TELEVISION SYNCHRONIZING, SIGNAL, AND CONTROL EQUIPMENT

The development and utilization of television synchronizing and image signals requires intricate electronic equipment. Shown above are: a synchronizing signal generator, a monitoring receiver, a monoscope video signal generator, sources of audio-frequency signals, and a group of video and audio transmitters selectable at will. The equipment is intended for development and line testing of television receivers.
Transformers for every application

Linear Standard

Hyperm Alloy

Ultra Compact

Commercial Grade

Ouncer

Sub Ouncer

Special Series

Variable Inductor

Foremost Manufacturers of Transformers to the Electronic Industry

United Transformer Corp.

150 Varick Street

New York 13, N.Y.

Export Division: 13 East 40th Street, New York 16, N.Y.

Cables: "ARLAB"
LAST WORD IN LONGER LASTING FLUORESCENT LAMP CAPACITORS

Check these advantages of SUPEREX*
Fluorescent Ballast Capacitors

✓ SMALL SIZE
✓ LONG LIFE
✓ STABLE CHARACTERISTICS AT HIGH TEMPERATURES
✓ LOW POWER FACTOR
✓ NON-FIAMMABLE
✓ UNDERWRITERS' LABORATORIES LISTED

Exceptionally long life at high ambient temperatures is the prime requirement for capacitors used in fluorescent lamp ballasts. Major ballast manufacturers have made certain of this extra reliability by specifying Superex Capacitors.

Designed for use at temperatures of 75°C, Superex treated capacitors show exceptional stability of electrical characteristics on long-term life tests.

Superex capacitors are available in a full line of ratings and container shapes for every lamp auxiliary. Write for Bulletin SPA-110.

Solar Manufacturing Corporation
285 Madison Ave., New York 17, N.Y.
1918. This “peanut” tube, the Western Electric 215A, was developed for service in World War I. It was the first commercial tube whose filament was powered by a single dry cell. It made possible compact, lightweight radio equipment.

1919. The introduction of the copper-to-glass seal made water-cooled tubes practical. The resulting high-power tubes were used for broadcasting and for transoceanic radio-telephony.

1923. This was one of the earliest photoelectric cells. It was made by Western Electric for use in commercial picture transmission over telephone wires.

1937. This microwave generator, the 368A, was the first commercial tube to generate frequencies higher than 1500 mc. This type of tube was used by Western Electric in the first absolute altimeter.

1940. The beating oscillator, used in the great majority of radar systems. This tube generated a wave in the receiver with which the received microwave was reduced in frequency for amplification.

1912. The first effective high-vacuum tube, developed by the Laboratories for long distance telephony, was capable of operation at both audio and radio frequencies, and thus marked the beginning of modern electronics.
1940. Bell Laboratories produced the first American multicavity pulsed magnetron from a British model. The team of Western Electric and Bell Laboratories developed 75 new and improved magnetron designs by extending operation into the 10 cm, 3 cm and finally the 1 cm bands, and produced over 300,000 of these wonder tubes of World War II.

1942. This tiny 6AK5, operating in the vicinity of 400 mc, proved itself invaluable as an amplifier in radar receivers. Design specifications were supplied to other manufacturers by Western Electric to speed war production.

1945. The Bell Laboratories traveling wave tube, still in the research stage, amplifies over a band 40 times wider than present tubes—may be able to amplify dozens of color or black and white television programs simultaneously.

TODAY. These new forced air cooled FM transmitting triodes are among the latest in the line of tubes designed by Bell Telephone Laboratories and made by Western Electric. Their thorialed tungsten filaments, rugged construction, flexible terminal arrangements and many other features make them tops in performance in the 88 to 108 mc band.

OVER 34 years ago in the laboratories of Western Electric, De Forest's Audion was improved and developed into the high vacuum tube and put to work for the first time amplifying telephone and radio frequency currents. And for over 34 years Western Electric and its research associate Bell Telephone Laboratories have been foremost in designing new and better electron tubes.

Every tube shown here and many developments basic to the tube art are examples of that leadership. More than 10 years ago, for instance, Bell Laboratories first used microchemistry to determine what gases were destructive to tube elements, and with Western Electric developed a manufacturing technique to keep these damaging elements out—thus increasing tube life many-fold.

Every one of the more than 300 codes of electron tubes now being made by Western Electric from Bell Laboratories' designs has the same unequalled background of research and manufacturing skill.

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They Lick Humidity and Vibration at High Frequencies

STACKPOLE
Polytite TRIMMER
ELECTRODE CORES

Placed in fitted metal sleeves, Stackpole Polytite Trimmer Electrode Core Forms serve as variable capacitors that assure honest-to-goodness capacity stability in high-frequency circuits where humidity and vibration must be considered. The molded Polytite has a high dielectric constant. Cores are moisture repellent and carry a heavy dielectric coating that establishes a path of high leakage resistance between the electrodes. Since these electrode surfaces have short, symmetrical current paths, the inductance may be kept low enough for use in the 200-megacycle range. Standard types provide easy capacity adjustment with a maximum from 20 to 40 mmf., depending on the size.

Write for Stackpole Polytite Trimmer Data Bulletin

STACKPOLE CARBON COMPANY
Electronic Components Division • St. Marys, Pa.

Stackpole Polytite Trimmer Electrode Capacitors are well suited for minimum capacity adjustments in tuned circuits, installed across the tuning capacitor as in Figure 1 or across the tuning inductance as in Figure 2. Trimmers may be mounted directly to the tuning capacitor.

A typical application using two Polytite Trimmer Electrode Capacitors in a circuit where band-spread tuning is desired. Various bands may be covered by the switching of coils and preadjusted trimmers.
Here is a new development of importance to all users of specialty capacitors. It is General Electric's new silicone bushing—available only on G-E capacitors.

This new bushing gives greater dependability and longer life for capacitors. Being elastic, it is self-sealing—permanent, for all practical purposes, in both physical and dielectric properties. Inserted through the openings in the top of the capacitor casing, it seals by compression—without adhesives or gaskets. It retains its elasticity over a wide range of temperatures and will not shrink, pull away, or loosen during the life of the capacitor.

This bushing has other advantages—all of which add to the reliability of G-E capacitors. The single piece construction provides permanently high dielectric strength and insulation resistance. It is highly resistant to oils, alkalies, and acids; it will not support fungus growth.

Silicone bushings will be used on all General Electric Pyranol* capacitors having solder-lug terminals. This new G-E first is one more reason for selecting General Electric capacitors. Others, all adding to dependability and long life, include the positive sealing of casings by double rolling or roll-crimping and soldering, the use of highest grade materials and superior processing methods, with strict quality control.

Apparatus Dept., General Electric Company, Schenectady 5, N. Y.

This bushing represents one of the newest uses for the recently developed G-E family of chemicals called silicones. Permanently elastic, formed to close tolerances, it seals itself by compression to the capacitor casing.

Announcing a new line of television capacitors

"Hi-Vo-Kaps"

made with Centralab’s original Ceramic-X

Three types of terminals for flexibility, convenience

ROD TYPE: .160” diameter rod type terminals. Designed for use with conventional fuse or clip-type connections. Terminals are solid brass, silver-plated and soldered directly to electrodes.

SLOT-AND-THREAD TYPE: .160” diameter with 3/16” x 3/16” slot in one terminal. Other terminal tapped 6-32, 3/16” deep for “twinning” or convenient chassis mounting.

DUO-THREAD TYPE: one terminal tapped 6-32, 3/16” deep full threads. Other terminal, 6-32, male thread 3/16” length. Designed for convenient series or tapped series connections.

The smallest high voltage capacitors ever designed exclusively for television circuits!

ANOTHER "FIRST" for Centralab! "Hi-Vo-Kaps" are made with Centralab's original Ceramic-X, combining high voltage, small size and terminal connections to fit virtually any television application!

Designed and developed by Centralab in response to stated requirements of television project engineers, "Hi-Vo-Kaps" are for use as filter and by-pass capacitors in video amplifiers — for high DC voltages with small component AC voltages (not for use in temperature compensation or resonant circuits).

Ratings: 10,000 WVDC, 15,000 VDC flash test, 500 mmf., — 50% — 20% capacity at 1 megacycle (2½% higher at 1 kilocycle).

Dimensions: diameter — .990”, length — .510”. Overall length varies with terminal types, maximum—1.597”. Send for Bulletin 946.
HERE’S FLAT RESPONSE UP TO 700 MC

410A VACUUM TUBE VOLTOMETER

with its new -hp- low-capacity diode probe, measures all the important radio voltages without disturbing circuits under test.

CHECK THESE FACTS ABOUT THE NEW -hp- PROBE*:

- Small size for ease in contacting hard-to-get-at components.
- Ultra-short leads, direct grounding assures high frequency response.
- Rugged, mechanical construction, dural shell, polystyrene insulation.
- Specailly-designed diode has short transit time, low input capacity, high resonant frequency of 2000 mc.
- Detachable tip lowers input capacity, shortens diode lead, utilizes maximum capabilities of diode.

The specially-designed diode, in combination with the -hp- probe design, makes possible the exceedingly flat frequency response shown graphically in Figure 1.

With this flat frequency response are combined the factors of low input capacity and high input resistance. The variation of these factors with frequency is shown in Figure 2. The input resistance and reactance are high throughout the entire range of the instrument, and thus measurements are made without appreciable detuning or loading of circuit. Maximum measuring accuracy is assured.

In addition to swiftly, easily, accurately making uhf radio measurements, this -hp- 410A is a convenient voltage indicator up to 3000 mc. And it serves equally well as an audio or d-c voltmeter, or an ohmmeter. A-c measurements are made in 6 ranges ...full scale readings 1 to 300 v. D-c full scale readings from 1 to 1000 v in 7 ranges. Input resistance all ranges =100 megohms. As an ohmmeter, the -hp- 410A measures resistances from 0.2 ohms to 500 megohms in 7 ranges.

In short, this -hp- 410A Vacuum Tube Voltmeter is ideal for obtaining most important parameters in radio design, manufacture, or servicing. Write today for full details. Hewlett-Packard Company, 1407D Page Mill Road, Palo Alto, California.

PROCEEDINGS OF THE I.R.E. June, 1947
See what your ears miss!

SHERRON
R.F. NULL
DETECTOR

Noise can't interfere with the indications registered on this new Sherron instrument. Where din and hubbub would nullify aural manifestations, you can count on the visual features of Model SE-518 to provide the findings, clearly, unmistakably. Instantaneously responsive to changes of signal level, the R. F. Null Detector is equipped with a Cathode Ray indicator... As a signal generator to provide power at 1 MC, the Sherron R. F. Null Detector is invaluable. It also serves as a sensitive detector at the same frequency. Both generator and detector are housed in the same cabinet.

Frequency: 1 MC
Generator Output: 0-3 volts
Detector Gain: 500,000 plus
Harmonic Suppression: 2nd down more than 100 db
Power Requirements: 115 volts, 60 cycles, 120 watts

SHERRON ELECTRONICS CO.
Division of Sherron Metallic Corporation
1201 FLUSHING AVENUE • BROOKLYN 6, NEW YORK
An enviable record based on advanced engineering and modern design

• More and more station owners every day are turning to Raytheon for the very finest in broadcast equipment. Raytheon is leading the way with simplified circuit design, thorough engineering and complete dependability.

Across the nation, enthusiastic station owners and engineers (both AM and FM) praise the high fidelity, servicing accessibility and low-cost maintenance of Raytheon broadcast equipment—from Single-Channel Remote Amplifiers to 5 KW Transmitters. With Raytheon equipment they find it far easier to set up programs—and operation is so simple and logical that errors are cut to a minimum.

Be sure you have all the facts before you buy. Investigate Raytheon’s complete line of speech input equipment and both AM and FM Transmitters ranging from 250 to 10,000 Watts.

These superb Raytheon products assure the most practical application to your specific broadcast problem . . . bring you the finest in modern high fidelity and engineering excellence. Write or wire for illustrated specification bulletins, including complete technical data.

Raytheon
Excellence in Electronics
RAYTHEON MANUFACTURING COMPANY
BROADCAST EQUIPMENT DIVISION
7475 N. ROGERS AVE., CHICAGO 26

Devoted to Research and Manufacturing for the Broadcasting Industry
COMPACT ENERGY FOR PHOTOFLASH CAPACITORS
Progress in practical flash photography has been greatly facilitated by new smaller, lighter capacitors incorporating the exclusive Sprague Vitamin Q impregnant. Write for engineering bulletin No. 201.

GUARDING AGAINST FLUORESCENT BALLAST FAILURES
A major fluorescent lighting problem has been one of finding ballast capacitors to withstand the combination of severe temperature and voltage conditions—and again Sprague Vitamin Q impregnant has proven the answer. Sprague Fluorescent Ballast Capacitors rated at 350v AC not only give maximum life under normal temperature and voltage conditions, but can be operated at 460v AC at 85° C. for 1,000 hours—without deterioration or major change in power factor. Thus they assure adequate safety factor under blink start conditions.

It's all done with ‘VITAMIN Q!

The history of capacitor progress is inseparably linked with the development of new and better dielectrics. Throughout the years, the aim has been to increase the amount of energy that can be stored in a capacitor of given size and to improve performance characteristics all along the line. The most remarkable advance in these respects has come with the development of the exclusive oil dielectric—Sprague Vitamin Q. Throughout industry, Sprague Capacitors impregnated with this material are setting new standards for smaller, lighter units for dependable operation at higher voltages and higher temperatures and for greatly improved insulation resistance.

The units illustrated are typical of the many new capacitor designs now available using Sprague Vitamin Q.
Friendly, tactful, impartial, trained to serve, these Hytron commercial engineers form the liaison between us—maker and user of electronic tubes. Few in the radio tube plant can be circuit specialists. Few outside the tube plant can be tube specialists. Both of us need these commercial engineers trained to see clearly both sides of our common problems and help us solve them.

Often their job begins with a request for advice in selecting a tube. Investigation of the circuit application helps them recommend an available type, a slight redesign, or a brand new type. If a new type is found to be the only practicable and economical solution, they cooperate with design and production engineers to achieve the performance desired.

Specification of adequate factory testing procedures and preparation of characteristics sheets do not end their work. Returns are closely checked. If trouble occurs, they go into the field, help dig out the facts, and offer possible solutions—improvements in tube or application. And they stick tenaciously with the problem until it is solved.

Using a wealth of test equipment and know-how, these boys really sweat to make it easy to make Hytron tubes which will make you happy. Busy as the one-armed paperhanger, yet they always welcome the tube problems of equipment engineers. They are nice guys, and we thought you would like to meet them.
HIGH VOLTAGE; NO DANGER

Portable - Rugged - Safe!

High voltage is the keynote of modern oscillography. Especially for brilliant traces at ultra-high speeds.

Type 263-A High-Voltage Power Supply was designed with present and future needs in mind. It provides a dependable yet inexpensive power supply for modernizing and extending the usefulness of certain types of cathode-ray oscillographs when examination of extremely high writing rates is required.

So here's a complete high-voltage power supply. Suitable for any application where high voltage at low current is called for. Consists of radio-frequency oscillator with its own power supply, an r.f. step-up transformer, a half-wave rectifier, and a high-voltage filtering and metering system.

Compact. Light. So designed that inexperienced personnel may handle it with safety. And it is made still safer in case of accidental contact with high voltage, because very little power is stored in its filtering circuit. Furthermore, no equipment damage will result if output is short-circuited. Rugged mechanical construction permits field or laboratory use.

Surely Type 263-A is a "must" instrument whether for high-voltage oscillography or general use!

Details on request!

DU MONT Type 263-A
HIGH-VOLTAGE POWER SUPPLY

Salient Oscillographic Features . . .
✓ 10,000 volt intensifier potential available for use with cathode-ray oscillographs.
✓ Visual observation of single transients hitherto invisible.
✓ Photography of extremely high writing rates (for example, 2000 km./sec. on SRP11 at 10 kivolt).
✓ Observation of entire waveshapes of short duration on long persistence screens.
✓ Conveneient use with Type SRP-A Multi-band High-voltage Cathode-ray Tube.

Working Details . . .
✓ Continuous variable d.c. output from 5,000 to 10,000 volts with loads up to 200 microamperes.
✓ Regulation within 2% from no load to 200 microampere load.
✓ Ripple voltage on output less than 0.5%.
✓ Power supply: 115 volts, 50-60 cps.
✓ Power consumption: 100 watts.
✓ Dimensions: 10½" h. x 8½" w. x 14¾" d.
✓ Weight: 24 pounds.
BROADCASTING that earns the approval of station managers and listeners alike under any and all local conditions for reliability, efficiency and economy.

Collins 21A 5kW Air Cooled Broadcast Transmitter made by Collins Radio Company, 11 West 42nd Street, New York 18, N.Y.

The new Collins 21A has been the choice of keen executives for close to a score of installations in recent months. Knowledge and experience gained by Collins engineers during war time are reflected in improved design, longer life, higher safety factors and unusual standards of trouble-free operation.

The operation

The equipment

The AMPEREX tubes

That do the job!

AMPEREX experience in communication goes back a quarter of a century. The same record of performance, long life and economy marks Ampex tubing for industrial, rectification, electro medical and special purpose use. As tube specialists concerned with all electronic developments Ampex engineers are in a position to give detached counsel and information.

Write Application Engineering Department.

Power Tube Specialists Since 1925

AMPEREX

Electronic Corporation

23 Washington Street, Brooklyn 1, N.Y., Cables: "ARLAB"

In Canada and Newfoundland: Rogers Majestic Limited, 622 Fleet Street West, Toronto 38, Canada
INTRICATE LABORATORY TECHNIQUES GUARD QUALITY OF TUNGSTEN IN SYLVANIA TUBES

Basic Studies of Wire Conducted at Each Stage of Production to Insure Electronic Tube Perfection

Two of the many metallurgical tests constantly carried on by Sylvania Electric are illustrated here.

To insure electronic tube perfection — to have Sylvania radio tubes measure up to long-established Sylvania standards — every important type of research technique is utilized.

Here electron microscopes, giving magnifications of thousands of times, are employed. Hardness testers, sag testers, gas analysis equipment, tensile testers are but a few of the methods used to guard the high quality of tungsten utilized.

Prior to sintering operation shown at left, tungsten bars of approximately ½" square are prepared by pressing finely divided metal powder under hydraulic pressures of up to 300 tons. The equipment used to pursue such studies is illustrated in the above photograph.

Both of the photographs shown here are indicative of the fundamental studies that have resulted in the development and maintenance of tungsten wire of superior quality.

Radio Tube Division, Emporium, Pa.
How AMPHENOL AN Connectors

Step Up Your Profit Potential

Standardized AN connectors provide a fast, fool-proof way to connect any industrial electronic equipment which frequently must be disconnected from associated equipment or power source.

Their use also permits the prefabrication of associated wiring to accommodate one or many circuits. This greatly simplifies and lowers the cost of electronic installations. AN connectors also permit such equipment to be completely tested at the factory before shipment to user. Upon arrival it then can be connected for operation in minutes.

These advantages combine to widen the field in which electronics may practicably be applied. Thus they offer an increased sales and profit potential to makers of electronic devices.

The Amphenol AN connector family offers you a number of important points of mechanical and electrical superiority. It is comprised of over 200 styles of dielectric inserts. These are interchangeable in any of the five major Amphenol metal shell designs (each of which is available in eighteen sizes). The practically endless variety of possible combinations offers an efficient solution to any industrial electronic connector problem.

Amphenol inserts handle currents up to 200 amperes, voltages up to 22,000. Housings include types which are pressure-proof, moisture-proof and explosion proof. Standard elements also are available for thermocouple installations.

Amphenol, long the leading builder of AN connectors for aircraft, ships, tanks and ordnance, is still completely tooled for large scale production. This makes these connectors available to industry at costs far below prewar levels. Write today for complete technical and cost data.

AMERICAN PHENOLIC CORPORATION
1830 SOUTH 54TH AVENUE, CHICAGO 50, ILLINOIS

Coaxial Cables and Connectors - Industrial Connectors, Fittings and Conduit - Antennas - Radio Components - Plastics for Electronics

PROCEEDINGS OF THE I.R.E. June, 1947
A REVOLUTIONARY NEW EIMAC TRIODE

YES... The 3X12500A3 is truly revolutionary... packaged power... that will fill not several, but all applications for a power-amplifier or oscillator from zero to 110 Mc. It will do a low frequency job better than “special low frequency” tubes. Its performance at vhf has long been the aim of vacuum tube researchers. The 3X12500A3 is smaller (over-all 11”x9”) and lighter (net 32 lbs.) than any comparable tube.... Yes, it is truly a revolutionary tube.

Audio
Induction heating
Broadcasting
Dielectric heating
Communication
Television
Industrial
FM Broadcasting
Research

RADIO FREQUENCY POWER AMPLIFIER
Grounded-Filament Circuit
Class-C Telegraphy (Key-down conditions, per tube)
MAXIMUM RATINGS (Frequencies below 85 Mc.)

<table>
<thead>
<tr>
<th>D.C. Plate Voltage</th>
<th>5000 Max. Volts</th>
</tr>
</thead>
<tbody>
<tr>
<td>D.C. Plate Current</td>
<td>8 Max. Amps</td>
</tr>
<tr>
<td>Plate Dissipation</td>
<td>12,500 Max. Watts</td>
</tr>
<tr>
<td>Grid Dissipation</td>
<td>600 Max. Watts</td>
</tr>
</tbody>
</table>

TYPICAL OPERATION (Frequencies below 50 Mc., per tube)

<table>
<thead>
<tr>
<th>D.C. Plate Voltage</th>
<th>3500 - 4000 - 5000 Volts</th>
</tr>
</thead>
<tbody>
<tr>
<td>D.C. Plate Current</td>
<td>-360 - 400 Volts</td>
</tr>
<tr>
<td>D.C. Grid Voltage</td>
<td>7 - 9 Amps</td>
</tr>
<tr>
<td>D.C. Grid Current</td>
<td>1.7 - 1.9 Amps</td>
</tr>
<tr>
<td>Peak R.F. Grid Input Voltage</td>
<td>1.3 - 0.95 - 1.35 kV</td>
</tr>
<tr>
<td>Grid Dissipation</td>
<td>400 - 350 - 590 Watts</td>
</tr>
<tr>
<td>Plate Input</td>
<td>25.2 - 25.6 - 40 kV</td>
</tr>
<tr>
<td>Plate Dissipation</td>
<td>5.7 - 5.6 - 10 kV</td>
</tr>
<tr>
<td>Plate Power Output</td>
<td>20 - 20 - 20 kV</td>
</tr>
</tbody>
</table>

RADIO FREQUENCY POWER AMPLIFIER
Grounded-Grid Circuit
Class-C FM Telephony or Telegraphy
MAXIMUM RATINGS (Frequencies below 110 Mc.)

<table>
<thead>
<tr>
<th>D.C. Plate Voltage</th>
<th>4000 Max. Volts</th>
</tr>
</thead>
<tbody>
<tr>
<td>D.C. Plate Current</td>
<td>8 Max. Amps</td>
</tr>
<tr>
<td>Plate Dissipation</td>
<td>12,500 Max. Watts</td>
</tr>
<tr>
<td>Grid Dissipation</td>
<td>600 Max. Watts</td>
</tr>
</tbody>
</table>

TYPICAL OPERATION (110 Mc., per tube)

<table>
<thead>
<tr>
<th>D.C. Plate Voltage</th>
<th>3700 - 4000 Volts</th>
</tr>
</thead>
<tbody>
<tr>
<td>D.C. Grid Voltage</td>
<td>-450 - 550 Volts</td>
</tr>
<tr>
<td>D.C. Plate Current</td>
<td>-7.2 - 7.4 Amps</td>
</tr>
<tr>
<td>D.C. Grid Current</td>
<td>0.9 - 1.1 Amps</td>
</tr>
<tr>
<td>Driving Power (approx.)</td>
<td>6.8 - 7.6 kW</td>
</tr>
<tr>
<td>Useful Power Output</td>
<td>27.4 - 30 kW</td>
</tr>
<tr>
<td>Apparent Overall Efficiency</td>
<td>102 - 103% per cent</td>
</tr>
</tbody>
</table>

EITEL-McCULLOUGH, Inc.
1653 San Mateo Avenue, San Bruno, California

Follow the Leaders to

EIMAC TUBES
The Power for R-F

EXPORT AGENTS: FRAZAR & HANSEN, 301 CLAY ST., SAN FRANCISCO 11, CALIFORNIA

PROCEEDINGS OF THE I.R.E. June, 1947
This page advertisement in electronic and communications magazines in 1939 announced that Major Edwin H. Armstrong's pioneer antenna structure for FM transmission was equipped with AlSiMag 196 insulators.

Most of the original AlSiMag insulators in W2XMN are still in use today. They are giving entire satisfaction in spite of the fact that one of the transmission lines up the tower, originally designed for 42 megacycles, is carrying 92 megacycles.

There has been no electrical failure of any AlSiMag insulator in W2XMN. A few have been replaced after heavy ice falls. There is no insulator in existence today which will stand up when squarely hit by a heavy ice fall with drops of several hundred feet. That is one of the problems challenging our Research Division.

In the spring of 1947, W2XMN will replace the vertical transmission line conductors with conductors of considerably larger size. These new and larger conductors will have new and larger insulators... of AlSiMag. Perhaps that is the best evidence of the satisfactory performance of AlSiMag insulators in the World's Pioneer FM Station.

ALSiMag Insulation in the Pioneer FM Antenna Structure

Still Going Strong!
A new binding post, Type DF30, manufactured by The Superior Electric Company, 47 Church Street, Bristol, Conn., meets the need for a multi-purpose electrical connector. In contrast to the usual connectors which permit only one or two methods of connection, the new binding post offers five ways of connecting leads; permanent clamping of wire up to size 1/2 through the center hole, looping of wire around the center shaft and clamping, plug-in connection of a standard 1/4 banana plug, clip-lead connection, and spade-lug connection. In addition to the versatility of connection, this unit provides complete insulation of the binding post from the mounting panel. Rated at 30 amperes, it may be mounted on any panel up to 1" thick.

Roto Switch

A new miniature rotary switch only 1" in diameter, with a contact pressure of 21/2 pounds, has been developed by Grayhill, 1 No. Pulaski Road, Chicago 24, Ill. Designated as the Series 5000 Roto Switch, it can be used in almost any circuit combination up to 5 amperes, breaking up to 1 ampere at 110 volts. The switch can be rotated 360° in either direction.

Plug-in Amplifiers

The Langevin Company, 37 W. 65 Street, New York 23, N. Y., announces two new types of plug-in amplifiers which, it is stated, will provide complete audio facilities with a minimum of different types of tubes, facilitate maintenance, and conserve space without effecting quality or overload safety factors.

Both amplifiers, Type 116-A (right) and Type 117-A (left), have identical frequency response characteristics of ±1 db. over the range of 30 to 15,000 cycles. Type 116-A has a 40 db. gain and may be used as a pre-amplifier or booster amplifier. Type 117-A has a 50 db. gain and may be used as a program, booster, or monitor amplifier.

Panel Meters

Shurite Meters, 61 Hamilton St., New Haven 8, Conn., announces a complete line of alternating and direct current, 2" and 2 1/2", panel meters. Two round cases and one rectangular case are available in AC and DC ammeters, milliammeters, voltmeters, and also resistance meters. All DC meters are polarized-vane solenoid type, and AC meters are double-vane repulsion type, with an accuracy within 5%.

All models are flush-mounting type of black-enamed brass construction. Bracket, ring or screw mounting, and narrow or wide flange denote design differences. Zero adjusters are supplied when required on two of the DC case types.

Radio Pack Set

Designed primarily for railroad two-way radio communications, the type MRT-2B VHF pack-set has been recently designed for portable use by Bendix Radio Corporation, E. Joppa Road, Baltimore, Md. The retractable vertical-rod antenna, when fully extended, measures 36 inches. The antenna is so designed that only when it is fully extended is the set turned "On."

The overall size of this unit is approximately 11 1/2" high, 9" wide, and 4" deep. It weighs slightly more than 15 pounds including the power supply. Two power supplies are provided so that one may be recharged while the other is in use. Transmitter and receiver are both crystal controlled.

