that $b_d^2 \ll g_r^2 \ll 1$, the apparent VSWR is:

$$VSWR_a \doteq \frac{1}{g_r}\left[1 - \frac{b_d^2 g_r^2}{1 + 4 g_r}\right]. \tag{18}$$

The fact that the susceptance appears squared signifies that the apparent VSWR will be identical for positive and negative susceptances of equal magnitude.

C. Condition for a Loosely Coupled Detector

If the detector has a susceptance which is less than

$$|b_d| < \frac{\pi(1+g_r)^2}{180} \doteq \frac{\pi}{180}; \quad g_r \ll 1, \qquad (19)$$

the error in the position of voltage maximum will be less than one degree.

It the susceptance is less than

$$b_d^2 < \frac{1+4g_r}{100g_r} \doteq \frac{1}{100g_r}; \quad g_r \ll 1, \qquad (20)$$

the error in the apparent VSWR will be less than one per cent.

Other Remarks

A study of the characteristics of coupled circuits reveals that for a coupling which is greater than critical, a maximum secondary current can be obtained that is the same as at critical coupling if both primary and secondary are simultaneously tuned to either longer or shorter wavelengths, with the oscillator frequency kept constant.[6] This is one test for determining the presence of over-coupled effects, and it has been verified in circuits having distributed constants.

An experimental method of determining if the detector is loosely coupled would be to change the coupling of the detector and compare the measurements. If loosely coupled, the measurements will be identical. The curves of Figs. 2, 3 and 4 have been verified qualitatively on a two-wire line operating at a frequency of 300 Mc.

The analysis thus far concerns the coupled detector and its effect; however, identical effects can be attributed to the generator which may be capacitively or inductively coupled to the line, for it, too, usually represents another coupled resonant circuit. The effect of a coupled generator is more easily corrected, provided, of course, the frequency and output voltage remain constant, since it is subtracted out in the resonance-curve width method and does not appear in the standing-wave-ratio method of measuring impedances.

Conclusion

Impedances measured by the SWR method or by the resonance-curve-width method are a function of the detector coupling if the detector is not loosely coupled. Impedances measured from the position and width of a distribution minimum will not be in error if $g_d \ll 1/g_r$.

Simple crystal detectors with sensitive microammeters, which often exhibit loading and over-coupled effects, can be used without correcting for any errors if the power levels on the transmission line are high. Moreover, such detectors as bolometers, spectrum analyzers, and superheterodyne receivers, which have considerably higher sensitivities, may also exhibit loading and over-coupled effects when the prevailing power levels are extremely low.

It is hoped that by recognizing the errors introduced by tight coupling, proper adjustments can be made to eliminate their adverse effects.

Acknowledgment

The author wishes to thank R. W. P. King of Cruft Laboratory for his valuable advice in the preparation of this paper.

Analogue Studies of Losses in Reflex Oscillator Cavities*

FREDERICK W. SCHOTT†, ASSOCIATE, IRE, AND KARL R. SPANGENBERG‡, FELLOW, IRE

Summary—**An analysis is made which shows the method of applying the network analogue to the investigation of the effect of dielectric and wall losses on cavity-resonator behavior.**

The *Q* and shunt resistance of re-entrant cavities operating in the first- and second-order *TM*$_0$ type modes are investigated. The condition for a zero of shunt resistance is determined. Experimental results are discussed.

I. Introduction

Previous papers[1,2] have described the design of network analogues for studying the electromagnetic field relationships in two-dimensional systems, and have indicated the wide variety of problems that may be handled. The present paper will be concerned with the application of such an analogue to the study of cavity-resonator losses in systems having axial symmetry. The analogue may represent either TM_0 or TE_0 type waves; however, the usual mode of coupling is such that the TM_0 mode is desired while

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[1] K. Spangenberg, and G. Walters, "An electrical network for the study of electromagnetic fields," Technical Report No. 1, Electronics Research Laboratory, Stanford University, Contract N6-ORI-106, Task III, May, 1947.

[2] K. Spangenberg, G. Walters, and F. W. Schott, "Electrical network analyzers for the solution of electromagnetic field problems: Part I—Theory, design, and construction," Proc. I.R.E., vol. 37, pp. 724–729; July, 1949. "Part II—Operation," Proc. I.R.E., vol. 37, pp. 866–872; August, 1949.

geometrically similar resonators operating in the same oscillation mode.

(B) Possibility of a Zero of Shunt Resistance

From (4) it may be seen that shunt resistance is proportional to $[\int_{\text{area}} H_\phi \, da_c]^2$. For the lowest TM_0 oscillation mode H_ϕ is in time phase over the entire region and the above integral always has a nonzero value. For the second-order oscillation modes, however, H_ϕ changes sign over part of the region so that essentially

$$\int_{\text{area}} H_\phi da_c = \int_A |H_\phi| da_c - \int_B |H_\phi| da_c \qquad (21)$$

where regions A and B are separated by the nodal surface of H_ϕ.

This indicates the possibility that the total integral, and hence the shunt resistance may be zero for the higher modes. A criterion may be established in this way: For the higher modes, there is always a surface (or line in the r–z plane) passing through a point of zero electric field along which there is no tangential component of electric field. Such a surface may be called a nodal surface in a restricted sense only, for along it, the normal component of electric field is not zero. If a contour of integration be made to coincide with this "nodal" surface and the axis, it is apparent that the only contribution to $\int \overline{E} \cdot \overline{ds}$ along such a path will be along the axis. It may be shown that this integral must be greater than zero in general.

The magnitude of the electric field is given by

$$|E| = \frac{1}{\omega_c \epsilon r_c} \text{grad } (r_c H_\phi). \qquad (22)$$

Therefore, the "nodal" surface of electric field must lie between the zeros of $r_c H_\phi$. In all TM_0 modes H_ϕ (and $r_c H_\phi$) is zero along the axis. Thus, the first type of nodal surface to be encountered in moving out radially is one of electric field. Since

$$\oint \overline{E} \cdot \overline{ds} = j\mu\omega_c \int_{\text{area}} H_\phi da_c \qquad (23)$$

and inasmuch as H_ϕ is of the same sign over this area because of the integration path which has been chosen, the integral cannot be zero. Consequently, the shunt resistance cannot be zero in any such general case. There remains but one situation that will permit a zero of shunt resistance; *it exists when the nodal surface of H_ϕ passes through the gap.* A very simple example of such a case is the TM_{011} mode in the elementary cylindrical resonator.

(C) Cavity Q

The work of the preceding section in the analysis of cavity shunt resistance points out the type of reasoning that must be pursued in the investigation of cavity Q.

The criterion for the wall loading has already been determined. Therefore, it is simply necessary to establish the analogue of cavity Q in terms of network quantities and thus obtain the desired conversion factor.

The stored energy in the cavity is

$$W_c = \int_{\text{vol}} \mu H_\phi^2 dv_c \qquad (24)$$

since H_ϕ represents the rms value of the magnetic field. As before, an integration in the ϕ direction may be performed immediately because TM_0 waves are being considered. Thus, the stored energy is

$$W_c = 2\pi \int_{\text{area}} \mu H_\phi^2 r_c da_c \qquad (25)$$

and recalling (1) which expresses the wall loss, a result for cavity Q is obtained in the form

$$Q_c = \frac{2\pi\mu\omega_c \displaystyle\int_{\text{area}} \frac{(r_c H_\phi)^2}{r_c} da_c}{2\pi R_s \displaystyle\int_{\text{wall}} \frac{(r_c H_\phi)^2}{r_c} dw_c}. \qquad (26)$$

This is to be compared with a corresponding ratio for the network. Here also the loss corresponding to the wall loss has been expressed. Rewriting (8)

$$P_N = \frac{1}{R_{s1}} \int_{\text{wall}} \frac{V^2}{r_N} dw_N. \qquad (27)$$

Hence, it is only necessary to determine the stored energy in the network. Because the time phase relationships are such that the junction voltage reaches its maximum everywhere at the same instant, the network stored energy is

$$W_N = \sum V^2 \Delta C \cong C_1 \int_{\text{area}} \frac{V^2}{r_N} da_N \qquad (28)$$

with C_1 referring, as before, to the capacity per unit area at unit radius.

The final expression for network Q then becomes

$$Q_N = \frac{\omega_N C_1 \displaystyle\int_{\text{area}} \frac{V^2}{r_N} da_N}{\displaystyle\frac{1}{R_{s1}} \int_{\text{wall}} \frac{V^2}{r_N} dw_N}. \qquad (29)$$

Taking the ratio of (26) to (29) and recalling (13), (14), (16), and (19), then

$$Q_c \frac{\delta}{\lambda} = \frac{1}{200\pi} Q_N. \qquad (30)$$

This result, as (20), is somewhat restricted but the same observation may be repeated: The quantities $R_{sh}(\delta/\lambda)$ and $Q_c(\delta/\lambda)$ are, for a given oscillation mode, functions of the geometry only.

(D) Dielectric Losses

The only losses previously considered were those associated with the finite conductivity of the resonator walls. However, when the resonator is used with a vacuum tube, the portion of the resonator containing the interaction gap must be evacuated; but it is generally not expedient to operate a tunable cavity entirely under a vacuum. Therefore, a glass seal is ordinarily placed around the gap region to permit the tunable portion of the cavity to operate in air. This dielectric material may occupy but a very small portion of the cavity volume and it may, in addition, have a loss factor of the order of 0.001, yet it may contribute very greatly to the total losses.

In order to represent the dielectric loss properly, it must be made to exist in the proper proportion relative to the wall loss. That is, once a factor has been decided upon for the wall loss, i.e., the amount of network wall loading, the same factor must be used for the dielectric loss. Thus

$$\left(\frac{P_{\text{wall}}}{P_{\text{dielectric}}}\right)_{\text{cavity}} = \left(\frac{P_{\text{wall}}}{P_{\text{dielectric}}}\right)_{\text{network}}. \quad (31)$$

From (30) it may be seen that with a wall loading represented by $R_{s1} = 10^5$ ohms, the wall loss in the network is increased by a factor of $\lambda/200\pi\delta$. Consequently, the series resistance used to represent the dielectric loss, which from Table I is seen to be the product of $\tan\theta$ by the corresponding coil reactance, must be given by

$$R_{d1} = \frac{\lambda\epsilon' \tan\theta\omega_N L_1}{200\pi\delta} = \frac{5\epsilon' \tan\theta}{\delta\beta} \quad (32)$$

where $R_d = (R_{d1})(r_N)$ and R_{d1} represent the series resistance at unit radius. Thus (20) and (30) are not inviolated by dielectric loss. Rather, (32) tells how to represent the dielectric loss to maintain their validity.

III. Method of Measurement

The theory developed in the previous section has shown that in order to determine the generalized cavity Q and shunt resistance, measurements of network Q and shunt conductance must be made. These measurements are quite simple in nature; however, to obtain a reasonably high degree of accuracy in making network measurements, certain precautions must be taken. Some of these have been discussed in footnote reference 2, but there are others pertinent to the type of observations made in this study.

When making observations of field configurations, the excitation source may be most conveniently connected to the network at a junction point through a large isolating resistance. The particular junction point used for this purpose is quite arbitrary, the only limitation being that the point chosen must not lie near a node of voltage. When the network is driven with this precaution it appears near resonance as a parallel R–L–C circuit to the exciting source; this is desirable, for any harmonics which may be present tend to be attenuated. The magnitude of the isolating resistance is also somewhat arbitrary. From the standpoint of obtaining a large amount of excitation, this resistance should be as small as possible; but the impedance to ground, as viewed at the network junction when looking toward the source, should be large if the field configuration is not to be affected. A suitable compromise is readily found.

(A) The Measurement of Network Q

In a network having a high ratio of stored energy to loss per cycle, the field distribution at resonance is independent of the manner or point of excitation. A suitably designed network meets this requirement and the Q therefore may be observed at any point. If the network is viewed between a junction point and ground, it appears as a parallel-resonant circuit and measurements may be made in the appropriate manner. Again, if the network is opened at any point, it appears series-resonant between the resulting pair of terminals, and the techniques for measuring the Q of a series circuit may be used.

Because of the method used to observe cavity field configurations and because of possible harmonics in the voltage source, the network Q is most readily determined by the former of the two methods which is illustrated in Fig. 2. If the isolating resistor is large, e.g., ten times the equivalent network resistance R_p, the current to the network will remain constant when the frequency of the exciting source is varied slightly. The network Q may then be observed by noting the fractional detuning necessary to reduce the response at point A by a specified amount. In certain cases where the network resonance being studied is far removed in frequency from an adjacent resonance, larger amounts of detuning may be used to advantage for obtaining increased accuracy.

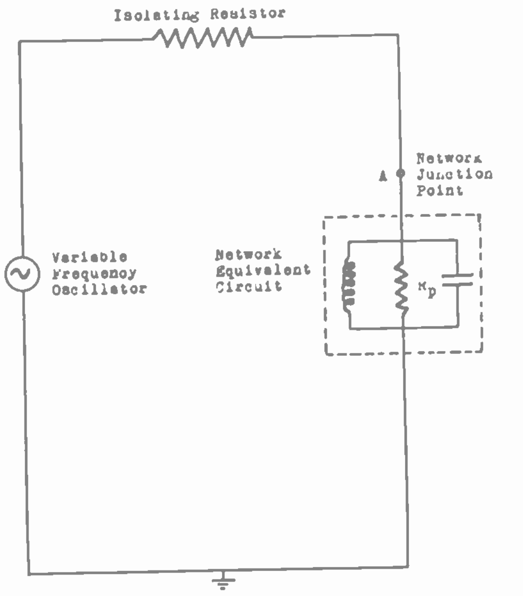

Fig. 2—Basic circuit for Q measurements.

The network Q which is desired is that associated with the losses which are represented by loading the network, as previously noted in the theoretical discussion. Unfortunately, there are always superimposed on these losses the effects of the coil resistance. These coil losses could also be performed to determine the effect of network spacing. Re-entrant cavities were represented first on a section of the analyzer having the larger spacing; the experiments were then repeated using the "fine" section in the vicinity of the gap region (where the fields tend to have the most rapid spatial variations). It was found that there was no appreciable difference in either the field distribution or the measured cavity parameters as long as the nominal wavelength was represented by roughly fifteen network sections or more.

As a result of these and other observations it is believed that the error associated with the measurements may be maintained generally within five per cent. The Q measurements and the shunt resistance measurements made by exciting the networks as a parallel-resonant circuit can be repeated within approximately five per cent of the calculated value. The bridge method for measuring shunt resistance produces values that may be consistently repeated within two per cent, but the results for the lowest-order mode are invariably low. The reason for this is not completely understood.

(B) *The Condition for a Zero of Shunt Resistance*

The verification of the criterion which has been established for the existence of a zero of shunt resistance is possible by a simple example. It has been stated that the shunt resistance may be zero when the node of H_ϕ lies in the interaction gap; thus it is expected that if a tunable coaxial-type cavity operating in a second-order TM_0 mode be investigated, this situation may be found.

In Fig. 6 is shown a particular re-entrant cavity for which this effect has been observed. If for large values of l the cavity operates in the mode usually termed the "three-quarter wave mode," the field configuration in the coaxial portion of the resonator resembles that of a TEM mode and the resonant frequency is relatively sensitive to the length l. When, on the other hand, the plunger is moved flush with the inner conductor, i.e., when $l=s$, the field is that of the TM_{020} mode in which the lines of electric field are purely axial in direction. Here the resonant frequency is relatively insensitive to the plunger position as may be seen from the normalized tuning curve, and as is expected from the perturbation theory of Slater.[5] In passing from the first condition to the second, the nodal surface of H_ϕ was carefully observed, and it was found that the shunt resistance became extremely small as the node approached the gap. It was of interest to note that the rate of change of the nodal position with respect to the length of the cavity was very great in this vicinity.

It was also of interest to note the behavior of the other second-order mode in this region. This other mode has characteristics indicated by the dotted curves in

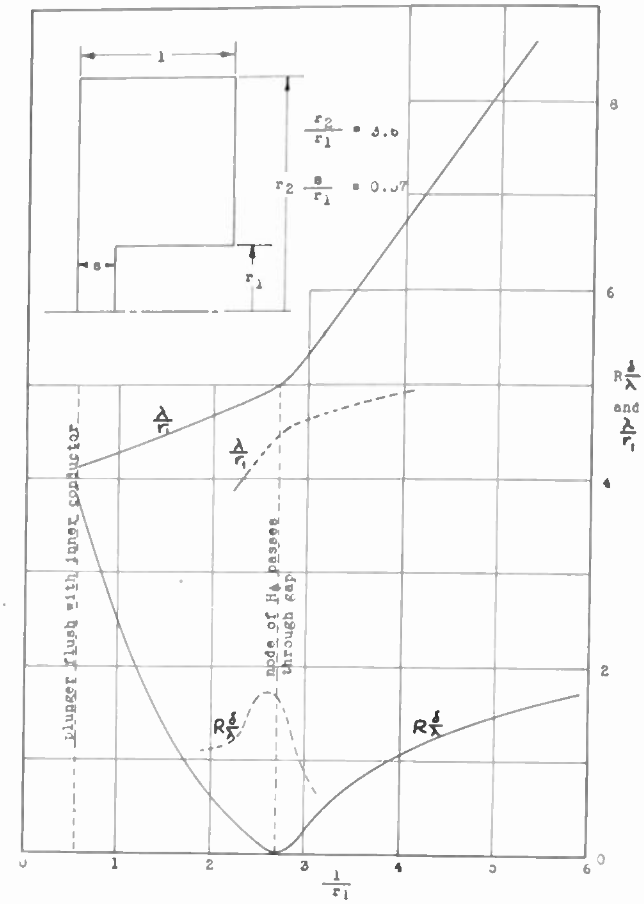

Fig. 6—Behavior of shunt resistance of the low-frequency second-order TM_0 mode.

Fig. 6. These show that although the desired mode might not be capable of excitation for $l/r\simeq 2.6$, there is for the same geometry a nearby resonance for which the shunt resistance is not zero.

V. Methods of Avoiding a Zero of Shunt Resistance

Before pursuing this consideration in detail, it is well to consider the possible oscillation modes that may be excited in a cavity when operating as an integral part of a reflex oscillator. The lowest mode, which has a zero of H_ϕ along the axis of the resonator only, is seldom used in tunable reflex oscillators because of the constructional limitations associated with the small sizes involved; this is perhaps unfortunate, not only because of the situation mentioned above, but also because the lowest mode is nearly always far removed in frequency from any others. The most commonly used mode is one of the second-order TM_0 modes, so designated because there are two nodal surfaces of H_ϕ, one of which is along the axis (and is strictly only a line) and another of which is removed from the axis. Now it must be emphasized that usually only one of the second-order modes is desirable, but that there are, in general, two possible second-order mode types, which may or may not be widely separated in frequency. As an illustration, there is shown in Fig. 7

[5] J. C. Slater, "Microwave electronics," *Rev. Mod. Phys.*, vol. 18, pp. 444–512; October, 1946.

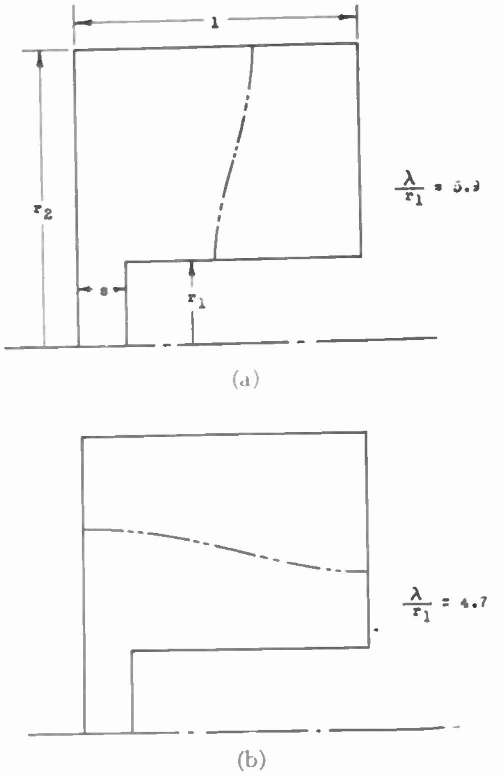

Fig. 7—Position of the nodal surface of H_ϕ for the two second-order modes in a simple re-entrant cavity.

a simple re-entrant cavity and the position of the nodal surfaces of H_ϕ for the two possible second-order modes. For this particular case, the frequencies differ by roughly 25 per cent. In addition, the shunt resistance is approximately the same for both field configurations so that the other factors, such as the characteristics of the electron beam passing through the gap would determine which of the modes would be excited.

It is now possible to investigate what may happen if a resonator is to be operated over a frequency range by tuning the cavity by means of an adjustable plunger. If the length of the resonator is large compared with the other dimensions, as is usually the case at the low frequency end of the tuning range, the desired field configuration will be of the type shown in Fig. 7(a). Also, for cavities of large axial length the second-order mode having the lowest frequency will be the one for which the direction of the lines of electrical field (which coincide with contours of constant rH_ϕ) are predominately radial. This is necessary if there is to be a reasonable change in resonant frequency with change in plunger position; the frequency of a configuration of the type of Fig. 7(b) is highly insensitive to the cavity length.

In a practical case then, if the cavity is in its low-frequency position the nodal surface of H_ϕ will lie in an almost radial plane. As the frequency is increased by moving the plunger toward the gap region, the nodal surface of H_ϕ will also move in that direction. This nodal surface will become distorted, however, as the frequency is further increased, until the field configuration has become that of the TM_{020} mode when the plunger is flush with the inner conductor. This is a general characteristic of simple re-entrant resonators, namely, that the nodal surface of H_ϕ shifts from a radial position to an axial position as the cavity length is decreased.

But there are two ways in which the nodal surface can progress. In Fig. 7(a) it may appear that the bottom of the nodal surface, or that portion of the node adjacent to the inner conductor, would move more rapidly away from the plunger face than the portion adjacent to the outer conductor. This is so. In fact, as this node changes its orientation as described above, the lower end will pass through the interaction gap as shown in Fig. 8(a). If excited suitably, this cavity could tune

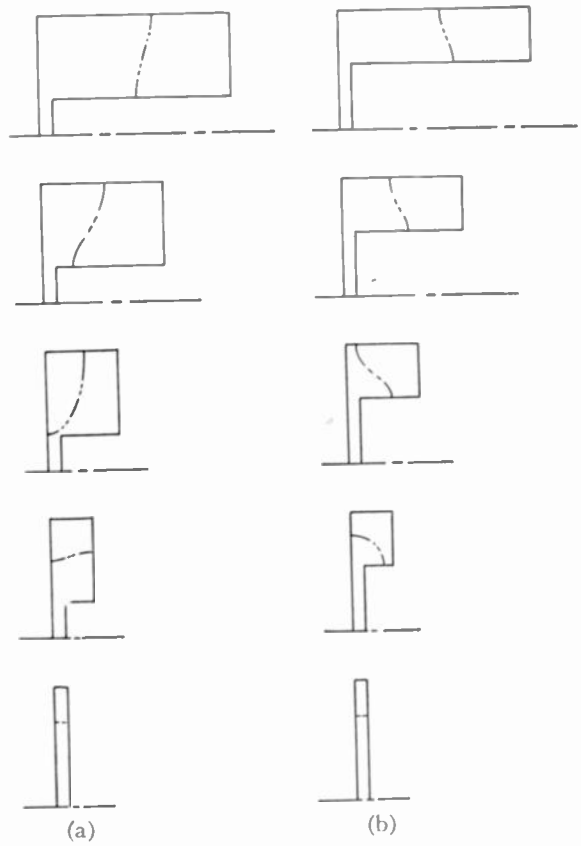

Fig. 8—Progression of the node of H in two similar cavities.

smoothly over this range with the field behaving in the manner indicated, but ***not if the cavity were excited in the normal manner***. The shunt resistance would become zero as the node of H_ϕ passed through the gap, and the cavity, if excited by an electron beam, would cease to oscillate in the desired mode. In practice, the shunt resistance would decrease as the node approached the gap, and when it became too low the cavity would cease to oscillate or jump to another mode.

Fortunately, this limitation on continuous tuning does not now seem to be of great importance, although it may become a factor to be considered if attempts are made to extend the tuning range of cavity resonators. Usually the nodal surface of H_ϕ does not pass through the gap until the plunger is very close to the gap, corresponding to a frequency above the high-frequency limit.

There is, however, a simple method which may be used to avoid this situation. If, as the plunger is moved toward the gap, the node of H_ϕ progresses in such a way that it does not pass through the gap, the shunt resistance cannot become zero. Thus in Fig. 8(b) is shown a slightly different geometry for which this second-order TM_0 mode always presents a reasonable shunt resistance.

Studies of actual reflex klystrons have revealed that, in at least those cases investigated, the geometry is such that the behavior is of the type of Fig. 8(b) and, therefore, this problem is avoided. As an example, there are shown in Fig. 9 the characteristics of the SD835 cavity type as the cavity length is varied. The generalized shunt resistance presented in this graph represents the effect of wall loss only and is, therefore, higher than the actual value, which is affected by dielectric loss.

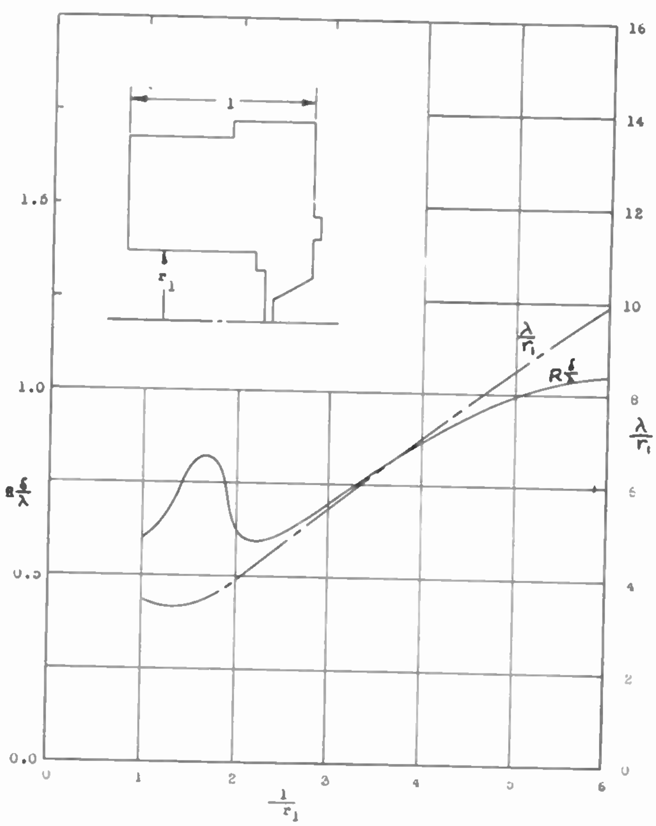

Fig. 9—Shunt resistance and tuning curve of the SD835 type cavity (without dielectric).

In order to obtain general information regarding the possibility of a zero of shunt resistance, a systematic study of cavities of the type of Fig. 7 was made. The objective was to find a relation between the parameters r_2/r_1 and s/r_1 such that, as the cavity is tuned on the desired mode, the node of H_ϕ will not pass through the interaction gap. As a result it has been observed that the desired nodal progression exists when $(r_2/r_1) < 2.2$ and is substantially independent of s/r_1 for practical structures in which $(s/r_1) < 1$.

A completely general study of this type becomes extremely lengthy if a cavity with a large number of geometric variables is to be considered. In spite of this, however, a somewhat more complex cavity of the type shown in Fig. 10 has been investigated over a limited range. The only significant observation is that the effect of the step associated with $(r_1/r_0) > 1$ is to permit utilization of $(r_2/r_1) > /2.2$ without encountering an undesirable type of nodal progression.

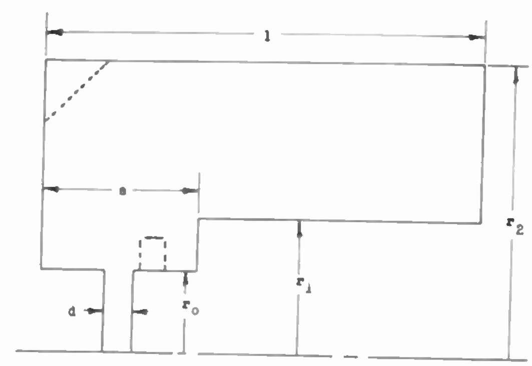

Fig. 10—A cavity with means for eliminating a zero of shunt resistance.

Of perhaps more importance is the determination of slight alterations that may be made to improve the behavior. Numerous methods were tried and it was found that either of the two methods shown by the dotted lines in Fig. 10 tended to eliminate a zero of shunt resistance.

VI. Conclusion

The network analyzer is very valuable for investigating any particular cavity and provides an answer in a short time. It would be desirable, for the benefit of those not having analyzers available, to present in reports such as these a large amount of generalized data, so that any cavity might be designed by interpolation among such data. Unfortunately, the number of variables is so great that such a study is hardly feasible. As data are amassed, however, certain trends may be noticed which will be helpful to designers of cavity resonators. Furthermore, it is feasible to investigate the effect of the variation of dimensions about certain mean configurations.

Beam-Loading Effects in Small Reflex Klystrons*

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Summary—**Beam-loading effects in small reflex klystrons have been measured experimentally by the use of electronic tuning data. They are found to be far greater than predicted by published analyses and to vary more or less linearly with total oscillator load, a variation not previously predicted.**

An analysis of beam-loading effects produced by secondary electrons ejected into the intergrid space by the main electron stream is found to predict large loading and linear variation with load, and to allow the possibility of negative beam loading.

I. Preliminary Consideration

SIMPLE THEORIES of the reflex klystron[1,2] usually neglect the power lost from the resonator fields in the process of changing the velocities of the electrons during their initial transit. When this is included in a more thorough analysis,[3] it is found that the calculated effect of this power loss is equivalent to a constant conductance across the tube grids which is, in most cases, completely negligible compared with the conductance embodying the effects of resonator losses and other loads. Although the beam-loading phenomena in a typical small reflex klystron are calculated by Pierce and Shepherd to be "small and probably less important than various errors in the theory," when measurements are made the beam loading conductances are found to be comparable in size to the resonator conductances, and not at all negligible. Furthermore, the beam loading does not remain constant as the loading of the resonator is varied. Apparently other phenomena than those considered cause loading effects which overshadow the loading described by these analyses.

Abraham has found[4] that beam loading effects in a single transit resonator with mesh grids are much greater than predicted by simple theory. By comparing these results with nongridded apertures, he has established that this is due to the presence of secondary electrons emitted from the grids.

Measurements of beam loading in small reflex klystrons indicate an even greater disagreement with predicted behavior than for the single transit resonator. This is thought to be due to the fact that, whereas in the single transit resonator the secondary electrons are produced at a relatively constant rate, in the reflex tube the bunched electrons of the returning beam strike the grid after passing through the intergrid space and produce a shower of secondary electrons more or less simultaneously. This "bunch" of secondaries traveling back across the grid space might be expected to have considerable effect on the tube characteristics.

Secondary electrons produced by the going and returning primary electron stream may, in turn, produce other secondaries. These are generally unimportant because of their fewness, except in the case of much higher rf voltages than are encountered in the low-power tubes under consideration, when the multipactor action described by Abraham may cause very heavy loading.

This paper contains an account of the methods and results of making measurements of beam loading and, as an Appendix, an analysis which attempts to explain observed results in terms of loading by secondary electrons ejected into the intergrid space by the main electron stream.

II. Measurements of Beam-Loading Q

The experimental method of measuring beam-loading Q is one of measuring separately the "cold" Q_L of the tube and cavity (with the accelerating voltage off but with the cathode hot) and the operating Q_t, and computing the beam loading from the assumption

$$\frac{1}{Q_b} = \frac{1}{Q_t} - \frac{1}{Q_L}. \tag{1}$$

The cold Q measurement is by now too standard an operation to mustify elaboration here.[5,6] The basic equation for the determination of the operating Q is easily derived from the familiar equivalent circuit for a reflex klystron and resonator.[1] The requirement for oscillation is that the sum of the electronic and resonator admittances be zero; hence the tangents of the phase angles of the admittances must be equal. Thus

$$\tan 2\pi(n + \tfrac{3}{4} - f\tau) = Q_t\left(\frac{f}{f_0} - \frac{f_0}{f}\right), \tag{2}$$

where τ is the repeller space transit time. Differentiating with respect to the repeller voltage and evaluating for $f\tau = n + \frac{3}{4}$, $f = f_0$, we have upon rearranging

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[1] E. L. Ginzton and A. E. Harrison, "Reflex klystron oscillators," Proc. I.R.E., vol. 34, pp. 97P–113P; March, 1946.

[2] D. R. Hamilton, J. K. Knipp, and J. B. H. Kuper, "Klystrons and Microwave Triodes," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 254, 311–351; 1948.

[3] J. R. Pierce and W. G. Shepherd, "Reflex oscillators," *Bell Sys. Tech. Jour.*, vol. 26, Appendix 8, pp. 663–672; July, 1947.

[4] W. Abraham, "Loading of resonant cavities by electron beams," *Phys. Rev.*, vol. 72, p. 741; October 15, 1947.

[5] J. C. Slater, "Operation and testing of reflex oscillators," MIT Radiation Laboratory Report No. 742.

[6] C. G. Montgomery, "Technique of Microwave Measurements," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 333–342, 1947.

$$Q_l = \pi f_0^2 \left. \frac{\left| \dfrac{d\tau}{dV_r} \right|}{\dfrac{df}{dV_r}} \right|_{f\tau = n+3/4} - \pi(n + \tfrac{3}{4}). \quad (3)$$

In order to obtain $d\tau/dV_r$ it is assumed that $f\tau = n + \frac{3}{4}$ when df/dV_r is a minimum. If measurements are taken on several modes at different frequencies a curve of τ

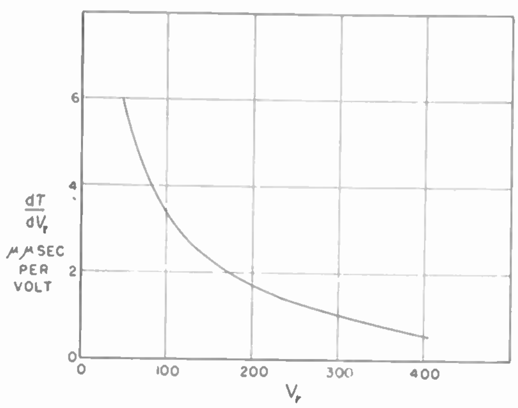

Fig. 1—2K28 repeller characteristic. Anode voltage, 300 volts.

versus V_r (Fig. 1) may be plotted which permits accurate determination of $d\tau/dV_r$.

Thus (3) provides a means of determining the operating Q_l from a measurement of the electronic tuning, df/dV_r. The second term in (3) is usually negligible.

The electronic tuning is conveniently measured by frequency modulating the klystron by superimposing a small sine or square wave on the repeller voltage and observing the resulting frequency spectrum when the output is beat down to a suitable frequency range. This may be done on a selective receiver for the square-wave modulation or with a spectrum analyzer type of circuit for the sine wave modulation. In the first case, the separation of the two output frequencies is measured: in the second, it is more convenient to measure the modulation coefficient of the frequency modulated output by varying the amplitude of the modulating voltage until the carrier first goes through zero.[7]

The experimental procedure may be facilitated by a method which requires but one cold Q measurement for a series of beam-loading measurements. The klystron and its resonator are coupled to a standing-wave detector through a coupling loop and short coaxial line which are left unaltered throughout the series of measurements. The cold Q_0 measurement is made by feeding power from a signal generator through the standing-wave detector to the resonator. In the process of this measurement a reference point x_0 on the standing-wave detector wlll be found such that the admittance of the resonator and connecting line at this point will vary with frequency near resonance in the same manner as the admittance of an equivalent parallel resonant circuit.[8] This reference point will be the position of a voltage minimum when the frequency is the original resonant frequency of the cavity, the cavity being detuned. At resonance the reference point is the position of a voltage minimum if the cavity is undercoupled, or a voltage maximum if it is overcoupled.

In making operating Q_l measurements the signal generator is replaced by an impedance transformer and load. The transformer is always adjusted so that either a voltage maximum or minimum appears at x_0 when the klystron is supplying power. In this case, neglecting the reciprocal of the voltage standing wave ratio (VSWR) when the cavity is far off resonance, it is easily shown that the resonator Q_0 is related to the Q_l of the load by

$$\frac{Q_l}{Q_0} = \frac{\rho}{\sigma_0}. \quad (4)$$

Here ρ is the reciprocal of the VSWR for a voltage minimum at x_0 or the VSWR for a voltage maximum at x_0. The symbol σ_0 denotes the reciprocal of the VSWR at resonance for the undercoupled case or the VSWR at resonance for the overcoupled case when power is fed into the cavity. Thus

$$\frac{1}{Q_L} = \frac{1}{Q_0}\left(1 + \frac{\sigma_0}{\rho}\right), \quad (5)$$

which permits the determination of the loading on the

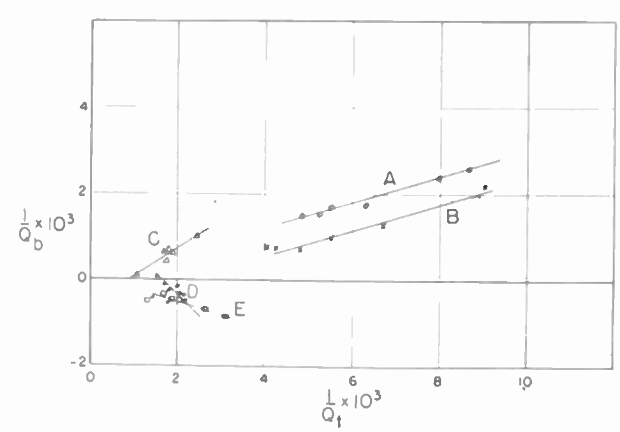

Fig. 2—Measured beam-loading of 2K28.

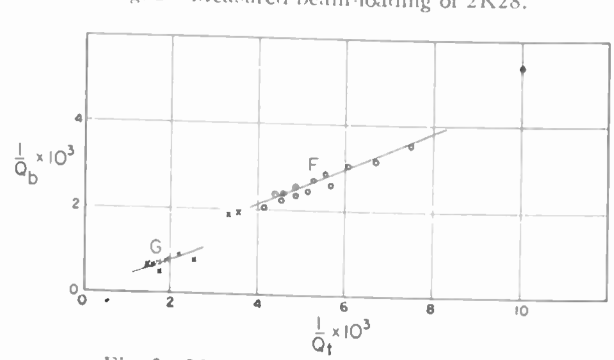

Fig. 3—Measured beam-loading of 6BL6.

[7] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., Inc., New York, N. Y., p. 578; 1943.

[8] J. C. Slater, "Microwave electronics," *Rev. Mod. Phys.* vol. 18, pp. 441–512; October, 1946.

cavity for a series of loads with only one Q_0 measurement.

III. Experimental Results

Results of beam-loading measurements are shown in Figs. 2 and 3 and interpreted in Table I. The resonators

TABLE I
Interpretation of Data in Figs. 2 and 3

Data set	Tube	Manufacturer	Freq. (Mc)	Transit time (cycles)	Estimated $C'(\mu\mu f)$
A	2K28	Raytheon	3240	$2\frac{3}{4}$	1.0
B	"	"	3240	$3\frac{3}{4}$	1.0
C	"	"	3440	$2\frac{3}{4}$	3.5
D	"	"	3455	$2\frac{3}{4}$	3.5
E	"	"	3455	$3\frac{3}{4}$	3.5
F	6BL6	Sylvania	3245	$2\frac{3}{4}$	0.8
G	"	"	4095	$2\frac{3}{4}$	4.0

used for runs A, B and F were simple cylinders fitting around the grid contacting rings. The remainder of the runs were taken in resonators which were closed sections of rectangular waveguides with the tube centrally located.

The beam-loading effects depend upon the rf voltage, and hence upon the net conductance at the tube grids. While the conductance is not easily measured directly, the product Q_tG_t is a function of resonator configuration which may be estimated by various analytical and experimental methods. This product is conveniently expressed as an equivalent capacitance C' defined by $\omega C' = Q_tG_t$. For a given resonator G_t is proportional to $1/Q_t$.

In appraising the accuracy of these measurements it should be remembered that the beam-loading is obtained as the difference between two reciprocal Q's which are both difficult to determine accurately, and which may be large compared with their difference. Probably the most serious uncontrollable error is due to the heating of the grids when the tube is operating, since this causes the resonator Q_0 to differ slightly for the two measurements.

IV. Conclusions

Analysis of beam loading due to primary electrons alone predicts a small effect, independent of the external circuit. The analysis of Appendix I, including secondary electron effects, predicts a linear dependence of beam loading with the conductance appearing across the tube grids, such as appears in the experimental results of Figs. 2 and 3. Certain of the parameters appearing in the beam loading expression cannot be measured directly, so a quantitative comparison of theory with experiment is difficult. However, inserting what seem to be reasonable values of parameters in the results of this analysis, it would appear that the center of the "bunch" of secondary electrons emitted from the first grid would, in crossing the gap, encounter fields which were, on the average, decelerating and plots of $1/Q_b$ versus $1/Q_t$ should have a negative slope. The experimental data shows most often a positive slope.

Experiment indicates that beam loading in tubes with grids is far from the negligible quantity it is often assumed to be. For cases of heavy external loading, the beam loading conductance may exceed the conductance due to resonator losses. One result of this is to completely invalidate methods of measuring beam loading which assume it independent of rf voltage level. These measurements definitely indicate that beam loading effects which are hard to calculate, and hence usually neglected, may be far more important than minor second-order corrections to reflex klystron theory which are included because they are calculable.

Acknowledgment

The authors wish to express their appreciation for the valuable suggestions and criticisms offered by Karl Spangenberg of the electrical engineering department and Marvin Chodorow of the physics department, of Stanford University.

Appendix I—Calculation of Secondary-Electron Beam-Loading Effects

The general scheme of analysis of secondary-electron beam loading is to compute the energy taken from or given to the rf field by one secondary electron which is emitted at an arbitrary phase angle $\omega t = \phi$ of the rf field

$$E = \frac{V_1}{d} \sin \omega t$$

between the grids at an average value of emission velocity. Assuming a constant proportionality factor between numbers of secondaries and primaries, this energy can be summed over an rf cycle to yield a net energy gain or loss, from which a conductance can be computed. Here the number of primary beam electrons striking the grids at any phase angle is obtained from klystron bunching theory.

For primary energies of less than a thousand electron volts experimental studies[9] of secondary emission indicate that the majority of secondaries emitted from the grids will have normal components of emission velocities which are only a few per cent of the primary electron velocity. A very few, perhaps a few per cent, may be expected to have velocities approximately equal to the primary velocity. Since a small change in velocity corresponds to an energy change approximately proportional to the velocity, the maximum energy contribution per electron will come from these "direct reflection" secondary electrons. However their number is so small that the major effect must be attributed to the more abundant low velocity secondaries.

If the normal component of initial velocity with which an electron leaves a grid be u_0, the velocity perpendicular to the grid at time t is given by

$$v_t = u_0 + \int_{\phi/\omega}^{t} \frac{e}{m} E dt = u_0 + \frac{V_1 e}{dm\omega} (\cos \phi - \cos \omega t). \quad (6)$$

[9] J. H. O. Harries, "Secondary electron radiation," *Electronics*, vol. 17, pp. 100–108, 180; September, 1944.

Substitution of representative values in this expression demonstrates that except for the very lowest velocity secondaries this velocity is always positive, i.e., always directed toward the opposite grid. This is important because in a moment we wish to average the kinetic energy per electron over a full cycle of transit time, and the averaging is more complicated if the electrons reverse their direction of travel at any time.

Now if t_1 is the time at which the electron reaches the opposite grid, we may set the distance at time t_1,

$$s_{t_1} = \int_{\phi/\omega}^{t_1} v_t dt = \frac{V_1 e}{\omega^2 dm}\left[\left(\cos\phi + \frac{\omega d m u_0}{V_1 e}\right)\omega t_1 + \sin\phi - \phi\cos\phi - \frac{u_0 \phi \omega d m}{V_1 e} - \sin\omega t_1\right], \quad (7)$$

equal to d, the grid spacing.

Considering this distance equation, we see (Fig. 4)

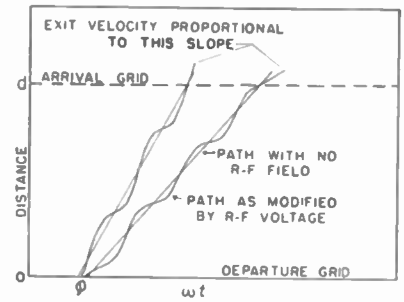

Fig. 4—Distance-time plot of secondary electrons in radio-frequency field between two grids.

that the path of an electron is represented in the distance-time plane by a straight line (first terms) of a slope dependent on u_0 with a superimposed sinusoidal variation (last term). For representative values of grid spacing and initial velocity, the electron will take at least several cycles to travel the intergrid distance. (This is reasonable, remembering that the time of transit for the primary electrons in these tubes is in the neighborhood of a third of a cycle.)

We consider for a moment two secondary electrons produced at the same phase angle with slightly different initial velocities following the trajectories shown in Fig. 4. The exit velocity and energy depend on the slope of this path at the point where the electron leaves the grid space, that is, at the distance d. We note that the exit velocity changes considerably for a small change in initial velocity. Put somewhat differently, over a small range of values of initial velocity (i.e., average slope of path) the transit time $(\omega t - \phi)$ will vary by a half-cycle and the change in electron velocity due to its experiences in the grid space will go from maximum increase to maximum decrease. Thus, averaging the energy change over a cycle of transit time is equivalent to averaging over a range of emission velocities. Extending this argument, we may obtain a mean value of energy averaged over the complete range of emission velocities by averaging the energy over one cycle of transit time and substituting an average value $\bar{u}_0$ of emission velocity.

The mean energy change per electron, as a function of ϕ, thus obtained is

$$\Delta W(\phi) = \frac{1}{2\pi}\int_0^{2\pi} \frac{m}{2}(v^2 - u_0^2)d(\omega t) = \frac{V_1^2 e^2}{2\omega^2 d^2 m}\left[\frac{2\omega d}{V_1}\sqrt{\frac{2U_0 m}{e}}\cos\phi + \cos^2\phi + \frac{1}{2}\right], \quad (8)$$

where we have introduced U_0, the mean emission energy in electron volts, i.e., $\frac{1}{2}m\bar{u}_0^2 = eU_0$. In this integration we have treated the mean velocity $\bar{u}_0$ as a constant, having shown that ωt is a rapidly varying function of u_0.

The product of this energy change with the rate (expressed as a function of ϕ) at which secondary electrons are produced, averaged over a complete cycle of ϕ, yields the power loss (or gain) due to secondary electron beam loading. It may be noted that if the rate of secondary production is constant, i.e., there is no bunching and the primary electrons are evenly distributed throughout the cycle, the quantity in the brackets averages to unity. We take the number of secondaries produced per second to be $\eta = (kI_T/e)$ where k is the ratio of total secondaries produced to number of primary electrons incident on the grids (perhaps 0.1).

The expression for the sum of the going (dc) and returning (bunched) currents from ordinary bunching theory is taken to be

$$I_T = 2I_0\left[1 + \sum_{n=1}^{\infty} J_n(nx)\cos n(\phi - \phi_1)\right], \quad (9)$$

where x is the bunching parameter and ϕ_1 is the phase angle of the rf voltage when the center of the returning bunch strikes the grid nearest the cathode. The total power loss is then

$$P_s = \frac{1}{2\pi}\int_0^{2\pi} \eta\Delta W(\phi)d\phi = \frac{keI_0V_1^2}{\omega^2 d^2 m}\left[1 + \frac{\omega d}{V_1}\sqrt{\frac{2U_0 m}{e}}J_1(x)\cos\phi_1 + \frac{J_2(2x)}{4}\cos 2\phi_1\right] \quad (10)$$

with higher-order terms in the summation integrating to zero because of the orthogonality of the trigonometric functions. With typical values, the last term in the brackets is negligible compared with the first two, and will be dropped.

We have now an expression for total power loss or gain resulting from the low-velocity secondaries. A secondary electron loading conductance G_s is defined by $G_s = (2P_s/V_1^2)$ whence the total beam loading conductance $G_b = G_p + G_s$ becomes

$$\frac{G_b}{G_0} = \frac{G_p}{G_0} + \frac{2ekV_0}{\omega^2 d^2 m}\left[1 + \frac{\omega d}{V_1}\sqrt{\frac{2U_0 m}{e}}J_1(x)\cos\phi_1\right] \quad (11)$$

where $G_0 = I_0/V_0$ and G_p is the equivalent conductance due to primary beam loading.

This expression still contains the quantities x and V_1 for which it is extremely difficult to obtain good theoretical or measured values. However, we note that in operation we shall be interested in the value of beam loading at the center of a mode, under which conditions

$$\frac{|Y_e|}{G_0} = 2\pi N \frac{J_1(x)}{x} = \frac{G_t}{G_0} \qquad (12)$$

where G_t is the conductance seen across the tube gap, which may be taken as essentially the conductance due to the resonator and load losses when beam loading effects are small.

Combining (11) and (12) and noting that $1/Q_t = G_t/\omega C'$,

$$\frac{1}{Q_b} = \frac{1}{\omega C'}\left[G_p + \frac{2ekV_0G_0}{\omega^2 d^2 m}\right] + \frac{k}{\beta d\omega}\sqrt{\frac{2U_0e}{m}}\,\frac{1}{Q_t}\cos\phi_1. \qquad (13)$$

It should be remarked that Q_b is implicit in Q_t by the relation (1). Although (13) could be simply solved explicitly for $1/Q_b$, the form given indidates clearly the dependence on the total Q_t as measured.

Equation (13) includes all the beam loading effects except those due to the "direct reflection" secondaries whose relative number is so small that they are found to contribute a minor portion of the total loading.

In order to obtain an idea of the relative importance of the various terms, we substitute some representative values for small external cavity klystrons into (13) and obtain

$$\frac{1}{Q_b} \approx 10^{-4} + \frac{10^{-2}}{Q_t}\cos\phi_1.$$

Examining voltage-current relations in the reflex klystron we find that at the center of a mode the bunch of electrons returning from the repeller space passes through the center of the grid space at a time when the field is maximum decelerating, i.e., $V_1 \sin \omega t$ is maximum, $\omega t = \pi/2$. We should expect, then, that the proper value to use for ϕ_1 would be $\pi/2$ plus $\delta/2$ (assuming negligible emission time of the secondary electrons). For the tube under consideration, this is in the neighborhood of 150 degrees, so that the cosine is nearly unity and negative.

Appendix II—Glossary of Symbols

C' = equivalent capacitance representing tube grids
d = grid spacing
e = charge of an electron
E = electric field
f = frequency
f_0 = resonant frequency of cavity resonator
G_b = beam loading conductance
G_0 = direct-current conductance equal to I_0/V_0
G_p = conductance due to bunching of electron beam
I_0 = direct-current beam current
I_T = total current
$J_n(x)$ = nth order Bessel function of the first kind
k = ratio of number of secondary to primary electrons
m = mass of an electron
n = transit time mode number
N = repeller space transit time in cycles equal to $f\tau$
Q = circuit factor defined as the ratio of energy stored in the resonator fields to the energy loss per radian
Q_b = Q due to beam loading losses
Q_l = Q due to external load
Q_L = Q due to resonator losses and load
Q_0 = Q due to resonator losses
Q_t = Q due to resonator losses, load, and beam loading
t = time
u_0 = secondary electron initial velocity
v = velocity
U_0 = secondary electron initial energy
V_0 = cathode-anode direct-current voltage
V_r = repeller-cathode direct-current voltage
V_1 = peak radio-frequency voltage across grids
x = bunching parameter equal to $\pi N\beta V_1/V_0$
x_0 = reference plane defined as the position of a voltage minimum with cavity detuned
Y_e = electronic admittance
β = gap modulation coefficient
δ = gap transit angle in radians
η = number of secondary electrons produced per second
ρ = reciprocal of voltage standing-wave ratio for voltage minimum at reference plane, voltage standing-wave ratio for voltage maximum at reference plane
σ_0 = reciprocal of voltage standing-wave ratio at resonance for undercoupled cavity, voltage standing-wave ratio at resonance for overcoupled cavity
τ = repeller space transit time
ϕ = phase angle of radio-frequency grid voltage
ω = angular frequency equal to $2\pi f$.

Fluctuation Phenomena Arising in the Quantum Interaction of Electrons with High-Frequency Fields*

D. K. C. MacDONALD† AND R. KOMPFNER†

Summary—On the basis of a quantum-mechanical analysis by Smith[1] of the interaction of an electron beam with an oscillating resonant cavity, the fluctuations in energy flow in the beam have been analyzed. The expressions for the classical and quantal cases are compared and it is concluded that, under extreme limiting conditions of operation, a just perceptible difference might be observed.

I. Introduction

RECENTLY Smith[1] has analyzed theoretically the energy exchange between a beam of electrons and a resonant cavity excited with electromagnetic energy. He deduces, quantum mechanically, that energy is always exchanged in units $h\nu$, where ν is the frequency of excitation, and has shown on this basis how the transition to a "classical" process occurs when the excitation energy is large compared with one quantum. On the other hand, for energies small compared with one quantum, many electrons will pass without suffering any energy change whatsoever, and thus the behavior in certain respects will differ considerably from that under classical conditions. In particular, the energy fluctuations in the electron beam will differ in the two cases and these we have essayed to analyze in this paper, assuming the validity of Smith's theory.

At conventional radio frequencies (say 10^6–10^8 cps), a quantum of energy is so small that any difference on a classical or quantal basis must certainly be unobservable. At a wavelength of one centimeter, however ($\nu = 3\times10^{10}$ cps) the quantum corresponds to a potential

$$V_q = \frac{h\nu}{e} = 1.23 \times 10^{-4} \text{ volts} \tag{1}$$

and this potential is of the same order of magnitude as the least perceptible signal voltage at the input of microwave receivers. It is, therefore, of interest to examine in principle the influence of the quantum hypothesis on the operation of microwave amplifying tubes and the sensitivity of such microwave receivers.

With electron streams originating at cathodes having temperatures T_c around 1,000°K, as used at present, it is to be expected à priori that such quantum effects will be unobservable since in this case

$$h\nu \ll kT_c. \tag{2}$$

Since, however, in principle a practically monoenergetic beam of electrons can be produced, it appears still worth while to investigate the question quantitatively.

II. Analysis

Let us then consider the phenomenon of an electron beam traversing a resonant cavity at a place where there is an electric field parallel to the direction of motion of the electrons. Let us further assume that the electrons cross the cavity in a small fraction of a cycle, so that—in the classical case—the voltage increment given to the electron is equal to the instantaneous value of the electric field times the distance the electron is traveling in the field. Smith considers also a case of long transit time, but for our purpose it will be sufficient to treat the passage of an electron across the resonant cavity, and to assume with him, the act of energy interchange as instantaneous. In this case, an electron traversing the field at time t will then classically acquire a voltage increment

$$V_t = V \sin \omega t \tag{3}$$

where

$$\omega = 2\pi\nu$$

and V is the peak voltage across the cavity. Thus the voltage increment is a smoothly varying function of time and any increment between zero and $\pm V$ is possible on this view.

In the quantum case, however, it appears from Smith's detailed analysis that energy will be interchanged only in units of

$$h\nu = eV_q \tag{4}$$

that is to say, only voltage increments of integral multiples of V_q are possible.[2] Smith has given general expressions for the probability of such interchange as a function of time, and has shown that the total energy exchange when averaged over a long time is equal to that in the classical case, and also that the voltage increments for large values of V approach those of the classical case.

The expression for the probability that N quanta be absorbed by an electron is

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† Clarendon Laboratory, Oxford, England.

[1] L. P. Smith, "Quantum effects in the interaction of electrons with high frequency fields and the transition to classical theory," *Phys. Rev.*, vol. 69, pp. 195–210; March, 1946.

[2] It is perhaps worth while to mention that the assumption is made here that energy is only exchanged with the electrons in the beam, and only with one of them at one time. In any practical resonant cavity, energy will also be dissipated as heat due to currents flowing in the walls.

It can be shown that conditions should be realizable where the possibility of quanta being shared in either of these ways can be excluded.

$$P_N = e^{-p\sin\omega t}\cdot\frac{(p\sin\omega t)^N}{N!}; \qquad (0 \leqq \omega t \leqq \pi)$$
$$= 0; \qquad (\pi < \omega t < 2\pi). \tag{5}$$

The probability that N quanta be emitted by an electron is

$$P_N = e^{p\sin\omega t}\cdot\frac{(-p\sin\omega t)^N}{N!}; \qquad (\pi \leqq \omega t \leqq 2\pi)$$
$$= 0; \qquad (0 < \omega t < \pi) \tag{6}$$

the ratio of the peak voltage to the "quantum" voltage, $P = V/V_q$

$$\left(V_q = \frac{h\nu}{e}\right).$$

We are chiefly interested in the case where $p \ll 1$, where it is seen that energy is exchanged practically only in single quanta and the probability for absorbing one quantum becomes simply

$$P_1 = p\sin\omega t; \qquad (0 \leqq \omega t \leqq \pi)$$
$$= 0; \qquad (\pi < \omega t < 2\pi) \tag{7}$$

and for emitting one quantum

$$P_1 = -p\sin\omega t; \qquad (\pi \leqq \omega t \leqq 2\pi)$$
$$= 0; \qquad (0 < \omega t < \pi). \tag{8}$$

Thus the most probable phase for an electron to absorb one quantum is around $\pi/2$ and to emit around $3\pi/2$.

Fig. 1(a) has been drawn with this in mind, indicating however that, in actual fact, the phase of absorption or emission will fluctuate in a rather random fashion. Fig. 1(b) shows the idealized classical case for the purpose of comparison, and the assumption has been made, arbitrarily, that the rf peak voltage V is $1/10\, V_q$. These figures bring out clearly the essential difference in the energy exchange process on the two views.

As mentioned above, the mean energy flow in both cases is the same; *the fluctuations in energy flow*, however, will clearly differ and this is the problem to be considered, since it is the fluctuation, or "noise" that will determine the limiting sensitivity of devices involving energy interchange between electron streams and electromagnetic fields.

The total fluctuation will arise from a number of sources:

(a) fluctuations in the rate of arrival of electrons ("shot effect") at the cavity.

(b) fluctuations in the energy content of electrons (thermal velocities of emission) arriving at the cavity. We propose to call this "chromatic noise."

(c) fluctiations in the amount of energy transferred. We propose to call this effect "transfer noise."

The first type of fluctuation is familiar; the second is less well known but is clearly of importance in problems of this type. The third contribution may arise in two different ways; in the classical case, if there were no fluctuation in the rate of arrival of electrons there

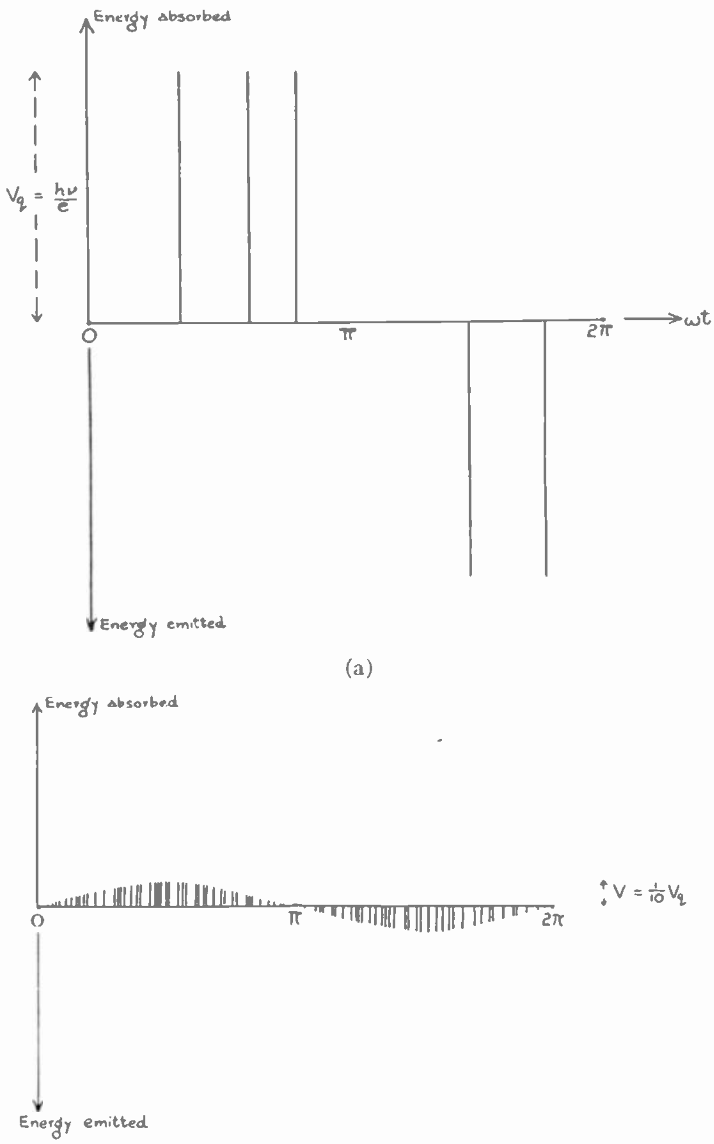


Fig. 1—The energy flow; (a) the quantum case, and (b) the classical case.

would be no fluctuation in the rate of energy transfer. Since, however, the rate of arrival of electrons always fluctuates to some extent, this will result in a fluctuation of energy transfer. Clearly, the amount of energy fluctuation introduced in this case will be a function of the degree of smoothness (quantitatively defined by the symbol Γ^2) of the incident beam.

On the other hand, in the extreme quantum case, where the individual probability of transfer of energy $h\nu$ to a single electron is small, the fluctuation in energy transfer will, therefore, be essentially random irrespective of the degree of smoothness of the electron beam.

It will be assumed in the following analysis that all types of fluctuations are uncorrelated and simply add.[3] Hence the total fluctuation will be the sum of the three types of fluctuation. Detailed analysis yields the fol-

[3] In a conventional thermionic tube, the degree of space-charge smoothing is correlated with the energy spectrum; i.e., Γ^2 is a function of temperature—nonetheless, the electrons leaving the potential minimum have still essentially full chromatic noise. In any event, in the general case, one can readily visualize a beam exhibiting no shot fluctuatuations which possesses full chromatic noise and, on the other hand, a monochromatic beam with full shot effect.

lowing result for the complete energy transfer spectrum:

(*i*) *Classical case*

$$\overline{(W)^2} = \frac{N^2e^2V^2}{2}\cdot\delta(\nu - \nu_0) + 2N^3e^4\Gamma^2Z^2$$

(signal) (induced shot noise)

$$+ 4N(kT_c)^2 + Ne^2V^2\Gamma^2$$

(chromatic noise) (transfer noise) (9)

(*ii*) *Quantum case*

$$\overline{(W)^2} = \frac{(Nph\nu)^2}{2}\cdot\delta(\nu - \nu_0) + 2N^3e^4\Gamma^2Z^2 + 4N(kT_c)^2$$

(signal) (induced shot noise) (chromatic noise)

$$+ \frac{4}{\pi}Np(h\nu)^2 - N(1 - \Gamma^2)p^2(h\nu)^2$$

(transfer noise) (10)

The symbols have the following meaning:

$\overline{(W)^2}d\nu$ = mean-square energy flow per unit time in an arbitrary frequency interval.

N = number of electrons crossing the cavity per second

e = charge of an electron

V = peak voltage across the cavity

$\delta(\nu-\nu_0)$ = Dirac function centered on ν_0 signifying the signal "line" in the spectrum

Γ^2 = shot-noise reduction factor

Z = effective shunt impedance of the cavity

k = Boltzmann's constant (1.37×10^{-23} joules per degree)

T_c = effective temperature of electron stream as defined by the spread of its energy spectrum

p = ratio of peak voltage to "quantum voltage"

$$\left(\frac{V}{\frac{h\nu}{e}} = \frac{V}{V_q}\right)$$

h = Planck's constant (6.55×10^{-34} joule second)

ν = frequency in cps.

The above expressions have been obtained by deriving the appropriate correlation function for the various terms. The frequency spectrum is then obtained as the Fourier inversion of the correlation function. This procedure, now a well-known tool in stochastic analysis, is based on the theorem of Wiener and Khintchine (Rice[4]). The transfer noise terms can be interpreted as "shot" fluctuations of the units of energy transfer; that is to say $\sim eV$ in the classical case and $h\nu$ in the quantum case.

The transfer fluctuations in the latter case may be likened to the problem of emission fluctuations in a photo cell; in that case we are dealing with fluctuations in the rate of arrival of incident photons and fluctuations arising from the fact that the probability of emission is less than unity.[5] If the photons arrived perfectly regularly, then this corresponds to $\Gamma^2=0$ in the above equation. If, however, the probability of emission is small, corresponding to $p\ll1$, then the rate of emission of photoelectrons is essentially random corresponding to neglect to the last term of (10). Formally, these problems are both analogous to the familiar phenomenon of partition noise in a positive-grid tube where the general formula for current fluctuations in the screen current is given by

$$\overline{(\delta i_{SG})^2} = 2eI_{SG}\cdot\frac{\Gamma^2 I_{SG} + I_A}{I_{SG} + I_A} \quad (11)$$

which may be written more fundamentally, for comparison with (10)

$$\overline{(\delta i_{SG})^2} = 2e\{p - p^2(1 - \Gamma^2)\}I_{\text{total}} \quad (12)$$

where p is the probability of capture of an electron by the screen grid.

Reverting to (9) and (10), it is clear that the signal power transferred in both cases is identical since $p=eV/h\nu$. The first two noise terms in either expression are the same. Therefore, for any significant difference between the two cases to be observable these must not be large compared with the transfer noise.

If we take the following values

$$I(=Ne) \approx 10^{-6} \text{ amp}$$

$$Z = 10^3\ \Omega$$

$$V \approx \frac{1}{10}V_q = 1.2\times10^{-5} \text{ volt} \quad (\therefore\ p = 0.1)$$

then the noise in the two cases may be written

$$3.2\times10^{-31}\cdot\Gamma^2 + 5\times10^{-33}\cdot\Gamma^2 + \begin{cases}2.4\times10^{-35}\ \Gamma^2 & (9a)\\ 3.2\times10^{-34} & (10a)\end{cases}$$

(shot noise) (chromatic noise) (transfer noise)

III. Conclusion

It is thus evident that the classical transfer noise will *always* be entirely negligible, while it would be necessary to achieve extreme limiting conditions $\Gamma^2<\sim.01$; $T<\sim1°$K.) for the quantum transfer noise to become perceptible. It is perhaps worth noting, however, that for *normal* operating conditions with such a beam current—say $T\sim10^{30}$K.; $\Gamma^2\sim.1$), the chromatic noise (5×10^{-27}) would apparently dominate the induced shot-noise power (3×10^{-32}). Such a case does not in fact usually arise in practice since the shot noise increases with the cube of the current while the chromatic noise only increases linearly.

[4] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. Jour.*, vol. 23, pp. 282–333, July, 1944; and vol. 24, pp. 46–157; January, 1945.

[5] One might also have to recognize the possibility of fluctuations in the delay between arrival of a photon and departure of an electron.

Design Procedures for Pi-Network Antenna Couplers*

LEO STORCH†

Summary—**The design of reactive pi-networks for transforming a wide range of complex load impedance into a fixed resistance shunted by a tuned circuit is subjected to a thorough investigation. A very significant result is the complete analogy which is established between the analysis and design of the pi-network and the equivalent manipulation of a group of simple geometrical figures.**

1. Introduction

THE AUTHOR was confronted recently with the task of designing an antenna-coupling network for a wide range of antenna impedance and frequency. A search of the literature of recent years disclosed the lack of a satisfactory account concerning coupling networks for the case of complex load impedances. It is the object of this paper to present the principal results of a comprehensive study of reactive pi-networks performing the functions of an antenna coupler.

The design equations are derived in compact form, convenient for use. It is then shown how to express the properties and behavior of the network by a set of circular and parabolic impedance contours. The design process is thereby reduced to the equivalent problem of finding a satisfactory arrangement of these geometrical loci. The pictorial map provides a powerful tool for the choice of the network configuration and determination of optimum design-parameters. The directness of approach and economy of effort in comparison to conventional methods are particularly in evidence when numerous values of antenna impedance and frequency (e.g. aircraft antennas) are involved.

Since the primary application is probably in matching the impedance of the antenna to that of the power amplifier of a transmitter, the terms "antenna coupler" and "antenna impedance" are used exclusively. Such a network finds use, of course, wherever it is necessary to transform a complex load impedance into a certain resistance, e.g., matching to a generator with resistive source impedance, and all the design procedures are directly applicable in such a case.

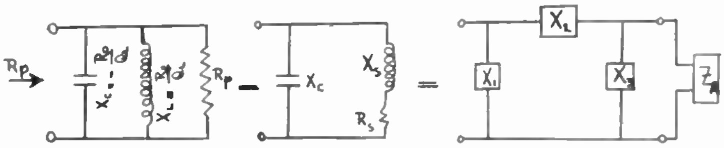
2. Analytical Design Procedure

For satisfactory operation of a radio transmitter, the load presented to the output stage should be a resistance shunted by a parallel-tuned circuit. Since this resistance R_p acts as plate-load for the power-amplifier tube, its value is determined by the tube characteristics and should be the same for all frequencies of operation. The tuned or "tank" circuit, which functions as an energy reservoir, is essential in producing a sinusoidal output signal in spite of class C operation. The values of its two reactances (Fig. 1(a)) depend on the desired value of "loaded Q" (Q_L). Since the antenna impedance Z_A, which symbol may also stand for the input impedance of the transmission line feeding the antenna, will hardly ever have the required value, it is necessary to transform it to this value by means of an antenna-coupling network.

The three-element, reactive, pi-network is a suitable choice for the antenna coupler. Economical in the number of required components, it is capable of handling an extremely wide range of impedance, allows close control of the loaded Q, and contributes to harmonic suppression when the shunt arm on the input side is capacitive.

2.1 Shunt Arms X_1 and X_3

The problem is to make the circuit of Fig. 1(c) equivalent to the loaded tuned circuit of Fig. 1(a) by the proper design of the nondissipative pi-network.[1]



(a) (b) (c)

Fig. 1—Antenna-coupler equivalent circuits.

Because of unavoidable stray capacitances to ground, and in order to contribute to the suppression of harmonics, let the input shunt arm be made capacitive:

$$X_1 = X_c = -\frac{R_p}{Q_L}. \tag{1}$$

Comparing Fig. 1(a) and Fig. 1(b)

$$R_s + jX_s = \frac{R_p \cdot \dfrac{jR_p}{Q_L}}{R_p + \dfrac{jR_p}{Q_L}} = \frac{R_p(1 + jQ_L)}{1 + Q_L^2} \tag{2}$$

or

$$R_s = \frac{R_p}{1 + Q_L^2} \cong \frac{R_p}{Q_L^2}$$

$$X_s = \frac{Q_L R_p}{1 + Q_L^2} \cong \frac{R_p}{Q_L} \tag{3}$$

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† Radio Corporation of America, Camden, N. J.

[1] Residual dissipation in the coils can be accounted for by adjusting R_s in (Fig. 1(c)), if desired.

since $1+Q_L^2 \cong Q_L^2$ (Q_L is about 10 to 20).

Proceeding from Fig. 1(b) to Fig. 1(c):

$$R_s+jX_s=\frac{Z_A \cdot jX_3}{Z_A+jX_3}+jX_2$$
$$=\frac{R_AX_3^2+jR_A^2X_3+jX_AX_3(X_A+X_3)}{R_A^2+(X_A+X_3)^2}+jX_2. \quad (4)$$

Solving the real part of (4),

$$R_s = \frac{R_AX_3^2}{R_A^2+(X_A+X_3)^2}, \quad (5)$$

for $1/X_3$ and inverting

$$X_3 = \frac{R_e}{-\frac{X_A}{R_A} \pm \sqrt{\frac{R_e}{R_s}-1}} \quad (6)$$

where

$$R_e = \frac{R_A^2+X_A^2}{R_A} \quad (7)$$

the equivalent, parallel-resistance component of Z_A.

Since X_3 must be real, it is necessary that

$$R_e \geqq R_s. \quad (8)$$

A graphical representation of this restriction, which hardly results in any practical difficulties, since usually $R_e \gg R_s$, is given in section 3.1. When (8) is satisfied, (6) generally supplies two possible-values for X_3. There will be both an inductive and a capacitive solution when

$$\left(\frac{R_A^2+X_A^2}{R_AR_s}-1\right) > \frac{X_A^2}{R_A^2}$$

or

$$R_A > R_s. \quad (9)$$

But when $R_A < R_s$, both solutions of X_3 are reactances of opposite type to X_A.

2.2 Series Arm X_2

From the imaginary part of (4), after eliminating X_3 by (6) and setting $X_s = -X_1$ by (3) and (1):

$$X_2 = -\left[X_1 \pm \sqrt{R_s(R_e-R_s)}\right]. \quad (10)$$

The choice of either the plus or minus sign in both (6) and (10) supplies a matched pair of values. The "minus" pair is referred to as $X_2'-X_3'$ and the "plus" pair as $X_2''-X_3''$.

2.3 Simplification of the Equations for X_3 and X_2

For values of Z_A not in the vicinity of quarterwave resonance or its odd multiples

$$R_e - R_s = \frac{R_A^2+X_A^2}{R_A} - R_s \cong \frac{X_A^2}{R_A}$$

usually is a very good approximation.

In that case

$$X_3 \cong \frac{X_A}{-1 \pm \sqrt{\frac{R_A}{R_s}}} \quad (11)$$

$$X_2 \cong -\left[X_1 \pm \frac{X_A}{\sqrt{\frac{R_A}{R_s}}}\right]. \quad (12)$$

Equations (11) and (12) reduce the labor of numerical calculations substantially over a wide range of frequency.

3. Impedance Loci and Geometrical Design Procedure

3.1 The X_3 Loci

In section 2, the solution of pi networks to transform specified antenna impedances into a resistance R_p has been obtained. The converse problem of determining what values of Z_A can be matched by a given range of X_3 is also of great practical importance when a wide range in antenna impedance is involved.

3.1.1 The P_1 Parabola. Equation (5) can be written as

$$R_A^2 - \frac{R_A}{R_s} \cdot X_3^2 + (X_A+X_3)^2 = 0$$

or

$$\left(R_A - \frac{X_3^2}{2R_s}\right)^2 + (X_A+X_3)^2 = \left(\frac{X_3^2}{2R_s}\right)^2. \quad (13)$$

This circle in R_A-X_A co-ordinates represents the locus of all Z_A's that require the value of X_3 given by the negative of the ordinate of the center (Fig. 2). It is tangent to the X_A-axis and has the center at

$$R = \frac{X_3^2}{2R_s}, \qquad X = -X_3.$$

Eliminating the parameter X_3, there follows

$$X^2 = 2R_s \cdot R. \quad (14)$$

The centers of all impedance circles (13) lie on this parabola P_1 of (14), which is fixed as long as R_s is constant.

Although X_3 is unchanged, X_2 varies in accordance with (10) as Z_A moves around any of the circles (13).

3.1.2 Graphical Solution for X_3. In order to solve (6) graphically, it is only necessary to find the circles (13) which pass through the given Z_A. Their centers lie on a parabola $\overline{P}_2$

$$(X - X_A)^2 = 2R_A \cdot \left(R - \frac{R_A}{2}\right) \quad (15)$$

since they must be equidistant from the X-axis and the point Z_A. Consequently, the two points of intersection of P_1 and $\overline{P}_2$ (Fig. 2) are the centers of the two circles which pass through Z_A and thereby determine the two possible values of $(-X_3)$. Actually, parabola $\overline{P}_2$ need not be drawn, since the two desired centers can be located very rapidly on P_1 by trial and error with the help of a compass.

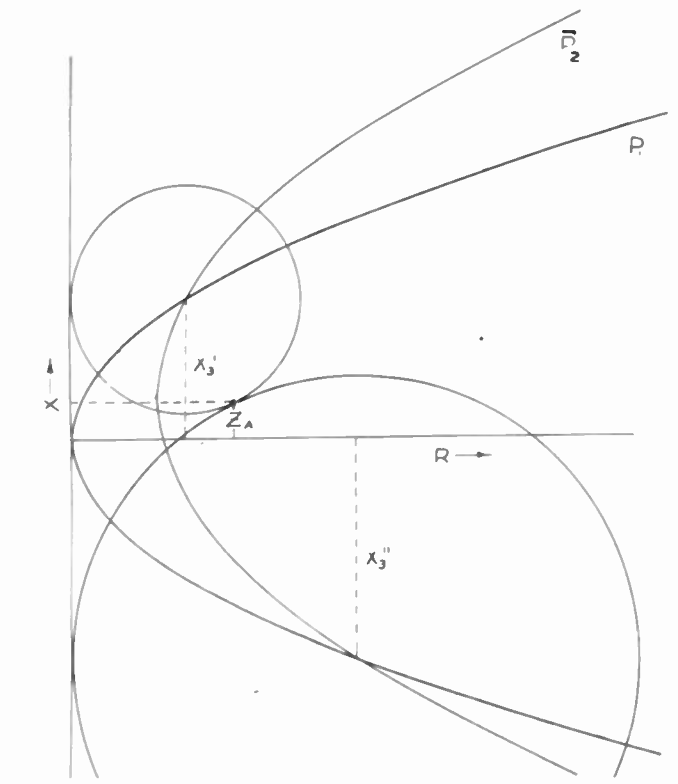


Fig. 2—Geometrical properties of X_3 solution.

3.1.3 Matching Region of Z_A for X_3 Variable Between Specified Limits. From (13) can be deduced the region of the Z_A plane which can be matched properly when X_3 is varied between assigned limits. It consists of the area swept out by the circles (13) as the parameter X_3 varies from the lower to the upper limit.

Part of the boundary is an arc of circle C_1 (Fig. 3), to which all circles C_2 of (13) are tangent. By (14), the directrix of P_1 is $R = -(R_s/2)$ and its focus is at $(R_s/2, 0)$. Since all C_2 circles are tangent to the X axis, $R_s/2$ short of the directrix, they also extend to within $R_s/2$ of the focus. Consequently, the circle C_1 with radius $R_s/2$ and centered at $(R_s/2, 0)$ is tangent to all the C_2 circles. Fig. 4 shows how to find the complete boundary by locating the transition (tangency) points $T_1 - T_4$ and joining the three circle arcs and a segment of the X axis; the shaded island between the intersecting arcs of the two C_2 limit circles is excluded from the matching region.

Since none of the C_2 circles crosses the boundary of C_1, its area is prohibited to Z_A. Since the equation of C_1 is

$$\left(R_A - \frac{R_s}{2}\right)^2 + X_A^2 = \frac{R_s^2}{4}, \quad \text{or}$$

$$R_A^2 + X_A^2 - R_sR_A = 0 \quad (16)$$

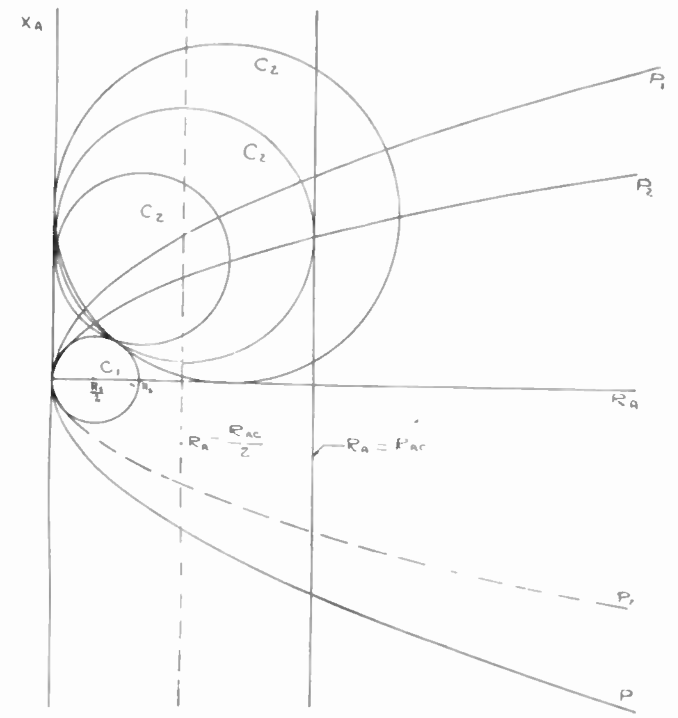


Fig. 3—The C_2 family of circles for X_3 and the P_2 locus.

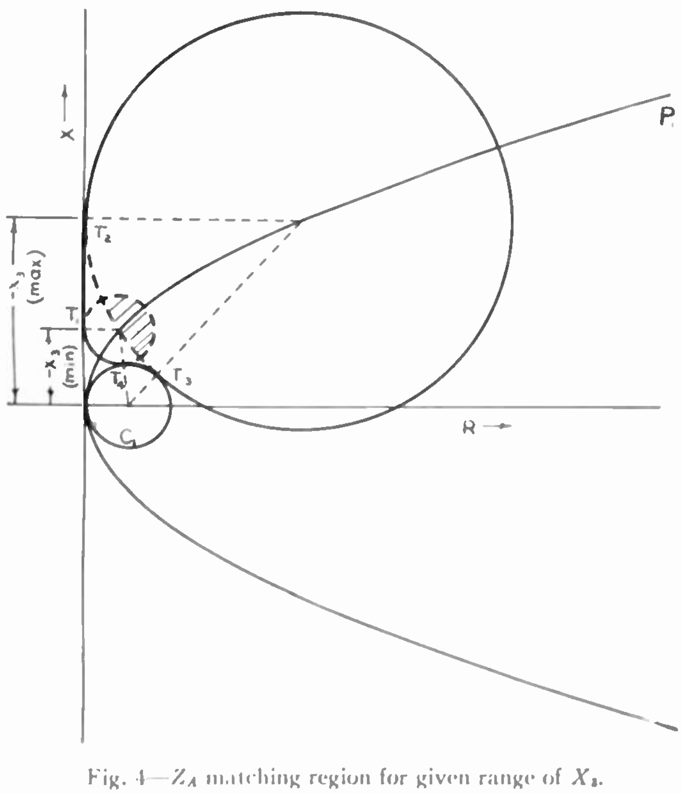


Fig. 4—Z_A matching region for given range of X_3.

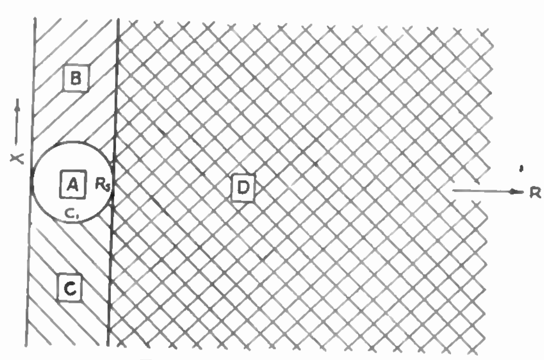
this restriction amounts to

$$\frac{R_A{}^2 + X_A{}^2}{R_A} - R_s \geq 0, \quad \text{or} \quad R_e \geq R_s. \tag{17}$$

This is identical with (8), represented graphically by the exclusion of Z_A from within circle C_1.

3.1.4 Matching Regions for Inductive or Capacitive X_3 Shunt Arm (Limits 0 and ∞). Evidently, letting X_3 vary from 0 to $-\infty$ in Fig. 4 should provide a graphical illustration of the totality of values Z_A may assume when X_3 is capacitive. The circle for X_3 min $=0$ is obviously the origin. The $X_3 = \pm\infty$ circle reduces to the straight line $R_A = R_s$, which follows by setting the denominator of (6) equal to zero.

Therefore, any Z_A located in regions B and D of Fig. 5 can be matched by a capacitance of required value in the X_3 arm. In the inductive case, the matching area consists of regions C and D, which is obtained by reflection about the axis of reals. This completes the geometrical representation of (6), (8) and (9) concerning the X_3 arm.



REGION	FOR EVERY Z_A, X_3 HAS
A	NO MATCH POSSIBLE
B	TWO CAPACITIVE SOLUTIONS
C	TWO INDUCTIVE SOLUTIONS
D	ONE CAPACITIVE AND ONE INDUCTIVE SOLUTION

Fig. 5—Possible X_3 solutions.

3.1.5 Characteristics of the X_3 Arm. In terms of Fig. 5, X'_3 is capacitive in B and D and inductive in C, while X_3'' is capacitive in B and inductive in C and D. On the boundary of C_1, there is only one solution $X_3 = (R_A{}^2 + X_A{}^2)/-X_A$ since $R_e = R_s$. Along $R_A = R_s$ one solution is $X_3 = (R_A{}^2 + X_A{}^2)/-2X_A$ and the other is ∞ (an open circuit).

3.2 The X_2-Loci

3.2.1 Graphical Solution for X_2 (C_{12} Circles). Evidently, X_2 in (10) is constant when R_e is fixed, X_1 and R_s being constants for a given problem. Writing (7) as

$$R_A{}^2 + X_A{}^2 - R_A R_e = 0$$

or

$$\left(R_A - \frac{R_e}{2}\right)^2 + X_A{}^2 = \frac{R_e{}^2}{4} \tag{18}$$

it is seen to be the equation of a circle, which has a diameter R_e, passes through the origin, and has its center on the R_A axis (Fig. 6). From similar triangles, as it is

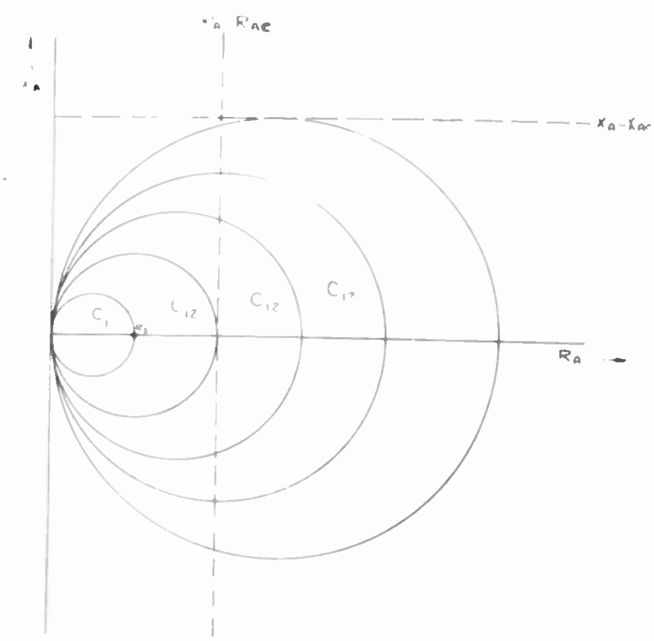


Fig. 6—The C_{12} family of circles for X.

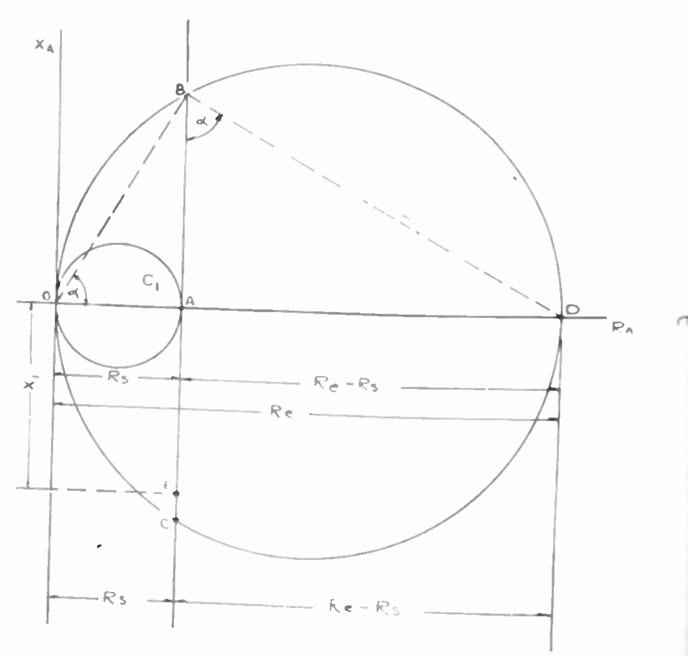


Fig. 7—Geometrical properties of X_2.

shown in Fig. 7.

$$\sqrt{R_s \cdot (R_e - R_s)} = \overline{AB} = |\overline{AC}|. \qquad (19)$$

Consequently, the two solutions of X_2 can be identified as

$$X_2' = \overline{EB}, \qquad X_2'' = \overline{EC}. \qquad (20)$$

The X_2 values for a given Z_A can be located graphically as follows:

(a) Find the C_{12} circle which passes through the given Z_A. The bisector of the Z_A vector intersects the R_A axis at its center.

(b) Move around this circle to locate the two points of intersection with the line $R = R_s$ (points B and C in Fig. 7).

(c) The X_2 values are the vertical segments along $R = R_s$ measured from $X = X_1$ (E) to these two points. X_2' is directed upwards and is inductive. X_2'' may be inductive or capacitive, depending on C being above or below E. It is zero (a short circuit) when C coincides with E.

3.2.2 Characteristics of the X_2 Series Arm. Important conclusions can be drawn from (10) and Fig. 7. X_2' is an inductive reactance larger than $|X_1|$ for any value of Z_A, while X_2'' may be inductive, a short circuit, or capacitive. When X_2'' is zero (short circuit), the equivalence of Fig. 1(a) and Fig. 1(c) requires that

$$R_e = R_p \qquad (21)$$

must hold. This also follows from a comparison of the first part of (4), $X_2 = 0$, with (2). Therefore, the $\overline{C}_{12}$ circle, for which $R_e = R_p$, constitutes a border line for values of X_2''. Inside, $R_e < R_p$ and X_2'' is inductive by (10), while outside of it X_2'' is capacitive since $R_e > R_p$. As a practical consideration, it is desirable to restrict the choice of the $X_2'' - X_3''$ pair to Z_A values inside the $\overline{C}_{12}$ circle. Otherwise, the X_2 arm may require excessively large capacitances for Z_A's just outside of the $R_e = R_p$ circle. The limit circle C_1, from which Z_A is excluded by (17), turns out to be the smallest allowable C_{12} circle with $R_e = R_s$.

3.2.3 Matching Region of Z_A for X_2 Variable Between Specified Limits. Any Z_A lying within the crescent-shaped area between the two C_{12} circles corresponding to X_2 min and X_2 max can be transformed into R_p, provided X_3 takes on the necessary values. It follows from the previous section that when X_2 may be any inductive reactance, $0 < X_2 < \infty$, any Z_A in the right-half plane outside the C_1 limit circle can be matched properly. Also, when X_2 may be any capacitive reactance, but is not allowed to become inductive, $0 > X_2 > -\infty$, any Z_A outside the $\overline{C}_{12}$ circle $(R_e = R_p)$ may be matched properly.

Knowing how to obtain the matching regions when X_2 and X_3 separately vary between specified limits, any Z_A lying within the area common to both of the regions can be matched properly without X_2 or X_3 exceeding its limits.

4. Minimum and Maximum Values of X_2 and X_3

When numerous values of Z_A are involved, their plot on the impedance plane will usually form a series of curves or cover a certain area. The extreme values of X_2 and X_3 can be deduced directly from each pair of C_2 and C_{12} circles tangent to it from the outside and inside (or passing through the extreme tip in case of a curve).

This geometrical approach will be outlined in connection with the rectangular grid in the Z_A plane. Lack of space does not permit including the complete derivation and other interesting aspects of this method.

4.1 Impedance Locus: $Z_A = R_{A_c} + jX_A$ (R_{A_c} fixed)

The C_{12} circle tangent to this vertical line is determined by $R_e = R_{A_c}$ (Fig. 6). Inserting this value in (10) leads to L_2' min and L_2'' max or C_2'' max, depending on $R_{A_c} < R_p$ or $R_{A_c} > R_p$, in terms of the operating frequency. When $R_{A_c} < R_s$, the smallest R_e occurs at the intersection with circle C_1 and is R_s, so that $\omega L_2'$ min-$\omega L_2''$ max $= -X_1$.

Since the C_2 circles tangent to the Z_A locus must also be tangent to the X_A axis, the abscissa of the center is $R_{A_c}/2$ (Fig. 3). Therefore, by (14) and the geometry of the figure

$$X_3 = \pm\sqrt{R_s R_{A_c}} = -X_A. \qquad (22)$$

Letting R_{A_c} vary from 0 to ∞, the Z_A values requiring minimum $|X_3|$ lie on parabola P_2 (Fig. 3) by (22):

$$X_A{}^2 = R_s R_A. \qquad (23)$$

C_3' max is obtained from the upper branch and L_3'' min from the lower one. Also, C_3'' max and L_3' min in the strip $R_A < R_s$ occur at the intersection with the upper and lower halves respectively of the C_1 circle.

4.2 Impedance Locus: $Z_A = R_A + jX_{A_c}$ (X_{A_c} fixed)

From Fig. 6, R_e min$/2 = |X_{A_c}|$ when $|X_{A_c}| > R_s/2$, which leads to an extreme value for L_2' and L_2'' or C_2''. For $|X_{A_c}| < R_s/2$, R_e min $= R_s$.

The pertinent $|X_3|$ min values follow from the tangency equations

$$\frac{X_3{}^2}{2R_s} = X_{A_c} + X_3, \quad \text{for} \quad X_{A_c} > 0 \qquad (24)$$

$$\frac{X_3{}^2}{2R_s} = -(X_{A_c} + X_3), \quad \text{for} \quad X_{A_c} < 0 \qquad (25)$$

since the vertical distance from the line $X_A = X_{A_c}$ to the center of a tangent C_2 circle is equal to its radius $X_3{}^2/2R_s$.

Letting X_{A_c} vary from $-\infty$ to $+\infty$, the Z_A's requiring $|X_3|$ min lie on two parabolas P_3 and P_4 (Fig. 8),[2] such that the solidly drawn segments apply to C_3' max, and their reflections about the axis of reals to L_3'' min.

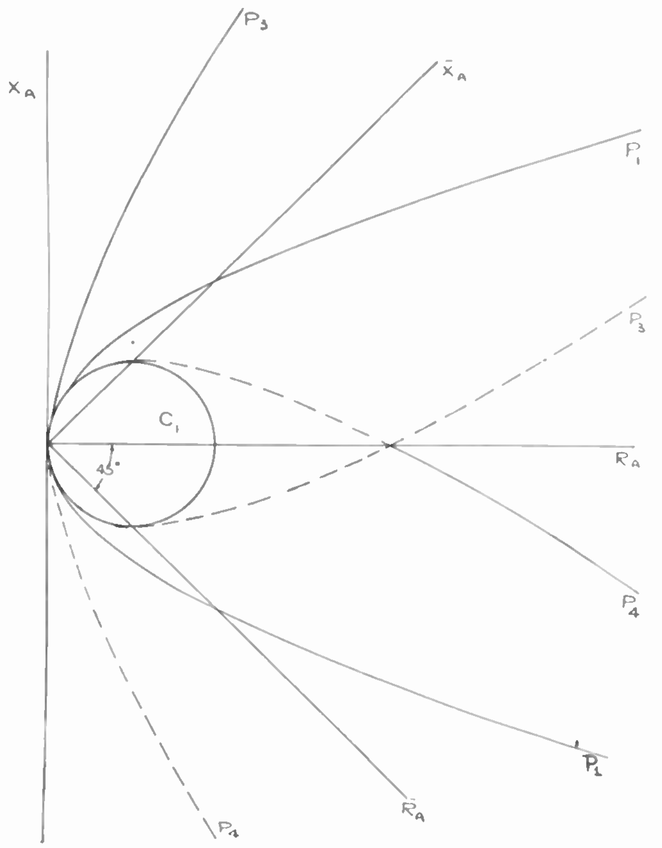


Fig. 8—The parabolic loci P_3 and P_4.

A vertical line through a point on these loci with ordinate X_{A_c} intersects the applicable branch of P_1 at $|X_3|$ min, valid for the whole Z_A locus $X_A = X_{A_c}$. The apex of P_3 is at $(R_s/8, (-3)/8\ R_s)$, its focal distance is $R_s/4\sqrt{2}$, and its axis of symmetry is inclined at 45° with respect to the axis of reals. P_4 is the reflection of P_3 about the axis of reals.

C_3'' max and L_3' min are obtained on approaching the X_A axis, and X_3 min $= -X_{A_c}$.

Since any area in the Z_A plane can be viewed as a bounded portion of the rectangular grid, it is instructive to be familiar with the behavior of X_2 and X_3 along the Z_A loci $R_A = R_{A_c}$ and $X_A = X_{A_c}$.

[2] Derived by substituting $X_3 = \sqrt{2R_sR_A}$ in (24) and (25), squaring, and transforming to principal axes.

5. Conclusions

Complete correspondence is established between the behavior of the pi-network antenna coupler and a set of simple geometrical figures. This reduces the design process to an equivalent problem in elementary analytic geometry, with the great advantage that the behavior of all design factors under any imposed set of conditions can be clearly visualized and studied with ease. Thus the way is opened to an optimum design in terms of the requirements (desired pi-network configuration, value of Q_L, restrictions on size and ratings of components, etc.). Even when numerous antenna impedance values are involved, numerical calculations are reduced to a minimum since the values required for the components follow directly from the tangency points of the geometrical representation. This contrasts very favorably with the laborious, point-by-point computations of the conventional design procedure, which, in addition, offers no assurance of obtaining the complete range of the components or devising the most economical design.

Equations (1), (3), (6), (7), and (10) with the approximate expressions (11) and (12) provide, in convenient form, the means for calculating and checking the reactance values where required.

Due to the lack of space, only the basic aspects of this geometrical technique could be presented. The author hopes to cover ramifications and some interesting applications to aircraft antennas in another paper.

Correction

It has been brought to the attention of the editors that an inadvertant error was made in the professional affiliations of Ronold King and K. Tomiyasu, authors of the paper, "Terminal Impedance and Generalized Two-Wire-Line Theory," which appeared in the October, 1949, issue of the Proceedings of the I.R.E. Dr. Tomiyasu has joined Sperry Gyroscope Company, Great Neck, L. I., N. Y., leaving Harvard University, where Dr. King is still engaged as a professor of applied physics.

The Application of IF Noise Sources to the Measurement of Over-all Noise Figure and Conversion Loss in Microwave Receivers*

LESLIE A. MOXON†

Summary—**The "noise diode" technique is now well-established for the absolute measurement of receiver noise figure at frequencies up to about 300 Mc. A method is described of extending this technique to much higher frequencies, using a frequency changer to produce the required rf test signal from an if noise source. An experimental procedure has been developed which enables the signal level to be accurately evaluated, subject to such limitations as those of Dicke's reciprocity theorem which are of little practical significance for the applications so far considered. As in the low-frequency case, measurements are in general independent of receiver bandwidth, there is no stray radiation problem, and the receiver output indicating device may follow any law within wide limits.**

The necessary components are easy to construct and can be calibrated without the use of additional apparatus. The system is particularly well suited to mixer crystal measurements since both conversion loss and over-all noise figure can be measured with equal facility, and the latter is readily analyzed in terms of conversion loss and noise ratio.

I. Introduction

The temperature-limited diode is now well-established as a method of measuring receiver noise figure at frequencies up to about 300 Mc, and possesses the merits of extreme simplicity and high accuracy. The diode anode current flows through a resistance R and (the diode impedance being normally much greater than R) the combination forms a signal generator of output impedance R and available power $eIBR/2$ where e is electron charge, I the anode current, and B the receiver bandwidth. If I is given the value required to double the noise power in the receiver, the receiver noise factor defined with respect to a temperature of 290° K, is given by the well-known expression

$$N = 20IR. \tag{1}$$

Some success[1] has been obtained in extending this technique up to about 3,000 Mc but suitable diodes are difficult to construct, corrections have to be made for transit time, and the absolute accuracy so far obtained is relatively poor. This paper describes a new technique of measurement[2] which has been evolved using an if noise source in conjunction with a mixer to provide an accurately known microwave rf signal for measurement of noise figure and mixer conversion loss.

The basic principle is as follows. Given two identical mixers, a known amount of if noise can be converted to rf and back again to if, the if output of the second mixer being measured by means of another noise generator connected across the output terminals to provide a reference signal. The total decibels loss in the double frequency-changing process being thus determined, it is merely necessary to divide by two to find the loss in either mixer subject to the assumption of reciprocity which will be justified later. From the losses in the first mixer and the known if noise level we can determine the rf noise level. Instead of two identical mixers, any three mixers may be used and the total loss measured for each of the three possible combinations; solution of the three equations gives the loss in each of the mixers. In practice, variations between mixers are almost entirely due to crystals so that it is only necessary to interchange crystals, and the full procedure is necessary only for the initial calibration, after which either noise-factor or conversion loss can be determined from a single reading of diode current.

There are certain practical difficulties to be overcome; for example, attenuation must be inserted between the mixers to enable them to be adjusted independently and, (preferably assisted by directive feeds) to confine each local oscillator to its own mixer. The amount of attenuation required is fairly small (10–15 db) but makes it advisable to use an amplifier between the first noise diode and mixer. To simplify the theory, the two local oscillators are tuned twice the if apart so that a single noise sideband is used.

This technique retains in comparison with continuous-wave signal generator methods of noise figure measurement many of the main advantages of the lower frequency diode noise sources; for example, there is no stray radiation problem, accurate knowledge of bandwidth is unnecessary, and difficulties connected with the indication of receiver output are minimized. The amount of attenuation required, and therefore the percentage accuracy with which it must be measured, is an order of magnitude less than in the cw signal generator case.

Another feature claimed for this technique is the readiness with which it lends itself to improvization. Given a receiver to be measured for noise factor, there is no need to have test equipment specially developed for

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[1] R. Kompfner, J. Hatton, E. E. Schneider, and L. A. G. Dresel, "The transmission-line diode as a noise source at centimetre wavelength," *Jour. IEE*, vol. 93, part IIIA, pp. 1436–1443, no. 9; March–May, 1946.

[2] Described in various unpublished reports dating from 1944, and briefly mentioned in an earlier publication (L. A. Moxon, *Wireless World*, January, 1947).

the purpose; the main requirements are a spare mixer and local oscillator, and the other items are easily constructed if not already at hand.

II. Typical Layout and Experimental Procedure for Precision Measurements

Fig. 1 illustrates an actual arrangement intended primarily for 10-cm mixer crystal measurements, but read-

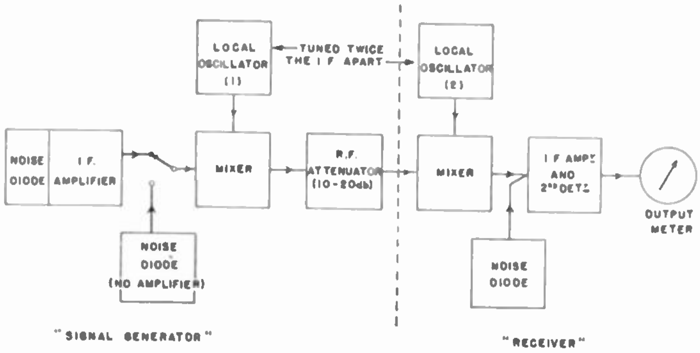

Fig. 1

ily adaptable for measuring the noise factor of other receivers, traveling-wave tubes, etc. In addition to the basic essentials outlined above, provision was made for the measurement of mixer noise ratio and if impedance. The rf circuitry employed waveguides throughout. The local oscillators were of the Heil type (CV230) and coupled to the mixers through variable attenuators and simple directive feeds giving about 6 db of directivity. The mixers were of a conventional waveguide type with double slug tuning and had a bandwidth of about 80 Mc for a standing-wave ratio better than 2.0 with most crystals. The if for both the receiver and the noise amplifier was 13.5 Mc the bandwidths being of the order of 1.5 and 10 Mc, respectively. The if noise figure was approximately 2 db, and the local oscillator contributed noise of the order of 0.05 to 0.2 units of crystal noise temperature ratio.

The Noise Amplifier

Let us first consider the case when the switch is in the lower position, i.e., no amplifier between noise source and mixer. The noise figure of the receiver is given by the usual expression

$$N_{rec} = L_{m2}(N_{if} + t - 1), \tag{2}$$

where N_{if} is the if noise factor, t is the mixer noise temperature ratio, and L_{m2} the conversion loss. A noise figure N' may be assigned to the entire system to the right of the switch, and is obviously given by

$$N' = N_{rec}L_{m1}L_{a}, \tag{3}$$

where L_{m1} is the conversion loss of the first mixer and L_a is the attenuation inserted between the mixers. If the noise source doubles the noise output power P_{no} of the receiver when adjusted to a current I_1, N' is given by $20I_1R$ in accordance with (1). The CV172 "noise diode" has frequently been run for short periods at 50 milliamps or more, and a load of about 400 ohms provides an impedance match for the average crystal; this corresponds to a value of 400 for N'. Lowest practicable values of N_{rec} and L_{m1} are of the order of 7 and 4 respectively, giving a permitted value of anything up to 400/28 or 11.5 db for L_a; this was just about sufficient from the point of view of preventing interaction between the mixers, and it was possible to work with a smaller change of receiver noise output, but there was not much in hand for dealing with anything short of the best possible performance, and an amplifier was therefore included. Using selected crystals it was possible to measure the insertion power gain G_p directly by operation of the switch, but in less favorable circumstances it might be necessary to resort to more elaborate measures.

The bandwidth of the noise source should of course exceed that of the receiver but difficulties may be encountered if it is unduly wide; thus with the amplifier out of circuit an appreciable error was found due to noise generated at three times the desired intermediate frequency, which produced a sideband at the receiver image frequency. This effect was eliminated by means of a 40.5 Mc series-tuned circuit across the noise diode load, but was not observed at all on another occasion when working with a 45 Mc if, due presumably to the greater shunting effect of the mixer rf by-passing capacitance.

Two types of noise amplifier have been used, the first consisting of a single stage having an anode load of 400 ohms to match the crystal impedance. The noise diode was connected in parallel with the grid circuit, which was also loaded with 400 ohms, the value in this case being determined by bandwidth considerations. The second type, with which most of the measurements have been made, had similar input and output arrangements but used 3 tubes in an inverse feedback circuit employing a 65-ohm resistance common to the cathodes of the first and last stages, the voltage gain being very approximately the ratio of the resistances (400/65) and therefore relatively stable. The internal noise level of the feedback amplifier was rather high, equivalent to 3 ma of noise diode current, but this was easily measured and allowed for. Although the feedback circuit was satisfactory at 13.5 Mc, difficulties arose at higher frequencies owing to phase shifts caused by stray capacitances.

The Attenuator

The attenuator consisted of two wedge-type sections, one variable and one fixed, calibrated by direct measurement of insertion loss. Attenuation should be sufficient to prevent the tuning of one mixer from affecting the other, and to ensure that the crystal current produced in either mixer by leakage from the local oscillator of the other does not exceed about 10–20 microamperes. If the latter condition is not satisfied, the second local oscillator may become appreciably modulated with if noise via the first mixer, and errors can also be caused by noise-modulated power from the first local oscillator reaching the second mixer and combining there with the two noise sidebands to produce unwanted if output. Additional attenuation acts against the leakage power as well as against the various noise sidebands, so that the unwanted effects decrease more rapidly than the wanted noise and a simple check is therefore to increase the local oscillator power and remeasure the apparent insertion loss of the attenuator which should be unchanged. Most of the 10-cm measurements were made with L_a equal to 14.5 db and with 6 db of directivity in each of the local oscillator feeds, the leakage effects being undetectable under these conditions.

Evaluation of Conversion Loss and Noise Factor

In order to evaluate the conversion loss it is necessary to know the available if noise output power P_n from the second mixer. If R_{m2} is the if output resistance of this mixer, and the noise diode connected across it is adjusted to a current I_2 such that it gives the same increase of receiver output as the current I_1 of the diode feeding the first mixer, P_n is proportional to I_2R_{m2}, and defining mixer conversion loss as available power output/actual input power, we have the relationship

$$L_{m1}L_{m2}L_a = I_1R/I_2R_{m2}. \qquad (4)$$

Perfect matching has been assumed between each mixer and the attenuator, and also between the first mixer (if impedance R_{m1}) and the load resistance of its noise diode. The easiest way to make L_{m1} equal to L_{m2} is to select crystals by comparison in the first mixer. Let L_{m1} can then be evaluated from (4).

Adjusting I_1 to double P_{no}, we can substitute $20I_1R$ for N' in (3) and obtain

$$N_{rec} = 20I_1R/L_{m1}L_a. \qquad (5)$$

It was found that adjustment of the mixers for maximum power output usually gave a standing-wave ratio better than 1.5, and taking $R=400$ ohms the value of R/R_{m1} or R/R_{m2} lay between 0.7 and 1.4 for normal crystals at 0.5 ma rectified current, so that mismatch errors are unlikely to be appreciable. On the other hand, ignorance of R_{m2}, since this quantity enters directly into equation (4), could lead to an error of ± 1.5 db in $L_{m1}L_{m2}$; i.e., with $L_{m1}=L_{m2}$ an error of 0.75 db in L_{m1}, and therefore in noise-figure measurements, would be likely. Measurement of R_{m2} is therefore necessary. It is to be noted that L_{m1}, L_{m2}, R_{m2}, and to some extent the noise ratio, are dependent on the rf impedance presented to the mixer, not only at the signal but also at the image frequency.[3] It is therefore desirable to have a good match at both frequencies, and the appropriate value of of if should be employed in measuring R_{m2}, dc or If bridge methods being unsuitable for accurate work.

If it is required to make measurements on, say, a number of crystals, the full process described above need only be carried out once. From inspection of (5) we see that the value of I_1 required (for example) to double the total noise power in the receiver is directly proportional to L_{m1} and to N_{rec}. I_1 is also proportional to L_{m2} if t is constant, but changing the crystal in the second mixer alters t as well as L_{m2}, and the simplest procedure is therefore to measure L_{m1} for any crystal, the value for other crystals being determined by comparison in the first mixer. Similarly, to determine N_{rec} the crystals are compared in the second mixer. If a noise amplifier is used, I_1 is of course replaced by the product of the power gain G_p and the observed current. L_{m2} may be obtained from (2), after t and N_{if} have been determined separately, but this is a more complicated procedure and, for a given crystal, L_{m1} should be equal to L_{m2}.

Measurement of the IF Impedance of the Mixer

Fig. 2(a) illustrates the if input circuit used in most

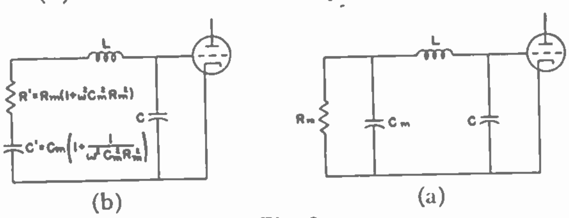

Fig. 2

of the measurements. At any given frequency it can be represented by the equivalent circuit of Fig. 2(b) by means of the following relationships:

$$R' = R_m(1 + \omega^2 C_m^2 R_m^2) \qquad (6)$$

$$C' = C_m(1 + 1/\omega^2 C_m^2 R_m^2). \qquad (7)$$

It is to be noted that C' and therefore the tuning point of the circuit, is dependent on both C_m and R_m, the slope dC'/dR_m being a maximum, for a given value of R_m, when $R_m=1/\omega C_m$. When the value of C_m was chosen to give minimum noise figure, R_m and $1/\omega C_m$ were found to be of the same order, and to make use of this fact for measurement of R_m it was only necessary to provide a calibrated tuning control on the input circuit, an if signal, and a sensitive resonance indicator.[4] For calibration, a range of small carbon resistors of known value was made up in crystal capsule form and inserted in place of the mixer crystal. The change of resonant frequency over the desired range of R_{m2} was small, but with care it was possible to measure R_{m2} to within about 5 per cent, leaving a possible error of the order of ± 0.1 db in the determination of L_{m1}.

It is not always possible to satisfy the input circuit conditions required by this method, and there is a simple alternative.[5] The crystal is replaced by a set of

[3] E. W. Herold, R. R. Bush, and W. R. Ferris, "The conversion loss of diode mixers having image frequency impedance," PROC. I.R.E., vol. 33, pp. 603–609; September, 1945.

[4] Method attributed to E. E. Schneider.

[5] H. C. Torrey and C. A. Whitmer, "Crystal Rectifiers," Radiation Laboratory Series, Vol. 15, McGraw-Hill Book Co., New York, N. Y., page 230; 1948.

resistances as in the previous method, and the incremental noise output power for some fixed value of diode current is observed for each resistance. The larger the resistance, the greater the increment, and a calibration curve may be drawn accordingly. Since mixer noise affects only the total output level and not the increment, observation of the latter for any mixer crystal enables its if impedance to be read from the calibration curve.

Noise Ratio Measurement

In a standard method of noise ratio measurement, the mixer is coupled to the first if tube by a circuit which gives a more or less constant output impedance as the input impedance is varied. The receiver noise output is then dependent only on the noise ratio and not on the impedance of the mixer, and is proportional to the quantity $(N_{if}+t-1)$. N_{if} is constant if the output impedance is constant and t can therefore be evaluated from observation of the change of noise output obtained on replacing the mixer by a resistance R_c, which need not be exactly equal to that of the mixer. The required conditions for this method should be satisfied for the circuit of Fig. 2(b) when $R_m=1/\omega C_m$, but an experimental check revealed some second-order anomalies attributable to using a small capacity across L as the tuning control, and to the effect of tuning on the phase relationship between tube shot noise and induced grid noise. With the tuning control adjusted for $R_c=400$ ohms and fixed, increase of R_c from 300 to 550 ohms caused an increase of N_{if} from 1.6 to 1.8 and a decrease of 10 per cent in output noise power, so that for accurate measurements by this method it was necessary to obtain a rough estimate of R_{m2} and apply corrections.

Alternatively, if R_{m2} is known accurately, we can increase the current I_2 of the diode connected across it until the total noise power is doubled; then if R_{m2} has a noise ratio t, we have

$$N_{if} + t - 1 = 20 I_2 R_{m2}. \qquad (8)$$

N_{if} is found by making $t=1$, i.e., replacing R_{m2} by an ordinary resistance, so that knowing I_2, R_{m2} and N_{if} we can obtain t.

Good agreement was obtained in practice between the two methods. It is of interest to note that if $(N_{if}+t-1)$ is determined by the first method, the value of R_{m2} can be deduced from (8).

Receiver Output Measurement

The second detector of the receiver consisted of a low impedance diode (Type VR92), and the rectified current was used as an indication of noise output giving a linear law over at least the desired range of 0.5 to 1.5 volts. The diode characteristic has proved stable over a period of years and holds for most tubes of the same type after ageing to reduce the standing current. The if amplifier itself was linear up to the highest output level required. Initially the diode was replaced by a thermistor bridge, but this was found to be an unnecessary complication.

The output indicating device can usually be calibrated, whatever its law (provided this is independent of gain setting), by means of the noise source using the following procedure suggested by L. A. G. Dresel. Let receiver noise alone give an output reading θ_1; to evaluate the noise factor N from (1) it is necessary to know I_x, the diode current which produces noise equal to the receiver noise. Let a current I_y increase the output to a reading θ_2 and let the gain then be turned down so that θ_2 is reduced to θ_1; if the current is now increased to a value I_z such that the output reading is again θ_2, the ratio r of I_x+I_y to I_x must be the same as that of I_x+I_z to I_x+I_y. This leads to the relation

$$I_x = I_y^2/(I_z - 2I_y). \qquad (9)$$

To obtain an accurate result from (9), $I_z/2I_y$ should not be less than about 2. This means that a value of at least 3 is desirable for r, and more diode current is required than if the detector law is already known, but this presents no difficulty with normal values of N_{if}.

III. Practical Results and Discussion

The double-mixing noise generator technique was first employed as a means of obtaining an accurate comparison of two receivers at 10 cms after difficulties had been experienced with alternatives which included crystal and klystron noise generators and cw signal generators. As a method of relative measurement it was found simple to construct and use, and comparisons with two cw signal generators gave agreement on absolute level within 1 db. At this stage the technique had not been fully developed and the probability of errors of the order of ±1 db would have been fairly high. The cw signal generators were certainly no more accurate than this, so the agreement may have been largely fortuitous. Working at 3 cms, G. Eichholz and T. J. Buchanan obtained agreement between the new technique and other methods of measurement within about 0.25 db, but this again may have been largely due to chance, and it was realized that various possible sources of error required investigation. For this purpose a 10-cm measuring equipment was set up as illustrated in Fig. 1 and already described, facilities being provided for thoroughly checking each stage of the procedure. These included standingwave measuring equipment for checking rf matching, and a resonant cavity for measurement (by elimination) of local oscillator noise. The procedure described above was evolved mainly on the basis of experience with this equipment.

In one typical experiment, two dozen crystals were measured for conversion loss in both the first and the second mixers, and comparison of the results showed a scatter of the order of ±0.5 db which it was at first thought might be due to reciprocity failure; the worst pairs of crystals were therefore taken, the over-all conversion loss was measured with one of the crystals in each mixer, and the crystals were then interchanged, and the measurements repeated. This was done several times with as much care as possible, and the total loss figures showed a spread of only about $\frac{1}{4}$ db. This means that, in general, reciprocity failure (if any) is not greater than $C \pm \frac{1}{8}$ db where C is the same for all crystals of the type tested (British "Red Dot" silicon crystals). The number of crystals was not large enough for all possibility of more serious variations to be excluded, but evidence from other sources[6] indicates (a) that for silicon crystals reciprocity failure is (by measurement) less than 0.2 db; (b) that any systematic or constant error C is inappreciable; and (c) for germanium crystals reciprocity does not hold in general, although even in this case it would appear that calibration would be reasonably accurate if carried out with crystals selected to have the lowest possible loss.

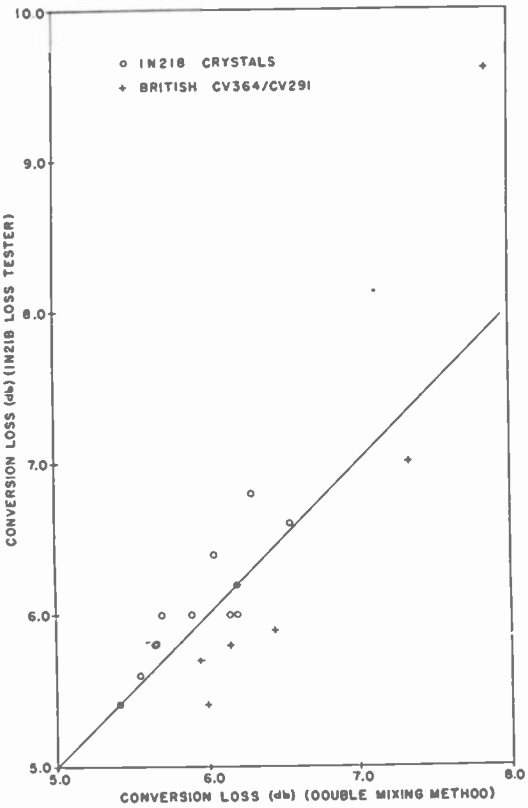

Fig. 3

Measurements of conversion loss by the above method on twelve 1N21B crystals and six British crystals which had previously been measured in one of the standard test sets[7] showed good agreement, as plotted in Fig. 3.

The crystal or diode mixer is commonly represented by an equivalent passive linear network, and reciprocity is implicit in any such representation. Reciprocity is also the basis of methods of conversion loss measurement based on the effect of if impedance on rf impedance or vice versa, such as that due to R. H. Dicke.[8] Dicke has pointed out, however, that full reciprocity holds only on certain assumptions, and a full discussion of this subject can be found elsewhere.[9] It is perhaps sufficient here to state that the necessary assumptions are not completely satisfied in practice, mainly because of variation of barrier capacitance with applied voltage. One would expect such an effect to be very variable from crystal to crystal, and the failure to observe any change in total loss when the crystals in the two mixers are interchanged confirms the more direct evidence, already mentioned, that reciprocity failure is of negligible practical significance for silicon crystals.

An apparent failure of reciprocity could arise from the practice of tuning the two local oscillators twice the if apart, assuming a lack of symmetry in the variation with frequency of the impedance presented to the rectifying contact of the crystal; this is because mixer conversion loss is dependent to some extent on the image frequency and (if appreciable) the harmonic frequency impedances seen by the nonlinear element. Lack of symmetry coupled with the staggering of oscillator frequencies means that these impedances can be different for the two mixers, even though the latter are physically identical. This effect should be detectable by comparing the over-all losses for the two possible tuning positions of each local oscillator in turn, but has not so far been observed.

Acknowledgment

This paper is published with the approval of the Lords Commissioners of the Admiralty, but the responsibility for any statements of fact or opinions expressed rests solely upon the author. The author is indebted to E. E. Schneider and L. A. G. Dresel, who played a large part in the development of the apparatus described.

[6] See pp. 125–127 and 212 of footnote reference 5.

[7] For the measurements in the 1N21 loss tester, and permission to publish them, the author is indebted to the Director, Naval Research Laboratory, Washington D. C.

[8] See page 203 of footnote reference 5.

[9] See chapter 5 of footnote reference 5.

The Effect of Antenna Size and Height Above Ground on Pointing for Maximum Signal*

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Summary—During April, 1947, the Electrical Engineering Research Laboratory of The University of Texas made measurements of the variation of signal strength and phase of 3.2-centimeter radio waves with heights up to 200 feet for a 27-mile desert path in Arizona.[1,2]

This paper presents the results of a study of the vertical angles-of-arrival which would be indicated by pointing antennas of various sizes and at various heights for maximum signal strength in three of these measured fields.

A comparison is also made of the response of the antennas for various angles of tilt in these measured fields with the response of the antennas in an assumed field made up of two plane waves components.

I. Introduction

THE PURPOSE OF this investigation is to study the effect of antenna size and height above ground on the apparent angle-of-arrival of the radio wave as indicated by pointing the antenna for maximum signal.

The study is made by determining the angle at which maximum signal is received as a function of antenna height and size for three field measured wave fronts. The three fronts selected for the study were cases in which low angle radiation had introduced additional wave components at the receiver due to trapping or earth reflections. The field measured data were taken for a wavelength of 3.2 centimeters over a 27-mile path at the Naval Electronics Laboratories' desert site near Gila Bend, Ariz.

The field measurements were made with the phase difference equipment built by the Electrical Engineering Research Laboratory of The University of Texas for measuring the difference in phase at 3.2-cm wavelength between the fields of two antennas separated vertically by ten feet and the signal strength at each antenna.[3] The procedure for obtaining the data involved raising the receiver vertically, while holding the transmitter at a fixed level.

II. Meteorological Considerations

The field measured data were taken such that each case analyzed represents a condition typical of a 24-hour meteorological cycle as observed at the desert site in Arizona.[4] Data for the 0021 curve, Fig. 1, were taken when a ground based duct was known to exist over the radio path. Data for the 0605 curve were taken when the duct was breaking up and meteorological conditions in general were irregular. Data for the 0857 curve were

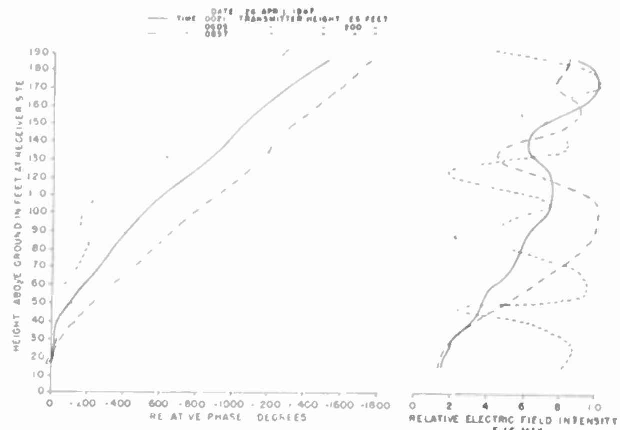

Fig. 1—Field measured phase-front and signal strength curves.

taken after the duct had disappeared and normal daytime conditions had set in. No attempt is made to correlate these meteorological observations with the indicated angle-of-arrival in this paper, except to note briefly the magnitude of the error in each case. This correlation is considered elsewhere.[1]

III. Method of Determining the Indicated Angle-of-Arrival

The method used is that of vector addition. This method consists of summing up the signal strengths at the appropriate phase angles along the face of the antenna for various angles of antenna tilt. The resulting signal strength is plotted against angle of antenna tilt, and from the plot, the angle which gives the maximum signal is determined. This angle of tilt is the indicated angle-of-arrival.

This method necessarily assumes that each portion of the wave front can be regarded as a secondary source or Huygen's source of known electric intensity, phase, and polarization and that the receiving antenna has transmitting characteristics such that it generates a wave

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† Electrical Engineering Research Laboratory, The University of Texas, Austin, Texas.

[1] A. W. Straiton, W. E. Gordon, and A. H. LaGrone, "A method of determining angle-of-arrival," *Jour. Appl. Phys.*, vol. 19, pp. 524–533; June, 1948.

[2] E. W. Hamlin and W. E. Gordon, "Comparison of calculated and measured phase differences at 3.2 centimeters wavelength," Proc. I.R.E., vol. 36, pp. 1218–1223; October, 1948.

[3] F. E. Brooks, Jr., and C. W. Tolbert, "Equipment for measuring angle-of-arrival by the phase difference method," Electrical Engineering Research Laboratory, University of Texas, Report No. 2; May, 1946.

[4] J. R. Gerhardt, "Meteorological measurements in Arizona during March and April, 1946," Electrical Engineering Research Laboratory, University of Texas, *Report No. 5*; February, 1947.

that is uniform in phase and magnitude in the plane of the face. As a receiver, equal signals at any two points along its face produce equal response. This method is described by Friis and Lewis.[5] In general, an antenna with variations in response to equal signals along the face would indicate approximately the same angle-of-arrival as a somewhat smaller antenna of the characteristics assumed above.

The antenna itself is assumed to have a rectangular opening, to be vertically mounted, and to have a width so small that horizontal variations in phase and signal strength in the wave front are negligible.

IV. Example of Method

Data: 0605 April 26, 1947. The data were taken from the phase front and signal strength curves in Fig. 1. A 100-foot antenna centered at 90 feet was assumed. The data consists of 50 signal strength magnitudes and relative time phase angles spaced evenly along the antenna face.

The solution is made by using the following equation:

$$E = \left[\sum_{1}^{N} E_n \cos\left(\Phi_n - \frac{360\, l_n \sin\theta}{\lambda}\right) + j\sum_{1}^{N} E_n \sin\left(\Phi_n - \frac{360\, l_n \sin\theta}{\lambda}\right)\right]\frac{L}{N},$$

where,

N = total number of field increments taken

E_n = relative electric field intensity of increment n (E_n/E_{max})

ϕ_n = time phase of E_n in degrees

l_n = distance in feet from the electrical axis of the antenna to E_n as measured along the antenna face

θ = antenna tilt angle. (For $\theta = 0°$, antenna face is vertical)

λ = wavelength = 0.105 ft. (3.2 cm)

L = vertical dimension of antenna face in feet

E = relative antenna response.

The curves in Fig. 5 show a plot of relative antenna response $|E|$, versus angle of antenna tilt, θ. The angle-of-arrival as indicated by maximum signal strength for the 100-foot antenna centered at 90 feet is $-0.050°$. This angle is shown plotted at 90 feet for the 100-foot antenna in Fig. 3.

V. The Indicated Angle-of-Arrival from Analysis of the Field-Measured Phase Front

The indicated angle-of-arrival versus height as obtained from the field-measured phase fronts is shown in curve form in Figs. 2, 3, and 4 for several antenna sizes. On each curve in this series, the angle-of-arrival of a single ray, arbitrarily positioned, is shown for comparison. The single-ray curve, for a homogeneous atmosphere and a path length of 27 miles is a straight line with a slope of 0.0406° per hundred feet.

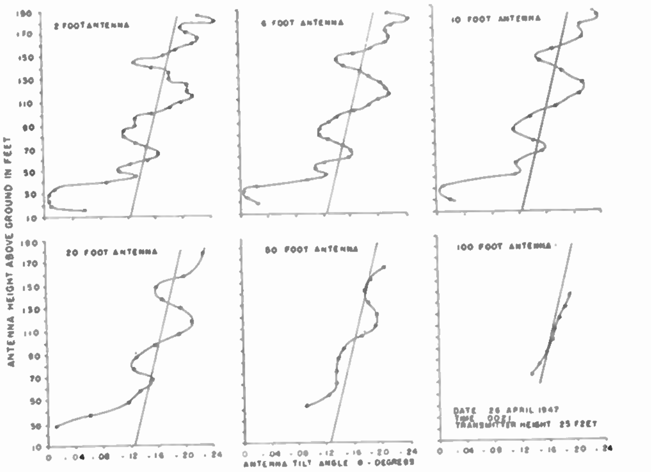

Fig. 2—Indicated angle-of-arrival at 0021.

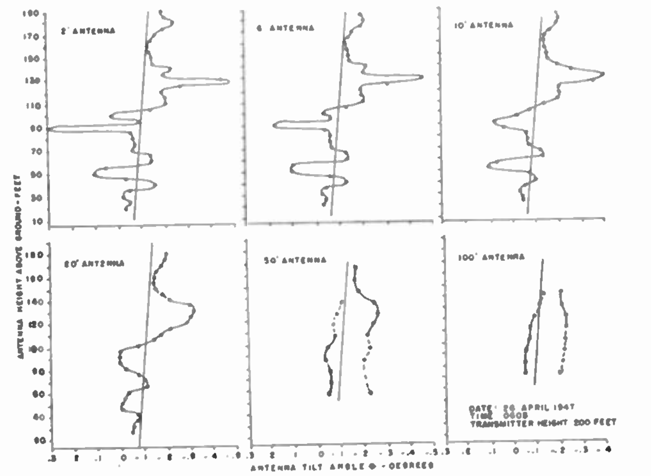

Fig. 3—Indicated angle-of-arrival at 0605.

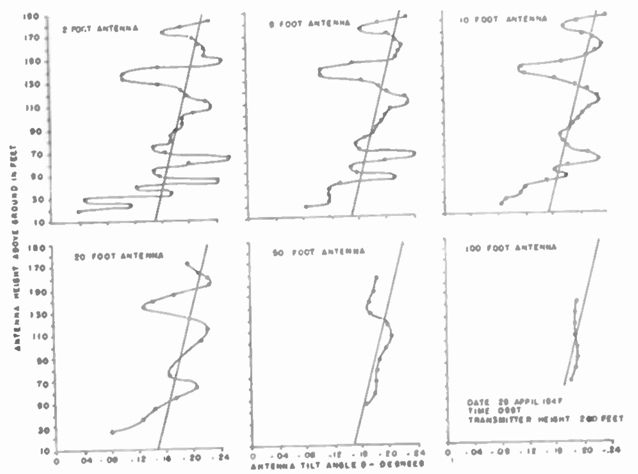

Fig. 4—Indicated angle-of-arrival at 0857.

[5] H. T. Friis and W. D. Lewis, "Radar antennas," *Bell Sys. Tech. Jour.*, vol. 26, p. 219; April, 1947.

A comparison of the field-measured and the single-ray indicated angle-of-arrival curves shows that the average slope of the field-measured curves agrees reasonably well with the slope of the single-ray curve, and that the field-measured curves fluctuate considerably with height. The fluctuation noted in the field measured indicated angle-of-arrival is evidence of the complexity of the wave front at the receiver.

In the field-measured cases, the greatest fluctuation in the indicated angle-of-arrival with height was observed in the two-foot antennas, and the least in the 100-foot antennas. This difference can be explained by the fact that the small antenna integrates over such a small portion of the wave front that any localized variation in phase has an important part in determining the tilt angle for maximum signal indication, and hence, the indicated angle-of-arrival. The large antenna, on the other hand, integrates over such a large portion of the wave front that localized variations are averaged out in the large signal received, and have little effect in determining the indicated angle-of-arrival.

Another point of interest in the field measured indicated angle-of-arrival is the sudden change in the angle observed in the 50-foot and 100-foot antenna curves in Fig. 3. (The indicated angle-of-arrival is shown by a solid line.) This phenomenon is not observed in any of the other antenna curves in this set nor in the sets shown in Figs. 2 and 4. This case is discussed in a later section.

VI. Error in the Indicated Angle-of-Arrival

The error in the field measured indicated angle-of-arrival, relative to the angle-of-arrival of a single ray in a homogeneous atmosphere, can be obtained by comparing the field-measured curves with the single-ray curves in Figs. 2, 3, and 4.

The error is shown tabulated in Table I for the different antenna sizes. The meteorological conditions which existed at the time the field data were taken are also shown. The error is calculated by taking an average of the two greatest deviations in the indicated angle-of-arrival, one above and one below, from the angle-of-arrival of the single ray.

Example: Fig. 2, two-foot antenna:

Greatest deviation above (at 30 feet) = 0.125°
Greatest deviation below (at 105 feet) = 0.050°
Error indicated (0.125+0.050)1/2 = ±0.088°.

TABLE I

ERROR IN THE INDICATED ANGLE-OF-ARRIVAL

Antenna Size L (Feet)	*Fig. 2 Ground Based Duct*	*Fig. 3 Transitional Period*	*Fig. 4 Normal Daytime*
2	±0.088°	±0.382°	±0.102°
6	±0.088°	±0.311°	±0.084°
10	±0.088°	±0.234°	±0.072°
20	±0.085°	±0.150°	±0.052°
50	±0.038°	±0.107°	±0.022°
100	±0.013°	±0.084°	±0.012°

VII. Antenna Response for Various Angles of Tilt in Measured Fields

1. *One Hundred-Foot Antenna.* The series of curves in Fig. 5 show in detail the signal strength received versus angle of antenna tilt of a 100-foot antenna for the field-measured wave front taken at 0605. The tilt angle at which maximum signal occurs is taken to be the indicated angle-of-arrival and the points are plotted at the indicated height on the 100-foot antenna curve in Fig. 3.

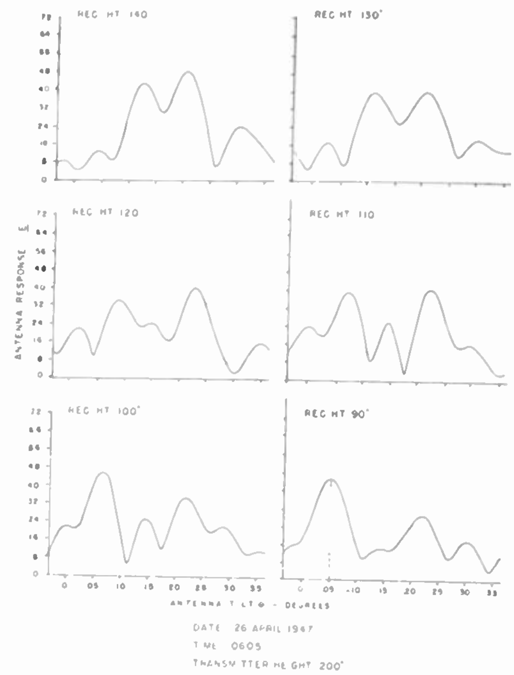

Fig. 5—One-hundred foot antenna patterns from field-measured phase front.

The curves in Fig. 5 show the interfering wave in the field-measured wave front which was suggested by the fluctuations in the indicated angle-of-arrival with height, Figs. 2, 3, and 4. The interfering wave appears as a strong second lobe in the antenna response pattern in this series of curves. In a similar study of the other two wave fronts, at 0021 and 0857, the interfering wave appeared relatively weak with respect to the main wave, yet strong enough to be identified as a second wave component and not as a minor lobe of the antenna response pattern.

The case shown by the curves in Fig. 5 is of particular interest because of the break observed in the angle at which maximum signal is received (solid line) in the 100-foot antenna curve in Fig. 3. A study of the curves in Fig. 5 shows the antenna maximum signal being determined by one wave components at the lower receiver levels and by another at the higher levels. This shift of the maximum signal from one wave component to the other is also observed to be abrupt, since both wave components are clearly defined at the height at which the change occurs. It should be noted that the plot in Fig. 3 includes the angle-of-arrival of each wave component for all antenna heights in this unusual case.

2. *Fifty-Foot Antenna.* The set of curves in Fig. 6 shows in detail the signal strength received versus angle of antenna tilt of a 50-foot antenna for the field-measured wave front taken at 0605. The tilt angle at which

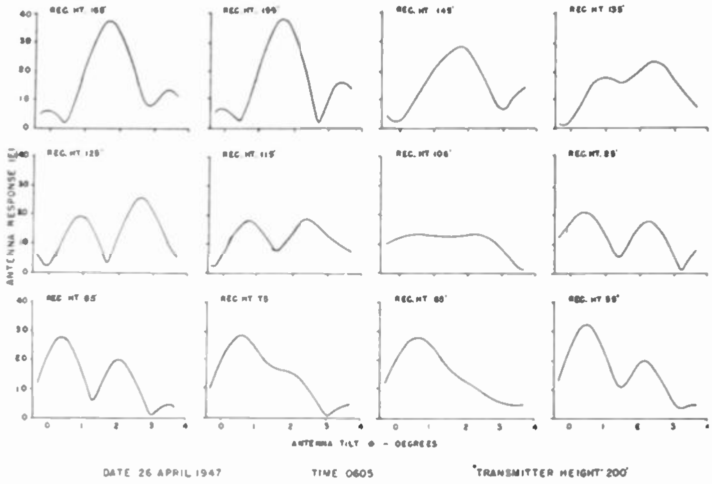

Fig. 6—Fifty-foot antenna patterns from field-measured phase front.

maximum signal occurs is taken to be the indicated angle-of-arrival and the points are plotted at the indicated receiver heights on the 50-foot antenna curve in Fig. 3.

As in the case of the 100-foot antenna analysis of this wave front, Fig. 5, these curves show two interfering waves of the same order of magnitude. The antenna maximum signal indication is again observed to shift from one wave component at the lower receiver levels to the other wave component at the higher receiver levels. The individual wave components are not as distinct at the point where the break occurs in this case as they were for the 100-foot antenna; however, the separate maxima are discernible.

The indicated angle-of-arrival curve for the 50-foot antenna in Fig. 3 includes the angle-of-arrival of both wave components for all antenna heights, where both were readable.

3. *Twenty-Foot Antenna and Smaller Antennas.* The small antennas which were used in the study of the wave fronts did not show a break in the indicated angle-of-arrival in the 0605 case as did the 50- and 100-foot antennas. A detailed study of the 20-foot antenna patterns did show evidence of the second wave component, but the antenna did not leave sufficient resolution to separate the components near the heights where their magnitudes were equal. The 10-, 6-, and 2-foot antenna patterns did not definitely show the second wave component.

4. *Discussion.* A study of the 0605 phase front curve, Fig. 1, for this unusual case suggests the possibility of two separate wave fronts, an upper and a lower, which merge at a height of about 100 to 110 feet. This is approximately the elevation at which the break in the indicated angle-of-arrival is observed in both the 50-foot and 100-foot antenna curves in Fig. 3. The two wave-front concept would also account for one signal being the stronger at elevations above 110 feet and the weaker below 100 feet.

VIII. Antenna Response at Various Angles of Tilt for an Assumed Field

For comparison with the field-measured wave fronts, an assumed front made up of two plane waves, α and β, was analyzed. The method of analysis was by vector addition. The result of the analysis is shown in curve form in Fig. 7. In the analysis, α and β were related as

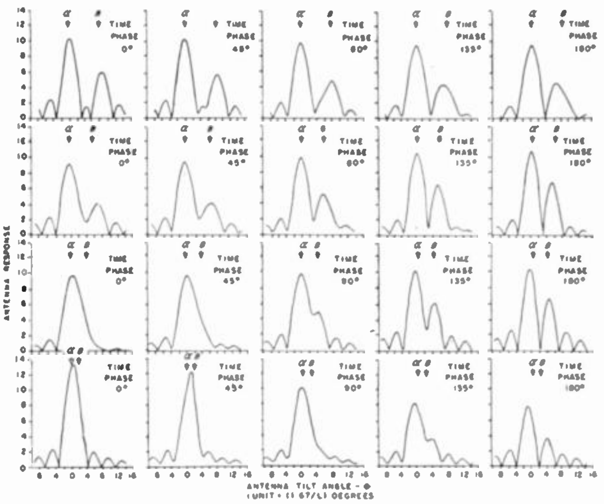

Fig. 7—Antenna response for phase front composed of two plane waves.

follows: the α wave was assumed to have a magnitude of 1.0 and to always arrive horizontally, i.e., at $\theta=0°$. The β wave was assumed to be one-half the magnitude of α and to arrive at some angle below horizontal (negative θ) as indicated by the position of the β arrow. The relative time phase of α and β was assumed as indicated on each curve. The antenna tilt scale is calibrated in terms of antenna size L, where L is the vertical dimension in feet; hence, the curves are not restricted to a given size antenna.

A study of the curves in Fig. 7 shows graphically that complex response patterns such as those in Figs. 5 and 6 can be obtained from two plane waves, and indicates the factors which determine the pattern. The effect of these factors on the indicated angle-of-arrival is also clearly shown. The controlling factors are found to be the relative time phase and the angular separation of the two wave components. Other factors, which would affect the indicated angle-of-arrival, but which were not shown directly by the analysis, would be the relative magnitude of the wave components and the introduction of further components.

IX. Comparison of the Theoretical Wave-Front Curves with the Field-Measured Wave-Front Curves

A comparison of the curves in Fig. 7 with those obtained from the field-measured wave fronts reveals a remarkable amount of similarity in some of the curves, and a few cases where the curves are almost identical. This fact strongly suggests that the wave components of the measured field were those of the assumed field. If this is true, then a field-measured wave front which yields a curve that agrees identically with a curve from an assumed wave front, will have wave components of the same relative magnitude, time phase, and angular separation as those of the assumed front. In this manner, wave components in a field-measured wave front could be separated.

X. Conclusions

1. For the field-measured data, antennas under 10 feet in size were found to be small enough to follow the localized height variation in phase; thus, the phase difference method (footnote reference (3)) agrees with the vector addition method in determining the indicated angle-of-arrival for these antennas.

2. The large antennas were found to be relatively insensitive to the localized phase variation, but were considerably affected by the major phase variations noted. For these larger antennas the phase difference method does not give the tilt angle which results in maximum signal and the vector addition method is necessary.

3. The effect of antenna size and height above ground on the angle-of-arrival as indicated by pointing for maximum signal is shown in curve form in Figs. 2, 3, and 4 and in Table I for three measured fields. The inaccuracies of this method of finding the angle-of-arrival of a radio wave are generally well known; however, this study, in addition to calculating the magnitude of some of these errors, shows how the error fluctuates with antenna size and height above ground.

4. Comparing the antenna signal strength response curves for field-measured data with similar curves obtained by analysis of an assumed wave front offers a possible means of separating the wave components in the field-measured wave front.

Design of Tunable Resonant Cavities With Constant Bandwidth*

L. D. SMULLIN†, ASSOCIATE, IRE

A PROBLEM of some interest in microwave circuit design is that of making a tunable filter whose loaded Q, or whose frequency bandwidth is constant over the tuning range. It is well known that a cavity using inductive irises as the coupling elements has a rapidly varying Q_L and that the bandwidth may change by almost a factor of 2 when the cavity is tuned over a 10 per cent range. It can be shown, however, that the use of capacitive irises results in a reasonably constant bandwidth.

If one computes the way in which the coupling susceptance should vary with frequency in order to make a cavity with constant loaded Q (Q_L) or constant bandwidth, one obtains the curves shown in Fig. 1. (The loading due to cavity losses has been neglected, which is reasonable for values of $Q_L \simeq 100-300$.) Here we have the abscissa proportional to λ/λ_c where λ is the free-space wavelength of the applied signal, and λ_c is the cut of wavelength of the wave guide. Obviously, a given value of loaded Q can be obtained by the use of either an inductive or a capacitive coupling iris, and these two branches are shown in Fig. 1. Also plotted in Fig. 1 are the curves showing the frequency dependence of waveguide inductive and capacitive irises. It is clear that a capacitive iris has a frequency dependence that will be in between that required for constant Q_L and constant bandwidth. The inductive iris, however, has a frequency characteristic of just the opposite slope from that required for constant Q_L or bandwidth.

Capacitive irises have the disadvantage that one can only get values of about $B/Y_0 \leq 10$ without going to very minute openings. Thus it does not appear possible to make cavities with Q_L greater than about 300 by this method.

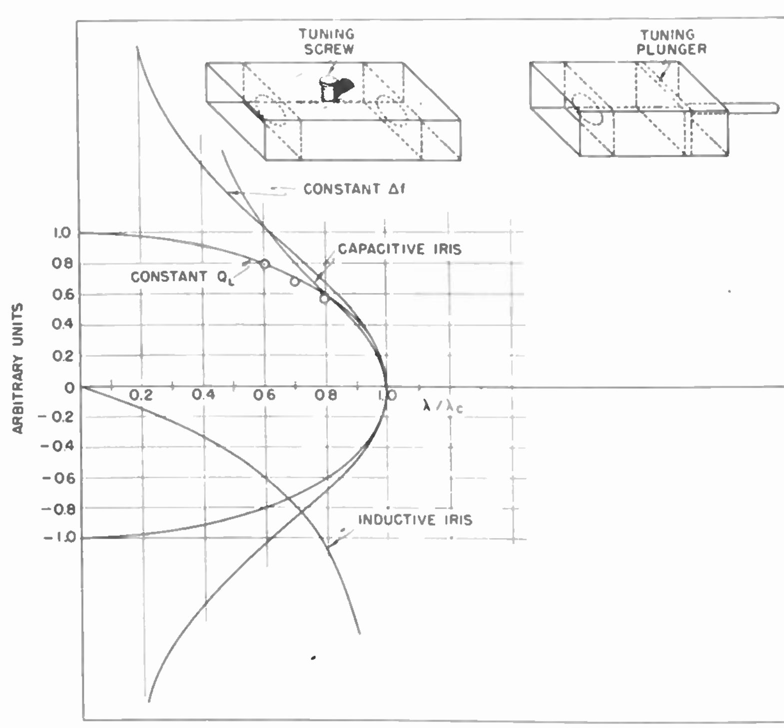

Fig. 1—Frequency dependence of coupling susceptance necessary to give constant Q_L or constant Δf.

* Decimal classification: R119. Original manuscript received by the Institute, June 13, 1949; abstract received, August 30, 1949. This work has been supported in part by the Signal Corps, the Air Materiel Command, and O.N.R., and is abstracted from Research Laboratory of Electronics Technical Report No. 106, April 5, 1949.

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Contributors to the Proceedings of the I.R.E.

R. H. Dishington was born on March 21, 1919, in Green Bay, Wis. He received the B.E.E.E. degree from the University of Southern California in 1942, and at that time entered the test course of the General Electric Co. Early in 1943 he did radio research for Bendix Aviation, Ltd., and taught night classes in electronics at the University of Southern California. In June of that year Mr. Dishington became a lecturer in electrical engineering at that University. Since the fall of 1947, he has been engaged in research work with The Rand Corp., Santa Monica, Calif.

R. H. Dishington

Mr. Dishington is an associate member of AIEE. He served on the electronics committee of the Los Angeles Section in 1943–1944. He is also a member of Eta Kappa Nu.

❖

Rudolf Kompfner was born in Vienna, Austria, on May 16, 1909. He attended the Realschule and Technische Hochschule in Vienna, and was graduated from the faculty of architecture in 1933. In 1934 he came to England to continue his study in architecture privately, and became the director of a building firm in 1937. Throughout these years, he devoted much of his spare time to television, radio, and physics.

Rudolf Kompfner

Mr. Kompfner entered the Admiralty service in 1941 as temporary experimental officer, taking up duty first in the physics department at Birmingham University. Since 1944 he has been associated with the Clarendon Laboratory at Oxford University, England.

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For a photograph and biography of W. W. Harman, see page 1435 of the November, 1949, issue of the Proceedings of the I.R.E.

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For a photograph and biography of Frederick W. Schott, see page 780 of the July, 1949, issue of the Proceedings of the I.R.E.

A. H. LaGrone (M'48) was born in Pauola County, Texas, on September 25, 1912. He received the degree of B.S. in electrical engineering from the University of Texas in 1938. After four years as distribution engineer with the San Antonio Public Service Co., San Antonio, Texas, he was commissioned in the U. S. Naval Reserve, and was ordered to active duty with the Navy in June, 1942. While in the Navy, Mr. LaGrone was instructor in radar at the Massachusetts Institute of Technology and later Radar Officer aboard the U.S.S. *Gillette*, D. E. 681, in the Atlantic.

A. H. LaGrone

At the conclusion of the war, Mr. LaGrone, then a lieutenant commander, was ordered to inactive duty and accepted the position of radio engineer with the Electrical Engineering Research Laboratory, the University of Texas. Mr. LaGrone also attended the University of Texas and was awarded the degree of M.S. in electrical engineering in 1948.

Mr. LaGrone is a member of Eta Kappa Nu and Tau Beta Pi.

❖

Keith MacDonald was born in Glasgow, Scotland, on July 24, 1920. He received the M.A. degree in mathematics and natural philosophy in 1941 and the Ph. D. degree in 1946, both from Edinburgh University. He served in the army as a radar and telecommunications officer from 1941 to 1946.

Keith MacDonald

Since 1946, Dr. MacDonald has been a research fellow in physics, Clarendon Laboratory, Oxford, England, working chiefly on properties of metals at very low temperatures, at the same time maintaining his interest in noise problems and fluctuations analysis.

❖

For a photograph and biography of K. Tomiyasu, see page 1156 of the October, 1949, issue of the Proceedings of the I.R.E.

Leslie A. Moxon was born on March 15, 1909. He received the London University B.Sc. degree in electrical engineering in 1929, following a 3-year course at the City and Guilds Engineering College. For the next two years he carried out research on high-frequency ammeters under the auspices of the Department of Scientific and Industrial Research, afterwards joining the staff of Murphy Radio Ltd., where he was responsible for the development of broadcast receivers and research on associated problems until 1940.

Leslie A. Moxon

In 1941 he joined H. M. Signal School in Portsmouth, England, where he took charge of a section concerned with the development of radar receivers; and he is now a member of the Royal Naval Scientific Service.

Mr. Moxon is the author of various technical papers, and of a book entitled, "Recent Advances in Radio Receivers," recently published by the University Press, Cambridge, England. He is an Associated Member of the British Institution of Electrical Engineers.

❖

Leo Storch was born on March 3, 1921, in Vienna, Austria. In January, 1944, he received the B.S.E.E. degree cum laude from the School of Technology, College of the City of New York. He was awarded the M.A. degree by the Graduate School of Stevens Institute of Technology in June, 1947.

Leo Storch

From February, 1944, to July, 1947, Mr. Storch was an assistant engineer in the Test Set Design and Development Department of the Western Electric Co. in Kearny, engaged in the circuit design for a wide variety of measuring equipment.

From August, 1947, to June 1948, he was associated in the capacity of engineer with the Teleregister Laboratories in New York, N. Y., where he was active in various phases of an electronic computer development project.

Since June, 1948, Mr. Storch has been a member of the advanced development group in the aviation radio section of RCA Victor in Camden, N. J. He has been principally concerned with the design of antenna tuning and loading networks for wide-band aircraft communications transmitters.

Contributors to the Proceedings of the I.R.E.

A. W. Straiton (M'47) was born in Tarrant County, Tex., on August 27, 1907. He received the B.S. degree in electrical engineering in 1929, the M.A., in 1931, and the Ph.D. in 1939, all from The University of Texas.

A. W. STRAITON

Dr. Straiton spent one year at Bell Telephone Laboratories, after which he taught at Texas College of Arts of Industries as assistant professor, associate professor, and professor of electrical engineering, successively. From 1941 to 1943, he was head of the Department of Engineering, Institutional Representative of E.S.M.W.T., and director of the Pre-Radar Training courses. Since 1943, he has been associate professor of electrical engineering at The University of Texas. He was recently made director of the Electrical Engineering Research Laboratory.

Dr. Straiton is a member of Sigma Xi, Tau Beta Pi, Eta Kappa Nu, the American Institute of Electrical Engineers, and the American Society for Engineering Education.

James H. Tillotson (S'45) was born in Oak Park, Ill., on November 14, 1923. He received the B.S. degree in electrical engineering from Purdue University in 1945, and the M.S. degree in 1947 from the same institution. Since 1947, he has been a graduate student at Stanford University, employed as a research assistant in the Electronics Research Laboratory, and also in the Microwave Laboratory at the University.

J. H. TILLOTSON

Mr. Tillotson is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

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For a photograph of KARL R. SPANGENBERG, see page 780 of the July, 1949, issue of the PROCEEDINGS OF THE I.R.E.

D. L. Waidelich (S'37–A'39–SM'44) was born on May 3, 1915, at Allentown, Pa. He received the B.S. degree in electrical engineering in 1936, the M.S. degree in 1938, both from Lehigh University, and the Ph.D. degree from Iowa State College in 1946. He was a teaching assistant at Lehigh University from 1936 to 1938, and since 1938 has been a professor of electrical engineering at the University of Missouri.

D. L. WAIDELICH

During 1944 and 1945, Dr. Waidelich was with the U. S. Naval Ordnance Laboratory in Washington, D. C., as an electrical engineer. For various summers he has also worked with the Bell Telephone Laboratories, New York, N. Y., the Westinghouse Electric Corporation at Bloomfield, N. J., and the U. S. Naval Electronics Laboratory at San Diego, Calif., as a consulting engineer. In 1946 he received mention by Eta Kappa Nu as one of the outstanding young electrical engineers in America. He is a member of the American Institute of Electrical Engineers, the American Society for Engineering Education, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

Correspondence

Chemical Composition and Structure of the Atmosphere*

Several year ago while investigating various atmospheric conditions and their effects upon high-speed racing engines while burning high-performance low-mileage alcohol blends, we ran across some interesting data upon the chemical composition and structure of the atmosphere. For one thing, there appears to be a gravitational stratification of the air, with a gradual change in chemical composition with altitude, until finally little more than the light hydrogen and helium remain at 1,000 to 2,000 miles.

* Received by the Institute, August 8, 1949.

Another interesting fact seems to be that many trace gases are contained in the air in appreciable percentages (ammonia, hydrogen, carbon dioxide, nitric oxides, etc.) other than water vapor and the inert gases. These vary in their concentration with locality, altitude, and time. We often wondered how much this condition has effected radio-wave propagation. Various papers on this subject do not appear to take the above factors into consideration.

Turning to the field of astronomy, the gravitational stratification of the terrestial atmosphere can be carried out still further. Recent theories of an "exploding universe" appear less credulous on the assumption that starlight spectroscopic frequency shifts are due largely to tenuous dust and gas clouds in interstellar space. Hence it might be reasoned that the earth's stratosphere does not end abruptly at several hundred miles, but gradually thins out into the highly rarefied gas and dust-filled space between the planets and the parent sun. The earth may also be ringed by faint dust rings similar to the prominent rings of Saturn, and the sun may be exploding columns of gases into space. These may account for some of the peculiar long-time-interval echoes and doppler shifts noted in radio-wave propagation studies. Jansky apparently suspected some such effects a good many years ago.

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Correspondence

Simplified Frequency Stabilization*

Previous attention to the problem of frequency stabilization of a reflex klystron has emphasized maximum stability rather than simplicity of operation and ease of AM. There is a large class of instrumentation problems in microwaves that does not require the maximum obtainable stability and does depend on AM. Hence there may be some general interest in one design of a compromise stabilizer that has proven itself in laboratory use.

Fig. 1 is a block schematic of a unit which is simple in operation and whose output is square-wave amplitude-modulated. Approximately 1 per cent of the power delivered by a klystron (such as a 2K39) is diverted by a directional coupler to a frequency discriminator composed of an unmatched hybrid junction, a phase changer, and a resonant cavity. The phase changer is adjusted to deliver equal modulated power to the detectors terminating the *E*- and *H*-arms of the hybrid when the cavity is detuned from resonance and the *AFC* switch is set to "off." The cavity is then tuned to resonance which again equalizes the power in the *E*- and *H*-arms, and the *AFC* switch is turned to "on." Subsequently, any tendency toward frequency change will cause an increase in power in one arm and a decrease in power in the other arm as a function of the cavity phase characteristic near resonance. Since there is a 180° phase reversal in the reflection from the cavity at resonance, the direction of frequency drift will determine whether the *H*-arm increases and the *E*-arm decreases or vice versa. Hence the amplified and rectified detector components may be used to control the g_m of the modulator tube and, consequently, the amplitude of the square-wave output. When this square wave is added to the steady-state klystron reflector voltage by derivative coupling, the total reflector voltage at the "on" oscillating condition is controlled by the discriminator and, consequently, the frequency of oscillation is stabilized.

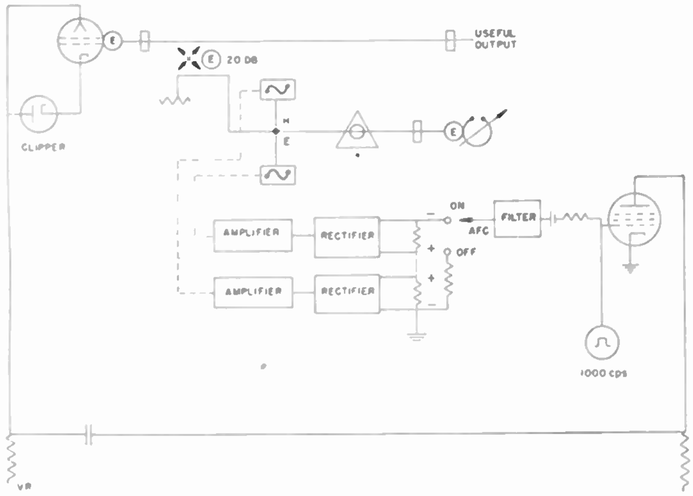

Fig. 1

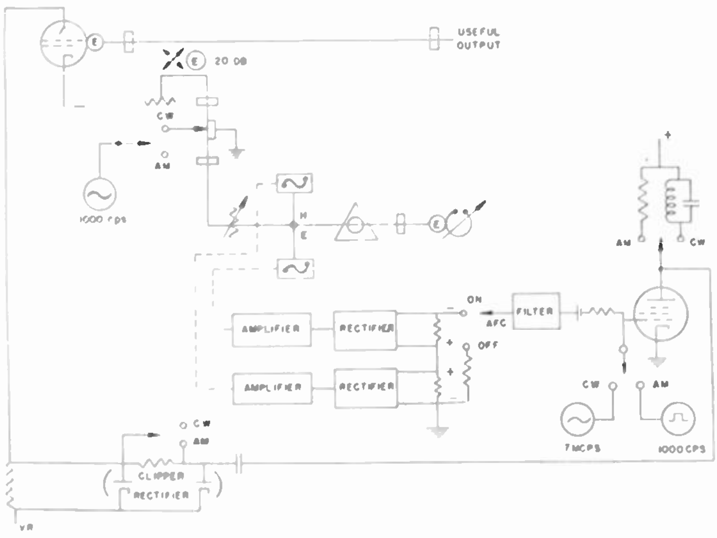

Fig. 2

With a slight modification, this stabilization scheme can be made to operate either with CW or with AM output (see Fig. 2). In this case the sampled power diverted to the discriminator is sinusoidally modulated by a crystal, amplified, rectified, and used to control the g_m of the modulator tube as before. However, for this case, a 7 Mc sine wave is derivatively coupled to a rectifier and the controlled direct current output is added to the steady-state reflector voltage. Conversion from AM to CW operation is accomplished by a single gang switch.

Stabilization of the order of 1 part in 10^5 is obtained for either the AM or CW case. The ease of operation in comparison to usual methods is, of course, the outstanding feature. Unskilled personnel are consistently able to "lock on" an arbitrary frequency in about 15 seconds after the klystron has been set.

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* Received by the Institute, July 20, 1949.

Patent References in Technical Papers*

It has become impressed upon me more and more through the past forty years, that papers published in the PROCEEDINGS OF THE I.R.E., and most other technical and scientific bodies, include very scant references to patents, either in footnotes or in appended bibliographies.

Patents constitute one of the largest, most comprehensive, and very frequently, the *only* source of technical information on many subjects, yet references to them by technical writers seem to be, for the most part, studiously avoided. This is true also of technical reference books.

Would it not be a very useful service if this were brought to the attention of our technical authors?

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N. J.

* Received by the Institute, August 15, 1949.

Correspondence

Resistance Attenuating Networks*

The equations for the components of a resistance attenuating network in terms of hyperbolic functions are well known. The derivation given below is of interest because of its simplicity, and because it leads to a novel form for the equations.

Consider first the T network of Fig. 1, where the series arms are each equal to R_0 and the shunt arm is a short circuit. This

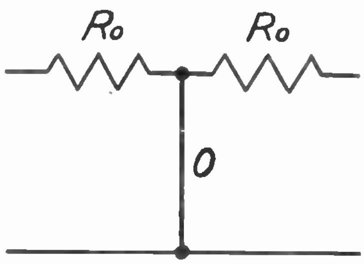

Fig. 1—Prototype resistance attenuating pad.

network obviously has a characteristic impedance equal to R_0 and infinite attenuation. Now consider the general "m" derived section, shown in Fig. 2, obtained from the

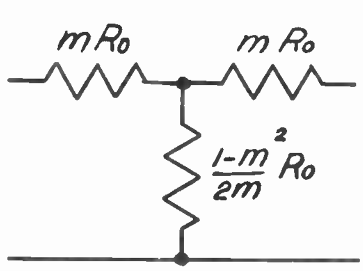

Fig. 2—"m" derived resistance attenuating pad loss $= 20 \log \left(\frac{1+m}{1-m}\right)$ db.

network of Fig. 1. This section has the same characteristic impedance (R_0) for all values of m (but only for $0<m<1$ will the network be realizable). It is readily shown that if the network of Fig. 2 is fed from a generator of impedance R_0 and terminated in a like impedance that the ratio of load current to generator current is

$$\frac{I_L}{I_G} = \frac{1-m}{1+m}.$$

Thus the loss of the network is

$$\text{loss} = 20 \log \left(\frac{1+m}{1-m}\right) \text{ decibels.}$$

SIDNEY BERTRAM
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* Received by the Institute, August 24, 1949.

Ground-Wave Field-Strength Calculation*

With regard to the paper by H. L. Kirke of the B.B.C. in your May, 1949,[1] issue, I should like to draw your attention to the fact that the recovery effect predicted in my paper[2] to which he refers has been substantiated by an experiment on a wavelength of 4m.[3] We have just completed a further test on 100 m. across a land/sea boundary also agreeing with the predicted result, and it is hoped to publish an account of these two experiments as part of another paper in the *P.I.E.E* later in the year.[4]

My purpose in writing is to say that although my method is empirical in the neighborhood of a boundary, these experiments suggest that it is in fact a close approximation to the truth in this important region, where the other methods described by Mr. Kirke are definitely inadequate.