Streamlined Microphone

A new microphone, "The Conneaut," Type 600-S, having a relatively high output and wide frequency range up to 10,000 cycles has recently been placed on the market by The Astatic Corporation, Conneaut, Ohio.

(Continued on page 48A)
The complete line of De Mornay-Budd standard test equipment covers the frequency range from 4,000 mcs. to 27,000 mcs. It provides all R. F. waveguide units necessary for delicate, precision test work requiring extremely high accuracy in attenuation measurements, impedance measurements, impedance matching, calibration of directional couplers, VSWR frequency measurements, etc.

To eliminate guesswork, each item of this De Mornay-Budd test equipment is individually tested and, where necessary, calibrated, and each piece is tagged with its electrical characteristics. All test equipment is supplied with inner and outer surfaces gold plated unless otherwise specified.

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

The three test set-ups illustrated above include:

- Tube Mount
- Flap Attenuator
- Frequency Meter
- Calibrated Attenuator
- Tee
- Stub Tuner
- Tunable Dummy Load
- Standing Wave Detector
- Type "N" Standing Wave Detector
- Directional Coupler
- High Power Dummy Load
- Cut-Off Attenuator
- Stands, etc.

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97% OF ALL
RADAR SETS

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Especially when it comes to radio parts. That's why National parts are precision-made with tolerances measured as close as .0002".

Operational results justify this close attention to detail for every National precision condenser is mechanically and electrically interchangeable and can be depended upon to fit the specifications called for. Production flows smoothly when you use National parts because their closely-tooled tolerances and sturdy construction make replacements unnecessary...

Send for your copy of the new National catalog containing over 600 parts today.

Please write to Department 17, National Company, for further information.

This PW Condenser is of extremely rigid construction with Stroolite stator insulation. The drive is through an enclosed preloadable worm gear with 20 to 1 ratio and the rotor shaft is parallel to the panel. Plate shape is straight-line frequency when the frequency range is 2:1.

PW Condensers are available in 2, 3, or 4 sections in either 160 or 225 mfd per section. A single-section PW Condenser with grounded rotor is supplied in capacities of 150, 200, 350 and 500 mfd, single spaced, and capacities up to 125 mfd, double spaced.

NPW-O uses parts similar to the NPW Condenser. Drive shaft perpendicular to panel. One TX-9 coupling supplied.

The PW-O uses parts similar to the PW Condenser. Drive shaft parallel to panel. Two TX-9 couplings supplied.

The NPW model is similar to the other PW Condenser models, except that the rotor shaft is perpendicular to the panel. Three sections... each 225 mfd.

The PW-D micrometer dial can be read directly to one part in 500. It revolves ten times in covering the complete range and fits a 1/4" diameter shaft.
HERE'S A NEW HF cable that will keep your FM and Television receivers working at peak performance—free from locally-induced interference, even in the most adverse locations. Where the performance of such costly equipment is at stake, it will pay you to specify Federal's KT 51—the finest high frequency lead-in cable available. More costly—but worth more!

The twisted, dual-conductor cable cancels any noise or signals not stopped by the double braided shields...because it's electrically balanced and stays that way in service, in any position. It's a rugged cable, too—remarkably resistant to abrasion, acids, alkalis, oils and greases, as well as smoky atmospheres and weather.

Don't let the lead-in wire be the "weak link" in otherwise perfect equipment. Be sure it's KT 51—the HF cable with the "right twist" to assure interference-free operation. For complete details, write to Dept.

### ELECTRICAL CHARACTERISTICS

<table>
<thead>
<tr>
<th>Frequency (mc)</th>
<th>Nominal Attenuation (db/100 ft.)</th>
<th>Maximum Capacity Unbalance</th>
<th>Nominal Characteristic Impedance (ohms)</th>
<th>Nominal Capacitance per ft. (uf)</th>
<th>Volts (rms)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10 mc</td>
<td>0.9</td>
<td>1%</td>
<td>95</td>
<td>16</td>
<td>2000</td>
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<tr>
<td>30 mc</td>
<td>1.7</td>
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<td>100 mc</td>
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<td>300 mc</td>
<td>7.0</td>
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<tr>
<td>400 mc</td>
<td>10.0</td>
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</table>

Federal Telephone and Radio Corporation

In Canada: Federal Electric Manufacturing Company, Ltd., Montreal
Export Distributor—International Standard Electric Corporation, 67 Broadway, N. Y. C.
This 6C22 tube, the result of a closely-guarded development during World War II, is a modified version of the tube used extensively for pulsing signals in radio transmission and may have had a vital influence in jamming enemy radar communications. Peace-time pursuits indicate that it will play an important part in furthering the development of television, having already proved of great value in a transmitter employed for color television. An unusual feature in the construction of this tube is to be seen in the one-piece formation of the anode and water-cooled radiator. The anode and grid ring are produced from Certified Oxygen Free High Conductivity Copper Bar, Revere Alloy 103-C, being formed by cold working in a 600-ton coining press.

Machining consists of drilling the center hole and milling the radiator slots. Each piece receives a special rolling operation in the area where it is sealed to glass. The grid ring which extends through the glass structure performs a dual function in supporting the grid internally and providing an external connection. As in other types of vacuum tubes Certified Oxygen Free High Conductivity Copper is used for ease of out-gassing and excellent glass bonding characteristics.

This type 6C22 vacuum tube was developed and is manufactured by the Federal Telephone and Radio Corporation, Clifton, New Jersey, and is rated at 1000 watts, plate dissipation at 600 mc.
Modern designs — new finishes — promises of greater performance. These are the things that sell today's products.

But the real features that keep the products sold are the hidden values — the parts inside the product that insure performance promises being kept.

SUCH FEATURES ARE INCORPORATED IN EL-MENCO CAPACITORS, WHOSE QUALITY IS BEYOND QUESTION.

THE ELECTRO MOTIVE MFG. CO., Inc. Willimantic, Conn.

Foreign Radio and Electronic Manufacturers communicate direct with our Export Department at Willimantic, Conn. for information.
Centralab reports to

1 Designed for peak AM and FM performance plus maximum reliability and long service life, Centralab's new slide switch now gives you flat, horizontal design that saves space, permits convenient location to coils, reduced lead inductances. "Twisted ear" mounting on base or panel from .038" min. to .052" max. Optional size or length of unit — min. 3 clips per side, max. 20 clips per side. 2 or 3 position, shorting type contacts. Movement of slide per position — ¼ inch. Send for bulletin 953.

2 For transmitters, power supply converters, X-ray equipment, etc., CRL's medium-duty power switches are now available. Efficient performance up to 20 megacycles.

3 First commercial application of the "printed circuit" and now available for the first time, Centralab's new Couplate offers a complete interstage coupling circuit which combines into one unit the plate load resistor, the grid resistor, the plate by-pass capacitor and the coupling capacitor.

JUNE 1947

Revolutionary new Slide Switch reduces lead inductance for improved AM-FM performance!
Electronic Industry

**NOW SEE HOW THIS REPLACES THIS**

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*Other Values Available*

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Integral Ceramic Construction: Each Couplate is an integral assembly of "Hi-Kap" capacitors and resistors closely bonded to a steatite ceramic plate and mutually connected by means of metallic silver paths "printed" on the base plate. Think of what that means in terms of time and labor savings! Send for bulletin 943.

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**There's none better than this line of ceramic capacitors which combines economy, small size and extreme dependability.**

**Only four soldered connections are now required by the Couplate instead of the usual eight or nine... (see above). That means fewer errors, lower costs!**

Look to Centralab in 1947! First in component research that means lower costs for electronic industry. If you're planning new equipment, let Centralab's sales and engineering service work with you. Get in touch with Centralab!

Centralab

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FOR FM AND TELEVISION

-110 to 220 mc frequency at max ratings
-1.5 to 6.4 kw typical Class C output

GENERAL ELECTRIC'S great 1947 series of ring-seal power tubes spells more efficient performance to those who build—or use—FM and television transmitters. Modern as tomorrow's telecast, these v-h-f tubes need minimum neutralization... are directly designed for grounded-grid circuits... meet in every way the new requirements of new station equipment going into service.

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FIRST AND GREATEST NAME IN ELECTRONICS

PROCEEDINGS OF THE I.R.E. June, 1947
A lot of useful new things in electronics have come out of the war. Take that 6AK5 cathode type, voltage amplifier. TUNG-SOL made millions of them for radar receivers. They are the perfect, popular priced tube for FM, television and most applications up to 200 Megacycles.

The TUNG-SOL 6AK5 is small and compact with the internal element structure mounted on short direct leads to the glass bottom base. There is a minimum of internal 'ginger bread.' This simple ruggedness of design, high electrical efficiency, low temperature cathode assure long trouble-free service.

"The wide band merit factor of the 6AK5 is as much as 25% greater than that for tubes previously used in this service. The small grid-cathode spacing, the low cathode lead inductance and low input capacitance all provide the exceptionally high input impedance of about 10,000 ohms at 100 megacycles. Together with the 5,000 micromhos transconductance, all these figures mean that 'low frequency' gain figures can become a reality in your new FM or television receiver design. Carried further, this tube offers an ideal solution to the frequency converter problem. In mixer operation with a separate oscillator, the 6AK5 gives good gain and remarkably low noise which will increase the usable sensitivity of your television receiver.

"You are planning on some new equipment, Joe. Why don't you ask the TUNG-SOL Engineers about using 6AK5's in it? Those boys know their stuff and will give you sound, impartial advice. You know TUNG-SOL is headquarters for Miniatures."
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For applications where low temperature rise, space and weight are vital factors. Encased in special phenolic compound for complete protection. Unique design of mounting bracket aids rapid heat dissipation. Multi-section feature permits exceptional flexibility for voltage dividing applications.

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PROCEEDINGS OF THE I.R.E. June, 1947 29A
There are still millions of dollars worth of war surplus transmitters, receivers, tubes and various other types of electronic equipment being offered for sale to manufacturers, jobbers and wholesalers.

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For your convenience we are listing the names and addresses of those companies appointed to serve you. They will be happy to quote items, price and delivery. Just call, write, or phone and see how you can “Save with Surplus.”

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Belmont Radio Corporation
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Chicago 9, Illinois

Carr Industries Inc.
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Brooklyn 16, N. Y.

Communication Measurements Laboratory
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New York 6, New York

Cole Instrument Co.
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Los Angeles 15, California

Electronic Corporation of America
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Electro-Voice, Inc.
Carroll & Cecil Streets
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Emerson Radio & Phonograph Corp.
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New York 11, N. Y.

Essex Wire Corporation
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General Electric Company
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Schenectady 5, New York

General Electronics Inc.
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New York 23, N. Y.

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New York 1, New York

Hoffman Radio Corporation
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Hytron Radio & Electronics Corp.
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RCA TUBE APPLICATION ENGINEERING LABORATORIES—
RCA maintains tube application laboratories at Harrison, Lancaster and Chicago. The services of these laboratories are at the disposal of all RCA tube customers.

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RCA works years ahead in tube design—anticipates future requirements. That's why you get the types of tubes you want when you want them.

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RCA DISTRIBUTION—
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RCA PRICING—
Mass-production techniques and the RCA "Preferred-Type Tube Plan" have consistently operated to reduce manufacturing costs—which means lower prices to you.

RCA ENGINEERING LEADERSHIP—
The vast resources of experience and ability that account for RCA's engineering leadership, are of direct benefit to RCA customers—a final reason why it pays to deal with RCA.
PROCEEDINGS OF THE I.R.E.

(Including the WAVES AND ELECTRONS Section)

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Copyright, 1947, by The Institute of Radio Engineers, Inc.
J. E. Brown was born on September 11, 1902, in Greenport, Long Island, and studied electrical engineering at Cornell University.

From 1924 to 1937 he was a radio inspector with the Radio Division of the United States Department of Commerce, the Federal Radio Commission, and the Federal Communications Commission. During this period he was engaged in the development of radio field-intensity measuring equipment and methods of measuring field intensities of broadcast stations. He resigned from the Federal Communications Commission in 1937 to assume direction of television activities of the Zenith Radio Corporation. In 1943 he became assistant vice-president and chief engineer of the company, and he still holds this position.

Mr. Brown joined The Institute of Radio Engineers as an Associate in 1924, transferred to Member in 1928, and to Senior Member in 1943. He is a charter member of the Detroit Section, where he served as vice-chairman in 1932; in 1937 he was chairman of the Chicago Section. On the I.R.E. Board of Directors in 1938 and 1947, he has also been a member of many Institute Committees which include currently Nominations, Television, Radio Receivers, and Modulation Systems.

Active in the Radio Manufacturers Association, the National Television System Committee, and the Radio Technical Planning Board, Mr. Brown has been chairman of committees on television in each of these organizations. He is a member of the Radio Engineers Club of Chicago.
Those who refuse cynically to believe that "language was given to man to conceal thought" will find much to applaud and to follow in the following clear guest editorial from the editor of Wireless World (London). Great benefits flow from the use of terminology which is clear, uniform, and unequivocal. And serious evils and confusion follow casual, vague, or confusing use of language.

—The Editor

Radio Jargon

H. F. SMITH

For better or worse, a very large proportion of the technical radio terms used in all countries are taken directly from English or else translated literally from that language. This would be cause for self-congratulation to both Americans and British, who share the English tongue, if we could rest assured that the terms we have chosen were descriptive, lucid, and free from ambiguity.

Few would argue that this object has been achieved; indeed, we find growing confusion and lack of descriptiveness. The position naturally deteriorated during the war, when intensive development took place under conditions where the coining of undescriptive or even deliberately misleading words was condoned or even positively encouraged, on the grounds that it would convey no useful information to the enemy.

The old-timer, who has grown up in the radio art, can generally look after himself, but let us not make life unnecessarily difficult for the rising generation and for the non-English speaker. What justification have we for using "oscilloscope" and "oscillograph" to describe the same thing? By the ordinary usage of language the words would not be taken for synonyms, and their use as such must have puzzled many students and outsiders.

There is not even uniformity between American and British terminology, but it would be unrealistic and over-idealistic to plead for a standard technical terminology throughout the English-speaking world. The American could no more easily be divorced from his "tube" than the Englishman from his "valve." Equally unrealistic at this late stage is the idea that well-established terms of long standing, however confusing or undescriptive they may be, can be cleared away.

But it is perhaps not too late to plead for some measure of agreement on the use of new words for describing new things. Take "radar"; an official United States publication tells us that a radio echo is inherent in all radar systems. But the word is already being used here (under protest from some of us, I should add) to describe devices that make use of a triggered response instead of a natural echo. More loosely, it is beginning to be applied to any new radio position-finding system.

This is not just a plea for pedantic accuracy, and still less a diatribe against mere inelegance in radio terminology. It is a matter of more than domestic or even Anglo-American concern. We who share the English language have a world-wide responsibility towards other users of our radio jargon. At a time when the language of radio is, through its electronic off-shoots, encroaching on so many fields of human endeavor, it is worth while to take some pains to avoid ambiguity and confusion.
The Generation of Centimeter Waves*

H. D. HAGSTRUM†

Summary—The electronic devices used most extensively, recently, for the generation of centimeter waves are discussed. The physical form, operating capabilities, and the basic physical principles of operation of the triode, velocity-variation, and magnetron oscillators are presented. An attempt is made to show how these oscillators are related to one another. For a variety of reasons, particular emphasis is placed on the magnetron oscillator.

INTRODUCTION

THE DEVELOPMENT of radar during the war years has included a thoroughgoing exploitation of all the previously known means of generation of very-high-frequency electromagnetic radiation, and has witnessed the appearance of these generators in new and revolutionary forms, capable of performance unknown with their predecessors. The discussion in this paper will be restricted to those oscillators which have been used as generators in the centimeter-wave region, that is, in the wavelength region for which the centimeter is the convenient unit of wavelength. As representative of about the middle of the region in which the devices to be described are of most use, one may take 10 centimeters wavelength, corresponding to a frequency of 3000 megacycles.

There are three types of generators which have been used extensively in the centimeter-wave region for radar and upon which considerable effort has been expended during the war. These are the triode in a special form known as the disk-seal or "lighthouse" triode; the velocity-variation oscillator in both the double resonator or klystron, and the single-resonator or reflex-klystron forms; and the magnetron oscillator.

THE TRIODE OSCILLATOR

In any oscillator in which electrons are used as the agents by which energy is transferred from the primary direct-current source to the radio-frequency oscillation, electrons which have gained kinetic energy of motion from the direct field transfer a part of this energy to the radio-frequency field set up by the radio-frequency circuit. A net transfer of energy is achieved because some means of electron selection and rejection, sorting, or bunching is operative. The necessary criterion is either that more electrons interact favorably with the radio-frequency fields than interact unfavorably, or that

Since the only stray coupling between output and input cavities is that from anode to cathode through the grid, the phase between grid-plate and cathode-grid cavities may be adjusted to take into account reactive effects when the transit time becomes an appreciable fraction of the period of oscillation. Beyond this, the success of the disk-seal triode is in no small part to be accounted for by the ability of the structure to maintain small clearances between the electrodes and to permit duplication of these dimensions from tube to tube. In Fig. 2 are shown several types of disk-seal tubes developed during the war.

**Fig. 2**—A group of disk-seal tubes developed in recent years. Tubes of this type have been operated to better than 3000 megacycles and can generate several watts of centimeter-wave energy. Reproduction by courtesy of the General Electric Company.

**VELOCITY-VARIATION OSCILLATORS**

In the velocity-variation-type oscillator, the fundamental requirement for oscillation, namely, that more electrons give energy to the radio-frequency oscillation than take energy from it, is achieved by a mechanism of velocity variation and drift. A uniform beam of electrons, homogeneous in velocity, after passing through a radio-frequency field at one point at which the electron velocities are varied, is allowed to drift through a field-free region in which the beam forms itself into bunches whose frequency of arrival at a second point is that of the radio-frequency oscillation. The interaction of this bunched beam with a second radio-frequency field (it may be the same field traversed in the reverse direction), in such phase that the electrons in the bunches are decelerated by the field, achieves the desired energy transfer to the radio-frequency oscillation. The advantages of this mechanism over that operative in the triode oscillator are these: The electrons are accelerated by the radio-frequency field in a radio-frequency field-free region, making it possible to effect the interaction with the radio-frequency fields over short distances at full electron velocity. This makes possible a short transit time in the radio-frequency fields and thus more effective use of the electrons in the beam. Secondly, unwanted interaction between the radio-frequency field which varies the electron velocity and that which extracts energy from the electrons is effectively eliminated or sidestepped either by the separation of the two at some distance as in the double-resonator klystron, or by making them identical as in the reflex klystron. These features, together with the facts that the velocity-variation oscillator quite naturally makes use of the cavity-type resonator, and because the drift distances and electron velocities necessary are of convenient magnitude, make it ideally suitable for the generation of radio-frequency energy in the centimeter-wave region.

The reader is undoubtedly familiar with the double-resonator klystron and single-resonator or reflex-klystron types. Both have been discussed exhaustively in the literature. It will be of interest to compare the electronic mechanism, phase relationships, the Applegate diagram, and electronic tuning of these oscillators with the similar features of the magnetron oscillators. Two representative klystron oscillators are shown in Figs. 3 and 4.

**The Magnetron Oscillator**

**General Description**

The multicavity magnetron oscillator has three principal components: an electron interaction space with concentric cathode and anode, a multiple resonator

---

system, and an output circuit. Each of these is illustrated schematically in Fig. 5. In the electron interaction space between the cathode and the multisegment anode, electrons emitted from the cylindrical cathode move under the action of the radial direct electric field, the axial direct-current magnetic field, and the radio-frequency field set up by the resonator system between the anode segments. These electronic motions result in a net transfer of energy from the direct electric field to the radio-frequency field. The radio-frequency interaction field is the fringing electric field appearing between the anode segments. The radio-frequency energy, fed into the resonator system by the electrons, is delivered through the output circuit to the useful load. The output circuit shown in Fig. 5 consists of a loop inductively coupled to one of the hole and slot cavities, feeding a coaxial line.

To operate such a magnetron oscillator, one must place it in a magnetic field of suitable strength and apply a voltage of proper magnitude to its cathode, driving the cathode negative with respect to the anode. This voltage may be constant or pulsed. With suitable values of the operating parameters, the magnetron oscillates as a self-excited oscillator whenever the direct voltage is applied.

**Electron Motions in Electric and Magnetic Fields — The Direct-Current Magnetron**

Before beginning a discussion of the electronics of the magnetron oscillator, it would be well to review briefly electron motions in various types and combinations of electric and magnetic fields, and the operation of the direct-current magnetron.18

An electron, of charge \( e \) and mass \( m \), moving in an electric field of strength \( E \), is acted upon by a force, independent of the electron velocity, of strength \( eE \), directed oppositely to the conventional direction of the field. If the field is constant and uniform, the motion of the electron is identical to that of a body moving in a uniform gravitational field like that of the earth near its surface.

An electron moving in a magnetic field of strength \( B \), however, is acted upon by a force which depends on the magnitude of the electron velocity \( v \) on the strength of the field, and on how the direction of motion is oriented with respect to the direction of the field. The force is directed normal to the plane of the velocity and magnetic-field vectors and is of magnitude proportional to

\[
\frac{eBv}{c}
\]

18 A. W. Hull, "The effect of a uniform magnetic field on the motion of electrons between coaxial cylinders," *Phys. Rev.*, vol. 18, pp. 31-38; July, 1921. This was the first report on the direct-current cylindrical magnetron.
the velocity, the magnetic field, and the sine of the angle $\theta$ between them. Thus the force is the cross or vector product of $v$ and $B$:

$$ F = e[v \times B], \quad F = Bev \sin \theta. $$

An electron moving parallel to a magnetic field ($\sin \theta = 0$) feels no force. One moving perpendicular to a uniform magnetic field ($\sin \theta = 1$) is constrained to move in a circle by the magnetic force at right angles to its path. Since this force is balanced by the centrifugal force, the radius $\rho$ of the circular path depends on the electron momentum and the strength of the field, that is,

$$ Bev = \frac{mv^2}{\rho}, $$

yielding

$$ \rho = \frac{mv}{eB}. \quad (1) $$

The time $T_e$ required to traverse the circle is independent of the radius of the path and, hence, of the velocity of the electron; $T_e = 2\pi \rho / v = 2\pi m / eB$. Thus, the frequency of traversing the circular path, the so-called cyclotron frequency, depends on magnetic field alone and is given by:

$$ f_c = \frac{\omega_c}{2\pi} = \frac{1}{T_e} = \frac{1}{2\pi} \frac{e}{m} B. \quad (2) $$

In the magnetron, electronic motion in crossed electric and magnetic fields is involved. Consider first such motion between two parallel-plane electrodes, neglecting space charge. If, as in Fig. 6, the electric field is directed in the negative $y$ direction and the magnetic field in the negative $z$ direction, and if the electron starts from rest at the origin, the orbit is a cycloid given by the parametric equations:

\[
\begin{align*}
x &= v_c t - \rho_c \sin \omega_c t = \rho_c (\omega_c t - \sin \omega_c t), \\
y &= \rho_c (1 - \cos \omega_c t),
\end{align*}
\]  

in which:

$$ v_c = \frac{E}{B}, \quad (4) $$

$$ \rho_c = \frac{mv}{eB^2}, \quad (5) $$

$$ \omega_c = \frac{eB}{m}. \quad (6) $$

This motion may be regarded as a combination of rectilinear motion of velocity $v_c$ in the direction of the $x$ axis, perpendicular to both $E$ and $B$, and of motion in the $xy$ plane about a circular path of radius $\rho_c$ at a frequency $\omega_c / 2\pi$, the cyclotron frequency. Fig. 6 shows the resulting cycloidal path and its generation by a point on the periphery of the rolling circle. Even for cylindrical geometry, it is often convenient to think in terms of the plane case.

In the case of cylindrical geometry with radial electric and axial magnetic fields, the electron orbit, neglecting space charge, approximates an epicycloid generated by rolling a circle around on the cylindrical cathode. The orbit is not exactly an epicycloid because the radial motion is not simple harmonic, which state of affairs arises from the logarithmic variation of the direct-current electric field with radius. The approximation of the epicycloid to the actual path is a convenient one, however, because the radius of the rolling circle, its frequency of rotation, and the velocity of its center, for the epicycloid, all approximate those for the cycloid of the plane case. These approximations improve with increasing ratio of cathode to anode radii. Several electron orbits in a direct-current cylindrical magnetron are shown in Fig. 7 for several magnetic fields.

![Fig. 6](image)

**Fig. 6**—The cycloidal path of an electron which started from rest at the cathode in crossed electric and magnetic fields for the case of parallel-plane electrodes. The mechanism of generation of the orbit by a point on the periphery of a rolling circle is depicted.

![Fig. 7](image)

**Fig. 7**—Electron paths in a cylindrical direct-current magnetron at several magnetic fields above and below the cut-off value $B_c$. The electrons are assumed to be emitted from the cathode with zero initial velocity. Below these orbits is plotted the variation of current passed by a cylindrical direct-current magnetron at constant voltage as a function of magnetic field. The orbits of electrons at four different magnetic fields are shown above the corresponding regions of the current characteristic.

It is clear from this simplified picture of the orbits in a direct-current cylindrical magnetron without space charge that, at a given electric field, an electron orbit for a sufficiently strong magnetic field may miss the
anode completely and return to the cathode. The critical magnetic field at which this is just possible is called the cut-off value $B_c.$ A current-versus-magnetic-field curve in addition to the electron orbits corresponding to the slotted field is shown in Fig. 7. For the case of parallel-plane electrodes, the cut-off relation between the critical anode potential and magnetic field $V_c$ and $B_c$ and the electrode separation $d$ is obtained by equating the electrode separation to the diameter of the rolling circle. Thus,

$$d = 2 \frac{m (V_c)^{-1} e}{d B_c^2},$$

from which

$$V_c = \frac{e B_c^2 d^2}{2m}.$$

For the cylindrical case, the relation may be shown to be:

$$V_c = \frac{e B_c^2 r_s^2}{8m} \left[ 1 - \left( \frac{r_s}{r_a} \right)^{2/3} \right]$$

in terms of cathode and anode radii $r_s$ and $r_a.$

The Fundamental Electronic Mechanism of the Magnetron Oscillator

The direct-current magnetron may be converted into an oscillator suitable for the generation of centimeter waves, if it is arranged to introduce radio-frequency fields into the anode-cathode region. This is done in the case of the type of magnetron oscillator in most common use today has been seen in the discussion of Fig. 5.

The electrons in the interaction space of the magnetron oscillator are the agents which transfer energy from the direct field to the radio-frequency field. As such, they must move subject to the constraints imposed by the direct radial electric and direct axial magnetic fields, considering, for the moment, the radio-frequency fields to be small. Under these conditions, as has been seen for the direct-current cylindrical magnetron (Fig. 7 for $B > B_c$), electrons follow approximately epicycloidal path which progress around the cathode. The mean velocity of this progression, that of the center of the rolling circle, depends upon the relative strengths of the electric and magnetic fields (see (4) for the plane case). By proper choice of direct voltage $V$ between cathode and anode and of magnetic field $B$, the mean angular velocity of the electrons may be set at any desired value.

The radio-frequency electric fields in the interaction space, with which the electrons moving as described above must interact, are the electric fields fringing from the slots in the anode surface. These fields are provided by the $N$ coupled oscillating cavities of which the magnetron-resonator system is composed. Such a system of resonators may oscillate in a number of different modes. At this point, however, only that mode in which the magnetron oscillator is generally operated, the $\pi$ mode, will be considered. This is the mode for which the oscillations in adjacent resonators are $\pi$ radians out of phase and for which the potential variation around the magnetron interaction space is a standing wave like that plotted in Fig. 8.

---

Fig. 8—A plot showing the $\pi$-mode anode potential wave at several instants in an eight-resonator magnetron and the mean paths of electrons which interact favorably with the radio-frequency field. The plot is developed from the cylindrical case, the shaded rectangles at the top representing the anode segments.