The experiments analyzed in his paper certainly imply that where the conductivity changes are complex, one method or another under the conditions assumed may happen to fit the measured values best, and that the Somerville method, which has the merit of simplicity, is adequate in many cases. As Mr. Kirke admits, however, it is purely empirical, having no other theoretical basis than that it is "a move in the right direction" from the P. P. Eckersley method, so that it obeys the reciprocity condition only approximately and not of necessity.

My own method is admittedly in effect the P. P. Eckersley method made reciprocal, but I have given considerable theoretical justification for it well away from the boundary, more particularly at the shortwave limit. Mr. Kirke points out that, as opposed to the other methods, it can even give an increase in field-strength on crossing a boundary from low to high conductivity. This is something of an understatement, for the increase can be spectacular, as is shown by the experiments referred to above, and can be of profound practical importance, as our further paper will reveal.

I feel, therefore, that in adopting any particular method in the absence of a complete mathematical solution of the problem, its possible limitations should be clearly realized, especially in the neighborhood of a boundary, and where the reciprocity condition is concerned. The recovery effect can now be regarded as a well-established propagation phenomenon, and one which must be taken into account in problems of ground wave coverage over a composite path.

Theoretically, with horizontal polarization the effect should be reversed, but it would be confined close to the surface, and so it would be difficult to detect and of little practical significance.

G. MILLINGTON
Marconi's Wireless Telegraph Co. Ltd.
Baddow Research Lab.
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* Received by the Institute, May 23, 1949.

[1] H. L. Kirke, "Calculation of ground-wave field-strength over a composite land and sea path," PROC. I.R.E., vol. 37, pp. 489–497; May, 1949.

[2] G. Millington, "Ground-wave propagation over an inhomogeneous smooth earth," *Proc. I.E.E.*, vol. 96, Part III, no. 39, pp. 53–64; 1949.

[3] G. Millington, "Ground-wave propagation across a land/sea boundary," *Nature*, vol. 163, p. 128; 1949.

[4] G. Millington, "Ground-wave propagation across a land/sea boundary 100 m," *Nature*, vol. 164, p. 114; 1949.

Kirke's Reply*

Mr. Millington, in his letter to the Editor, mentions that he has carried out experiments on wavelengths of 100 m. and 4 m. I have seen the results of these experiments, which show a very marked recovery effect on passing from land to sea. The experiments were carried out under more extreme conditions than were possible in the experiments mentioned in my paper, and also show that under certain conditions it is better to site a transmitter farther away from the service area if by doing so the initial path can be over sea.

While, as pointed out in my paper, the Somerville method gives better correspondence with practical results than the Eckersley method, there is no doubt that it does not give an adequate allowance for the recovery effect on passing from land to sea, nor does it give a sharp enough drop when passing from sea to land.

As pointed out in Mr. Millington's letter, while the Somerville method is adequate in many cases and has the merit of simplicity, it would be unwise to place complete reliance on that method in all cases.

Although the Somerville method is rapid in use, the additional labor involved in the use of the Millington method is not great, and even if the Somerville method is used it would be wise to carry out a few check calculations using the Millington method, particularly in the neighborhood of a boundary.

The results of Mr. Millington's experiments when published will be of considerable interest.

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* Received by the Institute, July 18, 1949.

Correspondence

Note on Transit-Time Deterioration*

It is the aim of this note to give a simple derivation and a generalization of some formulas which were published recently.

We start with transit time deterioration of the density modulation of an electron beam due to the velocity distribution of the electrons.[1] Let I be the dc beam current and $i_0 \exp(j\omega t)$ the ac convection current at the beginning of the beam. Let $p(V)\,dV$ be the probability of an electron velocity between V and $(V+dV)$ volts in the beam. Then the convection current $i(t)$ at the instant t at a distance d from the beginning of the beam is:

$$i(t) = \int_0^\infty i_0 \exp\{j\omega(t-\tau)\}\,p(V)dV$$
$$= i_0 \exp(j\omega t) f(\omega), \qquad (1)$$

if it is observed that the electrons having a velocity between V and $(V+dV)$ and arriving at this point at the instant t started at a time $(t-\tau)$, where τ is the transit time of the electrons along the distance d and:

$$f(\omega) = \int_0^\infty \exp(-j\omega\tau)p(V)dV. \qquad (1a)$$

We apply this result to an investigation of transit time deterioration of the space-charge reduction of shot effect in electron beams.[2] Consider an electron beam satisfying the conditions:

1. Saturated shot noise in the beam current.
2. No correlation between the fluctuations in electron velocity and in electron density at the beginning of the beam.
3. The velocities of the individual electrons are completely random.

After conditions 1 and 3, the arrival of an electron at any point of the beam is an independent event occurring at random, so that we have for the total noise current at any part of the beam:

$$\overline{i_0^2} = 2eI\Delta\nu. \qquad (2)$$

Traveling along the beam, the initial density fluctuations decrease due to the velocity distribution, but on the other hand the initial velocity fluctuations are transformed into density fluctuations, such that (2) remains valid. Let the first effect give a contribution i_1 and the second one a contribution i_2 to the total noise current i_0. After condition 2 these contributions are uncorrelated and have to be added quadratically, so that:

$$\overline{i_0^2} = \overline{i_1^2} + \overline{i_2^2}. \qquad (3)$$

As $\overline{i_1^2}$ follows from (1) we have:

$$\overline{i_2^2} = 2eI\Delta\nu[1 - |f(\omega)|^2]. \qquad (4)$$

* Received by the Institute, September 2, 1949.

[1] M. J. O. Strutt and A. van der Ziel, "Application of velocity modulation tubes for reception at uhf and shf," Proc. I.R.E., vol. 36, pp. 19–23; January, 1948.

[2] D. K. C. MacDonald, "Transit time deterioration of space-charge reduction of shot effect," *Phil. Mag.*, ser. 7, vol. 40, pp. 561–568; May, 1949.

Condition (4) was derived under the three conditions mentioned previously. But obviously i_2^2 only depends upon the *magnitude* of the velocity fluctuations, so that (4) remains valid even if conditions 1 and 2 are no longer satisfied. i_2^2 gives the correct expression for the transformation of velocity fluctuations into density fluctuations along the beam, as long as condition 3 is satisfied, even though it may happen that i_1 and i_2 are correlated and cannot be added quadratically, if conditions 1 and 2 do not hold.

We now apply this result to the case of a "shifted" Maxwellian velocity distribution discussed by several authors[2,3]: $p(V)\cdot dV=0$ for $V<V_0$; $p(V)dV=\exp(-\Delta V/V_T)\,d(\Delta V/V_T)$ for $V \geqq V_0$ in which T is the cathode temperature, $\Delta V=(V-V_0)$ and $V_T=kT/e=T/11{,}600$ volts.

Let τ_0 be the electron transit time along the distance d for those electrons having an initial velocity V_0, then the transit time τ for the electrons having an initial velocity $(V_0+\Delta V)$ is:

$$\tau = \tau_0[V_0/(V_0+\Delta V)]^{1/2}$$
$$= \tau_0 - \tfrac{1}{2}\tau_0(\Delta V/V_0), \qquad (5)$$

if $(\Delta V/V_0) \ll 1$. Introducing this result into (1a) we obtain:

$$f(\omega) = \exp(-j\omega\tau_0)\,\frac{2V_0/V_T}{2V_0/V_T - j\omega\tau_0}. \qquad (6)$$

Substituting this into (1) we obtain:

$$i(t) = i_0 \exp\{j\omega(t-\tau_0)\}\,\frac{2V_0/V_T}{2V_0/V_T - j\omega\tau_0}, \qquad (7)$$

equivalent to a result derived previously by Barlow.[3]

Substituting (7) into (4) we obtain:

$$\overline{i_2^2} = 2eI\Delta\nu\,\frac{(\omega\tau_0)^2}{(2V_0/V_T)^2 + \omega^2\tau_0^2},$$

a formula which was recently obtained by MacDonald.[2]

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[3] M. J. O. Strutt and A. van der Ziel, Discussion on: "Application of velocity modulation tubes for reception at uhf and shf," Proc. I.R.E., vol. 37, pp. 896–900; August, 1949.

Field-Strength Observations Made During the Total Eclipse of the Sun*

During the total eclipse of the sun which occurred on November 1, 1948, the following observations were made at Eastleigh Aerodrome, Nairobi, Kenya Colony, regarding the propagation of electromagnetic waves. It is thought that they may be of interest in view of the fact that this station was within the path of totality.

Observations were made on an army communications receiver Type R 206,

* Received by the Institute, July 12, 1949.

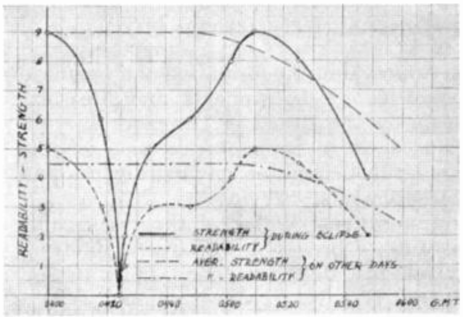

Fig. 1

Mark II, using a short, vertical antenna of small effective height; this equipment is in daily use in the electronics section of the East African Meteorological Department, at Eastleigh Aerodrome.

The electronics section had planned to release additional radio sondes for the purpose of sounding the atmosphere during the three days centered round the eclipse. Two days before the eclipse, it occurred to the writer that it would be interesting to observe the changes in field strength of a short-wave station whose line of propagation or some part of it would roughly coincide with the path of the total eclipse. The choice fell on the station WWV in Washington, D. C., broadcasting on 15 Mc because it seemed likely that, at the time of the eclipse (0424 G.M.T.), its transmissions would be received along a great-circle path via S.E. Australia and the Pacific ocean; i.e. approximately the same line as that of the eclipse. Unfortunately, it was not possible in so short a time to install a directional antenna. The receiver is not equipped with a field-strength indicator and the writer had to rely on his experience in judging strength of reception and readability.

The observations made are plotted in Fig. 1 and show distinctly the change in reception conditions. It will be seen that the reception of WWV disappeared completely during the period of the total eclipse, and reverted to normal by the time that the sun was completely unobscured. Field-strength observations (mean values) for the day preceding and the day following are also shown and it can be stated that these are quite typical.

It is realized that the observations described above were made without adequate preparation and, consequently, the results should be treated with extreme reserve. It is felt, however, that such a clearly defined change in reception conditions corresponding with the eclipse of the sun were worthy of report and that, on future occasions, it would be worth organizing a properly prepared series of observations.

The report is published by kind permission of the acting director of the East African Meteorological Department.

J. Gross
P.O. Box 973
Nairobi, Kenya Colony

Institute News and Radio Notes

Technical Committee Notes

Approval of the formation of a technical committee on Instruments and Measurements by the Board of Directors was announced at a September session of the **Standards Committee on Instruments and Measurements.** Serving as Chairman is Professor Ernst Weber of the Polytechnic Institute of Brooklyn. Millard A. Baldwin, Jr., of the Bell Telephone Laboratories, Murray Hill, N. J., was appointed Television Co-ordinator by the Standards Committee. He will be chairman of a small subcommittee for the co-ordination work within the IRE and with other organizations in order to expedite production of urgently needed standards on television and video techniques. Approval was granted for the *Standards on Reference Designations* (for the identification of electrical, electronic and mechanical parts and their associated graphical symbols) prepared by the **Symbols Committee.** . . . The **Electron Tubes and Solid-State Devices Committee,** at a meeting September 16, reported that it will complete definitions for publication in the February issue of PROCEEDINGS. . . . The **Electronic Computers Committee** met on September 14 at the Computational Laboratory of Harvard University with Jay W. Forrester, Chairman, presiding. At present the Committee is concerned chiefly with the formulation of the Committee's scope and subcommittee structure to give equitable coverage to analog and digital computers, with the compilation of a comprehensive computer bibliography, with the preparation of a dictionary of standard computer definitions, and with methods of interchange and dissemination of information on electronic computers. . . . L. C. Van Atta presided at the September 12 session of the **Antennas and Wave Guides Committee.** A. G. Fox, Chairman of the **Subcommittee on Wave Guides,** reported that his group had formulated several recommendations for presentation to the Committee. The definitions were adopted by the main committee. Other terms will be submitted to the attention of the **Wave Propagation Committee and the** Circuits Committee for definition. . . . The **Joint Technical Advisory Committee,** at an all-day session September 22, was concerned principally with the presentation of the JTAC's report to the FCC at the September hearing on Allocations for Television Broadcast Services. Donald G. Fink, Chairman, presented JTAC's testimony at the hearing.

National Bureau of Standards Publishes Electrical Measure

"Establishment and Maintenance of the Electrical Units" is the title of a new booklet published by the National Bureau of Standards and available from the U. S. Government Printing Office. It describes the system of electrical measurement using "absolute" units, adopted by the International Conference of Weights and Measures, officially instituted January 1.

Standards Notice

This issue of the PROCEEDINGS contains the following standards: **Standards on Piezoelectric Crystals, 1949; Standards on Radio Aids to Navigation, Definition of Terms, 1949; Standards on Railroad and Vehicular Communications; Methods of Testing, 1949; Standards on Tests for Effects of Mistuning and Downward Modulation, 1949.**

It is the aim of the Board of Directors to include the standards in the PROCEEDINGS unless unforeseen contingencies arise so as to make them available to all members. All issues of the PROCEEDINGS containing such standards will carry a mention of that fact on their front covers in red, and will be indicated by a red band on the spine of such issues. Reprints of such standards, while available, may be purchased from IRE headquarters.

A list of available standards and their prices may be found on the back page of the Annual Index facing page 33a in this issue.

IRE Announces Inception of Antennas, Propagation Group

An IRE Professional Group on Antennas and Propagation has been formed with membership to include those having professional interest in antennas and propagation (including waveguides). IRE members who wish to enroll may do so by forwarding a card to the Membership Committee, Professional Group on Antennas and Propagation, The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y.

The group held a three-day meeting in conjunction with the URSI in Washington, D. C., beginning October 31. A West Coast meeting is planned for next Spring.

Crystal Oscillator Plates Used for High Frequencies

Crystal grinding methods and machinery have been investigated by the National Bureau of Standards in order to overcome difficulties in supplying very thin quartz crystal oscillator plates having fundamental frequencies up to 100 Mc or even higher. Improved equipment, capable of producing 0.001-inch thick quartz crystals with a high degree of parallelism and flatness, can be used for grinding equally thin wafers from a variety of other materials. The Bureau has found that a promising application is the production of extremely thin dielectric plates for miniature radio condensers.

Members Are Appointed To Research, Development Board

Army, Navy, and Air Force representatives who will serve on the Research and Development Board, Department of Defense, have been announced by the Secretaries of the three Departments. They have been chosen in accordance with the recently revised directive from Secretary of Defense Louis Johnson to the Board, which provides that one member from each of the three military Departments shall be either an Under or Assistant Secretary of the department.

Army representatives are Archibald S. Alexander, the Assistant Secretary of the Army, and General Mark W. Clark, Chief, Army Field Forces. Navy members are Dan A. Kimball, Under Secretary, and Rear Admiral R. P. Briscoe, Director, Fleet Operational Readiness Division, Office of the Chief of Naval Operations. Representing the Air Force are Arthur S. Barrows, Under Secretary, and Lieutenant General Benjamin W. Chidlaw, Commanding General, Air Materiel Command, Wright-Patterson Air Force Base, Dayton, Ohio.

Dr. Karl T. Compton, chairman of the corporation of the Massachusetts Institute of Technology, who has been Chairman of the Research and Development Board since October 15, 1948, will continue to direct the activities of the seven-man board.

The Board, since its formation under the National Security Act of 1947, has been charged with preparing an integrated program of research and development for military purposes. It has been given the added responsibility of determining that such programs are carried out by military departments, and of directing changes in emphasis upon existing programs and projects, including the curtailment of programs or projects deemed to be unwarranted.

Haverford College Establishes S.S.R.S., New Science Society

The formation of the Society for Social Responsibility in Science has been established at Haverford College, Haverford, Pa., by a group of scientists and engineers from several states who convened at a week-end session. In part the aims of the group are stated as a wish "to foster throughout the world a tradition of personal moral responsibility for the consequences for humanity of professional activity, with emphasis on constructive alternatives to militarism; to embody in this tradition the principle that the individual must abstain from destructive work and devote himself to constructive work, according to his own moral judgment."

Buenos Aires Section Meeting

The celebration of the tenth anniversary of the inception of the Buenos Aires Section of the IRE was observed at its annual Engineering Week, held from October 31 to November 6.

Radio Station WWV's Signal Improves Propagation Services

Improvement of one of the technical radio broadcast services of the National Bureau of Standards' Central Radio Propagation Laboratory will result from the new broadcast signal of the Bureau's radio station WWV. This signal, a warning of unstable conditions in the ionosphere, provides additional data on ionospheric disturbances, information of vital significance to the Armed Services and the communications industry in maintaining uninterrupted long-distance radio communications. It became effective on November 1.

Heretofore two grades of propagation conditions have been recognized in the notices given at nineteen and forty-nine minutes past each hour by station WWV, which continuously broadcasts standard radio frequencies, time announcements, and the standard musical pitch in addition to the radio propagation disturbance notices.

The Letter "N" (in International Morse Code) repeated several times has signified normal conditions, while the letter "W" has constituted a warning that disturbed conditions were present or expected within 12 hours. A third category, indicating unstable conditions, and denoted by the letter "U," is now being used when the forecasters at the CRPL's warning center expect satisfactory reception of short-wave communication or broadcast services employing high-power transmitting equipment operating on the recommended frequency, but poor results on less well equipped services.

Such conditions often occur as major disturbances subside. Although point-to-point communication links are able to resume reliable operation, mobile services, and short-wave broadcasts continue to experience difficulty. The propagation disturbance notices, broadcast in International Morse Code, primarily refer to the North Atlantic Radio circuits.

Coast and Geodetic Survey Publishes New World Chart

Publication of a new world chart centered on the control tower at La Guardia airport has been announced by the United States Coast and Geodetic Survey.

The new chart, suggested by E. O. Cutler, consulting engineer, is produced on the azimuthal equidistant projection, valuable for radio and aeronautical operations. This projection has found wide applications in recent years for determining true distances and azimuths from a given point to any other point on the globe.

Distances and directions can be determined with a high degree of accuracy from the point of tangency which is the center point located at LaGuardia Airport control tower. A straight line drawn on the chart from New York City to any other point on the earth's surface shows the shortest route to that point and places traversed by such a route can be seen at a glance.

The chart, number 3042, is available at $0.40 a copy at any U. S. Coast and Geodetic Survey Office and at many booksellers.

Research Board Responsibility Extended by Defense Secretary

Extension of the authority and responsibility of the Research and Development Board in compliance with the terms of the recent amendments to the National Security Act will result from a directive issued by Secretary of Defense Louis Johnson.

According to Dr. Karl T. Compton, Chairman of the Research and Development Board, "Another significant step has just been taken toward the achievement of a truly effective program of military research in the United States. The Research and Development Board now has, in addition to the responsibility for formulating a complete integrated program of research and development for military purposes, the authority to determine whether its program is being carried out by the three military departments.

"In specific terms, this means that the board may, as it deems necessary, direct changes in the programs of the services, including the initiation of new projects, the increase of effort in certain areas, and the decrease or curtailment of effort in other areas."

Calendar of COMING EVENTS

1949 Annual Meeting, National Society of Professional Engineers, Houston, Texas, December 8–10

Southwestern IRE Conference, Baker Hotel, Dallas, Texas, December 9–10

AAAS 116th Annual Meeting, New York City, December 26–31

1950 IRE National Convention, New York, N. Y., March 6–9

1950 IRE Technical Conference, Dayton, Ohio, May 3–5

Armed Forces Communications Association 1950 Annual Meeting, April 26, Photographic Center, Astoria, L. I., N. Y.; May 12, New York City; May 13, Signal Corps Center, Fort Monmouth, N. J.

General Electric Television Equipment Installed in Italy

Installation of the first American television transmitter in Europe at Turin, Italy, has been announced by the General Electric Company. Telecasting began from Turin on September 11, and covers approximately 50 miles in Northwest Italy.

The main features of the installation are a studio equipped with three cameras and programming facilities, a microwave link to relay the programs to the transmitter site, and a 5 kw transmitter of the latest design.

SMPE Recommends Allocations for Television in Theaters

Frequency allocations for theater television have been requested from the FCC by the Society of Motion Picture Engineers under the chairmanship of D. E. Hyndman. According to the proposal, programs for theater television would be picked up from remote field locations, television studios or theaters, then would be sent to a central studio or transmitter, and then distributed to theaters that wish to present the program on their screens. Channels of radio frequencies would be required to carry the picture and sound from point of origin to theaters either on a local basis, between nearby cities, or on a nation-wide basis, depending on its commercial success.

Society engineers are of the opinion that picture quality would have to be as good as motion pictures are today. They recommended that the Commission provide wide enough channels to allow development in that direction.

Improved quality would have to begin from the present broadcast standards of 525-line black and white television. Channels 50 Mc wide, it was estimated, would be needed to give high-quality pictures in black and white, and subsequently in full color also.

According to the SMPE, as many as 60 different channels might be needed for a complete and thoroughly competitive nation-wide television system. In any given locality, fewer channels might provide adequate service.

Conferences have been held by SMPE recently with Theater Owners of America, the Motion Picture Association, and several other industrial groups in an attempt to provide the industry with a well-rounded picture of what theater television means technically.

New Coaxial Cable Now Serves Philadelphia-New York Area

Hundreds of additional telephone conversations and three more television channels between New York and Philadelphia have been placed in service by the introduction of a new Bell System coaxial cable, according to the Long Lines Department of the American Telephone and Telegraph Company. It will connect at Philadelphia with an already existing cable to form part of another communication link supplying more long-distance telephone and video program facilities both south and west of the city.

For television purposes it will be equipped to provide two more channels from New York to Philadelphia and one more in the reverse direction. A total of five channels will be available to carry programs in the southbound direction and two for northbound transmission.

At Philadelphia, the new cable joins another coaxial link leading to Baltimore and Washington. It is a joint project of the New York Telephone Company, the New Jersey Bell Telephone Company, the Bell Telephone Company of Pennsylvania, and the Long Lines Department of the AT&T.

IRE Executive Committee Approves New Groups

Approval of the Professional Group on Vehicular and Railroad Radio Communications and establishment of the Professional Group on Broadcast and Television Receivers has been granted by the Executive Committee of the IRE.

Officers of the Professional Group on Vehicular and Railroad Radio Communications are as follows: A. B. Buchanan, chairman, F. T. Budelman, vice-chairman, P. A. Penhollow, secretary; and administrative committee members: E. C. Denstaedt, C. N. Kimball, R. H. I. Lee, D. E. Noble, Waldo Shipman, R. C. Stinson, and H. E. Weppler.

Following a recommendation for the formation of a Professional Group on Broadcast and Television Receivers, an administrative committee was named as follows: J. E. Brown, Virgil M. Graham, R. A. Hackbusch, D. D. Israel, I. J. Kaar, and Henry C. Sheve.

Annual Spring Exhibition Of British Radio Industry

Announcement has been made of the seventh annual exhibition of British Components, Valves and Test Gear for the radio, television, electronic and telecommunication industries. It will be held Monday, April 17, to Wednesday, April 19, 1950, in the Great Hall, Grosvenor House, Park Lane, London, W.1.

Admission is by invitation of the organizers, the Radio and Electronic Component Manufacturers' Federation, 22 Surrey Street, Strand, London, WC2.

Radio Inventions, Inc. Now Is Named Hogan Laboratories, Inc.

The corporate name of Radio Inventions, Inc., John V. L. Hogan's research and development laboratory which specializes in facsimile, has been changed to Hogan Laboratories, Inc.

It is felt that the change in name is appropriate at this time, because Mr. Hogan is now devoting full time to the supervision of the engineering and development work of the laboratory, and because the organization has been called upon by the government and private industry to undertake projects far afield from that implied by the former corporate name. The company was founded in 1929, and will remain unchanged in personnel and location.

Structural Products Produce All-Glass Television Bulbs

All-glass rectangular television bulbs are being produced successfully by the American Structural Products Company, subsidiary of Owens-Illinois Glass Company, according to Stanley J. McGiveran, President of the Company, which is a subsidiary of Owens-Illinois Glass Company.

The new rectangular bulb is the result of extensive research, and will give television tube manufacturers an ideal all-glass bulb designed to receive 100 per cent of the transmitted television picture.

Industrial Engineering Notes[1]

Television News

The FCC has announced extension of the color phase of its television hearing into December with the **report** that **no important decisions** on color television, the lifting of the TV "freeze," or expansion of television broadcasting into the UHF **will be made before 1950.** Outlining a schedule for completing the color phase of the television hearing, the FCC said cross examination of the witness will not begin until December 5. . . . Five members of the FCC, members of the RMA Television Committee, leading industry engineers, and others, viewed prolonged demonstration of the proposed CBS color television system at the FCC television hearing in Washington. RCA gave a similar demonstration of its proposed electronic color television system before a similar audience. . . . It also viewed a demonstration by Color Television, Inc., at San Francisco . . . . **The Radio Technical Commission of the Ministry of Communications and Public Works** recently adopted regulations and standards for a Brazilian television service, according to information received by the U. S. Department of Commerce. The regulations provide for 12 channels and two different standards because the areas to be served have different power supplies. The government adopted a 525 line system with 60 cycles for the Sao Paulo area, and 625 lines with 50 cycles for the Rio de Janeiro area. Concessions already have been granted for two stations at Rio de Janeiro and one at Sao Paulo. . . . **United States** preparatory **work for** the sixth meeting of the **International Radio Consultative Committee** (CCIR) of the International Telecommunications Union has been started by the Study Group to further plan for the meeting at Praha, Czechoslovakia, in 1951. Curtis B. Plummer, of the FCC, is Chairman of the U. S. Preparatory Committee for Study Group No. 11, which will consider television problems, including questions relating to single sideband. The group is working on recommendations and questions to be considered in drafting the U.S. position.

FCC Actions

Appearing as Chairman of the RMA-IRE Joint Technical Advisory Committee, Donald G. Fink was the first witness when the FCC television **hearing opened** Monday morning, September 26. His direct testimony consumed only an hour, but his direct examination by several Commissioners and FCC legal and technical aides occupied the remainder of the day. After describing the methods utilized in evaluating some nine possible color television systems, Chairman Fink said, "JTAC is of the opinion that sufficient information is not available on these systems, and on their operation in the field, to permit a definite comparison of their suitability for public service." Mr. Fink cited numerous technical differences between the nine systems and emphasized that JTAC has not taken sides on the claims of the various proponents of color TV systems. He explained that JTAC does not oppose color television, but believes that further demonstrations and tests are needed before a final decision is made by the FCC. "Following the determination to standardize on a particular color system, and prior to the final adoption of standards for commercial use," he said "a public field test of at least six months duration should be undertaken to assure that the proposed service can, in fact, be rendered." . . . **The FCC has granted a petition of the Allen B. DuMont Laboratories, Inc. for comparative demonstrations** early in November of black-and white versus color receivers. The petition asked that proponents of color systems be required to include comparable demonstrations of black-and white commercial systems under "conditions controlled by the Commission and that this topic be the subject of public presentation before the Commission in advance of any such demonstrations." The FCC was also asked to consider the "availability of the equipment employed by the sponsored systems for utilization under commercial conditions." . . . **Total investment in taxi radio equipment is "nearing $30 million,"** FCC Commissioner George E. Sterling said in an address at the annual convention of the National Association of Taxicab Owners at Buffalo, N. Y. Two way radio for cabs has been authorized by the FCC for "two-thirds of all the taxicabs in the United States." There are approximately 2,700 authorized radio taxicab stations serving a total of 55,000 cabs. Many installations are completed and in use, others are underway, Commissioner Sterling added.

New Heater Compensation Method

Scientists at the National Bureau of Standards have developed a new method of compensating for line-voltage changes in stabilized current power supplies. The Bureau reports that in the new circuit arrangement, heater-voltage fluctuations are used to compensate for the line-voltage fluctuations, thus increasing the stability of the output voltage. It reports that the new method can be used to good advantage in power supplies for such constant-current devices as direct-current amplifiers and microwave oscillators.

Cathode-Ray Tube Sales Rising

Sales of television-receiver type cathode-ray tubes increased during the second quarter of the year, according to the RMA. Second quarter sales of 77,054 TV picture tubes values at $23,123,698 were reported by tube manufacturers, as compared with 686,620 units valued at $21,971,869 in the first quarter of 1949.

Radio Sales Decreased In July

July sales of appliance and specialty wholesalers, including radios, dropped 10 per cent under sales in June and 8 per cent below those of July, 1948, according to the Department of Commerce. Sales during the first seven months of 1949 were 6 per cent below the corresponding 1948 period.

[1] The data on which these Notes are based were selected, by permission, from "*Industry Reports*," issues of Sept. 19, Sept. 23, Sept. 30, Oct. 7, Oct. 14, published by the Radio Manufacturers' Association, whose helpful attitude is gladly acknowledged.

IRE People

Frank B. Jewett (F'20), for many years vice-president of the American Telephone and Telegraph Company, and former president of the National Academy of Sciences, has been awarded the 1950 medal of the Industrial Research Institute, Inc.

The medal is awarded annually for "outstanding accomplishment in leadership or management of industrial research which contributes broadly to the development of industry or the public welfare."

Dr. Jewett served as president of Bell Telephone Laboratories from its incorporation from 1925 until 1940. Previously he had served as assistant chief engineer, chief engineer, and vice-president of the Western Electric Company, manufacturing, and supply unit of the Bell System.

During World War II, Dr. Jewett was a major contributor to the activities of the National Research Council and the National Defense Research Committee of the Office of Scientific Research and Development, the top civilian agency which co-ordinated the efforts of thousands of civilian scientists. He has been honored by many universities, colleges, and professional societies, and was president of the National Academy of Sciences from 1939 to 1947.

❖

Clinton Richards Hanna (M'28-SM'43), associate director of the research laboratories, Westinghouse Electric Corporation, Pittsburgh, Pa., has been awarded the Howard N. Potts Medal in recognition of his initiative in the conception and development of the Tank Gun Stabilizer. The invention won for Dr. Hanna a Presidential Citation in 1942. This device helps to attain accuracy of fire while a tank is in motion on rough terrain, and secures a greater number of aimed hits than were formerly possible.

Dr. Hanna's achievements include the design of the Silverstat, an automatic voltage regulator first used for control of generators and used since 1938 on motors, turbines, or wherever automatic-voltage control is required. He directed development of the Westinghouse Photophone, one of the first successful methods of producing sound motion pictures.

More than ninety patients here and abroad are held by Dr. Hanna and he is the author of many technical works, included among which are "The Function and Design or Horns for Loud Speakers," "Design of Telephone Receivers for Loud Speaking Purposes," and "Loud Speakers of High Efficiency and Load Capacity."

Dr. Hanna, who is a native of Indianapolis, received the bachelor's degree from Purdue University in 1922. In 1926 Purdue awarded him the professional E.E. degree, and in 1945 an honorary doctor of engineering degree.

He joined the Westinghouse Corporation in 1922 and was active in the development of loud speakers and sound motion picture apparatus until 1930. He was then made manager of the development division of the research department, becoming manager of the electromechanical division in the same department in 1936. He has held his present post since 1944.

❖

Charles J. Breitwieser (A'37), chief of electronics and engineering laboratories for the San Diego, Calif., Division of Consolidated Vultee Aircraft Corp., has been awarded an honorary degree of Doctor of Science from the University of North Dakota, his alma mater.

He was cited for his contributions in the field of guidance and control of guided missiles; his efforts in the development of a high-voltage, alternating current, electrical system for large aircraft; his pioneering in radiotherapy and electrosurgery and in the general field of hyperpyrexia, or artificial fever; and his achievements in the fields of radio and television. The degree was conferred by his father, Dr. J. V. Breitwieser, Dean of the University's School of Education and summer school director.

A native of Colorado Springs, Dr. Breitwieser was graduated from the University of North Dakota in 1930 with a degree of bachelor of science in electrical engineering. He earned the M.S. degree from California Institute of Technology in 1933.

He became associated with Consolidated Vultee in 1942 as a staff engineer in charge of radio and electrical engineering. He holds patents for an aircraft wing flap synchronizer, a new system of radio communication, and also for electromedical equipment. He is a member of the Research and Development Board for the National Military Establishment.

Dr. Breitwieser's articles have been published in the *Archives of Physical Medicine*, the *Pacific Journal of Homopathy* the *Journal of the Osteopathic Association of America*, and other journals in this country and abroad.

❖

Everhard H. B. Bartelink (A'29-M'37-SM'43), has been named assistant to the director of research at General Precision Laboratory, Pleasantville, N. Y. Formerly he was head of the radio department of the General Telephone Corp.

Dr. Bartelink, who was born in Zutphen, Holland, received the degree in electrical engineering from Delft University, and the Ph.D. in physics from Munich University. In addition to the General Telephone Corp., he has served on the technical staffs of the Netherlands Telephone Company, General Electric, and Radiation Laboratory at MIT.

August J. Mundt (M'45), formerly general superintendent of training and personnel of the Western Union Telegraph Company, has been appointed Dean of Coordination at Walter Hervey Junior College. He retired in September from Western Union after twenty-seven years of service, both as an engineer and in personnel work.

Dean Mundt was graduated from Princeton University in 1915, and completed studies at Princeton University Graduate School of Engineering in 1917. He was an instructor in physics and electrical engineering at Princeton.

In his new association he will supervise the co-operative education program at Hervey and co-ordinate activities of the students in the work-study plan. He will be in charge of all student on-the-job employment in the engineering, business, and liberal arts curricula.

Dean Mundt is a member of the American Society for Engineering Education, and the American Institute of Electrical Engineering. For the past five years, he has been a member of the N. Y. Engineers Committee on Student Guidance; he was formerly a member of the Electrical Technology Advisory Commission for the New York State Institute of Applied Arts and Sciences.

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Ross H. Reynolds, Jr. (A'46) has been appointed a district representative for the General Electric Company's electronics department. He will be responsible for sales of marine electronic equipment in the New England, New York, and Atlantic districts of the department.

Mr. Reynolds, who has been with General Electric since 1940, is a native of Raleigh, N. C. He received the electrical engineering degree from North Carolina State College. When he joined General Electric, he was assigned to the instrument engineering section at the West Lynn, Mass., Works. From 1941 until 1946 he served in the U. S. Army as a member of the Signal Corps. He served overseas for two and one-half years, holding the rank of major.

Rejoining the electronics department of GE in April, 1946, he was assigned to radar engineering at Syracuse. In July, 1948, he was transferred to broadcast sales.

Mr. Reynolds is a member of the American Institute of Electrical Engineers.

❖

W. M. Gottschalk (S'41-A'45) has been appointed to the microwave and tube section of the newly created Research Division of the Raytheon Manufacturing Company, Waltham, Mass.

Joseph F. Bozelli (A'46) has been appointed assistant sales manager of the L. S. Brach Manufacturing Corp. of Newark, N. J. He will supervise and direct a new television antenna promotion for the Brach Corp., active in the electronic and electrical fields for over forty-two years.

Formerly, he was sales engineer with the JFD Mfg. Co. of Brooklyn, N. Y., in charge of manufacturing accounts, and earlier he had been sales production manager with the Fred Goat Co., also in Brooklyn. He has been associated with electronics for the last 10 years.

Mr. Bozelli holds a pilot's license and was a member of the Civil Air Patrol for three years. With 100,000 miles logged on Passenger Air Lines, he was recently admitted to United Air Lines' exclusive 100,000 Mile Club.

Mr. Bozelli's appointment coincides with a new television antenna and television accessory program soon to be announced.

William C. Hahn (A'36-SM'45), research associate in electronics at The Knolls research laboratory of the General Electric Company, died recently en route to the hospital. Mr. Hahn had been associated with General Electric since his graduation from Massachusetts Institute of Technology, where he earned the B.S. degree in 1923. Upon his completion of a student engineering course, GE sent him to its Chicago office. From 1933 he worked in the engineering general department at Schenectady.

A native of Illinois, Mr. Hahn was 48 years old at the time of his death. He attended high school in Kenosha, Wis., and for three years was a midshipman at the U. S. Naval Academy, Annapolis, Md.

Mr. Hahn was a member of the American Institute of Physics, the New York State Society of Professional Engineers, the Schenectady GE Engineers Association, the Whitney Club, and the GE Quarter-Century Club.

Martin M. Freundlich (A'38 SM'45) has been placed in charge of the newly established tube laboratory in the Applied Physics Section of Airborne Instruments Laboratory, Mineola, L. I., N. Y. Dr. Freundlich, who has been prominent for many years in vacuum-tube research and development, will conduct research on storage tubes and other vacuum devices which will help facilitate the laboratory's expanding program of electronic research and development.

He has been engaged in television research at Columbia Broadcasting System for more than ten years. There, in his tube laboratory, he did pioneer research on tubes for color television and also on early projection television tubes.

During the war Dr. Freundlich was occupied at North American Phillips for a period of one and one-half years with the development of radar and television cathode-ray tubes.

Books

Atmospheric Electricity by J. Alan Chalmers

Published (1949) by Oxford University Press, 114 Fifth Ave. New York 11, N. Y. 163 pages+6-page index+6 page references. 36 figures. 5¼×9. $3.75.

In preparing this book the author sets for himself a twofold purpose: first, of providing the reader with an introduction to the subject of atmospheric electricity; and second, to give the research worker a comprehensive literature survey of previous work in the field through citing and briefly discussing 248 references.

"Atmospheric Electricity" is divided into twelve chapters under titles: Historical Introduction; Fundamental Principles and General Summary; The Ions in the Atmosphere; The Earth's Vertical Field; The Conductivity of the Air; The Air-Earth Current; Point Discharge Currents; Precipitation Currents; The Transfer of Charge; The Thunder-Cloud; The Lightning-Flash; and the Separation of Charge.

Presentation of the subjects listed is largely through a chronological discussion of the references cited without seriously attempting to integrate all the various contributions into coherent reading. The omission of very important work recently done in the United States is indeed unfortunate. For example, the excellent contributions of Workman and Holzer have been neglected; the work of Byers and his group of thunderstorms has been omitted; no adequate discussion has been made of the detection and significance of radiation (sferics) originating in electrical storms or of the relationship of precipitation static to atmospheric electricity. In the same vein of criticism, there appears a lack of recognition of important research reported by German workers in the field. The deliberate omission of ionospheric exploration by radio methods and cosmic rays is understood and appreciated, for a lack of connection exists between such investigations and the area of activity which the book emphasizes.

It is not a lengthy treatise, and in the pages allowed it is perhaps unfair to expect a fully comprehensive treatment. The material presented is interesting, easily readable, and almost nonmathematical. Its reading public should be largely physicists interested in atmospheric phenomena, meteorologists, and, to a limited extent, radio engineers concerned with such activities as radio-sonde techniques, sferics, etc. However, almost any technically minded person would enjoy an evening browsing through the contents and acquiring more conversance with the tantalizing subject of atmospheric electricity.

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Pulses and Transients in Communication Circuits by Colin Cherry

Published (1949) by Chapman and Hall Ltd., London. 310 pages+5-page index+xvi pages. 120 figures. 5¼×8¾.

The most important function of this new volume will be to help engineers to put transient analysis in its proper perspective in the subject of communications theory. It is intended "as an introduction to circuit transient analysis for communications engineers . . . using, whenever possible, rigorous physical arguments and only elementary mathematics . . . . Electric waveforms are dealt with, rather than analytical functions, thus giving the book a geometrical or 'oscillographic' flavour."

The first three chapters of the book dealing with the basis of network analysis, the frequency spectra of modulated wave pulses and transients, and the steady-state, put the reader's knowledge of transients and networks on a sound footing. Chapter 4, which treats the transient response of networks, is followed by an excellent chapter which discusses the "use and abuse" of idealized response characteristics in transient analysis. A chapter on multistage amplifiers includes useful sections on the constancy of the bandwidth-gain product and on the relation between signal-to-noise ratio and frequency response. Chapter 7 deals with assymetric sideband channels, including suppressed sideband radio and television working. The final chapter on reflection and echo effects is largely relevant to the pulse-testing of networks, on which the author is an expert.

The book covers the subject quite comprehensively and with a practical emphasis that lends conviction. Although the theoretical treatment of Fourier-analysis in Chapter 2 may be bettered elsewhere, the clear exposition of its application to wave forms and their spectra is excellent. The statements of fundamental theorems of network analysis are also made very clearly in the early chapters.

A comprehensive list of references is provided at the end of each chapter and a list of symbols is also given. The book is well illustrated and is free from obvious errors. It is recommended as an up-to-date contribution to literature for the graduate engineer in radio, radar, television, and teaching.

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Books (continued)

Photoelectricity and Its Application by Vladimir K. Zworykin and E. G. Ramberg

Published (1949) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 478 pages+10-page subject index+6-page author index+xii pages. 6×9½.

In the early 1930's there were a number of books on photoelectricity which rather thoroughly covered the subject up to that time. Since then particular aspects have been covered, but there has been an increasing need for a more inclusive up to date summary.

"Photoelectricity and Its Application" emphasizes the great progress in recent years. For this reason the treatment is entirely new and not a revision of the former book on "Photocells and Their Applications" by Zworykin and Wilson.

The authors discuss radiant energy, thermal and gaseous sources, and photometric measurements. The general theory of the emissive effect is reviewed, and photocathode surfaces discussed in some detail. Two chapters are devoted to the materials and methods of preparing such tubes. The vacuum, the gas-filled, the multiplier, and the image tubes each have a chapter, as do the photoconducive, and the photovoltaic effects. More than half the book is devoted to numerous circuits and applications. Some of the newest applications are clearly described as the television camera tubes, light beam signalling, infrared detection, and the use of multipliers at very low light levels. There is a short chapfer on photo cells in the future, followed by an appendix of five tables covering the atomic elements, units and conversion factors, physical constants, and relative luminosity factors. MKS units are used throughout.

While the treatment is in general up to-date, a couple of comments may be made in this respect. On page 24 the new candle based on 60 candles per square centimeter for the brightness of a black body at the freezing temperature of platinum is referred to as an expected future use. Actually it has been an accepted standard since January 1, 1948, and corresponding to it, on page 477, 1 watt = 660 lumens of radiation at 5,550 A.U.

In the discussion of electrometer tubes and circuits on page 255 and following, one does not find a reference to an important bridge circuit of improved stability which utilizes a single electrometer tube containing two anode and grid structures with a common filament (G.E. Co. tube No. 5674). In this manner, fluctuations in the emission are balanced out, which is not possible with separate tubes.

The book covers so much ground that the treatment necessarily is condensed. Mathematical discussions are confined to footnotes, and the subject is clearly presented with excellent figures. It is well recommended for anyone interested in photoelectricity, and especially so for those concerned with the more practical aspects.

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Patent Law by Chester H. Biesterfeld

Published (1949) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 236 pages+5-page index+3-page appendix+18-page table of cases+2-page bibliography+viii pages. 6×9. $4.00.

Mr. Biesterfeld's book on patents, which is a second edition to the previous volume, covers the subject with a reasonable comprehension and explanation. It reviews the origin of the creation of the Patent Law of the United States under the Constitution and of the prior authorization in England, and gives the broad general background required to understand the theory and purpose of such legislation by Congress. It also covers, generally, each individual form of patent, although it stresses least those in the electronic field. The present trend toward electrical patents being applied in so many fields is not so thoroughly covered in the book as patents on other subjects.

The information given in the book under the topic heading is reasonably complete, and the author's explanation of varying decisions and of the trend of judicial determination of these conflicting considerations is most interesting, informative, and gives a very good view of how in many instances legislation and judicial construction is changing in the field, and how we might consider under each topic the present established rule and trend.

It is impossible, of course, in one volume to do any more than highlight the decisions indicating the judicial construction of the various questioned points involving patents and their adjucation and construction, but it would seem that within the confines of one small volume the author has done a very competent job, and the book does give any one interested in each of these disputed questions the general controlling decisions, and the author's conclusion as to their meaning.

The book would appeal to engineers, lawyers who practice not exclusively in patents, and inventors who need to know the general rules to aid them in their activities.

The book presents the facts clearly on each subject, to some extent following the general presentation of a college lecture course, and is easily readable and can be understood by anyone having the slightest background or education in the subject. It is carefully laid out on the various subjects and is in no way repetitious. The book, being a re-issue and a second edition of an existing one, has had all possible inconsistencies ironed out, and follows through on the various details of patent construction and decisions in a very smooth and uniform development.

This reviewer's conclusion of the book is that it is a worthwhile edition, bringing the first book up-to-date, and that the author's conclusions as to the varying decisions and trends of the courts construing patent law are worthwhile, and will reward any reader who examines the book thoroughly.

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Electrical Transmission of Power and Signals by Edward W. Kimbark

Published (1949) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 439 pages+21-page index+xi pages. 215 figures. 6×9. $6.00.

Written for undergraduate students in electrical engineering, the text assumes knowledge of the usual courses on dc and ac circuits with lumped parameters. The purpose of the text is to give the students their first detailed treatment of circuits with distributed parameters.

The book deals with basic transmission theory and its applications to three particular fields: power, telephony, and ultra-high frequencies. To avoid unnecessary duplication and save the student time, a feature of the book is the idea that these three topics may be advantageously developed from a single theoretical treatment, and the book has been written with this underlying objective.

The theoretical treatment has three main divisions: transmission line parameters, steady state phenomena, and transient phenomena. The chapters which deal with transmission-line parameters occupy 86 pages, and deal with various types of lines ranging from simple two-conductor lines to more complicated arrangements using stranded and ground-return conductors. The chapters which logically fall under steady state phenomena occupy 143 pages. The topics covered are: smooth transmission lines, lumpy lines, transmission line charts, impedance matching filters, and skin effect. The theory of smooth transmission lines is approached through setting up and solving the differential equations instead of the treatment wherein a smooth line is regarded as the limiting case of a recurrent ladder network. The treatment of the other topics is conventional. Transient phenomena are covered in a single chapter of 34 pages. Some of the more important transient phenomena are described and formulas are developed for the transients in loss-free lines.

The application of the theoretical treatment to particular types of transmission systems is dealt with in four chapters which occupy a total of 119 pages. These chapters are entitled,"Electric Power Transmission," "Telephone and Telegraph Transmission," "Radio Frequency Transmission Lines," and "Waveguides." The material presented in these chapters is largely of an introductory nature.

Another noteworthy feature of this book is the 51-page appendix which lists the significant characteristics of a wide variety of transmission lines. This should prove valuable not only to students, but also to those working in the fields of communication and power transmission.

In general, this new volume fulfills the objective of rigorously setting forth fundamental transmission theory in a straightforward manner, and shows how this theory may be applied to widely different types of transmission systems.

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Books (continued)

Electronic Time Measurements, Edited by Britton Chance, Robert I. Hulsizer, Edward F. MacNichol, and Frederick C. Williams

Published (1949) by McGraw-Hill Book Co., Inc., 330 W. 42 Street, New York, N. Y. 525 pages+10-page index+xviii pages. 356 figures. 6×9¼. $7.00.

One of the most important functions of a radar system is the accurate location of the objects from which reflections take place. In order to do this, an accurate measure of the time interval between emission of the pulse from the transmitting antenna and reception at the receiving antenna is required. In the so-called hyperbolic navigation system an accurate measure of the time interval between the arrival of two or more pulses is necessary. This book covers the fundamental factors involved in making both such measurements and describes the status of practical apparatus for making them during and up to the end of World War II. Although most of the material is based on the author's personal experience at the MIT Radiation Laboratory, a certain amount treated more sketchily is based on the developments in other laboratories.

The arrangement of material is logical and is exceptionally well co-ordinated for a book written by so many individuals. Although the book should be easy reading for technical graduates with a good understanding of nonlinear circuits, the less initiated will find a prior reading or at least reference to Vol. 19 of The Radiation Laboratory Series useful, if he is interested in an understanding of circuit details in addition to the broad aspects of the material that is covered. The treatment is descriptive rather than mathematical, and is replete with detailed circuit diagrams showing the application to numerous equipments.

In the early part of the book a general review is given of various systems for measuring distance and speed by the timed transmission of radio waves, including phase- and frequency-modulation systems, but the remainder of the volume is devoted almost entirely to the techniques used in pulse apparatus. The general techniques used for the timing of pulses is covered in one chapter. Following this are three chapters concerned with the generation of fixed and movable indices, i.e., covering essentially the generation of a series of accurately determined scale markers. Following this the methods are described which are used in systems where the determination of the timing of the received signal depends largely on operator manipulation or observation. The various types of cathode-ray tube displays are included in this section. The next two chapters deal with methods for determining the time of arrival of the received pulse by means of equipment whose accuracy depends in the main on an automatic positioning of the reading or operating index by means of the received pulse. In this section angular position measurement is also treated briefly. The last three chapters do not follow the logical sequence of the book too well, and could, with equal or perhaps better logic, have been placed in one of the other volumes of the Radiation Laboratory series. Chapters 10 and 11 cover Special Data Transmission Systems and Relay Radar Systems, and Chapter 12 deals with Delay and Cancellation of Recurrent Wave Trains. Although the use of storage tubes for this purpose is briefly referred to, it is not treated in detail and the chapter is mainly devoted to the mercury delay line system. Readers interested in storage tube or electrical circuit methods would do better to read chapters in Vol. 19 of the Radiation Laboratory Series.

"Electronic Time Measurements" will for many years be the outstanding reference work on the subject, and should have a place in any library where there is an interest in the development of circuits for this purpose.

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Invention and Innovation in the Radio Industry by W. Rupert MacLaurin

Published (1949) by the Macmillan Company, New York, N. Y. 265 pages+6-page index+7-page bibliography+xvi pages. 5¼×8¼. $6.00.

Based upon a series of studies of invention and innovation, financed through a Rockefeller Foundation grant, the volume is an excellent exposition of the impact of invention upon the structure of the radio industry. It is not a history of radio, but deals in part with the inventions of some of the pioneer scientists and engineers.

Distinctions are made between the research scientist, the inventor, and the business innovator in their respective relations to technological changes, and stress is placed on the author's opinion of the value of scientific research in industrial organizations. The author states: "In the United States it was not until the large and well-established electrical companies turned their attention to radio that research became more business-like and more co-ordinated. At the same time, it lost some of its spark and originality."

Mr. MacLaurin includes considerable authentic material dealing with the statistical history of radio manufcturing companies, and the economic, industrial, and patent circumstances which affected their growths, or which accelerated their liquidation. He presents useful information about the functioning of industrial laboratories, manufacturing processes, and merchandising and sales. The book includes a considerable amount of well-organized data on the general subject of radio patents. There is statistical material dealing with various of the early struggles in the courts in the connection with the legal establishment of priority of discovery.

The author certainly was not at pains to minimize Marconi's technological contributions to radio, even though in particular paragraphs he emphasizes Marconi's use of researches and inventions of others. The record presented of de Forest's radio achievements is excellent, but is not enhanced by reliance upon a discredited, scurrilous article which appeared in a popular weekly a few years ago.

There is a discrepancy in the name of the inventor of the Ultraudion circuit, but all in all the book is remarkably authentic in the coverage of invention and discovery. It should find wide acceptance among radio executives, research scientists, engineers, and students. In the text there is high inspiration, for in the accounting of what has thus been accomplished in radio, television, and related arts, a vision takes shape of what lies ahead in the way of opportunity.

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FM Transmission and Reception by John F. Rider and Seymour D. Uslan

Published (1949) by John F. Rider, Publishers, Inc., 404 Fourth Ave., New York 16, N. Y. 30 pages+10-page appendix+4-page bibliography+5-page index+vi pages. 201 figures. 5¼×8¼.

This is the ninth printing of a book which was first printed in 1948. It differs from previous printings chiefly in the inclusion of a set of questions at the end of each chapter, to make the text more useful to technical schools, as well as to the reader who uses the text for self-study.

The book is divided into two parts. In the first, the underlying theory of frequency and phase modulation is discussed as well as the propagation of FM signals, the basic characteristics of FM transmitters, and an analysis of those in use today. The coverage of FM transmitters is especially complete.

The second part of the book discusses the latest types of transmitting and receiving antennas. Every stage in an FM receiver is explained carefully, with special attention to the four different types of FM detectors. The last two chapters take up the alignment and servicing of FM receivers.

In discussing actual hardware and circuits in transmitters and receivers, the authors do a very workmanlike job. The same cannot be said of the first two chapters, on fundamental theory. Here many pages are spent trying to explain the difference between "phase modulation" and "frequency modulation." Since neither system of modulation is often used in its pure form, it seems unfortunate that the authors felt it necesary to perpetuate the apparent differences, which usually only lead to confusion in the mind of the student. In an effort, as stated in the preface, to keep mathematics at a minimum, the instantaneous frequency is erroneously defined as the reciprocal of the period. This definition leads to a wrong result in the example given on page 15. It would seem that a student in a technical school should be mature enough to be introduced to the concept of instantaneous frequency as the time rate of change of the carrier phase angle, but nowhere in the book is this brought out explicitly.

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Clarity in Technical Writing*

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Summary—Logical sequence of ideas and simplicity of literary style are the fundamental basis for clarity in technical writing. If a paper follows the general form of introduction, development, application, and conclusion, it should be readily understandable.

I. Introduction

IN TECHNICAL PAPERS intended for publication the material must be well organized and the ideas skillfully expressed if the all-important goal of clear, effective writing is to be achieved. Numerous postwar research and development projects have stimulated widespread publication of results in every field of engineering, and, because of this mushrooming growth of technical literature, all forms of engineering writing must now meet increasingly higher standards of quality. The purpose of this paper is to discuss briefly some of the principles of organic structure and literary style which are essential to clarity in technical writing.

II. Organization of Material

If a writer is to accept the responsibility for making his paper unquestionably clear, he must carefully arrange his written material in some orderly sequence and build it upon a logical framework. There is certainly no standard outline upon which all forms of scientific and engineering writing can be based; the broad principles discussed here, however, may serve as a general guide.

The Summary: The purpose of a summary or abstract is to give the reader an immediate understanding of the entire piece of writing. Since a good summary will often consist of only one or two short paragraphs, it must deal exclusively with the essential theme. There should be no attempt to provide introductory background material, to present experimental details, or to support the writer's conclusion or final results. The abstract instead should be a triple-distilled essence of all the material presented in the paper.

The Introduction: While the function of the summary is to condense the entire subtance of the paper into a few sentences, the introduction serves a different purpose. This section should begin with a general orientation of the reader by a short historical review of related work, or by an explanation of why the project was undertaken. These introductory remarks should then lead up to a precise statement of the problem. After the reader has been oriented and the problem well defined, the introduction can be brought to a close with a brief explanation of how the problem is to be attacked.

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The Process of Development: Having laid a sound foundation for the construction of the main body of the paper, the writer can next proceed to develop the fundamental idea. Whether the article is concerned with a basic research project, an engineering development, a general survey, a mathematical derivation, or any other technical treatment, the separate ideas should be presented in logical succession. The interests of clarity can best be served by arranging the subject matter so that there is an obvious relationship between cause and effect and a progressive growth of the central idea. An example of this type of development is a description of a piece of laboratory apparatus followed by a discussion of the actual experiment with the apparatus. Another example is design analysis of specified equipment, followed by a description of its construction. Still another is an evaluation of the factors comprising an engineering problem and a discussion which will lead up to recommendations for its solution. Regardless of the nature of the topic, its development will be aided by some orderly progression of ideas, such as from the known to the unknown, the simple to the complex, or the component parts to the integrated whole. In forging the chain of logic the writer must include every link.

The Application: After a fundamental idea or process has been introduced, explained, and logically developed, an indication may be given of how it may be applied, since the ultimate goal of all technical work is some sort of practical application. Frequently an abstract idea which is difficult to explain will be clarified when the writer shows its relationship to practical considerations that are easily understood. Moreover, a paper dealing with a tangible process, such as the development of a new measuring technique, should present a clear picture of the new method by showing how it may be applied in practice and how it is superior to previous techniques. Likewise, a report on design work should give an indication of the applications of the product or equipment. Since most technical writing is concerned with a highly specialized concept of science or engineering, the theme will often be more clearly understood if its relation to actual practice is made clear.

The Conclusion: While similar in scope to the abstract, the conclusion is written for a somewhat different reason. The abstract tells the reader in advance what the paper contains; the concluding section may either summarize, interpret, evaluate, or make recommendations. Usually, it will be weighted with opinion, which is seldom the case in the more factual abstract. Other functions of the conclusion are to indicate the trend of future work and to make any necessary acknowledgments. When the concluding section of a technical paper is inadequate, the writer has neglected to clarify his central theme by a recapitulation or at least a commentary on the significance of the work. No matter how well the information on the subject has been presented in the body of the article, the concluding portion will contribute added strength and meaning if properly written.

The proportions of the paper to be devoted to the introduction, development, application, and conclusion will depend entirely upon the scope of the subject and manner of treatment. In fundamental research, the process of development is generally the lengthiest portion, with little emphasis on applications. An article describing new products or machinery would, on the other hand, more likely devote most space to the section on application. Survey articles and reviews of the literature deal largely with information ordinarily found only in the introduction and conclusion. Regardless of how the various sections may be proportioned, however, the general principles of organization suggested here may be applied to any type of technical writing.

III. Relation of Style to Meaning

When the material for the paper has been organized on a sound basis, the writer can turn to considerations of literary style. There is an important relationship between style and intelligibility, and the merit of a technical article is based upon the manner of presentation, as well as the content. When dealing with difficult or complicated concepts, the tone must be adapted to the particular class of readers for which it is intended. Where a discussion of a highly specialized or difficult topic is written for popular consumption, the author can treat the subject as if it were pleasant and comprehensible, rather than create a barrier to understanding by stressirg the complexity of the problem.

In addition to tailoring the article to fit the requirements of clarity of any given group of readers, the writer should likewise emphasize some portions of the piece and subordinate others, depending on reader interest. For example, recent literature in physics journals on the transistor crystal amplifier deals mostly with physical theory, while the articles about the same crystal amplifiers in the engineering magazines emphasize practical applications. The process of subordination of ideas is vitally important not only in the over-all construction of the paper, but also in the details, down to the organization of paragraphs and even sentences. An example is the following sentence: *The gears did not mesh properly, and the machine would not operate.* The true relationship between proper meshing of the gears and the operation of the machine is not immediately clear. Subordination of the first clause to the second could be accomplished by a more lucid construction: *Since the gears did not mesh properly, the machine would not operate.* A little care in properly subordinating ideas throughout the paper will greatly enhance the clarity of meaning.

Another important aid to intelligibility is the use of transitional words, phrases, and sentences to serve as connecting bonds between separate ideas. Expressions such as *moreover, in addition to, however, for example, yet,* and *on the other hand* are especially helpful in the section dealing with the process of development. In technical writing, connectives should be used liberally as an aid to coherence and unity. The writer's words tend to reflect any discontinuity of thought or mental uncertainty; the deliberate use of transitional constructions will therefore serve the double purpose of clarifying the author's own thinking and of providing a smoothly flowing procession of ideas for the reader.

Finally, the keynote of all technical writing should be simplicity. If the subject is highly technical, the use of correct terminology is essential, but this is not to be confused with the use of ornate or intricate expressions. If, indeed, the chief concern of the author is to impress his readers with high-sounding wordage, he may well succeed in doing so, but the essential meaning of his paper is likely to be obscure. There is no effective substitute for a simple, direct style of writing if clarity is the ultimate aim.

IV. Conclusion

One of the prime requisites for lucidity in a technical paper is a basic structure consisting of introduction, development, application, and conclusion. If the ideas are presented in logical sequence, and if the literary technique is simple and direct, the goal of clarity will be achieved. One criterion of the intrinsic worth of any piece of technical writing is the manner in which it is constructed, because its completeness of information and creativeness of thought will be of little value to the reader if he cannot readily understand the written word.

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The Engineer and Industry*

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Summary—This paper is directed to the engineer just beginning his career in an industrial firm. Information regarding the organization of such firms and how the engineer fits into it is covered. His typical reactions observed over the years are shown. The importance of co-operation, human relations, and trust in others is emphasized. The time required between the birth of an idea and a product is reviewed in narrative form. The importance of the engineer in society is given.

There have been many discussions of the technical requirements of an engineer, but not enough has been said about the other parts of an organization into which an engineer must fit. This paper is intended to explain some of the reactions of an engineer in industry, and to present the views of both the management and the engineer.

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School to Industry

The engineer and his mental processes are important to himself and to management. The thoughts and reasoning of an engineer from the time of his graduation from an engineering college through several years of experience in an industrial firm can be arbitrarily divided into four parts for the purpose of this discussion. The four parts in chronological order are: (1) idealism, (2) disillusionment, (3) enlightenment, and (4) realism.

The period of time required for each part varies with individuals and conditions. It is not maintained that all engineers can divide their career into these four parts, because the problem is too complex and subtle to allow such an arbitrary attack. However, it is believed that the four parts are roughly typical of a large percentage of engineers. It also appears that veterans who have had war experience before completing college will enter industry with a more mature viewpoint. This condition alters the phases of the engineer's career, so that it does not so closely follow the parts outlined in this discussion. However, the parts listed are typical enough to warrant expansion and discussion.

Idealism

The engineer graduates. He is full of boundless enthusiasm and ideals. He is anxious to start, obtain his just rewards of satisfaction, and receive material gain for his long years of sacrifice and study. He also wants to make his contributions to science and engineering. Upon obtaining a job as a junior engineer in a firm doing electronic development and manufacturing, he is ready to "go."

He thinks of the firm that he has selected as a great giant of efficiency and marvelous undertakings. Everything is done perfectly, everyone understands him, untold facilities, instruments, and people are at his command. In other words, he is awed and impressed. He absolutely relies on the greatness, integrity, and perfection of the company of his choice. This blissful state may last one or two years and it may only last a few months, but it is not permanent for the average man. He rapidly enters the most depressing phase of his career.

Disillusionment

The engineer is now beginning to see the imperfections and gross inefficiencies in his firm. It is beyond his powers of comprehension how management can be so inefficient, short-sighted, heartless, "plain dumb," and still exist. He sees errors in judgment, wastefulness, unfairness, and lack of competence on the part of his associates, supervisor, and management, including all other departments. Everything is done wrong, no one will allow him to make it operate properly or listen to his ideas. If business is run this way, it is no wonder that the world is in such a sad state, he reasons. He complains about everything, but still goes along and tries. He either reaches the next stage soon, or gives up and tries another job. In one time out of 100 he was right, and has selected the wrong Company; but in most cases, if he adopts this philosophy and retains it, he will flit from one position to another and end a failure. However, most men outgrow this stage soon, and in a few years regain some of the ideals and enter the third part of their careers.

Enlightenment

The engineer has begun to realize the value of expediency and compromise. He has learned that any organization is composed of human beings, and can be only as good as the aggregate of these individuals. The value of other types of work becomes apparent to him. He finds that each person has his own ideals, beliefs, and troubles. All people are not equally capable, honest, industrious, or conscientious, and they never will become so. There are superior, average, and poor employees. Any organization which he can find will be similarly constituted, because people are human. The management is not perfect, but they also have their problems, and are usually honestly doing their best within their own experience and ability, and with the people and facilities available to them. He now has learned that a Company is not a machine. It is people, working together co-operatively for a common purpose: the sale of a product at a profit to all concerned. His profit is in the form of salary, prestige, satisfaction, and the feeling of accomplishment. The customers' profit is in the form of a useful product at a reasonable cost. The management's profit is the same as his; money, prestige, and satisfaction. The stockholder's profit is a return on the investment, which represent savings over a long period of time.

In other words, he has learned the value of his own work and the part it plays compared to others' work; the value of tolerance and understanding; the necessity for rules, procedures and regulations; the importance of the work of all of his associates who are not engineers, such as accountants, salesmen, assemblers, buyers, and many others. He cannot create or build a product alone, but all together, the people in his firm can and do. His technical judgment and knowledge have increased to the point that his superiors learn to rely on it. In other words he "belongs," perhaps not as idealistically as when he began, but his enthusiasm is more valuable. He is a recognized part of his organization, contributes his work as a part of a team knowing that he is of no value alone; and that the others are of no value without him.

The engineer is in the rewarding part of his career. He may not be at the peak in earning capacity and his use of leisure time; but he is ready for the fourth and ultimate part of his career.

Realism

The engineer is now a vital and smooth-working part of his organization. He may or may not be the chief engineer, but he is definitely responsible for certain products and truly has a wealth of facilities, equipment, and people at his command. Everyone is helping him and he is helping everyone else, be they engineers, salesmen, accountants, or toolmakers.

He listens to the other man's viewpoint, carefully evaluates it as related to his own, and then acts. True, he makes compromises every day. Sometimes a technical nicety is sacrificed to fit a cost estimate or a machine, or even an idiosyncrasy of a customer, salesman, or of his boss; but he knows that it is all for a purpose in the end—the sale of a satisfactory product at a profit.

He realizes that modern industry is complex and requires a well-integrated team of engineers, assemblers, toolmakers, salesmen, accountants, supervisors, buyers, and countless others; and he respects each one of them for his knowledge and experience in a particular field of modern business.

He has also learned how to make many decisions in a day without endless arguments on the pros and cons of the situation; and he is teaching less experienced people how to do the same.

He has learned the limits of his mind and body, and has reached a happy balance in work, leisure, and family relations. People like to work with him and know that if he is overruled or his idea rejected, he will not sulk and brood for days. In other words, he belongs! He is happy, respected and valuable. He is well-paid in salary, prestige, and a sense of well-being as a part of a successful organization. But above all, he is tolerant of others, because he knows that he is not perfect himself. He has reached the ultimate in his career and still has many years of successful living ahead. He has reached the fourth and last stage of his career.

The total time required to go through the three parts of the engineer's career will vary, but is usually from five to ten years. Of course, some engineers never reach the maturity and human understanding, power of expression, co-operativeness and compromise, combined with engineering proficiency, required to become happily and profitably a vital part of an industrial organization.

They either drop out, become a successful "lone worker," or are a continual source of irritation to themselves and management, and cannot understand why they are never promoted.

Adage

This is a good point at which to stop and reiterate this age-old advice to the young engineer: "You may be the best engineer in the world, you may have more ideas than anyone else, you may be conscientious, hard-working and brilliant; but if you stay in a corner and never sell your ideas to your associates, no one, neither your firm nor civilization, will get the benefit of your genius, nor seek you and your ideas. You must learn to express yourself, both in writing and verbally. You must work with and understand people. Too much emphasis cannot be placed on the necessity of adaptation and reasonable compromise."

From Idea to Product

A problem to the young engineer worthy of discussion because it is so often a cause for concern to him is: Why does it take so long from the proposal of an acceptable idea for a product, until it is commercially on the market; and why are so many ideas that management admits are good never used?

We must remember that there is far more to the successful sale of a product than its technical excellence. It must be available at a time when there is a demand for it. In other words, correct timing is essential. An idea or a product may be on the market either too early or too late. It does no good to offer the best product in the world if there is no demand for it. The demand can be created either by advertising and publicity, or it may naturally exist.

The product must also be competitive in price, it must be styled to be attractive to the particular customer, it must have performance and quality designed and built into it to do the job intended. It must be no better and no worse than required, or it can't be competitive.

The product must be manufacturable by the plant at your command, and it must be within the scope of capabilities and experience of your co-operating departments, such as sales and factory.

It is impossible to list all of the requirements that must be met before a particular firm can undertake to design, manufacture, and sell a new product. The inexperienced engineer cannot hope to understand and be competent in all of the necessary fields. Therefore, he must have the confidence in his management and his associates warranted by their greater experience and their successes, proven by the firm's existence.

To use a specific example which could occur, let it be assumed that the engineer works for a medium-sized firm that is engage in the design, manufacture, and sale of radio and television receivers. This firm employs several thousand people and has complete engineering, factory, and sales organizations. It also has complete manufacturing and laboratory facilities.

The engineer, during the course of his design work on a television set, has developed a new piece of test equipment for his firm's own use in the factory and laboratory. It does the job well and was developed because nothing was available on the market that would do the job satisfactorily. He has personally constructed four pieces of this equipment and made them operate properly. It appears to him that if his firm has the need for such a piece of equipment and it does the job so well, other firms and laboratories should need it and there should be a market for it. Accordingly, he proposes this to his chief, who admits it is a good idea and gives him permission to submit the idea to the president. The engineer is enterprising and thorough, so he has written up a complete description of the apparatus, obtained estimates of cost, and estimated the quantity that he believes could be sold in a two-year period. He believes that the firm should immediately start the manufacture of several hundred units, as he is positive that the design is complete and that he, personally, could sell them if the sales department cannot be made to do so.

The engineer, in this case, being unusually thorough and analytical, will be given the benefit of the doubt by assuming that his estimates are correct and that his enthusiasm is warranted. This is exceptional, because usually he only has an idea which is not thoroughly worked out. However, after consideration, the president rejects the plan, and sincerely attempts to explain to the engineer the reasons why the project cannot be undertaken profitably at that time.

Some of the reasons that he advances are (1) all departments are fully loaded and will be for another year; (2) the firm has all of the business that it can handle without expansion of facilities and personnel; (3) the sales department is fully occupied with a campaign and can give this no attention; (4) the sales organization is geared to large volume merchandising and would find it difficult to absorb a low volume, high cost, specialized unit; (5) the salesmen have no training in contacting the type of customer that would purchase this equipment and have no contacts with these customers; (6) the design is not complete enough to manufacture and would require considerable additional engineering and drafting man-hours to complete; (7) tooling, processing and cost estimates would require an investment out of proportion to total profit; (8) if it were agreed to overcome all of the above objections and go ahead, it would be over a year before the product would be available, and by that time the market would no longer exist in the same degree; (9 furthermore, there are specialty manufacturers in this field who can design, manufacture and market this product more effectively; (10) and unless a decision is made to go into this particular market and fully cover it competitively, it is unwise to market a single product in the field.

The engineer is disappointed and disillusioned because the president is well known for his progressiveness and willingness to take a chance. He cannot understand the objections because he is certain in his own mind that, if given the opportunity, he, alone, could overcome all of the objections and handle the sales himself, thereby making a profit for his firm that would not otherwise be made.

However, if he is a sensible young man, he finally realizes that his limited experience cannot compare with that of the president, even though he is not completely convinced that the president knows what he is doing. He realizes the president has operated the firm profitably for many years, and has attained excellent working conditions and wages for his employees.

On the other hand, the president, who was once young and inexperienced himself, feels a deep responsibility to his engineer. He has seen this happen many times before and he, himself, has tried ventures such as this previously, with success in some cases, but failure in many others. He is still tempted to take up the young engineer's idea, but finally lets his decision rest and the project is shelved. He knows that he has a problem with his engineer. He does not want him to lose such enthusiasm, but rather to encourage the engineer to propose further ideas and plans for products. However, if he is not careful, the engineer will retire into his own disillusionment and never propose another plan. The engineer, if he is really sensible, will see the common sense of the situation and continue to propose ideas. Nine out of ten of them will probably be rejected, but he cheerfully continues because the particular engineer is sensible and has the good judgment to rely on the experience of his management. To end the story, the years go by and several ideas of the engineer are accepted, one a tremendous success. The engineer is now the president, repeating the same story to his young engineers.

The moral of the story is: Rely on the better judgment and experience of your management and take the disappointments in your stride.

Conclusion

It is not my intention to repeat a long list of adages about how an engineer can be successful. Rather, I sincerely wish to give the young engineer an explanation of the operation of an industrial organization, and help him in some measure to fit himself into the system where he can accomplish the most—because in so doing he will be the most satisfied. The greatest satisfaction is accomplishment.

It will be of little value to describe in detail the operation and necessity for the various departments in a manufacturing firm; sufficient to state that accounting, inventory control, tool design, processing, purchasing, maintenance, inspection, cost control, sales, personnel, and many other departments are as essential as the engineering, sales, and manufacturing departments. Each has been adopted by experience and found essential. It is necessary for the engineer to realize the value of an organization, and learn that he cannot be an expert in all of these fields. He is an engineer and he must learn enough about the other departments to be able to co-operate with them; but he must leave inventory control, as an example, to someone trained in that field.

The engineers and scientists have the best chance in history to become a more important part of society and obtain the material things of life. Never has there been a more technical age. But they must do their part, learn their jobs well, do good technical work; but, above all, learn the value of good human relations. They must make use of their present advantage for their own happiness, prestige, and success, without harming other less fortunate people in society.

Beginning a Career in Engineering*

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Introduction

THE YOUNG MAN who has recently graduated from an engineering school, or is about to graduate, is frequently beset with many questions which his engineering education has failed to answer. Still other questions arise during the early stages of his career. These questions are not so much of a technical nature as they are concerned with the problems of self-management in relation to the field in which the engineer hopes to earn his livelihood.

This paper attempts to answer some of the more common questions on such subjects as "What kind of job should I try to get?" "How should I go about getting it?" "How can I improve my chances of success?" and "What about salary?"

Choosing a Job

Although the young engineer with little or no experience must approach the problem of choosing a job with a degree of open-mindedness, including a willingness to make his selection from the jobs which are available even though they may not appear to be exactly what he would like to have, the intelligent exercise of whatever degree of choice is permissible is extremely important. Wise or fortunate choice of the first job is a big step in the direction of a successful engineering career.

The most important criterion in the selection of the first job is probably the type of work involved and the opportunities which it offers for learning. Working for a sound and progressive company, location in an area providing pleasant living conditions, the starting pay offered, and opportunities for increased responsibilities and pay are all important, but learning opportunities should be the prime consideration. The first job is seldom the one in which an engineer will spend the majority of his life. He may change to another company providing higher pay and greater opportunities for advancement or located in a more suitable area, but the experience gained in the first job will always remain with him.

By the time of receiving his diploma an engineer has presumably decided, on one basis or another, which branch of engineering he prefers, whether electrical, chemical, mechanical, etc. He may also have selected a field more specifically, such as one of the branches of electrical engineering, perhaps radio engineering, power engineering, or illuminating engineering. In many cases, the young graduate has no further ideas as to preference, whether he would like research, development, production, or sales engineering, for example. Although initial preferences must of course be subject to later revision, a choice should be made in order that the field of endeavor may be that for which the individual is best suited by training and personality, and in which he will therefore probably be best satisfied.

The distinctions between research, development, production engineering, sales engineering, maintenance engineering, etc., have been explained in some detail in several published articles. In general, the engineer who particularly enjoys courses in mathematics and theory and who takes pleasure in new ideas and discoveries is probably well fitted for research or development. The man who enjoys doing something of a more obviously productive nature, such as actually building equipment for his own use, may prefer production engineering. One who has a gregarious nature and gets along well with others, particularly relative strangers, has a good start on a successful career in some field such as sales engineering, in which these abilities and talents play a large part.

The nature and amount of the supervision given the young engineer is of considerable importance in determining how much of value he will learn. Too much supervision is as bad as too little, since too much supervision tends to prevent the engineer from doing much work which is really his own in other than a routine sense. On-the-job learning should ideally be composed of about equal parts of learning by doing and learning from contact with others. Consequently, in selecting his first job the young engineer should ascertain that there is a real need for his engineering talents, but that he will not be required to solve difficult problems single-handed.

For the first job, it is wise to avoid the extremely small organization, employing only one or two engineers, as there is too little opportunity for learning from others. In organizations with perhaps half a dozen or more engineers, it is almost certain that the inexperienced man will have ample opportunity for contact with one or more persons from whom he can learn something of value. In the extremely large organization there will be several men who are authorities in their fields, but the very size of the organization ordinarily limits close contact of the young engineer to a relative few of these experts.

Some young engineers shy away from the very large organizations, feeling that there is danger of becoming lost among the many other engineering employees and that opportunities for advancement may therefore be limited. In general this is not the case, as almost all large engineering groups are subdivided into departments of more easily manageable size, the head of each department being autonomous to at least some degree. Advancement of the engineer occurs within his own department much as it would in a smaller organization.

On the other hand, suspicion is sometimes cast at smaller organizations on the basis that there are fewer jobs available at top levels. Here again, the suspicion is largely unfounded. It is true that the number of top-level jobs is limited, but at the same time the number of persons competing for these jobs is small, so that the individual has about as much chance of advancement in the small organization as in the large one. Even if the man should eventually outgrow the company, being capable of holding a job which is already occupied or does not even exist in that particular organization, he can leave the organization for another one with probable ultimate benefit to himself and his career as the result of the experience he gained in the smaller company.

Many of the larger organizations have training programs for young engineers, the purpose of these training programs being to familiarize the individual with all aspects of the company's business and to give him, in some cases, certain specialized technical or business training which he did not receive during his formal education. Such training programs are of great value to the young engineer.

Smaller companies, as a rule, have no such formal training provisions, but the small size of the organization makes it possible for the individual to become familiar with its over-all operation, including the various technical fields of concern, to a degree which is not possible in large organizations without some type of formal training program.

Getting the Job

Once one or more positions have been selected as promising in view of the young engineer's desires, the next step is that of actually securing one of them.

Most graduating engineers, particularly those from the larger schools, are offered opportunities for interviews with prospective employers at some time prior to graduation. This is an excellent method of making the initial contact, as the companies doing such interviewing obviously have a need for men of the type to whom interviews are offered. It is a mistake to assume, however, that interviews arranged through school authorities are the only suitable means of establishing contacts with employers.

Direct contact between the engineer and the employer is also a very promising possibility, as many organizations, particularly the smaller ones or those needing only a few new men, do not conduct college interviews. Direct contact is particularly desirable in the case of the man who has more to offer, especially in the form of actual engineering experience, than the average graduate.

Such direct contact should ordinarily be made by means of a letter to the personnel manager, chief engineer, etc., of organiza-

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tions for whom the engineer thinks he might like to work. The letter should preferably be rather brief, stating principally that the writer is interested in the possibility of obtaining a position and describing very briefly what he has to offer in the way of education and experience. Attached to the letter should be a more complete résumé of education, experience, etc. Instructions as to writing such letters of application are included in courses in engineering letter writing and business correspondence, which are available at almost all technical schools.

In any event, the engineer will almost always attend an interview with any employers who are interested in him before he is actually offered a position.

It is probable that no two executives conduct interviews in exactly the same manner and that no two executives look for exactly the same qualities in prospective employees. In general, however, the following qualities are considered important, the order of listing providing at least some indication as to relative importance in the eyes of most executives:

1. Capacity of the engineer to learn and to grow into bigger jobs.
2. Interest in his field.
3. Honesty and personal integrity.
4. Perseverance.
5. Personality.
6. Educational background and experience.
7. Extra-curricular activities while in college.

In addition, the applicant is expected to present a neat personal appearance at the interview. Very few things make such a poor impression on interviewers as the arrival at an interview of an applicant who has neglected to wear a fresh shirt or forgotten to comb his hair. The importance of neat appearance has been emphasized time and again, but a great many applicants apparently fail to pay it any heed.

The applicant should approach the interview with an air of assurance but without cockiness. He should see to it that the interviewer's attention is called to any special qualifications he may have, but should not brag about his accomplishments or attempt to make them seem more important than they actually are. Such subterfuges are readily detected by experienced interviewers and create a poor impression.

It is useless to attempt to hide such undesirable factors as poor grades, for example, as the interviewer who is interested can usually obtain the information he wants through the school authorities. The student with moderately low grades should not assume, however, that he is necessarily at a disadvantage when it comes to getting a job, as many organizations do not attach a great deal of importance to extremely high scholastic rank. Some, in fact, will not hire an engineer with the best grades for certain types of engineering positions because of a fear that the man may not be satisfied with work which does not demand the highest theoretical training and ability.

The man with mediocre grades should probably seek a position in production or sales engineering rather than research or development, as these latter fields require more ability and knowledge in such subjects as mathematics and other theoretical topics. In general, the man with mediocre grades, unless these grades have resulted from some such cause as the necessity for working long hours while attending school, will be neither happy nor successful in research or development even if he should be able to secure a position in these fields.

The engineer with very poor grades is almost always looked upon with considerable suspicion by employers. He will frequently do well to accept any position he is able to get and may, in fact, ultimately decide that he has no bright future as an engineer and should seek another type of work.

Discussion of salary will frequently arise during the first interview. In other cases, it may be left for later correspondence between the applicant and the employer. In any event, it is normal to express a healthy interest in the rate of pay involved, but too much importance should not be attached to it. Most organizations offer similar or identical salaries to all inexperienced engineers, since they generally have little information, other than that provided by educational background, on which to judge the individual.

The opportunity which offers the most in the way of possibilities for learning is usually the best, even though the pay may be a little lower. After a year or two on the job, the engineer will have had the opportunity to demonstrate his abilities to his employers and will almost always find that any initial inequality of salary will have been taken care of adequately.

Doing the Job

When he arrives to take over his new job, the engineer should resolve to make as good a first impression as possible. It is an ancient but accurate axiom that first impressions are the most lasting. The man who bobbles the first task to which he is assigned makes an impression that will be difficult to overcome, while the person who performs his first duties well can afford to make a few mistakes later without having his supervisors and associates regard him as necessarily stupid. A simple task done well is, of course, not as laudable as a more difficult one done equally well, but it is infinitely better than a simple task done in a mediocre fashion.

Employers frequently expect a new engineering graduate to be a liability for a time, so that the man who makes a good first impression comes as a pleasant surprise.

Of the many pitfalls on the road to engineering success, some of the most serious are lack of perseverance; lack of self-confidence; failure to accept responsibility or, almost equally bad, assuming too much responsibility; failure to keep the ultimate goal of the work in mind; jealousy of fellow employees; and unjustified antagonism toward supervisors or employers.

It has been truly said that brilliant minds easily become bored. Many otherwise capable engineers make an excellent start on each new job, but their interest gradually dies out as it loses its novelty. Perseverance is a very valuable quality in engineering, as in almost all fields of endeavor. A job well begun must be pursued to its end with equal diligence if it is to be of any real value.

Blind perseverance, however, is not enough. Each new job must be approached also with confidence. "If this job can be done, I can do it." The man who doubts his ability to do a job will almost certainly fail. It must be remembered that a supervisor will not assign a job to a man whom he does not consider capable, and that a difficult job is the best type in which to learn and to demonstrate ability.

A question which every engineer should ask himself frequently is "Why is this job being done?" In many cases, the basic answer is that the employer expects to make a profit from the work in some fashion. In other cases, such as in government or other nonprofit establishments, the profit motive is absent, the object being the protection of the country, the advancement of mankind, etc. In any event, the ultimate goal should exert a constant and significant influence on the course of the engineering work.

The engineer may frequently find it necessary to subordinate his own desires to do a technically perfect job to such considerations as time or cost. If this is the case, he should do so willingly, recognizing that it is necessary to the successful attainment of the actual goal of the work.

Continual recognition of the goal is difficult, the tendency being to lose oneself in the day-to-day details of a job. It is worthwhile to stop occasionally in order to reexamine the course of the work to date and to assess its probable future direction. In this way, it may frequently be found that difficulties which seem insurmountable can be overcome by virtue of a new approach which will serve the ultimate purpose equally well.

The matter of accepting responsibility is one which requires careful treatment. Most young engineers tend to accept too little responsibility rather than too much, but there are a few individuals who are willing to take the responsibility for anything, regardless of their ability or authority, and to give orders and suggestions to anyone who appears willing to take them.

Every engineer should familiarize himself with the details of his organization's operation in order to ascertain which functions are properly engineering functions and which belong to other departments. In no case should the engineer assume any responsibility or authority for functions, such as making purchase commitments, for example, which are properly within the jurisdiction of other departments. With regard to engineering functions, it is usually safe to assume responsibility in matters for which one is qualified and which are not known to be the responsibility of others. Supervisors should not be continually pestered for decisions on trivial matters. Except in instances in which an incorrect decision would have serious consequences, supervisors usually prefer that their subordinates do their work with a minimum of close supervision, and every man is, of course, allowed a certain percentage of mistakes.

A healthy attitude toward one's fellow

employees is important. Jealousy and hogging the credit have no place in a well functioning organization. The man who is always first in line when praise is in order is frequently the last to exert himself when there is work to be done, and he fools no one but himself. His associates resent his conduct and his supervisor will quickly find him out. The result may well be that he will fail to receive credit when it is due him, as others may suspect that he is again trying to steal the glory.

The man with a self-deprecating attitude may find that it takes a little longer for his achievements to come to the attention of those in authority, but he will be respected by his associates and will almost always ultimately receive the credit to which he is entitled.

Relations of the engineer with non-engineering employees are extremely important. In the course of his work, an engineer inevitably comes in contact with members of other departments whose training, abilities, and duties differ from his own. A haughty or dictatorial attitude must be avoided in such contacts, as it will result not only in loss of personal prestige but in failure to deal effectively with other departments. When co-operation or assistance from another department is desired, the engineer must be careful to present his needs through the proper channels, not short-circuiting established lines of authority. Once the necessary contact has been established, however, details may be handled on a person-to-person basis between the individuals involved. If the co-operation or service is unsatisfactory, the man responsible should be told first—not his supervisor. A difficult job well done, however, should be called to the attention of the supervisor if the opportunity is available. The person deserving the credit will be appreciative and will make every effort to do equally good work in the future.

In addition to maintaining a healthy attitude toward his fellows, the engineer must be careful not to develop resentment toward his supervisor or employer for injustices he fancies have been done him. If he takes the trouble to discover the facts behind the situation, he will usually find that the injustice was unavoidable and, frequently, that no injustice exists at all. In the event of a real injustice which appears unnecessary, it is still probable that it was not deliberate. In such a case, the matter should be called to the attention of the supervisor, who will almost always be found sympathetic and will do his utmost to have the situation corrected.

The engineer should always keep his supervisor's problems in mind. The supervisor has a large number of subordinates whose welfare he must look after, and he must be concerned with the smooth operation of his department as a whole, even if some inconvenience to individuals may occur in the process.

The man who expects to advance as an engineer must do his job to the best of his ability at all times. The dull, uninteresting tasks must be approached with the same industry which is applied to the more intriguing ones. Engineering is not a job one can leave behind at the end of the day. Overtime work, at home or at the office, is frequently necessary to the success of an endeavor. Competition for top engineering positions is keen, and only the most capable and most industrious achieve a high degree of success.

If advancement seems slow in coming, the engineer should not despair. It may be that a position of greater responsibility is not available immediately, but satisfactory performance, and particularly exceptional performance, is almost always rewarded.

Supervising Others

Sooner or later, the engineer can expect to be placed in a position where he will supervise other employees. Although this will not occur immediately upon taking his first job in most instances, some understanding of supervisory problems is important to the young engineer if he is to get along well with his own supervisor.

Supervision of others is one of the most difficult tasks for the engineer, possibly because his schooling seldom includes any training aimed specifically in this direction. Perhaps the most important thing to keep in mind is that all of one's associates, superiors, and subordinates are individuals like oneself, each having his own ambitions, likes, dislikes, and peculiarities. The successful supervisor must treat his subordinates accordingly, attempting to discover the particular abilities of each and assigning duties with these abilities in mind.

A common fault among poor supervisors is that they attempt to do too much of the work themselves, leaving only the simple tasks or the uninteresting duties to their subordinates. Such men fail to make the most of the abilities of their assistants.

A supervisor must not be afraid to admit that he is not omniscient. No man is expected to be the top expert on everything which takes place in his department. He is provided with a staff of carefully trained men and is required only to supervise their efforts in the interests of over-all effectiveness.

The supervisor should keep his subordinates informed of significant developments at all times. This tends to prevent the origin and spread of incorrect rumors and makes for better efficiency by placing the necessary information before those who are expected to act in accordance with it.

When praise is due, the supervisor should see that it is given to the persons deserving it and that these persons are aware that their performance has been noticed.

When criticism is in order, it must be carefully administered. Although praise may be given in public, successful criticism must be administered in private to avoid embarrassing the defaulter in front of others. The supervisor must approach the situation with a desire to be helpful rather than merely to castigate. He should attempt to explain the problem to the man at fault, to discover the true reasons for the difficulty, and to make helpful corrective suggestions. Above all, he should not threaten. A man who has been threatened with dismissal may as well be dismissed; he will be of little future use.

The Question of Pay

Frequently, the young engineer on his first job becomes disturbed because he feels that he is underpaid in comparison with other employees, particularly those having less education. He feels that he has invested several years of study, in addition to a considerable amount of money, in his education and that he is entitled to some compensation in excess of that received by other employees who have had little or no advanced formal education.

However, there are several factors which must be borne in mind for a true evaluation of the situation. In the first place, the engineer cannot expect to recover his investment in a short period of time from his first employer. Instead, returns will be received over the entire course of his career in the form of greater earnings in later years, a position demanding greater respect, and a job which gives him a large measure of satisfaction.

For the first several months, the young engineer is likely to be more of a liability than an asset, so that his salary during this time represents an additional investment in him on the part of his employer. In addition, the economics of business dictate that each employee must be paid in accordance with his worth to the organization. A good machinist, tool maker, or even a truck driver cannot be replaced by the average engineer at any price, and these men are frequently just as important to the organization as is the engineer.

Most organizations have established times for granting pay increases to deserving employees. The young engineer can expect, if he does his work satisfactorily, to receive periodic increases amounting to from one or two hundred to as much as four or five hundred dollars per year during the initial years of his career. After this period, he is on his own, receiving increases in both salary and responsibilities strictly on his own merits.

The man who feels that he is underpaid, after considering matters carefully, may take the subject up with his supervisor a month or two before any increases are due. He will usually find that his supervisor is aware of the injustice and that arrangements have already been made to correct it. At any rate, the supervisor will almost always be willing to listen to reasonable requests for salary increases if they are presented at such a time as to fit in with the organization's policies in such matters.

The engineer must not expect to grow rich in the engineering field. It is not a notably high-paying one, and the man whose principal interest is money would do well to cast his lot elsewhere.

An Experimental Large-Screen Television Projector*

P. MANDEL†, ASSOCIATE, IRE

Summary—A television projector using a high-voltage cathode-ray tube is discussed. The modulation circuits, the projector system, and the directive projecting screen are described in detail.

Introduction

THE PROBLEM of reproducing large television pictures covering a surface of about 10 square meters or more, with the same brightness as is usual in the case of ordinary motion pictures, has not been satisfactorily solved up to the present. This has been due to the economic necessity of first producing a highly developed field-tested home receiver; the necessity of developing a large-screen television projector being less urgent. However, there can be no doubt concerning the advantages of a simple and satisfactory television projector to the motion picture industry, to medical and technical institutions, and to educational organizations. With this in mind, research work was undertaken by the author and his coworkers in the Montrouge laboratories of the Compagnie des Compteurs several years ago with the purpose of obtaining a satisfactory solution to the projector problem.

As a result of this work, some questions could be solved in a satisfactory manner, while others are yet waiting for a more acceptable solution. The author feels, however, that some parts of the experimental large-screen projector would be of certain interest to the workers in the same field, and that the publication of the results obtained would contribute to the progress of the television art.

Choice of the System

It was felt at the beginning of the work that the use of a cathode-ray tube in the classical, or eventually in an improved, form would lead to a satisfactory solution, with the exception that, in the case of extremely large screens, the Eidophore projector would be better able to produce the necessary luminous flux. Consequently, the apparatus comprises the high-frequency picture- and sound-receiver, the sync-separator circuits, the projector, including the modulation, deflection, and various auxiliary circuits, the high-voltage generator, and the control desk, as shown in Fig. 1.

Two identical units form the complete projector so that, in the case of failure of any part of one unit, the second unit can be immediately switched on, thereby assuring continuity of projection.

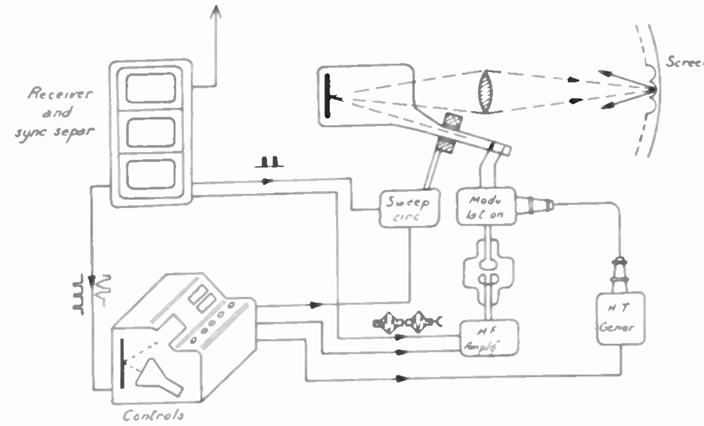

Fig. 1

The Video Channel

The apparatus is designed for the reception of the French low definition standard, i.e., 450 lines composing fifty fields and twenty-five complete pictures per second, and the picture being transmitted with both sidebands on the carrier of 46 Mc, with the accompanying sound on 42 Mc.

The high-frequency section of the video receiver is of the straight amplifier type, using vestigial sideband transmission, as represented by Fig. 2.

In order to obtain the highest possible signal-to-noise level, and in order not to deteriorate the background in certain cases by the spurious components generated by the harmonics of the carrier and of the local oscillator, no use was made of frequency changing. The

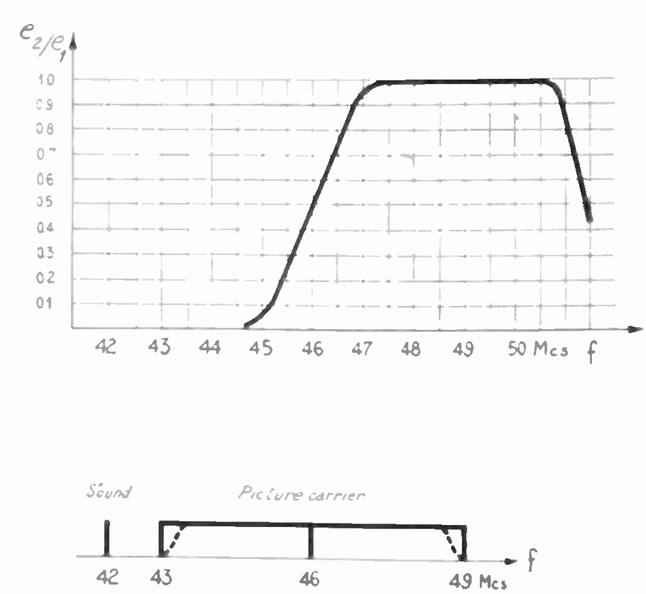

Fig. 2

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† Radio-Industrie, Paris, France.

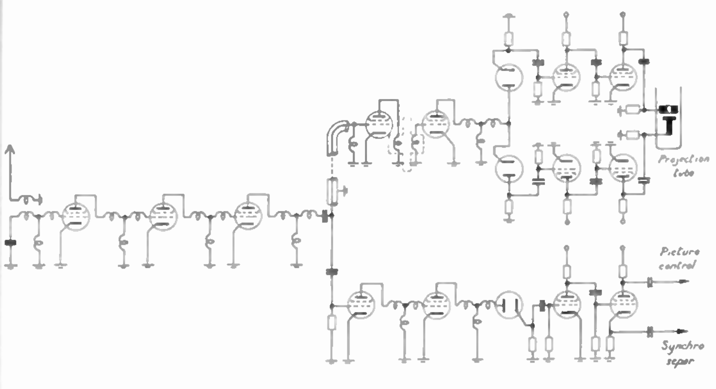

Fig. 3

general layout of the receiver is shown in Fig. 3. The receiver feeds, besides the projector, the control kinescope, and the sync separator.

The plate of the projection tube being at ground potential for reasons explained later, the introduction of the modulation signal on the tube, the cathode of which has a potential of some minus 80 kv, had to be made in a somewhat unorthodox manner. The video signal is transmitted by a high-frequency filter of a particular mechanical construction, as represented by Fig. 4.

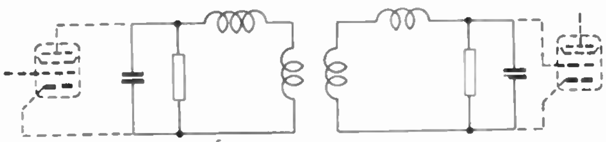

Fig. 4

The inductances of the critically coupled high-frequency transformer are disposed in the interior of an evacuated container. Short sections of a co-axial cable connect the inductances to the plate of the preceding, and to the grid of the following amplifier tube, whose output and input capacitances, shunted by resistors of convenient value, tune the filer. The form of the container and the distance between the inductances are chosen to support the potential difference of 80 kv with a sufficient margin of safety.

A slightly different coupling device, shown in Fig. 5, could be successfully used for the transmission of carriers with a bandwidth of 15 to 20 Mc. It consists of a half-wave section of a parallel high-frequency line, interrupted in the middle by a pair of small capacitors, whose separation is large enough to support the dc potential difference of 80 kv, once the device has been immersed in oil, or has been disposed in an evacuated container. The device is tuned by the output and input capacitances of the amplifier tubes.

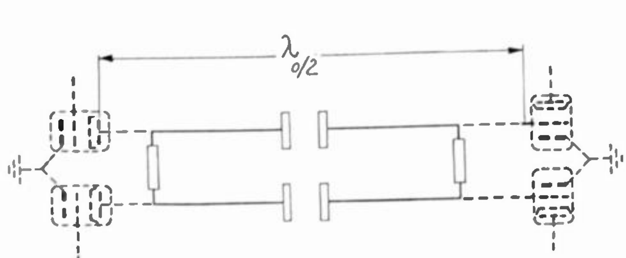

Fig. 5

In order to obtain the relatively high voltage (about 250 volts peak to peak) necessary for the full modulation of the projection tube without perceptible nonlinear distortion, a push-pull class-B arrangement was chosen, as shown in Fig. 6.

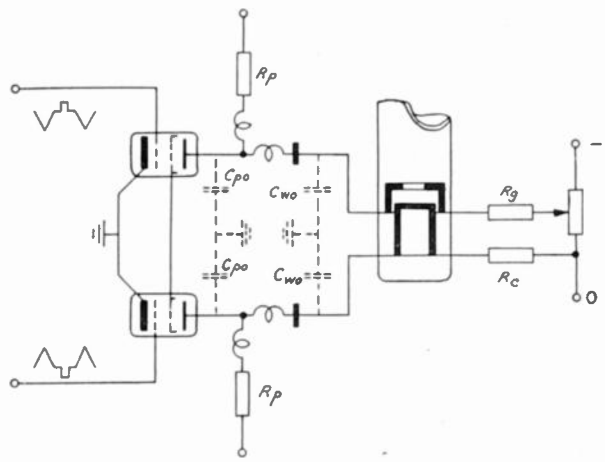

Fig. 6

By this means, not only may the admissible plate resistor (R_p) be doubled, but the power stage of the amplifier may be fully utilized from cutoff with a full grid swing without any appreciable distortion caused by the curvature of the tube characteristic. The two signals in opposition are obtained by two separate video detectors fed by the last stage of the high-frequency amplifier. The frequency characteristic of the video section is shown in Fig. 7.

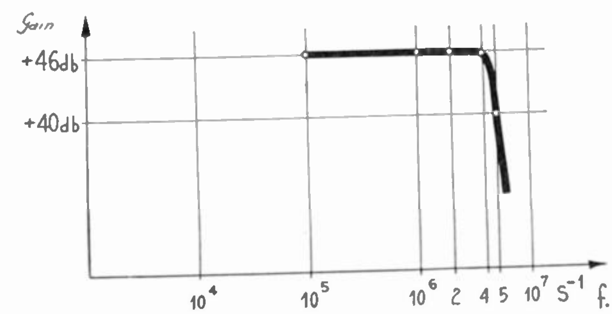

Fig. 7

The Sync Separator

It is a well-recognized fact that the advantage of interlaced scanning can not be fully appreciated without a correct interlacing of the two frames on the screen. It was recognized that in order to obtain this result under practically every conceivable circumstance, it is necessary that the whole system possess the following properties:

(a) The vertical sync-signal must have exactly the same shape after each field repetition. The foregoing concerns the form, the level, and the composition of the signal.

(b) After separation there should be absolutely no traces of horizontal sync pulses in the signal intended for vertical synchronizing.

(c) After separation there should be absolutely no traces of vertical sync pulses in the signal intended for horizontal synchronizing.

(d) There should not be the slightest interaction between the horizontal and vertical scanning generators.

(e) The vertical deflection must be constant within 2°/₀₀ of the total sweep amplitude.

(f) To insure the correct functioning of the sync separator, no manual adjustments are allowed. The sync separator should deliver synchronizing pulses absolutely identical in form and in amplitude, if the input of the receiver is superior to 500 μv, and inferior to 500 mv, regardless of the content of the picture. While (a) must be satisfied by the pickup equipment, (b) to (f) must be met by the receiving apparatus. Fig. 8 represents schematically the circuits used in the sync separator.

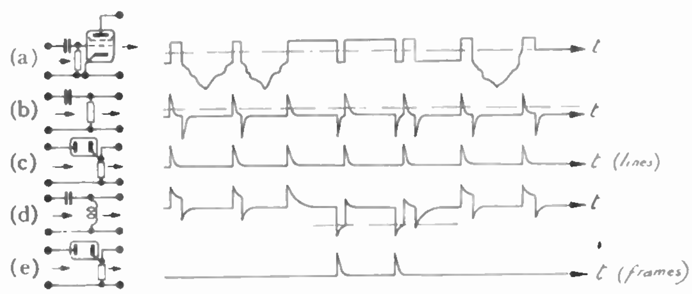

Fig. 8

The inverted video output is applied to the automatically biased grid of a high-slope pentode having a cutoff of only a few volts. In the case where the amplitude of the sync pulses is greater than the cutoff voltage, the output of the stage is independent of the amplitude of the input, and consists of the sync pulses as represented by the portion above the axis in Fig. 8(a). A differentiation by a very short time constant 8(b), followed by a preset clipping 8(c), permits the extraction of the horizontal sync pulses entirely free from any trace of the vertical sync pulses. On the other hand, the composed sync signal 8(a) is differentiated by a longer time constant 8(d), and is transformed after a preset clipping into a vertical sync signal 8(e), without any component possessing the line frequency.

The Projector Tube and its Auxiliary Circuits

Preliminary calculations concerning the size of the crossover, the obtainable current density in the spot, and the mutual repulsion of the electrons in the beam indicated that the necessary luminous flux can be obtained by increasing the final velocity of the electrons in the beam to very high values, say, up to 80 kv.

Experimental evidence indicated that, for a given type of electron gun, for a given number of lines, and for a given size picture, the highest beam current compatible with the resolution was, within very large limits, proportional to the total accelerating potential.

This property is well represented by Fig. 9, indicating the highest admissible beam current as a fnnction of the final accelerating voltage for a typical electron gun as used in a projection tube intended to be used for a resolution of 450 lines with a spot diameter on the order of 0.25 mm (0.01 inch).

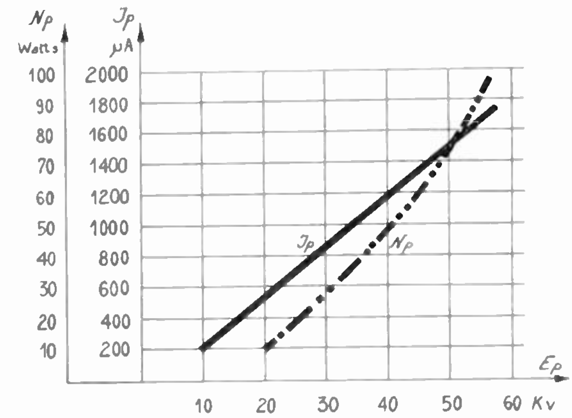

Fig. 9

By utilizing high accelerating voltages, it was necessary to put the last anode and the luminescent screen at ground potential in order to avoid excessive dielectric strain on the neck of the tube, and in order to increase considerably the spark-over distance between plate and cathode connections, as represented by Fig. 10.

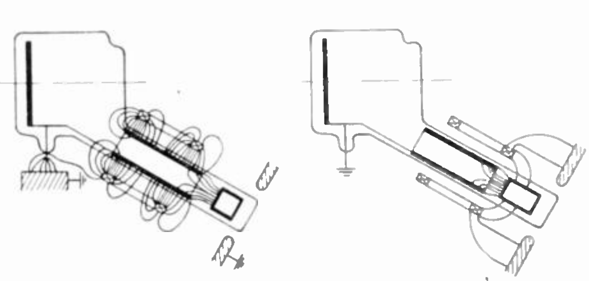

Fig. 10

The greater safety of the equipment seems to the author important enough to justify the somewhat unusual modulating apparatus as described below.

Fig. 11 is a picture of a typical projection tube. The principal characteristics are:

Admissible final accelerating voltage: 80 kv
Mean beam current (450 lines): 500 μamps
Necessary peak-to-peak modulation: 250 volts max.
Focusing and deflection: magnetic

The efficiency of the yellowish-colored luminescent screen, which can be cooled by water or by air, can attain the value of 4 candles per watt.

Research work is going on to obtain white-colored light and eventually higher luminous efficiency of the fluorescent screen.

Fig. 11

Keystone correction and vertical proportionality are obtained in a similar way to the case of iconoscopes, both tubes having substantially the same shape.

Owing to the fact that the length of the beam varies noticeably during the scanning of a complete field, the focus of the spot has to be automatically maintained during the whole sweep period. This is done by the superposition of conveniently shaped synchronous currents on the mean current flowing through the focusing coil. To obtain the greatest stability for this device, the necessary form of the focusing current is produced by the use of appropriated electric networks, excluding the curved parts of tube characteristics which are more subject to slow but inevitable variations.

The fluorescent screen is, for obvious optical reasons, plane. The resulting serious pin-cushion distortion of the scanned picture is compensated by conveniently formed constant magnetic field, generated by compensating coils arranged in a symmetrical manner in the immediate vicinity of the fluorescent screen.

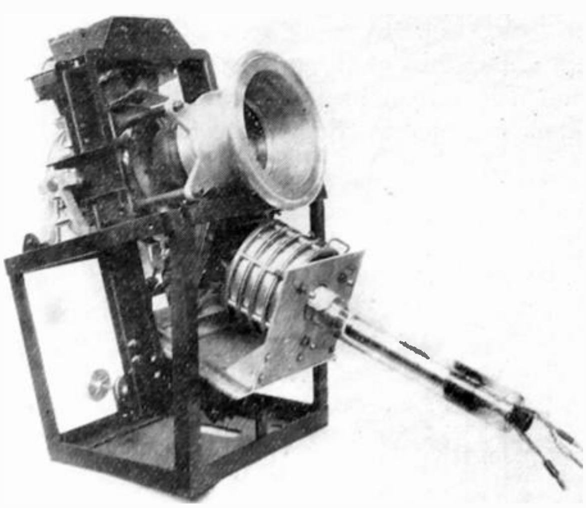

Fig. 12

Fig. 12 represents the physical aspect of the projector head, including the tube, the deflecting and concentration coils, the pincushion compensation, and the $f=1.9, f=200$ mm, coated projection lens.

Fig. 13 represents the cross section of the projector, showing from left to the right: the 50-cps power transformer for the video amplifier, the video amplifier, with the high-frequency coupling, the projector head, and the scanning power stages.

The screening necessary to absorb the X rays generated by the tube has been removed in order to make the head of the projector visible.

Fig. 13

Switching on, Control and Operation

The gradual switching on of the apparatus is automatically assured by cascaded relays. In case of failure of power, the apparatus is automatically switched off, and if desired, switched on again, after the return of the power. Voltage can not be applied to the projector tube if one of the scanning generators fails.

The controls for contrast, brightness, focus, picture-height and width, the eliptically swept control oscillographs for modulation depth and sync separation, and a 15-inch control viewing tube are included in the twin monitor desk, as represented by Figs. 14 and 15.

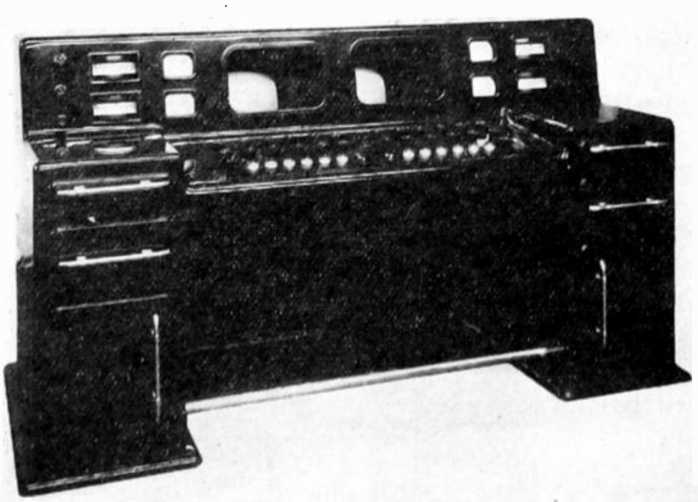

Fig. 14

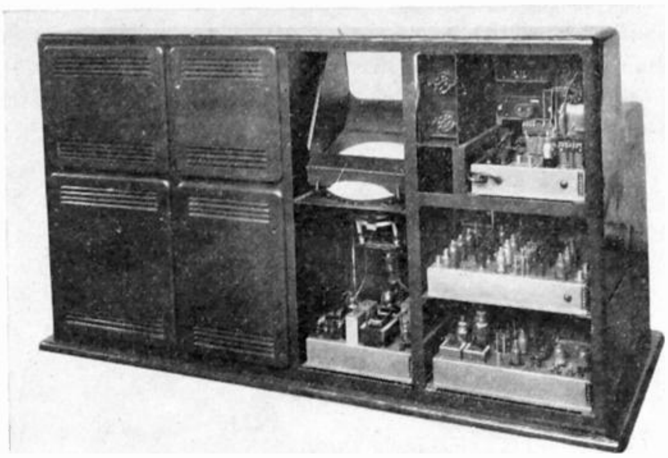

Fig. 15

The Projection Screen

Well-known economical conditions did not permit the use of the Schmidt type optical system in the place of the more conservative refractive lens system, which accepts only a few per cent of the total luminous flux issued by the luminescent screen. For this reason the use of a special projecting screen with a controlled directivity was decided upon during the development work.

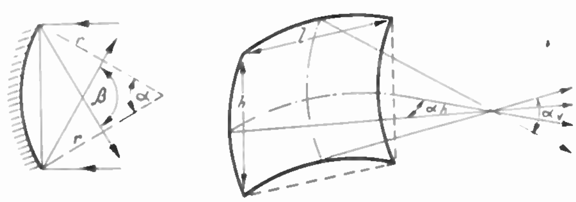

Fig. 16

The use of directional screens for photographic or projection work is by no means new. The realization of a directive screen having large dimensions, however, presents considerable difficulties. The inventiveness of our colleague, I. Saget, made it possible for us to undertake the construction of such a device.

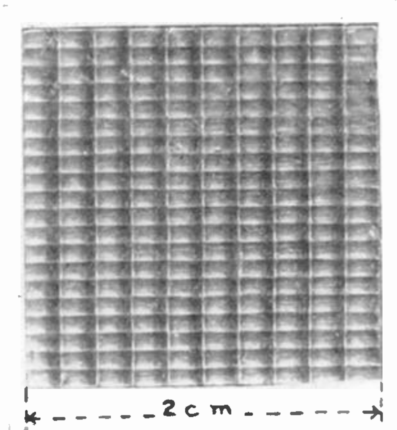

Fig. 17

The fundamental element of the screen is the spherical mirror. By the independent choice of the central angle in the horizontal and in the vertical plane, i.e., by the corresponding choice of r, l, and h, it is possible to control the aperture angles α_h and α_v by which a normally parallel incident beam is reflected. (Fig. 16).

Aperture angles were chosen for the screen intended to be used in medium-sized projection theatres, $\alpha_h = 70°$ and $\alpha_v = 36°$. The dimensions of the individual spherical mirrors, whose surfaces are considerably smaller than the size of one picture element are: $r = 3.15$ mm, $l = 2$ mm, $h = 1$ mm. The elementary screen consists of two hundred elementary mirrors stamped together in a highly polished aluminum sheet of 0.05 mm thickness

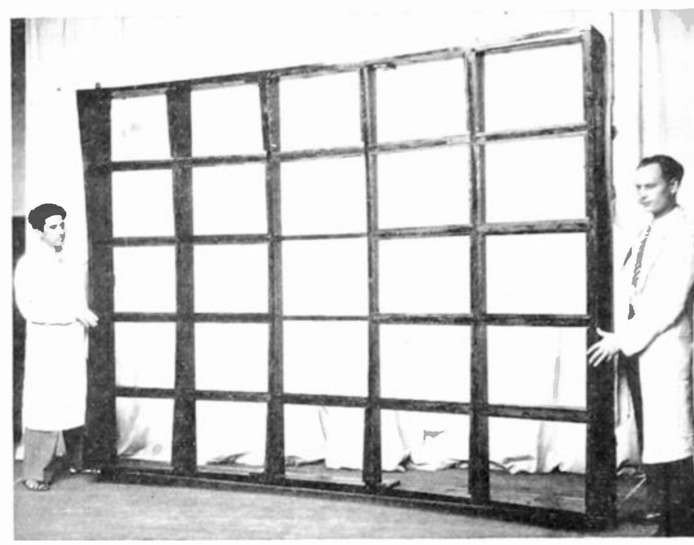

Fig. 18

(Fig. 17). About 15,000 of these elements are fixed in quincun on metallic supports held together by a wooden framework. (Fig. 18).

The dimensions of the projected picture are 3×2.25 meters. The screen itself has a radius of curvature of 9 meters in order to compensate for the divergence of

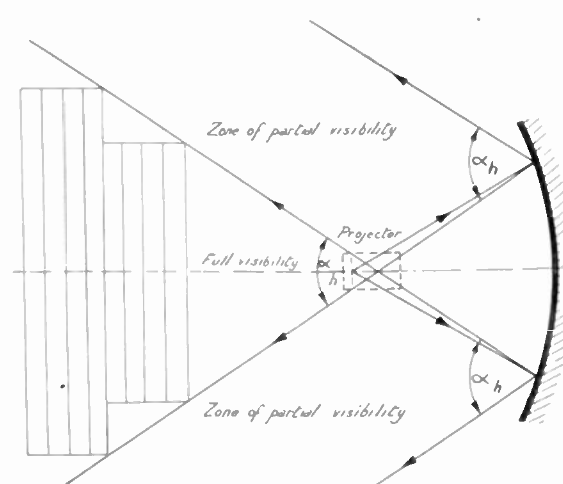

Fig. 19

the luminous beam issued by the projector, situated at a distance of 4.50 meters from the screen (Fig. 19). The measured gain of brilliance against a screen whose diffusion follows the cosine law is 4.25, in good agreement with the theoretical value.

In spite of the careful control of the different phases of the manufacturing process, the screen in its actual form is not wholly satisfactory. Owing to small irregularities resulting in minute local variations of the apparent screen brightness, the finest picture details are lost, giving the picture an undesirable granular aspect. Research work is being continued to improve this undesired effect.

ACKNOWLEDGMENTS

The author is deeply indebted to all his coworkers, and mainly to P. Thomas, of the Compagnie des Compteurs Laboratories, and to I. Saget, whose hearty collaboration made it possible to attain the results described in this publication.

An Analogue Computer for the Solution of Linear Simultaneous Equations*

ROBERT M. WALKER†, SENIOR MEMBER, IRE

Summary—**Linear simultaneous equations occur frequently in science and in engineering. Their solution by numerical methods is straightforward, but the amount of work required increases rapidly with the number of unknowns. A device is described for the solution of systems of linear simultaneous equations with not more than twelve unknowns. It is an electrical analogue computer which accepts the problem information in digital form from a set of punched cards. This facilitates the preparation, checking, and insertion of the input data and greatly reduces some of the usual liabilities of an analogue device. No special preparation of the problem is required, other than a simple one of scaling the coefficients. Solutions of well-determined problems are easily and rapidly attained and may be refined to any desired accuracy by a simple iteration procedure.**

I. STATEMENT OF THE PROBLEM

THE SIMPLEST case of linear simultaneous equations is that of two equations with two unknowns. For example

$$a_{11}x_1 + a_{12}x_2 + b_1 = 0$$
$$a_{21}x_1 + a_{22}x_2 + b_2 = 0. \tag{1a}$$

Such a case is very simply solved by any one of several computing methods. But the work required for solution increases rapidly with the number of the unknowns.

The general case with n unknowns may be represented by writing the i^{th} equation of the set of n equations as follows

$$\sum_{j=1}^{j=n} a_{ij}x_j + b_i = 0. \tag{1b}$$

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† Watson Scientific Computing Laboratory, New York, N. Y.

II. THE ELECTRICAL ANALOGUE

The device to be described solves a set of up to 12 equations (i.e., $n \leqq 12$) by means of an electrical circuit analogue. In setting up this analogue for 12 equations one requires 12 variables (x_j), 12 coefficients of each variable (making 144 of these a_{ij} coefficients), and the 12 constant terms, b_i.

The 12 variables are represented by 12 individually adjustable voltages, the magnitude and sign of each being controlled by the operator. The coefficients are three-digit decimal type voltage dividers which can be set for an output of from -0.999 to $+0.999$ times the input. (The method of obtaining the negative sign for a coefficient will be described later.) The constant terms (b_i) are obtained by feeding 12 more of these same decimal type voltage dividers from an additional (13th) voltage source. Addition of the terms in an equation is obtained by adding the voltage outputs of these dividers in a resistive network. A block diagram is shown in Fig. 1.

Since the values of the a_{ij}'s and the b_i's must all lie in the range from -0.999 to $+0.999$ for insertion into the machine, in general it will be necessary to prepare the problem by multiplying each equation through by a suitable constant so that the largest a_{ij} or the b_i in that equation has an absolute value of less than unity. Other types of transformations may be used if desired.

It is also necessary to make another transformation so that a finite range of voltage variation can represent an unlimited range of the unknowns. The transformed unknowns are designated as X_j and are related to x_j by a factor such that

$$x_j = \frac{X_j}{k}, \quad \text{where} \quad k \leqq 1,\ X_j^2 \leqq 1. \tag{2}$$

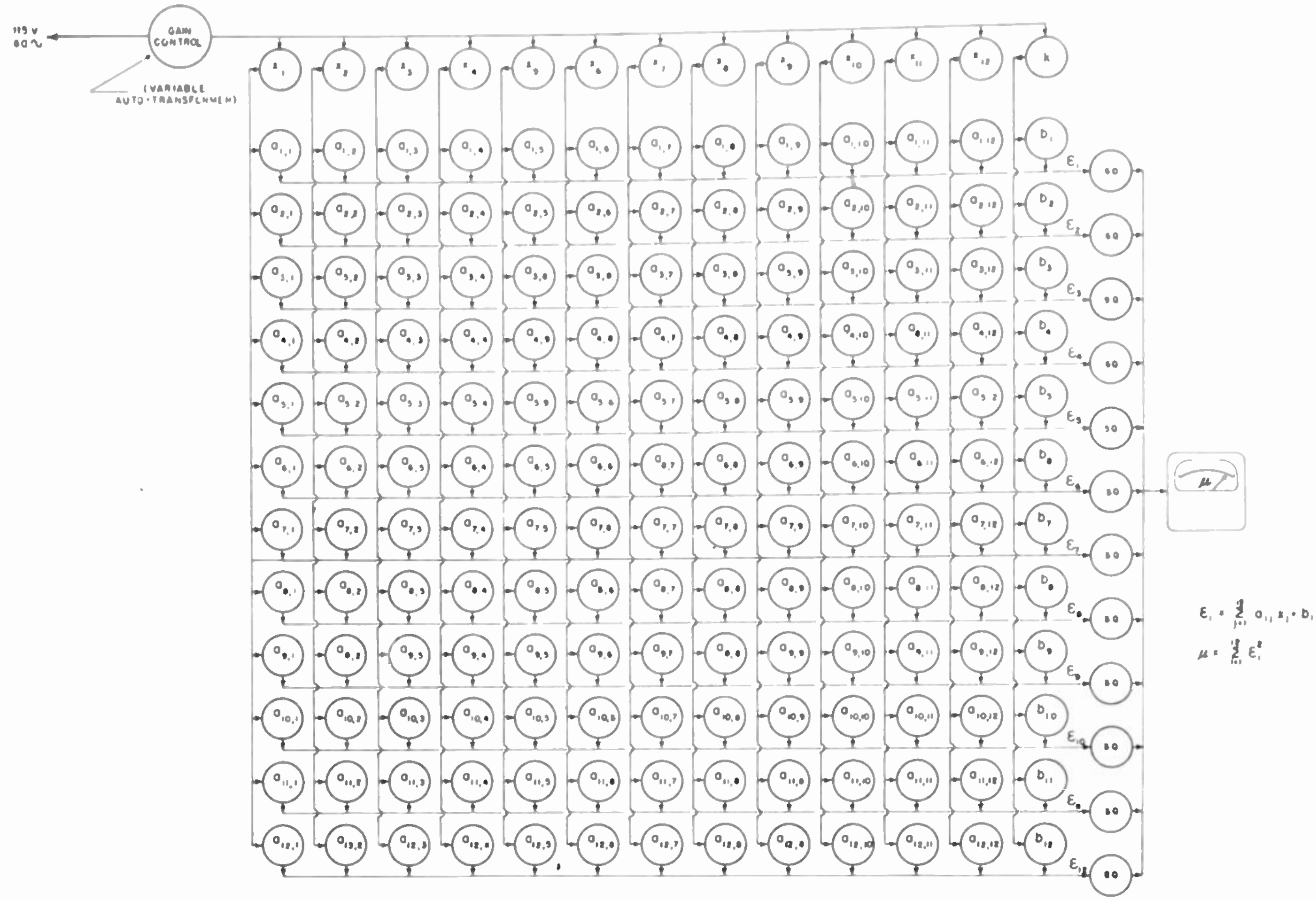

Fig. 1—Linear simultaneous equation solver.

Substituting this in the original set of equations gives

$$\sum_{j=1}^{j=n} a_{ij}X_j + b_i k = 0. \qquad (3)$$

It will be seen that k corresponds to the voltage source for the constant terms and that by making it adjustable it becomes a controllable scale factor for the unknowns. This scale factor may be changed as required during the progress of the solution.

Having thus set up the electrical analogue of a system of equations, when one sets in the proper voltages for the X_j's and the k corresponding to the correct solution of the problem, the voltage output will be zero for each of the equations. But until these correct values are obtained an error voltage does exist, in general, which for the i^{th} equation we designate as ϵ_i. The operator is furnished with information (in the form of a single meter reading) so that he can make adjustments leading to the reduction of all ϵ_i's toward zero. This process is similar to that of balancing a bridge.

III. The Analogue of X_j and k

Thirteen identical transmitter units feed out the 12 X_j voltages and the k voltage. These are schematically shown in Fig. 2. This unit takes the 60 cps voltage from the common supply bus, and provides a low-impedance balanced-output voltage of variable amplitude and sign. The voltage of the supply bus is adjustable from 0 to 120 volts by means of a variable autotransformer, to control sensitivity during the adjustment process. The reason for the division of the output

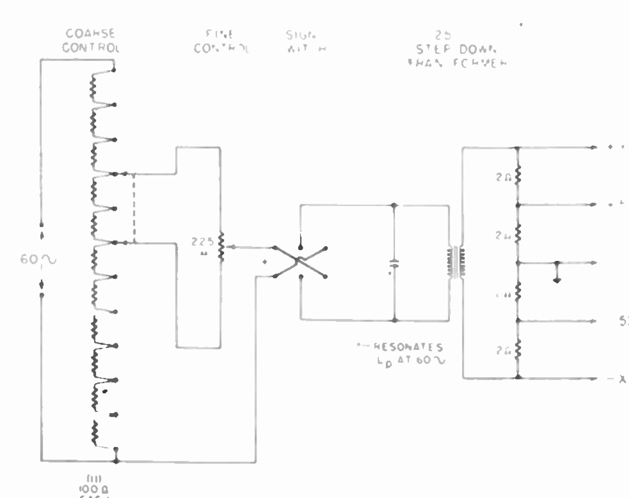

Fig. 2—Transmitter unit for 1 variable.

into four equal parts balanced with respect to ground will appear in connection with the coefficients. These four parts are adjusted to equality within better than 0.05 per cent.

For several reasons, it seems desirable to use ac for the voltage sources of this machine. First, because of the ease of obtaining a large amplification of the equation errors; second, because of the possibility of using transformers to go from unbalanced attenuators to balanced loads; third, because the commercial 60-cps power is economically available. There are consequent disadvantages to the use of ac. Care must be taken to keep phase shifts within narrow limits and, even with such control, special means are required in the indicating device to reject any quadrature components so that they do not obscure the balance indication.

TABLE I

E_1	E_2	$E_1+0.2E_2$
$-\ X$	$+0.5X$	$-0.9X$
$-\ X$	$+\ X$	$-0.8X$
$-0.5X$	$-\ X$	$-0.7X$
$-0.5X$	$-0.5X$	$-0.6X$
$-0.5X$	0	$-0.5X$
$-0.5X$	$+0.5X$	$-0.4X$
$-0.5X$	$+\ X$	$-0.3X$
0	$-\ X$	$-0.2X$
0	$-0.5X$	$-0.1X$
0	0	0
0	$+0.5X$	$+0.1X$
0	$+\ X$	$+0.2X$
$+0.5X$	$-\ X$	$+0.3X$
$+0.5X$	$-0.5X$	$+0.4X$
$+0.5X$	0	$+0.5X$
$+0.5X$	$+0.5X$	$+0.6X$
$+0.5X$	$+\ X$	$+0.7X$
$+\ X$	$-\ X$	$+0.8X$
$+\ X$	$-0.5X$	$+0.9X$

IV. The Analogue of a_{ij} and b_i

The 156 coefficient networks (144 a_{ij}'s and 12 b_i's) are all identical. Each is schematical as shown in Fig. 3. The horizontal busses are the output from one of the transmitter units. The three-digit nature of the device is evident from the formula shown.

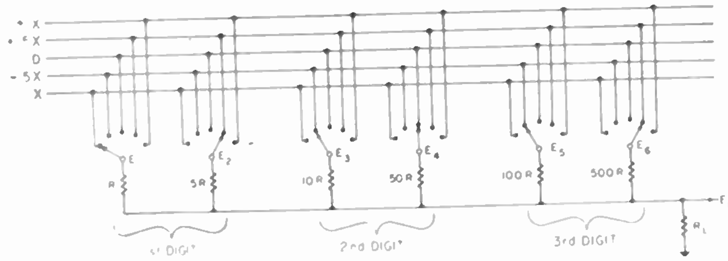

Fig. 3—Coefficient network (1 of 156)

$$E_L = \frac{(E_1 + 0.2E_2) + 0.1(E_3 + 0.2E_4) + 0.01(E_5 + 0.2E_6)}{K}$$

where

$$K = 1.332 + \frac{R}{R_L}.$$

Since the voltages E_1 to E_6 can each have any one of 5 discrete values, it would be possible to have 5^6 different switch combinations. However, for the first decimal digit only, the 19 combinations of E_1 and E_2 are used which give from $-0.9X$ to $+0.9X$. These are listed in Table I.

The same combinations are used for the second decimal digit (E_3 and E_4) and for the third digit (E_5 and E_6), but with the additional restriction that the second and third digits always have the same sign as the first digit. Thus only 1,998 combinations are actually used to represent values from $-0.999X$ to $+0.999X$.

The accuracy of the coefficient voltage dividers is better than ± 0.1 per cent. The lowest value resistor R of each coefficient is specified to ± 0.05 per cent, the $5R$ value to ± 0.25 per cent, the $10R$ to ± 0.5 per cent, the $50R$ to ± 2.5 per cent, the $100R$ to ± 5 per cent, and the $500R$ to ± 20 per cent. The three lowest values of resistors are precision wire-wound, and the three higher are a composition type.

The switching is accomplished by making contacts through holes in a punched card, in a card-controlled switching device. Fig. 4 shows the card in place preparatory to closing the reader. The spring-driven pins in the upper plate make contact through the holes punched in the card to switch the coefficient resistors to the appropriate voltage busses. The coding representing a digit (-9 to $+9$) is punched in a single column of the card, one punch in one of the upper five positions and another in one of the lower five. Successive columns represent the successive digits of a coefficient, and 36 columns are available on each card to accept the 12 three-digit coefficients of each variable. Thirteen cards are used in all, twelve for the coefficients of the twelve unknowns, and a thirteenth for the constant terms.

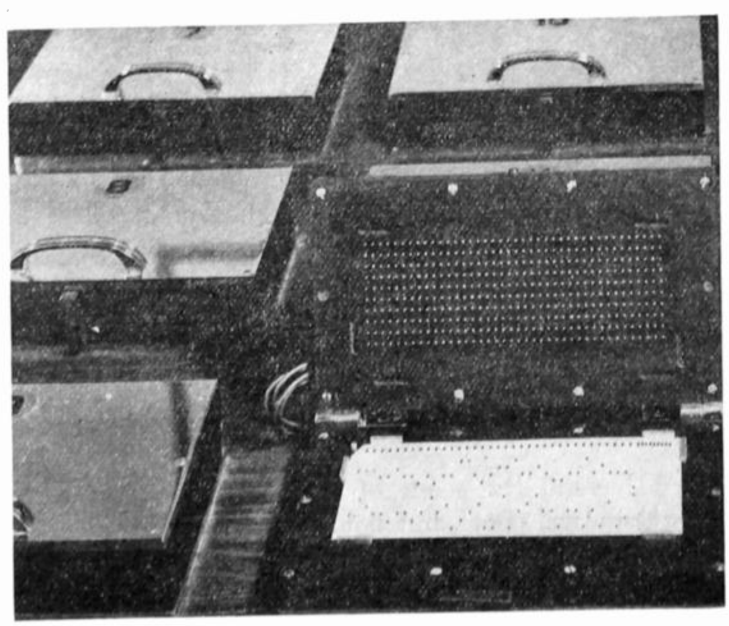

Fig. 4—Card reader open with coefficient card inserted.

Fig. 5 shows the appearance of one of the coefficient cards. For purposes of illustration, the punching has

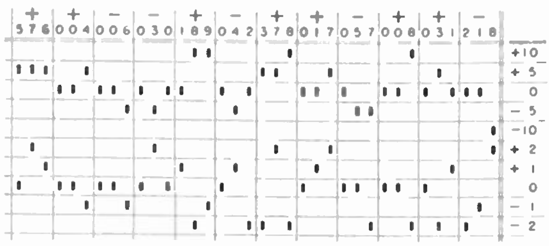

Fig. 5—Sample coefficient card with interpretation of punching.

been interpreted at the side and the top. This double-punch coding is done by a special keyboard attachment to the standard punch. The operator has only to press a digit key and the proper two holes are punched in that column.

The use of punched cards for the insertion of problem data into the machine has several important advantages over the use of switches or potentiometers. It allows the preparation of a new problem while the machine is in use for a previous problem, thorough checking of the data (by duplicate punching and comparison), easy permutation of variables as a check procedure, and the filing of a problem for subsequent reinsertion (as for iteration).

V. The Summation of Terms

The summing of voltage outputs from the appropriate coefficient networks to form the first equation is accomplished by connecting the outputs of all 13 coefficients associated with this equation ($a_{1,1}$, $a_{1,2}$, $\cdots$, $a_{1,12}$, b_1) to a common output bus. This means that the load resistor R_L for the coefficient network shown in Fig. 3 is the impedance seen looking back into 12 other identical networks in parallel. Therefore, $R/R_L = 12 \times 1.332$ and $K = 13 \times 1.332$.

The other 11 equations are summed in the same manner.

VI. Method of Solution

In order that that operator can adjust the variables to satisfy the conditions of the problem and thus obtain a null of voltage for each equation, he must be given an indication which will tell him in what direction to proceed. When the conditions of the problem are not satisfied, there will be a residual error which for the i^{th} equation is

$$\epsilon_i = \sum_{j=1}^{j=n} a_{ij}X_j + b_i k \tag{4}$$

It is beyond the scope of this paper to discuss the mathematical proof, but it has been shown that if one sets up the quantity

$$\mu = \sum_{i=1}^{i=n} \epsilon_i^2 \tag{5}$$

and adjusts the variables in a cyclic manner so as to reduce continuously the quantity μ, that μ will converge toward zero, which, of course, only occurs when all the ϵ_i's are zero.[1]

To carry this process out for this machine each of the ϵ_i's is electrically squared and the outputs added. This is not as simple as it might seem since the electrical voltages representing the X_j's are really complex, that is, they are not, in general, exactly in phase with the voltage representing k, but may have relative phase angles of the order $\pm 1°$. Considering the k transmitter as the reference phase, the situation is as follows:

The output voltage of the i^{th} equation is

$$\sum_{j=1}^{j=n} a_{ij}X_j \cos(\omega t + \theta_j) + b_i k \cos \omega t, \tag{6a}$$

or

$$\left(\sum_{j=1}^{j=n} a_{ij}X_j \cos\theta_j + b_i k\right)\cos\omega t - \left(\sum_{j=1}^{j=n} a_{ij}X_j \sin\theta_j\right)\sin\omega t. \tag{6b}$$

For values of $|\theta_j| \leqq 1°$, $X_j \cos\theta_j \cong X_j$, and

$$\epsilon_i = \sum_{j=1}^{j=n} a_{ij}X_j + b_i k \cong \sum_{j=1}^{j=n} a_{ij}X_j \cos\theta_j + b_i k. \tag{6c}$$

This equation error voltage is seen to contain a cosine term whose coefficient (or amplitude) is approximately proportional to ϵ_i, and also an undesired sine term. The amplitude of the cosine voltage term can be obtained by periodically sampling the output voltage at the times when $\cos \omega t$ is ± 1 and $\sin \omega t$ is zero. In practice, it is not necessary to remove completely the quadrature component (sine term) and what is done is to observe the output voltage for a short period each half-cycle when the cosine is large and the sine small; i.e., at the crests of the reference wave.

For the final balancing of a set of equations, it is necessary to be able to observe the presence of ϵ_i's as small as 0.001 which for this machine will mean voltages of

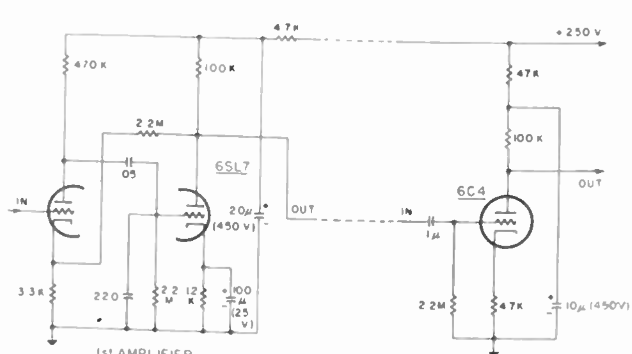

Fig. 6—Presquaring amplifier (1 of 12 channels).

[1] See reference 2 of the bibliography.

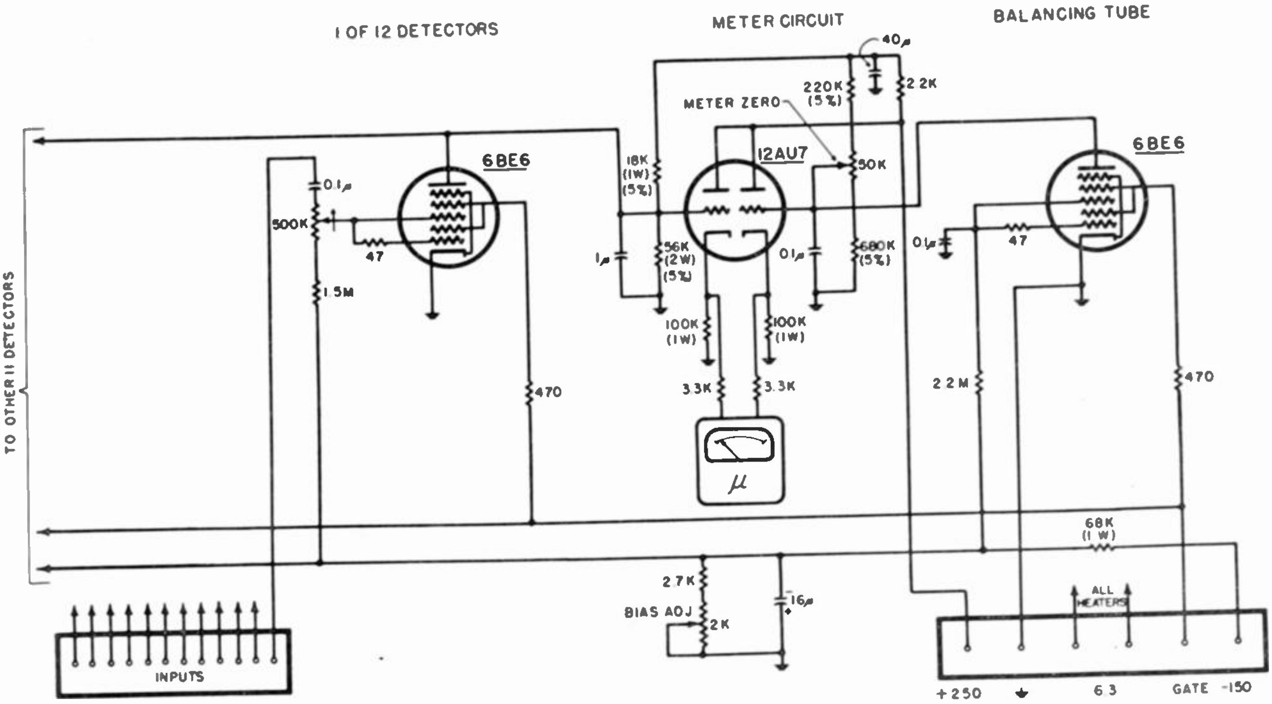

Fig. 7—Square-law detector and indicator circuit.

about 0.1 millivolt. Therefore, the output of each equation bus is fed into a 3-stage amplifier having a voltage gain of 3,000. One of these amplifier channels is shown schematically in Fig. 6. The output of each one of these amplifiers is then fed into a square-law detector. These detectors are pentagrid converters (6BE6) with the input applied to both the first and third grids to increase the square term. The tube is biased to virtual cutoff. Screen-grid gating is applied so that the tube is turned on only in the intervals when the cosine term of the input signal is large and the sine term is small. The dc increment of plate current is approximately proportional to the square of the input voltage, and by allowing all the detector plate currents to flow through a common load resistor, the drop in this resistor is a measure of the sum of the squares of all the ϵ_i's. A differential vacuum-tube voltmeter is used to display this quantity, which is proportional to μ. Fig. 7 is a schematic of the detector and indicator unit.

The screen-grid gating wave form for the 6BE6's is

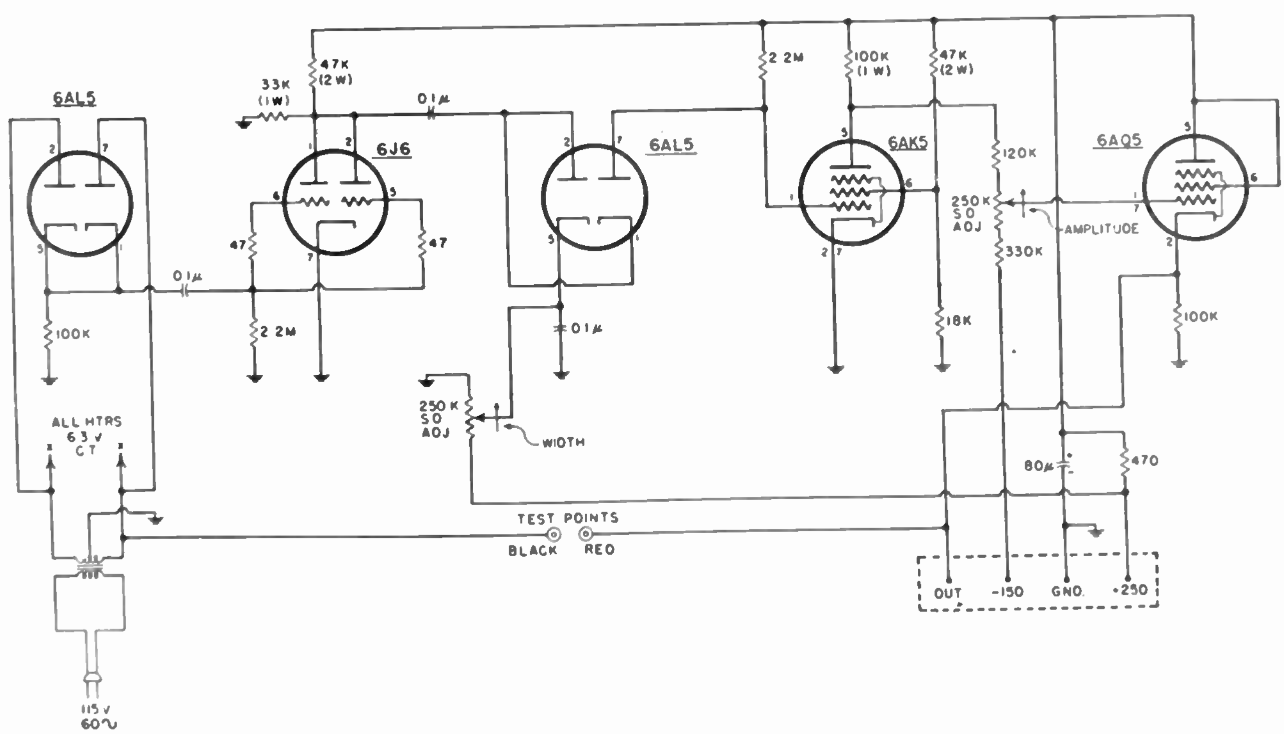

Fig. 8—120-cps gate generator unit.

obtained by full-wave rectification of the 60 cps, followed by clipping and squaring. The circuit of this gating unit is shown in Fig. 8. The output, which is applied to the screens of the 6BE6 detectors, is a 90-volt rectangular-shaped pulse occurring 120 times per second (symmetric about the positive and negative crests of the 60-cps sine wave). The time duration of each pulse is about 1.5 milliseconds.

VII. Read-Out Method

When all the variables have been adjusted to satisfy the conditions of the problem (as indicated by a null in the quantity μ) the values of the X_j and k voltages are successively read out by means of comparison with the output of a four-dial decade potentiometer fed from a reference voltage. In practice, these readings are made only to three digits, the fourth digit being used if necessary to determine the direction for rounding of the third digit. The comparison is made by means of a differential amplifier followed by a phase-detector, as shown in Fig. 9. The phase-detector is used so that the sign of the difference is indicated by the galvanometer, thus facilitating finding the balance point. Since the final x_j's are obtained as ratios of the X_j's to k, it follows that no absolute measurement is required and that the accuracy of the read-out is determined by the accuracy of the ratics of the decade potentiometer. This potentiometer has an accuracy of 0.05 per cent of full scale.

Fig. 10 is a view of the operating console, showing the power switches and gain control at the left, the μ meter at top center, the thirteen controls for the X_j's and k, and the read-out potentiometer and galvanometer on the right. The read-out selector is at the left of the galvanometer.

VIII. Operation

When the numerical data of a problem has been scaled (if necessary) to fit the range of the machine, it is punched into a set of 13 cards. If there are less than 12 unknowns, zeros are inserted for all the missing elements. These cards are placed in the card-reading units, and the machine is then ready for solution. The usual process is to start with $k=1$, and all the X_j's at zero. The common input voltage to all the X_j's and k is controllable by the large dial at the left of the console so that the large errors during the initial stages of solution do not overload the subsequent amplifiers and indicator.

The operator varies the X_j's in such a manner as to decrease the error and as necessary he increases the input voltage to increase effectively the scale sensitivity of the error indicator. A systematic cycling process is used which results in the error decreasing toward zero and the X_j's converging toward the solution. When the residual error is small enough to be consistent with the accuracy of the analogue, the results are read out in the manner previously described. The x_j's are then obtained by calculating the ratios X_j/k.

A numerical check can be obtained by substituting these values into the problem and calculating the actual residual errors. If greater accuracy of results is desired

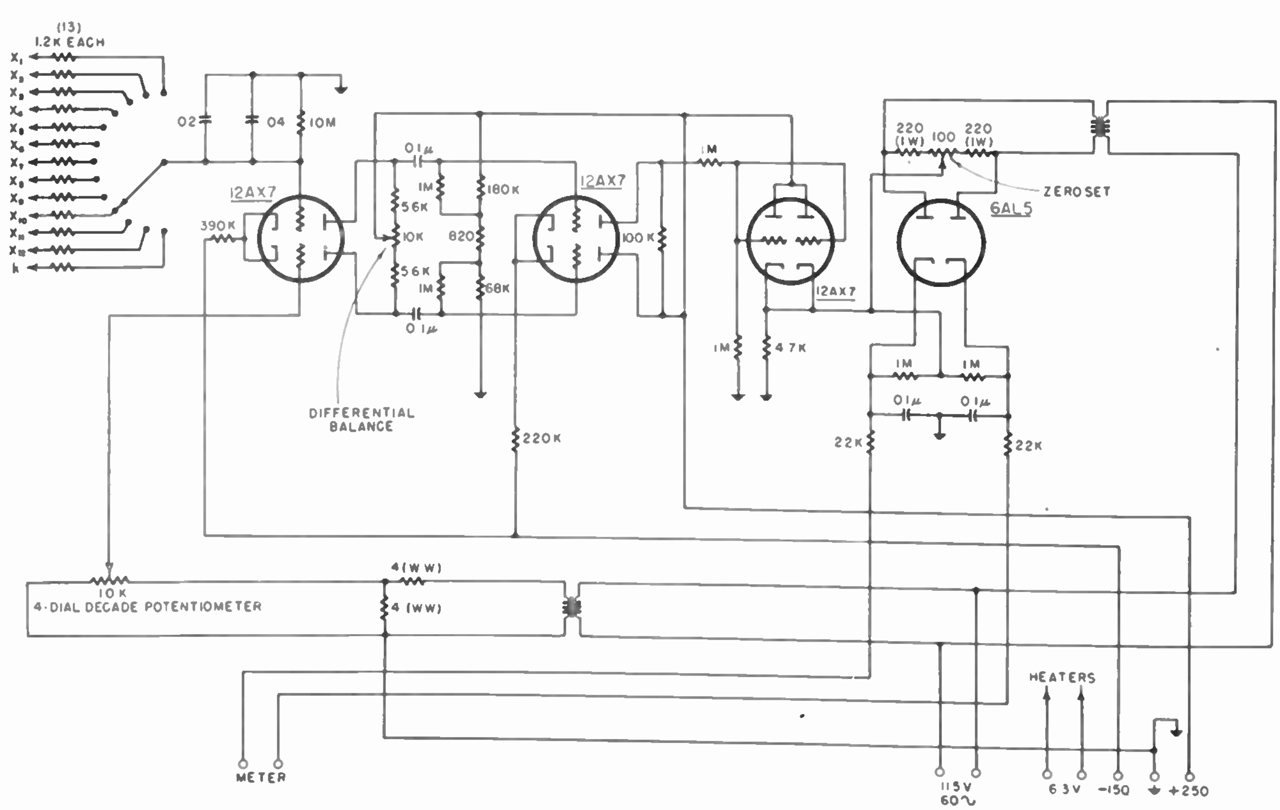

Fig. 9—Read-out potentiometer and galvanometer detector.

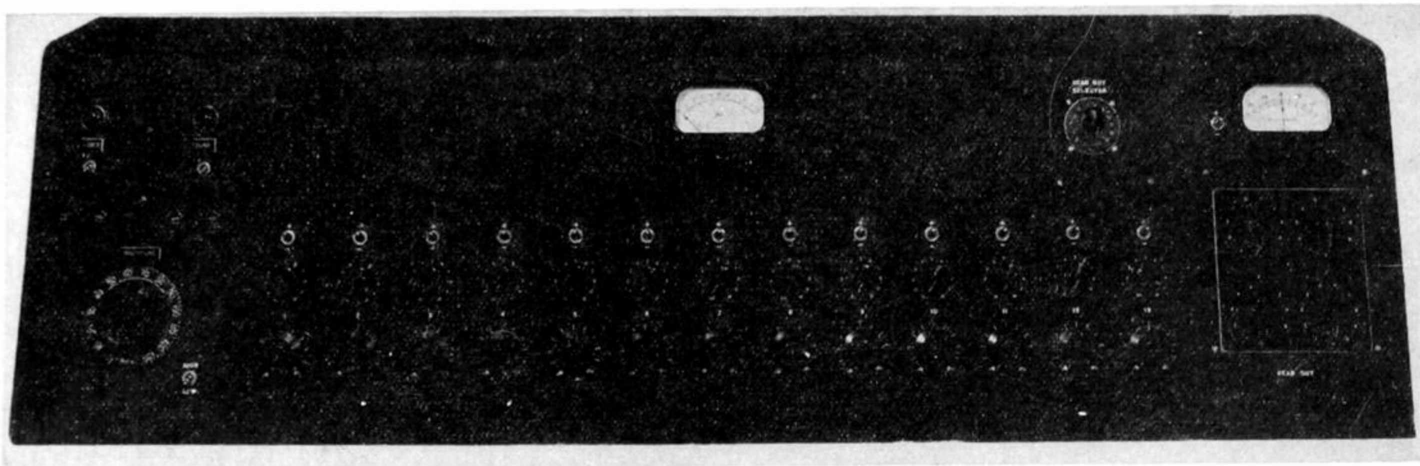

Fig. 10—The operating console.

than is obtainable from the initial solution, these residual errors may be used to form a new set of equations in which the new unknowns are the errors in the x_j's of the initial solution. The iteration procedure is simple and rapid, requiring the preparation of only one new punched card per step. As many iteration steps as required may be used to obtain the desired accuracy.

The usual type of problem with 12 equations requires about 15 minutes for solution and yields solutions of better than two-digit accuracy. Each iteration will then yield somewhat better than two additional digits.

IX. Capabilities

For most problems of linear simultaneous equations, this device affords a considerable saving of labor over the usual methods of computing and has the advantage when iteration is used of correcting any errors made by the machine operator. Errors in one step are corrected in the next so that instead of being cumulative they are successively diminished. Of course, the machine is of a particular utility when the accuracy required can be obtained with a single solution.

There are certain other advantages of an analogue device for this purpose. For example, it allows investigation of the effect of varying one or more of the coefficients, without requiring a complete new solution.

The machine has been used for inversion of matrices up to the twelfth order. An n^{th} order matrix requires n separate solutions to obtain the inverse.

Sets of up to 6 equations with complex coefficients may be handled separating the real and imaginary parts, thus doubling the number of equations and the number of unknowns.

X. Acknowledgment

The author wishes to acknowledge the invaluable assistance of Prof. F. J. Murray, of Columbia University, who is also responsible for the fundamental theory of this type of machine.

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An Electronic DC to AC Converter for Use in Servo Systems*

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Summary—The usefulness of alternating current servomechanisms has been restricted in many cases by the difficulty of obtaining a satisfactory ac error signal.

The conversion from dc signals to an ac error signal, i.e., the modulation system, is shown to be a predominant source of difficulty.

A modulation circuit having characteristics, much improved over conventional types, has been developed. Practical application to a servomultiplier with results observed is discussed.

I. Introduction

This paper describes a differential dc to ac converter or modulator for use in a servo system where the prime mover is an ac electric motor whose direction of motion is reversed by reversing the phase of the voltage applied to one of the windings, and whose torque is a function of the amplitude of the applied voltage. A linear relationship between voltage and torque for a given shaft rpm is usually desired. In addition, it is essential that the torque rpm characteristic be a decreasing monotone function over the full speed range

Almost invariably, some type of modulator is involved in the operation of an ac servo. In some systems this may be a mechanical modulator, (e.g., selsyn, vibrator, "chopper"); in others an electronic modulator is used.

II. Limitations of Electronic Modulators

Where dc signals are used, the modulator is usually the "weak link" in an ac servo system. Electronic modulators of usual design have very low efficiency because operation is dependent upon second-order variations in tube characteristics. The circuit configuration of a typical 400-cycle differential modulator is shown in Fig. 1.

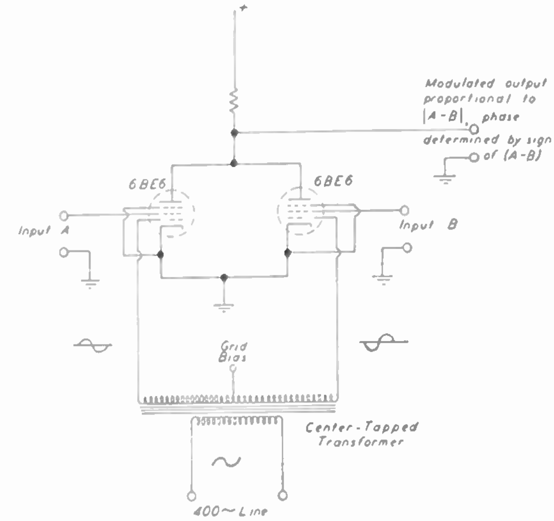

Fig. 1—G_m variation modulator.

This particular circuit produces about 28 millivolts peak-to-peak ac per volt dc differential input. This represents a voltage conversion efficiency of 2.8 per cent. This figure is typical of such modulators. Application of the signals to the control grids, to increase the conversion efficiency, restricts the usable input voltage range.

In addition, such a modulator is also unsatisfactory for many applications because the balance condition is not maintained for different dc levels with equal dc input signals. These effects may be overcome in some cases by very careful tube selection. In most cases, however, it is necessary to provide a dc differential error signal at a fixed level. The development of a dc error signal usually requires additional vacuum tubes, with attendant inaccuracies.

The presence of a second harmonic component of the order of 1 volt rms in comparison to a 10-millivolt carrier frequency signal for 1 volt differential input signal necessitates a high attenuation filter for removing this component, whose continuous presence would greatly overload amplifiers and motors.

Because of the low modulation efficiency, large gains are required in order to obtain sufficient voltage to operate a motor, and random and induced noises become a problem.

It is advisable for this reason to restrict the gain of the system for low frequency noise components. The use of a band-pass filter is indicated to remove simultaneously the second harmonic component. This filter should immediately follow the modulator, since noise may produce modulation in the amplifier due to nonlinearity.

III. A Square-Wave Modulator for AC Servomechanisms

A modulator has been developed which has many advantages over the type illustrated by Fig. 1, and appears to be superior in several characteristics to any which has previously come to our attention.

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In particular, the modulator to be described is almost completely free from zero drifts, has high modulation efficiency, and is extremely linear in operation. The range of operation may be made very large.

The principle of operation of the modulator is indicated by analysis of Fig. 2. Square-wave voltages of opposite phase and approximately equal amplitude are applied between points 1 and 2 and ground. Input signals, assumed positive with respect to ground, from low impedance sources are applied at signal points A and B. Slight circuit modifications permit the use of negative signals.

The maximum potential reached by point C is the value of the signal at A; similarly, the maximum voltage at D is the value of the signal at B. V_1 and V_4 prevent negative excursions of points C and D. Thus at point C is found a square wave whose amplitude is equal to the signal at A. At D is produced a square wave whose amplitude is equal to the signal at B, and of opposite phase to the signal at point C. Let E_x be the voltage to ground

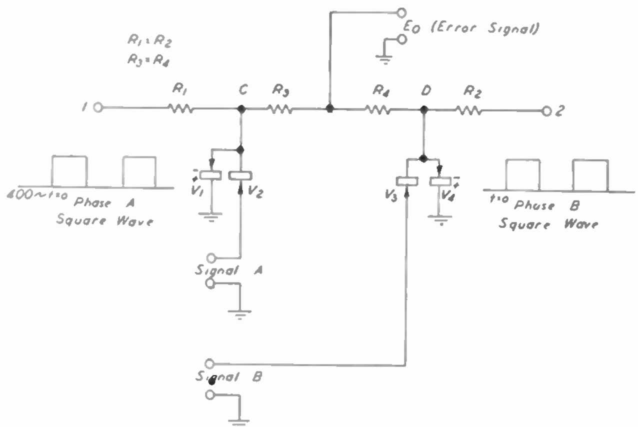


Fig. 2—Square-wave modulator.

at any point x. Since $R_3 = R_4$, the output signal E_0 is equal to $\frac{1}{2}(E_C + E_D)$.

Circuit operation is clarified by considering four cases:

Case I. $E_A = E_B = 0$ (A and B at ground potential)

The voltage at $C = 0$. The voltage at $D = 0$. $E_0 = 0$.

Case II. $E_A = E_B = E_{AB}$.

E_C and E_D are square waves of maximum amplitude E_{AB}, $E_0 = E_{AB}/2$ but there is no ac component. (This is shown in Fig. 3(a)).

Case III. $E_a > E_b$.

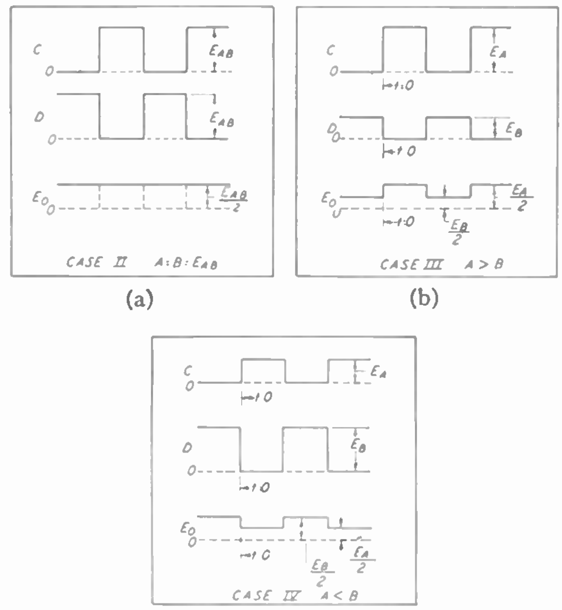
The square waves at C and D are unequal, and the output at E_0 is $E_b/2$ plus a square wave whose amplitude is $(E_a - E_b)/2$ of the same phase as phase 1. (See Fig. 3(b)).

Case IV. $E_a < E_b$.

This is similar to Case III, except that the output square wave is now of opposite phase. (Fig. 3(c).)

It is seen that the ac component of the output in every case is a square wave whose phase is determined by the larger signal, and whose amplitude is half the difference between the signals.

The output wave may be represented by the Fourier series, $E = E_{dc} + E_1 \sin \omega t + E_3 \sin 3\omega t + \cdots$, where E_{dc} is the dc component, E_1 is the amplitude of the funda-



(a) (b) (c)

Fig. 3—Theoretical wave forms in a square-wave modulator.

mental component and $\omega = 2\pi f$. The coefficient E_1 may be evaluated by standard methods[1] giving $E_1 = (E_a - E_b)/\pi$. Thus the peak-to-peak amplitude of the fundamental component of the output voltage (equal to $2E_1$) is 63 per cent of the dc difference between the two input signals. Performance is an order of magnitude better than that of g_m variation modulators of equal linear range.

If V_1, V_2, V_3, and V_4 (Fig. 2) be assumed perfectly conducting, and the impedance of signal sources A and B be assumed zero, the ac output of the device is entirely independent of the amplitude of E_1 and E_2, so long as neither signal voltage is larger than the amplitude of the respective square wave.

Since these conditions may not be met exactly, the balance condition is dependent upon the amplitude of E_1 and E_2. However, if R_1 and R_2 are kept large in comparison with the conduction resistance of V_2 and V_3 plus the respective source impedances of A and B, the unbalance error will be very small.

When A and B are of low amplitude, unequal resistance in the diode conduction paths may cause significant errors, particularly where the source impedances are high. In such cases the use of cathode followers for driving the diodes may be desirable.[2] If both square waves

[1] L. Mautner, "Mathematics for Radio Engineers," p. 298, Pitman Publishing Co., New York, N. Y., 1947.

[2] B. Chance, "Some precision circuit techniques used in waveform generation and time measurement," *Rev. Sci. Inst.*, vol. 17, pp. 396–416; October, 1946.

are not identical in wave shape, summing them produces pulses (at the changeover points) having fundamental frequency components in quadrature with the fundamental components of E_1 and E_2. This usually does not decrease the accuracy of the system, since components in quadrature with the normal driving signal produce no torque in most servomotors. The heating produced by these components may, however, limit maximum usable sensitivity. Square waves having appreciable rise and fall times produce pulses which occur at twice line frequency. These cause little difficulty, since they may be reduced to a tolerable level with a simple filter.

IV. Description of a Servomultiplier Using the Square-Wave Modulator

Figs. 4 and 5 show the application of the square-wave modulator to a servomultiplier used in a computer.

Fig. 4 is a simplified equivalent circuit of the multiplier. α is the percentage rotation of a balance potentiometer and E_{dc} the voltage across this potentiometer. It is seen that $E_1(t)$ is balanced by αE_{dc} produced by the

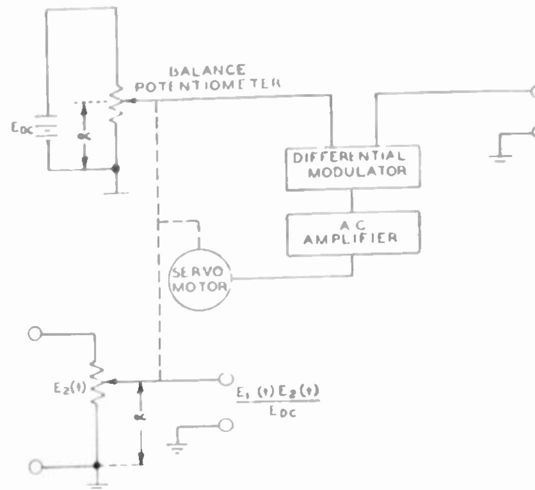


Fig. 4—Block diagram of servomultiplier.