As in any oscillator, oscillation in the magnetron is possible only if more energy is transferred to the radio-frequency field by electrons driven against it than is taken from the radio-frequency field by electrons accelerated by it. This can be accomplished only if the mean angular velocity of the electrons is such as to make them pass successive gaps in the anode at nearly the same phase in the cycle of the radio-frequency field across the gaps. Then it is possible for an electron, which leaves the cathode in such phase as to oppose the tangential component of the radio-frequency field across one anode gap, to continue to lose energy gained from the direct field to the radio-frequency field at successive gaps. Electrons which gain energy from the radio-frequency field are driven back into the cathode after only one orbital loop and are removed from further motion detrimental to the oscillation. This process of selection and rejection forms groups or bunches of electrons which sweep past the anode slots in phase to be retarded by the radio-frequency field component. The criterion that the electron drift velocity shall be such as to keep these bunches in proper phase is analogous to the condition that the drift angle in a velocity-variation oscillator be such as to cause the electron bunches to cross the gap of the second or catcher cavity in phase to lose energy to the radio-frequency field across the gap.

The condition placed upon the mean angular velocity of the electrons may be discussed more readily by reference to Fig. 8. Focus attention on an electron which
crosses the gap between anode segments 1 and 2 at the instant \( t \) when the radio-frequency field is maximum retarding, that is, the potential on segment 1 is maximum and on segment 2 minimum. It is clear that this electron can cross the next gap in the same phase if the time required to reach it is \( (|p| + 1/2) T \), in which \( p \) is any integer and \( T \) is the period of radio-frequency oscillation. In Fig. 8 four lines are drawn representing the mean paths of electrons moving with such velocities as to make \( p = 0, 1, 2, \) and 3. Each line crosses a gap when the radio-frequency field is maximum retarding, that is, when the potential has the maximum negative slope at the center of the gap. As will be seen later, a more convenient parameter, to be called \( k \), is that whose absolute magnitude \( |k| \) specifies the number of radio-frequency cycles required for the electron to move once around the interaction space. \( |k|/N \) is then the number of cycles between crossings of successive anode gaps, which for the \( \pi \) mode of Fig. 8 must take on the values:

\[
\frac{|k|}{N} = |p| + 1/2, \quad p = 0, \pm 1, \pm 2, \ldots
\]

or the values given by the more general expression, applicable to any mode:

\[
\frac{|k|}{N} = |p| + \frac{n}{N}, \quad p = 0, \pm 1, \pm 2, \ldots
\]

In this expression, \( n/N \) is the phase difference between adjacent resonators expressed as the fraction of a cycle; \( k \) may thus assume the values given by:

\[
k = n + pN, \\
p = 0, \pm 1, \pm 2, \ldots
\]

The mean angular velocity which the electrons must possess is then given by:

\[
\frac{d\theta}{dt} = \frac{2\pi}{kT} = \frac{2\pi f}{k}
\]

For the \( \pi \) mode \( (n = N/2) \) it is seen that the negative integers \( p \) give the same series of values for \( |k| \) as do the positive integers including zero. The sequence is \( |k| = 4, 12, 20, 28, \ldots \). Reference to Fig. 8 indicates that electrons may travel in either direction around the interaction space and interact favorably with the radio-frequency field, provided their mean angular velocity is given by (9) with values of \( k \) specified by (8). That this should be so is clear from the fact that the anode potential wave is a standing wave with respect to which direction has no meaning. Fig. 8 also makes clear how an electron moving with velocity different from that corresponding to the lines shown will fall out of step with the field and, on the average, be accelerated as much as it is retarded, thus effecting no net energy transfer.

The actual electron orbits do not correspond to simple translation but, as has been discussed, to rotation superposed on translation. However, the epicycloid-like scallops in the orbit are of no significance to the fundamental electronic mechanism. It is the mean velocity of the electron motion around the interaction space, specified by the relative values of \( V \) and \( B \), that is of importance.

The similarities between the electronic mechanism of the magnetron oscillator and those of the triode and velocity-variation oscillators may now be seen. In Fig. 9 an attempt has been made to depict schematically the parallelisms between these types of oscillators and a simplified equivalent lumped-constant circuit. In the magnetron, bunches or groups of electrons are formed by the interaction of the electrons and the radio-frequency field. These spokes in the space-charge cloud sweep past the anode gaps in phase to give up energy, gained from the direct field, to the radio-frequency fields across the gaps. The “bunching” field in the magnetron is thus the same field as that to which energy is transferred. In this sense the magnetron is analogous to the reflex klystron in which a single cavity is used as both buncher and catcher. How the bunches of electrons are formed in the magnetron interaction space will be discussed in greater detail when the traveling-wave picture of the electronic mechanism is presented.
Other Types of Magnetron Oscillators

The multicavity magnetron oscillator discussed above is one of three types of magnetron oscillators which may be distinguished by the nature of the electronic mechanism by means of which energy is transferred to the radio-frequency field. Oscillation of the so-called negative-resistance magnetron oscillator depends upon the existence of a static negative-resistance characteristic between the two halves of a split anode. The so-called cyclotron-frequency magnetron oscillator operates by virtue of resonance between the period of radio-frequency oscillation and the period of the cycloidal motion of the electrons (rolling-circle frequency which in plane geometry equals the cyclotron frequency). The so-called traveling-wave magnetron oscillator depends in its operation upon resonance, that is, approximate equality, between the mean translational velocity of the electrons and the velocity of a traveling-wave component of the radio-frequency interaction field.

The Negative-Resistance Magnetron Oscillator—Type I

In the negative-resistance magnetron oscillator the anode is split parallel to the axis into two halves between which the radio-frequency circuit is attached. The transit time from cathode to anode is not involved in the mechanism, except that it must be small relative to the period of the radio-frequency oscillation. The static negative-resistance characteristic arises from the fact that under certain circumstances the allowable orbits for the majority of electrons terminate on the segment of lower potential, irrespective of the segment toward which they start. These electrons, being driven against the radio-frequency component of the field, give energy gained in the direct field to the radio-frequency field.

The Cyclotron-Frequency Magnetron Oscillator—Type II

Not long after the invention of the direct-current magnetron, oscillations between anode and cathode were found to occur near the cut-off value of magnetic field. Later it was shown that the oscillation period is equal to the electron transit time from the vicinity of the cathode to the vicinity of the anode and back.

The electronic mechanism must be explained in terms of electrons moving in the direct radial electric and axial magnetic fields and the superposed radial radio-frequency electric field. This may be done as follows: An electron leaving the cathode in such phase as to gain energy when moving from the cathode toward the anode will also gain energy during its return, striking the cathode with more energy than it had when it left. There, such an electron is stopped from further motion during which it would continue to absorb energy from the radio-frequency field at the expense of the oscillation. The electron orbit is shown in Fig. 10. An electron leaving the cathode in the opposite phase, on the other hand, loses energy when moving toward the anode and again on its return toward the cathode. As is shown in Fig. 11, it reverses its direction after the first trip without reaching the cathode surface and starts over on a second loop of smaller amplitude, remaining in the same phase and continuing to lose energy to the field. This process continues until all the energy of the rotational component of the electron motion has been absorbed by the radio-frequency field. If the electron is not removed at this stage, in its subsequent motion the rotational component will build up, extracting energy from the radio-frequency oscillation. Means such as tilting the magnetic field or placing electrodes at the ends of the
Multicavity magnetron oscillators, developed from the British prototype, are now available at wavelengths ranging from approximately 0.5 to 50 centimeters. Both pulsed and continuous-wave generators of this type have been made. The upper limit of peak power is now about 100 kilowatts at 1 centimeter, 3 or 4 megawatts at 10 centimeters. Operating voltages may be less than 1 kilovolt or more than 40 kilovolts. The direct-current magnetic fields essential to operation range from 600 to 15,000 gauss.

The Interaction Field and the Modes of Oscillation of the Resonator System

The electronic mechanism of the traveling-wave-type magnetron oscillator like that of Fig. 5, oscillating in its so-called π mode, has already been discussed in terms of electron motions through the radio-frequency fields at the gaps in the multisegment anode. To extend the discussion to other points of view and for other modes of oscillation of the magnetron resonator system, it is necessary to treat in more detail the interaction field and its relation to the modes of oscillation.

The cylindrical magnetron anode structure is a series of $N$ resonators connected in a ring. The oscillation in each resonator of this array of coupled resonators is specified by a differential equation in terms of a parameter, such as current or voltage, the constants of the circuit itself, and the mutual interaction between the circuit and its neighbors. Each solution of the set of simultaneous differential equations for all the resonators involved corresponds to a definite phase shift between adjacent resonators. The allowed values of this phase shift depend upon the boundary conditions imposed by the connecting together of the resonators into a ring. Under these circumstances only those modes of oscillation are possible for which the total phase shift around the ring is $2\pi n$ radians, $n$ being any integer including zero. The oscillations in adjacent cavities then differ in phase by $2\pi n/N$ radians. This means that only those waves traveling around the anode block which constructively interfere are possible solutions. These are waves which, after leaving an assumed starting point and traversing the anode once, arrive back in phase with the wave, then leave in the same direction.

Each mode of oscillation of the multiresonator system has a resonant frequency different from the frequency of any other mode and from the frequency of one of the $N$ resonators oscillating freely and uncoupled from its neighbors. In the general case of $N$ coupled resonators, as in the case of two coupled resonators, the modes of oscillation have different resonant frequencies.
because of the effect of the mutual coupling between the resonators.

The interaction fields for the several modes of oscillation of the resonator system are thus to be distinguished by the number \(n\) of repeats of the field pattern around the interaction space. Since the potential at the anode radius is nearly constant across the faces of the anode segments and varies primarily across the slots, the azimuthal variation of the field cannot be purely sinusoidal but must involve higher-order harmonics. For a mode of frequency \(f = \omega / 2\pi\), corresponding to a phase difference between adjacent resonators of \(2\pi n/N\), the anode potential wave is of periodicity \(n\) around the anode, and may be written as a Fourier series of sinusoidal component waves traveling in opposite directions around the interaction space:

\[
V_{r-f} = \sum_{k} A_k e^{i(\omega t - k\theta + \gamma)} + \sum_{k} B_k e^{i(\omega t + k\theta + \delta)}
\]

where \(k = n + pN\), \(p = 0, \pm 1, \pm 2, \ldots\).

Note that the sums are taken over all integral values of \(k\) given by (8).

The interaction field for any mode of periodicity \(n\) is thus represented by two oppositely traveling waves, whose fundamentals are moving with angular velocities \(\omega/n = 2\pi f/n\), and whose component amplitudes \(A_k\) and \(B_k\) in general are not equal. \(\gamma\) and \(\delta\) are arbitrary phase constants.

The expression (10) may be reduced to the form:

\[
V_{r-f} = \sum_{k} (A_k - B_k) \cos (\omega t - k\theta + \gamma)
\]  

\[
+ \sum_{k} 2B_k \cos \left(\frac{\omega t + \gamma + \delta}{2}\right) \cos \left(\frac{k\theta - \gamma - \delta}{2}\right)
\]

which shows that the complete field pattern may be considered to consist of a rotating wave superposed on a standing wave, each having a fundamental component of periodicity \(n\).

The fact that the periodicities \(k\) of the harmonics in (10) or (11) are those for which \(k\) has the values given by (8) may be determined from a Fourier analysis of the complete anode potential waves like that of Fig. 8.

The terms in (10) and (11) for which \(|k| = n\) are the fundamental components; those for which \(|k| \neq n\) are called the Hartree harmonics. Any sinusoidal component for which the number of complete cycles around the anode is greater than \(N/2\) is thus a harmonic of the complete field pattern for one of the modes whose fundamental is of periodicity \(n\).

Physically distinguishable modes of oscillation exist only for the values of \(n\) less than or equal to \(N/2\) including zero. However, this accounts for only \(N/2 + 1\) of the \(N\) modes of oscillation which one expects a system of \(N\) resonators to possess, because in general the frequency of a mode specified by the parameter \(N\) (except for the value 0 and \(N/2\)) is a double root for a perfectly symmetrical anode structure. The mode is thus a doublet and is said to be degenerate. One would expect this on mathematical grounds from the fact that the general solution in (10) has four arbitrary constants, whereas a singlet solution of the system of second-order differential equations specifying the oscillations should have no more than two.

This degeneracy of the modes of the resonator system may be removed if the symmetry of the system is destroyed by the presence of a disturbance or perturbation at one point (a coupling loop in one of the cavities, for example) which provides the necessary additional boundary condition. Removal of the degeneracy makes possible only standing waves as complete solutions of the simultaneous differential equations specifying the oscillation. Thus, in (11) \(A_k = B_k\) and the first summation representing a rotating wave vanishes. The second summation may be broken up into two patterns: one, cosine-like with respect to the asymmetry as origin, whose frequency is altered from the degenerate value; and a second, sine-like with respect to the asymmetry as origin, whose frequency is the same as the degenerate value. This situation prevails for \(n = 1, 2, \ldots, N/2 - 1\), contributing \(N - 2\) modes. The remaining two modes of the resonator system, for which \(n = 0\) and \(N/2\), are singlet modes even in the symmetrical anode. For the \(n = 0\) mode, whose field pattern is independent of angle, the component whose frequency is undeviated from that of the degenerate pair corresponds to the trivial case of zero amplitude at all points. Similarly, for the \(n = N/2\) mode (the \(\pi\) mode), the cosine-like pattern gives zero potential at each anode segment, an equally trivial case. Thus each of the \(N\) modes of the multicavity resonator system has been accounted for.

As an example, plots of the field configurations for the modes of a magnetron having eight resonators are shown in Fig. 12. For clarity, only the electric field lines of the fundamental component (\(p = 0\)) of each mode are shown in the interaction space. Only the magnetic field lines are shown in the resonators. Below these is plotted the distribution in potential for each of the fundamentals, \(\sin n\theta\) and \(\cos n\theta\), \(n = 0, 1, 2, 3, 4\). For the \(n = 0\) mode the magnetic flux threads through all the resonators in the same direction and returns through the interaction space. That all the segments are in phase and the interaction space field is independent of angle may be seen. That there is but one \(\pi\) mode is also seen from the fact that the \(\cos 4\theta\) term corresponds to zero potential on all the anode segments. The first Hartree harmonic for the \(n = 1\) mode, namely, that for which \(p = -1\) (\(\sin 7\theta\) plotted instead of \(\sin -7\theta\)), having seven repeats (\(k = 7\)) or a total phase shift of \(14\pi\) radians around the anode, is also plotted in Fig. 12 in addition to the fundamental. The fact that it yields the same variation of anode-segment potential around the anode as the fundamental is apparent.

In recapitulation, one may say that for each value of \(n\) the total radio-frequency interaction field pattern is
generally composed of a rotating wave superposed on a standing wave. If the degeneracy of the mode is removed, only the standing wave remains. The electronic interaction with standing total wave on the anode for the π mode has already been discussed. This magnetron oscillator is called the traveling-wave type because its electronic interaction mechanism may be discussed in terms of electronic interaction with one of the sinusoidal traveling-wave components of periodicity k given in (10).

Electrons may interact favorably with the interaction fields of modes other than the π mode, resulting in a net transfer of energy from the direct to the radio-frequency fields. Plots similar to that of Fig. 8 may be drawn for these cases. There interactions, of interest in understanding the magnetron oscillator, have not been used much in practice and will not be discussed further here.

**The Traveling-Wave Picture of the Electronic Mechanism**

For each value of k in (10), whether or not \( A_k = B_k \), there are two oppositely traveling sinusoidal wave components of periodicity k. Since each such component requires k cycles of the radio-frequency oscillation to complete one trip around the interaction space, its linear velocity at the anode surface is \( 2\pi f_r / k \), corresponding to an angular velocity of \( 2\pi f / k \). The electronic mechanism of the traveling-wave or type-III magnetron oscillator may be discussed in terms of electron
interaction with these sinusoidal traveling-wave components present in the interaction field. This might at first appear to be difficult, in view of the many components of several possible modes. By mode-frequency separation, the means of which are mentioned later, it is generally possible to restrict oscillation to only one mode, usually the π mode. Further, the fact that the electronic motion in crossed direct electric and magnetic fields results in a mean drift of electrons around the interaction space enables one to restrict his attention to a single traveling wave corresponding to the fundamental or a single Hartree harmonic of the field of this mode; for it is possible, in principle at least, by proper adjustment of $V$ and $B$ to equate the mean angular velocity of the electrons to the angular velocity, $2\pi f/|k|$, of any one of the traveling-field components. When this is true, only the field of this component has an appreciable effect upon the electron motion. With respect to the fields of the oppositely traveling component of the same harmonic (same $k$), and the components of all other harmonics (different $k$), the electron finds itself drifting rapidly through regions of accelerating and decelerating field with no net energy transfer. From the point of view of the electron, the fields of the other components vary so rapidly as to average out over any appreciable interval of time. The only exception to these statements occurs when a harmonic of periodicity $k'$ of another mode of frequency $f'$ has the same angular velocity as the harmonic of periodicity $k$, that is, when $2\pi f'/|k'| = 2\pi f/|k|$. Should this occur, the magnetron may have a tendency to "mode," that is, to operate either steadily or intermittently in a mode other than the π mode. In the calculation of electron motions, the restriction to the field of a single traveling-wave component has been called the "rotating-field approximation."

The consideration of the electronic mechanism has thus been reduced to that of the motion of electrons under the combined influence of the radial direct electric field, the axial direct magnetic field, and a sinusoidal field wave traveling around the interaction space. From what has been said thus far it is clear that for energy to be transferred to the radio-frequency field it is necessary that the mean electron velocity very nearly equal that of the traveling wave. Then an electron, leaving the cathode in such phase as to find itself moving in a region of decelerating tangential component of the radio-frequency field, may continue to move with this region and lose energy to the field. In contrast to the type-II transit-time magnetron oscillator, the energy transferred to the radio-frequency field in this case is the potential energy of the electron in the radial direct electric field. The energy in the rotational component of the motion remains practically unaffected and the electron orbit from cathode to anode looks something like that plotted in Fig. 13, for the case with plane electrodes. On the other hand, an electron which leaves the cathode in such phase as to gain energy in a region of accelerating tangential radio-frequency field is removed at the cathode after only one cycle of the epicycloid-like motion. If this did not occur, the electron would continue to move with the field and absorb energy. Its orbit is shown in Fig. 14. It is instructive to compare the orbits of the two categories of electrons in the traveling-wave magnetron oscillator with the orbits of corresponding electrons in the cyclotron-frequency type of magnetron (Figs. 10 and 11). In each case, it is the fact that "favorable" electrons may interact for a considerably longer time than "unfavorable" electrons which makes possible a net energy transfer between the direct and radio-frequency fields.

![Fig. 14](image1)

**Fig. 14** — The orbit of an electron which gains energy from the radio-frequency field in a traveling-wave or type-III magnetron oscillator. The direct-electric force on the electron is directed from cathode to anode.

One may now compare the traveling-wave picture of the electronic mechanism with that presented earlier in which the motion of electrons past the gaps in the anode structure is considered. An electron, moving so that $|k|/N = |p| + (n/N)$ cycles of the radio-frequency oscillation elapse between its crossing of two successive anode gaps, is thus moving around the interaction space in synchronism with a traveling component of the $k$th harmonic of the interaction field. Both points of view are of value. That involving the motion of electrons past the anode gaps is more fundamental, physically. That in terms of a traveling-wave component, on the other hand, is more convenient in calculations of electron orbits including space-charge effects where, by transformation to a co-ordinate system rotating with the field, it is possible to deal with motions in static fields.

**Phase Focusing**

It has been seen from two points of view how groups of electrons which move around the interaction space of the magnetron oscillator are formed by a process of selection and rejection of electrons by the tangential component of the radio-frequency field. However, space-
charge debunching and the discrepancy at all but one radius between the mean velocity of translation of the electrons and the velocity of the interaction field would tend to disperse these groups and prevent efficient interaction, were it not for the phase focusing provided by the radial component of the radio-frequency field.

The mechanism of the phase focusing may be discussed either in terms of the interaction of electrons with the actual fields existing at the anode gaps or in terms of the traveling-wave picture of the electronic mechanism. The fundamental mechanism involved depends upon the effect of the radial component of the radio-frequency field in aiding or opposing the radial direct field. If the radial radio-frequency field increases the net radial field in which the electron finds itself at any instant, the mean velocity of the electron increases, as can be seen from (4) for the plane case. Similarly, a decrease in the net radial electric field, caused by the radio-frequency radial component, results in decreased electron translation velocity.

Consider an electron which crosses an anode gap at the instant the radio-frequency field there is maximum retarding, that is, an electron which is to be found on the plane marked $M$ in Fig. 15 at this instant. It experiences about as great an increase of velocity by virtue of the radial component aiding the direct radial field before crossing the gap as decrease by virtue of the radial component opposing the direct radial field after crossing the gap. Another electron which is lagging behind the electron just considered is to be found opposite a positively charged anode segment, as at $P$ in Fig. 15, when the radio-frequency field passes through its maximum value. Since the radio-frequency field component decreases with time after this instant, the effect of the radial component of the field on the electron velocity after crossing the gap will be less than its effect before crossing the gap, the net effect being one of increasing the mean velocity of translation, bringing the electron more nearly into the proper phase. An electron which leads the electron first considered, on the other hand, will be found opposite the negatively charged anode segment beyond the gap when the radio-frequency field is maximum, and for it the net effect of the radial component is to reduce the mean velocity of the electron, bringing it also more nearly into the proper phase.

In discussing the mechanism of phase focusing from the traveling-wave point of view, the field lines of Fig. 15 may be considered to be those of the traveling-wave component with which the electrons are interacting. Then the whole field pattern indicates moves to the right, as shown by the arrow above the plane of maximum retarding tangential field at $M$. An electron which falls behind the position $M$ to the point $P$, for example, finds itself in a stronger net radial electric field which increases its mean translational velocity, tending to bring it back to the position $M$. The reverse holds for an electron which runs ahead of the plane $M$.

**Space-Charge Configuration**

The over-all picture of the electronic mechanism in the type-III magnetron oscillator thus presents a spoke-shaped space-charge cloud of electrons wheeling around the cathode in synchronism with the anode potential wave, each spoke in a region of maximum retarding field. This picture of what is happening has been very handsomely confirmed by actual orbital calculations taking account of space charge. The result of one such calculation is shown in Fig. 16. The orbits of four electrons which were emitted from the cathode in different phases in one repeat of the anode radio-frequency field are plotted in a set of co-ordinates rotating with the

Fig. 15—A plot of lines of electric force on an electron (drawn for the plane case) for the fundamental of the $\pi$ mode. It is shown for the purpose of explaining the phase focusing property of the radial field component. The plane of maximum opposing force on the electron intersects that of the figure along the line $M$. The force on the electron due to the direct electric field is directed from cathode to anode.

Fig. 16—The orbits of four electrons which left the cathode in different phases in one period of the radio-frequency field, plotted in a co-ordinate system rotating with the anode potential wave. The dashed lines enclose the orbits of the electrons, and hence delineate the boundaries of the space-charge cloud which rotates around the cathode in synchronism with the anode-potential wave. Planes of maximum retarding tangential field are represented by the lines $M$ (see Fig. 15). This figure is reproduced by courtesy of the British Committee on Valve Development (CVD) and is taken from the CVD Magnetron Report No. 41.
Radio-frequency field component. One electron is returned to the cathode, and the other three reach the anode. The spoke-shaped structure is clear, and its position with respect to the rotating anode potential wave is as expected. The number of spokes of the cloud is equal to the order of the component of the mode with which the electrons are interacting. In the case of Fig. 16 there are four spokes, since the magnetron is operating in the fundamental of the \( n = 4 \) mode \((k = 4, p = 0)\).

**Induction by the Space-Charge Cloud**

Another view of the mechanism by which the electrons drive the resonator system may be obtained by considering the effect of the space-charge spokes in inducing current flow in the anode segments themselves. For example, the oscillation of the resonator block in its \( \pi \) mode corresponds to the periodic interchange of electric charge from each anode segment around a resonating cavity to the next anode segment. This oscillation is maintained, much in the manner of a pendulum escape ment drive, by the spoke of negative space charge appearing in front of an anode segment at that instant in the oscillation cycle when it can aid in building up the net positive charge on the segment. At the same instant the adjacent segments, being opposite a “gap” in the space-charge wheel, may build up a negative charge.

The radio-frequency current \( I_{rf} \), induced in the anode structure, thus results from the motion of the spoke-shaped space-charge cloud in the interaction space. It is not to be confused with the total circulating radio-frequency current in the resonator system. Whereas \( I_{rf} \) must be in phase with the space-charge cloud, it need not be in phase with the radio-frequency voltage \( V_{rf} \) between the anode segments. In terms of the electron motions, this means that the spokes of the space-charge cloud may lead or lag the maxima in the tangential field. In general, the electronic admittance defined by the ratio of \( I_{rf} \) to \( V_{rf} \) may thus include a susceptance as well as a conductance. The product of \( V_{rf} \) and the in-phase component of \( I_{rf} \), integrated over a period of one cycle of radio-frequency oscillation, equals the energy per cycle which is delivered to the load. This amount of energy is twice that transferred in the half cycle during which the spokes of space charge move against the field from positions in front of one set of alternate anode segments to similar positions in front of the adjacent anode segments.

In each spoke of the electron space-charge cloud, individual electrons progress from cathode to anode. The direct current \( I \) passed by the magnetron is made up of electrons which strike the anode from the ends of the space-charge spokes. If the magnetron is driven at greater direct current, the space charge in the interaction space increases but the phase of its structure with respect to the traveling anode wave does not change to a first approximation. Thus both the in-phase and quadrature components of \( I_{rf} \) increase with no change in electronic admittance. One of the second-order effects which arise from small shifts in the phase of the rotating space-charge structure is the shift at constant load of operating frequency with direct current passed by the magnetron. This shift is called the frequency “pushing” and is measured in megacycles per second per ampere.

**Necessary Conditions for Oscillation**

After having discussed the electron motions in the interaction space of the type-III magnetron oscillator, the viewpoint will now be changed to that looking from the outside in, so to speak, and it will be asked what conditions relating measurable parameters are imposed by the nature of the electronic mechanism. Beyond the geometrical parameters of cathode and anode radii \( r_e \) and \( r_a \), one can determine the direct voltage \( V \) applied between cathode and anode; the magnetic field \( B \) in which the magnetron is placed; the direct current \( I \) drawn by the anode; the frequency of oscillation \( f \); and, from impedance measurements, the radio-frequency load presented to the electrons by the resonator, output, and load.

Perhaps the most fundamental condition for oscillation of the traveling-wave magnetron is that imposed by the requirement of synchronism between the electron drift and the radio-frequency field. As has been indicated, the angular velocity of a rotating component of a Hartree harmonic of the interaction field of order \( k \) is \( 2\pi f / |k| \). An approximate expression for the mean angular velocity of the electrons may be determined by neglecting the variation of electric field with radius and calculating the angular velocity midway between cathode and anode, thus:

\[
\frac{V}{(r_a + r_e)/2} \approx \frac{E / B}{(r_a + r_e)/2} = \frac{V / (r_a - r_e)B}{(r_a + r_e)/2} = \frac{2V}{(r_a - r_e)B}
\]

Equating this to the angular velocity \( 2\pi f / |k| \), one obtains the relation

\[
V = \frac{\pi f}{|k|} r_a^2 \left[ 1 - \left( \frac{r_a}{r_e} \right)^2 \right]. \quad (12)
\]

In this derivation it should be recognized that the velocity \( 2\pi f / |k| \) may be considered either to be the velocity of traveling component of the radio-frequency field with which the electron interacts or the mean velocity which the electron must have to maintain proper phase with the total radio-frequency fields existing across the anode gaps.

Posthumousref derived an expression, assuming negligible cathode diameter, which is similar to (12). By the same method as that used above, Slater has derived an expression differing from (12) by a term which results from the use of a more accurate value for the electron translational velocity at the midpoint between cathode and anode in cylindrical geometry. Slater’s expression is

\[
V = \frac{\pi f}{|k|} r_a^2 \left[ 1 - \left( \frac{r_a}{r_e} \right)^2 \right] - 2 \frac{m}{e} \left( \frac{\pi f r_a}{|k|} \right)^2 \left[ 1 - \left( \frac{r_a}{r_e} \right)^2 \right]. \quad (13)
\]
Hartree has derived an expression from a consideration of the conditions under which electrons are just able to reach the anode with infinitesimal amplitude of radio-frequency voltage in the \( k \)th harmonic. It is:

\[
V = \frac{\pi f}{k} r_s B \left[ 1 - \left( \frac{r_e}{r_o} \right)^2 \right] - 2 \frac{m}{e} \left( \frac{\pi f r_e}{k} \right)^2. \tag{14}
\]

In a sense this condition represents a cut-off relation for the oscillating magnetron analogous to Hull’s cut-off relation for the direct-current magnetron [see (7)].

Plotted on a \( V-B \) graph, (12), (13), and (14) represent parallel straight lines. The line of (12) passes through the origin; the so-called Hartree line of (14) is tangent to the direct-current cut-off parabola; the so-called Slater line of (13) lies above the Hartree line but below the line of (12). Each of the above expressions indicates that the electrons will drive a given harmonic of the radio-frequency interaction field in a type-III magnetron oscillator only at values of direct voltage and magnetic field which bear a definite relation. This relation expresses the fact that \( V/B \) is very nearly constant [see (12)].

In Fig. 17 are plotted as an illustration the Hartree lines for the fundamentals \((\phi = 0)\) of the \( n = 1, 2, 3, \) and 4 modes and for the \( k = -5 \) harmonic \((\phi = -1)\) of the above have been of great value in the identification of the modes of operating magnetrons and as the starting point in the design of new magnetrons for given wavelengths, magnetic field, and voltage.

### The Performance Chart

Another fundamental performance characteristic of the operating magnetron is the \( V-I \) plot or performance chart. In Fig. 18 such a chart is plotted for the same magnetron used as the example for Fig. 17. In it are plotted contours of constant magnetic field, radio-frequency power output, and over-all efficiency. The fact that the constant magnetic-field contours are nearly horizontal and spaced as they are is a manifestation of the oscillation conditions of (13) and (14). The increase of voltage with current is an effect attributable to the space charge quite independent of the condition of synchronism between field and electrons, for if the magnetron is to deliver more power at a given magnetic field, the induced radio-frequency current must increase. This entails increased space charge and a greater direct-current flow. To maintain the increased space charge additional direct voltage is required.

### The Electronic Efficiency

The performance chart also shows the not too surprising fact that more power may be drawn from the magnetron as the voltage and current are increased. More interesting are the increases of the over-all efficiency with voltage and the maximum through which the efficiency passes with increasing current. This variation of over-all efficiency \( \eta \) is to be attributed to changes in the electronic efficiency \( \eta_e \) since the other factor involved in the over-all efficiency, the circuit efficiency \( \eta_c \), is essentially constant over the diagram \((\eta = \eta_e \eta_c)\).
The increase of electronic efficiency with voltage, and hence magnetic field, may be explained by the picture of electron motions in the interaction space. The highest electronic efficiency is attained when the electrons reaching the anode do so with least kinetic energy. The energy lost at the anode per electron is that gained as kinetic energy beyond the last cusp of the orbit. By bringing the last cusp closer to the anode, corresponding to a reduction of the amplitude of the rotational component of the electron motion, the fractional energy lost at the anode may be reduced. Thus, according to (5), for the radius of the rolling circle this energy loss should vary as \( V/B^2 \) or, since \( V/B \) is approximately constant, as \( 1/B, \eta_a \) increasing with \( B \). In terms of electron orbits for a plane magnetron, Fig. 19 shows how increase in voltage and magnetic field increases the electronic efficiency. The dependence of electronic efficiency on \( B \) predicted by this simple picture is in accord with the dependence predicted by more sophisticated theories.

**The Radio-Frequency Circuit of the Magnetron Oscillator**

Although the radio-frequency circuit of the magnetron oscillator must of necessity incorporate special features, its characteristics may be studied and specified in ways similar to those applied to other types of radio-frequency circuits in the centimeter-wave region. Thus, from suitable measurements on the nonoscillating magnetron, one may identify the modes of oscillation of the resonator system, which is connected to its load through a given output circuit, and, for each, specify a natural frequency of resonance, unloaded, loaded, and external \( Q \)'s, as well as a characteristic admittance of the system.

**The Resonator System**

In Fig. 20 is shown a series of resonator blocks for pulsed magnetron oscillators ranging in frequency from 700 to 24,000 megacycles. The rings or wires connected to the anode segments are the so-called straps to be discussed later. The smallest resonator system of Fig. 20 is a so-called "rising-sun" system having "vane"-type resonators of alternate size. It, also, is mentioned briefly later.

**Separation of Mode Frequencies**

The frequencies of several of the modes of oscillation possessed by the multicavity magnetron-resonator system would ordinarily be quite closely grouped near that of the \( \pi \) mode, were not steps taken to separate them. From the point of view of the electronics of the magnetron one might think such proximity of mode frequencies to be no problem, because the different modes, even if of the same frequency, generally require different conditions relating the operating parameters \( V \) and \( B \) for oscillation [see (14)]. From the circuit point of view, however, close proximity of the mode frequencies is clearly undesirable, for under such conditions it is possible that the electronically driven mode, usually the \( \pi \) mode, may excite oscillation in a second mode. The \( \pi \)-mode oscillation, coupled to the second mode through some symmetry in the resonator system, sets up forced oscillations in the second mode under these conditions. The interaction field pattern of the second mode then appears as a contamination of the \( \pi \)-mode pattern, adversely affecting the electronic interaction with the \( \pi \) mode.

Mode-frequency separation in magnetron-resonator systems has been accomplished by two methods. In one, conductive connections between the anode segments, called straps, are employed. In the other, resonant cavities of two sizes spaced alternately are employed in an unstrapped resonator system.

![Fig. 19—Orbits of electrons which transfer energy to the radio-frequency field, plotted for operation of the magnetron at two different magnetic fields.](image)

![Fig. 20—A series of resonator blocks for magnetron oscillators.](image)
in the strapping system. The original British strapping is not symmetrical around the anode. Other types are symmetrical, except for breaks which are usually incorporated at least on one end of the anode. These asymmetries in the strapping provide the most convenient method of incorporating the additional asymmetry needed in the resonator system to orient the standing-wave patterns of the doublet modes with respect to the output circuit of the magnetron so as to equalize their loading. In addition, the strap asymmetries are arranged so as not to affect the symmetrical distributions of voltage and current in the resonator system for the π mode, but to destroy such symmetry to an appreciable extent for other modes.

The "Rising-Sun" Resonator System

The second type of magnetron resonator system in which the mode frequencies may be separated sufficiently well to allow "clean" operation in the π mode is an unstrapped structure involving the use of resonant cavities of two sizes so arranged that adjacent cavities are alternately large and small. This resonator system, called the "rising-sun" system, accomplishes mode-frequency separation by means analogous to the increase in separation of the mode frequencies of a system of two coupled resonators achieved by relative detuning of the individual resonators. Such a resonator system was evolved at the Columbia Radiation Laboratory during a series of experiments with asymmetries in an unstrapped resonator system. It is particularly adaptable to use in magnetrons of short wavelength where straps become very small and extremely difficult to construct. A "rising-sun" resonator system for \( N = 18 \) is shown in Fig. 22 (compare with Fig. 20).

The Output Circuit and Load

In the general physical description of the centimeter-wave magnetron, whose constituent parts are shown in Fig. 5, there remains the discussion of the output circuit. The output circuit is the means of coupling the fields of the magnetron resonator to the load. As such, it must contrive to induce a voltage across a coaxial line or a wave guide to which the load circuit is connected.

The most common coaxial-line type is illustrated schematically in Fig. 5 and may be seen in the photograph of the cutaway magnetron model of Fig. 21. A variation of the loop and coaxial output in which the loop is placed above the end of the resonator and the coaxial terminated in a junction to wave guide is shown in Fig. 23.

The wave-guide type of output circuit is shown schematically in Fig. 24 and may be seen in a slightly different form in the photograph of a cutaway magnetron model of Fig. 25.

In both types of output circuit the necessary transformer action is now designed into the magnetron structure so that the magnetron may operate at satisfactory loading when connected to a matched output line or wave guide without the use of external transformers. In the coaxial output this is accomplished by adjustment of the loop size and by the use of coaxial
transformer sections inside the vacuum envelope. In the wave-guide output the usual form of transformer is a quarter-wavelength guide section, inside the vacuum envelope, of characteristic impedance equal to the root-mean-square of that of the external guide and the low impedance desired at the magnetron resonator (see Fig. 24).

The Rieke Diagram

The Rieke diagram is the third fundamental performance characteristic of the magnetron oscillator (the others are the $V-B$ and $V-I$ plots of Figs. 17 and 18). It represents the dependence of output power and operating frequency on load. It is usually plotted as contours of constant output power and operating frequency on a reflection coefficient plane. A typical example for centimeter-wave magnetrons is shown in Fig. 26. The Rieke diagram may be explained in terms of the theory of a simple lumped-constant circuit equivalent to that of the radio-frequency circuit of the magnetron.

The Pulling Figure

The Rieke diagram completely specifies the dependence upon load of the magnetron output power and frequency of operation. Nevertheless, it is convenient to be able to specify by a single parameter the dependence of operating frequency on load changes. It is customary to specify as the so-called pulling figure $PF$ the total excursion of frequency, $\Delta f = \Delta \omega / 2\pi$, resulting from a standard variation in load susceptance, namely, that obtained by the total possible phase variation of a standing wave of 1.5 voltage ratio in the line at the point in question. This is equivalent to traversing the dashed circle on the reflection coefficient plane shown in Fig. 26. The pulling figure of the magnetron is inversely proportional to its external $Q$.

Conclusion

In this paper it has been possible to present only a discussion of the fundamental physical basis of operation of the modern magnetron oscillator. It might be extended, for example, to include treatments of how a good magnetron design may be scaled to other frequencies, voltages, currents, and magnetic fields; how the frequency may be stabilized; what happens when the magnetron operates into a frequency-sensitive load ("long-line effect"); how the oscillator may be tuned; what the nature of the frequency spectrum of a pulsed magnetron is; how oscillations build up; and what special considerations enter into the design of cathodes and magnetic circuits for use in magnetron oscillators. Discussion of these topics may be found elsewhere.14
Selective Demodulation

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Summary—A method of demodulation is proposed in which the output current of the demodulator is a linear function of the input voltage, while at the same time provision is made for producing the necessary product terms which will result in demodulation. Demodulation is brought about by integrating the product of the instantaneous value of the modulated wave by the instantaneous value of a wave having the same frequency and phase as the carrier. Where this method of demodulation is used it is proposed that two carriers in quadrature on the same frequency may be employed, reducing the bandwidth to that required for single-sideband transmission.

It is suggested that the required linear demodulation characteristics may be obtained through the use of "electron-coupled" demodulators. Theoretical considerations indicate that, when demodulation of this type is employed, selectivity ahead of the demodulator may be dispensed with, the signal-to-noise ratio is improved, greater economy of spectrum space is obtained, the number of tubes required is materially reduced through the use of a common intermediate-frequency amplifier for a number of channels, and any impairment due to the instability of the carrier or oscillator frequency is reduced.

As an example of the possible application of the principles outlined, a hypothetical eight-channel transmission system is described.

INTRODUCTION

When a modulated wave of the form

$$e = F(t) \cos(\omega_c t + \phi_c)$$

in which $e$ is the instantaneous voltage, $F(t)$ is a modulating component, $\omega_c = \omega_k/2\pi$ is a carrier frequency, and $\phi_c$ is the phase angle of the carrier at time $t = t_0$, is multiplied by the instantaneous value of the carrier, the expression for the product is

$$i = \gamma e \cos(\omega_c t + \phi_c) = \gamma F(t) \cos(\omega_c t + \phi_c)^2$$

$$= \gamma \frac{F(t)}{2} + \gamma \frac{\cos(2(\omega_c t + \phi_c))}{2}$$

(2)

where $i$ is the instantaneous output current, and $\gamma$ is a constant depending upon the characteristics of the device used to bring about the multiplication.$^1$

If the result is integrated over at least one carrier cycle the second term becomes zero, and the useful result is

$$I_0 = \gamma \frac{F(t)}{2}. \quad (3)$$

The original modulating component is thus restored. While the operations of (1) to (3) usually occur in any analysis of modulation or demodulation and are some-

times incidentally employed in other problems, their general importance, and sometimes even their existence in the analysis, are usually obscured by the difficulties of the modulation problem. It is the purpose of this paper to point out that this relationship constitutes a fundamental principle, through the application of which useful results may be obtained. Transmission systems may be developed in which a number of modulating components may be transmitted in properly spaced and phased carriers, and separated at the receiver without recourse to the usual tuning arrangements. Two carriers of the same frequency but differing in phase may be employed, and their respective modulating components selected in the receiver as desired, to create a dual-channel transmission system. The result of (3) may be employed to produce a spectrum analyzer in which the frequency observed at any given instant is selected solely through controlling the frequency of a beating oscillator. Receivers of the superheterodyne type but without intermediate-frequency amplifiers, in which the output of the mixer is fed directly into the audio or video amplifier, may be designed for the purpose of eliminating difficulties due to image frequencies.

It is not claimed that the material in this paper is completely new. To some extent the analysis follows lines made familiar by other papers on modulation and demodulation, and some of the principles outlined have been applied in practical circuits for many years. In particular, A. V. T. Day and H. Nyquist have independently made suggestions as to the employment of carriers in quadrature.$^2$ The relationships derived are, however, presented in a somewhat novel manner which the author has found helpful in visualizing the general problem of demodulation. Perhaps this publication of the general concepts involved may help others to develop novel applications in practical devices as yet unthought of. The author regrets that circumstances have made it impossible for him to verify his theoretical conclusions experimentally, and hopes that this paper will be regarded primarily as an exposition of general principles.

Generalization of the Problem

The principle may be generalized as follows:

Assume that a transmitted wave, the instantaneous voltage value of which is $e_t$, is impressed on the input of a circuit element so designed that its output is proportional to the product of $e_t$ and the instantaneous voltage value $e_c$ of a modulating wave applied locally to another terminal of the circuit element. The instantaneous

$^1$ Formerly, National Defense Research Committee, Division 15, Cambridge, Massachusetts; now, Northwestern Bell Telephone Company, Des Moines, Iowa.


useful output current will be
\[ i_o = \frac{dQ}{dt} = \gamma e_e d t \]
(4)

where \( Q \) is the quantity of electricity existing in the circuit at time \( t \) and \( \gamma \) is a constant depending on the characteristics of the circuit element.\(^2\)

Then
\[ dQ = \gamma e_e d t \]
(5)

and the charge flowing out of the circuit element during the time interval between \( t = 0 \) and \( t = T \) will be
\[ Q = \gamma \int_0^T e_e d t. \]
(6)

The average output current during this period is
\[ I_0 = \frac{Q}{T} = \frac{\gamma}{T} \int_0^T e_e d t. \]
(7)

It is now assumed, for the first case, that the transmitted wave is of the form of (1):
\[ e_t = F(t) \cos (\omega_m t + \phi_m). \]
(8)

\( F(t) \) may take a variety of forms. In the case of carrier-transmitted amplitude modulation it takes the form, for a single modulating frequency:
\[ F(t) = m E_b [1 + \cos (\omega_n t + \phi_n)] \]
(9)

where \( E_b \) is the amplitude of the carrier, \( m \) is the modulation index, and \( \omega_n \) and \( \phi_n \) are the angular velocity and phase angle, respectively, of the modulating vector. The complete expression for the transmitted wave in this case is
\[ e_t = E_b [1 + \cos (\omega_n t + \phi_n)] \cos (\omega_m t + \phi_m). \]
(10)

Where carrier-eliminated transmission is involved,
\[ F(t) = m E_b [\cos (\omega_n t + \phi_n)] \]
(11)

and the expression for the wave becomes
\[ e_t = m E_b [\cos (\omega_n t + \phi_n)] \cos (\omega_m t + \phi_m). \]
(12)

If it now be assumed that
\[ e_d = \cos (\omega_d t + \phi_d) \]
(13)

and that we wish to integrate over one complete cycle of the carrier, the period of which is
\[ T_c = \frac{1}{f_c} = \frac{2\pi}{\omega_c} = T, \]
(14)

we have, by substitution of (8), (13), and (14), in (7):

\[ I_0 = \frac{\gamma \omega_k}{2\pi} \int_0^{2\pi/\omega_k} F(t) \cos (\omega_m t + \phi_n) \cos (\omega_k t + \phi_k) dt \]

\[ = \frac{\gamma \omega_k}{2\pi} \int_0^{2\pi/\omega_k} F(t) \cos (\omega_m t + \phi_n) dt \]

\[ + \frac{\gamma \omega_k}{2\pi} \int_0^{2\pi/\omega_k} F(t) \cos 2(\omega_m t + \phi_m) dt \]

\[ = \frac{\gamma \omega_k}{2\pi} \int_0^{2\pi/\omega_k} F(t) \cos (\omega_m t + \phi_m) \cos (\omega_m t + \phi_m) dt \]

\[ = \frac{\gamma F(1)}{2}. \]
(15)

This result is true if the period of \( F(t) \) is so long in comparison with \( T_c \) that \( F(t) \) may be considered to remain constant within the limits of integration.\(^6\)

Rewriting (15) in order to emphasize the salient features of the relationship demonstrated, we have
\[ I_0 = \frac{\gamma \omega_k}{2\pi} \int_0^{2\pi/\omega_k} F(t) \cos (\omega_m t + \phi_m) \cos (\omega_m t + \phi_m) dt \]

\[ = \frac{\gamma F(t)}{2}. \]
(16)

This equation states in effect that when the product of the instantaneous value of a modulated wave and the instantaneous value of a wave having the same frequency and phase as the carrier are integrated over one complete carrier cycle, the integrated output current is a linear function of the original modulation component; complete demodulation is therefore effected by this process.

It is now to be observed that the principle may be further generalized to take into account two carriers on the same frequency but in quadrature. If instead of the single carrier of (8) we have one cosine and one sine carrier, each with its own modulation, the equation of the wave becomes
\[ e_t = F_n(t) \cos (\omega_n t + \phi_n) + F_s(t) \sin (\omega_s t + \phi_s). \]
(17)

If in the receiving demodulator the product of \( e_t \) and \( e_d = \cos (\omega_d t + \phi_d) \) is integrated, we obtain for the average output current, in accordance with (7):
\[ I_0 = \frac{\gamma \omega_k}{2\pi} \int_0^{2\pi/\omega_k} \left[ F_n(t) \cos (\omega_n t + \phi_n) + F_s(t) \sin (\omega_s t + \phi_s) \right] \cos (\omega_k t + \phi_k) dt \]

\[ = \frac{\gamma F_n(t)}{2} \cos (\omega_n t + \phi_n) \sin (\omega_n t + \phi_n) dt \]

\[ + \frac{\gamma F_s(t)}{2} \sin (\omega_s t + \phi_s) \cos (\omega_s t + \phi_s) dt \]

\[ = \frac{\gamma F_n(t)}{2} \cos (\omega_n t + \phi_n) \sin (\omega_n t + \phi_n) dt \]

\[ + \frac{\gamma F_s(t)}{2} \sin (\omega_s t + \phi_s) \cos (\omega_s t + \phi_s) dt \]

\[ = \frac{\gamma F_n(t)}{2}. \]
(18)

\( \) It is to be noted that in the case of (15) integration over one-half carrier cycle would have been equally effective in eliminating the undesired alternating component. The limits \( t=0 \) and \( t=T_k \) are, however, employed for consistency, as it will later be demonstrated (see (21)) that they are required when the transmitted wave contains two or more frequencies in harmonic relationship.
While, if the demodulating voltage \( e_d \) is sin \((\omega_1 t + \phi)\), by a similar process not written out in detail we obtain,

\[
I_0 = \int_0^{2\pi/\omega_1} [F(t) \cos (\omega_1 + \phi)]
+ F(t) \sin (\omega_1 t + \phi_1)] \sin (\omega_1 t + \phi) dt
= \frac{\gamma F(t)}{2}.
\]

(19)

Either signal may therefore be demodulated at will merely by selecting the proper demodulating voltage \( \cos (\omega_1 t + \phi) \) or \( \sin (\omega_1 t + \phi) \), as required. Equations (18) and (19) may be applied in practice to effect a dual-channel transmission system, in which a single carrier frequency serves both channels. The benefits of single-sideband transmission may thus be realized without elaborate filtering means.

Also, if more than one modulated carrier frequency is received by the demodulator, as when,

\[
e_i = F(t) \cos (\omega_{31} t + \phi_{31}) + F(t) \sin (\omega_{31} t + \phi_{31})
+ F(t) \cos (\omega_{32} t + \phi_{32}) + F(t) \sin (\omega_{32} t + \phi_{32})
\]

(20),

the output current integrated over one complete period of carrier \( K1 \), for the case when the demodulating voltage \( e_d \) is \( \cos (\omega_{31} t + \phi_{31}) \), will be

\[
I_0 = \frac{\gamma \omega_{31}}{2\pi} \int_0^{2\pi/\omega_1} [F(t) \cos (\omega_{31} t + \phi_{31})
+ F(t) \sin (\omega_{31} t + \phi_{31})] \cos (\omega_{31} t + \phi_1) dt
= \frac{\gamma F(t)}{2} + \frac{\gamma \omega_{31}}{2\pi} \int_0^{2\pi/\omega_1} F(t) \sin (\omega_{31} t + \phi_{31}) dt
\]

(21)

If \( \omega_1 = N\omega_0 \), \( N \) being an integer, so that an exact multiple of a period of each carrier elapses during the time interval between \( t = 0 \) and \( t = T_{11} = 2\pi/\omega_0 \), all integrals are zero, and the expression reduces to

\[
I_0 = \frac{\gamma F(t)}{2}.
\]

(22)

Similarly

\[
I_0 = \frac{\gamma \omega_{31}}{2\pi} \int_0^{2\pi/\omega_1} \sin (\omega_{31} t + \phi_{31}) dt = \frac{\gamma F(t)}{2}
\]

(23)

and

\[
I_0 = \frac{\gamma \omega_{31}}{2\pi} \int_0^{2\pi/\omega_1} \cos (\omega_{31} t + \phi_{31}) dt = \frac{\gamma F(t)}{2}.
\]

(24)

Equations (22), (23), (24), and (25) show that if a modulated wave contains a number of carriers in harmonic relationship, each frequency having one cosine and one sine carrier, the modulating component of any one of the carriers may be restored in the receiver and all other components eliminated merely by multiplying the instantaneous value of the wave by the particular carrier to be selected, and integrating the result over one cycle of the lowest carrier frequency.

If this result be expressed in the most general possible terms, then

\[
I_0 = \frac{\gamma \omega}{2\pi} \int_0^{2\pi/\omega} [\sum F(t) \cos (\nu \omega_1 + \phi_n)]
+ \sum F(t) \sin (\nu \omega_1 + \phi_n)] \cos (\nu \omega_1 + \phi_0) dt
= \frac{\gamma F(t)}{2}
\]

(26)

and

\[
I_0 = \frac{\gamma \omega}{2\pi} \int_0^{2\pi/\omega} [\sum F(t) \cos (\nu \omega_1 + \phi_n)]
+ \sum F(t) \sin (\nu \omega_1 + \phi_n)] \sin (\nu \omega_1 + \phi_0) dt
= \frac{\gamma F(t)}{2}.
\]

(27)

Here, \( f = \omega/2\pi \) is taken to be the lowest, or fundamental, carrier frequency in the series of which all other carriers are harmonics. The number of the harmonic is indicated by the value of \( n \), which may be any integer.

Subscript \( a \) identifies modulating components that modulate cosine carriers; subscript \( b \), modulating components that modulate sine carriers.

Equations (26) and (27) are, of course, a statement of Fourier's theorem. In its most general form, therefore, the principle of (1) to (3) proposes the artificial creation, in the transmitter, of a Fourier series in which each cosine or sine function is a carrier, and each coefficient \( F_n(t) \) or \( F_b(t) \) is a modulating component. The resulting complex wave containing all components is transmitted to a receiver, and thus carries, in a single envelope, the intermingled intelligence of all channels. By performing in the receiving demodulator an automatic Fourier analysis, that is, by taking successively the product of the instantaneous value of the wave by the particular carrier selected for demodulation, and integrating, the modulation component of that carrier is restored and all other components are eliminated. It is
to be noted that recourse to the usual tuning methods is not taken in the receiver, the only requirement being that the pass band of the demodulator output shall extend from zero to an upper frequency equal to the highest frequency in the modulation spectrum.

Modulators and Demodulators

So far reference has been made only to hypothetical "circuit elements" constituted to bring about the results required. It was inferred in the note related to (4) that the requirements might be fulfilled by "balanced" demodulators of conventional design. Actually, of course, some improvement over such demodulators is required in order to perform the operations of (26) and (27) without introducing objectionable cross modulation. If a complex wave of the type defined by (20) were impressed on the input of a conventional balanced demodulator without previously filtering off the unwanted modulated carrier frequencies, interaction between the various carriers and sidebands in the impressed wave due to the nonlinear characteristic of the demodulator would produce spurious modulation products having frequencies within the acceptance band of the demodulator output filter, thereby causing cross talk and spurious responses. It is true that the use of a balanced demodulator will eliminate even-order cross-modulation products; but products of odd order will be unaffected by the balanced arrangement.

In order to eliminate this effect, it is necessary that the instantaneous plate current of the demodulator be a linear function of the impressed grid voltage \( e_i \) while, at the same time, means are provided for producing in the plate circuit terms representing the product of \( e_i \) and \( e_a \), the demodulating voltage. These requirements are rather satisfactorily met by demodulators in which the demodulation is performed by electron coupling. For example, in the pentagrid mixer, a suitable choice of operating parameters will result in a linear relationship between the signal-grid (grid 1) to plate transconductance \( g_{m1} \) and the oscillator-grid (grid 3) voltage \( e_a \) as indicated in Fig. 1. Here it is evident that if the characteristic is substantially a straight line in the operating range selected, as shown,

\[
g_{m1} = \gamma (e_i - e_a) \tag{28}
\]

where \( e_a \) is the voltage value at which an extension of the straight portion of the characteristic intersects the horizontal axis and \( \gamma \) is a constant representing the slope of the curve. This constant, which might be called the "gamma factor" of the tube, is evidently defined by

\[
\gamma = \frac{dg_{m1}}{de_i} \tag{29}
\]

and is the rate of change of the signal-grid to plate transconductance with change in oscillator-grid voltage.

The relationship expressed by (28) is, of course, strictly true only for a given, constant value of signal-grid voltage. Variations in signal-grid voltage will also cause \( g_{m1} \) to vary. However, the characteristics of the 6L7 tube are such that, if the amplitude of the voltage applied to the signal grid is kept small, variations in \( g_{m1} \) due to this cause will be negligible. Nesslage, Herold, and Harris have, in fact, suggested that the tube is suited for use as radio-frequency amplifier where a steep control characteristic is desired without sacrificing the benefits of remote-cutoff operation.\(^7\)

For small variations of \( e_i \) no more cross modulation should therefore be expected than would be obtained with a remote-cutoff pentode. It is assumed in the following analysis that \( e_i \) will be kept small at all times, and that (28) will accordingly be applicable.

The alternating component of the plate current, due to voltages impressed on the signal grid is, as is well known,

\[
i_p = \frac{g_{m1}e_i}{r_p + R_L} \tag{30}
\]

where \( r_p \) and \( R_L \) are the alternating-current plate resistance and the alternating-current load resistance, respectively. The oscillator frequency impressed on the oscillator grid will also be amplified and will appear in the plate circuit, so that the total alternating output current will be

\[
i_o = - (g_{m1}e_i + g_m'e_d) \frac{r_p}{r_p + R_L} \tag{31}
\]

where \( g_{m'd} \) is the oscillator-grid-plate transconductance.