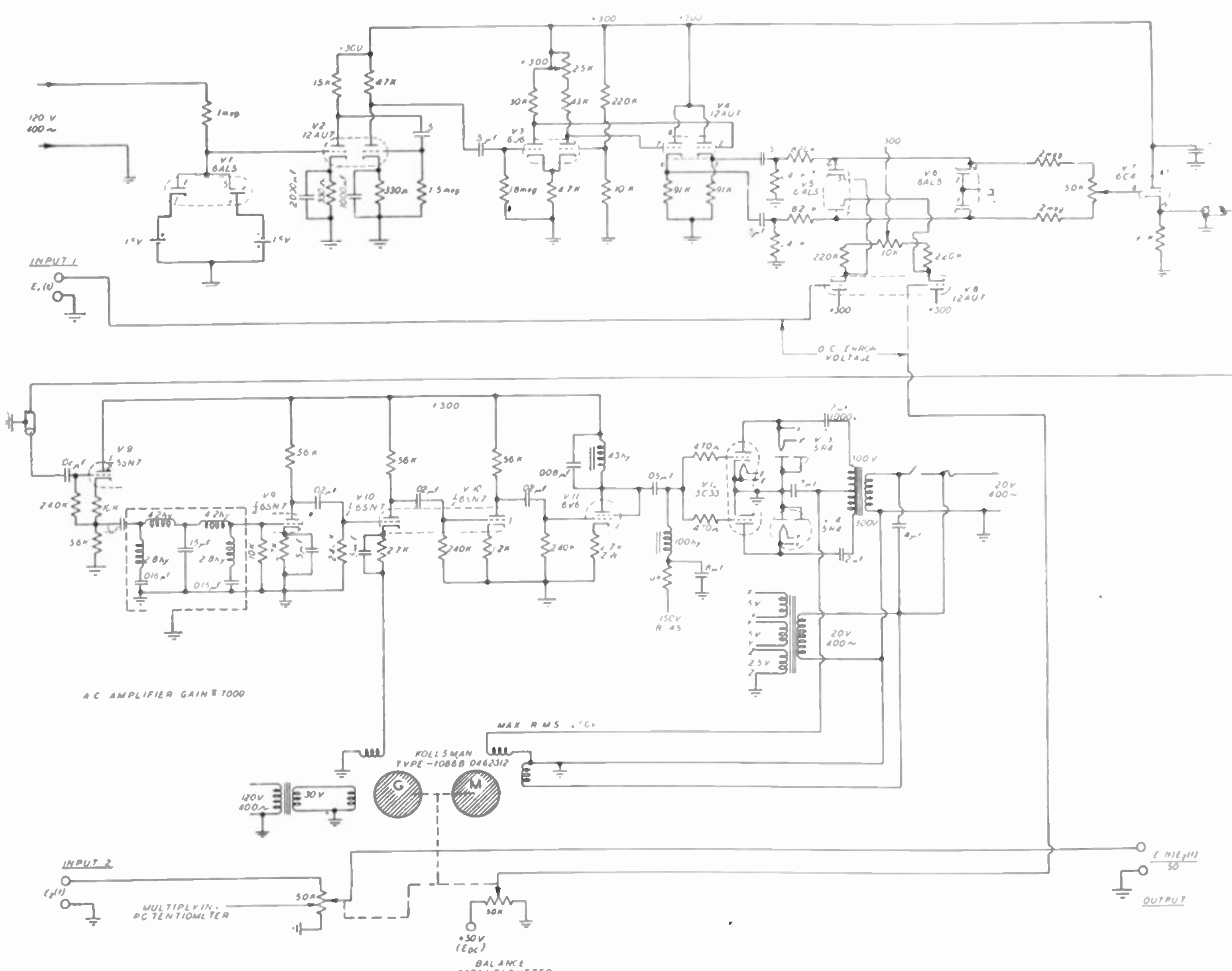


Fig. 5—Complete servomultiplier schematic.

rotation of the potentiometer. Balance is achieved by applying the amplified ac error signal from the differential modulator to the driving motor. In practice, the servo is able to position the potentiometer with sufficient accuracy to more than exhaust the accuracy of the potentiometer (0.1 per cent). Thus, very nearly, $\alpha = [E_1(t)/E_{DC}]$.

$E_2(t)$ is applied to a multiplying potentiometer, ganged with the balance potentiometer. Thus the output from arm to ground is $\alpha E_2(t)$ or $[E_1(t) \times E_2(t)/E_{DC}]$ The magnitude of E_{dc} becomes a scale factor.

The complete circuit exclusive of power supplies is shown in Fig. 5. This particular servo is shown because it was initially provided with a conventional modultaor similar to Fig. 1, and therefore provides an excellent opportunity for comparison between the two modulators. After tests had been made on the original system using the conventional modulator, the system was changed to use the square wave modulator.

Tubes V_1 through V_4 provide square waves for the modulation circuit. Square waves of approximately 70 volts are applied to the modulation circuits from the cathode followers V_4. V_5 and V_6 comprise the square-wave modulator. Comparison with Fig. 2 will indicate the method of operation. V_7 provides a low impedance output for the square-wave error signal. V_8 provides low impedance driving signals for the modulating diodes.

Section 1 of V_9 provides the desired driving impedance for the low pass filter. (V_7 could have been eliminated, but since the modulator section was on another chassis, its use simplified cabling.) A filter as elaborate as that shown is probably not necessary, but was originally incorporated for use with the g_m variation modulator. Section 2 of V_9, tubes V_{10}, and V_{11} comprise an ac amplifier having a gain of approximately 7,000. V_{12}, V_{13}, and V_{14} constitute a more or less conventional discriminator. The magnitude and phase of the voltages at the 3C33 tube grids determine the magnitude and phase of the current through the driven winding of the servomotor. The line phase of the servomotor is adjusted by means of a series capacitor so that the two windings are in phase quadrature. Damping is effected by introducing, in the cathode circuit of one section of V_{10}, the output of a derivative generator geared to the servomotor.

Performance data of the relative performance of the servomultiplier using both types of modulator are shown in Table I.

It is seen that considerable improvement has been made in the performance of the servo by substitution of the square-wave modulator for the original one.

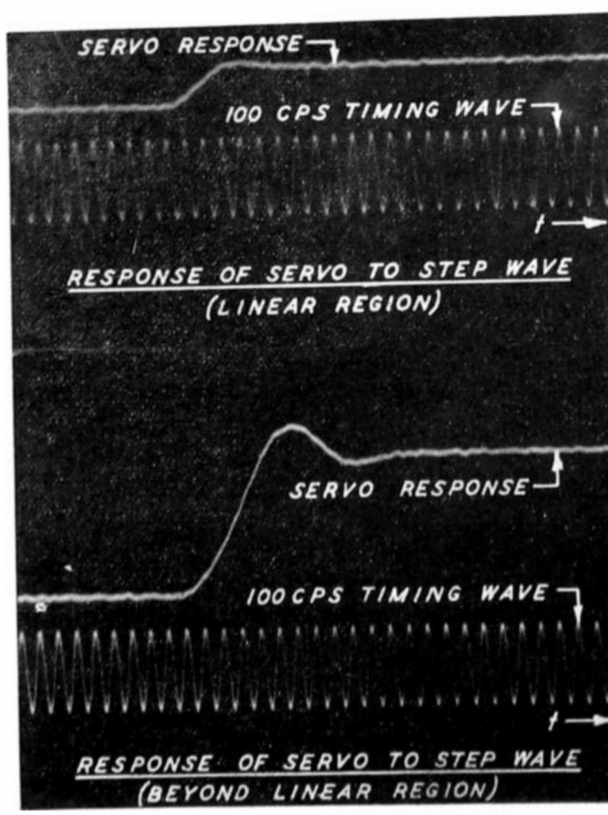
With the original modulator the system was limited in maximum usable loop gain, because noise and quadrature components produced heating in the 3C33 stage. Quadrature components, due to stray pulse pickup in the output circuit of the modulator, also limited the maximum usable loop gain after substitution in the square-wave modulator. The circuit could be markedly improved in this regard by careful attention to shielding. Even without such refinement, however, the maximum usable loop gain was greater by more than an order of magnitude when the new modulator was used.

Figs. 6(a) and 6(b) show the transient response of the system. It is seen that, whereas in Fig. 6(a), the rise time (10 to 90 per cent) is approximately 18 milliseconds, Fig. 6(b) shows a rise time of nearly 30 milliseconds. This anomaly is the result of difference in ampli-

TABLE I

RELATIVE PERFORMANCE OF SERVOMULTIPLIER

	Using G_M variation Modulator	Using Square-Wave Modulator
Position accuracy	1 part in 400	1 part in 10,000
Frequency response	6 db down at 5 cps	6 db down at 24 cps
Zero drift	1%	Not measurable better than 0.01%
Departure from linearity	5% over 50 volt range	Not measurable better than 0.01% over 70 volt range
Transient response to step function	Rise time (.1 to .9) 100×10^{-3} sec.	Rise time (.1 to .9) 18×10^{-3} sec. (See Fig. 7)
Loop gain factor (restoring torque per radian displacement)	1.8×10^6 dyne cm/radian (See Fig. 7)	4.0×10^6 dyne cm/radian (See Fig. 7)



(a)

(b)

Fig. 6—Oscillograms of transient response of servo.

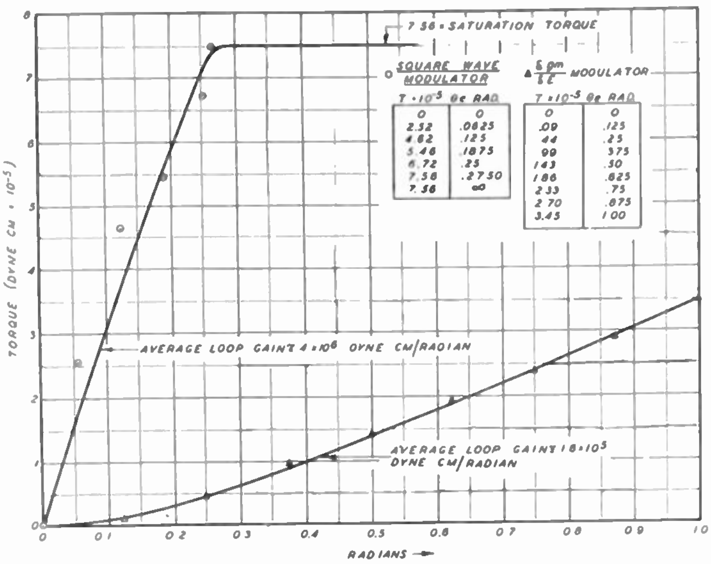


Fig. 7—Performance date of g_m variation modulator and square-wave modulator.

tude, and is a well-known phenomenon in servo systems. Fig. 7 shows that if a step request for shaft position change greater than 0.25 radian is given, the servomechanism will rapidly accelerate to maximum speed and run at this speed until a new balance point is reached. Thus the initial rate of increase in Fig. 6(b) is greater than in Fig. 6(a) for the first 10 milliseconds, and then no further increase is observed. The overshooting seen in Fig. 6(b) is due to the nonlinear behavior of the system. More damping is used than that desirable if the servo were never rapidly displaced more than 0.25 radian.

It should be emphasized that no particular attempt was made to exhaust the possibilities in this design, because performance was more than satisfactory in the intended application.

V. Conclusions

The performance indicates that the square-wave modulator system described is superior to types involving the use of curvature in tube characteristics.

Its sensitivity is not far from the maximum theoretically possible from non-amplifying devices.

The only modulators which appear to be capable of achieving better performance with regard to signal-to-noise ratio and limiting sensitivity are the mechanical types, such as the chopper and vibrator.

The use of the square-wave modulator should provide improved performance in many servo applications.

VI. Acknowledgments

The aid and advice of Gorman R. Nelson on the servo problems, and the care used in construction and wiring of the device by J. B. Ruble are gratefully acknowledged.

Resonant Circuits with Time-Varying Parameters*

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Summary—**With renewed interest in the superregenerator and similar circuits there has developed a need for a solution to the differential equation of a resonant circuit with time-varying parameters. An approximate solution is presented along with a criterion of error which determines the limits of accuracy of the solution. This error criterion shows that the solution may be applied to most communications problems with high accuracy. The method and ease of manipulation of the solution are demonstrated by application of the mathematics to the constant-parameter case. If direct manipulation of the formulas is not feasible, it may be easily applied to graphical analysis.**

Introduction

With the renewed interest in the superregenerator and the development of such systems as frequency modulation and servomechanisms, a great need has arisen for a useable solution to the differential equation of the resonant circuit with time-varying parameters. The basic problem is the mathematical formulation of an expression for the voltage across a resonant circuit as a function of the input current and the time-variation of the parameters G, C, and L. Previous papers[1–6] have treated the problem for the variation of one parameter only, but to the author's knowledge there has been no presentation of a solution

* Decimal classification: R141.2. Original manuscript received by the Institute, December 22, 1948; revised manuscript received, September 22, 1949.

† Massachusetts Institute of Technology, Cambridge, Mass.

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to the general case of a variation of all parameters simultaneously. Fox and Burns,[1] Bradley,[3,4] Riebman,[5] and Glucksman[6] all developed a quasistatic solution for the case of varying conductance only, but a determination of the accuracy of this solution has never been made. The purpose of this paper will then be twofold. First, the quasistatic solution shall be extended to the case of the variation of all parameters, and second, a criterion of error shall be developed for the solution presented. As to the advisability of depending upon a quasistatic solution, investigation seems to indicate that any attempt to arrive at an exact solution is extremely difficult and such a solution, if obtained, is of such a complexity as to have limited usefulness.

The Differential Equation and Its Reduced Solution

The differential equation for a parallel resonant circuit[7] with time-varying parameters may be written as

$$\frac{d}{dt}(Ce) + Ge + \frac{1}{L}\int e\,dt = i \tag{1}$$

which upon manipulation becomes

$$e'' + e'\left(\frac{2C'}{C} + \frac{G}{C} + \frac{L'}{L}\right) + e\left(\frac{C''}{C} + \frac{L'C'}{LC} + \frac{L'G}{LC} + \frac{G'}{C} + \frac{1}{LC}\right) = \left(\frac{i'}{C} + \frac{L'}{LC}i\right), \tag{2}$$

where primes denote differentiation with respect to time. Now the solution to this equation for zero driving current and constant parameters is given very nearly by

$$e = E_0\epsilon^{(\alpha + j\omega_0)t} \tag{3}$$

where

$$\alpha = -\frac{G}{2C} \tag{3a}$$

and

$$\omega_0 = \frac{1}{\sqrt{LC}}.$$

What we must do, then, is attempt to modify this equation to satisfy the new conditions, i.e., the time-variation of the previous constants G, C, and L. Physical reasoning presents us with two pertinent facts. First, we know that the real component of the exponent will now involve the capacitance and the inductance since they are now instantaneous sources and sinks of energy, because of their variation with time. Second, by comparison with the mathematics of frequency modulation we realize that we must now take the integral of the exponent rather than its instantaneous value. Our first guess is that the solution to the reduced equation is best given by

$$e = E_0\epsilon^{\int(\alpha + j\omega_0)dt} \tag{4}$$

or, more generally,

$$e = \mathrm{Re}\left[2E_0\epsilon^{\int(\alpha + j\omega_0)dt}\right] \tag{4a}$$

where α is to be determined and

$$\omega_0 = \frac{1}{\sqrt{LC}}.$$

Writing (2) in its reduced form as

$$e'' + Ae' + Be = 0, \tag{5}$$

we will now try to find an equation of the form

$$e'' + Ae' + B^*e = 0, \tag{6}$$

which satisfies the approximate formula. By placing this restriction on the approximate solution we shall determine α. Differentiating (4) and substituting into (6) we obtain

$$\alpha' + j\omega_0' + \alpha^2 - \omega_0^2 + 2j\alpha\omega_0 + A\alpha + jA\omega_0 + B^* = 0. \tag{7}$$

Now we know that A is given by

$$A = \frac{2C'}{C} + \frac{G}{C} + \frac{L'}{L}.$$

Thus, equating reals and imaginaries, we obtain

$$\alpha = -\frac{A}{2} - \frac{1}{2}\frac{\omega_0'}{\omega_0} = -\frac{G}{2C} - \frac{3}{4}\frac{C'}{C} - \frac{1}{4}\frac{L'}{L} \tag{8}$$

and

$$B^* = \frac{1}{LC} + \frac{7}{8}\frac{L'C'}{LC} + \frac{3}{4}\frac{C''}{C} + \frac{G'}{2C} + \frac{3}{16}\left(\frac{C'}{C}\right)^2 - \frac{1}{16}\left(\frac{L'}{L}\right)^2 + \frac{L''}{4L} + \frac{GC'}{2C^2} + \frac{1}{4}\left(\frac{G}{C}\right)^2 + \frac{1}{2}\frac{L'G}{LC}. \tag{9}$$

We now have an exact solution to the equation

$$e'' + Ae' + B^*e = 0, \tag{10}$$

which differs from the exact equation for the resonant circuit with time-varying parameters only in the linear coefficient B^*. We note that this error in the coefficient is

[7] Henceforth we shall treat the parallel circuit only. The solution to the series circuit is the same with the usual change of G to R, L to C, etc.

$$B^* - B = \frac{3}{16}\left(\frac{C'}{C}\right)^2 + \frac{L''}{4L} + \frac{GC'}{2C^2} + \frac{1}{4}\left(\frac{G}{C}\right)^2 - \frac{1}{16}\left(\frac{L'}{L}\right)^2 - \frac{1}{8}\frac{L'C'}{LC} - \frac{1}{4}\frac{C''}{C} - \frac{1}{2}\frac{L'G}{LC} - \frac{G'}{2C}. \quad (11)$$

We must now develop an error criterion to establish the relation of the magnitude of this error to the accuracy of the approximate solution.

The First-Order Error

Defining e as the approximate solution and δ as the first-order error in this solution, we write the following equations:

$$e'' + Ae' + B^*e = 0 \quad (12)$$

$$(e+\delta)'' + A(e+\delta)' + B(e+\delta) = 0. \quad (12a)$$

Now the above equations are linear; we may therefore subtract them, obtaining,

$$\delta'' + A\delta' + B\delta = (B^* - B)e. \quad (13)$$

Solving this equation by our approximate solution and the method of variation of parameters[8] the first-order error is found to be

$$\delta = Re\left[2E_1\epsilon^{\int(\alpha+j\omega_0)dt} + \epsilon^{\int(\alpha+j\omega_0)dt}\int\frac{(B^*-B)E_0\,t}{j\omega_0}\right]. \quad (14)$$

Here E_1 determines the initial value of the error, however this will of necessity be zero since we have knowledge of the initial voltage across the circuit, and thus make the error zero. Therefore the corrected solution to the reduced differential equation becomes

$$e + \delta = \mathrm{Re}\left[2E_0\epsilon^{\int(\alpha+j\omega_0)dt} \times \left(1 + \int\frac{B^*-B}{2j\omega_0}dt\right)\right], \quad (15)$$

where (B^*-B) is defined by (11).

Let us now examine the error term. Assuming the time-variation to be periodic, we may specify that (B^*-B) is of the form

$$(B^* - B) = K_0 + K_1\cos(pt+\phi_1) + K_2\cos(2pt+\phi_2) + \cdots + K_n\cos(npt+\phi_n), \quad (16)$$

where p is the fundamental frequency[9] of the time-variation of parameters. Now the constant K_0 may be dropped since we see that it actually produces a slight correction in the average frequency of the resonant circuits. This would be expected, since we have used the approximate formula for ω_0. Examination of the integral verifies this interpretation since K_0 produces a linearly-increasing phase shift with time whenever the integral is much smaller than one. Now the maximum instantaneous value of the remaining terms will be

$$|B^* - B - K_0|_{max} \leq K_1 + K_2 + \cdots K_n, \quad (17)$$

and the maximum possible value for the integral will be

$$\left|\int\frac{B^*-B-K_0}{2j\omega_0}dt\right|_{max} \leq \frac{1}{\omega_0}\left(\frac{K_1}{p} + \frac{K_2}{2p} + \cdots + \frac{K_n}{np}\right) \leq \frac{1}{\omega_0 p}|B^* - B - K_0|_{max}. \quad (18)$$

Thus the criterion for the accuracy of the approximate solution will be

$$\left|\int\frac{B^*-B-K_0}{2j\omega_0}dt\right|_{max} \ll 1 \quad (19)$$

or

$$|B^* - B - K_0|_{max} \ll \omega_{0min}p. \quad (20)$$

What we require for accuracy, then, is that the maximum deviation in the magnitude of (B^*-B) be much less than the product of p and the minimum value of ω_0. If a transient solution is desired, (19) may be applied directly. Application of this criterion to most problems encountered has shown that the formula in general produces accurate results.

The General Solution

Having obtained a reduced solution with error criterion, we may now derive the complete solution. By the method of variation of parameters[8] this is found to be

$$e = \mathrm{Re}\left[2E_0\epsilon^{\int(\alpha+j\omega_0)dt} + \epsilon^{\int(\alpha+j\omega_0)dt}\int\frac{\left(i' + \frac{L'}{L}i\right)\epsilon^{\int(\alpha+j\omega_0)dt}dt}{j\omega_0 C}\right]. \quad (21)$$

Placing limits on the proper integrals such that we may apply the equation to a circuit at time, $t=a$,

[8] L. R. Ford, "Differential Equations," McGraw-Hill Book Co., New York, N. Y., 1946, pp. 75–77.

[9] For simplicity we shall call ω frequency rather than angular velocity.

$$e = \mathrm{Re}\left[2E_a\epsilon^{\int_a^t(\alpha+j\omega_0)d\tau} + \epsilon^{\int(\alpha+j\omega_0)dt}\int_a^t \frac{\left(i' + \frac{L'}{L}i\right)\epsilon^{-\int(\alpha+j\omega_0)d\tau}d\tau}{j\omega_0 C}\right]. \qquad (22)$$

We must now place definite limits on the remaining integrals. Writing these as follows:

$$\int(\alpha + j\omega_0)dt = \int_0^t (\alpha + j\omega_0)dx + C_1 \qquad (23)$$

$$\int(\alpha + j\omega_0)d\tau = \int_0^\tau (\alpha + j\omega_0)dx + C_2, \qquad (24)$$

we see that $C_1 = C_2$ since the integrals were derived in similar fashion. Bringing the second exponential within the integral

$$e = \mathrm{Re}\left[2E_a\epsilon^{\int_a^t(\alpha+j\omega_0)d\tau} + \int_a^t \frac{\left(i' + \frac{L'}{L}i\right)\epsilon^{\int_\tau^t(\alpha+j\omega_0)dx}d\tau}{j\omega_0 C}\right]. \qquad (25)$$

Note that under the main integral $(\alpha+j\omega_0)$ in the exponent is a function of x, and the remaining terms, with the exception of t, are all functions of τ. This is the key equation for the solution of any problem to be treated.

To demonstrate the use of the above equation in the treatment of a steady-state problem, let us take the simple case of the constant-parameter circuit merely to show how the equation may be manipulated. We wish to find the response of the circuit to a current given by

$$i = I_0 \cos \omega_s t = I_0 \frac{(\epsilon^{j\omega_s t} + \epsilon^{-j\omega_s t})}{2}. \qquad (26)$$

Then

$$i' = j\omega_s I_0 \frac{(\epsilon^{j\omega_s t} - \epsilon^{-j\omega_s t})}{2} \qquad (27)$$

where ω_s is the frequency of the applied current. Substituting into (25), having dropped the first term, since we desire the steady-state solution,

$$e = \mathrm{Re}\left[\int_{-\infty}^t \frac{\left(j\omega_s + \frac{L'}{L}\right) I_0\epsilon^{\int_\tau^t(\alpha+j\omega_0)dx}\epsilon^{j\omega_s\tau}d\tau}{2j\omega_0 C} - \int_{-\infty}^t \frac{\left(j\omega_s - \frac{L'}{L}\right) I_0\epsilon^{\int_\tau^t(\alpha+j\omega_0)dx}\epsilon^{-j\omega_s\tau}d\tau}{2j\omega_0 C}\right]. \qquad (28)$$

We also set the lower limit of the integral equal to $-\infty$ for steady-state applications. At this point we specify that G, C, and L are constant. Then

$$e = \mathrm{Re}\left[\int_{-\infty}^t \frac{j\omega_s I_0\epsilon^{(t-\tau)(\alpha+j\omega_0)}\epsilon^{j\omega_s\tau}d\tau}{2j\omega_0 C} - \int_{-\infty}^t \frac{j\omega_s I_0\epsilon^{(t-\tau)(\alpha+j\omega_0)}\epsilon^{-j\omega_s\tau}d\tau}{2j\omega_0 C}\right]. \qquad (29)$$

The second integral is now dropped because of the $\epsilon^{-j(\omega_s+\omega_0)\tau}$ term. This is the same as the narrow-band approximation used in the more familiar treatment. The resultant steady-state expression for the voltage across the resonant circuit is

$$e = \mathrm{Re}\left[\frac{\omega_s I_0\epsilon^{-(G/2C)t}\epsilon^{j\omega_0 t}}{2\omega_0 C}\int_{-\infty}^t \epsilon^{(G/2C)\tau}\epsilon^{j(\omega_s-\omega_0)\tau}d\tau\right] \qquad (30)$$

which, when the integral is evaluated, is

$$e = \mathrm{Re}\left[\frac{\omega_s I_0\epsilon^{j\omega_s t}}{2\omega_0 C\left(\frac{G}{2C} + j\omega_s - j\omega_0\right)}\right]. \qquad (31)$$

This is more readily recognized if we cancel ω_s and ω_0, since they are approximately equal in the region of importance. The final result is

$$e = \mathrm{Re}\left[\frac{I_0}{G + 2j(\omega_s - \omega_0)C}\epsilon^{j\omega_s t}\right] \qquad (32)$$

which takes on the familiar form of the single-tuned response curve.

The application of the equations to actual time-varying parameter problems will not be presented here, as we feel that such a presentation cannot have sufficient generality in any specific case to warrant its inclusion in a paper of this type. The reader is referred to such papers as those of Bradley,[3,4] and Riebman,[5] for an application for one specific parameter only. It might be noted that the mathematical formulation is easily adaptable to graphical solution where direct integration is not feasible.

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Abstracts and References

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ACOUSTICS AND AUDIO FREQUENCIES

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534.84 3016

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Equivalent Electrical Circuit for the Piezoelectric Sound-Receiver—O. Schäfer. (*Akus. Zeit.*, vol. 6, pp. 326–328; November, 1941.)

621.392.51 3021

Piezoelectric Transducers—W. Roth. (Proc. I.R.E., vol. 37, pp. 750–758; July, 1949.) A piezoelectric transducer operating in the thickness mode is represented as a six-terminal network. The mesh equations, electromechanical impedance matrix, and equivalent circuit valid for any general conditions of loading and frequency are obtained. Equations for the electrical driving-point impedance are derived.

621.395.61/.62 3022

A Low-"Q" Directional Magnetostrictive Electroacoustic Transducer—L. Camp and F. D. Wertz. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 382–384; July, 1949.) Description of a lamination design for the magnetostrictive motors of a directional transducer array. Efficient operation is possible with a Q of 6 under a full water load.

621.395.61 3023

A New Dynamic Microphone—W. Baer. (*Akus. Zeit.*, vol. 8, pp. 127–135; August, 1943.) Theory is based on the equivalent circuit. Details of the construction of the Beyer moving-coil microphone are given.

621.395.61 3024

The B.B.C.-Marconi Ribbon Microphone Type AXBT—E. T. Wrathall. (*Marconi Rev.*, vol. 12, pp. 92–103; July and September, 1949.) Designed in 1933 by the research department of the BBC for studio use. The microphone consists of a crimped Al-foil ribbon supported between specially shaped pole-pieces of a powerful Alcomax permanent magnet. The ribbon resonates at a subaudio frequency; the amplitude of its motion is proportional to the particle velocity of the incident sound wave. The horizontal and vertical polar curves of sensitivity have the figure-of-eight shape. The ribbon is electrically screened and is protected from draughts and dust. Maintenance requirements are negligible over long periods.

621.395.61:621.396.822 3025

The Noise Level of High-Grade Microphones—W. Weber. (*Akus. Zeit.*, vol. 8, pp. 121–127; August, 1943.) The noise levels of microphones of the dynamic and capacitive types, made by several well-known firms, were calculated and determined experimentally. For the dynamic microphones the thermal noise of the internal resistance is about 10 times that of the first tube of the amplifier. The capacitor microphone itself produces no noise, but noise voltages are introduced by the load resistance in the low-frequency arrangement and by the resonance resistance of the oscillatory circuit in the high-frequency arrangement. The noise level in the low-frequency arrangement is again higher than that of the first amplifier tube, at the lower and middle frequencies by about 10 db, falling at the higher frequencies to about that of the tube. With suitable design of the high-frequency arrangement the noise level can be reduced by about 10 db.

621.395.623.7:791.45 3026

Theater Loudspeaker Design, Performance and Measurement—J. K. Hilliard. (*Jour. Soc. Mot. Pic. Eng.*, vol. 52, pp. 629–640; June, 1949.)

621.395.623.7.089.6 3027

Physical Measurements of Loudspeaker Performance—Veneklasen. (*See* 3207.)

621.395.625.2:621.392.52 3028

Crossover Filter for Disk Recording Heads—Roys. (*See* 3085.) (*Broadcast News*, no. 55, pp. 20–23; June, 1949.)

621.395.625.3 3029

Tape Characteristics for Audio Quality—R. Marchant. (*Tele-Tech*, vol. 8, pp. 30–33, 57; July, 1949.) A general survey covering the electrical and mechanical properties of the tape, mechanical requirements of the driving and tension devices and the playback head, the effect of the main supply voltage on the tape velocity, and precautions to avoid deterioration of the tape in storage.

621.395.625.3 3030

Factors Affecting Spurious Printing in Magnetic Tapes—S. W. Johnson. (*Jour. Soc. Mot. Pic. Eng.*, vol. 52, pp. 619–627; June, 1949. Discussion, pp. 627–628.) Study of the amount of signal printed through from adjacent layers in a roll of magnetic type. The effects of time, temperature, and output level of the original recording are considered.

621.395.625.3 3031

Optimum High-Frequency Bias in Magnetic Recording—G. L. Dimmick and S. W. Johnson. (*Broadcast News*, no. 55, pp. 4–7; June, 1949.)

534.321.9:620.179.16 3032

Ultrasonics [Book Review]—B. Carlin. Publishers: McGraw-Hill Book Co., New York, 1949, 270 pp., $5.00. (*Electronics*, vol. 22, pp. 209–210; August, 1949.) For readers with an engineering viewpoint. Piezoelectric and magnetostriction generators are considered as well as cw and pulsed systems. The use of ultrasonics for flaw detection is the application most thoroughly covered.

ANTENNAS AND TRANSMISSION LINES

621.315.1:531.221.8 3033

Graphical Calculation of Tension Tables for Overhead Lines and Horizontal Copper-Wire Aerials—C. M. A. Carranza. (*Rev. Telecommun.* (Madrid), vol. 3, pp. 16–23; June, 1948.) Abacs for determining tension, sag, and safety factor as a function of temperature for various lengths of span and diameters of wire.

621.315.212:621.3.09 3034

Coaxial Cables—P. Schiaffino and L. Albanese. (*Poste e Telecomun.*, vol. 17, pp. 85–104; February, 1949.) A study of propagation phenomena, particularly those affecting telephony, transmission characteristics, existing types of coaxial cable, and the problems associated with laying and joining such cables, fault clearance and the S.A.C.M. echo meter noted in 142 of February (Couanault and Herreng).

621.392.26† 3035

Waveguides—R. Malvano. (*Ricerca Sci.*, vol. 18, pp. 1595–1612, November and December, 1948.) Transmission theory with special reference to waveguides with dielectric inserts.

621.392.26† 3036

On the Theory of the Excitation of Radio Wave Guides—G. V. Kisunko. (*Compt. Rend. Acad. Sci.* (URSS), vol. 51, pp. 199–202; January 30, 1946. In English.) Formulas are derived for the field due to conductors with arbitrary current distribution and of arbitrary configuration within the waveguide.

621.392.26† 3037

Reflection Cancellation in Waveguides—L. Lewin. (*Wireless Eng.*, vol. 26, pp. 258–264; August, 1949.) A tapered section with uniform variation is commonly used to provide a smooth transition between waveguides of different cross-sections. For freedom from reflections, the length l of the tapered section should be as great as possible, and must exceed $\lambda/2$. Minima of total reflection occur when $l=n\lambda/2$, and maxima when $l=(2n+1)\lambda/4$, n being an integer. The case of small-angle tapers is discussed, with special reference to the diaphragm method of reflection compensation. The field-fitting method yields results which can be generalized and are applied to the general double taper.

621.392.26† 3038

Experimental Investigation of the Reflections produced in a Waveguide by any Dielectric—L. R. Noriega. (*Rev. Telecomun.* (Madrid), vol. 3, pp. 2–10; June, 1948.) Verification of theory developed by L. W. Holmboe in a thesis entitled "Reflections produced at the Junction of Two Rectangular Waveguides, one filled with a Single Dielectric and the other with Two Dielectrics."

621.392.26† 3039

Notes on the Excitation of Electromagnetic Waves in Cylindrical Metal Waveguides—A. Colino. (*Jour. Appl. Phys.*, vol. 20, pp. 576–577; June, 1949.) Starting from Maxwell's equations, general formulas applicable to cylindrical waveguides of any cross-section are derived. These formulas are of simple structure and for the case of a waveguide of circular cross-section excited by an antenna on the axis result in a formula identical with one given by Schelkunoff. For the original version (in Spanish) see *Rev. Telecomun.* (Madrid), vol. 3, pp. 34–37; June, 1948.

621.392.26†:621.396.611.4 3040

The Analogies between the Vibration of Elastic Membranes and the Electromagnetic Fields in Guides and Cavities—E. C. Cherry. (*Proc. IEE* (London), Part III, vol. 96, pp. 346–359; July, 1949. Discussion, pp. 358–360.) Detailed relations are shown between the fields in guides and cavities, and the vibrations of elastic sheets having similar boundaries. Only one of the two sets of electromechanical analogies commonly applied to circuits such as transmission lines and filters is applicable to distributed systems. In this, velocity corresponds to voltage (or E-vector) and force to current (or magnetic H-vector). Mass corresponds to capacitance (or κ) and elastic constant to inverse inductance (or $1/\mu$). A study of membrane vibrations may assist in the design of microwave components and in the derivation of new selective network forms.

621.392.26†:621.396.67 3041
Electromagnetic Radiation from Waveguides and Horns—L. Lewin. (*Nature* (London), vol. 164, p. 311; August 20, 1949.) The method of Levine and Schwinger (1845 of 1948) for analyzing the acoustic radiation from circular pipes is applied to the case of em radiation from rectangular waveguides.

621.392.26†:621.396.67 3042
The Electromagnetic Horn: Parts 1 and 2—W. D. Oliphant. (*Electronic Eng.* (London), vol. 21, pp. 255–258 and 294–299; July and August, 1949.) A survey paper, in which much of the information is abstracted from the 16 references given. Basic theory, design principles, and experimental results are discussed, with special reference to rectangular waveguides and the sectoral horn.

621.392.26†:621.396.67 3043
Transmission-Line Characteristics of the Sectoral Horn—H. S. Bennett. (Proc. I.R.E., vol. 37, pp. 738–743; July, 1949.) The sectoral horn is considered as one component of a microwave transmission system. Equivalent network functions are derived and plotted, the sectoral horn being regarded as a nonuniform transmission line. The physical significance of the derived normalized functions is discussed.

621.392.26†:621.396.67 3044
Laws of Potential Distribution along Slits [slots]—J. N. Feld. (*Compt. Rend. Acad. Sci.* (URSS), vol. 55, pp. 407–410; February 20, 1947. In English.) Approximate formulas, applicable to narrow radiating slots in waveguides, are derived.

621.392.43.012.3 3045
High-Frequency Transmission Line Chart—P. R. Clement (*Electronics*, vol. 22, pp. 104–105; August, 1949.) Determination of input impedances and matching-stub dimensions is simplified by means of a chart in which straight lines are used instead of curves as in circle or Smith diagrams.

621.396.67 3046
The Measurement and Interpretation of Antenna Scattering—D. D. King. (Proc. I.R.E., vol. 37, pp. 770–777; July, 1949.) The significance of scattering and back-scattering cross-sections in terms of antenna current distribution is considered, with particular reference to the influence of antenna load impedance on the magnitude and directional pattern of the scattered radiation. A method of measurement, which uses the standing waves set up by energy reflected toward the transmitter from any receiving antenna or parasite, permits direct study of the back-scattering from loaded and unloaded antennas. Approximate scattering data for several types of antenna are included.

621.396.67 3047
Experimental Determination of the Distribution of Current and Charge along Cylindrical Antennas—G. Barzilai. (Proc. I.R.E., vol. 37, pp. 825–829; July, 1949.) Using a wavelength of 1.90 m, the distributions are determined for center-fed straight cylindrical antennas of diameter 29 mm and lengths 1.25λ, 1.00λ, and 0.50λ respectively. In some cases parasitic antennas were added. The experimental accuracy was checked by means of measurements on a coaxial line whose inner conductor had the same diameter as the antennas.

621.396.67 3048
Radiating Surface Systems—J. N. Feld. (*Compt. Rend. Acad. Sci.* (URSS), vol. 51, pp. 203–206; January 30, 1946. In English.) A closed metal surface of dimensions comparable with the wavelength can in certain cases, if suitably excited, compare favorably with ordinary radiating systems. Formulas are derived for the em field and the surface distribution of current for a spherical-surface antenna excited by a known current distribution along a radial conductor inside the surface. The method adopted for the solution can easily be generalized for surface antennas of arbitrary form with an arbitrary arrangement of coupling elements.

621.396.67 3049
Diffraction Antennae with Axial Symmetry—J. N. Feld. (*Compt. Rend. Acad. Sci.* (URSS), vol. 51, pp. 115–118; January 20, 1946. In English.) The antennas considered are obtained by cutting the surface of endovibrators. Formulas are derived and applied to the determination of the field inside and outside a sphere from which a narrow belt has been cut out, the excitation being due to a dipole at the center.

621.396.67 3050
Excitation of a Hollow Spherical Resonator by a Dipole Placed at Its Centre—S. M. Rytov. (*Compt. Rend. Acad. Sci.* (URSS), vol. 51, pp. 111–115; January 20, 1946. In English.) Formulas are derived for the field within the resonator and for the energy dissipation. A small hole on the equator of the sphere radiates like a magnetic dipole parallel to the equator.

621.396.67 3051
Discone—40 to 500 Mc/s Skywire—J. M. Boyer, (*CQ*, vol. 5, pp. 11–15, 71; July, 1949.) The evolution of the discone from a flared open-ended waveguide is traced. Dimensions and construction details are given of three models whose frequency ranges are respectively 40 to 500 Mc, 400 to 1,200 Mc, and 800 to 5,000 Mc. The last provides a means of measuring the radiation pattern of the discone by the model technique. See also 303 of March (Kandoian, Sichak, and Felsenheld).

621.396.67.016.31 3052
A Power-Equalizing Network for Antennas—R. W. Masters. (Proc. I.R.E., vol. 37, pp. 735–738; July, 1949.) A bridge type of network which causes equal power to be delivered to two load impedances whose product is a predetermined real constant. The input impedance of the bridge is practically independent of suitably paired load-impedance values over a considerable band of frequencies. Application to the design of television broadcast antennas is indicated, examples are given, and power loss is discussed.

621.396.671 3053
On the Theory of the Radiation Patterns of Electromagnetic Horns of Moderate Flare Angles—C. W. Horton. (Proc. I.R.E., vol. 37, pp. 744–749; July, 1949.) A method attributed to Schelkunoff for the computation of radiation patterns is considered. For the case of transverse electric waves in a waveguide or horn of moderate flare angle, the radiation pattern is calculated in terms of two definite integrals. These integrals are evaluated for rectangular, circular, and semicircular horns for some common modes of vibration. Experimental results are in agreement with theory.

621.396.671 3054
The Radiation Resistance of Linear Aerials—A. A. Samarski and A. N. Tikhonov. (*Zh. Tekh. Fiz.*, vol. 19, pp. 792–803; July, 1949. In Russian.) The reactive component of the radiation resistance is calculated for a given current distribution. This component remains finite only in the case of a tuned dipole.

621.396.671 3055
The Transmitter Dipole—J. Müller-Strobel and J. Patry. (*Schweiz. Arch. Angew. Wiss. Tech.*, vol. 14, pp. 306–314; October, 1948.) Theory previously given (1334 of 1947) is applied to the calculation of the radiation impedance of vertical wire antennas with and without resistance losses. A practical formula is derived and its accuracy checked by measurements on a balloon-supported antenna. The discrepancies between theory and experiment can be easily explained. The problem of antenna resonance can be solved by means of the accurate formula, but in most cases a simple relation can be used.

621.396.671 3056
Evaluation of the Gain of a Microwave Radiating System—F. Bosinelli. (*Ricerca Sci.*, vol. 18, pp. 1009–1015; August and September, 1948.) For an antenna with a parabolic reflector the gain G is given approximately by the formula $G=4\pi I_1^2/\lambda^2 I_2$, where I_1 and I_2 are integrals involving the field distribution over the aperture of the reflector. Particular distributions are considered and a curve is given showing the diminution in gain for parabolic distributions more or less accentuated.

621.396.671 3057
Gain of Aerial Systems—D. A. Bell. (*Wireless Eng.*, vol. 26, pp. 306–312; September, 1949.) The maximum gain of an antenna of given aperture depends on the phase distribution of the illumination of the aperture. Three cases are considered, in order of increasing gain: (a) uniform-phase radiators (broadside arrays and "optical" radiators), (b) radiators, with effective phase-shift of π (end-fire antennas of all kinds), and (c) antennas with closely-spaced phase reversals (high-gain short antennas). See also 302 of February (Woodward and Lawson).

621.396.677 3058
Path-Length Microwave Lenses—W. E. Kock. (Proc. I.R.E., vol. 37, pp. 852–855; August, 1949.) Baffle plates extend parallel to the magnetic vector, and are suitably tilted or bent to force the waves to follow a longer path. The plate array is shaped to correspond to a convex lens. Advantages over other types of metallic lens are: broader-band performance, greater simplicity, and less severe tolerances. See also 2176 of 1948.

621.396.677:621.396.93 3059
Direction-Finding Site Errors at Very High Frequencies—Hopkins and Horner. (*See* 3145.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.2 3060
Impedance of Composite Conductors—(*Wireless Eng.*, vol. 26, pp. 267–275; August, 1949.) Schelkunoff's theorem for calculating this impedance (435 of 1935) is developed to derive useful working formulas. The theorem can be deduced by an extension of a method devised by Howe. Simplifications of the formulas have been made in the case of coaxial tubular conductors, where the radii of the conductors are electrically large and the curvature may be neglected.

621.3.016.35 3061
A New Harmonic Method for Studying the Stability of Linear Systems—Demontvignier and Lefèvre. (*See* 3127.)

621.3.087.4:551.510.535 3062
New Equipment for the Systematic Recording of Ionospheric Echoes—A. Bolle. (*Ann. Geofis.*, vol. 1, pp. 164–174; April, 1948.) Circuit details and description of apparatus with which the band 2.5 to 20 Mc is swept at 3-minute intervals. As in the equipment described by Sulzer (2983 of 1946) the v.f.o. controls both the frequency of the transmitter and that of the heterodyne stage of the receiver, the frequency being varied periodically by means of an electric motor driving gearing attached to a variable capacitor. Peak pulse power is about 20 kW at the lower frequencies.

621.314.2:621.396.611.33/.34 3063
A Design for Double-Tuned Transformers—J. B. Rudd. (*Jour. Brit. I.R.E.*, vol. 9, pp. 306–316; August, 1949.) The term "transformer" is used in the broad sense to include networks consisting of a pair of L-C circuits with either inductive or capacitive coupling. A

method is described of designing such transformers to provide uniform power transmission over a given frequency range; the insertion-loss curve is then approximately symmetrical when plotted on a linear frequency scale. The frequency variable used in the design equations allows a common representation of both inductively and capacitively coupled transformers. The extent of the uniform-transmission band and the transformation ratios possible with various types of coupling are discussed. Charts are presented which simplify the design procedure. Practical circuits satisfying specified conditions are given. See also 2177 of 1948 (Rideout).

Reprinted from *Proc. I.R.E.* (Australia), vol. 10, January, 1949.

621.314.3† 3064

An Analysis of Magnetic Amplifiers with Feedback—D. W. Ver Planck, M. Fishman, and D. C. Beaumariage. (Proc. I.R.E., vol. 37, pp. 862–866; August, 1949.) Full paper; summary noted in 2448 of October

621.314.3† 3065

The Transductor, D.C. Pre-Saturated Reactor, with Special Reference to Transductor Control of Rectifiers—U. Lamm. (*Acta Polyt.* (Stockholm), no. 17, 215 pp.; 1948. In English.) Fundamental principles are outlined and various methods of using transductors for regulation and control purposes are described. Similarity laws for the practical design of a series of transductors from measured values for one unit of the series are derived. A theory of transductor-controlled multiphase rectifiers is developed. Static and dynamic conditions are treated and general equations for rapid calculation of transductor-regulator parameters are derived. Experimental results confirm the theory.

621.316.313 3066

Electrical Network Analyzers for the Solution of Electromagnetic Field Problems: Parts 1 and 2—K. Spangenberg, G. Walters, and F. Schotts. (Proc. I.R.E., vol. 37, pp. 724–729 and 866–872; July and August, 1949.) 1948 IRE Convention paper noted in 2475 of 1948. Discussion of the design and construction of two analyzers for solving the wave equation in two-dimensional axially-symmetric cylindrical co-ordinates and in rectangular co-ordinates. The use of these analyzers is also considered for determining various modes of concentric lines, waveguides and resonators, field-strength distributions, resonant frequencies of cavities, etc.

621.316.8:621.396.822 3067

Noise from Current-Carrying Resistors 20 to 500 kc/s—R. H. Campbell, Jr., and R. A. Chipman. (Proc. I.R.E., vol. 37, pp. 938–942; August, 1949.) The dc noise voltage for some resistors has fluctuations much larger than those characteristic of thermal noise. This effect was investigated experimentally for solid carbon-composition, metallized palladium film, and pyrolitic-carbon resistors, for resistances from 1 to 30 kΩ and currents from 1 to 10 mA. The fluctuations are large when current is first applied to a resistor, before its resistance reaches an equilibrium value at a higher temperature. The effect is thus analogous to the Barkhausen magnetization effect, but cannot as yet be correlated with other factors.

621.316.86 3068

Thermistors—G. Pierry. (*Toute la Radio*, vol. 16, pp. 240–242; September, 1949.) A short account of different types, their properties and uses.

621.317.35 3069

On Some Properties of Signals with Limited Spectra—J. Oswald. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 21–22; July 4, 1949.) Every signal function $x(t)$ whose corresponding Fourier function $X(f)$ is zero outside a finite interval $(-f_1, f_1)$ can be developed in a series of orthogonal functions. The equidistant ordinates x_n completely determine such a signal; they are the components of a vector of the Hilbert subspace defined by the segment $(-f_1, f_1)$ in the space (f) derived from the space (t) by the Fourier unitary transformation. All the parameters of the signal $x(t)$ can be expressed in terms of the coefficients x_n and all the transformations of $x(t)$ by linear operators can be studied equally well by means of these coefficients. Examples are given.

621.318.572 3070

New Design for a Secondary-Emission Trigger Tube—C. F. Miller and W. S. McLean. (Proc. I.R.E., vol. 37, pp. 952–954; August, 1949.) 1948 IRE National Convention paper. A triode input section produces a primary electron beam which impinges on a dynode to produce secondary electrons. These are collected by two different output elements which may be used separately or as a unit. A dynode surface having long life and stability is described. Suggested applications include its use as a relaxation oscillator, multivibrator, pulse inverter, triangular-wave generator, and dynatron. See also 1567 of 1942 (Skellett).

621.318.572 3071

Admittance of the 1B25 Microwave Switching Tube—R. W. Engstrom and A. R. Moore. (Proc. I.R.E., vol. 37, pp. 879–881; August, 1949.)

621.319.4:621.315.614 3072

Paper Capacitors using Chlorinated Liquid Impregnants—C. G. Farley. (Proc. I.R.E. (Australia), vol. 9, pp. 13–17; July, 1948. Discussion, pp. 17–18.) Trends in the development and use of synthetic impregnants are discussed, the characteristics of chlorinated naphthalene, chlorinated diphenyl, and natural impregnating compounds such as castor oil are compared, and the physical and electrical properties of capacitors impregnated with pentachlorodiphenyl are tabulated and discussed.

621.392 3073

Bridged Reactance-Resistance Networks—G. R. Harris. (Proc. I.R.E., vol. 37, pp. 882–887; August, 1949.) 6-arm, 6-element *R-C* bridged networks are considered. Six symmetrical structures exist having the infinite attenuation property of the parallel-T network. The duality of certain pairs of these structures is demonstrated.

621.392 3074

Effective and Circuit Band-Widths—W. J. Kessler. (*Elec. Eng.*, vol. 68, p. 590; July, 1949.) Summary only. The effective bandwidth of any network of maximum response A is defined as the bandwidth of an equivalent network whose response is A throughout the transmission band, provided the noise powers developed across the output terminals of the two networks are equal for the same noise-signal input. The term "circuit bandwidth" is reserved to specify the selectivity or frequency-discriminating properties of a network. The effective bandwidth is equal to the area under the squared response curve divided by the square of the maximum response. For a single elementary *L-C* network the ratio of effective to circuit bandwidth (3-db attenuation) is equal to $\pi/2$. This ratio approaches 1.07 as the number of such elementary circuits in cascade increases and approaches unity as the response curve approaches a rectangular form.

621.392 3075

The Gyrator, a New Circuit Element—H. Feigs. (*Funk. und Ton.*, vol. 3, pp. 459–465; August, 1949.) A shortened version of Tellegen's recent work (980 of May and 2745 of November).

621.392 3076

Miller Effect—"Cathode Ray." (*Wireless World*, vol. 55, pp. 307–312; August, 1949.) A step-by step resolution of some of its paradoxes for resistive and for reactive loads.

621.392:517.433 3077

Operational Approach to Nonlinear Circuit Analysis—G. H. Cohen. (*Jour. Frank. Inst.*, vol 247, pp. 573–581; June, 1949.) The operational method can be extended to nonlinear-circuit problems by first expanding the expression for the unknown variable i in a power series of the driving function e. Each term of the series represents a component of the total current. Each component is an operational expression for a linear differential equation involving the current component i_r to the first power, the impedance C_r corresponding to this current component, and the rth power of the driving voltage. The nonlinearity is thus shifted from the unknown dependent variable i to the known independent variable e, making it possible to find and solve the transformed equations for each current component.

621.392.4 3078

Constant-Phase-Shift Networks—R. O. Rowlands. (*Wireless Eng.*, vol. 26, pp. 283–287; September, 1949.) To every phase-shift network there corresponds an attenuation network whose attenuation is directly related to the phase shift of the first network. This attenuation network, being easier to design, is designed first, and from its parameters those of the phase-shift network are deduced.

621.392.4:621.3.015.3 3079

The Energy of a Passive Linear Two-Terminal Network in the Transient Regime—M. Abele. (*R. C. Accad. Naz. Lincei*, series 8, vol. 1, pp. 1321–1324; December, 1946. In Italian. Reprint.) General treatment for the case where the applied emf is a periodic function of time. This is extended to the case of a nonperiodic emf of short duration, by considering it as a single period of a periodic emf.

621.392.43 3080

Compact Antenna-Coupling Device—S. Wald. (*Radio and Telev. News, Radio-Electronic Eng. Supplement*, vol. 12, pp. 7, 30; March, 1949.) Description, with illustrations, of a continuously variable inductor with two independent tappings which can be used for antenna tuning and matching over a wide frequency range.

621.392.5:621.3.015.3 3081

Transients in the Low-Pass Filter—G. Newstead and D. L. H. Gibbings. (*Proc. IEE* (London), part III, vol. 96, pp. 264–268; July, 1949.) Formulas for the termination current are given and plotted for various impulsive voltage inputs. Limitations of the usual approximate treatments are discussed. A solution is obtained for a uniformly dissipative low-pass filter terminated in a resistance of $\sqrt{(L/C)}$.

621.392.5:681.142 3082

Mercury Delay Line Memory Using a Pulse Rate of Several Megacycles [per second]—I. L. Auerbach, J. P. Eckert, Jr., R. F. Shaw, and C. B. Sheppard. (Proc. I.R.E., vol. 37, pp. 855–861; August, 1949.) The possible pulse rate has been effectively doubled by means of the pulse envelope system of representing data. The control of signals at high pulse rates has been achieved by means of crystal gating circuits. A multichannel memory using a single pool of mercury has simplified mechanical construction and temperature control, and has reduced the size. Intelligence can be transmitted and received by the memory system described at 5×10^6 binary digits per second.

621.392.52 3083

A Valve-Assisted Filter for Audio Frequencies—J. D. Storer. (*Jour. Brit. I.R.E.* vol. 9, pp. 268–275; July, 1949.) The filter combines the function of voltage limiting and wave filtering, and consists of a flip-flop oscillator which is inoperative until triggered. Its oscillation frequency is controlled by that of the triggering voltage, which is developed across a reactive circuit. Characteristics are summarized and circuit diagrams illustrating the applications of the filter are given.

621.392.52 3084

RC Filter Networks—A. Sabbatini. (*Poste e Telecomun.*, vol. 16, pp. 83–88; March, 1948.) Detailed analysis of (a) a bridge-type circuit for phase variation due to Scott (1802 of 1938), (b) a resonance potential divider due to Willoner and Tihelka (*Hochfrequenztech. u. Elektroakust.*, vol. 61, p. 48; February, 1944), (c) a R-C amplification stage, and (d) R-C low-pass and high-pass filters.

621.392.52:621.395.625.2 3085

Crossover Filter for Disk Recording Heads—H. E. Roys. (*Broadcast News*, no. 55, pp. 20–23; June, 1949.)

621.395.665.1 3086

Contrast Expansion—G. Mitchell and J. G. White. (*Wireless World*, vol. 55, pp. 315–316; August, 1949.) Comment on 2171 of September (Wheeler).

621.396.611.1 3087

Iterative Impedance and Resonance Curve of Symmetrical Homogeneous Recurrent Circuit—P. Kalantarov and L. Zeitlin. (*Compt. Rend. Acad. Sci.* (URSS), vol. 51, pp. 281–284; February 10, 1946. In English.)

621.396.611.1 3088

On Approximate Integration for Oscillatory Systems with One Degree of Freedom—V. V. Kazakevitch. (*Compt. Rend. Acad. Sci.* (URSS), vol. 51, pp. 107–110; January 20, 1946. In French.) A method enabling the building-up process, the form and the period of the oscillations of a system to be determined for the case where there are no external perturbations.

621.396.611.1 3089

Resonance Phenomena in Homogeneous Symmetrical Recurrent Circuits—P. L. Kalantarov and L. A. Zeitlin. (*Compt. Rend. Acad. Sci.* (URSS), vol. 51, pp. 357–360; February 20, 1946. In English.)

621.396.611.1:621.3.015.3 3090

A Note on the Transient Response of an Oscillatory Circuit with Recurrent Discharge—A. M. Hardie. (*Phil. Mag.*, vol. 40, pp. 748–759; July, 1949.) Such circuits were discussed by Wilkinson (864 of 1948). The general characteristics and duration of the transient are here considered and illustrated graphically. The transient response to a voltage step-function is calculated. Account is also taken of circuit losses.

621.396.611.4 3091

Approximate Integration of Maxwell's Equations [for stationary e.m. fields] Inside a Cavity Resonator—M. Abele. (*Atti Accad. Sci. Torino*, vols. 81 and 82, pp. 159 167; 1945 and 1947. Reprint.) Calculations applied in 3823 of 1947.

621.396.611.4 3092

Nodal Planes in a Perturbed Cavity Resonator: Parts 2 and 3—K. F. Niessen. (*Appl. Sci. Res.*, vol. B1, pp. 251–260 and 284–298; 1949.) A mathematical paper. The resonator considered is rectangular and has one movable wall. In part 1 (994 of May) a vibration without nodal planes was considered. In part 2 a vibration with a single nodal plane (a) perpendicular, (b) parallel to the movable wall is discussed. Part 3 is concerned with the case where the vibration in the unperturbed resonator has two nodal planes, one perpendicular and the other parallel to the movable wall. The perturbed field is determined in each case. See also 1330 of June.

621.396.611.4:621.392.26† 3093

The Analogies between the Vibration of Elastic Membranes and the Electromagnetic Fields in Guides and Cavities—Cherry. (*See* 3040.)

621.396.615 3094

Study of the Transmission-Line Oscillator with Ordinary Valves—R. de Magondeaux. (*Radio Franç.*, nos. 6, 7, and 8, pp. 21–24 and 13–19; June, July, and August, 1949.) Theory of the operation of simple oscillators for wavelengths between 4 and 40 cm. Relations between the various currents and voltages are shown graphically, and output efficiency and radiated power are considered. Such oscillators are particularly suitable for demonstration purposes.

621.396.615:621.317.083.7 3095

Transistor Oscillator for Telemetering—F. W. Lehan. (*Electronics*, vol. 22, pp. 90–91; August, 1949.) An oscillator used for FM of the telemetering transmitter. Advantages are noted. Variation of frequency with transistor temperature is undesirable. Temperature compensation is being investigated.

621.396.615.029.64:621.316.726 3096

An Analysis of the Sensing Method of Automatic Frequency Control for Microwave Oscillators—E. F. Grant. (Proc. I.R.E., vol. 37, pp. 943–951; August, 1949.) Circuits using a simple cavity resonator for the stable element and either FM of the controlled oscillator or modulation of the cavity resonance frequency are analyzed to obtain effective discriminator curves which give a null output for the average cavity resonance frequency. The complete afc loop gain, the best method of decreasing the pulling of the oscillator frequency by the cavity, and the pulling of the cavity frequency by a variable-susceptance load are discussed.

621.396.615.17 3097

Pulsed Stimulator Aids Medical Research—L. A. Woodbury, M. Nickerson, and J. W. Woodbury. (*Electronics*, vol. 22, pp. 84–85; August, 1949.) A multivibrator-controlled constant-current pulse generator. Pulse duration, 0.025–1.5 ms. Repetition rate, 0.1–1,000 pulses per second. Output continuously variable from zero to over 1,000 mA.

621.396.615.17:621.317.755:621.397.6 3098

Television Time Base Linearisation—A. W. Keen. (*Electronic Eng.* (London), vol. 21, pp. 195–198, 223; June, 1949.) Linearity correction by the integration method is discussed. A simple sawtooth generator consisting of a series *R-C* circuit connected across a source of constant dc voltage, with a discharge device connected across the capacitor, has an exponential output when the discharger is inoperative. An additional *R-C* circuit can be associated with the generator so that a suitable proportion of the output of the second circuit is added to that of the generator to make the resultant essentially linear. Details of practical correction circuits are discussed.

621.396.615.17:621.397.645.001.4 3099

Video Amplifier Testing—Using a Square-Wave Generator—T. B. Tomlinson. (*Electronic Eng.* (London), vol. 21, pp. 204–208; June, 1949.) The square-wave generator described is of a conventional type using a multivibrator whose output is clipped by means of a tube operating near cut-off. The output of the squaring tube is fed into a cathode follower to prevent waveform deterioration when working into a considerable load-capacitance. Modifications of the square wave by the more common types of distortion are shown and discussed.

621.396.615.18 3100

High-Ratio Multivibrator Frequency-Divider—M. Silver (*Radio and Telev. News, Radio-Electronic Eng. Supplement*, vol. 13, pp. 7–9; 20; July, 1949.) Theory and description of a stable circuit capable of division ratios as high as 300:1. Component details are furnished of a circuit giving a 15-kc output from a 4.5-Mc input; only two 6SN7 double triodes are required.

621.396.645 3101

High-Quality Amplifier: New Version—D. T. N. Williamson. (*Wireless World*, vol. 55, pp. 282–287; August, 1949.) Modifications of an earlier model (2715 of 1947) with construction data and details of the necessary adjustments to give linear response with low harmonic and intermodulation distortion. The impedances for various connections of the output transformer secondary are tabulated. Negative feedback and the prevention of instability are discussed.

621.396.645 3102

Some Aspects of Cathode-Follower Design at Radio Frequencies—F. D. Clapp. (Proc. I.R.E., vol. 37, pp. 932–937; August, 1949.) Simple design charts, derived by approximations which are applicable over a wide range of frequency and of circuit parameters, for determining at hf the circuit gain, the gain phase angle, the input impedance in resistive and reactive components, the maximum allowable input signal voltage, etc. Various circuit changes which reduce or eliminate the undesirable effects of the grid/cathode capacitance are discussed.

621.396.645 3103

A Wide-Band Amplifier (100 cls to 20 Mcls)—J. C. Plowman. (*Electronic Eng.* (London), vol. 21, pp. 338–340; September, 1949.) A two-stage, filter-coupled amplifier with cathode-follower output, giving an over-all gain of 38 db in the frequency range 100 cps-16 Mc, with a slight falling off at higher frequencies.

621.396.645:621.3.015.3 3104

Design of Optimum Transient Response Amplifiers—P. R. Aigrain and E. M. Williams. (Proc. I.R.E., vol. 37, pp. 873–879; August, 1949.) The method described is derived from operational analysis using Laplace transforms. It is based on transient considerations, and not derived from steady-state theories. Applications to video amplifiers, symmetrical bandpass amplifiers, and unsymmetrical bandpass amplifiers with low-level modulation are discussed.

621.396.645:621.3.015.3 3105

Transient Response of Wideband Amplifiers—W. E. Thomson. (*Wireless Eng.*, vol. 26, pp. 264–266; August, 1949.) A suitable 2-terminal load for a wideband amplifier stage is the "infinite-order critically-damped load" discussed in 671 of 1947. This load gives the fastest unit-step response without overshoot. Any desired approximation to the compensating reactance can be obtained by using one of a certain series of networks. The second member of this series consists of one inductor and one capacitor, and gives a result adequate for most practical purposes.

621.396.645:621.396.615.142 3106

Application of Velocity-Modulation Tubes for Reception at U.H.F. and S.H.F.—M. J. O. Strutt and A. van der Ziel. (Proc. I.R.E., vol. 37, pp. 896–900; August, 1949.) Discussion on 1890 of 1948.

621.396.645:621.396.813 3107

On Criteria for the Permissible Non-Linear Distortion of Amplifiers—V. F. Schut and C. W. Kosten. (*Appl. Sci. Res.*, vol. B1, pp. 261–267; 1949.) The sum of the amplitude of the second harmonic and twice that of the third harmonic appears to be a good criterion. For reproduction of moderate quality this sum should not exceed 24 per cent of the amplitude of the fundamental.

621.396.645.029.3:621.385.3 3108
A Low-Noise [audio] Input Tube—Knight and Haase. (*See* 3303).

621.396.645.029.4/.5 3109
On the Amplification of the Low Frequencies in Wide-Band Amplifiers—W. Dillenburger. (*Funk. und Ton.*, vol. 3, pp. 423–428; August, 1949.) An extension of the frequency band towards the lower frequencies is made possible by a circuit which increases the effective coupling time-constant between two amplifier stages. The effect of the time-constant of the cathode circuit can be completely compensated with a suitably designed anode circuit, so that the frequency characteristic of the amplification is determined only by the coupling member for the stage. A design is given in which the effect of the filter capacitors of the supply unit on the amplitude characteristic at low frequencies is reduced, as well as the kipp oscillations which may occur with more than two stages.

621.396.645.37 3110
Feedback Amplifier Design—H. Mayr. (*Wireless Eng.*, vol. 26, pp. 297–305; September, 1949.) Discussion of design when the response curve is pre-selected and has no spurious peaks. A simple general equation is derived which gives the frequency response of the amplifier with feedback, if the response of the amplifier without feedback and the frequency characteristics of the feedback network are known. The special case of amplifiers with up to four stages of resistance-capacitance or tuned-circuit coupling, with constant feedback and equal center frequencies for all stages is considered; design formulas are given.

621.396.645.371 3111
Negative Feedback Amplifiers—T. S. McLeod. (*Wireless Eng.*, vol. 96, pp. 312–313; September, 1949.) Comment on 2768 of November (Brockelsby).

621.396.69 3112
Circuit Techniques for Miniaturization—P. G. Sulzer. (*Electronics*, vol. 22, pp. 98–99; August, 1949.) Controlled positive feedback between stages can often be used to avoid the necessity for bulky circuit components, such as cathode and screen by-pass capacitors and video-amplifier compensating inductors, with corresponding reduction in both size and cost.

621.397.62 3113
Transit-Time Effects in Television Front-End Design—H. M. Watts. (*Electronics*, vol. 22, pp. 158, 170; August, 1949.) The effect of transit-time is to add about 4 to the noise figure near the frequency where the transit-time loading conductance equals the desired input conductance. Transit-time effects tend to level out the differences between various circuit combinations, so that as the frequency increases, the reduction of noise from sources other than transit-time decreases in importance

621.397.645 3114
Television Stabilizing Amplifier—J. L. Schultz. (*Radio and Telev. News, Radio-Electronic Eng., Supplement*, vol. 12, pp. 12–15, 28; May, 1949.) Full circuit details and special features of a unit which can be used in the studio or at the transmitter as a picture-line amplifier, or as a program amplifier for a line or radio link.

621.396.69 3115
Components Handbook [Book Review]—J. F. Blackburn (Ed.). Publishers: McGraw-Hill Book Co., New York, 1949. 626 pp., \$8.00. (*Electronics*, vol. 22, pp. 212, 214; August, 1949.) Vol. 17 of the M.I.T. Radiation Laboratory series. Only components developed by or under the sponsorship of the Radiation Laboratory or of primary importance in its work are covered thoroughly, and several important classes of these components have been left out. Most of the data included have not been published before.

GENERAL PHYSICS

535.3 3116
Quantitative Evidence for Boundary-Layer Waves in Optics—H. Maecker. (*Ann. Phys.* (Lpz.), vol. 4, pp. 409 431; June 25, 1949.) The existence of such waves is established. Experimental results are in good agreement with Ott's theory (18 of 1943 and 3117 below). The connection between optical boundary-layer radiation and the ray shift in total reflection described by Goos and Hänchen (*ibid.*, vol. 1, p. 333; 1947.) is examined.

535.3 3117
On the Reflection of Spherical Waves—H. Ott. (*Ann. Phys.* (Lpz.), vol. 4, pp. 432–440; June 25, 1949.) Previous calculations for the Schmidt "head wave" [Kopfwelle] for the vertical dipole (18 of 1943) are extended to dipoles with any direction whatever. See also 3116 above.

535.42 3118
On the Theory of Diffraction—W. Franz. (*Z. Phys.*, vol. 125, pp. 563–596; March 15, 1949.) An approximation method is given for the solution of acoustical and optical diffraction problems, which includes Kirchhoff's diffraction theory as a special case of the first approximation. While Kirchhoff's method is concerned only with diffraction at a black screen, with the present method reflection and refraction can also be treated and its application is not limited to short wavelengths. The first and second approximations are applied to the case of the semi-infinite plane and the higher approximations to that of the small sphere.

535.42 3119
An Asymptotic Treatment of Diffraction Problems—N. G. van Kampen. (*Physica's Grav.*, vol. 14, pp. 575–589; January, 1949. In English, with French summary.) An asymptotic development of Kirchhoff's integral for $\lambda \rightarrow 0$ is given. The first term corresponds to geometrical optics, and includes intensity and phase. The other terms are the corrections for diffraction. The diffraction at the edge of the opening is treated quasi-geometrically. The theory is applied to optical systems, and the third-order aberration constants are calculated.

537.291+538.691]:537.525.92 3120
Electron Flow in Curved Paths under Space-Charge Conditions—B. Meltzer. (*Proc. Phys. Soc.* (London), vol. 62, pp. 431–437; July 1, 1949.) A general, synthetic method of obtaining rigorous solutions of steady electron flow subject to space-charge forces is presented. The solutions are not obtained for given boundary conditions, but the boundary conditions are deduced from the solutions. Two examples of such solutions, involving strongly curved two-dimensional electron trajectories, are given; the method is in principle capable of giving the solutions of all possible electron flow patterns in three dimensions except perhaps those involving intercrossing trajectories. It is suggested that the subject offers scope for applied mathematical research at least on the same scale as potential theory.

537.311.4 3121
Contact Resistance and Its Variation with Current—S. Rudeforth. (*P. O. Elec. Eng. Jour.*, vol. 42, pp. 65–69; July, 1949.) Empirical relationships have been derived for the non-linear resistance/current characteristics of specified contacts.

537.523.4 3122
Calculation of Spark Breakdown Voltages in Air at Atmospheric Pressure—A. Pedersen. (*Appl. Sci. Res.*, vol. B1, pp. 299–305; 1949.) Discussion of a new semi-empirical criterion for breakdown, which depends on the ion density.

538.3 3123
The Experimental Basis of Electromagnetism: Parts 3 and 4—N. R. Campbell and L. Hartshorn. (*Proc. Phys. Soc.* (London), vol. 62, pp. 422–429 and 429–444; July 1, 1949. Discussion, pp. 444–445.) The principles outlined in previous parts (3091 of 1947 and 1909 of 1948) dealing with the dc circuit and electrostatics are here applied to magnetism to show how the basic concepts are defined in terms of the operations performed in measuring them. The vector B is established as measurable everywhere, even within solid bodies. The vector H and the scalar $\mu = B/H$ are shown to be measurable in special circumstances by means of the magnetometer and permeameter, but in general their values depend on a hypothesis, which is stated. The significant facts concerning the magnetic properties of real materials are briefly reviewed.

538.566 3124
Diffraction of Electromagnetic Waves by a Perfectly Conducting Plane Screen—J. P. Vasseur. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 179–181; July 18, 1949.) Copson, in his treatment of the problem, omitted from his equations a curvilinear integral which, though zero in the examples he considered, may be important in other cases. The correct solution is here given, the formulation of the equations being analogous to that of Bethe (706 of 1945).

538.569.4.029.64 3125
Microwave Spectroscopy—W. Gordy. (*Rev. Mod. Phys.*, vol. 20, pp. 668–717; October, 1948.) General discussion of instruments, experimental methods, spectra of gases, vapors, liquids and solids, and applications.

538.569.4.029.64+537.226.2]:546.212 3126
Electrical Properties of Water—J. A. Saxton. (*Wireless Eng.*, vol. 26, pp. 288–292; September, 1949.) Anomalous dispersion occurs mainly between the frequencies of 10^3 and 10^6 Mc. Over this interval the permittivity of water falls from 80 to 5.5. The ionic conductivity of fresh water is important only at frequencies below 10^3 Mc, and that of sea water at frequencies below 2×10^4 Mc. The effect of anomalous dispersion on the reflection coefficient of fresh water surfaces is considered See also 1912–1915 of 1948.

621.3.016.35 3127
A New Harmonic Method for Studying the Stability of Linear Systems—M. Demontvignier and P. Lefèvre. (*Rev. Gén. Élec.*, vol. 58, pp. 263–279; July, 1949.) The mathematical basis and the physical significance of the usual harmonic methods are reviewed and Nyquist's criterion of stability is generalized. The principles are explained of a new method, of very general application, which can be applied to any linear system, starting from its generalized phase diagram. Several abacs are given which enable the phase and amplitude diagrams to be traced quickly. The method is applied to the theory of the stability of servomechanisms. See also 1568 of 1948 (Rocard) and back references, for which the above U.D.C. would have been preferable.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

521.15:538.12 3128
Theory of the Relations between Gravitation and Electromagnetism and Their Astrophysical and Geophysical Applications—A. Gião. (*Jour. Phys. Radium*, vol. 10, pp. 240–249; July and September, 1949.) Application to space-time of Codazzi's equations for a hypersurface leads to fundamental relations between gravitation and electromagnetism, when the external metric tensor of space-time is interpreted as a tensor of the em field. This interpretation is a consequence of the fundamentals of the author's unitary theory, according wo

which all the em properties of the universe are described, directly or indirectly, by the external metric tensor of space-time.

Application of general formulas to the particular case of a sphere in rotation without permanent magnetization gives an important relation between the magnetic moment and the moment of inertia, and proves that the general magnetism of large rotating bodies, such as the stars, is a fundamental consequence of their rotation. Formulas are obtained for the em field of a sphere in rotation which explain the general magnetic field of the earth both external to the surface and underneath it. The same formulas can also be applied to the permanent and periodic magnetism of stars. Codazzi's equations lead to a relation between gravitation and the es field which serves to explain both the mean es field of the earth and the maintenance of its charge. See also 1023, 1634, and 2776 of 1948.

523.72.029.63"1949.05.08":523.75 3129

Exceptional Solar Radio Emission during 8th May 1949—M. Laffineur and R. Servajean. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 110–112; July 11, 1949.) Records of solar radiation on a wavelength of 54.5 cm, obtained towards sunset at Meudon observatory, showed large variations of intensity, which at times exceeded 5 times that of the quiet sun. Simultaneous spectroheliscope observations revealed an intense solar eruption.

An increase in the number of atmospherics on a wavelength of 11,500 m was recorded at the same time at Bagneux, Bordeaux, Poitiers, and Rabat. A fade-out of the sw transmissions from Leipzig on 9.732 Mc and from WWV (Washington) on 15 Mc was noted at Bagneux.

A very small crochet in the record of the vertical component of the earth's magnetic field was noted at Chambon-la-Forêt observatory. This crochet coincided with a 10-second jump of 54.5-cm intensity to 4.7 times that of the quiet sun.

Phenomena probably associated with this solar activity were (a) a lowering of the critical frequency at 0400 on May 11, which was the start of a perturbation of the height of the ionized layers, particularly the F_2 layer, which reached an abnormal height between 0600 and 0700 on May 13; (b) a sudden drop in the value of the horizontal component of the earth's magnetic field at 0200 on May 11, and a violent magnetic storm on May 12 and 13.

550.383 3130

The Magnetic Field within the Earth—E. C. Bullard. (*Proc. Roy. Soc. A*, vol. 197, pp. 433–453; July 7, 1949.) A discussion on the magnetic effects of motion in the earth's core. Tidal friction, fluctuations in the rate of rotation, nutation, and the variation of latitude have negligible magnetic effects. Radioactivity of core material will greatly affect the internal field as a result of thermal convection. This field is larger and more complex than was previously believed and its existence confirms the induction theory of the origin of the secular variation. See also 381 of March

551.510.535 3131

The Ionosphere over Mid-Germany in June 1949—Dieminger. (*Fernmeldetech. Z.*, vol. 2, p. 244; August, 1949.) Continuation of 2792 of November. A whole series of weak and medium disturbances of the F_2 layer were observed during the month

551.510.535 3132

Ionospheric Virtual Height Measurements at 100 kc/s—R. A. Helliwell. (Proc. I.R.E., vol. 37, pp. 887–894; August, 1949.) A simple high-power sounding equipment is described. Results of intermittent night-time measurements of virtual height at vertical incidence are discussed. The virtual height varied between 84 km and as much as 106 km. At night the reflecting layer appears to consist of ionized clouds, in contrast to the more uniform ionization of the regular layers which affect hf waves. A rotation of the polarization of the reflected signal relative to that of the transmitted signal was observed.

551.510.535:525.624:550.384.4 3133

Lunar Oscillations in the D-Layer of the Ionosphere—E. V. Appleton and W. J. G. Beynon. (*Nature* (London), vol. 164, p. 308; August 20, 1949.) The daily measurements of ionospheric absorption made at Slough during the period 1943–1948, using a frequency of 2 Mc, indicate a lunar oscillation in D-layer absorption. If the D-layer electrons move up and down between levels of different electron collision frequency, the oscillation would be almost exactly out of phase with that known to exist in the higher E layer.

551.510.535:621.3.087.4 3134

New Equipment for the Systematic Recording of Ionospheric Echoes—Bolle. (*See* 3062.)

551.510.535:621.396.11 3135

Correlation of Sporadic E Region Ionization over Short Distances and Comparison with Magnetic Disturbances—V. B. Gerard. (*N.Z. Jour. Sci. Tech.*, vol. 30, pp. 27–37; July, 1948.) Simultaneous observations of sporadic-E ionization were made at points separated by distances up to 40 km, using two fixed stations and portable recording equipment. An approximately linear relationship was found between the correlation coefficient of simultaneous sporadic-E critical frequencies at two points, and the distance between the points. A method of calculating the muf for sporadic-E communication over distances up to 100 km is outlined. One observation suggests that a particular sporadic-E cloud had a diameter of 540 km and a velocity of 270 km per hour. No relationship between sporadic-E changes and changes in the earth's magnetic field could be detected. See also 3117 of 1948 (Ferrell) and 3410 of 1948 (Revirieux: Lejay).

551.524.4 3136

The Vertical Temperature Gradient in the Lower Atmosphere under Daylight Conditions—G. W. C. Tait. (*Quart. Jour. R. Met. Soc.*, vol. 75, pp. 287–292; July, 1949.) An empirical relationship is derived for the first 10–20 m of the atmosphere, in terms of the position of the sun and cloud cover. This relationship is independent of wind speed, and applies to reasonably level surfaces of soil or short vegetation during daylight, but does not apply to open water surfaces.

551.524.7 3137

The Thermal Equilibrium at the Tropopause and the Temperature of the Lower Stratosphere—R. M. Goody. (*Proc. Roy. Soc. A*, vol. 197, pp. 487–505; July 7, 1949.) Continuity of temperature at the tropopause is a necessary condition for stable transition from a state of convective equilibrium to one of radiative equilibrium.

551.594 3138

On the Fundamental Problem of Atmospheric Electricity—T. Schlomka. (*Z. Phys.*, vol. 125, pp. 733–738; March 5, 1949.) It is shown that Michel's theory (2086 of 1941) is untenable, since it is based on an erroneous assumption. The maintenance of the earth's negative charge requires a process continuously supplying a negative charge to the earth's surface. This, however, is not the case for unipolar induction with a rotating earth, which can only produce a static charge distribution in the atmosphere and within the earth.

551.594.5 3139

Auroral Radiation in the 3000-Mc/s Region—P. A. Forsyth, W. Petrie, and B. W. Currie. (*Nature* (London), vol. 164, p. 453; September 10, 1949.) Short pulses of radiation were observed on the indicator of a 3,000-Mc radar during an auroral display even when the transmitter was off. These pulses arrived in a random manner, in bursts lasting a small fraction of a second. Individual pulses lasted 1–5 μs. See also 2522 of October (Petrie, Forsyth, and McConechy).

LOCATION AND AIDS TO NAVIGATION

621.396.9 3140

A Forward-Transmission Echo-Ranging System—D. B. Harris. (Proc. I.R.E., vol. 37, pp. 767–770; July, 1949.) 1949 IRE Convention paper noted in 1672 of June (No. 92). A proposed system with p.p.i. display for detecting targets such as atmospheric irregularities, which have a low reflection coefficient except at grazing incidence. Transmitter and receiver are about 100 miles apart. Microwave pulses lasting about 0.01 μs are required.

621.396.93 3141

V.H.F. Direction Finder for Light Planes—G. Wennerberg. (*Electronics*, vol. 22, pp. 118, 140; August, 1949.) This omni-range system provides azimuth information directly in degrees for an aircraft in any position within the line-of-sight range of the transmitting station. The frequency used is within the band 108 to 132 Mc and the system has a useful working range of 50 to 100 miles. The basic principle is the same as that of the German Sonne system; navigational information is supplied as the time difference at the receiving point between a nondirectional signal and one transmitted on a rotating beam from the same transmitter.

621.396.93 3142

The Relative Merits of Presentation of Bearings by Aural-Null and Twin-Channel Cathode-Ray Direction-Finders—S. de Walden and J. C. Swallow. (*Proc. IEE* (London), part III, vol. 96, pp. 307–320; July, 1949.) The visual method of bearing display is shown to be superior in nearly all respects except for its relative ineffectiveness at very low signal-to-noise ratios.

621.396.93 3143

The Specification and Measurement of Polarization Errors in Adcock-Type Direction Finders—W. Ross. (*Proc. IEE* (London), part III, vol. 96, pp. 269–277; July, 1949.) For instruments erected not more than about λ/4 above the ground, the "standard wave error" is the best specification of polarization error, while the "pick up ratio" for wanted and unwanted fields is appropriate for more elevated systems. The method of test using a nearby elevated transmitter is described in detail; in the frequency range 3 to 30 Mc a loop up to about 1.6 m in diameter at a distance of not less than 100 m may be used. For frequencies below 3 Mc the "local-injection" method of test may be more practicable. The performance of a direction finder is very dependent on the electrical properties of the site.

621.396.93:621.396.11 3144

Scattering of Radio Waves by Metal Wires and Sheets—Horner. (*See* 3237.)

621.396.93:621.396.677 3145

Direction-Finding Site Errors at Very High Frequencies—H. G. Hopkins and F. Horner. (*Proc. IEE* (London), part III, vol. 96, pp. 321–332; July, 1949. Discussion, pp. 340–345.) Theoretical and practical investigations are described, concerned mainly with Adcock-type direction finders. The variation in error with the position of various types of reflecting obstacle is examined and the use of error charts for locating such sources of error on a site is described. Of several practical methods suggested for locating the obstacles, two have been used with success, namely the variation of the azimuth or of the frequency of the transmitter. Methods of suppressing unwanted reflections are considered. A criterion is suggested to express the susceptibility of a direction finder to site error, and is applied to well-known instrumental types.

621.396.932 3146

The Design and Characteristics of Marine Radar Equipment—A. Levin and A. C. D Haley. (*Jour. Brit. I.R.E.*, vol. 9, pp. 202–219; June, 1949.) Discussion of the use of radar for coastal navigation and collision warning. Information made available by radar is compared with that available visually. The p.p.i. display is almost always used. The effect of meteorological phenomena on the choice of wavelength, and the limitations imposed by vessel size and the space available for the installation are considered. Experimental procedure for determining the main constants of the radar equipment is described, and typical results are discussed and illustrated.

621.396.933+629.139.83 3147

What We Learned from the Berlin Airlift—M. A. Chaffee and R. B. Corby. (*Electronics*, vol. 22, pp. 78–83; August, 1949.) The control of the aircraft was achieved by means of a combination of conventional radio ranges and homing beacons with appropriate airborne range receivers, long-range surveillance radar, precision landing-approach radar and vhf voice-communication equipment. The radar system used was the American CPS-5, with a range of more than 100 miles, capable of high accuracy and incorporating moving-target indication. The technique of video-mapping was also used.