Now the alternating-current plate resistance of the tube is nominally rated at greater than 1 megohm, and may be expected to remain extremely high for all values of oscillator-grid voltage. Under these conditions, \( R_L \) may be chosen so that it is negligibly small with respect to \( r_p \), and the factor \( r_p/r_p+R_L \) will approximate unity. Variations in \( r \), which may occur during the cycle will therefore have a negligible effect on \( i_o \) and (31) will reduce to

\[
i_o = -(g_{m1}e + g_{md}e_d).
\]

Substituting the value of \( g_{m1} \) previously derived, we have

\[
i_o = -[\gamma(e_d - e_0)e + g_{md}e_d]
= -[\gamma e_d e - \gamma e_0 e + g_{md}e_d].
\]

(33)

Inspection shows that the first term on the right-hand side of the equation is the product term desired (see (4)), the second term is the amplified input wave, and the third term is the oscillator frequency. All terms of the first order, and cross modulation will not result between various components of \( e \). It is also noted that the second and third terms represent radio-frequency components which will not pass through the demodulator output filter, so that to all intents and purposes the output of the tube may be represented as

\[
i_o = -\gamma e_d e_d,
\]

an equation identical with (4) except for the change in sign brought about by conventional considerations as to the direction of current flow in the plate circuit.

It is therefore suggested that the operations of (26) and (27) may readily be performed practically, and without the creation of undesirable cross modulation, by employing a pentagrid mixer or similar electron-coupled tube, operated in the center of the straight part of the \( g_{m1} - e_d \) curve, in which the input wave

\[
e_i = \left[ \sum F_{on}(t) \cos (\omega t + \phi_n) \right.
+ \sum F_{bn}(t) \sin (\omega t + \phi_n) \left. \right]
\]

(35)

is impressed on the signal grid (grid 1), and the demodulating wave

\[
e_d = \cos (\omega t + \phi_n) \quad \text{or} \quad e_d = \sin (\omega t + \phi_n)
\]

(36)

is impressed on the oscillator grid (grid 3), as indicated in Fig. 2.

It may be of interest to note in passing that the modulation process described in this paper is actually carried out in an indirect manner in conventional nonlinear demodulators. For example, consider the demodulation of a carrier-transmitted wave,

\[
e_i = [1 + F(t)]e_k
\]

(37)

where \( F(t) \) has for brevity been substituted for the more usual \( m \cos (\omega t + \phi_m) \) and \( e_k \) for \( E_k \cos (\omega t + \phi_k) \). If this wave is applied to the input of a square-law demodulator, the second-order terms of the alternating component of the output current will be

\[
i_o = ke_k^2 = k\left[1 + F(t)e_k\right]^2
= k\{e_k + F(t)e_k\}^2
= k\{e_k^2 + 2F(t)e_k^2 + [F(t)e_k]^2\}
\]

(38)

where \( k \) is a constant depending on the characteristics of the demodulator.

![Fig. 2—Linear demodulator employing electron coupling.](image)

It is to be observed that the second term, \( 2F(t)e_k^2 \), is identical with the expression under the integral sign in (15), except for the coefficient and the fact that in this case \( F(t) \) is specifically defined as \( m \cos (\omega t + \phi_m) \), while in (15) it may take this meaning or may also be \( [1 + m \cos (\omega t + \phi_m)] \). This term is the only term in the output of a square-law demodulator which leads to useful demodulation products, as the first term produces merely a direct-current component and a component having twice the frequency of the carrier; and the last term contains a large number of cross-modulation products between the various sideband components.

It may therefore be said that (15) merely provides a direct method of carrying out the process which is brought about indirectly, and with the creation of undesirable spurious cross modulation and distortion products, by ordinary nonlinear demodulation techniques.

**Practical Applications**

Practical applications of this principle would seem to lie in two fields. First, the creation of a complete series containing a large number of different carrier frequencies would appear to be limited by the radio-frequency pass band of transmitting and receiving equipment. The usefulness of this approach is, therefore, restricted to cases where a limited number of carriers can be employed in a multichannel transmission system, or where the "carrier" frequencies can be made very small, so that a large number of adjacent frequencies can be transmitted in a relatively narrow spectrum. If, for example, the fundamental carrier frequency is taken to be 20 cycles, a total of 500 carriers, each with its own modulation, can be transmitted in a spectrum 10,000 cycles wide. The latter possibility might have application in a facsimile...
system where each point in the object could be represented by a separate carrier frequency, all carriers being transmitted simultaneously. In such a system the condition qualifying (15), that the period of $F(t)$ remain large with respect to $T_a$, would be valid in spite of the low frequency of the carrier, as $F(t)$ would be constant.

Second, the presence of both cosine and sine terms in the series offers possibilities in connection with the reduction of the bandwidth of systems operating at conventional radio frequencies.

The following example of an application of the general principle involved is presented by way of clarification and emphasis, and is not necessarily intended to describe a practical workable system possessing advantages with respect to existing systems such that its development and construction would be desirable. It is felt, however, that some of the considerations discussed in connection with this example might be useful in calling attention to general practical approaches which could be employed as a solution of other problems not yet actively being investigated.

**Multichannel Communication System**

Fig. 3 shows in block form the receiving portion only of a communication system utilizing the implications of (26) and (27). In this system provision is made for eight voice-frequency channels, of which four are operated with cosine carriers, and four with sine carriers. Four carrier frequencies only are provided for the eight channels.

The frequency employed in the radio-frequency section of this receiver may be selected as required, and is accordingly not shown in the figure. It is, however, assumed that the intermediate-frequency amplifier will center at 603 kilocycles, and that the spectrum of the received signal in the intermediate-frequency circuit will be as indicated in Fig. 3. It is seen from this figure that the four carriers have frequencies of 594, 600, 606, and 612 kilocycles, respectively. Double-sideband transmission is used, so that the modulation spectrum extends 3000 cycles to the side of each carrier, and the acceptance band of the intermediate-frequency amplifier must include frequencies from 591 to 615 kilocycles, a bandwidth of 24 kilocycles. It is assumed that the transmitted wave is of the form of (20), extended to include four carrier frequencies. The carrier may or may not be transmitted, depending upon the degree of cross-talk balance required. It is noted that, since the demodulators are all of the linear type mentioned under the heading “Modulators and Demodulators,” demodulation will not result unless the carrier is applied to the oscillator grid, even though the transmitted wave applied to the signal grid contains the carrier. Transmitter design is simplified if the carrier is transmitted, as balanced modulators in the transmitter are not required under these conditions. On the other hand, if a high degree of freedom from cross talk is mandatory it may be necessary to suppress the carrier, as a slight curvature in the signal-grid to plate transconductance characteristic of the demodulator might result in some demodulation independently of the action of the oscillator grid, if the carrier were transmitted.

The received wave, containing the eight modulation envelopes, is impressed in parallel on the signal grids of all the demodulators, a separate demodulator being provided for each channel. Demodulation is effected by applying the appropriate demodulating voltage $e_{a}$ to the oscillator grid of each demodulator. This demodulating voltage will, of course, be of the same frequency and phase as the carrier of the particular channel to be demodulated.

Demodulating voltages of the proper frequencies are derived in the following manner:

The output of the intermediate-frequency amplifier is delivered in parallel with the demodulators to a carrier amplifier tuned to the frequency of one of the carriers. This particular carrier must, of course, be transmitted and will be referred to hereafter as the "control carrier." It is not necessary or desirable, however, that a large amount of carrier be provided, and in general it would be expected that the modulation index would be very large. In the figure, the 600-kilocycle carrier has been chosen as the control carrier. It is assumed that all carriers in the transmitter will also have been derived from this same control carrier, in order to maintain a uniform phase standard.

The carrier amplifier is then tuned exactly to 600 kilocycles, by means of crystal tuning if necessary, in order to eliminate the sidebands to the greatest possible extent. The 600-kilocycle output is delivered to a multivibrator operated at the tenth subharmonic, to derive a 60-kilocycle frequency. This wave in turn is impressed on an amplifier sharply tuned to 60 kilocycles for the purpose of further attenuating any sidebands which may have resulted from modulation of the multivibrator.

The output of the 60-kilocycle amplifier is delivered to a second multivibrator, which again reduces the frequency by a factor of 10. The resulting 6-kilocycle wave
passes through a third amplifier tuned sharply to 6 kilocycles. The output of this amplifier should now be entirely free of any sideband components, and should consist of a pure 6-kilicycle wave.

The output of the 6-kilicycle amplifier is now employed to control a third multivibrator at the fundamental frequency of 6 kilocycles. The output of this multivibrator will, therefore, contain harmonics at 6-kilicycle intervals, and if a square wave shape is assumed, the amplitudes of these harmonics will be inversely proportional to the orders of the harmonics. Useful amplitudes may therefore be expected for harmonics as high as the several hundredth.

The 99th harmonic of the final multivibrator stage, having a frequency of 594 kilocycles, is delivered to a carrier amplifier tuned exactly to this frequency, in order to eliminate adjacent harmonics at 600 and 588 kilocycles. The output of the carrier amplifier, which is assumed to be a cosine wave and can be made exactly so by adjusting the "cosine" phasing network, is applied to the oscillator grid of cosine demodulator 1a. The output of this demodulator will, therefore, be the modulation attached to the cosine carrier having a frequency of 594 kilocycles, or the nomenclature of (20), $F_{99}(t)$. (It is to be noted that strict adherence to the subscript definitions previously established would require that this component be identified as $F_{99s}(t)$, but this nomenclature is avoided here because $F_{99}(t)$ more satisfactorily associates the modulation with the demodulator involved.)

The output of the 594-kilicycle carrier amplifier is also applied to a "sine" phasing network, which alters the phase 90 degrees. The resulting sine wave is delivered to the oscillator grid of demodulator 1b, so that the output of this demodulator will be the modulation attached to the sine carrier having a frequency of 594 kilocycles, or $F_{99}(t)$.

The 100th harmonic of the final multivibrator stage, having a frequency of 600 kilocycles, is, in a similar manner, selected by a carrier amplifier and applied as a cosine function to demodulator 2a and as a sine function to demodulator 2b, in order to derive modulation components $F_{100s}(t)$ and $F_{100c}(t)$. Similar treatment with respect to the 101st harmonic, 606 kilocycles, and the 102nd harmonic, 612 kilocycles, results in producing at the output of demodulators 3a, 3b, 4a, and 4h, modulation components $F_{101s}(t)$, $F_{101c}(t)$, $F_{102s}(t)$, and $F_{102c}(t)$, respectively.

The demodulators all have acceptance bands extending from 200 to 3000 cycles and cutting off sharply above 3000 cycles in order to eliminate frequencies outside the band. Each is provided with an audio amplifier for the purpose of delivering the audio output to a line or another required termination.

A general appraisal of the merits of this system reveals that a complete eight-channel radio receiver is provided with a total of 30 tubes (double tubes are assumed to be used in the multivibrators), or a ratio of 3½ tubes per channel. Full advantage of the available spectrum width is taken, as the eight channels occupy a spectrum 24 kilocycles wide, 3 kilocycles per channel. In part, this spectrum economy is made possible by avoiding the use of input filters and placing the channels immediately adjacent in the spectrum. In part, it is due to the employment of cosine and sine carriers on the same frequency, so that the spectrum is reduced to that required for single-sideband operation. It should be noted that channels may be added to the receiver as required merely by providing an additional demodulator, audio amplifier, and ½-carrier amplifier per channel, and broadening the intermediate-frequency stages. For example, 16-channel operation may be arranged with a total of 50 tubes, a ratio of 3½ tubes per channel. Under these conditions, the bandwidth will be 48 kilocycles. In spite of the wide intermediate-frequency band, the signal-to-noise ratio should be excellent, as only the noise contained in that part of the spectrum occupied by the sidebands being demodulated will appear in the audio output.

Considerable simplification in design is afforded by eliminating filtering ahead of the demodulators. The only filtering employed is in connection with the derivation of the demodulating carriers, and as these carriers contain no modulation components, a simple high-Q tuned circuit is all that is needed to bring about effective filtering, the difficulties of band-pass filter design being avoided. In fact, except for the simple tuned circuits at the demodulator inputs, no band-pass filters are employed in the demodulating section of the receiver, the demodulator output filters being essentially of a low-pass character.

The performance of this system will, of course, depend in practice to a very large extent on the degree of balance that can be maintained. Any variations in phase or balance may produce cross talk or the generation of other spurious products. It appears that cross talk of this type may originate principally from two sources.

First, it is expected that some cross modulation may be produced in the demodulator, due to nonlinearity of the control-grid to plate characteristic. With careful design of the tube it should be possible to limit such cross modulation to a low level, probably 40 decibels below the level of the wanted components. Further attenuation of intelligible cross talk can be brought about by suppressing the carriers, as restoration of the original modulation will not be effected under any conditions unless the carrier is present. If an adequate degree of balance can be maintained in the transmitting balanced modulators under these conditions, the level of the transmitted carrier should not be greater than 30 decibels below the level of the sidebands. It is, therefore, not unreasonable to expect that the level of intelligible cross talk, due to curvature of the demodulator characteristic, should not be higher than 70 decibels below the level of the wanted modulation at the output of the demodulator. Spurious components arising from nonlinearity might be further reduced by employing...
a balanced arrangement in the receiving demodulators.

Second, cross talk between the cosine and sine channels on a given frequency will result if exact quadrature is not maintained between the modulated carriers at the transmitter, or if the demodulating carrier does not remain exactly in phase with the modulated carrier at all times. Phase variations in any of these carriers may result from a number of factors, such as selective fading, changes in circuit constants due to temperature variation, and instability of the multivibrators. Of these factors, the last is probably the most important. It is, however, known that it is possible to construct multivibrator systems capable of generating pulses at a pulse repetition frequency of 6000 with a maximum variation in the time position of the pulse not exceeding $2 \times 10^{-10}$ seconds. At the chosen intermediate frequency of 600 kilocycles, a time interval of this magnitude represents a phase displacement of $2 \times 10^{-5} \times 6 \times 10^{10}$ cycles, or about 0 degrees 2.5 minutes. If a phase displacement of this magnitude takes place in the demodulating carrier, the relative amplitude of the undesired modulation having the same carrier frequency will be raised from 0 to sin $(\varphi 2.5') = 0.00073$ with respect to the amplitude of the desired modulation, which will remain at 1.00000. The current ratio between the desired and undesired components will be 1370, which represents a difference in level of 63 decibels. It therefore seems reasonable to suppose that, with careful design, the level of undesired modulation components resulting from variations in carrier phase might be kept more than 60 decibels below the level of the desired modulation at the output of the demodulator. It is recognized, of course, that some difficulty might be experienced in practice with the first multivibrator stage, as the initial frequency of 600 kilocycles is somewhat high for reliable multivibrator operation. This difficulty might be surmounted by refinement in the multivibrator design, or by substituting some other type of frequency divider for the first multivibrator stage. It might also be desirable to provide some sort of automatic phase control in order to assure that the multivibrator output will at all times remain in phase with the control carrier. It would appear that a phase control of this type could easily be arranged to operate from any residual direct current in the demodulator output of the "sine" demodulator on the control carrier frequency which might result from phase displacement in the cosine control carrier.

It might, of course, be found desirable to rearrange the frequency and phase allocations indicated in Fig. 3, in order to suit the particular operating conditions required. For example, in two-way service it might be found desirable to reserve all cosine channels for east-to-west operation, allocating the corresponding sine channels to west-to-east operation. Under these conditions, operations in both directions would be on the same frequencies, and any undesired demodulation products resulting from phase irregularities would appear as side tone or singing rather than as cross talk.

At first glance, it might appear that irregularities in the phase or frequency of the local oscillator of the receiver might produce such a degree of phase unbalance as to render the system inoperative. Consideration of this question confirms, however, that any such irregularities imparted to the modulated carriers will equally affect the control carrier, and hence the demodulating carriers, so that the net effect on the balance of the system will be nil if it be assumed, of course, that the multivibrator system can be designed to keep in step with these irregularities at all times.

Variations in the carrier or local-oscillator frequency will, of course, render separation of the control carrier from the sidebands difficult, as the carrier-amplifier-multivibrator system must be very sharply tuned to produce the required result, and carrier or local-oscillator instability may cause the control carrier to deviate outside the acceptance band of the carrier system. Where extremely high carrier frequencies are involved, it may be necessary to separate the control carrier entirely from the sidebands, placing it several kilocycles distant in the spectrum. The acceptance band of the carrier-amplifier-multivibrator system might then be widened so that deviations of several kilocycles in either carrier or oscillator frequency could be tolerated. It is to be noted that, in a system of this type, expedients of this kind employed to counteract instability will have no effect on the selectivity of the system, as the control carrier and the sidebands deviate together and the selectivity is obtained in the audio system. Difficulties in the design of the intermediate-frequency system may, therefore, be avoided by making the bandwidth of the intermediate-frequency amplifiers much wider than the spectrum of the received signal without impairing the selectivity of the system. It has previously been shown that an increase in intermediate-frequency bandwidth will not increase the signal-to-noise ratio.

In summary, it would appear that, although no system of the type described has been constructed, there are no insuperable obstacles to its successful performance. It is to be expected, of course, that many difficulties not anticipated herein would be encountered if an attempt were made to construct a prototype. Nevertheless, in view of the evident advantages offered, the effort might be justified.

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Input Admittance of Cathode-Follower Amplifiers

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Summary—General expressions are derived for the input admittance, conductance, and susceptance of a cathode-follower amplifier, and curves are given that show the effect of frequency and circuit parameters upon the input conductance and the effective input capacitance for typical values of plate resistance, transconductance, and interelectrode capacitances. The analysis shows that a capacitance shunting the load resistance of a cathode-follower amplifier may result in a negative input conductance. If the load capacitance is of the order of magnitude of the interelectrode capacitance or greater, this negative conductance is of the order of 5 X 10⁻⁴ mho at frequencies at which ωC₂ is of the order of 2 X 10⁻⁴ mho and above. For typical values of C₂ this value of ωC₂ corresponds to frequencies of the order of 50 megacycles and above. Negative conductance of this magnitude may readily cause oscillations in the input circuit unless preventive measures are taken. A number of such measures are discussed.

One important function of cathode-follower amplifiers is to provide a very high input impedance. For this reason a knowledge of the effect of circuit and tube parameters upon the input admittance of cathode-follower amplifiers is useful. Furthermore, since the input conductance of a cathode-follower stage may be negative and may therefore lead to oscillation in input circuits, it is important to determine the manner in which low values of negative input conductance may be avoided.

Fig. 1—Cathode-follower amplifier with tapped cathode resistor.

Fig. 2—Cathode-follower amplifier in which the grid resistor is connected to the cathode.

Fig. 3—Equivalent plate circuit for the circuit of Fig. 2 when the reactance of C₁ is negligible.

Theorem to the circuit of Fig. 2 yields the parallel equivalent plate circuit of Fig. 3 at frequencies below those at which electron-transit time must be considered. In this circuit, rₚ is the alternating-current plate resistance of the tube, gₑ the transconductance, and zₑ the admittance of the parallel combination of rₚ, zₑ, and Cₑ. The tube factors rₚ and gₑ are assumed to be constant. The total load admittance 1/zₑ may be written in the form

\[
\frac{1}{zₑ} = gₑ + j bₑ = gₑ + j \left( ωCₑ - \frac{1}{ωLₑ} \right)
\]

in which Cₑ and Lₑ are the capacitance and the inductance of a parallel combination of resistance, capacitance, and inductance equivalent to the actual total load, including rₚ and Cₚₑ.

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Summation of voltages in the circuit of Fig. 3 yields the equation

\[ E = I \frac{r_c}{j\omega C_{eq} + 1} + \left( I + \frac{1}{C_{eq}} E_v \right) \frac{1}{y_b + g_p} \quad (1) \]

where \( g_p = 1/r_p \). It is apparent from Fig. 2 that the alternating grid voltage \( E_v \) is the voltage drop produced by the flow of the current \( I \) through the parallel combination of \( r_c \) and \( C_{eq} \). Therefore,

\[ E_v = I r_c / (j\omega C_{eq} + 1). \quad (2) \]

Solution of (1) and (2) gives the following relation for the input admittance:

\[ y_s = g_i + jh = j\omega C_{eq} + \frac{1}{E} \]

\[ = j\omega C_{eq} + \frac{y_b + g_p}{1 + j\omega r_c / (j\omega C_{eq} + 1)} \quad (3) \]

Manipulation of (3) gives the following expressions for the effective input conductance \( g_i \) and the input susceptance \( b_i \):

\[ g_i = \frac{1}{(1 + 1)} \left( g_v + g_s \right) \left( 1 + \frac{r_s + 1}{1} \right) \frac{m^2}{r_c} \]

\[ = \frac{1}{(1 + 1)} \left( g_v + g_s \right) \left( 1 + \frac{r_s + 1}{1} \right) \frac{m^2}{r_c} \]

\[ b_i = \omega \left( C_{eq} + C_{Rf} \right) \left( 1 + 1 \right) \]

\[ = \omega \left( C_{eq} + C_{Rf} \right) \left( 1 + 1 \right) \]

\[ \text{where} \]

\[ m = \frac{y_b + g_p}{\omega C_{eq}} \left( C_{eq} + 1 \right) \quad (6) \]

\[ \lambda = \omega r_c C_{eq} \quad (7) \]

\[ \gamma = g_v + g_s + \omega \lambda \quad (8) \]

Examination of (4) shows that the input conductance may be negative if \( m \) is positive, i.e., if the load is capacitive, but is always positive if the load is inductive. Equation (5) shows that the input susceptance is always positive, or capacitive.

Several limiting cases are important:

(a) When \( r_c \) becomes infinite, (4) and (5) reduce to

\[ g_i = \frac{1}{(m + 1)^2} \left( g_v + g_s \right) \quad \frac{m^2}{r_c} \quad (4a) \]

\[ b_i = \omega \left( C_{eq} + C_{Rf} \right) \left( 1 + \frac{1}{(m + 1)^2} \right) \quad \frac{m^2}{r_c} \quad (5a) \]

(b) As the frequency is increased, \( g_i \) approaches the limiting value

\[ g_i \rightarrow \frac{g_v + g_s + m^2 g_e - m^2 \gamma}{(m + 1)^2} \quad (9) \]

(c) As the frequency approaches zero, \( g_i \) approaches the value

\[ g_i \rightarrow \frac{g_v + g_s + m^2 g_e}{g_m + g_p + g_b + g_e} \quad (10) \]

where \( g_p = 1/r_p \).

Because sustained oscillation will be set up in an oscillatory circuit shunted by a negative conductance if the magnitude of the negative conductance exceeds the positive conductance of the circuit at resonance, it is important to analyze in some detail the input conductance when the load is capacitive, i.e., when \( m \) is positive. In Figs. 4 and 5 are shown typical curves of the input conductance at infinite frequency as a function of \( m \) for capacitive load. The type 6AG5 tube, connected as a triode and as a pentode and used with low and high

![Fig. 4—Limiting value of input conductance \( g_i \) approached at high frequency (transit-time effects neglected). Triode-connected 6AG5, \( r_c = 4 \times 10^{-4} \) mho, \( g_p = 10^{-4} \) mho, \( g_b = 10^{-4} \) mho.](image-url)

values of load resistance, gives four combinations of tube and circuit parameters typical of those likely to be encountered in practice.
The minima in the curves of Figs. 4 and 5 occur for a value of \( m \) given by the relation

\[
m = \frac{C_b/C_{eb}}{g_m + 2(g_p + g_b)}. \tag{11}
\]

By substituting (11) into (9) and making \( r_c \) infinite, the maximum negative value of \( g_i \) is found to be

\[
\max g_i = -\frac{g_m^2}{4s} = -\frac{g_m^2}{4(g_p + g_b + g_0)}. \tag{12}
\]

The magnitude of the negative conductance indicated by Figs. 4 and 5 is sufficiently great to make possible sustained oscillation in oscillatory circuits having values of \( Q \) normally encountered. Such oscillations are frequently observed in cathode-follower amplifiers not provided with damping resistance in the grid circuit.\(^3\)

From (5) the effective input capacitance is seen to be

\[
C_i = C_{ep} + C_{eb}(A+1) - \frac{(g_p + g_b)sr_c^2}{A(m+1+sr_c+1)^2 + (m-r_s)^2}
\]

\[= C_{ep} + C'. \tag{13}\]

![Graph showing input conductance as a function of \( \omega C_{eb} \)](image)

Since \( C_{eb} \) is independent of load and frequency, it is sufficient to analyze the expression for \( C' \), which may be written in the form

\[
\frac{C'}{C_{eb}} = (A+1) \frac{m(m+1+sr_m+1)}{A(m+1+sr_c+1)^2 + (m-r_s)^2}. \tag{14}\]

![Graph showing \( C'/C_{eb} \) as a function of \( \omega C_{eb} \)](image)

As \( \omega C_{eb} \) becomes infinite, (14) reduces to

\[
\frac{C'}{C_{eb}} = \frac{m}{m+1}. \tag{15}\]

\(^3\) K. Schlesinger, "Cathode-follower circuits," Proc. I.R.E., vol. 33, pp. 843-855; December, 1945. It is of interest to note that the circuit may also be analyzed as a form of Colpitts oscillator.
As \( \omega C_{\phi h} \) approaches zero, (14) reduces to

\[
\frac{C'}{C_{\phi h}} = \frac{m(r_g + 1) + (g_p + g_b)sr_p^2}{(r_g + 1)^2}.
\]

(16)

Fig. 7 shows curves of \( C'/C_{\phi h} \) for the same tube and circuit parameters as those used for Figs. 4 and 6. Curves for other values of tube factors and load resistance are of the same form, but differ in the value of \( C'/C_{\phi h} \) approached at low values of \( \omega C_{\phi h} \). Examination of these curves or of (6) discloses that the effective input capacitance decreases with increase of \( r_e \) and of \( 1/g_b \), which is usually nearly equal to \( r_p \). At values of \( \omega C_{\phi h} \) below \( 10^{-4} \) mho, \( C' \) becomes very small for large values of \( r_e \) and \( r_p \), and the total effective input capacitance approaches the value \( C_{\phi h} \).

The tendency of a cathode-follower stage to oscillate can be prevented either by insuring that the shunt impedance of any oscillatory circuit shunting the input is low or by making the magnitude of the total negative input conductance of the tube small in one of the following ways:

1. By using tubes of low transconductance (see (12)). This method is in general impractical because high transconductance is desirable in cathode-follower amplifiers for other reasons.

2. By using a low value of cathode load resistance (see (12)). This method is feasible in some applications of cathode-follower amplifiers.

3. By using low capacitance across the cathode load resistor (see Fig. 4). The load capacitance cannot, however, be reduced below the sum of the plate-cathode capacitance and the minimum circuit-wiring capacitance.

4. By using a sufficiently low value of \( r_e \), so that the positive conductance resulting from \( r_e \) is equal to or nearly equal to the negative conductance resulting from \( C_{\phi h} \). The principal objection to this method is that it increases the conductance at low frequencies to an excessively large positive value.

5. By using resistance in series with the grid. Ordinarily a series resistance of less than 100 ohms is sufficient to prevent oscillation. This value is small enough so that frequency distortion resulting from voltage drop in the series resistance at high frequency is negligible. This method is obviously the most feasible.

Oscillation may also be prevented by insuring that the sum of the tube input susceptance and the susceptance of the input circuit is not zero, i.e., that resonance does not occur, in the frequency range in which the input conductance of the tube is negative. Since Fig. 6 shows that the magnitude of negative input conductance increases with frequency, it is apparent that this can be accomplished only by lowering the resonance frequency of the input circuit.

**An Exponential Transmission Line Employing Straight Conductors**

WILBUR NORMAN CHRISTIANSEN†

**Summary**—An exponential transmission line is useful for impedance transformations over a wide band of radio frequencies.

It is shown that a four-wire line of the "side-connected" type employing uniform conductors, and in which the rates of taper change only once along the line, may be designed to approximate closely to an exponential line.

Design equations and charts are given which aid in determining the wire sizes, values of taper, and change in taper for building some of these transformers.

**I. INTRODUCTION**

In recent years the increasing use in short-wave radio communication of semiaperiodic antennas, principally of the rhombic type, has resulted in attention being given to impedance-transforming devices useful over a large range of radio frequencies.

Wide-band transformers can now be constructed to achieve any normally required impedance transformation in the useful frequency range of the antennas. These transformers are designed to have a high coefficient of coupling between primary and secondary windings, and this makes difficult their application where high radio-frequency voltages are present. Hence broad-band transformers using closely coupled coils are normally not used with transmitting antennas.

The useful impedance-transforming properties of the exponential horn in acoustics suggested† the analogous application of the principle to electrical transmission lines. Various types§ of open-conductor exponential

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II. Electrical Characteristics of an Exponential Line

An exponential transmission line has the capacitance and inductance per unit length of line varying in a manner such that

\[ Z_0z = Z_{00} e^{\delta x} \]  

where \( x \) is the distance from the origin of the point considered, \( Z_0z \) is the "nominal characteristic impedance" of the line at a point \( x \) along its length, and \( Z_{00} \) is the nominal characteristic impedance at \( x = 0 \), i.e., at the end of the line. (The "nominal characteristic impedance" is equal to the characteristic impedance of a uniform transmission line having the same dimensions as the variable line at the point considered.)

From (1) it follows that

\[ \delta = \log_e \frac{Z_{0z}}{Z_{00}} \]

where \( l \) is the physical length of the line and \( Z_{0z} \) refers to the remote end of the line. It has been demonstrated\(^4\) that such a line behaves as an impedance-transforming high-pass filter with a cut-off frequency \( f_c \), given by

\[ f_c = \frac{1}{2\pi} \cdot \frac{\delta}{v} = \frac{v \log_e \frac{Z_{0z}}{Z_{00}}}{4\pi l} \]

where \( v \) is the phase velocity of wave propagation along the line for very high frequencies, i.e., for frequencies where the change in \( Z_{0z} \) per wavelength is very small. It may be noted that \( v \) is assumed to be constant along the length of the exponential line. This implies that the line has low dissipation and unchanging dielectric.


For an open-wire line, (3) becomes approximately

\[ f_c = \frac{55.0 \log_{10} Z_{0z}/Z_{00}}{l} \text{ megacycles} \]

where the unit of length is the meter. The useful impedance-transforming property of the exponential line appears as follows. If one end of the line is terminated in a load equal to \( Z_{0z} \), then for frequencies much greater than \( f_c \) the driving-point impedance at the input to the line approximates very closely to \( Z_{0z} \), the nominal characteristic impedance at the input end of the line. As the frequency is decreased towards \( f_c \) increasing deviations occur in the driving-point impedance from the value of \( Z_{0z} \).

From the analysis of Burrows\(^5\) it may be shown that the frequencies at which these deviations occur are related mainly to the length of the line, while the magnitude of the deviations depends on the rate of line taper. For the line terminated with a resistance equal to \( Z_{0z} \), it was shown by Burrows that the ratio of the driving point impedance \( Z_{0z}' \), at the input to the line to the nominal characteristic impedance at that point is

\[ \frac{Z_{0z}'}{Z_{0z}} = \frac{1 + \sqrt{1 - v^2} - j \nu - (1 - \sqrt{1 - v^2} - j \nu) e^{-2\pi \sqrt{1 - v^2} f}}{1 + \sqrt{1 - v^2} + j \nu - (1 - \sqrt{1 - v^2} + j \nu) e^{-2\pi \sqrt{1 - v^2} f}} \]

where \( \nu = f_c / f \) and is less than unity, \( \gamma \) the propagation constant of the line at frequencies very large compared with \( f_c \) and for the line considered is approximately equal to \( j\beta \), \( \beta \) being the phase-change coefficient, at such high frequencies.

On putting the exponentials into the trigonometric form, we obtain

\[ Z_{0z}' = 1 + 
\]

When \( \nu = 0 \), \( Z_{0z}' / Z_{0z} = 1 \).

When

\[ \sin 2\pi \sqrt{1 - v^2} = 0, \]

i.e., \( 2\pi \sqrt{1 - v^2} = n\pi, \) \( n \) being an integer, \( \langle \rangle \)

\[ Z_{0z}' / Z_{0z} \]

is equal to unity or \( (1 - j\nu)/(1 + j\nu) \), depending on whether \( n \) is even or odd. In either case \( |Z_{0z}' / Z_{0z}| \) is equal to one.

The magnitude of the input impedance is, therefore, equal to the nominal characteristic impedance of the line at the input for frequencies

\[ f = \left\{ \frac{(n\pi/4)}{2} + f_c \right\}^{1/2}. \]

It approaches unity also for all values of \( \beta l \) as \( \nu \) approaches zero, i.e., as \( f \) approaches infinity.

If \( f \) is large compared with \( f_c \), (8) becomes approximately

\[ f = n/4 \cdot v/l; \]

i.e., frequencies \( f \) are those for which the line is an integral number of quarter waves in length.
If \( f \) is large compared with \( f_0 \), we may use the approximation \( \sqrt{1 - \frac{f}{f_0}} \approx 1 \), and (6) then becomes
\[
\frac{Z_{0'}}{Z_{00}} = \frac{2 + \nu \sin 2\beta l - j\nu(1 - \cos 2\beta l)}{2 - \nu \sin 2\beta l + j\nu(1 - \cos 2\beta l)}
\]
(10)
and
\[
\left| \frac{Z_{0'}}{Z_{00}} \right| = \left\{ \frac{2 + \nu^2(1 - \cos 2\beta l) + 2\nu \sin 2\beta l}{2 + \nu^2(1 - \cos 2\beta l) - 2\nu \sin 2\beta l} \right\}^{1/2}
\]
(11)
and this has maximum values of approximately \((1 + \nu)\) when
\[
2\beta l = \frac{4n + 1}{2}\pi,
\]
i.e., when
\[
f = \frac{4n + 1}{8}\frac{\nu}{l}.
\]
Similarly, minimum values of approximately \(1/(1 + \nu)\) occur when
\[
f = \frac{4n - 1}{8}\frac{\nu}{l}.
\]
The above calculations are for \( \delta \) positive, i.e., the impedance \( Z_{0'} \) is considered at the low-impedance end of the line. For impedances at the high-impedance end, \( \delta \) is negative, since the line is then convergent. The sign of \( \nu \) is changed in (3), thereby inverting it. Hence the values of \( Z_{0'}/Z_{00} \) corresponding to the frequencies (12) and (13) are also inverted.

It was shown by Wheeler\(^1\) that if the exponential line is placed between the elements of a half section of a constant-\( K \) low-pass filter, and in addition an \( M \)-derived half section is connected at each end of the system, it is possible to keep the impedance (resistance) deviations within 5 per cent of the required value for all frequencies 15 per cent or more above the cut-off frequency.

Where it is desired to limit the length of the exponential line, or where exact matching is required, such terminating sections have useful application. In many cases, however, it is simpler to use a line of such a length that the working frequency is always very high compared with the cut-off frequency of the line so that a purely resistive line termination may be used.

111. THE DESIGN OF AN OPEN-WIRE EXPONENTIAL LINE FOR A 2-TO-1 IMPEDANCE TRANSFORMATION

(a) Previous Designs

The transformation from 600 to 300 ohms with an open-wire exponential line is made difficult by the fact that the construction of a two-wire line with the latter value of characteristic impedance involves the employment of inconvenient physical dimensions for the line, while if a multiple-wire line is used, the same difficulty is experienced at the high-impedance end.

If a two-wire line is designed to provide such a transformation, the ratio of wire separation \( d \) to the radius \( r \) must change from 150/1 to 12/1 over the length of the line. In many applications of such a line the wire separation cannot be reduced below a value of several inches if mechanical instability and the danger of dielectric breakdown are to be avoided. Hence, for \( d/r = 12 \), tubing rather than wire must be used in the construction of the line.

Burrows\(^2\) constructed a two-wire exponential transmission line in which conductors with large radii were used at the low-impedance end of the line, the conductor radius being reduced at intervals along the line towards the high-impedance end. By this means the conductor spacing was kept at convenient values throughout the length of the line. Small discontinuities existed at the points where the conductor size was altered, but these were not serious as was shown by the fact that Burrows successfully approached the performance predicted by theory for the exponential line.

Another design for an exponential line for use where one feeder line branches into two was suggested by Neiman.\(^3\) In this line the two 600-ohm-line pairs approach each other from a great distance in such a manner that the resultant four-wire line has an exponential characteristic. The pairs theoretically are required to coalesce to form a single 600-ohm pair at the high-impedance end of the line, but in practice a special two-wire section could be used to overcome this. The shaping of the line is done with tension spacers connecting the two pairs of lines. No change in the size of conductors is required with this line, except possibly for the section at the high-impedance end.

(b) Design Employing Straight Conductors

The exponential line to be described here\(^4\) is based on the type of four-wire transmission line in which parallel connections are made between the wires of each adjacent pair, instead of between diagonal wires as in the more commonly used four-wire transmission line. The arrangement of wires is shown in Fig. 1. If it is assumed that the spacings \( a \) and \( b \) are large compared with the wire radius \( r \), and that dissipation in the line is negligible, then the characteristic impedance may be calculated from the expression
\[
Z_0 = \sqrt{\frac{L}{C}} = \frac{1}{\epsilon C} \text{ ohms,}
\]
(14)

Australian Patent No. 121,850.
L and C being respectively the inductance (henries) and capacitance (farads) per centimeter length of line, and 
\[ c = 3 \times 10^6 \text{ centimeters} \] 
The calculation of the capacitance per unit length of line may be done by the method of logarithmic potential to give 
\[ \frac{1}{C} = 2 \log_e \left( \frac{d \sqrt{b^2 + d^2}}{br} \right) \text{ statfarad per centimeter} \]
\[ = 2c^2 \log_e \left( \frac{d \sqrt{b^2 + d^2}}{br} \right) \times 10^6 \text{ farad per centimeter.} \quad (15) \]
Therefore,
\[ Z_0 = 138 \log_e \left( \frac{d \sqrt{b^2 + d^2}}{br} \right) \quad (16) \]

For the production of an exponential line, d and b may both vary with the distance x along the line, while r preferably is fixed.

![Fig. 2—Characteristic impedance of a four-wire (side-connected) transmission line for various values of conductor spacing.](image)

It is obvious that the construction of an exponential line is greatly simplified if b and d can be arranged to vary linearly with x over appreciable ranges of variation of x, without causing Z_0 to depart appreciably from the exponential form. To investigate this, use is made of the graphical construction of Fig. 2, in which Z_0 is plotted on a logarithmic scale against d/r for various values of b/r. (For the ratio of b/r = 5 it might be suspected that equation (16), which was based on the assumption that b and d were both very large compared with r, would no longer be accurate. However, a field plot for this spacing shows that the error in Z_0 resulting from the use of (16) is only about 1 per cent.)

If it is stipulated that d/r is to change linearly with the distance x along the feeder, then a straight line drawn on the graph represents a linear change of log Z_0 with respect to x, or Z_0 = A e^{Bx} (A and B being constants), which represents an exponential transmission line.

If, moreover, a straight line can be drawn on Fig. 2 so that the contours representing equal arithmetic intervals of b/r make equal intercepts on it, then the straight line on the graph represents an exponential transmission line in which b/r is related linearly to d/r and hence to x. Such a transmission line, therefore, would be constructed wholly of straight conductors.

It is found that a large number of lines can be drawn to fulfill approximately this condition over part of the impedance range of 300 to 600 ohms. Consideration must be given, however, to the use of the values of d/r and b/r that are convenient in practice. The straight lines drawn on Fig. 2 represent a transformation from 300 to 600 ohms done in two sections, the division being made at the point where \[ Z_{0x} = \sqrt{Z_{00}Z_{01}} = 424 \text{ ohms approximately} \] which is the physical center of the exponential line.

![Fig. 3—Outline diagram of the exponential line described.](image)

The configuration of the line is illustrated in Fig. 3. In Table I are shown the values of d/r and b/r at the ends and center of the exponential line. To illustrate the physical dimensions of a typical line, spacings are given for a line with conductors of No. 10 American Wire Gauge wire (0.102 inch).

<table>
<thead>
<tr>
<th>Distance along feeder</th>
<th>X = x/f</th>
<th>0</th>
<th>0.5</th>
<th>1.0</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal characteristic impedance Z_0</td>
<td>Z_0</td>
<td>300</td>
<td>424</td>
<td>600</td>
</tr>
<tr>
<td>Spacing b/r</td>
<td>d/r</td>
<td>95</td>
<td>185</td>
<td>385</td>
</tr>
<tr>
<td>Spacing b/r</td>
<td>d/r</td>
<td>80</td>
<td>30</td>
<td>6.5</td>
</tr>
<tr>
<td>For No. 10 A.W.G. wire</td>
<td>d</td>
<td>4.8&quot;</td>
<td>9.4&quot;</td>
<td>19.6&quot;</td>
</tr>
<tr>
<td>For No. 10 A.W.G. wire</td>
<td>b</td>
<td>4.1&quot;</td>
<td>1.53&quot;</td>
<td>0.33&quot;</td>
</tr>
</tbody>
</table>

In the exponential line represented by the straight lines on Fig. 2, b/r changes almost linearly. Hence we may approximate it with a feeder line in which b/r is actually linear, i.e., we may use straight conductors between the center and each end of the line. That the impedance variation along the resultant line follows very closely the exponential form is demonstrated in Fig. 4. The straight line represents an exponential change of impedance along the transmission line, while the points adjacent to the line are those calculated for a line having the form described above. The maximum departure of the impedance level along the "straight-wire" line from the corresponding values for a true exponential line is only one part in a hundred.

The exponential line described above requires to be supported only at the ends and the center. In a practical design, of course, it may be found necessary to use a greater number of supports.
Consider the design of a line to be used for impedance transformation in the range of frequencies from 6 to 20 megacycles. If terminating filter sections of the type indicated by Wheeler are used, then 5 megacycles may be chosen as the cut-off frequency for the line. From (4) we find that \( l = 3.3 \) meters, approximately. It should be noted, however, that with such a short length of line the size of conductors would have to be small; otherwise the conductor spacings would become comparable with the length of the line, and the transmission-line equations would no longer apply.

![Fig. 4—Variation of nominal characteristic impedance along an ideal exponential line (full line) and "straight-wire exponential line" (small circles).](image)

If it is not desired to use filter sections at the line terminations, then the cut-off frequency must be made considerably lower than the lowest working frequency. Where impedance deviations of 10 per cent from the mean can be tolerated, then from (12) and (13) we see that for the above range of frequencies 0.6 megacycle may be taken as the frequency of cut-off. This gives \( l = 27.6 \) meters, approximately.

IV. EXPERIMENTAL LINE

A line of the type described above was constructed to allow an experimental check to be made of the system. The length of the line was 40 meters and the conductors used were of No. 12 Standard Wire Gauge (approximately No. 10 Brown and Sharpe) copper wire. Supports were spaced at 22-foot intervals, this being the spacing of poles used in associated uniform transmission lines. The insulated supports (see Fig. 5) were constructed to make possible easy adjustment of the wire spacings \( d \) and \( b \). (The insulators shown in the photograph were not designed for horizontal mounting, but no better type was available when the line was being built.)

Short lengths of 300- and 600-ohm uniform transmission line were attached to the respective ends of the line.

For this line the cut-off frequency given by (4) is 0.415 megacycle. If the appropriate values for \( l \) and \( f_c \) are substituted in the expressions (9), (12), and (13), then the significant points on a curve of input impedance versus frequency are obtained. The calculated variation of input impedance of the line with frequency is shown in Fig. 6.

Measurements of input impedance were made at each end of the line, the remote end in each case being terminated with a carbon resistor of appropriate value.

![Fig. 5—Typical feeder pole used with an experimental line. Both horizontal and vertical separations of wires are variable.](image)

For the measurements, a portable impedance meter was employed, this unit being composed of a radio-frequency oscillator, balanced amplifier, and tuned output circuit. Across the latter circuit is connected a pair of diodes, the rectified current from which actuates a

![Fig. 6—Theoretical relationship between input impedance and frequency for an exponential line when a resistive termination is used. Length of the line is 40 meters, and the line is designed for a 2-to-1 impedance transformation. Curve 1 is for the high-impedance end of line; curve 2 for the low-impedance end.](image)
with an appropriate fixed resistor (of magnitude greater than the impedance to be measured) connected across the output circuit, adjusting the output from the oscillator until a full-scale deflection is seen on the diode-current meter. The fixed resistor is then replaced by the load to be measured. The reactive component of the latter (in terms of shunt reactance) is calculated from the change in capacitance required to tune the circuit when the external impedance is connected. The equivalent shunt resistance of the load is then read directly from the calibrated scale on the diode-current meter.

In Fig. 7 are shown the measured values of input impedance of the line. The impedances shown are resistive, since in all measurements the reactive component was small enough to be neglected. It is seen that the maximum deviation from the required transformation is 10 per cent. The deviations at the higher frequencies could, doubtless, have been reduced if more detailed attention had been given to the line terminations.

It may be noted in Fig. 7 that the short length of uniform transmission line attached to each end of the exponential line has caused the frequencies of maximum and minimum impedance to be displaced slightly from the positions shown in Fig. 6.

V. CONCLUSION

It has been shown that a four-wire line may be designed to provide a very close approximation to an exponential line in the range of impedances from 300 to 600 ohms. Experimental tests have confirmed that a satisfactory impedance transformation may be obtained with such a line. The employment of a single size of conductor throughout, the absence of elaborate shaping, and the convenient physical dimensions, are the useful features of the line. It has particular application to the problem of supplying power to multiple rhombic or other aperiodic antennas.

VI. ACKNOWLEDGMENT

The writer wishes to express his appreciation for the assistance given by the staff of the Rockbank Beam Receiving Station in the construction and testing of the experimental line.

Correspondence

Magnetic-Wire Response

The integral contained in equation (17) of Marvin Camras' paper on magnetic-wire record response in the August, 1946, issue of the PROCEEDINGS OF THE I.R.E. AND WAVES AND ELECTRONS can be found analytically.

Basset gives its value as

$$\int_{0}^{\infty} \frac{\cos nx}{\sqrt{x^2 + 1}} dx = K_0(n),$$

where $K_0$ is the modified Bessel function of the second kind of order zero.\(^1\)

A table of $K_0(n)$ versus $n$ is available in British Association for the Advancement of Science, Mathematical Tables, vol. VI, Cambridge, 1937. A graphical solution of the aforementioned integral is thus unnecessary.

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New York

Demagnetizing Coefficient

In the issue of August, 1946, there was a paper by Mr. Camras\(^2\) whose equation (17) expressed the demagnetizing coefficient $D$ in terms of a certain integral, which he evaluated by graphical methods, presumably because he did not recognize it. In fact, however,

$$\int_{0}^{\infty} \frac{\cos nx}{\sqrt{x^2 + 1}} dx = K_0(n)$$

where $K_0$ is the modified Bessel function of the second kind which has been fairly completely tabulated. Evaluation of $D = \frac{1}{\rho} K_0(n)$ shows that the author's Fig. 5 obtained by graphical methods is of good accuracy.

R. E. BURGESS
National Physical Laboratory
Teddington, Middlesex


NOTICE

The new I.R.E. television standard, "Standards on Television: Methods of Testing Television Transmitters—1947," is now available. The price is $0.75 per copy, including postage to any country.

Orders may be sent to The Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y., enclosing remittance and the address to which copies are to be sent.
Contributors to Proceedings of the I.R.E.

Wilbur Norman Christiansen

Wilbur Norman Christiansen was born on August 9, 1913, at Melbourne, Victoria, Australia. He received the degree of B.S. from the University of Melbourne in 1934 and M.S. in physics in 1935. In 1935 he joined the staff of the Commonwealth X-Ray and Radium Laboratory and also continued research work in the physics department of the University of Melbourne. From 1937 to the present he has been employed as a development engineer in communication engineering in the Research Laboratories of Amalgamated Wireless (Australasia) Ltd., Sydney, and since 1940 has specialized in problems connected with the overseas communication services of the company.

H. D. Hagstrum was born in 1915 at St. Paul, Minnesota. He received the B.E.E. degree in 1935, the B.A. degree in 1936, the M.S. degree in 1939, and the Ph.D. degree in physics in 1940 from the University of Minnesota. From 1936 to 1940 he had a research and teaching assistantship in the physics department at the University of Minnesota.

In 1940 Dr. Hagstrum joined the staff of the Bell Telephone Laboratories. During the war he was engaged in the development of magnetron oscillators for radar use. He is now in the physical research department at Bell Laboratories.

Donald B. Harris (SM’45) was born at Minneapolis, Minnesota, on February 10, 1901. He received the B.A. degree from Yale University in 1922, after completing a physics major. During the school year of 1922 to 1923 he occupied the position of master in physics and mathematics at the Adirondack-Florida School, Onchiota, New York, and Miami, Florida. In 1923 he became associated with the Cutting and Washington Radio Corporation, Minneapolis. He joined the Northwestern Bell Telephone Company in 1924, subsequently occupying various technical and administrative positions. During this period he independently developed a number of electronic devices on which patents were granted or are now pending.

In 1943 Mr. Harris became technical aide of Division 15 of the National Defense Research Committee stationed at Radio Research Laboratory, Harvard University, where he was responsible for the administration of contracts of Division 15 in the Cambridge, Massachusetts, area. He is at present the transmission and protection engineer of the Northwestern Bell Telephone Company, stationed at Des Moines, Iowa.

Herbert J. Reich (A’26–M’41–SM’43) was born on October 25, 1900, at Staten Island, N. Y. He received the M.E. degree from Cornell University in 1924 and the Ph.D. degree in physics in 1928. He was an instructor in machine design at Cornell University during 1924 and 1925; instructor in physics, Cornell University, from 1925 to 1929; assistant professor of electrical engineering, University of Illinois from 1929 to 1936; and associate professor from 1936 to 1939, when he was made a professor. He was on leave at the Radio Research Laboratory at Harvard University from 1944 to 1946, and in 1946 he became professor of electrical engineering at Yale University.
Institute News and Radio Notes

Report of the Secretary—1946

The report of the Secretary for the calendar year 1946 is submitted, in accordance with a requirement of the By-Laws. As usual, important statistics are presented to indicate growth, the state of general activities, fiscal condition, and the distribution of members and Sections geographically.

It is now clear that our Institute's growth over the past few years, both as to size and member activities, has continued without any signs of abatement. This expansion, combined with the completion and occupancy of new permanent headquarters, strengthening of the Board of Directors through the adoption of the Regional Representation Plan, and the further rounding out of headquarters staff organization, signals the achievement of a significant stability as the technical society that uniquely serves the field of radio communication, electronics, and allied activities throughout the world.

The attention of readers is particularly directed to the increased size of the Proceedings and to the increased cost of printing (Fig. 1). The maintenance of an enlarged Proceedings, so necessary to serve the rapidly growing electronic art, poses a problem of major magnitude.

Membership

At the end of the year 1946, the membership of the Institute, including all grades, was 18,154, a 15 per cent increase over the previous year. Before the war, the annual increase in the number of members was less than 5 per cent; in the first three years of the war it was about 25 per cent. In 1944 and 1945, the figure dropped to 20 per cent, and to 15 per cent in 1946. The membership trend from 1912 to date is shown graphically in Fig. 2.

The distribution of members in the various grades for the years 1945 and 1946 is shown in the accompanying plot, Fig. 3. Actual figures are shown in Table I. Note that the percentage of Associates has dropped and that the percentage of Members and Senior Members has increased. The membership ratio (Associates/Higher grades) was 6 to 1 in 1944, 4 to 1 in 1945, and less than 3 to 1 in 1946, a very satisfactory trend.

Table I—Membership Distribution by Grades

<table>
<thead>
<tr>
<th>Grade</th>
<th>As of Dec. 31, 1946</th>
<th>As of Dec. 31, 1945</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number</td>
<td>Per Cent of Total</td>
<td>Number</td>
</tr>
<tr>
<td>Fellow</td>
<td>219</td>
<td>1.2</td>
</tr>
<tr>
<td>Senior Member</td>
<td>1,703</td>
<td>7.7</td>
</tr>
<tr>
<td>Member</td>
<td>2,330</td>
<td>12.6</td>
</tr>
<tr>
<td>Student</td>
<td>21,510*</td>
<td>128.3</td>
</tr>
</tbody>
</table>

Total 18,154 10,779

* Includes 1,701 Voting Associates.
| Includes 2,048 Voting Associates.

Table II shows an analysis for the past five years of the distribution of members at home and abroad. It may be noted that the foreign membership has increased rapidly since the end of the war.

It is with deep regret that this office records the death of the following members of the Institute during the year 1946:

Senior Members

George Edward Cabot (A'16-M'39-SM'43)
Austin M. Curtis (M'13-SM'43)
Paul Franklin Johnson (A'23-M'26-SM'43)
Ralph A. Powers (M'41-SM'43)
Frank Clifford Stockwell (M'38-SM'43)
John Wadhams Watson (A'44-SM'45)

Members

A. Nelson Butz, Jr. (A'42-M'45)
Leonard T. Carlson (A'41-M'46)
Frank M. Davis (M'44)
Karl Stielje (A'39-M'45)

Associates

Howard I. Hickel (A'43)
David L. Bigley (A'45)
Albert Preston Brieden, Jr. (S'43-A'45)
Truman Preston Brewster (A'44)
Albert P. Cartier (A'45)
Stephen Donald Custodero (A'45)
Herhol V. Fitz Charles (A'41)
Allan Jack Hoenig (S'42-A'45)
George Hunt (A'41)
Raymond Hutchens (VA'36)
George Jacobsen (VA'36)
Frank Reginald Lambton (A'44)
Alfred W. Moxon (VA'27)
Russell Louis Nielsen (S'40-A'41)
Tainter Parkinson (VA'26)
Ward Curtiss Priest (VA'26)
Edward Albert Schluter (VA'26)
William James Thomas (A'45)
Philip S. Walsh (A'42)
Ralph A. Webster (VA'37)

Students

William Russell Lach (S'44)

Editorial Department

In 1946 there were published in the Proceedings 1256 editorial pages, 891 of which were technical. This compares with 944 pages in 1945, with 702 technical pages. Accordingly, total editorial pages increased 41 per cent. There were 122 papers published from 160 authors from different organizations, academic institutions, and the military services, as contrasted with 103 papers in 1945, submitted by 138 authors. In 1946, 47 of our authors were nonmembers of the Institute; in 1945 this number was 33. Advertising in 1945 comprised 976 pages and in 1946, 974 pages. The total number of pages (including covers) published during 1946 was 2240; in 1945, 1912. This averaged 187 pages per issue; in 1945, 157.

After many delays occasioned by shortages, printing difficulties, and other problems, the 1946 Yearbook was published in November. This volume consists of 224 pages of editorial material and 188 pages in the advertising section.

Rising prices increased the page-rate cost of the Proceedings from $31.00 in 1945 to $40.00 in 1946, and the rate per copy from $0.29 to $0.36. The high for the year were $43.00 and $0.40, respectively. These figures are print costs, and do not include overhead or salaries.

At the end of 1946, the Proceedings "bank" showed 1072 pages either accepted for publication or in the hands of the readers. This compares with 411 pages at the end of 1945. The Editorial Department, with great hesitancy, found it necessary to return many papers to the authors, asking for condensation, sometimes as much as 60 per cent. This was an absolute necessity in order to preserve much useful material from as many authors as possible.

In January, 1946, the journal became known as "Proceedings of the I.R.E. and Waves and Electrons." Separate contents pages and separate pagination, denoted by the letters P and W, were first used. In May, the separate pagination was discontinued, although the three parts of the Proceedings continued to be indicated as Proceedings of the I.R.E., Institute News and Radio Notes, and Waves and Electrons Section. This form continued through December.

By a special arrangement with Hille and Son, and the Times, London, England, as well as the Department of Scientific and Industrial Research of the British Government, the Institute, in June, began to publish "Abstracts and References." Reprints of these in an edition printed on one side only are available to members on subscription.

In the early fall copy was sent to the printer for "Standards on Television: Methods of Testing Television Transmitters."