621.396.933 3148

The Course-Line Computer for Radio Navigation of Aircraft—F. J. Gross. (Proc. I.R.E., vol. 37, pp. 830–834; July, 1949.) 1948 IRE Convention paper noted in 2524 of 1948. The range and bearing of an aircraft from a vhf omnirange station are converted into distance from destination and lateral deviation from a selected course. Circular courses may also be flown with the range station at the center.

621.396.933 3149

Modern Air and Ground Instrumentation in America's Air Navigation Program—D. W. Rentzel. (*Instruments*, vol. 22, pp. 492–493, 542; June, 1949.) The omni-range system provides a visual indication of the bearing of the aircraft from a fixed station. Associated distance-measuring equipment at the station and a lightweight airborne "course-line computer" enable the pilot to fly a straight course between any two selected points. Two landing aids—Instrument Landing System and Precision Beam Radar—are briefly discussed. For another account see 2236 of September (Sandretto).

621.396.9 3150

Principles and Practice of Radar [Book Review]—H. E. Penrose. Publishers: G. Newnes, London, 42s. (*Engineer* (London), vol. 188, p. 71; July 15, 1949.) "... a book for practical men rather than for theorists ... deserves a place on any radio engineer's shelves."

621.396.9 3151

A Textbook of Radar [Book Review]—Staff of the Radiophysics Laboratory, Council for Scientific and Industrial Research, Australia. Publishers: Chapman and Hall, London, 1948, 579 pp., 50s. (*Proc. Phys. Soc.*, vol. 62, pp. 465–466; July 1, 1949.) The work of the wartime radiophysics team in Australia. The whole field is covered in 20 chapters by 21 contributors. The editing has been well done and in consequence there are few obscurities and few definite mistakes. It should be of value to any serious student of radio.

621.396.9 3152

Microwaves and Radar Electronics [Book Review]—E. C. Pollard and J. N. Sturtevant. Publishers: J. Wiley und Sons, New York, 1948, 414 pp., $5.00; Chapman and Hall, London, 30s. (Proc. I.R.E., vol. 37, p. 785; July, 1949. *Wireless Eng.*, vol. 26, p. 313; September, 1949.) The book presents to the engineer, who has had little or no experience with microwaves, the fundamental and practical aspects of microwave and radar engineering. Only a working knowledge of physics, calculus, and tube theory and practice is assumed.

621.396.9 3153

The War History of the Radio Branch [Book Notice]—Publishers: National Research Council of Canada, Ottawa, Report No. ERA-141, 131 pp. An account of Canadian radar research up to the end of 1945.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.56 3154

Design Calculations for Molecular Vacuum Pumps—R. Risch. (*Schweiz. Arch. Angew. Wiss. Tech.*, vol. 14, pp. 279–285; September, 1948.)

535.37 3155

On the Quenching of the Luminescence of Certain ZnS–Cu and CaS–Bi Phosphors—F. Bandow. (*Ann. Phys. (Lpz.)*, vol. 1, pp. 399–404; October 1, 1947.)

535.37 3156

Decay and Quenching of Fluorescence in Willemite—F. A. Kröger and W. Hoogenstraaten. (*Physica's Grav.*, vol. 14, pp. 425–441; September, 1948. In English.)

535.37 3157

The Influence of Temperature Quenching on the Decay of Fluorescence—F. A. Kröger, W. Hoogenstraaten, M. Bottema, and T. P. J. Botden. (*Physica,'s Grav.*, vol. 14, pp. 81–96; April, 1948. In English.) Temperature quenching increases the rate of decay considerably. The probabilities of fluorescence transition and of radiationless transition are determined separately as functions of temperature; these two probabilities determine both the efficiency of fluorescence and the decay. Results favor the theory of Mott and Seitz for the radiationless process.

535.371.07:621.385.832 3158

The Physics of Cathode Ray Tube Screens—G. F. J. Garlick. (*Electronic Eng.* (London), vol. 21, pp. 287–291; August, 1949.) Discussion of the characteristics of screen materials, and of the mechanism of processes involved in screen luminescence.

538.22 3159

Magnetic Properties of Ferrites; Ferrimagnetism and Antiferromagnetism—L. Néel. (*Ann. Phys.* (Paris), vol. 3, pp. 137–198; March and April, 1948.) Comprehensive discussion, with detailed theory and comparison with experimental results.

538.22 3160

Dispersion and Absorption in Magnetic Ferrites at Frequencies above One Mc/s.—J. L. Snoek. (*Physica,'s Grav.*, vol. 14, pp. 207–217; May, 1948.) The contribution of the Bloch boundaries to magnetization is neglected at frequencies above 1 Mc. For pure and unstrained polycrystalline aggregates of cubic crystals, the critical frequency ω_0 and the initial susceptibility χ satisfy the equation

$$\omega_0\chi = 3/2^{\frac{1}{2}}g|M$$

where $g = e/mc = 1.76 \times 10^6$, M is the magnetic moment per cm³, and the damping is assumed small. Internal stresses tend to increase the losses at lower frequencies.

538.221 3161

Tentative Theory of the Magnetic Properties of Rhombohedric Sesquioxide of Iron—L. Néel. (*Ann. Phys.* (Paris), vol. 4, pp. 249–268; May and June, 1949.)

538.6 3162

Systematic Relations Existing between the Properties of Solid Materials—C. Zwikker. (*Physica,'s Grav.*, vol. 14, pp. 35–47; January, 1948. In English.) Volume changes due to electrostriction, magnetostriction, or Barrett effect are discussed. Four independent relations are found between the following eight quantities: Hall effect, Seebeck effect, Nernst effect, Peltier effect, Ettinghausen effect, Leduc-Righi effect, electric resistivity, and thermal resistivity.

538.652 3163

Longitudinal Magnetostriction of the Ferrites of Nickel and Magnesium—R. Vautier. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 177–179; July 18, 1949.) The ferrites investigated all contained 50 per cent Fe_2O_3, with either NiO or MgO in proportions up to 50 per cent, the remainder being ZnO. Results are shown graphically.

546.212-16:621.317.335.3† 3164

The Electrical Behaviour of Ice—F. X. Eder. (*Ann. Phys.* (Lpz.), vol. 1, pp. 381–398; October 1, 1947.) See 1077 of 1948.

548.0:537.228.1 3165

Piezoelectric Resonator of Ethylene Diamine Tartrate with Zero Temperature Coefficient of Frequency—R. Bechmann. (*Nature* (London), vol. 164, pp. 190–191; July 30, 1949.) For the contour shear mode in square plates with two sides parallel to the y axis and with the normal in the xz plane, the temperature coefficient of frequency is zero when θ, the angle between the normal and the x axis, is 17° or 77°.

621.315.5/.6+669 3166

The Development of New Materials—F. E. Robinson. (*Marconi Rev.*, vol. 12, pp. 108–116; July and September, 1949.) Discussion of metals, alloys, and insulating materials developed during the war and only recently released to industry.

621.315.5/.6 3167

Materials Section—(*Electronics Buyers' Guide Issue*, vol. 22, pp. M1–M32; June, 1949.) Electrical, mechanical, and other significant characteristics of various materials used by the electronic industry are tabulated. Some of the tables are new; others are revised forms of tables such as those noted in 3131, 3133, 3142, 3146, 3150, and 3152 of 1948.

621.315.59:546.281.26 3168

The Structure and Electrical Properties of Surfaces of Semiconductors: Part 1—Silicon Carbide—T. K. Jones, R. A. Scott, and R. W. Sillars. (*Proc. Phys. Soc.*, vol 62, pp. 333–343; June 1, 1949.)

621.315.59:621.3.011.2 3169

The Temperature Dependence of the Resistance of Semiconductors—J. H. Gisolf. (*Ann. Phys.* (Lpz.), vol. 1, pp. 3–26; January 3, 1947.)

621.318.22 3170

Permanent Magnets and the Electrical Industry—A. Edwards. (*Electrician*, vol. 142, pp. 1567–1571; May 20, 1949.) The magnetic energy of an Alcomax-III magnet is nearly 20 times that of the best magnet of equal size available 30 years ago. Corresponding advances in stability have also taken place. The properties and treatment of Alcomax and other materials are discussed, and the magnet shapes required for various applications are considered.

621.318.42:538.213 3171

A Method, Based on the Gaus Function, for Calculating the Effective Permeability of Premagnetized Choke Cores—A. Weis. (*Funk. und Ton.*, vol. 3, pp. 438–448; August, 1949.) Calculation from the B/H curve gives values for the effective permeability which are much too high. The method here described gives results in good agreement with measured values for air-gap chokes.

621.775.7 3172
Methods of Iron Powder Manufacture and Their Influence on Powder Properties—H. Bernstorff. (*Metal Treat.*, vol. 16, pp. 93–102; Summer, 1949.) A review of various German methods, including grinding, atomization, chemical reduction, and electrolysis. The influence of particle size and shape (shown in micrographs) on some of the physical properties of the finished product is discussed.

669.018:621.3.011.2 3173
Pressure and Temperature Coefficients of the Electrical Resistance of Certain Alloys—H. Ebert and J. Gielessen. (*Ann. Phys.* (Lpz.), vol. 1, pp. 229–240; May 22, 1947.) The results of measurements on a large number of alloys, including invar, thermostan, constantan, manganin, and series of Ag/Mn, Au/Mn, and Cu/Cr alloys, are presented graphically. The greatest pressure coefficient noted was 3.7×10^{-6} per atmosphere for a Ag/Mn alloy with about 15 per cent (by weight) of Mn. The results indicate a relation between the two coefficients for a series of alloys, the pressure coefficient decreasing with increasing temperature coefficient.

533.5 3174
Scientific Foundations of Vacuum Technique—[Book Review]—S. Dushman. Publishers: J. Wiley and Sons, New York, 1949, 882 pp., $15.00. (*Rev. Sci. Instr.*, vol. 20, p. 453; June, 1949.) The book covers the applications and fundamentals of vacuum technology in the fields of physics, chemistry, and metallurgy. The completeness of the work recommends it as a reference book.

MATHEMATICS

517.512:621.3.015.3 3175
Contribution to the Study of Transient Phenomena by Means of Time Series—M. Cuénod. (*Bull. Tech. Suisse Romande*, vol. 75, pp. 201–209; July 30, 1949.) The practical advantages of this method of calculation are considered. The method is outlined and applied to the determination of the response curve of a system and to integration, differentiation, the solution of linear differential equations and the determination of the conditions of stability of an automatic regulator. The relation between operational calculus and time-series methods is also indicated.

681.142 3176
Electronic Techniques Applied to Analogue Methods of Computation—G. D. McCann, C. H. Wilts, and B. N. Locanthi. (Proc. I.R.E., vol. 37, pp. 954–961; August, 1949.) The electronic devices and principles developed for the California Institute of Technology general-purpose, large-scale computer are described. This computer can be used for solving algebraic, ordinary differential, or partial differential equations, both linear and nonlinear.

681.142 3177
Principles and Progress in the Construction of High-Speed Digital Computers—A. D. Booth and K. H. V. Britten. (*Quart. Jour. Mech. Appl. Math.*, vol. 2, pp. 182–197; June, 1949.) Consideration of: (a) the basic principles underlying the mathematical design of high-speed digital computers, (b) the necessary components of such machines, (c) scale of notation, the form of the "memory," the action of the control, and other practical details, (d) the exact arithmetic functions of which these machines must be capable, (e) current computer projects in America, including Aiken's second relay computer at Harvard, the Bell relay machine, E.D.V.A.C., and the Princeton electronic computer, with reference to their state of completion in 1947.

681.142 3178
A Digital Computer for Scientific Applications—C. F. West and J. E. DeTurk. (Proc. I.R.E., vol. 37, p. 861; August, 1949.) Correction to 1107 of May.

681.142 3179
A Magnetic Digital Storage System—A. D. Booth. (*Electronic Eng.* (London), vol. 21, pp. 234–238; July, 1949.) The storage device consists of a cylindrical drum coated with magnetic material and rotating under a series of read/record heads arranged along a generator of the cylinder. Numbers are recorded in sequence as the drum rotates; to distinguish between them an extra track is added which contains a set of equally spaced positive "clock" pulses. The start of the clock pulse track is indicated by leaving a small gap free from pulses and using this as the zero position from which the position of any number can be obtained. Circuit and practical details are given.

681.142:621.392.5 3180
Mercury Delay Line Memory Using a Pulse Rate of Several Megacycles [per Second].—Auerbach, Eckert, Shaw, and Sheppard. (*See* 3082.)

517.564.4 3181
Spherical Harmonics [Book Review]—T. M. MacRobert. Publisher: Dover Publications, New York, 2nd edn 1948, 367 pp., $4.50. (Proc. I.R.E., vol. 37, p. 785; July, 1949.) Fourier series and Bessel, Legendre, and hypergeometric functions are covered as well as spherical harmonics. There is insufficient explanatory material for the engineer, but the treatment is thorough and useful for the applied mathematician.

517.63 3182
An Introduction to the Laplace Transformation [Book Review]—J. C. Jaeger. Publishers: Methuen and Co., London, 132 pp., 7s.6d. (*Wireless Eng.*, vol. 26, p. 276; August, 1949.) "The book contains the substance of a course of lectures delivered to engineers and physicists at the National Standards Laboratory, Sydney, in 1944 . . . [It] contains as little theory as possible; it is, in fact, largely a collection of worked examples illustrating the methods of solution of the various types of problem commonly arising in circuit theory."

MEASUREMENTS AND TEST GEAR

531.764.5:621.396.615.18 3183
A Compact Piezoelectric Chronometer—J. E. Benson and E. M. Dash. (*Proc. I.R.E.* (Australia), vol. 9, pp+ 4–8; August, 1948. Discussion, p. 8.) See 1113 of May.

621.317+083.7 3184
Radio Telemetering—G. L. Hinckley. (*Electronic Eng.*, vol. 21. pp. 209–211, 223; June, 1949.) Factors influencing the choice of system.

621.317.3 3185
Measurement of Impedance, Capacitance, Inductance and Frequency by the Method of Proportional Currents—A. I. Fürstenberg. (*Compt. Rend. Acad. Sci.* (URSS), vol. 51, pp. 277–280; February 10, 1946. In English.) A method consisting essentially in equalizing the potential drop across constant nonreactive resistances in series with the impedances. A source of constant voltage and frequency is required.

621.317.3:621.385.38 3186
The Deionization Time of Thyratrons: A New Method of Measurement—H. de B. Knight. (*Proc. IEE* (London), part III, vol. 96, pp. 257–261; July, 1949.) A circuit providing two firing pulses at an adjustable interval is used to measure the time required to reestablish grid control. Typical deionization times and grid-current decay curves are given for various tube structures and fillings. See also 2553 of October (Birnbaum.)

621.317.312:621.314.632 3187
Use of Copper-Oxide Rectifiers for Measuring the Smallest Alternating Voltages—H. Ifland. (*Funk. und Ton.*, vol. 3, pp. 449–454; August, 1949.) Temperature effects in the bridge type of rectifier normally used with moving-coil instruments for ac measurements can be partially compensated by connecting a resistor of suitable value in series with the instrument. Circuit details are given of a multirange meter with full-scale readings from 0.1V to 300V.

621.317.33:621.396.611.33 3188
Simplified Measurement of L and k—V. A. Sheridan. (*Electronics*, vol. 22, pp. 146, 154; August, 1949.) The coupling coefficient k is determined from the change in effective inductance of one winding of a pair of inductively coupled circuits when the other winding is first open-circuited and then short-circuited. Accuracies within 1 per cent are obtained with the bridge described, which is fed through a double-tuned transformer from a 23-kc oscillator and is suitable for measurements on most rf transformers.

621.317.333.4:621.315.23 3189
Cable Fault Finder—F. E. Planer. (*Elec. Rev.* (London), vol. 145, pp. 57–58; July 8, 1949.) Short description of portable inductive test equipment. A conductor carrying 1 mA ac can be detected at a distance of 45 ft. A direct indication of cable depth is tiven.

621.317.335.3† 3190
Construction of Apparatus for Very Accurate Measurement of the Dielectric Constant of Liquids—Mouradoff-Fouquet. (*Ann. Phys.* (Paris), vol. 4, pp. 310–367; May and June, 1949.) A double-beat method using oscillations of medium wavelength, with capacitors of extremely accurate mechanical construction for containing the liquids and with arrangements for maintaining the temperature at any desired value, enabled the dielectric constant of various organic liquids to be measured to about 1 part in 20,000.

621.317.335.3†:546.212-16 3191
The Electrical Behaviour of Ice—F. X. Eder. (*Ann. Phys.* (Lpz.), vol. 1, pp. 381–398; October 1, 1947.) *See* 1077 of 1948

621.317.336:621.317.372 3192
The Development of Q-Meter Methods of Impedance Measurement—A. J. Biggs and J. E. Houldin. (*Proc. IEE* (London), part III, vol. 96, pp. 295–302; July, 1949. Discussion, pp. 303–305.) Three usual definitions of Q are shown to be equivalent for a system at simple resonance. Circuit magnification factor is discussed in relation to Q, and circuits for measuring Q are critically examined. A new high-impedance injection meter is described and illustrated, with details of precautions necessary for hf measurements.

621.317.336.1:621.392.52 3193
Measurements on Intermediate-Frequency Transformers—E. Stern. (*Jour. Brit. I.R.E.* vol. 9, pp. 157–166; April, 1949.) A method is described for testing double-tuned transformers with a Q-meter or any other rf resistance meter. The resonance transfer impedance of two coupled circuits is expressed in a form independent of the nature of the coupling reactance. Charts are provided from which the transfer impedance of a transformer can be determined from three Q-meter readings. These readings can also be used to determine the frequency response curve from published generalized response curves if the tuning capacitances are known. Simple formulas for the bandwidth at —60 db of composite systems containing single- and double-tuned circuits of different dynamic resistances and coupling are also

given. Reprinted from *Proc. I R.E.* (Australia), vol 9, pp 4 11; January, 1948

621.317.361:621.385.832 3194

A Cathode-Ray Tube Frequency Comparator for 1 kc/s Sub-Standard Tones—J. F. M. Laver. (*P. O. Elec. Eng. Jour.*, vol. 42, pp. 61–64; July, 1949.) The nominal 1-kc tone transmitted by land line from a distant source is applied to the X plates of a cr tube which is modulated in brilliancy by means of a 100-kc frequency standard. The movement of the resultant dot pattern is used to compare the two frequencies rapidly. The method is more reliable than a heterodyne method in the presence of noise or interference voltages. Applications, sources of error, and accuracy are discussed.

621.317.382.029.64 3195

Broad-Band Power-Measuring Methods at Microwave Frequencies—L. E. Norton. (Proc. I.R.E., vol. 37, pp. 759–766; July, 1949.) The first method uses the forces due to the em fields in a transmission system to cause displacements of a diaphragm which are proportional to the square of the actuating field. In the second method, thin films are inserted in a transmission system so as to cause only small discontinuities. The small fraction of the power dissipated in the film raises its temperature and changes its resistance, which is measured. In both methods the output of the indicator system is proportional to the power within ±1 db between 1,000 and 10,000 Mc.

621.317.66 3196

Measurement of Microwave-Transmission Efficiency—A. L. Cullen. (*Wireless Eng.*, vol. 26, pp. 255 257; August, 1949.) The transmission efficiency of any transmission device is defined as the ratio (power out)/(power in) when the device is inserted in an otherwise matched transmission system. If the reflection coefficient for waves incident on the normal output end of the device, with the input end closed by a movable short-circuiting plunger, is plotted in the complex plane for several positions of the plunger, the points obtained will lie on a circle of radius equal to the transmission efficiency.

621.317.66:621.39 3197

Measurement of Telecommunications Efficiency—(*Proc. IEE* (London), part III, vol. 96, pp. 277 278; July, 1949.) Report of an IEE discussion meeting. No satisfactory objective test of quality is as yet available; existing subjective tests of loudness and intelligibility are compared.

621.317.7.029.63 3198

Measurement Apparatus for Decimetre Waves—H. H. Meinke. (*Fernmeldetech. Z.*, vol. 2, pp. 197–200, July, 1949.) Illustrations and short general description of (a) equipment comprising supply unit, transmitter, circular transmission line for impedance measurement (see 3203 below), receiver and cro indicator, (b) capacitive voltage divider, (c) diode, (d) transmission line of length variable as in a trombone, (e) reactive transmission line of characteristic impedance 70Ω, and (f) bolometer for power measuremetn in the range 10^{-2}–10^{-6} W.

621.317.715 3199

The Alternating Current Galvanometer—J. M. W. Milatz, P. M. Endt, C. T. J. Alkemade, and J. T. Olink. (*Physica*, *'s Grav.*, vol. 14, pp. 260–268; May, 1948. In English.) Theory is given which takes into account the induction current caused by the vibration of the moving system. The galvanometer can be made aperiodic and "field independent" with a combination of ac and dc magnetic fields, or with another damping device replacing the dc field. Measurements confirming the theory are described.

621.317.715 3200

On the Limit of Sensitivity of Galvanometers—M. Surdin. (*Jour. Phys. Radium*, vol. 10, pp. 253 254, July and September, 1949.) A formula is established giving the spectral intensity of the brownian couple which acts on a mechanical or an electrical system satisfying a linear differential equation of the second order and having a single degree of freedom. It is deduced that the sensitivity limit of a galvanometer remains the same whether the measurement circuit is open or closed.

621.317.715 3201

Valve Galvanometer with Ordinary Valves—J. Krenzer. (*Z. Phys.*, vol. 125, pp. 707–714; March 15, 1949.) Special electrometer tubes are normally used in apparatus for the measurement of very small currents such as that given by a photocell, but with ordinary tubes currents as low as 10^{-13}A can be measured and a limit of 10^{-14}A may be reached with selected tubes. Details of practical equipment using a KF4 pentode, which has a well-insulated grid, are given.

621.317.73:549.514.51 3202

The Measurement of the Series-Resonant Resistance of a Quartz Crystal—I. A. Rosenthal and T. A. Peterson, Jr. (*Rev. Sci. Instr.*, vol. 20, pp. 426–429; June, 1949.) Two methods of measuring the equivalent resistance of commercial plated or pressure mounted quartz crystals in the frequency range 80 kc to 100 Mc are discussed, namely (a) the substitution method, in which a resistor replaces the crystal unit, the amplitude of oscillation being maintained constant, and (b) the calculation method in which the quantities actually measured are the voltage drop across the crystal and the current through the unit. Three instruments are described in which combinations of these methods are used. Typical results are discussed.

621.317.73.029.63 3203

A Measurement Line with Visual Indicator—H. H. Meinke. (*Fernmeldetech. Z.*, vol. 2, pp. 233 241; August, 1949.) A device for measuring impedance in the dm-λ range. It consists essentially of a transmission line forming nearly a complete circle, with a motor driven rotary radial arm making contact with both the inner and outer conductor. The actual instrument is accurately machined and the end of the rotary arm makes contact with the inner conductor through a slot on the inner face of the line. A generator is connected to one end of the line, which is terminated by the impedance to be measured. Capacitive couplings to the two contacts on the rotary arm enable the shape of the voltage wave along the line to be displayed on a cro. The voltage picked up by the contacts is amplified, heterodyned to give a difference frequency of 3 Mc and, after further amplification and rectification, applied to the cro. Typical oscillograms are given, the method of calibration is described and also applications to phase measurement and to various measurements on receivers and transmitters. The apparatus is one unit of the equipment noted in 3198 above.

621.317.772 3204

Resistive Phase Shifters—J. E. Bryden (*Electronic Eng.* (London), vol. 21, pp. 322–326; September, 1949.) Description of a phase-shifter for which power is supplied from an electronic single-phase/6-phase conversion unit. Advantages, possible errors and their elimination, and applications are discussed.

621.317.78.029.64 3205

Broadband Bolometric Measurement of Microwave Power—H. J. Carlin. (*Radio and Telev. News, Radio Electronic Eng. Supplement*, vol. 13, pp. 16–19; July, 1949.) Theory and description of broad band units for the ranges 20 to 1,000 Mc, 1,000 to 4,000 Mc, and 4,000 to 10,000 Mc. One type, using Wollaston wire, is suitable for low-power measurements from 25μW to 1 mW, with extension to 10 mW if an attenuator is used. Another type uses metal film for powers from 1 mW to 50 mW, with extension to 5 W. Typical curves for voltage swr are given.

621.319.4.001.4 3206

Direct Voltage Performance Test for Capacitor Paper—H. A. Sauer and D. A. McLean (Proc. I.R.E., vol. 37, pp. 927 931; August, 1949.) Discussion of a testing procedure requiring about a day for preparation of samples and about another day for the actual life test. See also 965 of 1948 (McLean)

621.395.623.7.089.6 3207

Physical Measurements of Loudspeaker Performance—P. S. Veneklasen. (*Jour. Soc. Mot. Pic. Eng.*, vol. 52, pp. 641 656, June, 1949.) Facilities for outdoor calibration of loudspeakers and microphones at the Altec Lansing Corporation, California, are described. Techniques are illustrated by measurements of frequency response, angular distribution, and distortion for a typical loudspeaker. Methods for uniform presentation of performance data and specifications are also suggested. Smooth and clean reproduction over a limited range of frequencies should be achieved before wide range reproduction will be worth while

621.396.615 3208

Beat-Frequency Oscillator for the Carrier-Frequency Range—H. Boucke and H. Lennartz (*Fernmeldetech. Z.*, vol. 2, pp 245 248, August, 1949.) Details of a 1944 model, Type SR200R, which is not limited to carrier frequencies, its frequency range being from 50 cps to 200 kc. Curves show the output voltages into 150-Ω and 600 Ω loads for the two ranges 50 to 15,000 cps and 10 to 200 kcs. Distortion curves are also given.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

538.63 3209

Focusing Properties and Separating Power of a Magnetic Field bounded by Parallel Planes—R. Vauthier. (*Compt. Rend. Acad. Sci.* (Paris), vol 229, pp. 181–183; July 18, 1949.) Such a magnetic field can be used in the construction of a mass spectrometer, the separating power being nearly equal to that of a magnetic sector.

539.16.08 3210

An Automatic Timer for Radioactivity Measurements—B. D. Corbett and A. J. Honour. (*Electronic Eng.* (London), vol. 21, pp. 341 345; September, 1949.) Circuit and construction details for a timer to be used in conjunction with a commercial electronic scaler developed by the Atomic Energy Research Establishment. Hundreds of counts, minutes, and seconds are displayed on three easily reset dial-type registers. The duration of operation may be preset in the range 1 second to 200 minutes, or alternatively the number of counts may be preset in the range 100 to 20,000. Units and tens of counts are displayed by means of neon lamps.

539.16.08 3211

Geiger Counter Tubes—H. Friedman. (Proc. I.R.E., vol. 37, pp 791–808; July, 1949.) Discussion of various types and their special characteristics and applications. 68 references are given

539.16.08 3212

Geiger-Müller Counters with External Cathode—D. Blanc and M. Schérer. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 2018–2020; June 22, 1949.) The performance of this type of counter is improved by connecting the graphite coatings on the ends of the tube to the axial wire, thus eliminating end effects The curve for counting rate as a function of the length of the central graphite cathode, for a given voltage, is then a straight line passing through the origin. The plateau is lengthened by 50 per cent or more compared with normal counters. Argon-alcohol and argon-methane counters were investigated, of lengths from 20 cm to 1 m and diameters from 0.9 cm to 3.5 cm. Improved results were obtained in all cases. Counters which were defective or had poor plateaux could be reconditioned by the graphite-coating technique.

539.16.08 3213
Point Counters and Counter Tubes for Surface Investigations in Metallography—J. Kramer. (*Z. Phys.*, vol. 125, pp. 739–756; March 15, 1949.) Many examples are given showing the wide scope of counter methods in such research.

539.16.08 3214
Experiments in the Possibility of Increasing the Efficiency of Gamma-Counters—II—Slätis (*Rev. Sci. Instr.*, vol. 20, pp. 353–354, May, 1949.) Abridged version of 2873 of November.

549.211:539.16.08 3215
Removal of Space-Charge in Diamond-Crystal Counters.—A. G. Chynoweth. (*Phys. Rev.*, vol. 76, p. 310; July 15, 1949.) The use of radiation from a Nernst filament, of wavelength 1–10 mμ, is quite satisfactory for removing the space charge which causes decay of counting rate.

620.179.16:534.23 3216
On Sound Transmission through Metal Plates in Liquids for Oblique Incidence of Plane Waves—Götz. (See 2998.)

621.396.615:621.317.083.7 3217
Transistor Oscillator for Telemetering—Lehan. (See 3095.)

621.38.001.8 3218
Electronic Equipment for the Production Engineer (*Jour. Brit. I.R.E.*, vol. 9, pp. 222–237; June, 1949.) Report of a discussion. Examples demonstrate the wide range of operations that, by the development of new or the adaptation of existing instruments, can be controlled or carried out, rapidly and reliably, by electronic means. Specialization is essential in so wide a field. Selected electronic control units are illustrated and described.

621.38.001.8 3219
The Fourth Manchester Electronics Exhibition (*Electronic Eng.* (London), vol. 21, pp. 336, 337; September, 1949.) Brief descriptions of various exhibits including evaporating and sputtering plant, a servodyne, a cro, a tube test panel, an electron microscope, an inductance bridge, a microsecond chronometer, etc.

621.38.001.8 3220
Electronic Equipment at the B. I. F. (*Electronic Eng.* (London), vol. 21, pp. 224–227; June, 1949.) Brief descriptions of various exhibits.

621.38.001.8:621 3221
Electronics in Heavy Engineering—W. Wilson (*Jour. Brit. I.R.E.*, vol. 9, pp. 278–305; August, 1949.) A comprehensive review including discussion of high-power rectifiers, industrial hf generators, amplification and control apparatus, servomechanisms, motor control, and applications of the cro.

621.384.6 3222
Beam Extraction for the Electron Centrifuge—K. Gund and H. Reich. (*Z. Phys.*, vol. 126, pp. 383–398; May 27, 1949.) Discussion of the problem for the betatron, synchrotron, and FM cyclotron, and description of a method permitting 70 per cent of the electrons to be extracted in the form of a narrow beam.

621.384.611.2† 3223
Hospital Synchrotron—J. H. Martin. (*Elec. Rev.* (London), vol. 145, pp. 277–279; August 12, 1949.) The first of two 30-MeV synchrotrons designed for the Medical Research Council has recently been installed at the Royal Cancer Hospital, London. Photographs and a few technical details of the equipment are given. The total weight is about 3 tons.

621.384.611.2† 3224
The Synchrotron Accelerator—its Potentialities as a Generator of X-Rays and Electrons of 10-50 MeV Energies for Medical Use—D. W. Fry. (*Brit. Jour. Radiol.*, vol. 22, pp. 462–472; August, 1949.) The principle of the electron synchrotron is considered and the main characteristics of a β synchrotron are illustrated by reference to the operation of a 14-MeV and a 30-MeV electron synchrotron. With the 30-MeV machine an output of 13 r per minute at 1 m has been obtained. The X-ray characteristics are described and the possibility of extracting the electron beam for clinical use is discussed. The main factor controlling the design of both betatrons and synchrotrons at present is the injection process. If the efficiency of injection could be increased, a major improvement would result in synchrotrons and betatrons designed for clinical work.

621.384.62† 3225
The Microwave Linear Electron Accelerator—G. R. Newbery. (*Brit. Jour. Radiol.*, vol. 22, pp. 473–496; August, 1949.) Various types are briefly described and a detailed account is given of traveling-wave, standing-wave, and multicavity accelerators. The theoretical and practical limitations of the performance of each type at 3,000 Mc are discussed. The design of linear accelerators, suitable for medical use at this frequency is considered; tentative designs for 5-MeV and 10-MeV machines are given.

621.385.832:[535+77 3226
Cathode-Ray-Tube Applications in Photography and Optics—C. Berkley and R. Feldt. (*Jour. Soc. Mot. Pic. Eng.*, vol. 53, pp. 64–85; July, 1949.) Reprint of article noted in 2596 of October.

621.397.3:539.211 3227
The Electron Scanner—an Image Method Using Secondary Electrons—J. te Gude. (*Funk. und Ton*, vol. 3, pp. 373–383; July, 1949.) Principles and construction of apparatus using an electron beam for scanning a surface to be examined, and obtaining an image by means of the secondary electrons emitted from the surface. Applications to the investigation of surface films, such as those of Cu_2O rectifiers, and to corrosion research are mentioned. See also 4111 of 1939 (Knoll and Theile) and 2509 and 3593 of 1941 (Knoll).

620.179.16:534.321.9 3228
Ultrasonics [Book Review]—Carlin. (*See* 3032.)

PROPAGATION OF WAVES

538.56:535.421 3229
On the Diffraction of a Plane Wave by a Semi-Infinite Conducting Sheet—C. W. Horton. (*Phys. Rev.*, vol. 75, p. 1263; April 15, 1949.) At radar frequencies the detecting unit is much smaller than the region in which Sommerfeld's diffraction formulas are invalid. Approximate expressions are here given for the electric vector of the diffraction field within the excluded regions, when the electric vector is parallel to the edge of the screen, with similar expressions for the magnetic vector when this is parallel to the diffracting edge. The symbols used are those defined by Baker and Copson (2907 of 1940).

538.566+621.396.11 3230
On the Propagation of Radio Waves around the Earth—H. Bremmer. (*Physica's Grav.*, vol. 14, pp. 301–318; June, 1948. In English.) A general discussion of existing theories and the extensions required by the discovery of new phenomena like superrefraction. For a fuller account see 3242 below.

538.566 3231
Reflection of Electromagnetic Waves at an Inhomogeneous Layer—W. Kofink. (*Ann. Phys.* (Lpz.), vol. 1, pp. 119–132; January 3, 1947.) Mathematical study of several methods of calculating the reflecting power of a layer in which the dielectric constant varies with depth in the layer. The essentials of the following methods are presented: (a) the functional method; (b) the method of van Cittert; (c) the method of Gans (WKB method); (d) a differentiation method. (c) and (d) are only applicable to layers of finite thickness.

538.566.3 3232
On the Theory of the Double Refraction of Electromagnetic Waves in an Ionized Gas under the Influence of a Constant Magnetic Field (Ionosphere)—H. Lassen. (*Ann. Phys.* (Lpz.), vol. 1, pp. 415–428; October 1, 1947.) The complex refractive index and the waveform are calculated in a simple manner. A relation is established between previous calculations by Försterling and the author and the formulas of Appleton, Goldstein, and Hartree.

621.396.11 3233
Ground-Wave Propagation across a Land/Sea Boundary—G. Millington and N. Elson. (*Nature* (London), vol. 164, pp. 114–116; July, 16, 1949.) An increase of field strength on crossing the coastline was observed by Millington for 3.13-Mc radiation along a path passing for about 100 km overland and then across the English Channel. The field strengths along this path are shown graphically; the results confirm the theory previously given (1758 of July). See also 2307 of September.

Measurements of the field strength of 1.122-Mc transmissions from Crowborough were made by Elson in an aircraft flying at a height of 1,000 ft across East Anglia and then over the North Sea. At the frequency used, both land and sea are essentially conducting. A major recovery of field strength was noted about 25 miles beyond the coastline. The results confirm Millington's theory. A small recovery effect was noted on crossing the Thames estuary.

621.396.11:551.510.535 3234
Correlation of Sporadic E Region Ionization over Short Distances and Comparison with Magnetic Disturbances—Gerard. (*See* 3135.)

621.396.11:551.510.535:518.3 3235
Nomograms for Ionosphere Control Points—J. C. W. Scott. (Proc. I.R.E., vol. 37, pp. 821–824; July, 1949.) An abac for determining the latitudes of the ionospheric control points, given the length of a radio circuit and the latitudes of its terminals. This simplifies the use of world muf charts.

621.396.11:621.396.812.3 3236
On the Fading of Short Waves—W. Menzel. (*Fernmeldetech. Z.*, vol. 2, pp. 243–244; August, 1949.) Discussion with special reference to Ratcliffe's theory (193 of February).

621.396.11:621.396.93 3237
Scattering of Radio Waves by Metal Wires and Sheets—F. Horner. (*Proc. IEE* (London), part III, vol. 96, pp. 333–340, July, 1949. Discussion, pp. 340–345.) Formulas for the scattered fields are derived, using transmission-line theory for wires and diffraction theory for sheets. Measurements of the scattered fields have been made at a frequency of 600 Mc, using a direction finder as the indicating instrument, the results are in substantial agreement with theory. For wires of the order of 1 mm in diameter and more than 5λ long, resonance effects at 600 Mc are small. Such effects are negligible in sheets whose dimension normal to the electric vector is greater than λ.

621.396.81: 3238
Reception at over 16000 km from the Transmitter—R. G. Sacasa. (*Rev. Telecommun.* (Madrid), vol. 3, pp. 11–15; June, 1948.)

Graphs show the variations during March, 1948, of receiver output power at Madrid for Australian broadcasting stations on wavelengths of 16.82, 19.74, and 25.49 m. Measurements were made daily at 0730 GMT at peak modulation. Comparison with measurements on the 32-m signals from the Arganda station, Madrid, only 17 km from the receiving station, revealed definite correlation with the variations of the Australian 25 49-m signals

621.396.812 3239
Anomalous Radar Propagation over Land in the Period November 29 to December 1, 1948—R. F. Jones. (*Met. Mag.*, vol. 78, pp. 233–234; August, 1949.) The abnormal ranges obtained during this period at a radar station near Dunstable for $\lambda = 10$ cm were associated with a rapid lapse of water-vapor content above fog. Radiosonde data do not indicate precisely the boundaries of the dry layers, because of the rate of ascent of the balloon and the time-lag in the humidity element.

621.396.812.029.62 3240
U.H.F. Propagation Characteristics—E. W. Allen, Jr. (*Electronics*, vol. 22, pp. 69–89; August, 1949.) From the results of 13 vhf surveys made by the National Bureau of Standards, correction factors have been determined for expected median field-strengths; these corrections are applicable to the FCC groundwave signal/range charts for frequencies from 63 to 195 Mc.

621.396.812.029.64 3241
Microwave Phase Front Measurements for Overwater Paths of 12 and 32 Miles—A. W. Straiton. (Proc. I.R.E., vol. 37, pp. 808–813; July, 1949.) Continuous curves of phase and signal strength at a wavelength of 3.2 cm are shown for a range of transmitter and receiver heights from 10 ft to 55 ft above mean sea level The results for the two paths are compared, and deviations from those commonly expected for oversea propagation are noted.

538.566+621.396.11 3242
Terrestrial Radio Waves [Book Review]—H. Bremmer. Publishers: Cleaver-Hume Press, London, 344 pp., 36 s. (*Wireless Eng.*, vol. 26, pp. 275–276; August, 1949.) The book is based on researches of the author originating in classic papers written in collaboration with van der Pol. It will appeal most to the mathematical physicist who can follow the general line of the analysis, but there are sections of direct use to the engineer. ". . . this book will be a mine of information to the few to whom will fall the task of tackling the outstanding problems of propagation theory."

RECEPTION

621.396.62.029.58 3243
The Orchestra in Your Home. The TR138.—R. Geffre. (*Toute la Radio*, vol. 16, pp. 243–248; September, 1949.) Complete circuit details of a high-fidelity sw receiver with ample sensitivity for good reproduction in France of transmissions from the U.S.A. Special features of the various stages are discussed.

621.396.621 3244
A High-Performance Dual-Conversion Superhet—R. C. Cheek. (*CQ*, vol. 5, pp. 16–23, 77; July, 1949.) Complete details of a receiver which operates directly from the antenna on 3.5 and 7 Mc, but which is preceded by an hf converter when operating on 14, 21, 27, or 28 Mc. Alignment procedure is described.

621.396.621 3245
Philips Model 681A—(*Wireless World*, vol. 55, pp. 289–290; August, 1949.) Test report. Normal tuning is provided in a superheterodyne circuit for wavelength ranges 11.1–34.2 m, 34.2–110.5 m, 192–560 m, and 900–2,000 m. There are also 8 selected sw broadcast bands of width about 0.5 Mc centered at wavelengths 11, 13, 16, 19, 25, 31, 41, and 49 m, for which a double superheterodyne principle is used in the bandspread circuits, so that the local oscillator on each band works at a fixed frequency and is thus easier to stabilize.

621.396.621:621.396.619.13 3246
The Demodulation of a Frequency-Modulated Carrier and Random Noise by a Discriminator—N. M. Blachman. (Proc. I.R.E., vol 37, p. 895; August, 1949.) Summary only. See also 1772 of July.

621.396.621:621.396.65.029.58 3247
The Receiving System at Cooling [Kent] Radio Station—C. F. Booth. (*P. O. Elec. Eng Jour.*, vol. 42, pp. 84–89; July, 1949.) The factors limiting the performance of long distance R/T links operating in the frequency range 3 to 30 Mc are outlined with particular reference to the downcoming angle at the receiver. The receiver system described uses a highly directive steerable antenna arranged to feed four parallel receiver branches, one of which gives an energy/downcoming-angle diagram on a cr tube from which the other three are manually set to three different optimum angles. Performance is compared with that for a receiver having a single antenna of possible future developments of directive receiving systems are considered. The system is similar to the M.U.S.A. system noted in 3016 of 1940 (Polkinghorn).

621.396.823 3248
Car-Ignition Interference—W. Nethercot. (*Wireless Eng.*, vol. 26, pp. 251–255; August, 1949.) The wide-band continuous radiation from the ignition circuit is due to traveling waves set up in the hv cables when the distributor and sparking-plug gaps break down. The current through the sparking-plug gap consists of a series of very steep-fronted steps, the intervals between which are determined by the time the waves take to travel twice the length of the hv cables. The envelope of these current steps is oscillatory and its frequency lies between 30 and 50 Mc. Resistors at the sparking-plug and distributor terminals should give suppression over the whole frequency band.

621.396.828 3249
Goniometer Arrangement for the Suppression of Interfering Transmissions by means of Angle Measurement with Beam-Aerial Systems—H. Fricke. (*Fernmeldetech. Z.*, vol. 2, pp. 249–253; August, 1949.) Two antennas, whose beams are directed towards the wanted station, are connected to the two stator coils of the goniometer, a phase-shifter being interposed between antenna and coil in one case. The receiver is connected to the goniometer search coil. With suitable adjustment of the phase-shifter and of the position of the search coil, signals from an unwanted transmitter operating on the same wavelength as that of the wanted station can be completely eliminated

STATIONS AND COMMUNICATION SYSTEMS

621.39:061.053 3250
The Third Session of the Administrative Council of the International Telecommunications Union (I.T.U.) [Geneva, 1948]—G Gneme. (*Poste e Telecomun.*, vol. 16, Supplement, pp. 26–34; November, 1948.)

621.395.44:621.396.619.2 3251
A 48-Channel Carrier Telephone System: Part 2—Mechanical Construction—G. H. Bast, D. Goedhart, and J. F. Schouten. (*Philips Rech. Rev.*, vol. 10, pp. 353–358; June, 1949.) Part 1: 236 of 1948.

621.396 3252
Modern Tendencies in Commercial Long-Distance Radio Communications—A. Niutta. (*Poste e Telecomun.*, vol. 16, pp. 241–251; June and July, 1948.) A review covering single-sideband transmission, frequency-shift keying, and multiplex teleprinter systems

621.396:061.3 3253
The Fifth Meeting of the C.C.I.R. [Stockholm, July 1948]—T. Gorio. (*Poste e Telecomun.*, vol. 16, pp 493–502; December, 1948.)

621.396.1 3254
The European Broadcasting Conference at Copenhagen [1948]—G. Gneme. (*Poste e Telecomun.*, vol. 16, Supplement, pp. 1–16; November, 1948.) Detailed report, with special consideration of the position of Italy

621.396.1 3255
The Copenhagen Maritime Regional Radiocommunication Conference [1948]—G. Gneme (*Poste e Telecomun.*, vol. 16, Supplement, pp 17–25; November, 1948.) Full details on allocations to coastal stations and to ships

621.396.61.029.54:621.396.712 3256
B.B.C. Transmitting Station at Postwick Grange—(*Engineer* (London), vol. 188, p. 77, July 15, 1949.) Further details of the new station near Norwich. The antenna system consists of two 126-ft tubular masts spaced $\lambda/4$ apart. The easterly mast is energized; the resulting directional system brings Yarmouth within the service area. See also 2942 of November.

621.396.619.11/.13 3257
F.M. vs A.M.—D. J. Braak. (*Electronics*, vol. 22, pp. 218, 220; August, 1949) Comment on some of the statements made in the paper abstracted in 1504 of June (Toth.)

621.396.619.16:621.396.41 3258
A Time-Division Multiplexing System—W. P. Boothroyd and E. M. Creamer, Jr (*Elec. Eng.*, vol. 68, pp. 583–588; July, 1949.) A system using pulse-amplitude modulation with a filtering arrangement for minimizing the required transmission bandwidth.

621.396.65.029.58:621.396.621 3259
The Receiving System at Cooling [Kent] Radio Station—Booth. (*See* 3247.)

621.396.931 3260
Portable F.M. Equipment—H. V. Carlson. (*FM-TV*, vol. 9, pp. 14–16; July, 1949.) Description of a unit weighing less than 10 lb, which operates at a fixed frequency in the ranges 25 to 50 Mc or 152 to 165 Mc and provides 2-way communication over distances of several miles under noisy conditions.

621.396.931 3261
A 28-Mc/s Installation for. the Car.—G. P McGinnis. (*QST*, vol. 33, pp 11–16; August, 1949.) Construction and installation details for amateur equipment which does no damage to the car and only requires an input of 17 W.

621.396.932 3262
Automatic Station Call Selector—W. W McGoffin and R. H. Schulz. (*Electronics*, vol. 22, pp. 75–77; August, 1949.) An instrument which sounds an alarm at a radio station when its own call letters are received at any sending speed from 6 to 34 words per minute.

SUBSIDIARY APPARATUS

621.314.58 3263
Thyratron Replaces Vibrator—(*Electronics*, vol. 22, pp. 140, 144; August, 1949.) Description, with circuit diagram, of a simple dc/ac converter which operates from a 6-V battery and has no moving parts.

621.316.722:621.396.682 3264
Pre-Calculation of Magnetic Voltage Stabilizers—W. Taeger. (*Funk. und Ton.*, vol. 3, pp. 429–437; August, 1949.) Design procedure, with numerical calculations for an output power of 75 W at 220V. With mains voltage

variations from 160V to 260V, the stabilized voltage only varied from 215V to 224V, in good agreement with theory.

621.316.726 3265
Frequency Correction of Electric Signalling Power Supplies—E. Friedlander and R. A. Duncan. (*GEC Jour.*, vol. 16, pp. 130–137; July, 1949.) Detailed description of the equipment noted in 2054 of August, with particular reference to special features such as the frequency relay and protective devices for tripping in case of hunting or failure to correct the frequency. The frequency relay is based on the principle of phase change in a resonant circuit; its construction and operation are clearly explained.

621.396.68:539.16.08 3266
Miniature Counter-Tube Power Supply—D. L. Collins. (*Electronics*, vol. 22, pp. 170–173; August, 1949.) The high voltage is obtained across a miniature transformer in a blocking-oscillator circuit using a 1V5 tube. The oscillator pulses are rectified by a low-power hv half-wave VX-21 rectifier. Regulation is obtained by means of a 900-V corona voltage regulator.

621.396.682 3267
30-kV D.C. Regulated Power Supply—W. Spellman. (*Radio and Telev. News, Radio-Electronic Eng. Supplement*, vol. 12, pp. 16–17, May 30, 1949.) Circuit diagram, without component values, of a supply unit giving an output from 25 to 30 kV; regulation is within 1 per cent under load variations from zero to 1 mA and line-voltage variations from 95 to 125V.

621.396.682:621.316.722.1 3268
An Electromechanically Stabilised Mains Supply Unit—A. E. Maine. (*Electronic Eng.* (London), vol. 21, pp. 319–321; September, 1949.) A neon-tube bridge and a voltage-sensing circuit with gas triodes are used to control a bidirectional motor which adjusts the tappings on a variable-ratio transformer, thus correcting any deviation from the required voltage. The output voltage is regulated within 1 per cent, for loads up to 1.5 kVA and for voltage deviations of −15 per cent to +5 per cent.

621.314.632 3269
Metal Rectifiers [Book Review]—H. K. Henisch. Publishers: Oxford University Press, 168 pp., 15 s. (*Electronic Eng.* (London), vol. 21, p. 229; June, 1949.) The book will be useful both to students as an introduction to electronic properties of solids and to practical users of dry rectifiers who require a critical but compact account of the subject. The book contains no difficult mathematics.

TELEVISION AND PHOTOTELEGRAPHY

621.397.331.2 3270
Slow-Electron Television Cameras—J. J. M. Moral. (*Rev. Telecomun.* (Madrid), vol. 3, pp. 38–52; June, 1948.) Operating principles and characteristics of the iconoscope, orthicon, and isoscope tubes.

621.397.331.2 3271
Distortion of Scanning Waveforms—G. G. Gouriet. (*Electronic Eng.* (London), vol. 21, pp. 327–331; September, 1949.) Requirements for a linear scan are discussed. The distortion due to insufficient bass response and means of correcting this distortion are also considered.

621.397.5 3272
Televising Moving Images—R. W. Hallows. (*Wireless World*, vol. 55, pp. 291–293; August, 1949.) Calculations of balanced definition for moving images should not be based entirely on data for still images, since moving images introduce many new problems. The greatest immediate need is for developing methods of producing wide-band transmittng and receiving apparatus at reasonable cost

621.397.5:535.88 3273
Three-Dimensional Picture Screens for Television and the Cinema—E. G. Beard. (Proc. I.R.E. (Australia), vol. 9, pp. 4–16; June, 1948). Previous attempts to produce 3-dimensional pictures on a flat screen are discussed and a practical screen is described. Manufacturing tolerances and the modifications necessary to adapt this screen for use in cinema theaters are considered. An experimental screen gave promising results.

621.397.5:535.88 3274
A Projection System for Domestic Television Receivers—(*Electronic Eng.* (London), vol. 21, pp. 314–318; September, 1949.) Discussion of a Mullard folded-Schmidt system giving adequate resolution for picture sizes up to 15 in. by 12 in., which can be used in conjunction with most existing receiver chasses. See also 2387 of 1948 (Rinia et al.)

621.397.5:535.88 3275
Home Projection Television: Parts 1–3—H. Rinia, J. de Gier, P. M. van Alphen, G. J. Siezen, J. Haantjes, and F. Kerkhof. (Proc. I.R.E. (Australia), vol. 9, pp. 9–18; August, 1948.) Abridged version of 2387 of 1948.

621.397.5:617 3276
Television as an Aid to Teaching Operative Surgery—(*Electronic Eng.* (London), vol. 21, pp. 212–213; June, 1949.) Short description of specially designed equipment installed in Guy's Hospital, London, which enables the progress of operations to be followed by many observers without crowding the operating theater. At present four viewing sets with 15-in. screens are used.

621.397.6:621.385.832 3277
High-Power Cathode-Ray Tubes—Moss. (*See* 3308.)

621.397.6:621.395.625 3278
Television Recording: Simplified System—D. A. Smith. (*Wireless World*, vol. 55, pp. 305–306; August, 1949.) An economical system using a television receiver and a 16-mm film projector modified for use as a camera and sound recorder. Of every three television frames, two consecutive ones are photographed and the third is missed. A recording lamp for the sound signal is fed from the receiver output tube. For reproduction a standard projector with a 3-bladed shutter and running at 16 frames per second can be used.

621.397.6:621.396.615.17:621.317.755 3279
Television Time Base Linearisation—Keen. (*See* 3098.)

621.397.62 3280
Television Receiver with Rimlock Valves and Automatic Frequency and Phase Control—F. Juster. (*Radio Prof.* (Paris), vol. 18, pp. 11–15; June, 1949.) Complete circuit and component details for a receiver with an MW31-7 cr tube.

621.397.62 3281
Designing a TRF [tuned radio frequency] Television Receiver—W. H. Buchsbaum. (*Tele-Tech*, vol. 8, pp. 36–39; August, 1949.) The 15-tube receiver uses three metal rectifiers "B" supply and operates from ac or dc. A 2-tube rf amplifier and a four-stage video amplifier give high gain and performance nearly equal to those of superheterodyne receivers. Advantages and limitations of this type of receiver are discussed.

621.397.62 3282
Transit-Time Effects in Television Front-End Design—Watts. (*See* 3113.)

621.397.645 3283
Television Stabilizing Amplifier—Schultz. (*See* 3114.)

621.397.7 3284
The Television Studio—D. C. Birkinshaw. (*BBC Quart.*, vol. 4, pp. 105–117; July, 1949.) An account of studio technique used at the London television station, Alexandra Palace.

621.397.828 3285
Television Interference Suppression—"Spenny." (*RSGB Bull.*, vol. 25, p. 44; August, 1949.) Suppression of a 90-Mc harmonic from an amateur transmitter was effected by shunting the output by means of a $\lambda/2$ line.

621.397.828 3286
TVI Reduction—Western Style—C. E. Murdock. (*QST*, vol. 33, pp. 24–27, 82; August, 1949.) Interference caused by the harmonics of a 1-kW amateur transmitter 40 ft away from a television receiver was reduced by using high-capacitance tank circuits in the anode of the driver tube and single-turn coaxial pickup loops and high-capacitance grid circuits in the final amplifier. Antenna tuners were also built for each amateur band. Methods of suppressing interference due to line voltage fluctuations and keying clicks are also discussed.

621.397.828 3287
The H.R.O. and T.V.I.—R. L. Varney. (*RSGB Bull.*, vol. 25, pp. 41–42; August, 1949.) A strong third harmonic at the television frequency of 45 Mc is produced by the first heterodyne oscillator of the HRO communications receiver when operating in the 14-Mc band. This is suppressed by connecting a series resonant circuit between the oscillator tube cathode and earth.

TRANSMISSION

621.396.61 3288
A New 150-kW A.M. Transmitter—T. J. Boerner. (*Broadcast News*, no. 55, pp. 42–49; June, 1949.) An efficient compact transmitter for the frequency range 540 to 1,600 kc, using class-B modulation of a class-C final amplifier. Details of design, layout, and installation are given and the results of performance tests are shown graphically and tabulated.

621.396.61 3289
The Types TGM.651 and TGZ. 651 Transmitters—W. J. Morcom. (*Marconi Rev.*, vol. 12, pp. 104–107; July and September, 1949.) Type TGM.651 is a 3-kW mf transmitter, and Type TGZ.651 a 3-kW mf and lf transmitter. These complete the series of which other members were noted in 553 of March (Cooper) and 1219 of May.

621.396.61:621.392.52 3290
A Filter Design for the Single-Sideband Transmitter—F. M. Berry. (*QST*, vol. 33, pp. 29–35; June, 1949.) A highly selective first if bandpass filter. Sharp cut-off is restricted to the hf side of the passband which extends from 17 to 20 kc. Basic design formulas are given for the filter which consists of two combined *m*-derived π sections. Modifications permit operation directly from a ring modulator into the grids of a balanced tube modulator. The filter can be aligned with the minimum of special equipment.

621.396.61:621.396.8 3291
Operation of A.M. Broadcast Transmitters into Sharply Tuned Antenna Systems—W. H. Doherty. (Proc. I.R.E., vol. 37, pp. 729–734; July, 1949.) Investigation of the clipping of sidebands and distortion of the voltage envelope at high modulation frequencies. The effects may be reduced by suitable coupling methods.

621.396.61:621.396.97 3292
The New 100-kW [broadcasting] Transmitter at Naples—S. Bertolotti. (*Poste e Telecomun.*, vol. 17, pp. 44–46; January, 1949.) Distortion at 100 per cent modulation is within 2 per cent, the frequency curve is linear to within 2 db up to 10,000 cps, and background noise is at least 60 db down. Over-all efficiency is 39 per cent

621.396.619.23 3293
A High Voltage Ring Modulator—M. J. Tucker. (*Electronic Eng.* (London), vol. 21, pp. 239–242; July, 1949.) A form of biased ring modulator using diodes provides a satisfactory precision "phase-conscious" rectifier for use at voltages high enough to enable the output to be applied directly to the grid of an output stage. Practical circuits are discussed. See also 3542 and 3543 of 1948 (D. G. Tucker).

TUBES AND THERMIONICS

537.291 3294
Control of a Beam of Electrons by an Intersecting Electron Beam—J. L. H. Jonker and A. J. W. M. Van Overbeek. (*Nature* (London), vol. 164, pp. 276–277; August 13, 1949.) Two electrode systems are arranged to produce two mutually perpendicular electron beams. The voltage applied to grid 1 determines the current in beam 1; this in turn controls the current in beam 2. A graph of the voltage of grid 1 against current to anode 2 is shown. The slope of this curve is more than twice that obtainable with normal direct grid control. It is suggested that replacing anode 2 by a secondary-emission multiplier would result in a slope of several amperes per volt at currents of the order of 10^{-2} A.

537.291+538.691]:537.525.92 3295
Electron Flow in Curved Paths under Space-Charge Conditions—Meltzer. (*See* 3120).

621.383 3296
On a Method for the Production of Photo-Sensitive Layers of Very High Resistance with PbS as Infra-Red-Sensitive Semiconductor,—K. Frank and K. Raithel. (*Z. Phys.*, vol. 126, pp. 377–382; May 27, 1949.) The layers are produced by vaporization of PbO in an atmosphere of sulphur vapor at low pressure, with subsequent heat treatment. Resistance is of the order of 10^{11}–$10^{12}\Omega$ cm. The results of an investigation of such films by electron diffraction methods are discussed.

621.383:621.385.15 3297
Electron-Multiplier Tubes. Developments. Use—A. Lallemand. (*Jour. Phys. Radium*, vol. 10, pp. 235–239; July and September, 1949.) The fluctuation of the output current is determined for the ideal electron multiplier as a function of the number of stages and of the multiplication factor per stage. Other causes of current fluctuation are considered and a tube designed to eliminate such fluctuations as far as possible is described. The use of multiplier tubes with a very stable symmetrical amplifier, and also in a simple circuit including a neon lamp shunted by a capacitor, is discussed.

621.383:621.385.15:621.396.822 3298
On the Variation of the Background Noise of a Photomultiplier RCA 931A with the Potential of the Glass Envelope—C. Taylor. (*Jour. Phys. Radium*, vol. 10, pp. 255–256; July and September, 1949.) The background noise is multiplied by about 10 when the potential of the envelope differs from that of the photocathode by +500 or 1,000 V. A theory of the effect is proposed. For applications requiring low background noise, such as the counting of particles by scintillations, the envelope should be maintained at the potential of the photocathode.

621.383.4 3299
Lead Sulfide Photoconductive Cells—S. Pakswer. (*Electronics*, vol. 22, pp. 216, 218; August, 1949.) Correction to 2370 of September.

621.385 3300
The Electron Wave Tube—A. V. Haeff. (PROC. I.R.E., vol. 37, pp. 777–778; July, 1949.) Discussion on 1825 of July

621.385 3301
The Development of Radio Transmitting Valves—J. Bell and J. W. Davies. (*GEC Jour.*, vol. 16, pp. 138–149; July, 1949.) The limitations of early types are discussed. New construction and manufacturing techniques which have largely overcome these limitations are described. Typical tubes are illustrated and their ratings and performance are tabulated.

621.385.032.29 3302
The Calculation of the Electrode Temperatures in Radio Valves—S. Wagener and I. A. Harris. (*Jour. Brit. I.R.E.*, vol. 9, pp. 318–319; August, 1949.) Comment on 1232 of May (Harris) and the author's reply.

621.385.3:621.396.645.029.3 3303
A Low-Noise [audio] Input Tube—C. R. Knight and A. P. Haase. (*Radio and Telev News, Radio-Electronic Eng. Supplement*, vol. 12, pp. 15 18, 31; March, 1949.) Description of the double triode 12AY7, in which microphony effects and other noises are particularly low. A balanced amplifier, using these tubes in the first two stages and a 12AU7 in the output stage, with cross neutralization and inverse feedback, is also described and a detailed circuit diagram is given. The response curve is essentially flat from 30 cps to 20 kc.

621.385.38:621.317.3 3304
The Deionization Time of Thyratrons: A New Method of Measurement—Knight. (*See* 3186.)

621.385.832 3305
Projective Three-Dimensional Displays: Parts 1 and 2—D. M. MacKay. (*Electronic Eng.* (London), vol. 21, pp. 249–254 and 281 296; July and August, 1949.) A circuit is discussed which will perform two stages of rotation, one about the Y axis moving X to X′, and the second about the X′ axis. These rotations remove any structural ambiguities present in a given projection. Three methods of obtaining perspective convergence are given. A stereoscopic switching unit, and various methods for the relative measurement of projections are described. See also 577 of March (Parker and Wallis) and 1242 of May (Berkley)

621.385.832:535.371.07 3306
The Physics of Cathode Ray Tube Screens—Garlick. (*See* 3158.)

621.385.832:621.396.9 3307
Three-Dimensional Cathode-Ray Tube Displays—E. Parker and P. R. Wallis. (*Proc. IEE* (London), part III, vol. 96, pp. 291 294; July, 1949.) Discussion on 577 of March.

621.385.832:621.397.6 3308
High-Power Cathode-Ray Tubes—H. Moss. (*Wireless Eng.*, vol. 26, pp. 293–296; September, 1949.) A preliminary survey of the design of tubes with screen diameters up to 30 in. for direct viewing. The relation between the response of cascade screens and the beam voltage is uncertain and is of critical importance. Mechanical design difficulties are briefly discussed.

621.396.615.141.2 3309
The Cavity Magnetron—H. A. H. Boot and J. T. Randall. (*Proc. IEE* (London), part III, vol. 96, pp. 261–263; July, 1949.) Discussion on 890 of 1948.

621.396.822 3310
Valve Noise and Transit Time—C. J. Bakker. (*Wireless Eng.*, vol. 26, p. 277; August, 1949.) Comment on 2420 of 1948 (Campbell, Francis, and James). See also 255 of February, 583 of March (Houlding), and 3311 below.

621.396.822 3311
Measurement of Induced Grid Noise—F. L. H. M. Stumpers. (*Wireless Eng.*, vol. 26, pp. 277–278; August, 1949.) Discussion of recent experimental results which agree with those discussed by Bakker (3310 above and back references).

621.396.822 3312
Transit-Time Effects in U.H.F. Valves—J. Thomson. (*Wireless Eng.*, vol. 26, pp. 192 199; June, 1949, corrections ibid., vol. 26, p. 278; August, 1949.) Mathematical technique suitable for cases, such as total-emission damping where space-charge distortion of the es field can be neglected. See also 3313 below.

621.396.822 3313
Transit-Time Effects in U.H.F. Valves—R. E. Burgess and J. Thomson. (*Wireless Eng.*, vol. 26, p. 313; September, 1949.) Burgess suggests that Bakker and de Vries (3374 of 1935) covered much of the work noted in 3312 above. Thomson regards their work as valid only for a simple idealization.

MISCELLANEOUS

621.396 Popov 3314
Alexander S. Popov—G. W. O. H. (*Wireless Eng.*, vol. 26, pp. 249–250; August, 1949.) Reply to comment by Thornton (2113 of August) on 1842 of 1948.

652.6 3315
Directory of Translators—(*Jour. Franklin Inst.*, vol. 248, p. 104; July, 1949.) A directory of language specialists competent in various technical fields has been established by the Science-Technology Group of Special Libraries Association under the management of Mr. W. Kalenich, Librarian at the Southwestern Research Institute, San Antonio, Texas. About 3,000 technical translations already available in private files have also been listed.

621.396(031) 3316
The Radio [amateur's] Handbook (Le Manuel Radio.) French edition [Book Review]—Headquarters Staff of the American Radio Relay League. Publishers: P. H. Brans Antwerp, 1948, 350 pp., 240 fr. (Belgian). (*Alta Frequenza*, vol. 18, p. 88; April, 1949.) See also 2682 of October.

681.2 3317
The Instrument Manual [Book Review]—Publishers: United Trade Press, London, 548 pp., 70s. (*Electronic Eng.* (London), vol. 21, p. 310; August, 1949.) The book covers a very wide field. The majority of instruments and control gear described are mechanical, although electronic devices are not ignored. The text is in simple descriptive language, and many illustrative diagrams are given.

621.39 3318
Wheeler Monographs—A series of monographs appearing at two-monthly intervals, available on a subscription basis from Wheeler Laboratories, Inc., 122 Cutter Mill Road, Great Neck, N. Y. Single copies of each issue cost $25.00. All the monographs are by H. A. Wheeler except where otherwise stated. Titles of the first 11 monographs are:—1. Transmission Lines and Equivalent Networks. 2. Slide Rule Operations for Radio Problems. 3. A Simple Theory and Design Formulas for Superregenerative Receivers. 4. Geometric Relations in Circle Diagrams of Transmission-Line Impedance. 5. Generalized Transformer Concepts for Feedback Amplifiers and Filter Networks. 6. A Simple Theory of Powdered Iron at all Frequencies. 7. Superselectivity in a Superregenerative Receiver. 8. The Piston Attenuator in a Waveguide below Cut-off. 9. Measuring the Efficiency of a Superheterodyne Converter by the Input Impedance Circle Diagram, by H. A. Wheeler and D. Dettinger. 10. The Transmission Efficiency of Linear Networks and Frequency Changers. 11. The Maximum Speed of Amplification of a Wide-Band Amplifier

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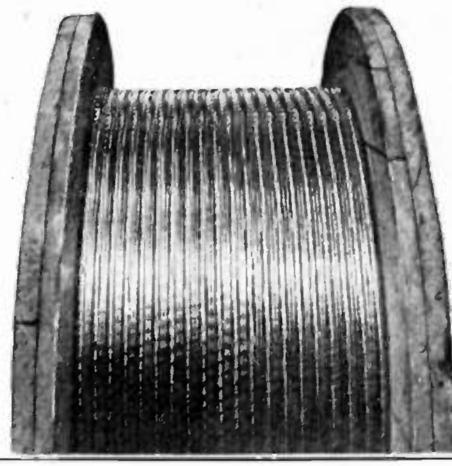

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"Television Pickup Tubes," by R. B. Janes, Radio Corporation of America; and Business Meeting; September 29, 1949.

Baltimore

"Speeding up the Determination of Dielectric Properties of Radio Frequencies," by C. F. Miller, Faculty of Johns Hopkins University; and Business Meeting; October 12, 1949.

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"Fundamental Particles of Modern Physics," by J. S. Ham, Jr., Graduate Student, University of Chicago; September 20, 1949.

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"Summary of Proposed Color Television Systems," by M. G. Nicholson, Colonial Radio Corporation; September 19, 1949.

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"High-Gain and Directional Antennas for Television Broadcasting," by L. J. Wolf, Engineering Products Division, RCA Victor Division; September 22, 1949.

"Performance of Broadcast Directional Antenna Systems," by J. S. Brown, Chief Engineer, Andrew Corporation; October 18, 1949.

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"Television Station Installation Problems," by Garo Ray, Radio Station WNHC; September 15, 1949.

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"Inspection Trip of Transmitting Facilities of KRLD-TV," by J. F. Kluttz and Roy Flynn, Radio Station KRLD; October 6, 1949.

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"High Permeability Materials," by T. D. Yensen, Westinghouse Electric Corporation; October 26, 1949.

"History of Mathematics and the Digital Computer," by Gunter Nelson and E. V. Gulden, National Cash Register Company; November 10, 1949.

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"Design of Electronics Equipment Using Subminiature Components," by M. L. Miller, Capehart Farnsworth Company, October 10, 1949.

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"Some Recent Developments in Electronic Applications to Geophysics," by Eugene Frowe, Robert H. Ray Company; September 27, 1949.

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"Magnetic Recording," by M. J. Stolaroff, Ampex Electric Corporation; and Business Meeting; September 27, 1949.

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"The Effect of Contact Potential Difference in Electron Tube Characteristics," by G. D. O'Neill, Sylvania Electric Products Inc.; September 16, 1949.

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ADVENTURES IN ELECTRONIC DESIGN

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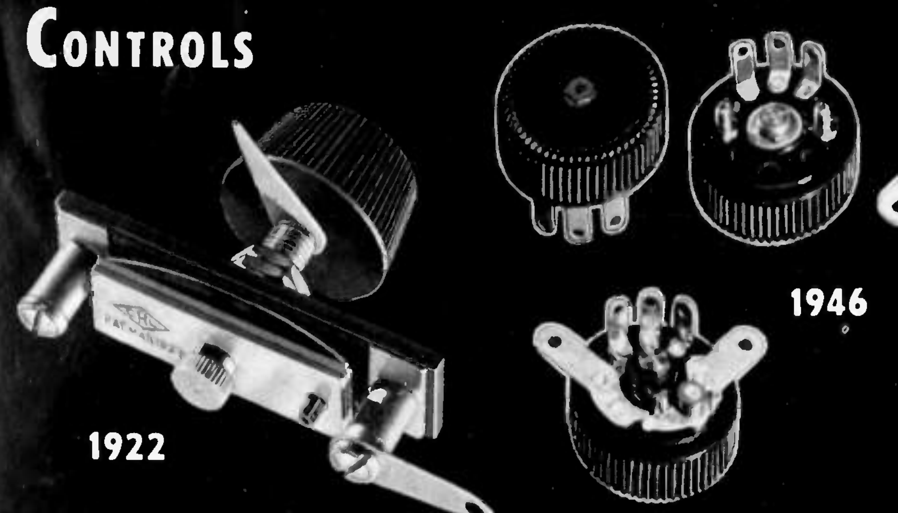
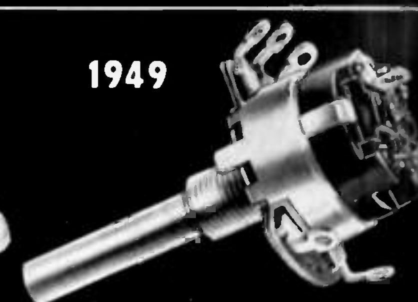
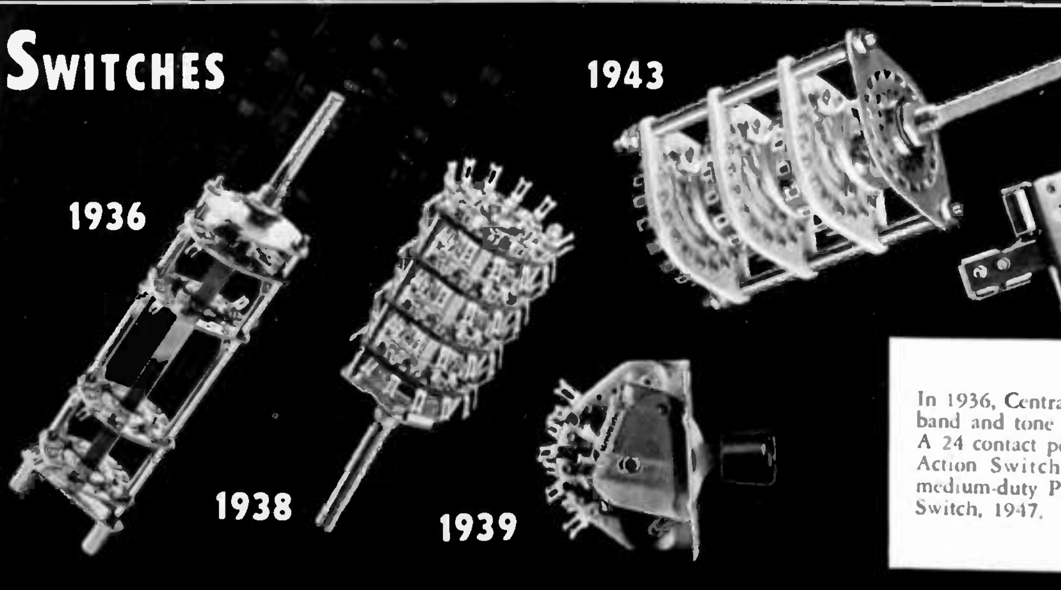
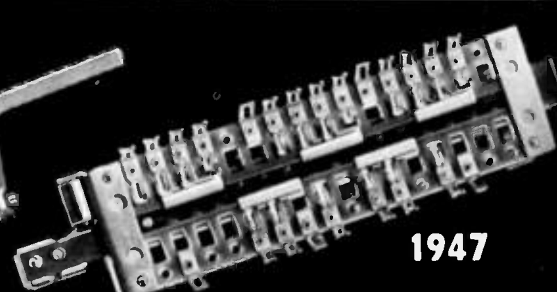




Centralab has come to be known in the electronics industry as "first" in the majority of the products it manufactures and sells. Originator of the Composition Type Volume Control, in 1922. The first Barnacle Type Line Switch attached to a volume control, in 1929. Model 1, smallest switch type hearing aid control, 1946. Now, 1949, Model 2 Radiohm, the most modern high quality variable resistor.

SWITCHES

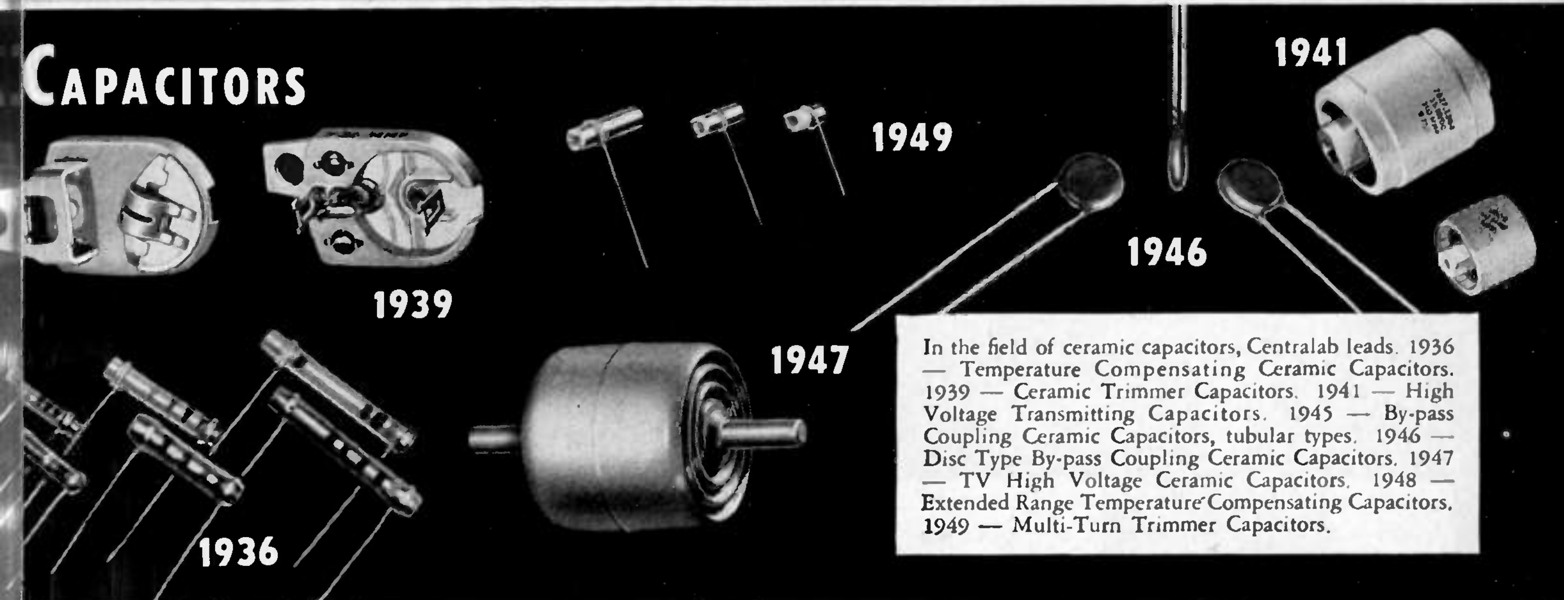




In 1936, Centralab introduced a complete line of wave band and tone switches. First in switch development. A 24 contact per single section switch in 1938. Lever Action Switch in 1939. Low-loss, high-frequency, medium-duty Power Switch, 1943. FM and TV Slide Switch, 1947.

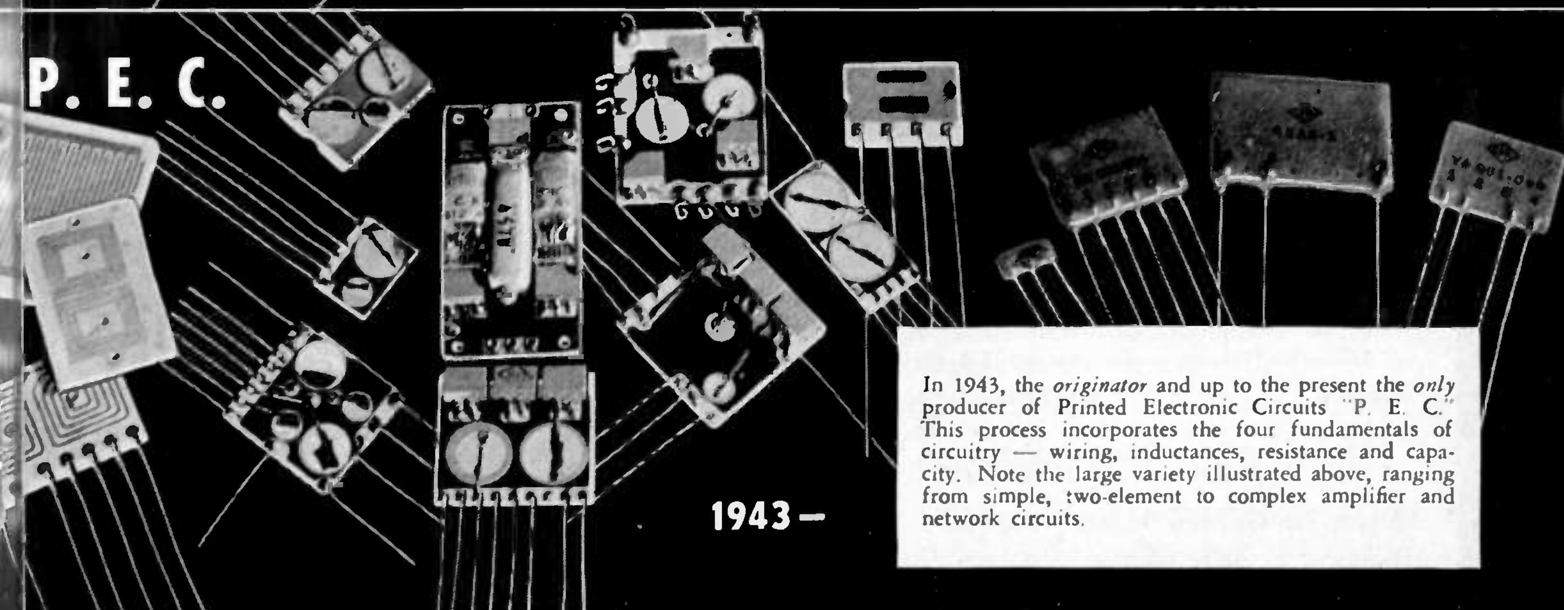
Electronic Industry

CAPACITORS



In the field of ceramic capacitors, Centralab leads. 1936 — Temperature Compensating Ceramic Capacitors. 1939 — Ceramic Trimmer Capacitors. 1941 — High Voltage Transmitting Capacitors. 1945 — By-pass Coupling Ceramic Capacitors, tubular types. 1946 — Disc Type By-pass Coupling Ceramic Capacitors. 1947 — TV High Voltage Ceramic Capacitors. 1948 — Extended Range Temperature Compensating Capacitors. 1949 — Multi-Turn Trimmer Capacitors.

P. E. C.



In 1943, the *originator* and up to the present the *only* producer of Printed Electronic Circuits "P. E. C." This process incorporates the four fundamentals of circuitry — wiring, inductances, resistance and capacity. Note the large variety illustrated above, ranging from simple, two-element to complex amplifier and network circuits.

CERAMICS

In 1942 Centralab developed a grade L-5 Steatite Ceramic superior to the then existing navy grade "G" specification. In 1944 a grade L-6 was introduced. In 1945, Cordierite and Zirconite bodies with grade L-4 rating were developed. Since 1943, Centralab has led in the metalizing of ceramic bodies for special purposes.

IMPORTANT BULLETINS FOR YOUR TECHNICAL LIBRARY!

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Centralab Printed Electronic Circuits

973 — Ampec — three-tube P. E. C. amplifier.
42-6 — Couplate — P. E. C. interstage coupling plate.
999 — Pentode Couplate — specialized P. E. C. coupling plate.
42-9 — Filpec — Printed Electronic Circuit filter.

Centralab Capacitors

42-3 — BC Tubular Hi-Kaps — capacitors for use where temperature compensation is unimportant.
42-4 — BC Disc Hi-Kaps — miniature ceramic BC capacitors.
42-10 — Hi-Vo-Kaps — high voltage capacitors for TV application.
695 — Ceramic Trimmers — CRL trimmer catalog.
981 — Hi-Vo-Kaps — capacitors for TV application. For jobbers.
42-18 — TC Capacitors — temperature compensating capacitors.
814 — Capacitors — high-voltage capacitors.
975 — FT Hi-Kaps — feed-thru capacitors.

Centralab Switches

953 — Slide Switch — applies to AM and FM switching circuits.
970 — Lever Switch — shows indexing combinations.
995 — Rotary Switch — schematic application diagrams.
722 — Switch Catalog — facts on CRL's complete line of switches.

Centralab Controls

42-7 — Model "1" Radiohm — world's smallest commercially produced control.
697 — Variable Resistors — full facts on CRL Variable Resistors.