The unfortunate illness of the Editor, Dr. Alfred N. Goldsmith, in the early part of the year, worked severe hardship on the Editorial Department, since without his guiding hand, problems were magnified. Early in 1947...
VOLUME OF TECHNICAL AND EDITORIAL MATTER PUBLISHED IN THE PROCEEDINGS 1913 TO 1946

SMALL FORMAT (42% WORD CONTENT OF LARGE FORMAT) LARGE FORMAT (IN USE SINCE 1939)

NUMBER OF MEMBERS 17,500

TREND OF PAID MEMBERSHIP 1912 TO 1946

Figs. 1 and 2
INCOME AND EXPENSE
1914 TO 1946

KEY:

INCOME
EXPENSE
DEFICIT
SURPLUS

ANNUAL SURPLUS
OR DEFICIT
1914 TO 1946

Fig. 4


April, Clinton B. DeSoto accepted the position of Technical Editor, made vacant by the resignation of Ray D. Rettenmeyer. At about this same time, Mary L. Potter became Assistant Editor.

The Editorial Department was comfortably housed in a loft building, having the entire fourth floor at 26 West 58 Street, where it had moved in April of 1945 in order to relieve the congestion at Headquarters. In July, it was necessary to move the Department to the second floor of the same building, where only a half floor was available. In November, the Department again moved, our permanent home at 1 East 79 Street. Attractive and commodious quarters enable efficient and pleasant operation of the Department.

Administratively, in matters involving personnel and fiscal affairs, the Editorial Department has been integrated with Headquarters staff under the direction of the Executive Secretary, George W. Bailey. In matters involving editorial policy, content, format and the like, of all IRE publications, the Editorial Department continues to function under the direction of the Editor, Dr. Alfred N. Goldsmith, with the support and counsel of the Board of Editors, the Papers Review Committee, the Papers Procurement Committee, and the Editorial Administrative Committee.

Fiscal

A condensed comparison of income and expenses for the years 1945 and 1946 is shown in Table III.

<table>
<thead>
<tr>
<th>Income</th>
<th>1945</th>
<th>1946</th>
</tr>
</thead>
<tbody>
<tr>
<td>Membership Dues</td>
<td>$313,716.64</td>
<td>$569,131.18</td>
</tr>
<tr>
<td>Advertising</td>
<td>160,406.41</td>
<td>129,028.85</td>
</tr>
<tr>
<td>Subscriptions</td>
<td>20,006.79</td>
<td>20,333.82</td>
</tr>
<tr>
<td>Others</td>
<td>67,724.14</td>
<td>73,514.57</td>
</tr>
<tr>
<td><strong>Total Income</strong></td>
<td>$520,222.98</td>
<td>$655,284.46</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Expenses</th>
<th>1945</th>
<th>1946</th>
</tr>
</thead>
<tbody>
<tr>
<td>Printing Proceeds</td>
<td>$120,121.68</td>
<td>$50,307.27</td>
</tr>
<tr>
<td>Salaries</td>
<td>108,266.01</td>
<td>79,017.99</td>
</tr>
<tr>
<td>Advertising Commissions</td>
<td>48,407.89</td>
<td>26,059.51</td>
</tr>
<tr>
<td>Others</td>
<td>104,007.14</td>
<td>99,841.21</td>
</tr>
<tr>
<td><strong>Total Expenses</strong></td>
<td>$377,802.82</td>
<td>$251,205.97</td>
</tr>
<tr>
<td><strong>Net Income</strong></td>
<td>$ 4,200.16</td>
<td>$15,078.49</td>
</tr>
</tbody>
</table>

Section Activities

We were glad to welcome five new Sections into the Institute during the past year. They are as follows:

<table>
<thead>
<tr>
<th>City</th>
<th>Month</th>
<th>Year</th>
</tr>
</thead>
<tbody>
<tr>
<td>Columbus</td>
<td>Jan.</td>
<td>1946</td>
</tr>
<tr>
<td>Houston</td>
<td>Feb.</td>
<td>1946</td>
</tr>
<tr>
<td>Milwaukee</td>
<td>Feb.</td>
<td>1946</td>
</tr>
<tr>
<td>N.C.-Va.</td>
<td>June</td>
<td>1946</td>
</tr>
<tr>
<td>Syracuse</td>
<td>Dec.</td>
<td>1946</td>
</tr>
</tbody>
</table>

The total number of Sections is now 38. There has been a substantial increase in membership of these Sections, with a few exceptions where there has been a slight decrease. In addition, during the year there

has been formed the following group, unofficially designated as a Sub-Section:

Winnipeg Sub-Section (Toronto)

Technical Activities

The Winter Technical Meeting of the Institute was held in January, 1946, in the Hotel Astor, in New York City. There were 7020 persons actually registered, and 87 technical papers were presented during the meeting. The unexpected large attendance overtaxed the facilities of the hotel. Provision will be made in 1947 for commodious quarters. An interesting feature of the meeting was the exhibition of products by a large number of manufacturers, many being exhibited for the first time since the veil of wartime secrecy had been lifted.

During the year, the Technical Committees showed increased activity, and preparations are being made for a very active 1946-1947 winter season.

Our commodious new quarters at 1 East 79 Street are ideal for a meeting place for our Technical Committees.

We were reluctant to accept the resignation of Dr. William H. Crew, but realized that in so doing he was accepting a position of large responsibility as Assistant Dean of Rensselaer Polytechnic Institute in Troy, New York.

We are fortunate in having secured the services of Laurence G. Cumming as Technical Secretary. Mr. Cumming was formerly a commander in the Office of Naval Research and took active part in important technical electronic developments during the war. He enjoys a wide acquaintance among radio engineers and scientists.

The Institute's New Home

In November, 1946, the Institute staff moved into our new home at 1 East 79 Street. The Institute owes a debt of gratitude to its friends and members who generously contributed to the Building Fund, which made the new home possible, and to the Office Executive Committee, which has so successfully prepared the building for occupancy. All the departments of the Institute are now under one roof with room for expansion. Under such ideal working conditions, the staff is prepared to function smoothly in serving members.

Respectfully submitted,

Haraden Pratt
Editor

April 9, 1947

A.I.E.E. Pacific General Meeting

An entire division of the 1947 Pacific General Meeting of the American Institute of Electrical Engineers, at Hotel San Diego, San Diego, California, August 26-29, 1947, will be devoted to communications.

N. E. C. Proceedings Available

Volume 2 of the Proceedings of the National Electronics Conference (1946) has been published recently. While they are still available, copies of Volumes 1 and 2 may be obtained from R. E. Beam, Secretary (for 1947), c/o Electrical Engineering Department, Northwestern University, Evanston, Illinois. Price, $3.00 per copy for Volume 1 and $3.50 per copy for Volume 2.

CONNECTICUT VALLEY SECTION ANNUAL MEETING

The Connecticut Valley Section will hold its annual meeting at New London, Connecticut, Saturday, June 7. There will be a symposium on "Frequency Modulation Receivers" by several speakers, each giving a short talk describing his company's developments. The technical session in the morning will be followed by a business meeting and election of officers for the ensuing year. After a luncheon, there will be an inspection trip to the Submarine Base.

NATIONAL ELECTRONICS CONFERENCE OFFICERS

The National Electronics Conference, Inc., has released a list of the officers who have been elected to serve for the coming year. This corporation, whose purpose is to serve as a national forum on electronic developments and their application, is sponsored jointly by Illinois Institute of Technology, Northwestern University, American Institute of Electrical Engineers, Institute of Radio Engineers, and the University of Illinois, with the Chicago Technical Societies Council a co-operating organization.

Among the new officers of the corporation, the following are members of The Institute of Radio Engineers:

Chairman of the Board of Directors—W. O. Swinnyard (A'37-M'39-SM'43-F'45), Hazeltine Electronics, Inc.

President—A. B. Bronwell (A'39-SM'43), Northwestern University, Evanston, Illinois

Executive Vice-President—W. L. Everitt, (A'25-M'29-F'38), University of Illinois

Vice-President in Charge of Program—G. H. Fett (SM'45), University of Illinois

Vice-President in Charge of Publicity—H. S. Renne (A'41-M'45), Radio-Electronic Engineering Co.

Secretary—R. E. Beam (S'37-A'41-SM'44), Northwestern University

Treasurer—A. H. Schulz (A'38-SM'46), Armour Research Foundation

Chairman, Exhibits Committee—O. D. Westerberg (A'45), Commonwealth Edison Co.

Chairman, Hotel Committee—R. J. Donaldson (A'36), Commonwealth Edison Co.

Plans are now being formulated for the 1947 National Electronics Conference, which will be held on November 3, 4, and 5 at the Edgewater Beach Hotel, Chicago. An exceptionally fine program of technical papers will be presented, and numerous exhibits of electronic equipment are being planned.
APPROXIMATELY 600 guests registered for the one-day Chicago I.R.E. conference held under the sponsorship of the Chicago Section of The Institute of Radio Engineers at Northwestern University Technological Institute in Evanston, Illinois, on April 19.

The meeting was opened by Alois W. Graf, Chairman of the Chicago Section of the I.R.E. The welcoming address was delivered by Dr. O. W. Eschbach, Dean of the Technological Institute; Dr. W. R. G. Baker, President of I.R.E., delivered the opening address.

The morning was then divided into three concurrent sections:

**Radio Receivers**

**Chairman:** Hugh S. Knowles, Jensen Manufacturing Company, Chicago, Ill.


   A compact metallic ladder of mechanically resonant elements, linked by compliant members and coupled to electrical circuits by magnetostriuctive terminations, transmits uniformly within a 4- to 14-kilocycle band with very rapid attenuation outside. Data for design and performance of this filter in a receiver were given. It is adapted to economical production. A demonstration was included.


   The characteristics of a perfect phonograph pickup were outlined. Upon the basis of this theoretical unit, a practical unit is derived, including only such compromises as are required by the state of the art and considerations of a proposed medium price. The pickup thus derived includes a "bimorph" crystal of the new ammonium-phosphate type, rigidly lever-coupled to a sapphire stylus, and all floating freely in a viscous gel.

**Electronics**

**Chairman:** A. B. Bronwell, Northwestern University, Evanston, Ill.


   The elements of a control system were outlined and electronic systems were emphasized. Sensing devices, servomotors, and computers were described in detail.

2. "Comparison and Use of Photosensitive Devices," H. S. Snyder, Northwestern University, Evanston, Ill.

   A general description and introduction to the subject, including a qualitative discussion of photosensitive devices, classification of various types, and notes on uses and limitations.

**Engineering Profession**

**Chairman:** A. Crossley, Consulting Engineer, Chicago, Ill.

1. "Patents and the Engineer," Curtis F. Prangley; Moore, Olson and Trexler, Chicago, Ill.

   An analysis of the engineer's relation to the patent system and of the keeping of records by engineers to facilitate the

   securement and enforcement of patents.


   A review of the current situation in industrial research, pointing out the reasons for increased emphasis in both industrial and fundamental research today, the opportunities which exist in the industrial research field, and industrial research as a national asset; a discussion of the opportunities for scientific men in research and the requirements for capable research men in industrial research laboratories; suggestions regarding the responsibility of industry to assist in the training and preparation of those men, and one or two definite suggestions for a training program.

A buffet-style luncheon was served in two sections at 12:30 and again at 1:15 in the gymnasium. The afternoon sessions were covered in three panels:

**Radio Receivers**

**Chairman:** Nelson P. Case, The Hallicrafters Company, Chicago, Ill.


   The objectives to be reached in television production testing were outlined and related to the general provisions of the Farnsworth plan. Details of this plan were presented, as well as reasons for rejection of other possible methods.


   Practical design information was discussed on linear permeability-tuned oscillators with a demonstration of their application in the Collins 75A receiver and 310B and 310C exciter units.


   Several receiving-type tubes may be used to advantage in television receivers designed to tune all 13 channels. The performance of these tubes in radio-frequency amplifiers, mixers, and local oscillators was discussed. Both push-pull circuits and single-ended circuits were treated. Data were presented for over-all gain, noise, image rejection, and oscillator frequency stability. These data were taken at two respective points in the band: 60 and 200 megacycles.

**Communications Equipment**

**Chairman:** D. E. Noble, Galvin Manufacturing Corporation, Chicago, Ill.

1947

Institute News and Radio Notes

ON STAGE FOR THE OPENING SESSION

Left to right: Alois W. Graf, Chairman, Chicago Section, I.R.E.; Dr. Ovid W. Eschbach, Dean of the Technological Institute, Northwestern University; Dr. W. R. G. Baker, National President, I.R.E.; Robert E. Samuelson, Executive Chairman, Chicago I.R.E. Conference; William Jarzembski, President, Electro Tech Society at Northwestern University.

Three types of mobile common-carrier service are: urban, highway, and marine. Mobile systems were discussed in general, with special emphasis on their connection with telephone lines. Coverage and other results were summarized.

2. "A Variable-Frequency Oscillator with Narrow Band FM," L. A. Mayberry, The Hallcrafters Company, Chicago, Ill. The design of a low-power variable-frequency exciter unit to be used for amateur radio communications. The device provides three types of output: variable-frequency continuous-wave, crystal-controlled continuous-wave, and the variable-frequency frequency-modulation phone. Particular attention was given to the properties of narrow-band frequency-modulation and to frequency stability.

3. "Personal Plane Radio," D. H. Mitchell, Galvin Manufacturing Corporation, Chicago, Ill. Radio manufacturers have done an outstanding job of solving the problems of private aircraft radio during the last year. Solution of these problems by economical application of radio techniques common to the broadcast field were discussed. There have been more radios manufactured and made available for the post-war airplane than any other accessory. The engineer's most difficult problem is to convince the private flyer what he needs in aircraft radio and why.

BROADCAST EQUIPMENT

Chairman: W. E. Phillips, Raytheon Manufacturing Company, Chicago, Ill.


The new frequency-modulation broadcast transmitters require a precision monitor for performance tests. For the past two years development of such a monitor has been under way in the General Radio Company laboratories. The design problems that were involved and details of their solution were described.


The general problem of the generation of relatively large amounts of video-modulated radio-frequency power was reviewed, and the need for a high-efficiency modulating system was shown. The design of such a transmitter was illustrated by description of the Zenith television transmitter, based on the Doherty-Terman system of high-efficiency grid modulation. Approximately 4 kilowatts peak power output is obtained, using four type 425OA tetrodes in an unconventional grounded-grid version of the Doherty-Terman circuit.

There were 22 manufacturers who had educational exhibits open throughout the day.

Inspection trips of a "typical" lecture room, and electrical and mechanical engineering laboratories, were arranged at twenty-minute intervals and conducted in groups of about twenty-five guests throughout the entire day. These trips were conducted by Northwestern University students, members of the Electro Tech Society.

There were five continuous laboratory demonstrations supervised by students, including an electron microscope, electronic control and servomechanisms, a high-voltage generator, microwave optics, and a laboratory display of circuit analogies.

The Chicago Section of The Institute of Radio Engineers is greatly indebted to Dr. O. W. Eschbach, dean of Northwestern Technological Institute; Dr. John F. Calvert, chairman of the Electrical Engineering Dept., Northwestern Technological Institute; and Dr. Robert E. Beam, adviser for the Student Section A.I.E.E.-I.R.E. at Northwestern University, for their splendid co-operation and active assistance in planning and carrying out this conference.

The Chicago Section wishes to thank the Electro Tech Society (Northwestern University Student Section of the A.I.E.E.-I.R.E.) for its active participation and help in making the conference a success.

The conference was another successful activity of the Chicago Section, and a similar session is planned for next year.
I.R.E. People

VLADIMIR K. ZWORYKIN

Vladimir K. Zworykin (M'30–F'38), recently elected vice-president and technical consultant of the RCA Laboratories Division, Radio Corporation of America, was awarded the Howard N. Potts medal by the Franklin Institute for his invention of the iconoscope and kinescope, which are essential to modern commercial television.

Born in Mouron, Russia, in 1889, Dr. Zworykin received his E.E. degree from the Petrograd Institute of Technology in 1912. Continuing there in research work under Professor Boris Rosing, he started his first experiments in television; later he spent two years in X-ray research at the College de France under Professor P. Langevin. In 1920 he came to the United States and joined the research staff of Westinghouse Electric and Manufacturing Company at Pittsburgh, attending at the same time the University of Pittsburgh from which he received his Ph.D. degree in 1926. He became an American citizen in 1924.

Dr. Zworykin joined the Radio Corporation of America in 1930 as director of their Electronic Research Laboratory, where his activities have covered many phases of radio and electronics. For his research work in the development of the electron microscope, he received the Rumford award of the American Academy of Arts and Sciences in 1941.

In World War II he performed distinguished service as a member of the Scientific Advisory Board to the Commanding General of the United States Army Air Forces, the Ordnance Advisory Committee on Guided Missiles, and three subcommittees of the National Defense Research Committee, and directed important research work.

Dr. Zworykin received the Morris Liebmann Memorial Prize in 1934 from The Institute of Radio Engineers; the Overseas Premium of the British Institution of Electrical Engineers in 1937; the honorary degree of Doctor of Science from the Brooklyn Polytechnic Institute in 1938; and in 1940 the Modern Pioneers Award of the American Manufacturers Association.

He is the co-author of "Photocells and Their Application," "Television," and "Electronic Optics and the Electron Microscope." In January, 1947, he disclosed at a joint meeting of the American Meteorological Society and the Institute of Aeronautical Sciences that he is directing work on an electronic calculator which may possibly accurate weather prediction and control.

Dr. Zworykin is a member of the American Association for the Advancement of Science, the American Physical Society, the American Institute of Electrical Engineers, American Academy of Arts and Sciences, National Academy of Sciences, Franklin Institute, the French Academy of Sciences, and Sigma Xi.

ROBERT D. TEASDALE

Robert D. Teasdale (S'45–A'46) has been awarded a Gerard Swope Fellowship for the academic year 1947–48. His problem involves analytical research on high-frequency wave propagation with particular emphasis on boundary value problems. A graduate of Carnegie Institute of Technology, where he held a George Westinghouse Scholarship, Mr. Swope is instructing and studying for his M.S. at the Illinois Institute of Technology in Chicago. He is an associate member of the American Institute of Electrical Engineers, and a member of Tau Beta Pi, Eta Kappa Nu, Phi Kappa Phi, and Pi Delta Epsilon.

MERRILL A. TRAINER

Merrill A. Trainer (A'35) was recently appointed manager of television equipment sales for the Radio Corporation of America. Prior to his appointment, he was in charge of television terminal-equipment development where he supervised the company's development of airborne-television equipment and television-guided missiles.

Born in Philadelphia on July 25, 1905, Mr. Trainer received the B.S. degree in electrical engineering from the Drexel Institute of Technology in 1927. Upon graduation, he became associated with E. W. Alexander in television research at the General Electric Company. Since 1930 he has been on the RCA television engineering staff.

Mr. Trainer is a member of Tau Beta Pi, and has served on several I.R.A. committees.

Otto H. Schmitt

Otto H. Schmitt (SM'44) returned to the University of Minnesota on April 1, after a five-year wartime leave of absence with the Airborne Instruments Laboratory. He resumes his position as associate professor in the departments of physics and zoology where a major portion of his efforts will be devoted to research in biophysics. Dr. Schmitt played a leading part in the development of the magnetic airborne detector, a wartime-developed equipment for locating submerged enemy submarines from the air, and in the designing, building, and operating of an aircraft-antenna radiation-pattern-measuring system employing the airplane-modeling method.
Minutes of Technical Committee Meetings

The following brief abstracts of I.R.E. technical committee minutes are intended to keep the membership informed as to the activities of such groups. Members having views or proposals of interest to the committees, or desiring possibly available information from them, should write directly to the chairman of the particular committee, sending a copy of the letter to Mr. Lawrence C. Cumming, Technical Secretary, The Institute of Radio Engineers, 1 East 79 Street, New York 21, N.Y.—The Editor.

Electron Tube Conference

Date: March 3, 1947
Place: Hotel Commodore, New York, N.Y.
Chairman: E. W. O'Neil

Present

G. W. O'Neil, Chairman

L. Malte
E. D. McArthur
J. A. Morton
A. L. Samuel

Syracuse University was accepted as the meeting place of the Conference. Mr. McArthur was empowered to organize a local group at Syracuse, which will take care of all necessary arrangements. Professor Reich was asked to replace Mr. McArthur as chairman of the entertainment committee. The technical program of the Conference was discussed and additions made.

Radio Transmitters

Date: March 3, 1947
Place: Hotel Commodore, New York, N.Y.
Chairman: E. A. Laport

Present

E. A. Laport, Chairman

Cledo Brunetti
H. R. Butler
A. E. Kerwin

The subcommittee structure of the Committee was reviewed in the light of the decision of the previous meeting concerning the scope of work of the Committee, and it was decided that four subcommittees, already existed, would suffice for the present program. Dr. Brunetti read the list of terms which his group presently has in process of definitions. All but about six terms were accepted for inclusion in the list for Transmitters. In order to divide the work of defining other terms, I. R. Weir, J. E. Young, and Robert Surrell, as subcommittee chairman, will select from the catalog of terms issued in January those which are most properly related to their special activities, and overlaps can be eliminated at the next meeting.

Circuits

Date: March 3, 1947
Place: Hotel Commodore, New York, N.Y.
Chairman: E. A. Guillemin

Present

E. A. Guillemin, Chairman

H. W. Bode
J. G. Brauerd
Cledo Brunetti
C. R. Burrows
W. L. Everitt
L. A. Kelley
W. N. Tuttle

The subcommittee of which Mr. Neitzert is chairman will continue under the new chairmanship of Professor J. B. Russell (Mr. Neitzert dropping out) and with the addition of Messrs. E. H. Perkins, O. J. Zobel, and R. M. Foster as soon as the latter can be added to the membership of the Circuits Committee. Mr. Brunetti pointed out that other committees whose work overlaps that of this one may give better attention to the arrangements unless the Committee announces its desire to pass upon their tentative results. This matter will be looked into by Mr. Brunetti and he will prepare a list of items upon which the Committee will prefer to pass judgment before their final adoption by the Standards Committee. It was also pointed out that the Committee must concern itself with the work of the A.S.A. and that of the Symbols Committee insofar as overlapping items are concerned.

Navigation Aids

Date: March 3, 1947
Place: Hotel Commodore, New York, N.Y.
Chairman: D. G. Fink

Present

D. G. Fink, Chairman

H. G. Busingnies
L. G. Cumming
C. J. Hiri
D. R. MacQuivey
H. R. Minno

The Chairman stated that he had written to General Rives and General Arnold requesting representation from the Army Air Force Material Command and from the Air Transport Association, respectively, but thus far had not received replies. He reviewed the Committee's functions and stated that it would reserve the right to make definitions on radio terms, but that those of navigation terms will be passed on to the Radio Technical Committee for Aviation and the Radio Technical Committee for Marine, Institute of Navigation, and Institute of Aero- nautical Sciences for comments. The report of the subcommittee was read and detailed comments were made on the work of the various sections. Professors Minno and Pierce were again nominated to make suitable corrections in the definitions. It was agreed that a classification covering the electronic technique for navigation might constitute a valuable addition, and the secretary was asked to attempt these definitions.

Standards

Date: March 3, 1947
Place: Hotel Commodore, New York, N.Y.
Chairman: A. Chamberlain

Present

A. B. Chamberlain, Chairman

R. S. Burnap
I. S. Coggeshall
M. G. Crosby
L. G. Cumming
Eginhard Dietze
R. F. Guy

The best method of presenting standards for practical use in the field was discussed. Mr. Laport, Chairman of the Transmitter Committee, strongly recommended a thorough study of the activities of similar technical organizations, aimed toward the reduction of duplicating man-hours in the preparation of Definitions and drafting of Standards for Test Methods. Mr. Laport briefly described his chart, which indicates the relationship of I.R.E.'s Technical Committees to electronic and communication systems. Mr. McIwain recommended a joint A.I.E.E.-I.R.E. subcommittee for the segregation of definitions common to both organizations. Such a subcommittee has been formed and held its first meeting on the 27th of March, under the chairmanship of Mr. A. B. Chamberlain, the Chairman of the Standards Committee. Mr. Burnap said that the American Standards Association, under the guidance of its Mr. Chester Dawes at Harvard, is increasing its liaison with A.I.E.E., I.R.E., and similar organizations. In contrast to this, Mr. Dietze advised that the American Acoustical Society is now setting up own standards committee of six or seven members, Mr. Chamberlain advised that suggested revisions to the Standards Committee Manual currently in progress would be made available to all members of the Standards Committee.

Railroad and Vehicular Communication

Date: March 4, 1947
Place: Hotel Commodore, New York, N.Y.
Chairman: D. E. Noble

Present

D. E. Noble, Chairman

A. E. Abel
G. H. Phelps
G. M. Brown
F. M. Ryan
F. T. Baudeman
W. W. Salisbury
W. A. Harris
W. J. Young
W. G. Hawkins
W. R. Young

In order to clarify the position of the various committees dealing with communication service, Mr. Noble gave a brief description of their respective functions. The main object of the meeting was to formulate definitions and methods of measurement for various transmitter and receiver characteristics. Mr. Noble suggested that the definitions and methods be distributed to individual committee members for concentrated work and presented to the entire committee for approval at the next meeting. Desirable methods of writing the various definitions and test methods were discussed.
PROCEEDINGS OF THE I.R.E.

Date........ March 4, 1947
Place.......Hotel Commodore, New York, N.Y.
Chairman...P. S. Carter

Present
P. S. Carter, Chairman
G. H. Brown W. E. Kock
L. G. Cumming D. C. Ports
Sidney Frankel J. C. Schelleng
R. F. Holz M. W. Scheldorf
W. J. E. Young George Sinclair
R. B. Jacques S. A. Schelkunoff
E. C. Jordan P. H. Smith
L. C. Van Atta

Dr. George Sinclair submitted a definition of "Echoing Area." The members were unable to reach agreement concerning this term. A subcommittee of Doctors Van Atta, Schelkunoff, and Kock then worked up a definition under the title "Back Scattering Coefficient" and this was approved. This completes the list of definitions for the present. Since no objections were raised by any members of the Committee, the list will be forwarded to the Standards Committee shortly. The subject of transmission lines was brought up and Dr. Schelkunoff, Chairman of the Wave Propagation Committee, stated that he considered this work to come within the scope of his Committee. Dr. Sinclair submitted a revised edition of "Methods of Testing." Dr. G. H. Brown raised questions concerning the accuracy of the mutual-impedance measuring method suggested. Because of lack of time it was agreed that all comments would be submitted to Dr. Sinclair within a month. At the end of that period, the report will be considered approved and will be submitted to the Standards Committee. It was agreed that no more meetings of this Committee will be held until it has been advised of the action taken by the Standards Committee concerning the work so far completed.

RADIO WAVES Propagation AND Utilization

Date........ March 4, 1947
Place.......Hotel Commodore, New York, N.Y.
Chairman...S. A. Schelkunoff

Present
S. A. Schelkunoff, Chairman
S. L. Bailey A. G. Fox
C. R. Burrows D. E. Kerr
T. J. Carroll H. O. Peterson
A. E. Culum J. A. Pierce
L. G. Cumming George Sinclair
H. W. Wells

Dr. Schelkunoff requested the Committee's opinion on how frequently meetings should be held during the period in which definitions are being prepared. It was agreed that meetings of the full Committee would be held only when decisions are to be made. There was extensive discussion of the criteria which should govern the preparation of the definition of terms. It was generally agreed that the present program will be limited to terms concerned primarily with wave propagation: that they should be terms that are useful and commonly employed in more than one sense, or new and not sufficiently well defined in the literature. Controversial terms should be considered particularly. The Committee will recommend to the I.R.E. Standards Committee that the rationalized meter-kilogram-second-coulomb system of units be adopted, and that the Standards Committee take the steps necessary to ensure adoption by the I.R.E. This Committee will prepare and submit for publication in the PROCEEDINGS OF THE I.R.E. a paper explaining the rationalized meter-kilogram-second-coulomb system, and discussing its advantages. Dr. Bailey suggested that testing methods be placed on the schedule for discussion at future meetings, with the purpose of deciding to what extent the Committee should prepare definitions or recommend test procedures.