Centralab Ceramics

967 — Ceramic Capacitor Dielectric Materials.
720 — Ceramic Catalog — CRL steatite, ceramic products.

General

26 — General Catalog — Combines Centralab's line of products for jobber, ham, experimenter, serviceman or industrial user.

Look to CENTRALAB in 1950! First in component research that means lower costs for the electronic industry. If you're planning new equipment, let Centralab's sales and engineering service work with you. For complete information on all CRL products, get in touch with your Centralab Representative. Or write direct.

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SECTION MEETINGS

(Continued from page 34A)

NEW YORK

"Trends in Air Navigation Instrumentation," by Harry Davis, Watson Laboratories, and George Comstock, Airborne Instrument Laboratories; September 14, 1949.

"Scientific Calculation by Electronic Means," by W. J. Eckert, Watson Computing Laboratory, Columbia University; October 5, 1949.

NORTH CAROLINA-VIRGINIA

"Design Problems of Triodes and Tetrodes for High-Frequency Operation," by C. M. Smith; Machlett Laboratories; September 23, 1949.

PHILADELPHIA

"Visual Sensitivity—How It Is Influenced by Ultra-Violet," by Ernst Wolf, American Optical Company; October 6, 1949.

PITTSBURGH

"The X-Ray Image Amplifier Tube," by F. Marshall, Westinghouse Research Laboratories; October 8, 1949.

PORTLAND

"The Measurement of Nonlinear Distortion," by A. P. G. Peterson, General Radio Company; September 8, 1949.

"Antenna Impedance," by F. E. Miller, Pacific Telephone and Telegraph Company; September 22, 1949.

PRINCETON

"Television by Pulse Code Modulation," by W. M. Goodall, Bell Telephone Laboratories; October 13, 1949

ROCHESTER

"High-Speed Movies with Grid Photography," by W. C. Newcomb, Eastman Kodak Company; October 6, 1949.

SACRAMENTO

"Latest Developments in Color Television," by N. D. Webster, McClatchy Broadcasting System; "Radio Range Facilities," by A. H. Rhode, United States Air Force, McClellan Field; "Direction Finding Facilities," by I. L. Dutton, United States Air Force, McClellan Field; "United States Air Force Television," by Raymond Fisher, United States Air Force, McClellan Field; and "New Developments in Ground Control Approach," by L. A. Querolo, United States Air Force, McClellan Field; September 13, 1949.

SAINT LOUIS

"Universal Phonograph Styli," by J. D. Reid, Crosley Division, AVCO Manufacturing Corporation; September 21, 1949.

SALT LAKE

"Microwaves for Instrument Landing of Aircraft," by D. F. Folland, Sperry Gyroscope Company; October 4, 1949.

TOLEDO

Film: General Science, Electrical Radio; October 3, 1949

WASHINGTON

"Application of Electronic Techniques in Medical Research," by S. A. Talbot, Johns Hopkins Hospital; October 10, 1949.

SUBSECTION MEETINGS

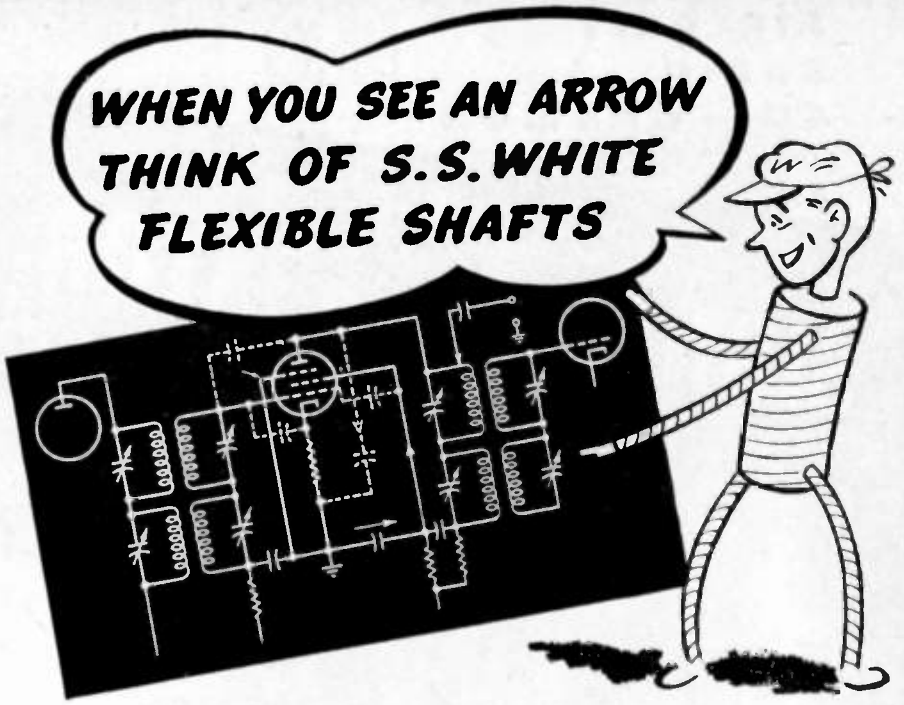
AMARILLO-LUBBOCK

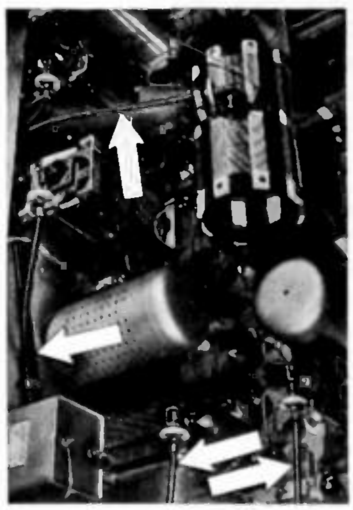
"FCC Rules and Regulations," by J. H. Homsy, FCC District 10; September 23, 1949.

LANCASTER

"RCA (Ground Control Approach) Past, Present and Future," by C. W. Hicks, Bendix Radio Division; October 12, 1949.

(Continued on page 40A)





The designer of this broadcast transmitter circuit thought of S.S.White flexible shafts—and used them effectively to get optimum circuit efficiency and conveniently centralized control.

"As you know, an arrow through a symbol means a variable element. In a circuit diagram variable elements are no problem. You just draw them where you want them. But it's quite different when you come to design the actual equipment. Then, you have to consider electrical efficiency, operating convenience, ease of wiring, space economy, appearance and servicing.

"You can satisfy every one of these requirements by coupling variable elements to their controls with S.S.White flexible shafts, because this simple arrangement gives you complete freedom to put both the elements and their controls where it's best for them to be.

"So, whenever you see an arrow, think of S.S.White flexible shafts. And here's how you can get complete information about them. . . ."

WRITE FOR THIS FLEXIBLE SHAFT HANDBOOK

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SIGNAL GENERATOR

900-2100 MEGACYCLES

–Simplified
–Compact
–Portable

- 900-2100 megacycles, single band
- Directly calibrated, single dial frequency control
- Directly calibrated attenuator, 0 to –120 dbm
- CW or AM pulse modulation
- Internal pulse generator with controls for width, delay, and rate. Provision for external pulsing
- Controls planned and grouped for ease of operation
- Weight: 42 lbs. Easily portable—ideal for airborne installations
- Immediate delivery

Built to Navy Specifications for research and production testing

Write for specifications — investigate the advantages of this outstanding new instrument.

DEPENDABLE ELECTRONIC EQUIPMENT SINCE 1928

Aircraft Radio Corporation
BOONTON, New Jersey

SECTION MEETINGS

(Continued from page 39A)

LONG ISLAND

"The Mine Detector of World War II," by L. F. Curtis, Hazeltine Electronics Corporation, and H. A. Wheeler, Wheeler Laboratories; October 11, 1949.

MONMOUTH

"Programming for a Large-Scale Digital Computer," by Joseph Weinstein, Signal Corps Engineering Laboratories; September 21, 1949.

NORTHERN NEW JERSEY

"A Report on the Scheduled FCC Hearing to Consider Proposals for a Change in TV Transmission Standards," by T. T. Goldsmith, Allen B. Du-Mont Laboratories; September 21, 1949.

STUDENT BRANCH MEETINGS

CASE INSTITUTE OF TECHNOLOGY—IRE BRANCH

"Problems and Prospects of Television," by Russell Olsen, Chief Engineer, Radio Station WEWS-TV; October 4, 1949.

COLLEGE OF THE CITY OF NEW YORK—IRE BRANCH

"The Engineer's Job and the Engineering Society," by E. K. Gannett, Technical Editor, The Institute of Radio Engineers; October 4, 1949.

CLARKSON COLLEGE OF TECHNOLOGY—IRE BRANCH

Election of Officers; October 6, 1949.

UNIVERSITY OF COLORADO—IRE BRANCH

"A New Long-Range Navigational Aid; Console," by G. E. Glass, Faculty of University of Colorado; October 6, 1949.

CORNELL UNIVERSITY—IRE-AIEE BRANCH

"Solar Radiation and its Effect on Long-Distance Power Lines," by J. T. Wilson, Chief Physicist, Allis Chalmers Corporation; October 7, 1949.

"A Welcome to Freshmen," by H. B. Hansteen, Faculty of Cornell University; September 30, 1949.

UNIVERSITY OF FLORIDA—IRE-AIEE BRANCH

"Public Opinion the Ultimate Source of Power," by Seldon Waldo; September 29, 1949.

GEORGIA SCHOOL OF TECHNOLOGY—IRE BRANCH

"Wartime Radar," by M. A. Honnell, Faculty of Georgia School of Technology; October 13, 1949.

UNIVERSITY OF ILLINOIS—IRE-AIEE BRANCH

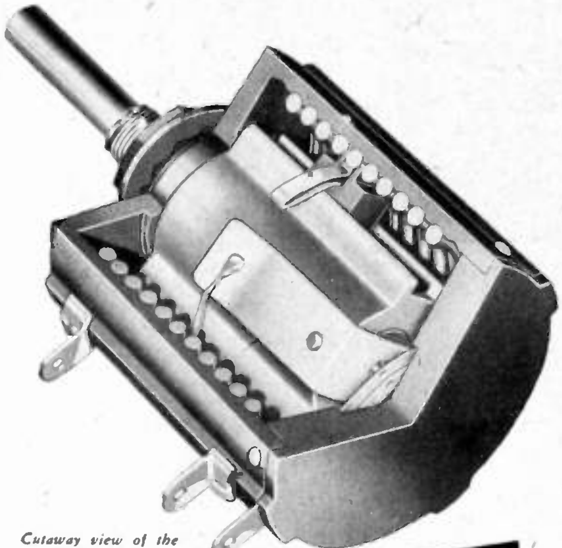
Speakers; Dr. Ryder, Head of Electrical Engineering Department, University of Illinois, and W. L. Everitt, Head of College of Engineering, University of Illinois; October 4, 1949.

IOWA STATE COLLEGE—IRE-AIEE BRANCH

"Industry and Your AIEE and IRE," by G. R. Town, Head of Experiment Station, Iowa State College; October 5, 1949.

(Continued on page 42A)

For new simplicity, wide range, and high accuracy in the control of modern electronic circuits...



Cutaway view of the HELIPOT (Model A—10 Turn—1¾" Diameter)

THE BECKMAN Helipot

(Trademark of the HELical POTentiometer)

Provides *many times* greater resistance control in *same* panel space as conventional potentiometers!

IF YOU are designing or manufacturing *any type* of precision electronic equipment be sure to investigate the greater convenience, utility, range and compactness that can be incorporated into your equipment by using the revolutionary HELIPOT for rheostat-potentiometer control applications... and by using the new DUODIAL turns-indicating knob described at right.

Briefly, here is the HELIPOT principle... whereas a conventional potentiometer consists of a *single* coil of resistance winding, the HELIPOT has a resistance element *many times* longer coiled *helically* into a case which requires no more panel space than the conventional unit. A simple, foolproof guide controls the slider contact so that it follows the helical path of the resistance winding from end to end as a single knob is rotated. Result... *with no increase in panel space requirements,* the HELIPOT gives you as much as 12 times* the control surface. You get far greater accuracy, finer settings, increased range—with maximum compactness and operating simplicity!

COMPLETE RANGE OF TYPES AND SIZES

The HELIPOT is available in a complete range of types and sizes to meet a wide variety of control applications...

MODEL A: *5 watts, 10 turns, 46" slide wire length, 1¾" case dia., resistances 10 to 50,000 ohms, 3600° rotation.*

MODEL B: *10 watts, 15 turns, 140" slide wire length, 3¼" case dia., resistances 50 to 200,000 ohms, 5400° rotation.*

MODEL C: *3 watts, 3 turns, 13½" slide wire length, 1¾" case dia., resistances 5 to 15,000 ohms, 1080° rotation.*

MODEL D: *15 watts, 25 turns, 234" slide wire length, 3¼ case dia., resistances 100 to 300,000 ohms, 9000° rotation.*

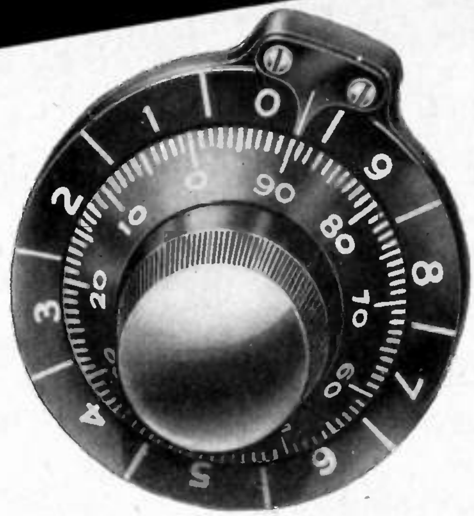
MODEL E: *20 watts, 40 turns, 373" slide wire length, 3¼" case dia., resistances 150 to 500,000 ohms, 14,400° rotation.*

Also, the HELIPOT is available in various special designs... with double shaft extensions, in multiple assemblies, integral dual units, etc.

Let us study *your* potentiometer problems and suggest how the HELIPOT can be used—possibly is already being used by others in your industry—to increase the accuracy, convenience and simplicity of modern electronic equipment. No obligation, of course. Write today outlining your problem.

**Data for Model A, 1¾" dia. Helipot. Other models give even greater control range in 3" case diameters.*

THE BECKMAN Duodial



The inner, or Primary dial of the DUODIAL shows exact angular position of shaft during each revolution. The outer, or Secondary dial shows number of complete revolutions made by the Primary dial.

A multi-turn rotational-indicating knob dial for use with the HELIPOT and other multiple turn devices.

THE DUODIAL is a unique advancement in knob dial design. It consists essentially of a primary knob dial geared to a concentric turns-indicating secondary dial—and the entire unit is so compact it requires only a 2" diameter panel space!

The DUODIAL is so designed that—as the primary dial rotates through each complete revolution—the secondary dial moves one division on its scale. Thus, the secondary dial counts the number of complete revolutions made by the primary dial. When used with the HELIPOT, the DUODIAL registers *both* the angular position of the slider contact on any given helix as well as the particular helix on which the slider is positioned.

Besides its use on the HELIPOT, the DUODIAL is readily adaptable to other helically wound devices as well as to many conventional gear-driven controls where extra dial length is desired without wasting panel space. It is compact, simple and rugged. It contains only two moving parts, both made entirely of metal. It cannot be damaged through jamming of the driven unit, or by forcing beyond any mechanical stop. It is not subject to error from backlash of internal gears.

TWO SIZES—MANY RATIOS

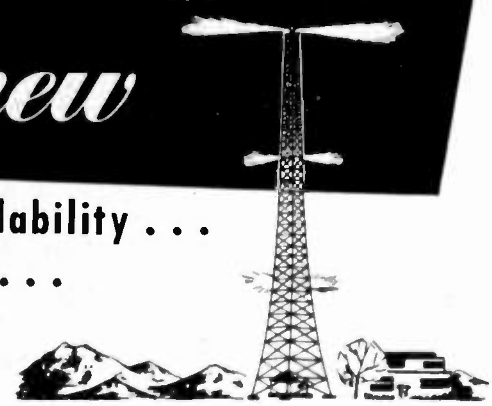
The DUODIAL is now available in two types—the Model "R" (illustrated above) which is 2" in diameter, and the new Model "W" which is 4¾" in diameter and is ideal for main control applications. Standard turns-ratios include 10:1, 15:1, 25:1 and 40:1 (ratio between primary and secondary dials). Other ratios can be provided on special order. The 10:1 ratio DUODIAL can be readily employed with devices operating *fewer* than 10 revolutions and is recommended for the 3-turn HELIPOT. In all types, the primary dial and shaft operate with a 1:1 ratio, and all types mount directly on a ¼" round shaft.

Send for this HELIPOT AND DUODIAL CATALOG!

Contains complete data, construction details, etc., on the many sizes and types of HELIPOTS... and on the many unique features of the DUODIAL. Send for your free copy today!

THE Helipot CORPORATION, SOUTH PASADENA 6, CALIFORNIA

a COMPLETE LINE of CAA APPROVED* TOWER LIGHTING EQUIPMENT BY Andrew



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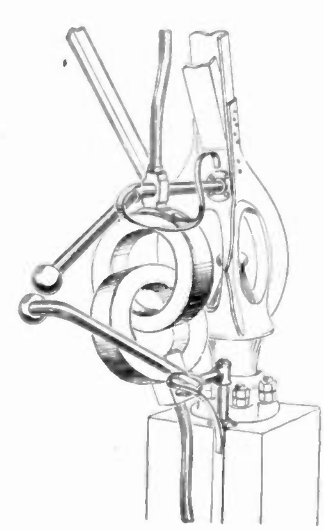
300 MM CODE BEACON, Type 660. Sturdily constructed, completely dependable. To provide steady, uninterrupted service for many years of exposure to rigorous weather conditions, metal parts are made of cast aluminum with hardware of corrosion resistant bronze. Insects are kept out by screens placed in ventilating openings.



ISOFORMERS, Types 2015 and 2030. Interlocking ring, air-insulated lighting transformers; particularly adapted for use with towers that develop a high voltage across the base insulator.

REPLACEMENT LAMPS, for code beacons and obstruction lights. Carried in stock in variety of filament voltages.

LIGHTING FILTERS, for use with insulated towers developing moderate voltages above 1 MC. Models available unhoused or in weatherproof steel housing.

BURNOUT INDICATORS, to show lamp failure.

PHOTOELECTRIC CONTROL SWITCHES, to turn tower lights ON and OFF.

FLASHERS, for code beacons.

COMPLETE TOWER LIGHTING KITS, including conduit, wire, and all fittings for towers of any height.

SINGLE (Type 661A) and DOUBLE (Type 662A) OBSTRUCTION LIGHTS. Easy to service, rugged, reliable. To replace burned out lamps, just loosen one thumb screw and open the two piece cast aluminum housing.

Write for descriptive bulletins or further information—today.

**CAA approvals cover only lighting fixtures themselves. Associated equipment is not subject to CAA regulations but more than meets all local regulations.*

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ANTENNA TUNING UNITS · TOWER LIGHTING EQUIPMENT · CONSULTING ENGINEERING SERVICES
WORLD'S LARGEST ANTENNA EQUIPMENT SPECIALISTS

STUDENT BRANCH MEETINGS

(Continued from page 40A)

State University of Iowa—IRE Branch

"Introduction to the Student Branch of the IRE," by L. A. Ware, Faculty of State University of Iowa; September 28, 1949

Films: Power and Light, and Curves of Color October 5, 1949

Field Trip; October 12, 1949.

Business Meeting: October 19, 1949

Kansas State College—IRE Branch

"Precipitation Static in Aircraft," by R. C Ayres, Chief Testing Engineer Bendix Aviation Corporation: October 6, 1949

University of Kentucky—IRE Branch

Film: Unfinished Rainbows; October 11, 1949

Lafayette College—IRE-AIEE Branch

"The AIEE," by William Rohland, and "The IRE," by Robert Kudlich, Student Branch Officers, September 27 1949.

Manhattan College—IRE Branch

Business Meeting: October 12, 1949

University of Michigan—IRE-AIEE Branch

"Preparing for Interviews," by W. C. Bergman, Michigan Bell Telephone Company; October 5, 1949

University of Missouri—IRE-AIEE

"How Co'Op Electricity Works," by R. J. Martin, Manager, Boone County R.E.A.; October 13 1949

Missouri School of Mines and Metallurgy—IRE-AIEE Branch

"Midgets of Telephone Science," by Mr. Mattick, Bell Telephone Company; and Election of Officers; October 6, 1949

Newark College of Engineering—IRE Branch

Nomination of Officers; October 4, 1949
Election of Officers; October 11 1949

University of New Mexico—IRE Branch

Business Meeting; September 30, 1949

New York University—IRE Branch

Film: Adventures in Research; October 6, 1949

University of Notre Dame—IRE-AIEE Branch

"The National Electronic Code," by John Brehmer, Electrical Contractor; September 28 1949

Ohio State University—IRE-AIEE Branch

"Recent Developments in Electron Tube Research," by E. M. Boone, Faculty of Ohio State University; "Magnatrons," by John Moll; "Klystron Oscillators," by James Ebers; and "Traveling Wave Tube," by George Muller; October 6 1949

Oregon State College—IRE Branch

"The AIEE and IRE and Student Engineers," by F. O. McMillan, Head of School of Engineering Oregon State College; September 28, 1949

"A New Personnel Program in the Pacific Telephone and Telegraph Company," by Bruce Pickett, Pacific Telephone and Telegraph Company; October 13, 1949

University of Pennsylvania—IRE-AIEE Branch

Business Meeting; October 17, 1949

(Continued on page 44A)

A page from the note-book of Sylvania Research

Sylvania advances development and production of Planar Triode Tubes

The production of planar triodes in the Advanced Development Laboratories of Sylvania Electric at Kew Gardens, N. Y. bears small resemblance to standard receiving tube manufacture. The unusual design of the tubes eliminates conventional stem and mounting techniques. Instead individual tube elements, including plate, grid, and cathode are sealed directly into the glass envelope!

Sylvania is now producing several types of tubes in the planar triode family for direct application to microwave systems where frequencies range between 300 and 3000 mc. Used as pulse generators, some of these types will develop peak power outputs up to 200 watts.

Such results are typical of Sylvania's continuing research and determination to create new and better products.

In the planar triode tubes, disc-shaped grid and plate leads and rod-shaped cathode leads are sealed directly into glass. Internal grids consist of fine parallel wires under tension to assure dimensional stability. Individual metal parts must be fabricated to precise dimension and contour for the requirements of microwave applications. Throughout manufacture, the greatest of care in handling, storage and application is required and maintained.

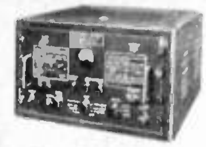
SYLVANIA ELECTRIC

ELECTRONIC DEVICES; RADIO TUBES; CATHODE RAY TUBES; FLUORESCENT LAMPS, FIXTURES, WIRING DEVICES, SIGN TUBING; LIGHT BULBS; PHOTOLAMPS

If the Problem is ELECTRONIC

... look to B&W for the answer

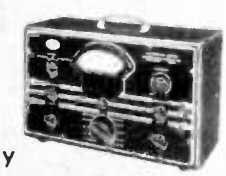
2 KW AMPLIFIER



CONTROL UNIT

DUAL DIVERSITY CONVERTER

RECEIVER ASSEMBLY



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From amplifiers to electronic repeaters ... from diversity converters to complete receiver assemblies, test instruments and frequency multiplier units, B & W equipment is backed by men who have spent a lifetime in the electronic field ... men who know the problems of radio and electronics and how to solve them. Modern, up-to-date facilities are ready to convert the designs of these engineers into components or complete assemblies, designed and fabricated to withstand the toughest assignments and carrying the B & W trade-mark, recognized the world over for excellence.

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BARKER & WILLIAMSON, Inc.

237 Fairfield Avenue — Upper Darby, Pa.

STUDENT BRANCH

MEETINGS

(Continued from page 42A)

Pratt Institute—IRE Branch

Business Meeting; September 18, 1949.

"The Consulting Engineer," by T. C. Gams Industrial Electronics School; October 18, 1949.

Rhode Island State College—IRE-AIEE Branch

Business Meeting; September 27, 1949.

Rutgers University—IRE-AIEE Branch

"Function of Personnel Placement Division," by William Hobbie; October 11, 1949.

Syracuse University—IRE-AIEE Branch

"Engineers in Russia," by Igor Plusc, Faculty of Syracuse University; Nomination of Officers; September 28, 1949.

"Working in a Power Plant," by D. Y. Brouse, Student; October 12, 1949.

South Dakota School of Mines and Technology—IRE Branch

Business Meeting; September 28, 1949.

Texas Agricultural and Mechanical College—IRE-AIEE Branch

Business Meeting; Election of Officers; October 11, 1949.

University of Utah—IRE-AIEE Branch

"Orientation to Activities of AIEE-IRE," by O. C. Haycock, Faculty of University of Utah; October 4, 1949.

Virginia Polytechnic Institute—IRE Branch

Business Meeting; September 27, 1949.

"Graduate Work in Electrical Engineering," by F. C. Vilbrandt, Head of Chemical Engineering Department, Virginia Polytechnic Institute; October 4, 1949.

Films: Coaxial Cable, Antennas and Microwaves, and Station Installer; October 18, 1949.

"Analaogy Between Fluid Flow and Electric Current," by V. G. Szebehely, Faculty of Virginia Polytechnic Institute; October 11, 1949.

Wayne University—IRE-AIEE Branch

Business Meeting; September 27, 1949.

MEMBERSHIP

The following transfers and admissions were approved and will be effective as of December 1, 1949:

Transfer to Senior Member

Black, R. R., 2203 U Pl., S. E., Washington 20 D. C.

Fisher, W. C., RCA Victor Company, Ltd., 168 Market Ave., E., Winnipeg, Man., Canada

Gregory, L. W., 108 Oaklee Village, Baltimore 29, Md.

Hedeman, W. R., Jr., 908 Greenleigh Rd., Baltimore 12, Md.

Hughes, W. R., 11600 Sherman Way, N. Hollywood, Calif.

Kramer, A. S., 377 S. Second St., Lindenhurst, L. I. N. Y.

(Continued on page 46A)

Solve your **SEALED LEAD** *problems in* **SECONDS**—

— WITH THE NEW
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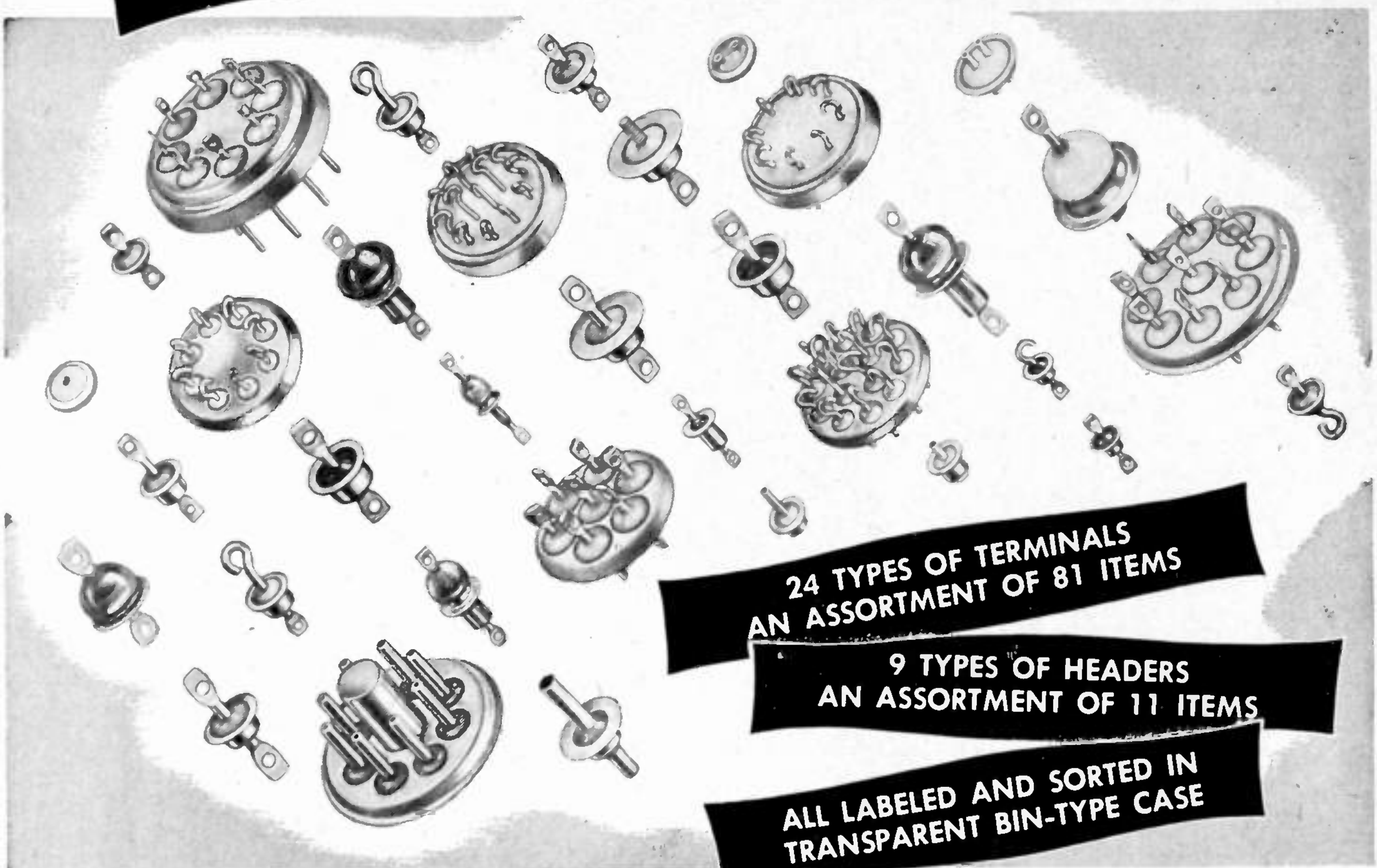
E-I sealed lead and multiple header kit!



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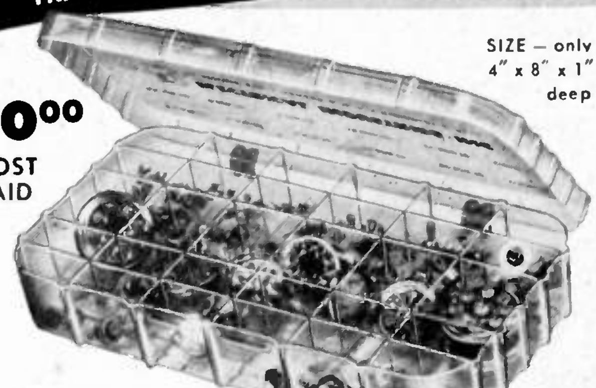

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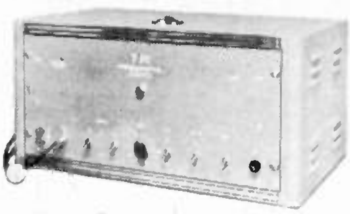
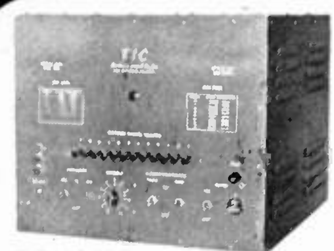


TYPE 2120
R.F. PICTURE SIGNAL GENERATOR

Provides picture and sound carrier. Modulated by standard R.M.A. composite picture signal. Sound carrier stability suitable for testing Inter Carrier type receivers. Internal 400 cycle FM and External audio with 75 microsecond pre-emphasis. Output max. 0.1v p-p across 75 ohm line. Available channels 2-13.



TYPE 1200
12 CHANNEL
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Intended for precise adjustment of R.F. head oscillator coils and R.F. band pass circuits. Pulse type markers of picture and sound carrier frequencies extend to zero signal reference base line. Accuracy of markers 0.02% of carrier frequency. 12 to 15 MC. sweep on all channels. Max. 1.V peak output across a 75 ohm line. Provisions for balanced input receivers. Instant selection by push button.

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(Continued from page 44A)

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Mayo-Wells, W. J., 5938 13 Pl. N.W., Washington 11, D. C.
Millman, J., 144-54-69 Ave., Flushing, L. I., N. Y.
Schwetman, H. D., Physics Department, Baylor University, Waco, Tex.
Soria, R. M., 1830 S. 54 Ave., Chicago 50, Ill.
Stralton, A. W., Department of Electrical Engineering, University of Texas, Austin, Tex.
Thorson, H. L., 1517 Wyoming Ave., Schenectady, N. Y. (effective November 1)
Vogt, E. J., 2200 N. Tejon St., Colorado Springs, Colo.
Webber, S. E., Research Laboratory, General Electric Company, Schenectady, N. Y.
Welge, V., 432 LaCrescentia Dr., San Diego, Calif.
Wolf, H., 114-01—86 Ave., Richmond Hill, L. I., N. Y.

Admission to Senior Member

Beck, V. R., 254 Grantley, Elmhurst, Ill.
Pack, L., 128 Russell Lane, London N. 20, England
Watson, S. H., Chews Landing Rd. cor. Hutchinson Ave., Haddonfield, N. J.
Willoughby, J. A., Federal Communications Commission, Washington 25, D. C.

Transfer to Member

Brewer, G. R., 1359 Erving Ct., Willow Run Village, Mich.
Cholmondeley-Smith, D. R., Transmitting Station New Zealand Broadcasting Service, Opapa Hawke's Bay, New Zealand
Douglass, C. F., 19 W. Fifth St., Emporium, Pa.
Fulmer, N. C., 16 Forest St., Montclair, N. J.
Gerlough, D. L., Department of Engineering, University of California, Los Angeles, Calif.
Hanft, H., 711 New Jersey Ave., Brooklyn 7, N. Y.
Hyland, F. G., 1394 N. Fifth St., Columbus, Ohio
Johnson, L. E., The Pines, R. R. 3, Wayzata, Minn.
Kong, Y., 26-3 W. Wai Oy Rd., Canton, China
Martin, R. A., 210 North Ave., N.W., Atlanta 3, Ga.
McCall, E. A., 3504 E. 26 St., Kansas City, Mo.
Nord, R. H., Box 6, Bayside, L. I., N. Y.
Occhiogrosso, T., 84 Barker Ave., Eatontown, N. J.
Roney, E. L., 479 Pacific St., San Luis Obispo, Calif.
Stultz, L. R., 46 Westcliff Rd., Colonia, N. J.
Waxler, B., 1956 Bathgate Ave., New York 57, N. Y.

Admission to Member

Brenneman, D. E., Bell Telephone Laboratories, 463 West St., New York 14, N. Y.
Cameron, D. B., National Carbon Company, Inc., Box 6087, Cleveland, Ohio
Fox, W. C. O., Thornycroft Apartments, Garth Rd., Scarsdale, N. Y.
Hewitt, F. J., 4 Orange St., Sandringham, Johannesburg, South Africa
Hults, E. H., 18 Milburn Ct., Baldwin, N. Y.
James, W. S., Jr., 200 Highland Blvd., Brooklyn 7, N. Y.
Khouri, J. O., Direction Telephonique, Beyrouth, Lebanon
LaVielle, W. R. R., 495 Lightfoot Rd., Louisville, Ky.
Loreto, D. R., 62 Sicard St., New Brunswick, N. J.
Mayerson, M. I., 41 Barker Ave., Eatontown, N. J.
Mleczko, E. L., 7619 Hinds Ave., N. Hollywood, Calif.
Perry, R. L., 259 Saratoga Ave., Los Gatos, Calif.
Pratt, C. B., 11 Halsey Dr., Dayton 3, Ohio
Robinson, W. H., 29 Bartlett Ave., Arlington 74, Mass.
Ruark, A. E., 1315 St. Paul St., Baltimore 2, Md.
Shaw, A. B., 316 W. 44 St., New York 18, N. Y.
Sirota, N., 10 E. 57 St., Brooklyn, N. Y.
Thomas, L. R., 830 Park Ave., Wilmette, Ill.

(Continued on page 47A)

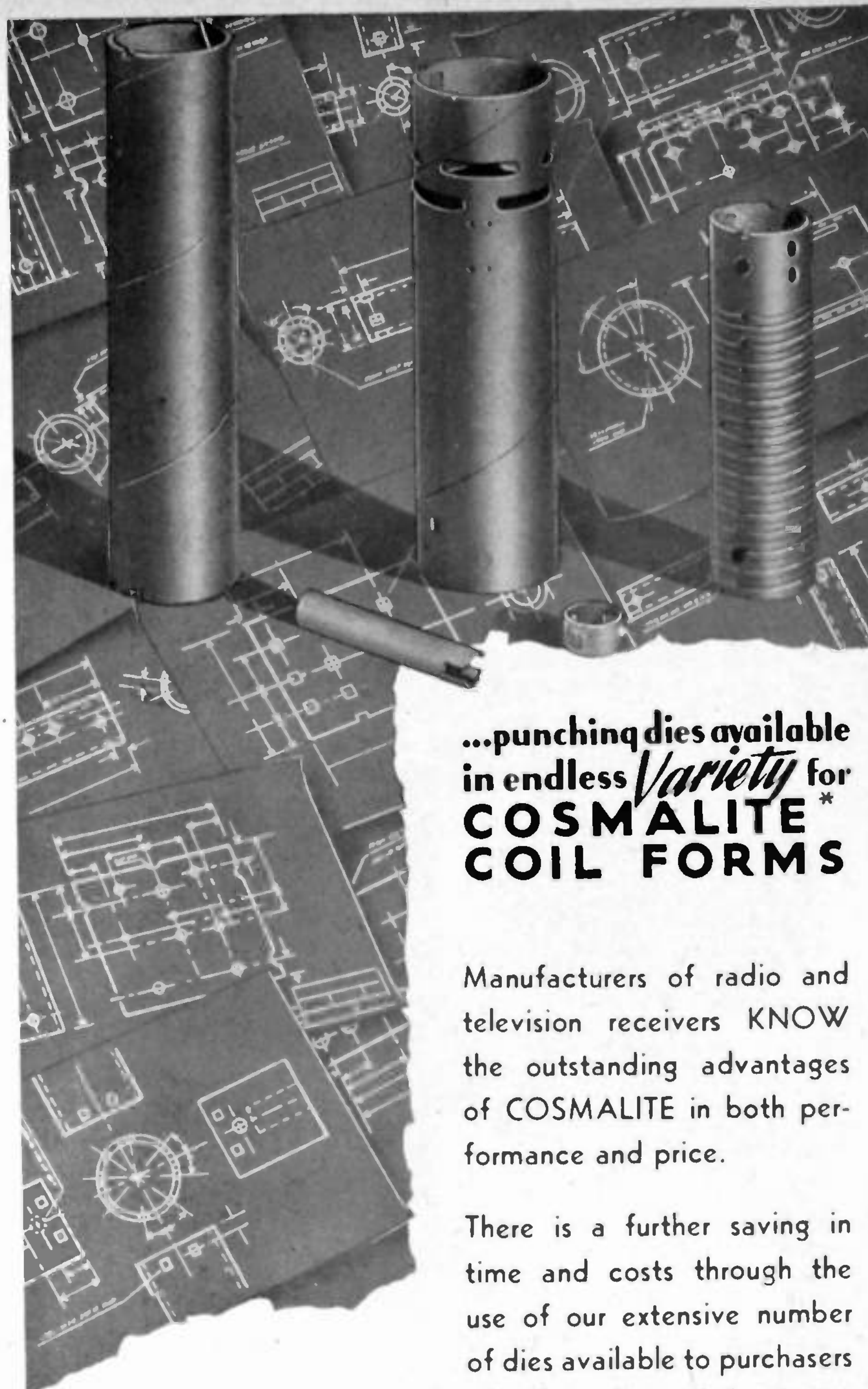
MEMBERSHIP

(Continued from page 46A)

The following elections to the Associate grade have been approved and will be effective as of November 1, 1949:

Abert, E. P., 108 Forest Dr., North Syracuse, N. Y.
Accardo, N. A., 267 Montclair Ave., Vaux Hall, N. J.
Amaitis, E. J., 310 S. Third St., Brooklyn 11, N. Y
Baker, W. L., Box 881, State College, Pa.
Benson, E. D., 70 Forsyth St., Boston, Mass
Bettis, W. E., 3328 T. S., Box 220, Scott Air Force Base, Ill.
Binkholder, C. E., c/o American Television Institute, 5050 Broadway, Chicago, Ill.
Brandt, R. W., 4545 W. Augusta Blvd., Chicago 51, Ill.
Bross, C. F., 5 E. P. Bross, R.R. 3, Palmyra, Mo
Buersmeyer, C. R., Box 371, Washington, Mo.
Burgar, D. G., 2178 Ave. H, Redondo Beach, Calif
Burgard, G. J., Maple Ave., Evans City, Pa
Clark, D., Jr., 40270 Abourne Rd., Los Angeles, Calif.
Cloutier, D. A., 4545 W. Augusta Blvd., Chicago 51, Ill.
Coffey, J. L., 32 Oak Island Rd., Revere 51, Mass
Cooper, W. C., 3960 Ogden, Beaumont, Tex.
Cotellessa, R. F., Box 162, 472 Van Emburgh Ave., Ridgewood, N. J.
Crupi, D. W., 5050 Broadway, Chicago, Ill.
Das, J., West Command Signal Regiment, New Delhi, India
Daykin, D. R., 1283 Clifton Prado, Lakewood Ohio
De Cubas, J. D., 20 Christopher St., New York 14, N. Y.
De Socio, G., 702 Evesham Ave., Baltimore 12, Md
Dilger, L. E., 3617 N. 79 St., Milwaukee 13, Wis.
Dobbins, W. E., 217—17 St., Manhattan Beach, Calif.
Dolence, J. J., 918 E. 14 Pl., Chicago 15, Ill.
Duschenchuk, L., 148 Newport Rd., E. Hempstead L. I., N. Y.
Dyett, E. G., Jr., 29 William St., Cambridge 39 Mass.
Elliott, J. W., 1157 W. 11 St., San Pedro, Calif
Ellis, C. E., Jr., 204 Fostoria Ave., Springfield, Ohio
Farrell, P. T., 495 E. 188 St., New York 58, N. Y
Fegley, K. A., 200 S. 33 St., Philadelphia 4, Pa.
Fields, T., 6917 S. Crandon Ave., Chicago 49, Ill.
Fong, B. W., 1129 Stockton St., San Francisco, Calif.
Friedberg, I. S., 86 W. 183 St., New York 53, N. Y
Fuller, B. L., 4024 Hillcrest Dr., Los Angeles 43, Calif.
Gerold, L., 53 Prairie Lane, Levittown, Hicksville, L. I., N. Y.
Green, C. A., Sioux Ordnance Depot, Sidney, Nebr.
Greer, L. H., 26 W. 47 St., New York 19, N. Y.
Hardy, W. G., 807 Columbus Ave., New York 25 N. Y.
Harvey, F. K., Bell Telephone Laboratories, Murray Hill, N. J.
Hausenbauer, C. R., College of Engineering, University of Arizona, Tucson, Ariz.
Heckert, R. E., 1203 S. Berendo St., Los Angeles 6, Calif.
Hedges, H. G., 519 Roy Ave., Dayton 9, Ohio
Holt, T., 2301 Gantz Rd., Grove City, Ohio
Honda, H., 36 N. 34 St., Philadelphia, Pa.
Hopf, E. W., 324 William St., Boonton, N. J.
Houlroyd, G. F., Boonton Radio Corp., Intervale Rd., Boonton, N. J.
How, F., South African Broadcasting Corporation Commissioner St., Johannesburg, Transvaal, South Africa
Jones, S., 104-45—121 St., Richmond Hill, L. I. N. Y.
Kantor, F. I., 3869 Sedgwick Ave., New York 63, N. Y

(Continued on page 48A)



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Lanning, W. M., 6820 W. 88 St., Los Angeles, Calif
Lock, R. W., 6228 N. Winthrop Ave., Chicago 40 Ill.
Marek, L. W., 215 W. 23 St., New York 11, N. Y.
Martin, P. W., 3019 Walnut St., Huntington Park, Calif.
Maass, C. F., 516 E. South St., Anaheim, Calif.
McKay, M. W., 2368 Victory Pkwy., Cincinnati 6 Ohio
Myers, E. G., 102½ S. Flenoaks Blvd., Burbank, Calif.
Neill, W. R., Box 131, Medway, Ohio
Newman, H. S., R. D. 1, Kent, Ohio
Nitsche, J. E., Corning Glass Works, Corning, N. Y.
Patrick, P. J., Box 2386, Anchorage, Alaska
Perkins, G. M., General Delivery, China Lake, Calif.
Pickens, G. O., Box 295, Point Loma, San Diego Calif.
Press, M., 46-40—189 St., Flushing, L. I., N. Y.
Quavraux, H. F., 1031 N. Bonnie Brae, Los Angeles Calif.
Rai, R. K., 5 Ridge, Nagpur, India
Rhode, F. S., Box 22, Valparaiso, Ind.
Rice, J. W., 272 Union St., Ashland, Mass.
Roberts, L. L., Jr., 4353 S. Van Buren Pl., Los Angeles, Calif.
Rooten, A., 702 Smith Ave., Xenia, Ohio
Sauer, R. D., 57 Anchor Lane, Levittown, L. I. N. Y.
Scott, D. E., 16508 Eastburn, Detroit, Mich.
Sear, H., Gladstone Hotel, 11 and Pine St., Philadelphia, Pa.
Seybold, R. E., 2803 Summit, Toledo 11, Ohio
Shaw, B., 321 Ave. C, New York 9, N. Y.
Shuman, H., 39 Baird St., Dorchester 24, Mass.
Smith, J. G., 15026 Talman Ave., Harvey, Ill.
Soloway, M. D., 1929 Haight Ave., New York 61, N. Y.
Sparks, W. J., Jr., 134 W. 21 St., Indianapolis, Ind.
Sprengel, W. L., 1130 Noble St., Toledo, Ohio
Stanton, W. R., 225 Mt. Pleasant Ave., Stratford Conn.
Taischoff, J., 30 Rockefeller Plaza, New York 20, N. Y.
Thresh, J. L., 375 Carroll Park, E., Long Beach 14, Calif.
Tippott, D. W., 1405—36 St., Sacramento, Calif.
Venketakrishnan, R., 40 Edward Elliot Rd. Mylapore, Madras, India
Volpe, D. F., 316 E. 22 St., Paterson 4, N. J.
Walker, E. T., 1627 Russell Ave., Dayton, Ohio
Weber, G. J., 1525 Sampson St., Los Angeles 33, Calif.
Weber, S. E., 205 W. Second St., Arcanum, Ohio
Wightman, B. A., National Research Council Sussex St., Ottawa, Ont., Canada
Wilson, C. R., 91-21—195, Hollis, L. I., N. Y.
Wong, G. W., 66 University Rd., Brookline 46, Mass.
Wynn, J. D., Jr., Box 1186, Port Hueneme, Calif.
Yarbrough, K. A., 6038 Bryan Pkwy., Dallas 5, Tex.
Young, R. O., 620 W. 115 St., New York 25, N. Y.
Zegers, T. A., 2652 Hudson Blvd., Jersey City 6, N. J.

The following transfers to the Associate grade were approved to be effective as of October 1, 1949:

Aburano, F., 2007 Main St., Seattle 44, Wash.
Adams, M. M., 404 Choctaw, Bartlesville, Okla.
Adovnik, F. W., 1013 Lee St., Rock Springs, Wyo.
Alley, R. P., 173 Burbank, Pittsfield, Mass.
Brill, G. J., 836 Masonic Ave., San Francisco 17, Calif.
Buescher, W. E., 220 West Fourth St., Emporium, Pa.
Burgwald, G. M., 10628 Ave. F., Chicago 17, Ill.
Clemens, J. F., 612 College Highway, Evansville, Ind.

(Continued on page 49A)

Membership

(Continued from page 48A)

Crill, P. D., 125-12 Street, Oakland, Calif.
De Carlo, G. J., 1133 Waring Ave., New York 67, N. Y.
Dick, A. I., 283 E. 171 St., New York 57, N. Y.
Espenlaub, W. C., 4607-260 St., Great Neck, L. I., N. Y.
Findley, R., 515 Taylor Ave., Avalon, Pittsburgh 2, Pa.
Ganahl, P. J., 975 Blossom Dr., Santa Clara, Calif.
Gilroy, R. B., 52 Adams Pl., South Weymouth, Mass.
Hamann, O. F., 708 S. George Mason Dr., Arlington, Va.
Hammond, W. H., 1 Collier Rd., N.W., Atlanta, Ga.
Hart, M. T., 437 S. Poplar, Centralia, Ill.
Hartnett, G. C., 25 Kemp Ave., Troy, N. Y.
Hayes, B., 1121C Eastern Ave., Baltimore 21, Md.
Heintz, R. M., 601 University Ave., Los Altos, Calif.
Hollander, J. M., 140 West 86 St., New York 24, N. Y.
Horton, E. J., 435 W. Fairview Ave., San Gabriel, Calif.
Kafalas, C., 22-18—24 St., Astoria 5, L. I., N. Y.
Kugler, F., 291 Rockaway Pkwy., Brooklyn 12, N. Y.
Lamb, H. M., 58 Barnyard Lane, Levittown, Hicksville, L. I., N. Y.
Lang, H. J., 100 Franklin St., Bldg. 1, Apt. A3 Morristown, N. J.
Manry, L. V., Jr., 4324 Betty St., Bellaire, Tex.
Mew, H. Y., 1326 Alewa Dr., Honolulu 29, Hawaii
Meyer, D. R., 7260 Bellaire Ave., North Hollywood, Calif.
Miller, R. L., Box 95, Parkland, Pa.
Mueller, H. W., 2428 N. Holton St., Milwaukee 12 Wis.
Perkins, G. O., 314 W. Zeralda St., Philadelphia 44, Pa.
Peterson, C. D., 11117 Edbrooke Ave., Chicago 28 Ill.
Rau, F. J., 1015 Uvilla St., Pittsburgh 20, Pa.
Reber, J. H., 120 W. Maple Ave., Osborn, Ohio
Robinson, W. C., c/o W. B. Manchester, R.F.D. 3, Ithaca, N. Y.
Ryff, A. S., 22915 Pleasant, St. Clair Shores Mich.
Scher, G. P., 529 Allen St., South Bend 16, Ind.
Schnebbe, A. D., Michael St., Menlo Park, N. J.
Schreiner, R. J., 34 Pier St., Yonkers 5, N. Y.
Stroble, R. R., Box 104, La Porte, Tex.
Strong, J. J., Jr., 28-19 Hobart St., Woodside, L. I., N. Y.
Thompson, C. E., Box 89, Canyon, Tex.
Wells, W. S., Box 521, Rapid City, So. Dak.
White, J. F., Jr., 5459 Iowa St., Chicago 41, Ill.
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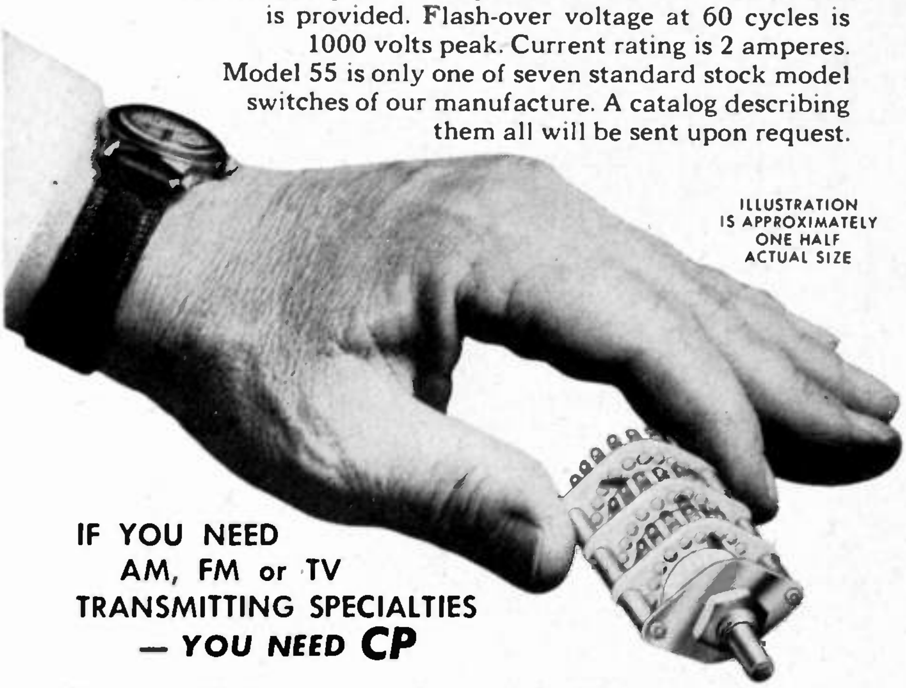

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(Continued on page 51A)

PHYSICISTS AND ENGINEERS

This expanding scientist-operated organization offers excellent opportunities to alert physicists and engineers who are interested in exploring new fields. We desire applicants with experience in the design of electronic circuits (either pulse or c. w.), computers, gyros, antennas, or precision mechanical instruments. A few openings for Junior Engineers and Technicians also exist. This company specializes in research and development work. Laboratories are located in suburbs of Washington, D.C.

JACOBS
INSTRUMENT CO.
4718 Bethesda Ave.
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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

POSITIONS OPEN

(Continued from page 50A)

Agency, Washington 25, D.C. in the field of electronics for electrical engineers, grades P-3, $3727.50 per annum through P-5, $6235.20 per annum. Minimum requirements are a Masters degree or a Bachelors degree from an accredited college or university, plus 1 year of professional engineering experience. All positions are permanent in so far as the Agency is concerned. Address reply to: Chief, Army Security Agency, CS GAS-61, The Pentagon, Washington 25, D.C.

RADAR ENGINEER—PHYSICIST

Must have heavy experience in basic study and research on new radar systems and similar electronic equipment. Excellent opportunity for senior man. Juniors please do not apply. State particulars, reply confidential, to A. Hoffsommer, The W. L. Maxson Corp., 460 West 34 Street, New York 1, N.Y.

PATENT ENGINEER

A manufacturer of electrical and mechanical devices in the New York City area, is in need of a man with legal, patent and engineering training and some experience, to act as liaison between engineers and sales executives and outside patent counsel. In reply give full particulars of experience and training. Address reply to Box 582.

(Continued on page 52A)

PROJECT ENGINEERS

Real opportunities exist for Graduate Engineers with design and development experience in any of the following: Servomechanisms, radar, microwave techniques, microwave antenna design, communications equipment, electron optics, pulse transformers, fractional h.p. motors.

SEND COMPLETE RESUME TO EMPLOYMENT OFFICE.

SPERRY
GYROSCOPE CO.
DIVISION OF
THE SPERRY CORP.
GREAT NECK, LONG ISLAND

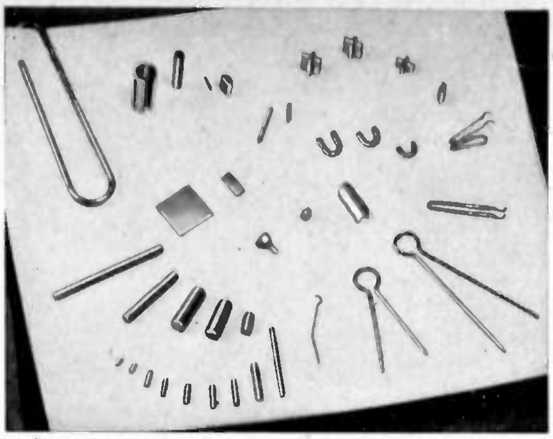
An EASY and ACCURATE Way to Measure Audio Frequency Voltages

● Designed for the measurement of AC Voltages from .001 Volt to 100 Volts over a frequency range of 10 to 150,000 cycles. ● Accuracy of readings is ±2% at any point on the scale. ● Very stable calibration—unaffected by changes in line voltage, tubes or circuit constants. ● Range switching in decade steps—easy to use—only ONE scale to read. ● Output jack and output control provided so that Voltmeter can be used as a high-gain (70 DB) high-fidelity amplifier. ● Accessories available to extend readings up to 10,000 Volts and down to 10 microvolts. ● Precision Shunt Resistors convert Model 300 Voltmeter to very sensitive direct-reading milliammeter. ● Write for complete data.

PRICE $200.00

BALLANTINE LABORATORIES, INC.
BOONTON N. J. • U. S. A.

In addition to the Model 300 Voltmeter, Ballantine Laboratories also manufacture Battery Operated Electronic Voltmeters, R. F. Electronic Voltmeters, Peak to Peak Electronic Voltmeters, and the following accessories—Decade Amplifiers, Multipliers, Precision Shunt Resistors, etc.



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ELECTRICAL CONTACTS ON POTENTIOMETERS, SLIP RINGS, RELAYS AND SWITCHES

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SLIDING CONTACTS FOR POTENTIOMETERS

PALINEY #7 is being used for a contact material on potentiometers wound with a nickel-chrome alloy resistance wire. This combination is consistently producing units with life of better than one million cycles and maintained accuracy of 0.1% or better throughout the life of the unit.

NEY-ORO #28

SLIP RING BRUSHES

NEY-ORO #28 is a special alloy developed as a contact brush material for uses against coin silver slip rings. Laboratory tests and reports from users indicate life of better than 10 million revolutions with no electrical noise.

Write or telephone (HARTFORD 2-4271) *our Research Department*

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SPECIALISTS IN PRECIOUS METAL METALLURGY SINCE 1812

20NY49

POSITIONS OPEN

(*Continued from page 51A*)

ELECTRONIC ENGINEER

College graduate, E.E., with 3 or more years experience in electronic circuit design. Expanding consulting firm, specialists in custom building of industrial electronic instruments, has opening for fast-thinking engineer who can apply theoretical background to practical problems. Location Detroit, Michigan. Box 584.

ELECTRICAL ENGINEERS

Graduates in electrical engineering or physics with at least 3 years design and manufacturing experience on radio transmitters or radar equipment. Non-graduates having at least 6 years similar practical experience will be considered also. Address reply to Personnel Manager, RCA Victor Company, Ltd., Montreal, Canada.

SALES ENGINEER

Sales Department of small company engaged in research, development and manufacture of instruments specializing in the radiation field, requires an engineer with sales promotion experience demonstrating qualities of aggressive leadership, ability in business, business correspondence and a good knowledge of electronics. Must be capable of planning sales functions and following them through to a successful conclusion. Con-

(*Continued on page 53A*)

V.L.F.!

Very Low Frequencies covered by the . . .

STODDART NM-10A RADIO INTERFERENCE AND FIELD INTENSITY METER

- MEASURES radiated and conducted signals, including pulse or random interference.
- RANGE—14 kc to 250 kc.
- SENSITIVITY — Field strength using rod antennas one microvolt-per-meter to 2 volts-per-meter. Field strength using shielded loop antennas 10 microvolts-per-meter to 100 volts-per-meter. As a two-terminal voltmeter, either balanced or unbalanced, one microvolt to one volt.
- READS directly in microvolts and db.
- A.C. POWER SUPPLY REQUIREMENTS 105 to 125 volts or 210 to 250 volts A.C. Single phase source may be ANY FREQUENCY BETWEEN 50 CPS AND 1600 CPS. No shock hazard.
- GRAPHIC RECORDER included with versatile complement of accessories.

Write for complete technical data

STODDART AIRCRAFT RADIO CO.

Main office and plant: 6644 Santa Monica Blvd. Hollywood 38, Calif. Phone: Hillside 9294

8-247 General Motors Bldg. Detroit 2, Michigan Phone: Trinity 1-9260

1346 Connecticut Ave. duPont Circle Bldg. Washington 6, D. C. Phone: Hudson 7313

ENGINEERING OPENINGS

at

DuMONT TELEVISION

- **ELECTRONIC PROJECT ENGINEER**
  Development & design of UHF tuning devices. Must have related experience.
- **ELECTRONIC ENGINEER**
  Television receiver circuit design & development; preferably experienced in synchronizing problems.
- **MECHANICAL PROJECT ENGINEER**
  Development & design of UHF tuning devices. Must be familiar with die casting & injection molding technique.
- **MECHANICAL ENGINEER**
  Experienced in product design of home radios or television receivers.

Call in person or write:

Allen B. DuMont Laboratories

Personnel Dept., Television Receiver Division

East Patterson, N.J.

Interviews outside of N.Y. area may be arranged.

Positions Open

(Continued from page 52A)

tact, Berkeley Scientific Co., 6th & Nevin, Richmond, California.

TELEVISION ANTENNA ENGINEER

Well established New York City manufacturer has an immediate opening for engineer with thorough up-to-the-minute knowledge of commercial television receiving antenna theory, design and construction. Duties include sales engineering and execution of graphs and test patterns. Write Box 583.

VACUUM TUBE ENGINEER

4-5 years experience microwave measurements, circuits or electronics. Some tube engineering and construction experience helpful. Ability to direct development projects. B.S. in E.E. or physics minimum. Large company No. New Jersey. Write in detail Personnel Dept. Box 586.

Positions Wanted By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants, and to avoid overcrowding of the corresponding column, the following rules have been adopted:

(Continued on page 54A)

RCA VICTOR
Camden, N. J.

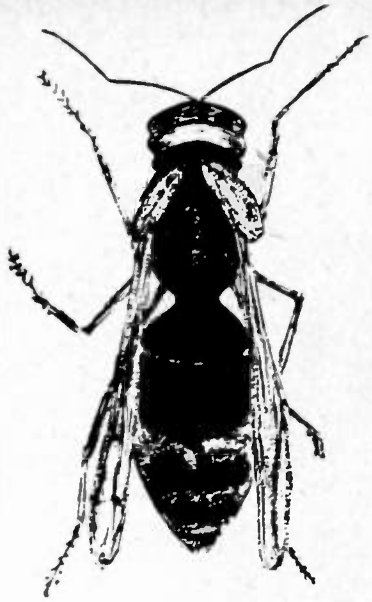
Requires Experienced Electronics Engineers

RCA's steady growth in the field of electronics results in attractive opportunities for electrical and mechanical engineers and physicists. Experienced engineers are finding the "right position" in the wide scope of RCA's activities. Equipment is being developed for the following applications: communications and navigational equipment for the aviation industry, mobile transmitters, microwave relay links, radar systems and components, and ultra high frequency test equipment.

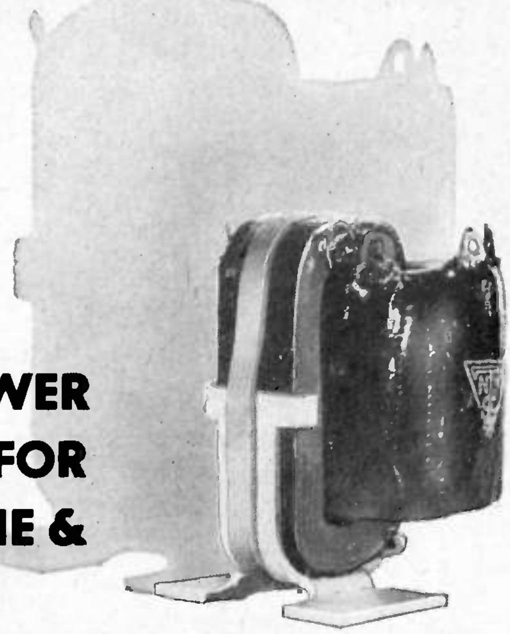
These requirements represent permanent expansion in RCA Victor's Engineering Division at Camden, which will provide excellent opportunities for men of high caliber with appropriate training and experience.

If you meet these specifications, and if you are looking for a career which will open wide the door to the complete expression of your talents in the fields of electronics, write, giving full details to:

National Recruiting Division
Box 147, RCA Victor Division
Radio Corporation of America
Camden, New Jersey



THIS IS THE HORNET



A NEW MINIATURE POWER TRANSFORMER FOR USE IN AIRBORNE & PORTABLE EQUIPMENT

Illustration shows relative size of HORNET and conventional transformers of comparable capacity.

FEATURING

SMALLER SIZE
than any previous design, through the use of newly developed class H insulating materials, and design techniques. As shown above, HORNET transformers are only about one-fourth the size of similarly rated conventional transformers.

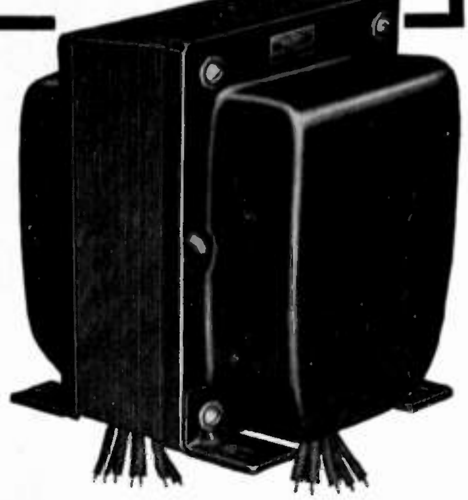
GREATER POWER OUTPUT
because of improved design and construction. HORNET transformers operate with unimpaired efficiency at high temperatures, and are suitable for operation at ambient temperatures as high as 150 deg. C. High output plus smaller size and lighter weight make these units ideal for use in airborne and portable equipment.

MEETS JAN SPECIFICATIONS
HORNET transformers are designed and built to meet requirements of current JAN T-27, and equivalent specifications.

Write for descriptive bulletin of sizes and specifications

NEW YORK TRANSFORMER CO., INC.
ALPHA, NEW JERSEY

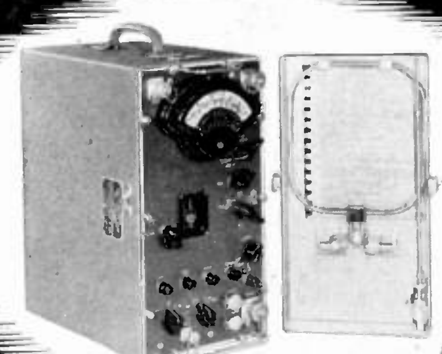
NEW T. V. IDEAS
NEED ACME ELECTRIC TRANSFORMER *performance*



New engineering ideas, to advance the reception qualities of Television, need better than average transformer performance. Acme Electric engineers will assist your ideas by helping you design a transformer, exactly in accordance with your needs.

Acme ACME Electric
TRANSFORMERS

ACME ELECTRIC
Corporation
4412 Water St. Cuba, N.Y., U.S.A.



Portable Precision for the Field Engineer!

STODDART NM-20A
RADIO INTERFERENCE AND FIELD INTENSITY METER

- A portable unit that you can DEPEND upon! Designed especially to withstand the rigors of all-weather field operation and yet provide reliable performance.
- Measures FIELD INTENSITIES of radio signals and r.f. disturbances using either a rod antenna or a rotatable loop antenna.
- May be used as a two-terminal r.f. voltmeter (balanced or unbalanced), frequency selective over the CONTINUOUS RANGE 150 kc to 25 mc.
- ONE MICROVOLT SENSITIVITY as a two-terminal voltmeter; 2 microvolts-per-meter using rod antenna.
- Operates from self-contained dry batteries or external A.C. power unit providing well-regulated filament and plate supplies.

Write for complete technical data

STODDART AIRCRAFT RADIO CO.

Main office and plant:
6644 Santa Monica Blvd.
Hollywood 38, Calif.
Phone: Hillside 9294

8-247 General Motors Bldg.
Detroit 2, Michigan
Phone: Trinity 1-9260

1346 Connecticut Ave.
duPont Circle Bldg.
Washington 6, D. C.
Phone: Hudson 7313

Positions Wanted

(Continued from page 53A)

The Institute publishes free of charge notices of positions wanted by I.R.E. members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The Institute necessarily reserves the right to decline any announcement without assignment of reason.

DEVELOPMENT ENGINEER

Eight years broad experience, research and development. Servos, auto, controls, analogue computers, industrial electronics. Excellent theoretical ability well balanced by laboratory and experimental work. High scholastic standing in college. Desires position as Senior Development Engineer or Assistant Director of Research. Box 321 W.

ENGINEER

B.E.E. Cum Laude, C.C.N.Y. February 1948. Tau Beta Pi and Eta Kappa Nu. Age 24. 1½ years design, development and production experience as project engineer. Main field antennas. Desires position in New York Metropolitan area. Box 322 W.

ENGINEER

B.E.E. 1943 C.C.N.Y., M.S. 1948 Columbia University. Age 28. Married. 6 years experience in electronic research and product engineering. 1st class radio telephone license. New York area. Box 323 W.

(Continued on page 55A)

NATIONAL UNION RESEARCH DIVISION

There are several desirable openings for experienced

PHYSICISTS
and
ENGINEERS

capable of handling the design and development of electron tubes and UHF circuits.

Our growing organization can offer excellent prospects for security and advancement to qualified personnel.

Interested applicants are invited to send their résumé to:

Divisional Personnel Manager
National Union Research Division
350 Scotland Road, Orange, N.J.

Positions Wanted

(Continued from page 54A)

JUNIOR ENGINEER OR LABORATORY TECHNICIAN

Graduate of R.C.A. Institute Technology course. 4 years commercial experience in electronic laboratory technique. Worked for Alexander Fowler and Allen B. Dumont laboratories. Former Air Corps instrument instructor and specialist. Hold 1st class F.C.C. radio and telephone license. Experienced in all phases of laboratory work on video or electronic circuitry. Age 27. Married. Desires New York City area or Long Island. Box 325 W.

ENGINEER

B.E.E. University of Florida, September 1949. Communications major. 5 years Army service. Age 27. Married, no children. Willing to travel. Box 326 W.

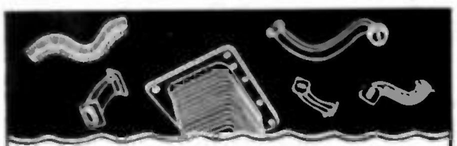
RADIO ENGINEER

B.S.E.E. University of California, September 1949. Communications major. Age 23. 2 years Navy ETM. Desires position in radio, television, or technical writing. Prefer West coast. Box 327 W.

MICROWAVE ENGINEER

B.E.E. 1943, graduate evening student. Married. Age 27. 3½ years research and development experience on microwave transmission components and systems. 2 years Army P.P.M. radar link work. Desires research or development work vicinity New York City. Box 328 W.

(Continued on page 56A)



FLEXAGUIDE

A flexible waveguide with an electrically continuous FLEXIBLE CONVOLUTED bellows innercore protected by a specially LOW TEMPERATURE flexible molded jacket. It is PRESSURE tight and electrically correct for all conditions of bending and flexing. Standing Wave Ratio is equivalent to standard rigid waveguide assembly throughout the flexing cycle and has excellent attenuation characteristics.

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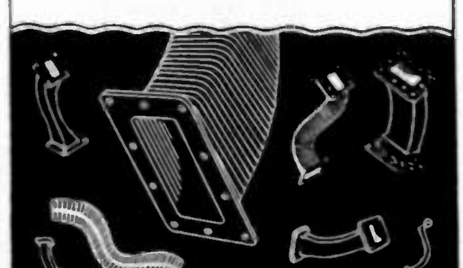
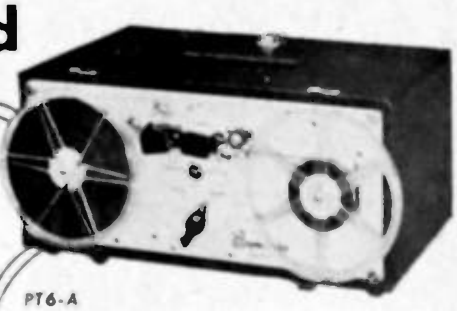


FOR GREATEST *Flexibility* IN PROFESSIONAL TAPE RECORDING

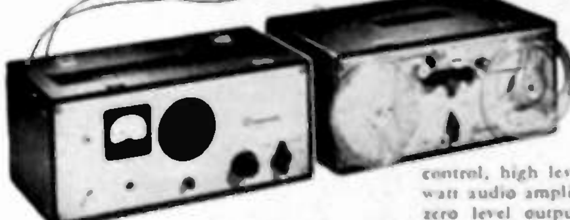
Get Magnecord UNIT CONSTRUCTION

FM STUDIO QUALITY
PLUS PORTABILITY
AND ECONOMY

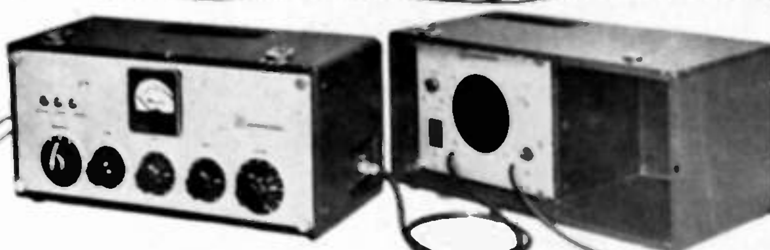
Only Magnecord gives you the economy and adaptability of unit construction plus really high fidelity. The several Magnecord units combine to meet every studio and remote recording demand. Buy only those you need. Carry and use them only where and when you need them.



PT6-A
Recorder Mechanism
is the heart of Magnecord combinations for studio and remote recording. Weighs 26 lbs. in carrying case, easily removable for rack mount. Quick-change capstans for recording at 7½ or 15 inches/sec. High speed rewind. Frequency response 40 to 15,000 cps ± 2 db. $278

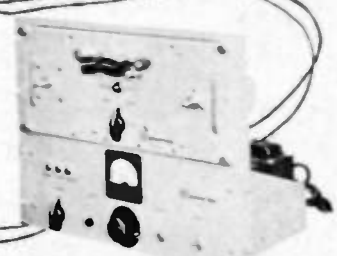


PT6-JA Recorder & Amplifier provides complete portable facilities for professional quality reproduction at a new low price. Includes PT6-A Recorder plus amplifier containing low impedance mike input with gain control, high level input, monitor speaker and 10 watt audio amplifier with jack for external speaker, zero level output terminal, VU type meter. $499.50



PT6-P Portable Amplifier is a light weight record-playback-remote amplifier plus power supply designed for use with PT6-A Recorder. 3 low level independently mixed mike inputs plus bridging input for use with a line level input. Monitor amplifier and small speaker in power supply section. $462

PT6-R Rack-Mount Amplifier is a high fidelity, single channel amplifier for use with existing audio amplifiers and PT6-A Recorder. Uses only 14 inches of rack space. Recorder can be removed from carrying case and fastened to flush mounting in seconds. (Recorder not included). $383



Write for complete specifications and name of nearest dealer.

Magnecord, INC., CHICAGO 1, ILL.
360 NORTH MICHIGAN AVENUE
World's Largest and Oldest Manufacturers of Professional Magnetic Recorders.

SOLDERING IS A CINCH

when you know these SIMPLE TRICKS!

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No matter how much you know about soldering, there's always a trick that will make it easier. This little 20-page pocket guide is crammed full of such time-and-trouble savers.

Without wasting words, it covers the whole soldering operation—points out DO's and DON'T's—refreshes your memory on difficult points—suggests methods that help you work faster. Yet there's no hard studying, no tough technical talk. Every word is plain everyday English and every point is made clear by easy-to-understand illustrations.

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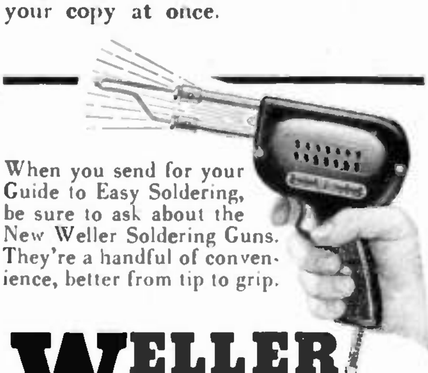


When you send for your Guide to Easy Soldering, be sure to ask about the New Weller Soldering Guns. They're a handful of convenience, better from tip to grip.

WELLER MANUFACTURING COMPANY

Weller Mfg. Co., 821 Packer St., Easton, Pa.

Enclosed find ten cents (10c) for which please send my copy of the Weller "Soldering Tips".

☐ *I am also interested in the new Weller Soldering Guns. Please send Catalog Bulletin.*

Name ______

Address ______

City ______ *State* ______

Positions Wanted

(Continued from page 55A)

ENGINEER

B.S. in Radio Engineering. 3 years Airborne radio and radar maintenance with U.S.M.C. 19 months audio repair with Sound Scriber Distributor. Age 27. Married, 1 child. Desires work in U.H.F. field. Box 329 W.

ELECTRICAL ENGINEER

B.S.E.E. 1949. Single. Age 23. University of Illinois graduate, upper quarter. Knowledge of Greek. Desires position in Engineering Department of American firm in Europe, preferably Greece. Available immediately. Box 330 W.

JUNIOR ENGINEER

B.S.E.E. Columbia University, June 1949. Age 28. Single. Desires promising starting position in design development or production, anywhere in United States. Box 331 W.

COMMUNICATIONS ENGINEER

B.S.E.E. 1947. 2 years carrier telephony, Signal Corps radio-link. Age 27. Married, 2 children. Now employed in Boston, wants research, design, station construction, sales engineering, teaching or technical writing in central to southern Maine. 2 years design of high-frequency and microwave antennas. Box 332 W.

ELECTRONIC TECHNICIAN

High school graduate. 2 years U. S. Coast Guard radio and radar school. 2 years at RCA Institutes. 5 years experience in radio and radar maintenance and installation with U. S. Coast Guard. 3 years with American Airlines as radar technician in transmitter band experimental radar laboratory. Box 334 W.

JUNIOR ELECTRICAL ENGINEER

B.S.E.E. June 1949, Bucknell University. Married. 26 months experience as Navy electronic technician. No other experience, but willing to learn. Desires position in electronics in New York City area. Box 352 W.

COMMUNICATIONS ENGINEER

Graduate of Ohio State University, December 1947 with B.E.E. Married. Age 23. Experience: 1½ years with automatic switching equipment, 1 year part-time in electronics development. Desires position in electronic or communications field in Northern New Jersey area. Box 353 W.

ELECTRONIC ENGINEER

B.S.E.E. June 1949. State College of Washington. Single. Age 29. 2 years radio operation experience in Army. Desires position in electronic field. Prefer Pacific coast area. Box 355 W.

ENGINEER

M.S. in physics, June 1949, Fordham University. 2 years experience as part time instructor in general physics lab. Graduate of Navy electronic training program. Age 24. Engaged. Desires position in sales or development, preferably in New York area. Box 356 W.

ENGINEER

B.S.E.E. June 1949. University of Missouri. Tau Beta Pi, Eta Kappa Nu, Pi Mu Epsilon. Age 25. Single. Some radio experience in Signal Corps. Desires communications or electronic work. Anywhere in U.S. Box 357 W.

(Continued on page 57A)

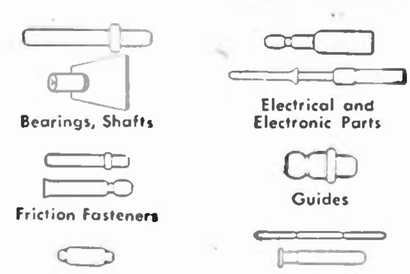
Which pin saves the money?

Choose right and make Big Savings on SMALL METAL PARTS

COSTS HALVED! Instead of turning and drilling parts like these from solid rod, or stamping and forming them, the BEAD CHAIN MULTI-SWAGE Process automatically swages them from flat stock. By doubling the production rate and eliminating scrap, this advanced process can save you as much as fifty percent of the cost of other methods.

The BEAD CHAIN MULTI-SWAGE Process produces a wide variety of hollow or solid metal parts—beaded, grooved, shouldered—from flat stock, tubing, rod, or wire—of any metal. Sizes to ¼" dia. and 1½" length.

GET COST COMPARISON ON YOUR PARTS—If you use small metal parts in quantities of about 100,000, don't overlook the almost certain savings of this high-speed, precision process. Send sketch, blueprint or sample part and our engineers will furnish facts about Multi-Swage economy. Or, write for Catalog. The Bead Chain Manufacturing Co., 60 Mountain Grove St., Bridgeport, Conn.



BEAD CHAIN MULTI-SWAGE PROCESS

TRADE MARK

Positions Wanted

(Continued from page 56A)

JUNIOR ENGINEER

Graduated June 1949 from Newark College of Engineering, Newark, New Jersey, with a B.S.E.E. degree. 3 years training and experience as radio technician in the Signal Corps. Age 30, married. Desires employment as electrical or electron engineer particularly in the high frequencies anywhere in U.S. or Canada. Box 358 W.

SALES ENGINEERING OR TECHNICAL ADMINISTRATION

Well-known, highly experienced and realistic engineer-consultant-writer in radio-electronic nucleonic fields, presently in California, seeks new employment anywhere. Particularly qualified in sales engineering, public relations and promotion of new developments. Very familiar with governmental, industrial and educational electronic circles nationwide. Box 359 W.

JUNIOR ENGINEER

Graduate student of radio and television desires junior engineering position in electronics industry. Particularly interested in audio or recording field. Broadcast experience. Age 23, married, child. Willing to travel occasionally. Prefer midwest or south. Box 360 W.

ENGINEER

B.E.E. June 1949, New York University, communications major. Age 23. Single. Eta Kappa Nu, Tau Beta Pi. Would like to start career in any of following fields: Hf, VHF and/or micro-

(Continued on page 58A)

Senior Electronic Circuit Physicists

for advanced Research and Development

Minimum Requirements:

1. M.S. or Ph.D. in Physics or E.E.
2. Not less than five years experience in advanced electronic circuit development with a record of accomplishment giving evidence of an unusual degree of ingenuity and ability in the field.
3. Minimum age 28 years.

HUGHES AIRCRAFT COMPANY

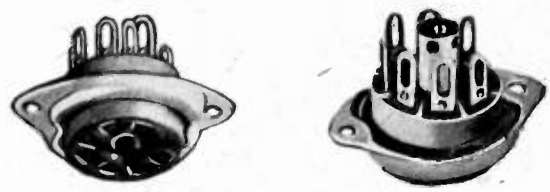
(Mr. Jack Harwood)

Culver City, California

Announcing

MYCALEX 7 Pin Miniature Tube Sockets

For the first time a miniature tube socket of glass-bonded mica has been produced successfully by injection molding. It permits closer tolerances, low dielectric loss with high dielectric strength, high arc resistance and dimensional stability over wide humidity and temperature ranges. The technical skill and research of Mycalex Corp. of America has made it possible to produce insulating materials with extremely low loss factors at competitive prices.



Above: Complete 7 pin miniature Mycalex socket. Actual size, two views.

"Mycalex 410" was developed for applications requiring close dimensional tolerances not possible in ceramics and with much lower loss factor than mica filled phenolics with the advantage in economy.

"Mycalex 410X" was developed to compare favorably with general purpose bakelite in economy but with a loss factor of only about one-fourth of that material.

The following ratings show the difference between Mycalex 410 and Mycalex 410X miniature tube sockets.

MYCALEX 410 (color grey)		MYCALEX 410X (color lt. green)
600 V.ac	Rated Working Voltage	600 V.ac
.015	Insulation loss factor (at 1 M.C.)	.083
10,000 megohms	Insulation resistance (Minimum)	10,000 megohms
	Safe operating temperatures:	
80° C.	Brass contacts	80° C.
375° C.	Socket body	375° C.

These superior sockets are now available, manufactured to high quality standards and fully meet RMA recommendations. We would be glad to have our engineers consult with you on your particular design problems. Write for prices, complete data sheet and samples to:

Mycalex Tube Socket Corporation

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

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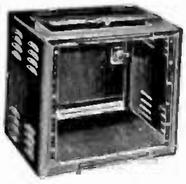
R-1744 · AC-406 · CR-1728 · AC-407 · CR-1745 · AC-408 · CR-1741 · AC-409 · CR-17
CR-1739 · CR-1743
R-1740 · AC-403 · CR-1742 · AC-421 · CR-1739 · AC-404 · CR-1743 · AC-422 · CR-17

BUD HAS YOUR NUMBER IN CABINETS AND RACKS

In response to wide spread demand Bud has now augmented its already large line of Deluxe Cabinet Racks and Aluminum Chassis by the addition of several new sizes. The table below lists these new sizes as well as the old ones. Now, more than ever, Bud is able to meet your needs in sheet metal as well as other radio and electronic components.



BUD DE LUXE CABINET RACKS

These cabinet racks have rounded corners and attractive red-lined chrome trim. There is a recessed, hinged door on the top with a snap catch. These racks are made of heavy gauge steel and are of sturdy construction. The five large sizes have a hinged rear door, while the small sizes have a welded panel in the rear.

Adequate ventilation is assured by means of louvered sides and a two inch opening in the bottom of the back extends the entire width.

"NO-SCRATCH" EXTENDED METAL FEET ARE EMBOSSED ON THE BOTTOM TO MINIMIZE MARRING OF A TABLE TOP. Racks are furnished in either black or grey wrinkle finish. Depth 14¾" width 22" Will fit standard 19" panels.

Catalog No.	Overall Height	Panel Space	Shipping Wt.	Dealer Cost
CR-1741	10 9/16"	8¾"	29 lbs.	$10.05
CR-1740	12 5/16"	10½"	31 lbs.	11.32
CR-1742	14 1/16"	12¼"	32 lbs.	12.25
CR-1739	15 13/16"	14"	36 lbs.	13.85
CR-1743	19 5/16"	17½"	40 lbs.	16.77
CR-1727	22 13/16"	21"	45 lbs.	18.00
CR-1744	28 3/16"	26¼"	50 lbs.	19.20
CR-1728	33 5/16"	31½"	55 lbs.	21.20
CR-1745	36 13/16"	35"	60 lbs.	21.57

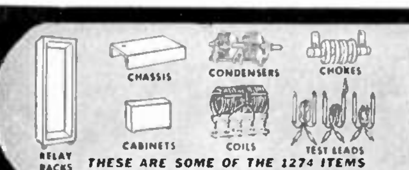
BUD ADD-a-RACK SERIES
Write for literature on this newest Bud product. Find out how you can get more panel space in less floor area at lower cost.

BUD ALUMINUM CHASSIS

The construction and design of these chassis is exactly the same as our steel chassis. The aluminum chassis are welded on government approved spot welders that are the same as used in the welding of aluminum airplane parts. The gauges in table below are aluminum gauges. As a result, you can depend on **BUD** Aluminum Chassis to do a perfect job. Etched Aluminum finish.

Catalog Number	Depth	Width	Height	Gauge	Dealer Cost
AC-402	5"	7"	2"	18	$ .69
AC-403	5"	9½"	2"	18	.81
AC-421	5"	9½"	3"	18	.89
AC-404	5"	10"	3"	18	.99
AC-422	5"	13"	3"	18	.98
AC-405	7"	7"	2"	18	.81
AC-406	7"	9"	2"	18	.90
AC-407	7"	11"	2"	18	.96
AC-408	7"	12"	3"	18	1.14
AC-409	7"	13"	2"	18	1.02
AC-411	7"	15"	3"	16	1.68
AC-423	7"	17"	3"	16	1.43
AC-424	8"	12"	3"	16	1.38
AC-425	8"	17"	2"	16	1.52
AC-412	8"	17"	3"	16	1.77
AC-413	10"	12"	3"	16	1.44
AC-414	10"	14"	3"	16	1.92
AC-415	10"	17"	2"	16	1.80
AC-416	10"	17"	3"	16	2.04
AC-426	11"	17"	2"	14	1.89
AC-417	11"	17"	3"	14	2.40
AC-418	12"	17"	3"	14	2.52
AC-419	13"	17"	2"	14	2.25
AC-420	13"	17"	3"	14	2.67
AC-427	10"	17"	4"	14	2.30
AC-428	13"	17"	4"	14	3.05

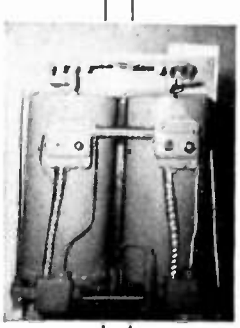
Prices are 10% higher west of the Mississippi River.



THESE ARE SOME OF THE 1274 ITEMS AVAILABLE FROM BUD RADIO, INC.

BUD RADIO, INC.
BUD
2110 E. 55th ST. • CLEVELAND 3, OHIO

The
"KELLOGG DEHYDRATOR"



An all purpose self-reactivating dehydrating unit. To be used for removing moisture from gases. Numerous applications in the fields of Physics, Electronics and Chemistry. Dual insulated tanks with thermostatically controlled heating elements. Complete with 20 lbs. of Silica gel., heating elements, shut-off and safety valves.

$62.50
F.O.B., N.Y.

INTERSTATE Appliance Co., Inc.
Dept. KD, 600 Broadway, NEW YORK 12, N. Y.

Positions Wanted

(Continued from page 57A)

wave communications; instrumentation; general electronics. Will work in design, development or manufacture. Location immaterial. Richard A. Davidson, 2043 Holland Ave., New York 60, N.Y.

ELECTRONIC ENGINEER

B.E.E. Polytechnic Institute of Brooklyn, June 1948. Experience includes 1 year as communications project engineer and several months test and development of radiation circuits and equipment. Desires position as project or development engineer in New York City or vicinity. Box 361 W.

ELECTRONIC ENGINEER

M.S.E.E. University of Illinois, 1949; B.S.E.E. Purdue University, 1948. Age 27. Single. 2 years teaching experience in electronics. Member Eta Kappa Nu. 1st Radiotelephone and Class A Amateur Licenses. Desires employment in electronic research and development. Box 362 W.

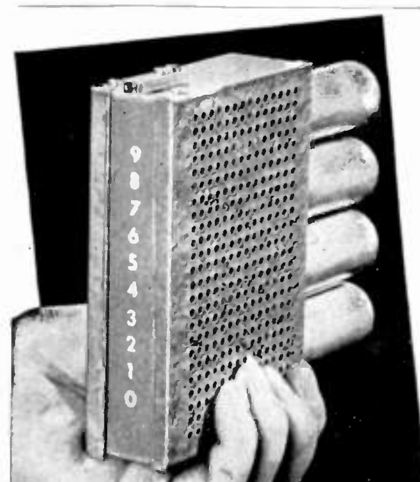
ENGINEER

B.E.E. 1943. Postgraduate work in servomechanisms, circuit analysis, etc. 6 years experience in electronic instrumentation, guidance and control of guided missiles, digital and analogue computers. Box 363 W.

TELEVISION ENGINEER

B.S.T.E. November 1948. American Television Institute of Technology. Age 29, married. 3 years Naval experience in electronics; 6 months in industry. 1st class FCC license. Desires position as TV station engineer. Box 364 W.

(Continued on page 61A)



THE NEW BERKELEY MODEL 700 DECIMAL COUNTING UNIT

This unique packaged component is *easily built into* your apparatus. It has *true decimal reading*, and simple binary circuit with reliable *automatic interpolation*. Miniature size. Moderate price. Immediate shipment.

Send for Bulletin DCU-116

Berkeley Scientific Company
SIXTH AND NEVIN AVE · RICHMOND, CALIFORNIA

RAWSON METERS

Types:
501A
501C

Accuracy:
½ of 1%

MULTIMETERS and REGULAR METERS
AC and DC types, high accuracy, multiple ranges. 2 microamperes to 1 ampere DC. 2 milliamperes to 3 amperes AC.

ELECTROSTATIC VOLTMETERS
Ranges 100 v. to 35,000 v. AC or DC Resistance exceeds million megohms. Can measure static electricity.

FLUXMETERS
Laboratory and production measurements on magnets and magnetic circuits. Single push button return-to-zero.

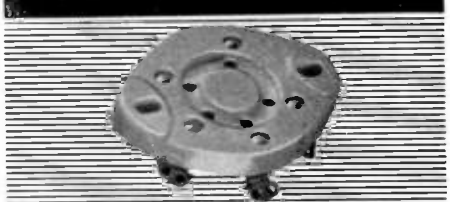
WATTMETERS
High sensitivities, low power factors. New types soon to be announced.

SINE—COSINE POTENTIOMETERS

RAWSON ELECTRICAL INSTRUMENT COMPANY
118 Potter St. Cambridge, Mass.

Representatives
Chicago Los Angeles

JOHNSON SOCKETS . . .



always a perfect choice!

The JOHNSON 122-225 wafer pictured above is insulated with glazed grade L-4 steatite. Contacts are brass with steel spring, cadmium plated. Mounted against phenolic washers in molded recesses to prevent movement. Rivets countersunk, mounting holes bossed to permit sub-panel mounting. Locating grooves facilitate tube insertion. Available also in 4, 6, and 7 contact as well as octal.

All JOHNSON sockets are of equally outstanding quality. Get the best—get JOHNSON sockets! See Your JOHNSON Dealer

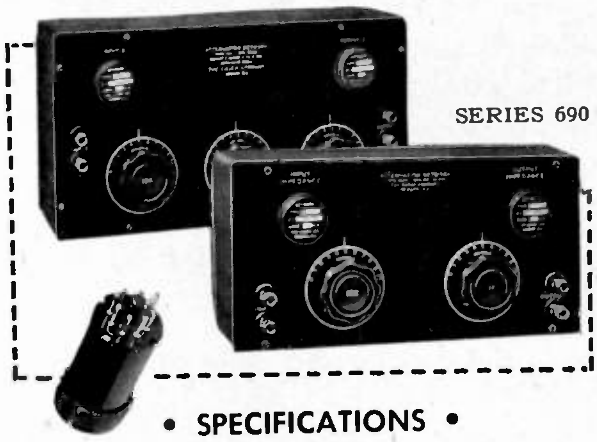
E. F. JOHNSON CO.
WASECA, MINNESOTA

DAVEN

PORTABLE ATTENUATION NETWORKS

The ELECTRONIC INDUSTRY'S STANDARD

These standards of attenuation are designed for use in general laboratory and production testing, where ease of operation and reliability are important. An outstanding feature of these units is the use of "plug-in" impedance adjusting fixed pads on both input and output. Thus, either, or both, input and output terminal impedances can be readily altered by inserting the proper fixed network.



SERIES 690

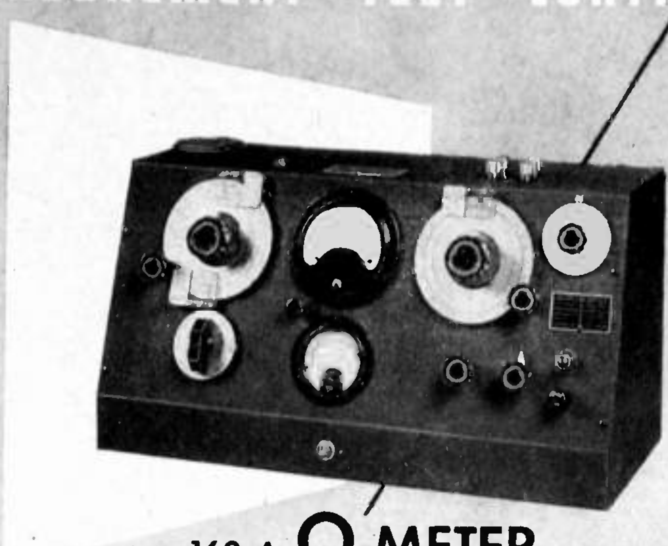
• SPECIFICATIONS •

- **CIRCUITS:** "T" or "Balanced H".
- **IMPEDANCES:** Three base impedances of 150, 500, and 600 are available; however, input and output impedances may be changed by varying the "plug-in" pads.
- **ACCURACY:** Resistors are calibrated to ±1%. Greater accuracy on request.
- **ATTENUATION RANGE:**
  2 dial models—0 to 110 DB in steps of 1 DB.
  3 dial models—0 to 111 DB in steps of 0.1 DB.
- **FREQUENCY RANGE:** 0 to 50,000 cycles. Other models available to 200 KC.
- **MOUNTING:**
  Portable models in hand rubbed walnut cabinets.
  Rack models with slip-on metal dust covers.

For further information write to Dept. IE-8

THE **DAVEN** CO.
191 CENTRAL AVENUE
NEWARK 4, NEW JERSEY

MEASUREMENT • TEST • CONTROL



160-A Q METER

The 160-A Q-Meter is unexcelled for laboratory and development applications, having received world wide recognition as *the* outstanding instrument for measuring Q, inductance, and capacitance at radio frequencies.

Frequency Range: 50 kc. to 75 mc. (8 ranges)
Q Measurement Range: 20 to 250 (20 to 625 with multiplier)
Range of Main Q Capacitor: 30-450 mmf.
Range of Vernier Q Capacitor: +3 mmf., zero, −3 mmf.

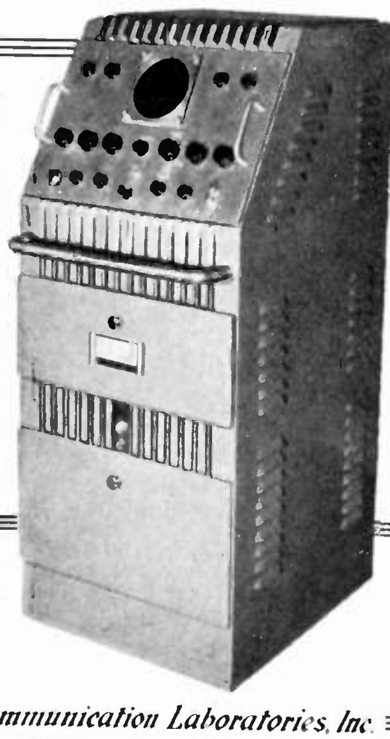
BOONTON RADIO *Corporation*
BOONTON · N · J · U · S · A

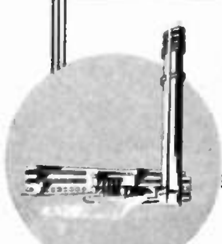
A limited quantity of these instruments is available for immediate delivery.

ACCURATE OBSERVATION OF WAVEFORMS FROM 10 CYCLES TO 50 MC PER SECOND

50 MC WIDEBAND VIDEO OSCILLOSCOPE FTL-32A

- Vertical amplifier bandwidth of 10 cps to 50 mc.
- High deflection sensitivity over the entire bandwith.
- Low-capacity probe maintaining high sensitivity.
- Sweep time as fast as one micro-second.

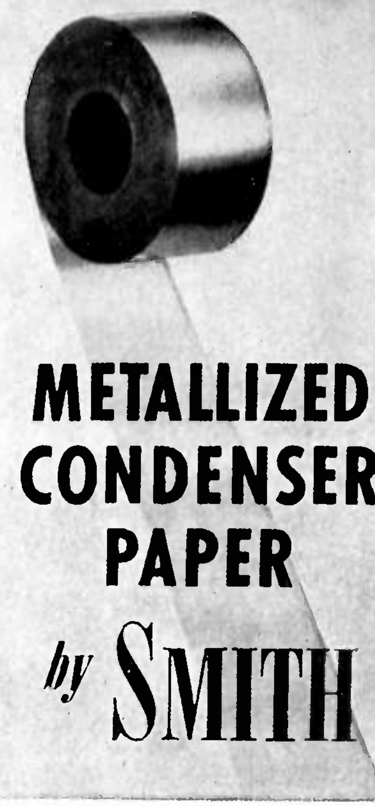




Write for complete FTL-32A brochure.

Federal Telecommunication Laboratories, Inc.

An I.T.&T. Associate
500 Washington Avenue
Nutley 10, N.J.



METALLIZED CONDENSER PAPER *by* SMITH

When you use Metallized Condenser Paper, it is possible to:

1. Manufacture a one-layer condenser.
2. Save 75% space.
3. Use more economical neutral oils.
4. Use less impregnating materials.
5. Eliminate the use of foil electrodes.
6. Use simpler capacitor winding machines.
7. Eliminate breakdown effects through self-healing.
8. Obtain higher insulation resistance, higher dielectric strength.

Get all the facts on Smith Metallized Condenser Paper as it applies to your industry. A card, letter or phone call to Smith Paper, Inc., Lee, Massachusetts (telephone Lee 7) will bring you the information you seek, without obligation on your part.

Manufacturers of Condenser Papers

SMITH PAPER, INC.
LEE, MASSACHUSETTS

SMITH

Positions Wanted

(Continued from page 58A)

ELECTRONIC ENGINEER

B.S. (Physics/Math) February 1949. Married, two children. Graduate CREI 2 years experience test and research. 2 years experience Navy radio technician. 3 years experience Broadcast radio engineer 1st class Radiotelephone license. Amateur License. Box 365 W.

ELECTRONIC ENGINEER

Electronic engineer, M.E.E. 2 years research and development experience in servomechanisms electronics and radar technique plus laboratory instruction in electrical engineering and Navy radar experience. Top scholastic record, Tau Beta Pi, Eta Kappa Nu. Box 369 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 22A)

DC Driven DC-AC Chopper

A recent addition to the line of dc-ac choppers manufactured by **Stevens-Arnold Inc.**, 22 Elkins St., South Boston, Mass. is a dc driven type.

This self-excited chopper is an electromechanical vibrator to be used as a modulator, rectifier demodulator, or square-wave generator. This design was arranged for those applications where dc is preferred to ac as the source of coil excitation. Actually it is a combination of an SPDT and an SPST chopper.

(Continued on page 63A)

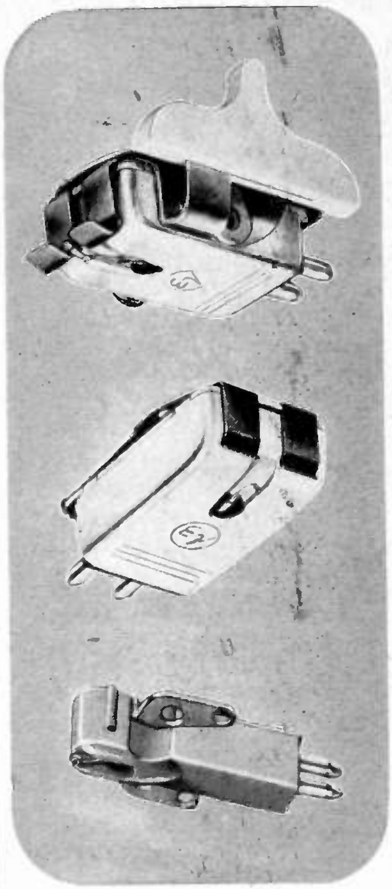
Recent Catalogs

• • • A new booklet describing the advantages and new applications of vacuum coating chambers for industrial purposes has been printed by **Vacuum Equipment Div., Distillation Products, Inc.**, 755 Ridge Rd. W., Rochester 13, N. Y.

• • • The 1950 Twin-Trax catalog, a 16-page listing of more than 30 new tape recorders, is obtainable from **Twin-Trax Div., Amplifier Corp. of America**, 398-1 Broadway, New York 13, N. Y.

EVERYTHING you want . . . with EV Modern Phono Cartridges

- TAILORED FREQUENCY RESPONSE
- HIGHER COMPLIANCE
- LOWER MECHANICAL IMPEDANCE
- HIGHER VOLTAGE-COMPLIANCE PRODUCT
- IMPROVED TRACKING
- LOW DISTORTION TORQUE DRIVE
- LOWER STYLUS FORCE



Dual Needle The Twilt, proven in thousands of high quality receivers, tracks perfectly at six grams in *all* grooves and speeds. E-V's inline twin tips provide higher compliance (softness of needle touch to record) at three mil tip . . . where it is needed for playing 78 RPM records without weight change. Twin-tilt cartridges are available with a standard frequency response or can be made to your response specifications.

Single Needle For one or three mil grooves. Stylus easily replaceable. Response tailored to your specific needs. TORQUE DRIVE* gives higher voltage and compliance product. Choice of snap-in or rigid mounting.

Universal Needle A compromise stylus that will give satisfactory service on both three mil and one mil records. Minimum tracking force only eight grams. Frequency response made to your requirements.

The *Universal Needle* is the result of months of intensive research and an alert product development program.

• Higher compliance than hitherto obtainable from old style crystal harnessing is an inherent characteristic of TORQUE DRIVE, exclusive to all E-V cartridges. Multiplication of stylus force produces a higher product of voltage and compliance. Accurate frequency response standards are maintained in QUANTITY PRODUCTION because all pads, bearings and high parallel compliances are eliminated. Special E-V SILICONE moisture-proofing gives the crystal many times greater protection against humidity than normal moisture treatment . . . considerably increasing cartridge life.

• *E-V single play tone arms are built to the same exacting standards as E-V cartridges. Arm dimensions permit excellent tracking across any size record. Feedback from base-board to cartridge is reduced because of scientific tone arm design.*

**Patent Pending*

ELECTRO-VOICE, INC., BUCHANAN, MICH. Export: 13 East 40th St., New York 16, U. S. A. Cables: Arlab

NO FINER CHOICE THAN Electro-Voice

CRYSTAL AND MAGNETIC PHONO PICKUPS • CRYSTAL, DYNAMIC, CARBON AND VELOCITY MICROPHONES • STANDS AND ACCESSORIES

TWIN Power Supply

Electronically Regulated for Precise Measurements

Two independent sources of continuously variable D.C. are combined in this one convenient unit. Its double utility makes it a most useful instrument for laboratory and test station work. Three power ranges are instantly selected with a rotary switch:

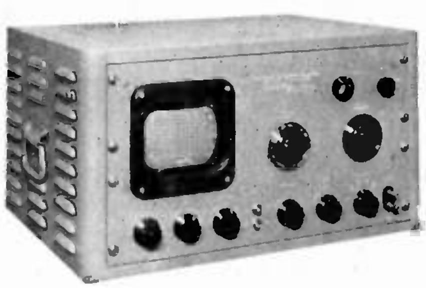
175-350 V. at 0-60 Ma., terminated and controlled independently, may be used to supply 2 separate requirements.
0-175-V. at 0-60 Ma. for single supply.
175-350 V. 0-120 Ma. for single supply.

In addition, a convenient 6.3 V.A.C. filament source is provided. The normally floating system is properly terminated for external grounding when desired. Adequately protected against overloads.

- *Output voltage variation less than 1% with change from 0 to full load.*
- *Output voltage variation less than 1 V. with change from 105 to 125 A.C. Line Voltage.*
- *Output ripple and noise less than .025 V.*

Twin Power Supply Model 210
Complete $130.00
Dimensions: 16" X 8" X 8" Shipping Wt. 35 lbs.
(Other types for your special requirements)

FURST ELECTRONICS
14 S. Jefferson St., Chicago 6, Illinois



Available Now!

Easy, Fast
Ultrasonic Spectrum Analysis
WITH
MODEL SB-7

PANORAMIC ULTRASONIC ANALYZER

An invaluable new direct reading instrument for simplifying ultrasonic investigations, the SB-7 provides continuous high speed panoramic displays of the frequency, amplitude and characteristics of signals between 2KC and 300KC. The SB-7 allows simultaneous observation of many signals within a band up to 200KC wide. Special control features enable selection and highly detailed examination of narrower bands which may contain signals separated by less than 500c.p.s. The instrument is unique in that it provides rapid indications of random changes in energy distribution.

USES

- Ultrasonic Vibration Measurements
- Harmonic Analysis
- Cross Modulation Studies
- Noise Investigations
- Determining Transmission Characteristics of Lines and Filters
- Monitoring Communications Carrier Systems
- Checking Interference, Spurious Modulation, Parasitics, Effects of load changes, shock, humidity, component variations, etc. upon frequency stability
- Telemetering

SPECIFICATIONS

Frequency Range: 2KC—300KC, stabilized linear scale
Scanning Width: Continuously variable from 200KC to zero
Four Input Voltage Ranges: 0.05V. to 50V. Full scale readings from 1 millivolt to 50 volts
Amplitude Scale: Linear and two decale log
Amplitude Accuracy: Within 1db. Residual harmonics suppressed by at least 50db.
Resolution: Continuously variable. 2KC at maximum scanning width, 500c.p.s. for scanning widths below 8KC

***WRITE NOW** For Complete Information, Price and Delivery*

PANORAMIC RADIO PRODUCTS Inc.
10 SOUTH SECOND AVE. MOUNT VERNON, N. Y.

ANNOUNCING A NEW 8" SPEAKER WITH A NEW HARD-HITTING RANGE

JIM LANSING SIGNATURE
8" SPEAKER
100 TO 12,000 C. P. S.

WHERE QUALITY REPRODUCTION IS A "MUST" and space is at a premium—the Jim Lansing 8" Speaker answers the problem! High efficiency and good over-all performance. For improved radio, phonograph and custom television sound reproduction. Designed especially for commercial or industrial use. Ideal for music distribution and paging systems. At all better dealers and distributors.

MODEL D-1002
Two-Way System

For FM monitoring and high quality home sound reproduction. Console type cabinet.

See your Jobber or write to:

JAMES B. LANSING SOUND INC.
2439 Fletcher Drive
Los Angeles, California

HIGH CAPACITY TO SPACE RATIO

With Johnson Pressurized Capacitors

JOHNSON Pressurized Capacitors are so carefully engineered that they provide the desired capacity and voltage rating with minimum pressure and condenser height. Because of their efficient electrical and mechanical design, they also provide the utmost in stable operating conditions.

Available as "standard" are variable, fixed and fixed-variable units — in a wide variety of capacitance and current rating. In addition, JOHNSON can build any pressure condenser to individual specifications.

FEATURES

- Low Loss
- High KVA Rating
- Shielded From External Electrostatic Fields
- Low Internal Distributed inductance
- Complete Dependability

Write For Illustrated JOHNSON Catalog and Prices

JOHNSON
a famous name in Radio!

E. F. JOHNSON CO. WASECA, MINN.

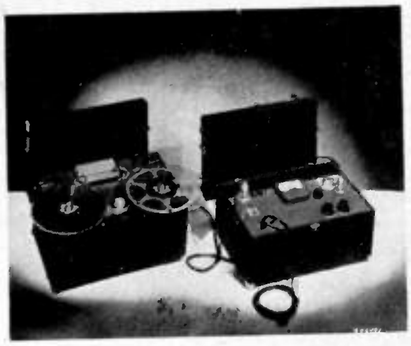
News—New Products

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)

Portable Magnetic Tape Recorders

Two new portable professional models of Ekotape magnetic tape recorders, Model 105, consisting of a single unit contains both record and playback amplifiers in addition to the magnetic tape recorder mechanism, Model 107, consisting of two units, the recording mechanism and the amplifier chassis, are announced by **Webster Electric Co.**, Racine, Wis.



Both models are provided with a single knob control for record, stop, listen, and rewind. The synchronous two-speed motor of Model 107, shown in illustration, provides a tape speed of 15 inches per second for a full half-hour program, or a tape speed of 7½ inches per second for an hour program. Tape speed of 7½ inches per second with Model 105 provides for a full half-hour program. Fast forward and fast rewind speeds permit rapid selection and replay of any part of a recording without removing the tape reels.

DC Amplifier and Millivoltmeter

A new dc amplifier and millivoltmeter, Model DCA-3, with a sensitivity of 1 mv and 165 micromicroamperes for full-scale deflection, has been developed by **Millivac Instruments**, P.O. Box 3027, New Haven, Conn.

The DCA-3 has a direct input impedance of 6 megohms—16 and 60 with a multiplier. The input impedance selector operates between 100 ohms to 4 megohms parallel burden. There is an output shunt selector—1 to 1,000 ohms parallel burden, or protection of sensitive external meters.

Maximum dc output is 5 milliamperes, sufficient for ink and photographic recorders.

(Continued on page 64A)

FM SIGNAL GENERATORS

MODEL 78-FM

86 Mc.—108 Mc.

1 MICROVOLT TO .1 VOLT

DEVIATION: Directly calibrated dial. Two ranges, 0 to 30 kc., 0 to 300 kc. Internal 400 cycle oscillator. Can also be modulated from external source.

DIMENSIONS: 10"x13"x7". Weight 20 lbs.

POWER SUPPLY: 117 volts, 50-60 cycles. 36 watts.

• SPECIAL GENERATORS

One-band Model 78-FM generators, with a tuning ratio of approximately 1.2 to 1, are available for use within the limits of 30 to 165 megacycles.

MODEL M-275

I. F. CONVERTER

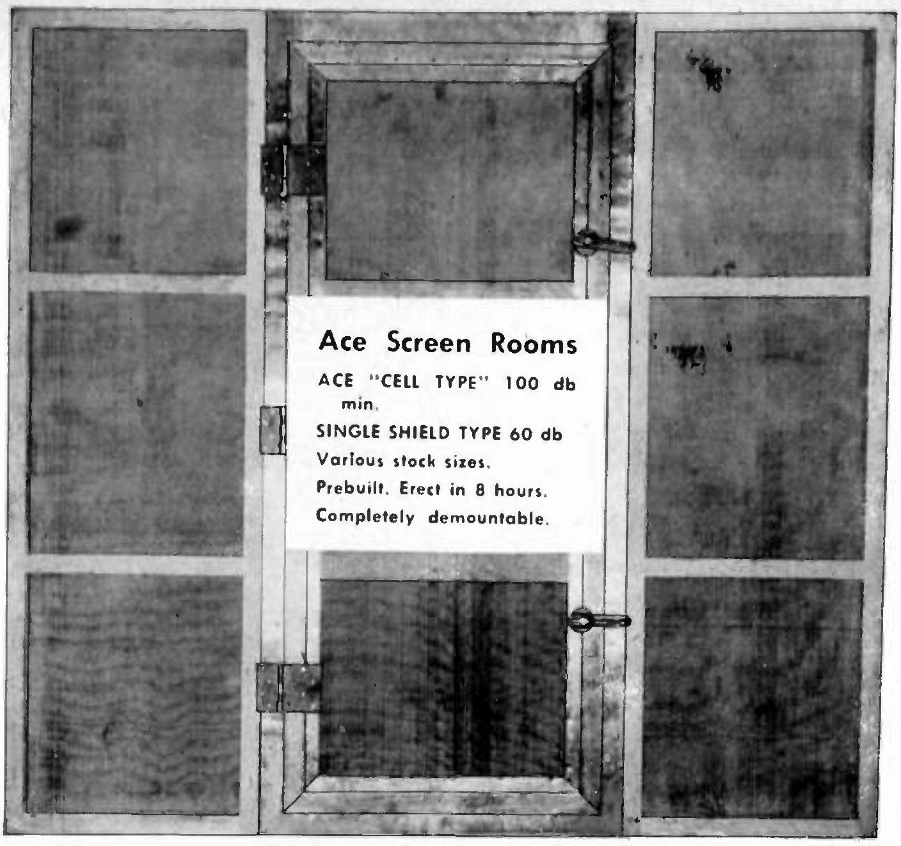
For Use With Model 78-FM.

CARRIER FREQUENCIES: 4.5 Mc.; 10.7 Mc.; 21.7 Mc. (Provision for one extra frequency).

OUTPUT: When used with Model 78-FM the output voltage is variable from 10 microvolts to 1 volt.

POWER SUPPLY: 117 volts, 50-60 cycles, 45 watts.

MEASUREMENTS CORPORATION
BOONTON NEW JERSEY



Ace Screen Rooms

ACE Engineering & Machine Co.

3644 N. Lawrence St. Philadelphia 40, Pa.

NOW AVAILABLE!

A Z-ANGLE METER FOR RADIO FREQUENCY MEASUREMENTS

Featuring the same ease and efficiency of operation as the widely used audio Z-Angle Meter, the new R-F Z-Angle Meter simplifies the problem of measurements at radio frequencies of broadcast antennas, transmission lines, complex networks and impedance elements.

Write today for bulletins on other T. I. C. products: Z-Angle Meter . . . R-F Oscillator . . . Electronic Phase Meter . . . Precision Linear and Non-Linear Variable Resistors . . . Translatory Variable Resistors . . . Slide Wire Resistance Boxes.

Frequency Range: 100 kc to 2 mc
Direct Reading:
Impedance Range: 10 to 5,000 ohms up to 200 kc
10 to 1,000 ohms up to 1 mc
Phase Angle: +90° (X_L) thru 0° (R) to −90° (X_C)
Q Range: 0.2 to 10
Portable—for field or laboratory use
Self-contained balance indicator
Operates at high level—overrides extraneous antenna pickup

TECHNOLOGY INSTRUMENT CORP.
1058 MAIN STREET, WALTHAM 54, MASS.

ENGINEERING REPRESENTATIVES

Chicago, Ill.—STate 2-7444 Cambridge, Mass.—ELiot 4-1751 Canaan, Conn.—Canaan 649
Rochester, N.Y.—Charlotte 3193-J Dallas, Tex.—Logan 6-5097

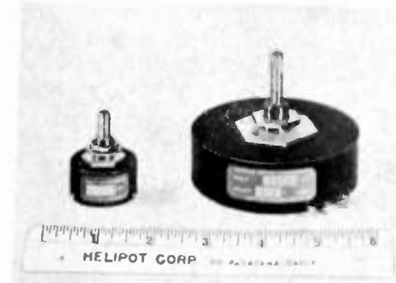
News—New Products

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

(Continued from page 63A)

Two New Precision Linear Potentiometers

Two new potentiometers, Models F and G, with guaranteed linearity accuracies of ±0.5 per cent, and 1.0 per cent respectively are being manufactured by **Helipot Corp.**, South Pasadena 6, Calif.



With the Model F, by substitution, it is possible to obtain accuracies of 0.1 per cent, and in the higher values of resistance 0.05 per cent. All ratings of the Model G are held within linearity tolerances of ±1.0 per cent, and in the higher resistances to accuracies of ±0.25 per cent.

Both models have been tested in excess of a million revolutions at speeds as high as 100 rpm. Slide wire and slip ring contacts are of precious metal alloys. Resistance values of both models are held within ±5 per cent, but can be maintained at tolerances as low as ±1 per cent, if required.

Dynaural Converter

Based upon the H. H. Scott Dynamic Noise Suppressor, a new line of equipment for Dynaural reproduction of music, particularly from phonograph records has been developed by **Hermon Hosmer Scott, Inc.**, 385 Putnam Ave., Cambridge 39, Mass.

In a Dynaural system the bandwidth is automatically and continuously adjusted to conform with the requirements of the music, thus combining maximum fidelity with minimum noise level.

The first unit available is the Type 111-A Dynaural Converter. Designed for use with standard amplifiers, phonographs and combinations, it has a 14 kc range.

Simplicity of installation and operation have been stressed in the design of the Dynaural Converter, which requires only two connections. A single remote Dynaural control is provided with a special dial plate and bracket for mounting on the panel of the phonograph, in the phonograph compartment, or any other convenient location.

(Continued on page 65A)

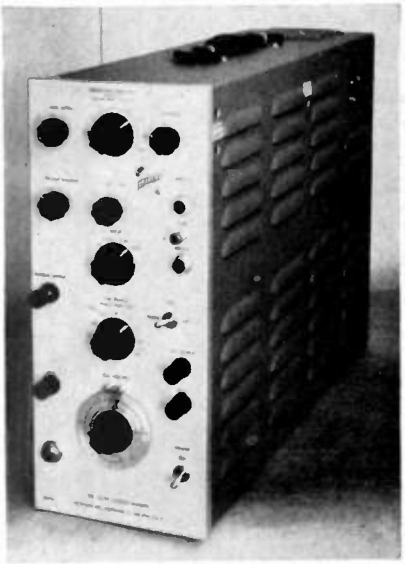
News—New Products

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

(Continued from page 64A)

New Model Direct Coupled Amplifier

The Type 112 dc amplifier, with a bandwidth of 1 Mc when used at a maximum gain of 5,000 volts, is obtainable on order from **Taktronix, Inc.**, 712 S.E. Hawthorne Blvd., Portland 14. Ore.



For gain requirements of 166 volts or less, the bandwidth extends to 2 Mc. An output of approximately 150 volts (peak to peak) is available to a high impedance load such as cathode-ray tube deflection plates. Control of gain is continuously variable from 0.5 to 5,000 volts by use of a step and variable attenuator.

The amplifier has an input impedance of 1 megohm—45 $\mu\mu f$ each side to ground, or 10 megohms—14 $\mu\mu f$ each side to ground when using the supplied probes.

A 1 kc square wave calibrating voltage from 0 to 50 volts is available by a nine position range switch in conjunction with a calibrated potentiometer providing an accuracy of ±5 percent.

New Electrosensitive Recording Paper

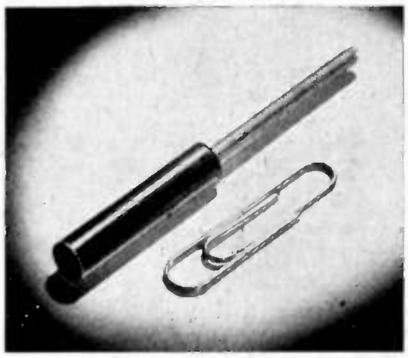
Designed for use with Helix-type recorders, and applicable to record radar, sonar, and infrared signals, computer totals, and telemetered dispatch, a new electrosensitive paper is available from **Alfax Paper and Engineering Co.,** 46 Riverside Ave., Brockton, Mass.

The manufacturer has stated that when using a Helix-type and Alfax paper, inertia-free instantaneous recording at speeds to 300 inches per second and up are available. With this inkless, permanent paper, recording of four simultaneous overlapping signals is accomplished with no difficulty.

New Cores For TV Image Width Control

Molded of a powdered material assuring high permeability, these iron cores have been designed for television horizontal image deflection circuits by **Electronic Components Div., Stackpole Carbon Co.,** St. Marys, Pa.

In screen areas where there is a sudden voltage drop, the manufacturer claims that these cores give ratios of from 1 to 8 or more, compared to 1 to 5 for previous high permeability cores.



Known as Stackpole Ceramag II cores, the new units are of standard screw-type construction and are available in a complete range of frequencies for modern television applications.

(Continued on page 68A)

A.R.C.'s VHF Communication and Navigation Equipment is a

REVELATION

Get static-free communication and the added reliability of omni range navigation with A. R. C.'s Type 17 2-way VHF Communication and Type 15B Omni Range Navigation Equipment. With the 15B tuned to VHF omni stations, you fly directly in less time. You can receive weather broadcasts simultaneously with navigation signals —static free! It simplifies navigation and gives long, trouble-free life. The Type 17 adds an independent communication system for use while the 15B is providing navigational information. Installations for both single and multi-engined planes are made only by authorized agencies.

CLEAR COMMUNICATION ON VHF

All A.R.C. airborne equipment is Type Certificated by CAA. It is designed for reliability and performance—not to meet a price. Write for further details or name of your nearest A.R.C. representative.

Aircraft Radio Corporation
BOONTON, NEW JERSEY

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tell you what they make, in your IRE Yearbook. Products are organized in 68 easy-to-use, fundamental engineering groups. It saves you time and money to "look it up" in your IRE Yearbook!

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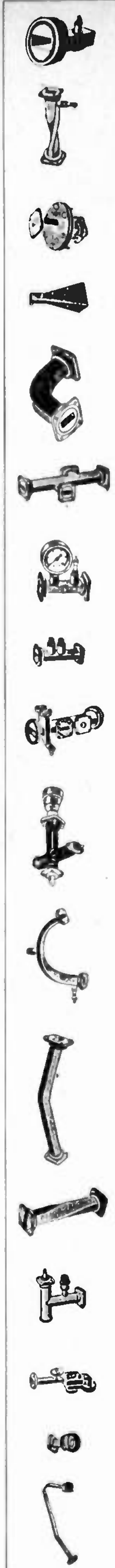
R. F. COMPONENTS—MICROWAVE—TEST EQUIPMENT



10 CENTIMETER

WAVEGUIDE TO ⅞" RIGID COAX "DOORKNOB" ADAPTER. CHOKE FLANGE. SILVER PLATED BROAD BAND $37.50
WAVEGUIDE DIRECTIONAL COUPLER, 27 db. Navy type CABV-47AAN, with 4 in. slotted section $32.50
SQ. FLANGE to rd choke adapter, 18 in. long OA 1½ in. x 3 in. guide, type "N" output and sampling probe $27.50
Crystal Mixer with tunable output TR pick up loop, Type "N" connectors. Type 62ABH $14.50
Slotted line probe. Probe depth adjustable, Sperry connector, type CPR-14AAO .. $9.50
Coaxial slotted section, ⅝" rigid coax with carriage and probe $25.00
Right Angle Bend 6" radius E or H plain $15.00
Right Angle Bend 3" radius E or H plain—Circular flanges $15.00
AN/APR5A 10 cm antenna equipment consisting of two 10 CM waveguide sections, each polarized. 45 degrees .... $75.00 per set
APN-7 TR McNally Cavity for 707B, with tuning slugs .... $5.50 ea.
"S" Band Crystal Mount, gold plated, with 2 type "N" connectors $12.50
PICKUP LOOP, Type "N" Output $2.75
TR BOX Pick-up Loop $1.25
POWER SPLITTER: 726 Klystron input dual "N" output $5.00
MAGNETRON TO WAVEGUIDE coupler with 721-A duplexer cavity, gold plated $27.50
10 CM WAVEGUIDE SWITCHING UNIT, switches 1 input to any of 3 outputs. Standard 1½" x 3" guide with square flanges. Complete with 115 vac or dc arranged switching motor. Mfg. Raytheon. CRP 24AAS. New and complete $150.00
"S" BAND Mixer Assembly, with crystal mount, pick-up loop, tunable output $3.00
721-A TR CAVITY WITH TUBE. Complete with tuning plungers $12.50
10 CM. McNALLY CAVITY Type SG $3.50
Type SF $4.50
WAVEGUIDE SECTION, MC 445A, rt. angle bend, 5½" ft. OA. 8" slotted section .. $21.00
10 CM OSC. PICKUP LOOP, with male Homedell output $2.00
10 CM DIPOLE WITH REFLECTOR in lucite ball, with type "N" or Sperry fitting .. $4.50
10 CM FEEDBACK DIPOLE ANTENNA, in lucite ball, for use with parabola ⅞" Rigid Coax Input $8.00
PHASE SHIFTER. 10 CM WAVEGUIDE, WE TYPE ES-683816. E PLANE TO H PLANE, MATCHING SLUGS $95.00
721A TR cavities. Heavy silver plated $2.00 ea.
10 cm. horn and rotating joint assembly, gold plated .... $65.00 ea.

⅞" RIGID COAX—⅜" I.C.

⅞" rigid coaxial tuning stubs with vernier stub adjustment. Gold Plated $17.50
⅞ RIGID COAX ROTARY JOINT. Pressurized. Sperry #810613. Gold Plated $27.50
Dipole assembly. Part of SCR-584 .... $25.00 ea.
Rotary joint. Part of SCR-584 ...... $35.00 ea.
RIGHT ANGLE BEND, with flexible coax output pickup loop $8.00
SHORT RIGHT ANGLE BEND, with pressurizing nipple $3.00
RIGID COAX to flex coax connector $3.50
STUB-SUPPORTED RIGID COAX, gold plated 5' lengths. Per length $5.00
RT. ANGLES for above $2.50
RT. ANGLE BEND 15" L. OA $3.50
FLEXIBLE SECTION, 15" L. Male to female $4.25
FLEX COAX SECT. Approx. 30 ft. $16.50

⅞" RIGID ¼" IC

CG 54/U—4 foot flexible section ¼" IC pressurized $15.00
⅞ RIGID COAX. Bead Supported $1.20
SHORT RIGHT ANGLE BEND $2.50
Rotating joint, with deck mounting ... $15.00
RIGID COAX slotted section CU-60/AP $5.00

WAVEGUIDE

½" x ¼" ID	$1.00 per foot
1" x ½" OD	1.50 per foot
⅝" x 1¼" OD	1.65 per foot
⅝" x 1¼" OD Aluminum	.75 per foot
1½" x 3" OD	3.00 per foot
2½" x 3" OD	3.50 per foot
1" x ½" OD Flexible	4.00 per foot
⅞" rigid coax ¼" IC	1.20 per foot

(Available in 10FT to 15ft. lengths or smaller.)

UG 65/U 10CM flanges	$6.75 each
UG 53/U Cover	4.00 each
UG 54/U Choke	4.50 each

SCR 584 SPARE PARTS AVAILABLE

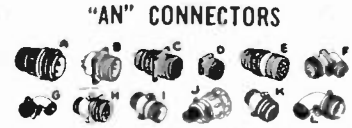
3 CENTIMETER

(STD. 1" x ½" GUIDE UNLESS OTHERWISE SPECIFIED)

723 A/B Klystron mixer section with crystal mount, choke flange and Iris flange output $22.50
TR-ATR Section for above with 724 ATR Cavity $8.50
Cutler 105/APS31 Directional Coupler 25 DB $25.00
90 degree twist, 6 inches long $8.00
723 AB Mixer—Beacon Dual Oscillator Mount with Crystal holder $12.00
2 Way Wave Guide directional coupler, type N fitting 1⅛" x ⅝" guide 26DB $18.50
CG 98B/APG 13, 12" flexible section 1¼" x ⅝" OD $10.00
TR-ATR Section, APS 15, for 1B24, with 724 ATR Cavity with 1B24 and 724 tubes. Complete $21.00
Crystal mount in waveguide $17.50
SO-3 Echo box, Xmsn. type cavity w/ bellows $28.50
3 cm. 180° bend with pressurizing nipple .... $6.00 ea.
3 cm. 90 bend, 14" long 90° twist with pressurizing nipple .... $6.00 ea.
3 cm. "S" curve 18" long .... $5.50 ea.
3 cm, "S" curve 6" long .... $3.50 ea.
3 cm. right angle bends. "E" plane 18" long cover to cover .... $6.50 ea.
3 cm. Cutler feed dipole. 11" from parabola mount to feed back .... $8.50 ea.
TWIST 45 deg.. 6" long $8.00
APS-31 mixer section for mounting two 2K25's. Beacon reference cavity 1B24 TR tube. New and complete with attenuating slugs $42.50 ea.
DUPLEXER SECTION for 1B24 .... $10.00
CIRCULAR CHOKE FLANGES, solid brass .55
SQ. FLANGES, FLAT BRASS .... ea. .55
FLEX, WAVEGUIDE .... $4.00/Ft.
TRANSITION 1 x ½ to 1¼ x ⅝, 14 in. . $8.00
"X" BAND PREAMPLIFIER, consisting of 2-723 A/B local oscillator-beacon feeding waveguide and TR/ATR Duplexer sect. incl. 30MC Pre Amp with tubes $67.50
Random Lengths wavegd, 6" to 18" Lg. $1.10/Ft.
WAVEGUIDE RUN. 1¼" x ½" guide, consisting of 4 ft. section with Rt. angle bend on one end 2" 45 deg. bend other end .. $8.00
WAVEGUIDE RUN, 1¼" x ½" guide, consisting of 4 ft. long $10.00
12" SECTION 45 deg twist, 90 deg bend $6.00
11" STRAIGHT WAVEGUIDE section choke to cover. Special heavy Construction, silver plated $4.50
15 DEG BEND 10" choke to cover $4.50
5 FT. SECTIONS choke to cover, Silver Plated $10.00
18" FLEXIBLE SECTION $17.50
"X" BAND WAVEGUIDE 1¼" x ⅝" OD 1/16" wall Alluminum .... Per Foot $ .75
WAVEGUIDE 1" x ½" I.D. .... Per Foot $1.50
TR CAVITY For 724 A TR Tube $3.50
3" FLEX SECT. sq. flange to Circ. Flange Adapt $7.50
724 TR TUBE (41 TR 1) $2.50
SWR MEAS. SECTION, with 2 type "N" output probes MTD full wave apart. Bell sizes guide. Silver plated $10.00
WAVEGUIDE SECTION, 12" long choke to cover 45 deg. twist & 2½" radius, 90 deg. bend $4.50
SLUG TUNER/ATTENUATOR, W. E. guide, gold plated $6.50
TWIST 90 deg. 5" choke to Cover w/press nipple $6.50
WAVEGUIDE SECTIONS 2½ ft. long silver plated with choke flange $5.75
ROTARY JOINT choke to choke $17.50
ROTARY JOINT choke to choke with deck mounting $17.50

UG TYPE CONNECTORS
FULL LINE OF UG CONNECTORS
IN STOCK. IMMEDIATE DELIVERY.

"AN" CONNECTORS



LARGE VARIETY AVAILABLE
AT GREAT SAVINGS
Send your specs and let us quote

1.25 CENTIMETER

"K" BAND DIRECTIONAL COUPLER CU104/APS-34 20 DB .... $49.50 ea.
"K" BAND FEEDBACK TO PARABOLA HORN, with pressurized window $30.00
MITRED ELBOW cover to cover $4.00
TR/ATR SECTION choke to cover $4.00
FLEXIBLE SECTION 1" choke to choke .. $5.00
ADAPTER, rd. cover to sq. cover $5.00
MITRED ELBOW and S sections choke to cover $4.50
WAVE GUIDE ½ x ¼ per ft. $1.00
K BAND CIRCULAR FLANGES .... 50¢
3J31 "K" BAND MAGNETRON $55.00

TEST EQUIPMENT

MODEL TS-268/U
Test set designed to provide a means of rapid checking of crystal diodes 1N21, 1N21A, 1N21B, 1N23, 1N23A, 1N23B. Operates on 1½ volt dry cell battery. 3x6x7.
New $35.00

CRYSTAL DIODES

No.	Each	2 for	10 for
1N21	$1.00	$1.79	$ 8.30
1N22	1.50	2.79	14.00
1N23	1.50	2.79	14.00
1N26	3.00	5.90	27.50

10CM ECHO BOX CABV 14ABA-1 of OBU-3, 2890 MC to 3170 MCS direct reading micrometer head. Ring prediction scale plus 9% to minus 9%. Type "N" input. Resonance indicator meter. New and Comp. w/access. Box and 10 CM Directional Coupler $350.00
3 CM RECEIVER. SO-3. Complete with W.G. Mixer Assy (723-A/B). Reg. Fil. Power Supply 6 stages IF (6AC7) $99.50
10 cm. horn assembly consisting of two 5" dishes with dipoles feeding single type "N" output. Includes UG28/U type "N" "T" junction and type "N" pickup probe. Mfg. cable. New $15.50
10 cm. cavity type wavemeters 6" deep, 6½" in diameter. Coax. output. Silver plated .... $64.50 ea.
10 cm. echo box. Part of SFI Radar W/115 volt DC tuning motor Sub Sig 1118AO $47.50
THERMISTOR BRIDGE: Power meter 1-203-A. 10 cm. mfg. W.E. Complete with meter, interpolation chart, portable carrying case $72.50
W.E. I 138. Signal generator. 2700 to 2900 Mc. range. Lighthouse tube oscillator with attenuator & output meter. 115 VAC input reg. Pwr. supply. With circuit diagram $150.00
3 cm. wavemeter. Ordnance type micrometer head. New: Absorbtion type $85.00
9000-9500 MCS Transmition type $92.50
SL. wavemeter. Type CW60ABM $125.00
TS 89/AP Voltage Divider. Ranges 100: ½ for 2000 to 20000v. 10:1 for 200 to 2000v. Input Z 2000 ohms. Output Z 4 meg ohms flat response 150 cy to 5 meg cy $42.50
10 cm Wavemeter. WE type B 435490 Transmission type. Type N Fittings. Vecder Root Micrometer dial. Gold Plated W/Calib. Chart. P/o Freq. Meter X66404A. New $99.50
AS14A/AP—10cm Pick up Dipole with "N" Cables $4.50
TS 235 UP Dummy Load $87.50

COMPLETE RADARS AVAILABLE

SO1 Used	$1200.00
SO13 Used	$1200.00
SE New	$1200.00
SFI New	$2800.00
SCR 533 Trailer unused	$950.00
CXBR Beacon Used	$1500.00
CPN3 Beacon Used	$1500.00
APS4	
APS15 Major Components	$500.00
Airborne Radar Altimeter	$175.00

ALL MERCHANDISE GUARANTEED. MAIL ORDERS PROMPTLY FILLED. ALL PRICES F.O.B. NEW YORK CITY. SEND MONEY ORDER OR CHECK ONLY. SHIPPING CHARGES SENT C.O.D. RATED CONCERNS SEND P. O. MERCHANDISE SUBJECT TO PRIOR SALE

COMMUNICATIONS EQUIPMENT CO.
131 "I 12" Liberty St., New York, N.Y.
Cable "Comsupo" Ph. Digby 9-4124

MAGNETRONS - RADAR - PULSE EQUIPMENT

PULSE EQUIPMENT

MIT. MOD. 3 HARD TUBE PULSER: Output Pulse Power: 114 KW (12 KV at 12 amp.). Duty Ratio: .001 max. Pulse duration: 5. 1.0 2.0 microsec. input voltage: 115 v. 400 to 2400 cps. Uses 1-715-B, 1-829-B, 3-'72's, 1-'73. New $110.00

APQ-13 PULSE MODULATOR. Pulse Width .5 to 1.1 Micro. Sec. Rep. rate 624 to 1348 Pps. Pk. pwr. out 35 KW. Energy 0.018 Joules $49.00

TPS-3 PULSE MODULATOR. Pk. power 50 amps. 24 KV (1200 KW pk): pulse rate 200 PPS, 1.5 microsec., pulse line impedance 50 ohms. Circuit—series charging version of DC Resonance type. Uses two 705-A's as rectifiers. 115 v. 400 cycle input. New with all tubes $49.50

APS-10 MODULATOR DECK Complete, less tubes $75.00

APS-10 Low voltage power supply, less tubes $18.50

PULSE TRANSFORMERS

G.E.K.-2745 $39.50

G.E.K.-2744-A. 11.5 KV High Voltage, 3.2 KV Low Voltage @ 200 KW oper. (270 KW max.) 1 microsec. or ¼ microsec. @ 600 PPS $39.50

W.E. #D166173 Hi-Volt input transformer, W.E. Impedance ratio 50 ohms to 900 ohms. Freq range: 10 kc to 2 mc. 2 sections parallel connected, potted in oil $36.00

W.E. KS 9800 Input transformer. Winding ratio between terminals 3-5 and 1-2 is 1.1:1, and between terminals 6-7 and 1-2 is 2:1. Frequency range: 380-520 c.p.s. Permalloy core $6.00

G.E. #K2731 Repetition Rate: 635 PPS, Pri. Imp: 50 Ohms, Sec. Imp: 450 Ohms. Pulse Width: 1 Microsec. Pri. Input: 9.5 KV PK. Sec. Output: 28 KV PK. Peak Output: 800 KW. Riflar 2.75 Amp. $64.50

W.E. #D169271 Hi Volt input pulse Transformer $27.50

G.E. K2450A. Will receive 13KV, 4 micro-second pulse on pri., secondary delivers 14KV Peak power out 100KW G.E. $4.50

G.E. #K2748A. Pulse Input, line to magnetron $36.00

#9280 Utah Pulse or Blocking Oscillator XFMR Freq. limits 790-810 cy-3 windings turns ratio 1:1:1 Dimensions 1 13/16 x 1⅛" 19/32 $1.50

Pulse 131-AWP L-421435 $6.00

Pulse 134-BW-2F L-440895 $2.25

PULSE NETWORKS

15A—1-400-50: 15 KV, "A" CKT, 1 microsec., 400 PPS, 50 ohms imp. $42.50

G.E. #6E3-5-2000-50P2T, 6KV. "E" circuit, 3 sections, .5 microsecond, 2000 PPS, 50 ohms impedance $6.50

G.E. #3E (3-84-810; 8-2.24-405) 50P4T; 3KV, "E" CKT Dual Unit: Unit 1, 3 Sections, .84 Microsec. 810 PPS, 50 ohms imp.: Unit 2, 8 Sections, 2.24 microsec. 405 PPS, 50 ohms imp. $6.50

7.5E3-1-200-67P. 7.5 KV, "E" Circuit, 1 microsec. 200 PPS, 67 ohms impedance, 3 sections $7.50

7.5E4-16-60.67P. 7.5 KV, "E" circuit, 4 sections, 16 microsec. 60 PPS, 67 ohms impedance $15.00

7.5E3-3-200-6PT. 7.5 KV, "E" Circuit, 3 microsec. 200' PPS, 67 ohms imp., 3 sections $12.50

DELAY LINES

163169 Delay Line Small quantity available $50.00

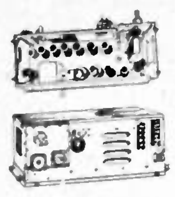
D-168184: .5 microsec. up to 2000 PPS, 1800 ohm term. $4.00

D-170499: .25/.50/.75. microsec. 8 KV. 50 ohms imp. $16.50

D-165997: 1¼ microsec. $7.50

THERMISTORS		VARISTORS	
D-167332 (tube)	$.95	D-170225	$1.25
D-170396 (bead)	$.95	D-167176	$.95
D-167613 (button)	$.95	D-168687	$.95
D-164600 for MTG in "X" band Guide	$2.50	D-171812	$.95
D-167018 (tube)	$.95	D-171528	$.95
		D-168549	$.95
		D-162482	$3.00
		D-163298	$1.25
		D-99428	$2.00
		D-161871A	$2.85
		D-171121	$.95
		3A(12-43)	$1.50
		D-167020	$3.00

WRITE FOR C.E.C. MICROWAVE CATALOG NOW AVAILABLE



3 CM RECEIVER

SO-3. Complete With W.G. Mixer Assy (723 A/8 Reg. Fil. Power Supply, 6 Stages IF 6AC7) $99.50

MAGNETRONS

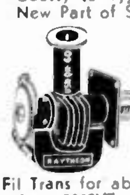
Tube	Frq. Range	Pk. Pwr. Out	Price
2J31	2820-2860 mc.	265 KW	$25.00
2J21-A	9345-9405 mc.	50 KW.	$25.00
2J22	3267-3333 mc.	265 KW.	$25.00
2J26	2992-3019 mc.	275 KW.	$25.00
2J27	2965-2992 mc.	275 KW.	$25.00
2J32	2780-2820 mc.	285 KW.	$25.00
2J37			$45.00
2J38 Pkg.	3249-3263 mc.	5 KW.	$35.00
2J39 Pkg.	3267-3333 mc.	87 KW.	$35.00
2J40	9305-9325 mc.	10 KW.	$65.00
2J49	9000-9160 mc.	58 KW.	$85.00
2J34			$55.00
2J61	3000-3100 mc.	35 KW.	$65.00
2J62	2914-3010 mc.	35 KW.	$65.00
3J31	24,000 mc.	50 KW	$55.00
5J30	-		$39.50
714AY			$25.00
718DY	2720-2890 mc.	250 K.W.	$25.00
720BY	2800 mc.	1000 KW.	$50.00
720CY	2860 mc.	1000 K.W.	$50.00
725-A	9345-9405 mc.	50 KW.	$25.00
730-A	9345-9405 mc.	50 KW.	$25.00
728	AY, BY, CY, DY, EY, FY, GY		$50.00
700	A, B, C, D		$50.00
706	AY, BY, DY, EY, FY, GY		$50.00
Klystrons.	723A/B $12.50; 707B W/Cavity		$20.00
	417A $25.00	2K41	$65.00

MAGNETRON MAGNETS

Gauss	Pole Diam.	Spacing	Price
4850	¾ in.	⅝ in	$12.50
5200	21/32 in.	¾ in.	$17.50
1300	1⅝ in.	1 5/16 in.	$12.50
1860	1⅝ in.	1½ in.	$14.50

Electromagnets for magnetrons $24.50 ea

GE Magnets type M7765115, GI Distance Between pole faces variable. 2 1/16" (1900 Gauss) to 1½" (2200 Gauss) Pole Dia. 1⅝" New Part of SCR 584 $34.50



"CW" MAGNETRONS

QK 62 3150-3375 mc.
QK 59 2675-2900 mc.
QK 61 2975-3200 mc.
QK 60 2800-3025 mc.
New, Guaranteed Each $65.00
QK 915 Raytheon $150.00

Fil Trans for above 115v 60 cy Pri Four 6.3v/4A Sec: 5000VT $27.50

Magnetronkit. of four QK's 2675-3375 mc W/trans. Special $250.00

SUPER SONICS

QCU Magneto striction head RCA type CR 278225. New $75.00

Stainless Steel streamlining housings for above $18.50

QBG Driver Amplifier. New $200.00

QCU Magneto striction head coil plate assembly, new $14.50

QCQ—2/QCB Magneto striction head coil plate assembly $14.50

APS 4 AIRBORNE RADAR

Complete sets and various components available.

AS-17/APS—10cm. Airborne Antenna complete with all plumbing, dish, etc. $89.50

SCR 584 SPARE PARTS AVAILABLE

AN/APG2 Servo Amplifier less tubes new $35.00

R. F. EQUIPMENT

LHTR. LIGHTHOUSE ASSEMBLY. Part of RT-39/APG-5 & APG-15. Receiver and Transmitter Lighthouse Cavities with assoc. Tr. Cavity and Type N CPLG. To Rcvr. Uses 2C40, 2C43, 1B27 Tuneable APX 2400-2700 MCS. Silver plated $49.50

APS-2 10CM RF HEAD COMPLETE WITH HARD TUBE (715B) Pulser, 714 Magnetron 417A Mixer all ⅞" rigid coax. incl. rcvr. front end $210.00

Beacon lighthouse cavity 10 cm with miniature 28 volt DC FM motor. Mfg. Bernard Rice $47.50 ea

T-128/APN-19 10 cm. radar Beacon transmitter package, used, less tubes $59.50 ea.

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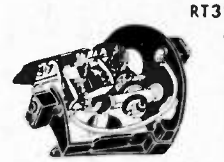
Pre-amplifier cavities type "M" 7410590GL to use 446A lighthouse tube. Completely tunable. Heavy silver plated construction $37.50 ea

RT32/APS 6A RF HEAD



Compl. with 725A Magnetron magnet pulse xfmr. TRA-ATR. 723 A/B local osc. and beacon mount, pre amplifier. Used but exc. cond. $97.50

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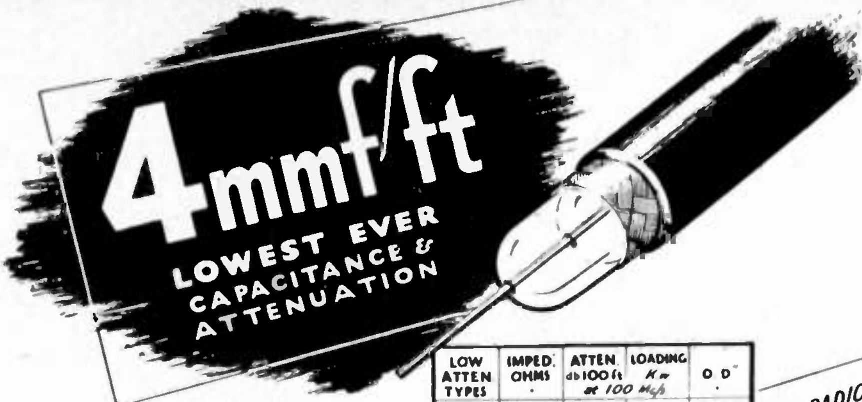

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A.1	74	1.7	0.11	0.36
A.2	74	1.3	0.24	0.44
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C.11	6.3	173	3.2	0.36
C.2	6.3	171	2.15	0.44
C.22	5.5	184	2.8	0.44
C.3	5.4	197	1.9	0.64
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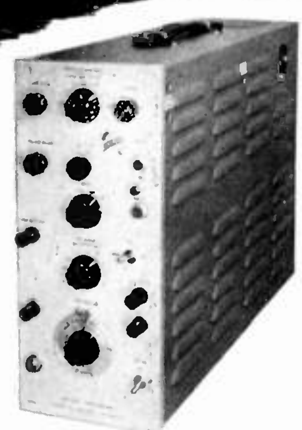

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Detailed Specifications on request.

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News—New Products

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

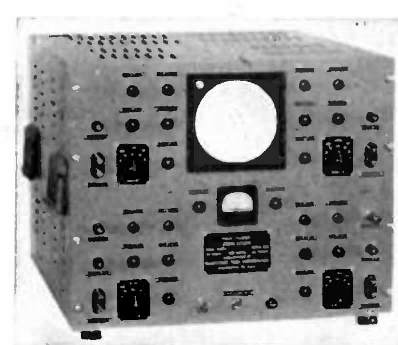
(Continued from page 65A)

HF Noise Generating Diode

A new miniature noise generating diode, suitable for measurements at frequencies up to 500 Mc, has been announced by the Radio Div., **Sylvania Electric Products Inc.**, 500 Fifth Ave., New York 18, N. Y.

The new T 5½ tube Type 5722 is designed for standard laboratory noise measurement. It is operated with 150 volts on plate and at filament voltages ranging between 2 and 5.5 volts, depending on desired plate current or noise output. In intermittent service, maximum plate dissipation is 5 watts.

Record Oscilloscope Displays Four Independent Variables



A single 5-inch cathode-ray tube oscilloscope capable of indicating four separate phenomena simultaneously is currently available from **Electronic Tube Corp.**, 1200 E. Mermaid Lane, Philadelphia 18, Pa.

(Continued on page 69A)

the first line of STANDARD electronic voltage regulators

Sorensen electronic voltage regulators offer accuracy under simultaneous line and load changes.

IMPORTANT SORENSEN FEATURES:

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The Sorensen Catalog contains complete specifications on standard voltage regulators and nobatrons. It will be sent to you upon request.

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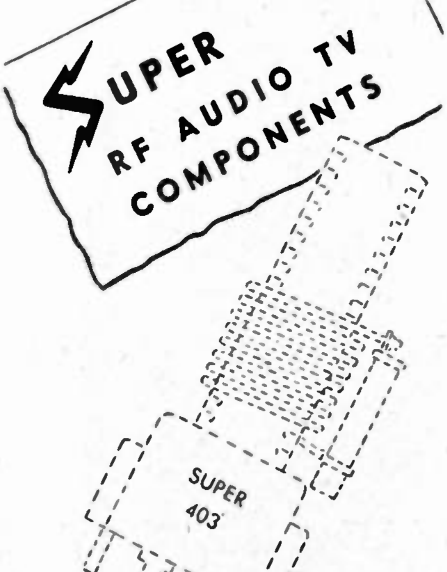

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(Continued from page 68A)

Separate focus, intensity, and positioning controls are provided in each of the four channels, and positioning in both horizontal and vertical directions is possible.

The four signal amplifiers have a frequency range from 0 to 200 kc ± 3 db. and a gain of about 120 times.

Power supplies for the tube and the amplifiers are contained in the cabinet, and require a primary supply of 105–120 volts, 50 to 60 cps.

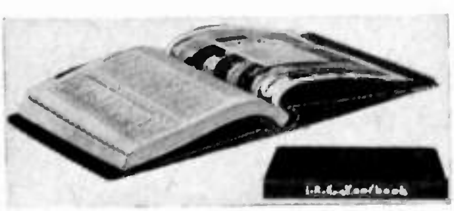
Designated as Model H-43, this oscilloscope is normally supplied with a Fairchild Type, F-246A, Oscillo-Record 35-mm camera, but may be used in conjunction with larger drum type cameras if preferred.

Microwave Relay System for Industrial Purposes

New commercial microwave relay equipment for high-frequency point-to-point communication systems is now available to industrial users from the designer, RCA Engineering Products Dept., **Radio Corp. of America, Camden, N. J.**

Stations are approximately 35 miles apart, and house two receivers and transmitters for simultaneous two-way operation. The Type CWTR-5A radio relay equipment provides a modulation channel extending from 300 to 30,000 cps, and is designed for unattended operation in the 940- to 960-Mc frequency band. However, by use of the channeling equipment, each radio circuit is capable of carrying four voice conversations simultaneously. Channeling equipment is also available to break each of these voice bands into as many as 16 signaling circuits for telemetering, signalling, or supervisory control functions.

Yearbook Binder



A permanent, springback binder that snaps onto the IRE YEARBOOK transforming the book into a handsome, bound volume, and that can be transferred to each new annual issue, lasting for years, is offered for $2.75 by **R. V. Metcalf, 52 Chauncy St., Boston 11, Mass.**

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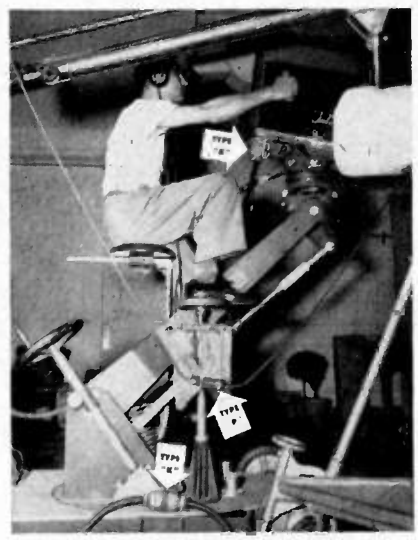
. . . get on the gratis subscription list of *"The Cannon Electric Cannonade"* bi-monthly, eight-page, technical house organ and preview of Cannon Plugs, news and information about all Cannon Electric products.



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. . . be assured of "good connections." That's why television stations, for instance, use Cannon Electric Type K, P, and other series for cameras, microphones and transmission equipment that *must not fail.* Shown above is a camera at KTLA—Hollywood.

Cannon Plugs are available through a network of radio parts dealers all over the U. S. A. Buy them from Rochester Radio Supply, *Rochester, N. Y.;* Warren Radio, *Sioux Falls, N. D.;* Electra Dist., *Nashville, Tenn.;* Radio Specialties, *Detroit;* The Hargis Co., *Austin, Tex.;* Radio & Electronic Parts, *Cleveland;* and more than 400 other radio parts distributors. Write for new C-48 Condensed General Catalog.

Cannon Electric Development Company, Division of Cannon Manufacturing Corporation, 3209 Humboldt Street, Los Angeles 31, California. Canadian factory: Toronto. World Export: Frazar & Hansen, San Francisco, New York, Los Angeles.

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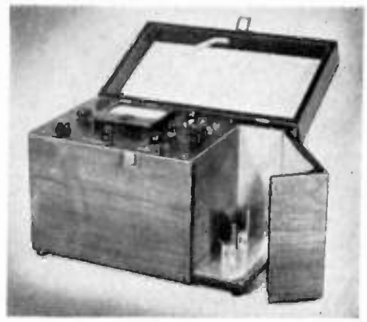

Photo courtesy of National Technical Laboratories, So. Pasadena, Calif.

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It gives essential data about S.S.White Resistors including construction, characteristics, dimensions, etc. Copy with price list on request.

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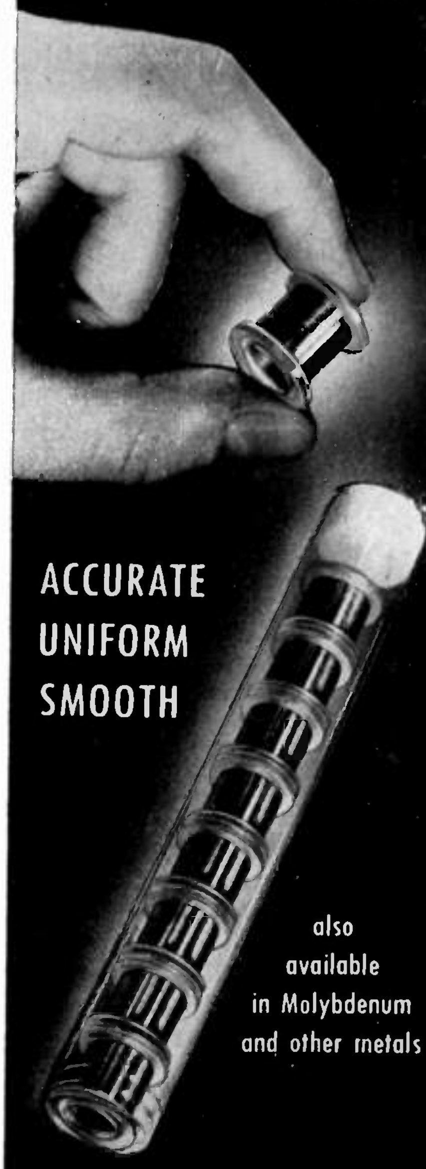

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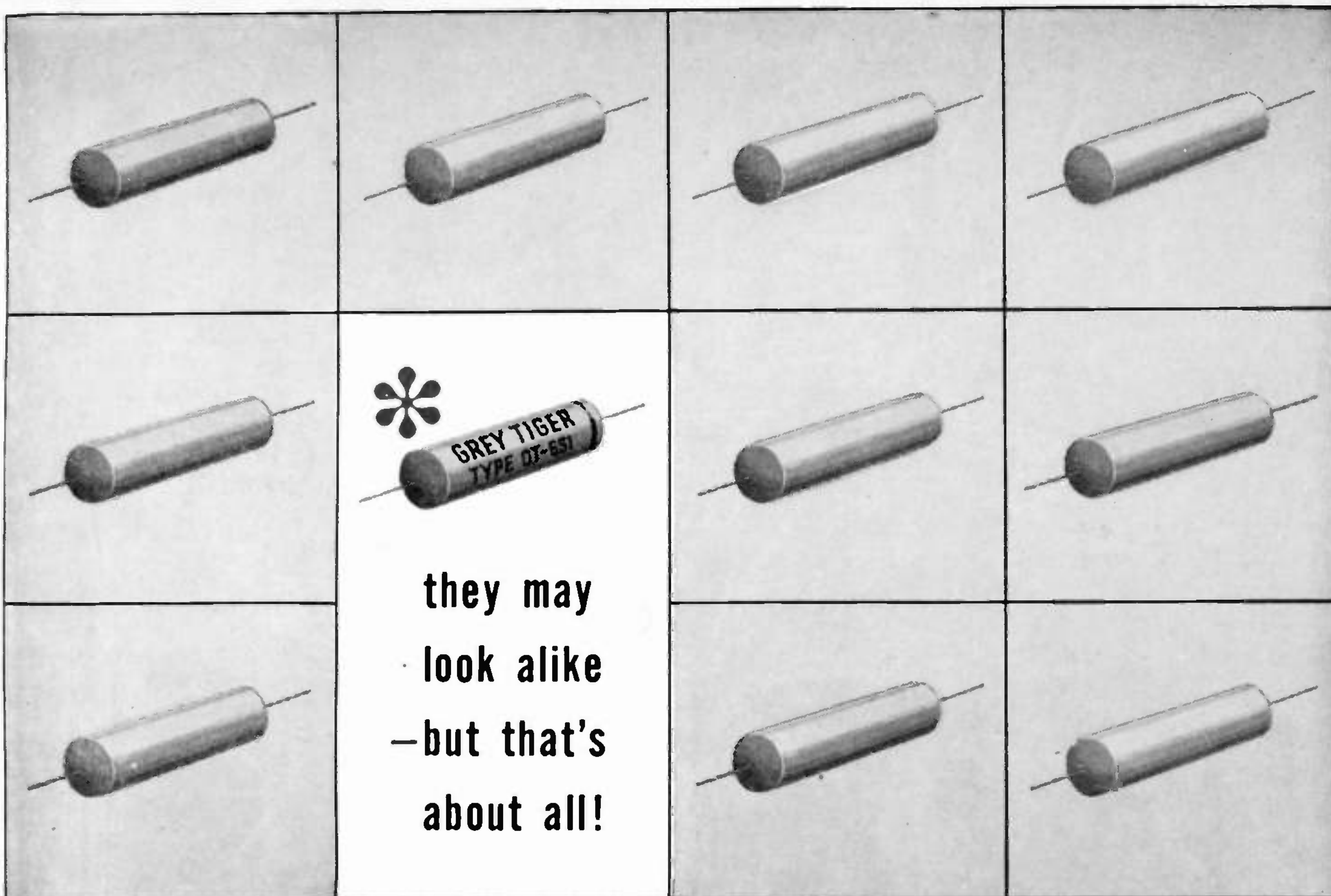

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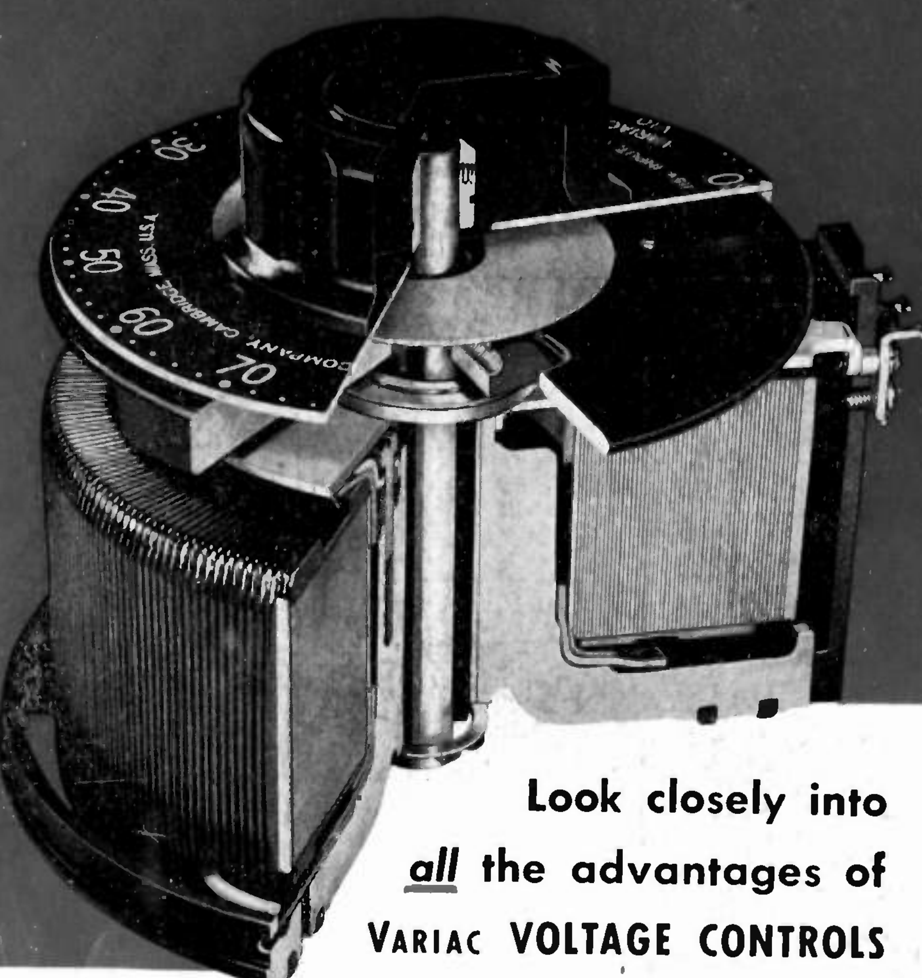

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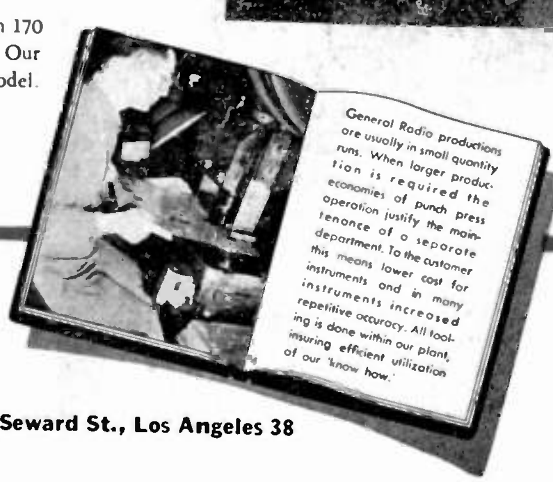

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