MODULATION SYSTEMS

Date........ March 4, 1947
Place.......Hotel Commodore, New York, N.Y.
Chairman...M. G. Crosby

Present
M. G. Crosby, Chairman
H. S. Black C. T. McCoy
F. L. Burroughs E. M. Ostlund
D. M. Hill G. R. Town
V. D. Landon B. Trevor
B. D. Loughlin J. E. Young

Chairman Crosby exhibited the galley proofs of the report, "Radio Progress During 1946." Suggestions of research problems suitable for submission to the I.R.E. Technical Committee on Research were requested by Mr. C. T. McCoy. A discussion developed concerning the policies, scope, and method of operation of the Research Committee which should be clarified before the other committees can be of much service to it. Considerable discussion centered around methods of reducing the time now required to get new standard definitions into use.

The Chairman read a program of the committee's work for the next several years. The committee passed the following motion: "Due to the fact that this I.R.E. committee embraces many fields in which other societies are interested, it will submit to the Executive Committee the proposal: in the case of a material prepared by this committee, it will submit to other societies and groups which might be interested, the prepared material for their comments, and will delay a maximum of two months before submitting the material to the Standards Committee." A subcommittee was appointed to revise the proposed program as read by the chairman. The following were appointed as members of this subcommittee: Messrs. Dietze, DiToro, Nixon, and Seeley.

ELECTROACOUSTICS

Date........ March 5, 1947
Place.......Hotel Commodore, New York, N.Y.
Chairman...R. S. Burnap

Present
R. S. Burnap, Chairman
E. L. Chaffee Louis Malter
K. C. DeWalt J. A. Morton
W. G. Dow I. E. Mouroumteff
J. E. Gorham L. S. Nergaard
J. W. Greer G. D. O'Neill
D. R. Hull H. J. Reich
S. B. Ingram A. L. Samuel
C. M. Wheeler

The Chairman reviewed the revised draft of the proposed Standards on Methods of Testing Frequency-Modulated Broadcast Receivers and several points were agreed upon for the first issue of the Standards. Subcommittees were set up to investigate and report on additional items. A subcommittee under the chairmanship of Mr. R. F. Shea will make the necessary revisions in the 1938 Standards on Testing Broadcast Receivers to bring them in line with current practice and with modern designs of amplitude-modulation sets. Mr. A. R. Hodges was appointed Chairman of a Subcommittee to compile information for annual review of the literature which will be included in the PROCEEDINGS OF THE I.R.E.

ELECTRON TUBES

Date........ March 5, 1947
Place.......Hotel Commodore, New York, N.Y.
Chairman...R. S. Burnap

Present
R. S. Burnap, Chairman
E. L. Chaffee Louis Malter
K. C. DeWalt J. A. Morton
W. G. Dow I. E. Mouroumteff
J. E. Gorham L. S. Nergaard
J. W. Greer G. D. O'Neill
D. R. Hull H. J. Reich
S. B. Ingram A. L. Samuel
C. M. Wheeler

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The purpose of the meeting was to determine procedures leading to revisions of Electronic Terms sponsored by A.S.A. Sectional Committee C-42. It was agreed that I.R.E. would provide all members present a report of the status of terms and definitions work of each of its nineteen Technical Committees. The various groups and subcommittees of A.S.A. Sectional Committee C-42 were checked for relationship to the I.R.E. definitions and recommendations for further appointments were made. A report on the status of the definitions work of pertinent A.S.A. groups will also be distributed to I.R.E. Technical Committees for comment.

RMA-I.R.E. CO-ORDINATION
Date……..April 9, 1947
Place……..Engineer’s Club
Chairman……V. M. Graham

Present
G. W. Bailey J. J. Farrell
A. B. Chamberlain Keith Henry
L. G. Cumming L. C. F. Horle

The meeting opened with a discussion of the possibility of further co-ordination between RMA and I.R.E. for the purpose of the conservation of man-hour expenditure in creating lists of Definitions and Standards. Mr. Graham presented for consideration, an excerpt from the foreword of RMA’s Standards and Engineering Information covering the respective scopes of I.R.E. and RMA in standardization work. Mr. Farrell suggested the RMA’s fundamental work in facsimile. Mr. Horle suggested that I.R.E. Technical Committees have representation at the meetings of the Radio Technical Planning Board. It was agreed that I.R.E. will provide RMA a schedule of its Technical Committees’ program of work, in advance, when available. After considerable discussion, it was agreed that in the future, RMA will request I.R.E. to provide available definitions and standards prior to initiating work on its own standards. The matter of standardization of disc recording systems was discussed.

(Continued on p. 595)
### Sections

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<td>Mrs. G. L. Curtis</td>
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**Sections**

**SUBSECTIONS**

**Chairman**

J. D. Schantz
Farnsworth Television (Chicago Subsection) and Radio Company
3700 E. Pontiac St.
Fort Wayne, Ind.

T. S. Farley
74 Hyde Park Ave. (Toronto Subsection)
Hamilton, Ont., Canada

K. G. Jansky
Bell Telephone Laboratories
Box 107
Red Bank, N. J.

**Secretary**

S. J. Harris
Farnsworth Television and Radio Co.
3702 E. Pontiac
Fort Wayne 1, Ind.

E. Ruse
195 Ferguson Ave., S.
Hamilton, Ont., Canada

L. E. Hunt
Bell Telephone Laboratories
Deal, N. J.

**Chairman**

C. W. Mueller
RCA Laboratories
Princeton, N. J.

**Secretary**

A. R. Kahn
Electro-Voice, Inc.
Buchanan, Mich.

W. A. Cole
323 Broadway Ave.
Winnipeg, Man., Canada

**Chairman**

A. V. Bedford
RCA Laboratories
Princeton, N. J.

A. M. Wiggins
Electro-Voice, Inc.
Buchanan, Mich.

C. E. Trembley
Canadian Marconi Co.
Main Street
Winnipeg, Man., Canada

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**Minutes of Technical Committee Meetings**

**SUBCOMMITTEES**

**ELECTRON TUBE CONFERENCE**

Date.............April 4, 1947
Place............I.R.E. Headquarters, New York, N. Y.
Chairman........I. E. Mouroumteff

The appointments of Mr. Schaffner as Vice Chairman of the Local Arrangements Committee and Professor Galbraith as Treasurer and of Mr. H. W. Parker as a member of the Entertainment Committee were approved. Mr. Schaffner reported that Mr. C. A. Priest had agreed to take charge of the banquet on June 9 following the Conference meeting. Mr. Mouroumteff submitted a list of questions regarding the program which he would like to circulate among the prospective members of the conference. This list will be mailed by Dr. Nergaard together with the invitations.

**POWER OUTPUT HIGH-VACUUM TUBES**

Date.............March 28, 1947
Place............I.R.E. Headquarters, New York, N. Y.
Chairman........I. E. Mouroumteff

Present
I. E. Mouroumteff, Chairman
L. G. Cumming
J. A. Morton
R. A. Galbraith
L. S. Nergaard
Louis Malter
G. D. O'Neill
S. G. Schaffner

The matter of additional assignments was discussed. The Chairman announced that the subjects of magnetron and transmitter-tube (T.R. tubes) had been assigned to this Subcommittee. It was suggested that since the present membership was not composed of many competent to deal with magnetrons, a small group be formed to consider the matter of desirable definitions and methods of test for magnetrons. The proposed definition for perversity was discussed and agreed upon. Methods of Testing, Section 3, Emission Tests, was approved as revised in recognition of comments received from the parent committee and the Small-Signal (SS) Tube Committee. A new definition for field-free emission current which is required because of new material included in Section 3, was approved. A revised draft of Section 5, leakage currents, as submitted by Mr. Mouroumteff, was discussed and some changes made.

**STUDENT BRANCHES**

The Board of Directors, at its April 2, 1947 meeting, approved the petitions that Student Branches be formed at the following universities: University of Alberta, Edmonton, Alberta, Canada; University of Michigan, Ann Arbor, Michigan; University of Syracuse, Syracuse, New York.

**INTERNATIONAL CONGRESS**

**ITALIAN NATIONAL RESEARCH COUNCIL**

The Italian National Council of Research will hold an International Congress in Rome, from September 28 to October 5, 1947, to celebrate the fiftieth anniversary of the discovery of radio by Marconi. It proposes to give a complete picture of the present development of radio studies in the technical and scientific field, and the possibilities foreseen for the future; and to gather radio experts from all over the world, so as to increase the scientific and technical international collaboration in the field of radio communications.

A program of meetings, receptions, and excursions, including visits to scientific Institutes and Italian industries working in the radio communications field, will be offered. The technical and scientific reports and papers presented will be discussed in special meetings and then collected into a volume which, in showing the results of the Congress, will document the development of radio applications up to date.

The Congress is open to all who are interested in radio studies. Participation forms should reach the General Secretariat of the Congress (Consiglio Nazionale delle Ricerche) at Rome by May 31, 1947; manuscripts of papers, in duplicate, must arrive there before June 30, 1947.

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**Note from the Executive Secretary**

Fellow-Members:

To some of the I.R.E. members, Headquarters is "just around the corner." These are the members living and working in this section of the country. To others, Headquarters seems a long way off.

But this is a message to all members. It aims to tell you that each of us at Headquarters regards every Section of the Institute and every member of the Institute as being right here. We feel very close to each of the Sections. Their fine accomplishments in the past and their vigorous work at present are well known at Headquarters.

Over and over again, at Board of Directors meetings, at Executive Committee meetings, and in conferences of the Institute staff, the thought has come up and been applauded that the members are the Institute, and that the Sections, which represent them, are guiding lights in the navigation of the Institute.

Naturally, an engineering society of over 20,000 members might seem a bit impersonal. But if you know how each individual member is regarded as "one of the family" here, and how pleased all of us are when matters of personal interest appear in the "I.R.E. News and Radio Notes" Section of the PROCEEDINGS, you would understand that the I.R.E. has managed to retain its youthful spirit and its friendly attitude toward all.

The latch string is always out here at your I.R.E. home.

George W. Bailey
Lawrence R. Quarles was born in Charlottesville, Virginia, on January 26, 1908. He received the degree of Bachelor of Science in engineering in 1929 from the University of Virginia. Following this he spent three years as a research engineer with Westinghouse Research Laboratories at East Pittsburgh, where he worked on various electron-tube circuits for industrial control. He returned to the University of Virginia in 1932 as a Service Fellow in physics and electrical engineering, and was awarded the degree of Doctor of Philosophy (physics) in 1935. He joined the staff of the electrical engineering department as instructor, became an assistant professor in 1939, associate professor in 1942, and professor in 1947. Professor Quarles is currently serving as acting dean of the department of engineering at the University of Virginia.

During the war Professor Quarles was associated as consultant on electronics problems with the Rouss Physical Laboratory, University of Virginia, which was operating under the Office of Scientific Research and Development.

He joined The Institute of Radio Engineers as a Member in 1942, becoming a Senior Member in 1943 when the new grading system was inaugurated. He is also a Member of the American Institute of Electrical Engineers, and a member of Sigma Xi and Tau Beta Pi.
The Industrial Scientist as Citizen

JOHN MILLS

Scientists and engineers are accused of exerting little influence upon the uses which society makes of their technical developments. They produce while others dispose.

Why is it that those who devise the physical mechanisms of our civilization are, as a rule—and the exceptions only prove the rule—so unconcerned and speechless as to the social utilization? They are driven by the creative urge of intellectual curiosity and the instinct of workmanship. Curiosity points the way and craftsmanship follows. But the compelling drive of curiosity becomes canalized. The rigorous training which they undergo, despite some courses in the humanities, soon limits their questions to their chosen fields. They tend to become specialists. They do not explore human relationships, social, economic, and political. In these matters, blocked by their own inhibitions, they tend to accept unquestioningly the current platitudes or the prejudices and doctrines they absorbed as children.

This tendency is pronounced in industry where organization is functional. Much of the power of industrialized science can be ascribed to its co-ordination of highly specialized experts. For any problem outside one expert's field there is, at least nominally, another thoroughly qualified expert. Each has vested interest in his speciality; and when technical advice is needed beyond his range he learns, on a live-and-let-live basis, to rely on co-ranking experts. He accepts their opinions the more completely the further their fields are from his own.

He bows to the authority of experts, and keeps strictly out of preserves that are not his own. Nor does he ask embarrassing questions about matters “which aren’t his business.” And thus by a natural transition he becomes sterile in matters social, economic, and political. He will damn a Communist, and justly, for following his party line and taking his opinions from Marx or Moscow; but he can remain blissfully unconscious while he is following a company’s “policy” line.

He has the brains and the analytical ability to handle the larger questions of social import; but he rarely speaks out and almost never acts. Despite his unique value to society, due to his facility in the scientific method, he pulls his punch and fails to follow through.

Today, in the words of the Biblical paradox, to save his life he must lose it. At any rate, he must divert time and energy from his scientific work to the social problems that vitally concern his future and that of all his fellow men.

This world of ours already has enough products of science to wreck it good and plenty. What it needs is some hair of the dog that bit it: an application to its present problems of the same methods of science as unwittingly provided the mechanisms by which it got where it is. And engineers and scientists must insure that the prescription is filled.
The Job Ahead*

CHARLES R. DENNY†

The bed of the ocean. The brakenian in the caboose of a speeding freight can speak with the engineer, and the engineer can speak with the dispatcher or with the conductor in another train. Buses, trucks, and taxicabs keep in touch with their central office by radio. And recently I rode down 14th Street in Washington with the Chief Engineer of the British Post Office as he talked by telephone with his wife on a farm in Manchester, England. Local and long-distance telephone calls in the United States from moving automobiles have become commonplace.

But to me one of the most awe-inspiring things that you have produced is the radio tube that not only can add, subtract, multiply, and divide but also can solve a series of complex simultaneous equations and retain in its memory as it goes along the answers to the equations which it has solved.

This is but a summary of the evidence against you, and you are involving yourselves deeper and deeper every day. You are experimenting with the use of airplanes circling in the stratosphere to relay signals from place to place and to broadcast frequencies-modulation programs and television pictures. And tomorrow morning at this convention you are going to discuss ways and means of bouncing radio waves off the moon, thereby using the moon as a passive repeater for ultra-high-frequency radio transmission. And there is even talk among you of artificial moons.

Your experiments have even gone beyond the present bounds of the radio spectrum. You are sending the voice by infrared waves and television by light waves.

Gentlemen, plainly there is sorcery in your science. My best legal advice to you is to plead guilty. Luckily the punishment for these days of such technological black magic is being awarded an honorary degree, a Fellowship, a Medal of Honor, the Morris Liebmann Memorial Prize, or the Browder J. Thompson Memorial award.

However mysterious your methods are, the end product of your work has won you the admiration of the world. Your progress continues at an ever-accelerating rate. The most daring prophecies of yesterday prove too conservative and become today's actuality. Things which today are experiments in your laboratories, simply ideas in the back of your minds, and even things which you have not yet dreamed of, will, I am confident, in the next ten years be in practical use, contributing importantly to the health, safety, culture, comfort, and well-being of men everywhere. Clearly, we are on the threshold of an immense expansion in the use of radio in our day-to-day lives.

In this expansion, the radio engineer, the radio industry, and the Commission must work together, closely and co-operatively.

Even before the war ended it was evident that the Commission was faced with an enormous expansion which was bound to come there would have to be important changes in the basic plan which allocates bands of frequencies to the various radio services. Developments since 1938, particularly in avia-
either operate on frequencies assigned to them, or some method must be devised for shielding them so that they do not radiate.

The Commission is attempting to solve this problem by setting up graveyards at strategic points in the radio spectrum where all radio heating devices can operate without causing interference to radio communication services. Thus far we have established four such graveyards in the 13-, 27-, 40-, and 2450-megacycle regions. The radio heating people advise us that they need still more frequencies and wider bands. We are endeavoring to provide for them as best we can. But we cannot come anywhere near giving them all they ask for without doing great damage to essential communication services.

Thus, we have here two great industries developing side by side in the radio spectrum. Their problems must be solved so that they can both go forward. On the one hand radio heating devices cannot be permitted to roam the spectrum indiscriminately and cause interference to radio communications. On the other hand provision must be made for the orderly growth of the vast new electronic heating industry. This, gentlemen, is going to be one of the biggest headaches of the next decade. I urge you now to place this problem on the agenda of things to be tackled in your laboratories.

Radio Heating—Radio was born and grew because of man's desire to communicate. But today radio is performing another service—a service growing so rapidly that it soon may boast of a larger investment than radio communications. This is radio heating. It is now being used for such diverse purposes as welding metals, melting plastics, vulcanizing rubber, curing tobacco, fusing glass, drying penicillin, relieving aches and pains, inducing artificial fever, and grilling frankfurters.

With thousands of what are, in effect, powerful transmitters in operation and radiating radio waves, we are all faced with a serious problem. These machines must

A Technical Audit*

An Address by PRESIDENT W. R. G. BAKER†, FELLOW, I.R.E.

PROBABLY each of us looks at the National Convention of The Institute of Radio Engineers from a different viewpoint. To me it represents the annual audit of the technical progress of the radio and electronics industry and to what extent the 20,000 scientists and engineers comprising The Institute of Radio Engineers have fulfilled their responsibility to the industry and to the public.

Just as an audit exhibits the facts and shows the fiscal status of a business, so do the exhibits and the technical papers show the progress of a science and the status of an industry in terms of physical accomplishments. Regardless of hopes, ambitions, and desires, an industry can advance no faster than its engineers and its productive facilities can make available the actual products on which new systems and services must depend. For this reason The National Convention of The Institute of Radio Engineers forces a realistic appraisal of what has been accomplished to date, and what may reasonably be expected in the near future.

It may be well to call your attention to a simple fundamental concerning the art and science of electronics. Back of all the countless developments in electronics and basic to all the applications, regardless of whether industrial or entertainment, is the electron tube in a circuit. This holds true whether we are considering the most simple form of broadcast receiver or the most complex radar equipment. The electron tube, which can detect, identify, amplify, regulate, and control, is the common denominator of the electronics industry.

We all recognize that new services now in the process of commercialization—frequency-modulation broadcasting and television—may well have far-reaching effects on the prosperity of our country and the standard of living of our people. Further than that, new systems and services still in the laboratories and hence not subject to this "audit," show in themselves that this science, art, and industry is so young that its future cannot be realistically projected.

The I.R.E. is the pre-eminent association in the field of electronics. It was organized in 1913. Its growth has been continuous and sound. At present the Institute of Radio Engineers represents 20,000 scientists and engineers engaged in all phases of the research and engineering of the electronics industry.

The future of the Institute is the future of electronics. The future of electronics is beyond man's ability to forecast.


† General Electric Company, Schenectady, N. Y.
One-Millionth-Second Radiography and Its Applications*

CHARLES M. SLACK†, AND DONALD C. DICKSON, JR.†, STUDENT, I.R.E.

Summary—The making of ultraspeed radiographs using exposure times of the order of one-millionth of a second requires the passage of electron currents approaching 1000 amperes. Such currents are supplied by an electron source utilizing field emission from a cold-cathode electrode which degenerates into a metallic arc in a high vacuum. The recording of such high-speed transients is briefly reviewed. The development of this equipment has been greatly accelerated because of the war. Illustrations showing its applications to various radiographic problems requiring short exposure times which have recently been released by the War Department are included; among these are radiographs taken at Frankford Arsenal and Aberdeen Proving Grounds of exploding shells and bombs, and at Princeton University showing the wounding mechanism of high-velocity fragments.

Introduction

Although X rays and “radio” waves are part of the same electromagnetic spectrum, it is not uncommon to find that engineers well versed in “radio” and allied electronic fields do not feel at home when discussing X rays. For this reason it is thought best to review some pertinent basic principles regarding the generation and utilization of X rays.

Fig. 1 is a diagram of a rather common type of hot-cathode, radiographic X-ray tube. The tungsten filament supplies electrons which are focused by the focusing cup and accelerated by the voltage drop across the tube to strike the tungsten anode or target approximately in the rectangular area shown. This bombardment of the tungsten by the electrons results in the emission of X rays with approximately equal intensity in all directions from each atom of the bombarded tungsten. However, the X rays which are directed into the anode are more or less absorbed depending upon the thickness of the anode, so that useful X rays are emitted in all directions to the cathode side of the plane of the anode. Because there is no known practical means of focusing X rays, it follows that all X-ray pictures must be shadowgraphs. For perfect definition in a shadowgraph it would be necessary to have a point source of X rays. It is impossible to obtain a true point source of X rays, so that it is then necessary to determine what maximum source size can be tolerated for a given application. Once a specific maximum bombarding area has been established there remains to be determined the maximum power that can be dissipated in the focal spot for a given time without raising the temperature of the tungsten surface sufficiently to cause excessive melting or vaporization of tungsten. The following formula applies in making this determination:

$$W_m = T_m \times (\pi KCt)^{1/2}$$

where $W_m$ is the loading in total energy per unit of focal-spot area which will raise the surface temperature an amount of $T_m$ degrees centigrade, $K$ is the thermal conductivity, $C$ is the heat capacity of the anode material, and $t$ is the exposure time.

When an X-ray tube is operating with constant accelerating potential, a certain number of milliamperc-seconds will be required to take a particular radiograph. It may be found that an exposure of 1-milliamperc second at 300 kilovolts is required to produce a certain desirable film blackening through 1 inch of steel placed 1 meter from the X-ray tube anode. This could be done by using an anode current of 1 milliamperc and an exposure time of 1 second, or a 10-milliamperc anode current and a 1/10-second exposure. Following this reasoning, it is seen that $10^6$ milliamperes or 1000 amperes would be required at 300 kilovolts to take the same picture in 1-millionth of a second.

Equation (1) indicates that for the same surface temperature of the tungsten this 1-millionth of a second exposure focal-spot area would have to be 1000 times larger than would be required to take the same picture in 1 second. Thus we see that when an X-ray shadowgraph is being taken of a stationary object there must be a sacrifice in definition as the exposure time is shortened. If a problem should require taking a radiograph of an object moving at high velocity it would be necessary to use an exposure time short enough to limit blurring to a small amount, which means that for moving objects a compromise must be made between lack of definition due to blurring and lack of definition due to focal-spot size.

* Decimal classification: 621.375, 623. Original manuscript received by the Institute, May 21, 1946; revised manuscript received, September 3, 1946. Presented, New York Section, New York, N. Y., January 26, 1946.
† Westinghouse Electric Corporation, Bloomfield, N. J.
From Fig. 1 it can be seen that by using the X rays coming from the anode in the direction shown the effective focal-spot size is considerably reduced along one dimension. This is known as the line-focus principle.

**EARLY ATTEMPTS TO DEVELOP ULTRASPEED RADIOGRAPHY**

Steenbeck, and also Kingdon and Tanis, solved the problem of high-speed radiography in the laboratory by utilizing a mercury-pool cathode tube, but the method had the limitations of a single-tube position plus the necessity of maintaining low mercury-vapor pressure by cooling to low temperatures. Oosterkamp succeeded in obtaining rather short exposures by suddenly raising the cathode temperature of an ordinary X-ray tube to near the melting point, but this was a rather dangerous procedure and yielded currents of only about 20 amperes.

**DEVELOPMENT OF THE FIELD EMISSION ARC TUBE**

A few years ago experiments were begun to determine whether or not field-emission currents from a cold metallic cathode could be used to obtain microsecond X-ray exposures. The first efforts were directed towards an investigation of the simple point-to-plane electrode arrangement shown in Fig. 2 (a). It was found that when a high positive voltage was suddenly applied to the flat tungsten electrode with respect to the pointed electrode, there was occasionally a very short burst of high-intensity X rays. For a given impressed voltage the operation was very dependent upon the spacing between the electrodes. If the spacing was too great there was no breakdown at all, but with very close spacing vaporized tungsten filled the gap so quickly that a low-voltage tungsten-vapor arc formed and all the voltage drop occurred in the circuit external to the tube, so that no X rays resulted.

In order to stabilize the breakdown characteristic, the electrode arrangement of Fig. 2 (b) was experimented with. The intention here was that, with the sudden application of voltage as shown, there would be an initial breakdown between the point cathode and a very closely spaced auxiliary anode. The resistor between the auxiliary anode and the main anode would limit the current through this initial arc, and then the discharge would transfer to the main anode. This principle was found to work very well experimentally and was incorporated into the design of a commercial tube whose electrodes had the configuration shown in Fig. 2 (c). The auxiliary anode has been made concave so as to give some focusing effect on the main electron stream.

![Fig. 3—Photograph of the electrodes of the ultraspeed X-ray tube](image)

Fig. 3 is a photograph of the electrodes in the commercial high-speed X-ray tube. A sharp-edged piece of metal G serves as the cold cathode. The entire structure H is the auxiliary anode, which is often loosely referred to as the auxiliary or starter cathode because its voltage is so nearly the same as that of the cathode after the low-voltage metallic arc has been formed between it and the cathode G. The tungsten anode I is about \( \frac{1}{2} \) inch thick, \( \frac{1}{4} \) inch wide, and \( \frac{1}{4} \) inches long, and is the source of the microsecond burst of high-intensity X rays.

Fig. 4 is a photograph of a standard high-speed tube. The main anode connection is the small cap at the bottom, while the cathode lead is brought out at the top. The side-arm connection is to the auxiliary anode, and the close external spacing between this and the cathode lead is possible because of the rapidity with which the starting metallic arc forms, bringing the two leads to nearly the same potential. Excluding the long flexible cathode lead, the tube is about 26 inches in length and 5 inches in diameter at the center.

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The slight blackening of the glass bulb around the discharge area is the result of anode tungsten being vaporized and then condensed on the bulb walls. This could be eliminated by using a larger focal spot, but the disadvantage of poorer definition with a larger focal spot outweighs the advantage of longer life. Under normal conditions the life of one of these high-speed X-ray tubes is limited by this bulb deposit. There is some reduction in X-ray output as a result of absorption by the deposit of tungsten, but the worst effect is to alter the electrostatic fields inside the tube so that defocusing of the electron stream occurs, resulting in a large part of the discharge missing the anode. Finally, after the bulb deposit has become sufficiently great, it will "short circuit" the entire discharge away from the cathode-anode region.

**Power-Supply and Control Circuit for High-Speed X-Ray Tube**

The Marx-type surge-generator circuit is ideally suited to supply the millionth-of-a-second burst of energy at high voltage which is necessary to operate the tube. Fig. 5 is a simplified schematic diagram of the surge-generator power supply and the firing-control circuit which are incorporated in the standard commercial unit. The six capacitors to the right of C are charged in parallel to a maximum of 50 kilovolts. At the same time, with switch B closed, C₁ is charged to about 1000 volts. When the circuit is broken at B, as a result perhaps of a bullet breaking a metalized glass fiber, the negative blocking voltage on the grid of thyratron T begins to decrease at a rate dependent upon the time constant of \( R_2 C_8 \). When the thyratron becomes conducting, capacitor C₄ discharges through the primary of an induction coil E which causes a high-voltage impulse from the secondary of E to break down the triggering gap L. Once this gap is broken down the other sphere gaps in the surge generator follow very quickly, which effectively puts all of the capacitors in series across the X-ray tube. Resistors \( R_b \) and \( R_d \) and the rest of the charging resistors are of sufficiently high resistance so that very little energy is lost in them during the microsecond or so that it is necessary to discharge the capacitors through the X-ray tube. Resistor J, which stabilizes the starting arc, has a resistance usually of between 5000 and 20,000 ohms, while K is 100,000 ohms or more and serves only to keep the cathode and anode at the same voltage before the breakdown of gap \( F_1 \) results in the application of high voltage to the tube.

An actual installation of this equipment at Picatinny Arsenal is shown in Fig. 6. This is a photograph of two
high-speed radiographic units set up side by side so that simultaneous or sequence pictures might be obtained of the same or of related objects. One charging unit containing the high-voltage transformer, two rectifying tubes, and associated relays and resistors is designed to supply four surge generators simultaneously. The drain on the 60-cycle power line is very small because it is always convenient to spread the 300-watt-second maximum charge on the capacitors of one surge generator over several seconds, and after reaching full charge the power line has only to supply the rectifier filament power and some small losses.

**Applications**

The applications of this ultra-high-speed radiographic technique are in general similar to those of the better-known flash photographic techniques, with the important addition that the X-ray method can be used to reveal internal behavior of material opaque to light. Also, the X-ray method is not at all influenced by visible light originating with the object under study.

![Fig. 7—Series showing the penetration of a steel plate by an explosive 20-millimeter shell. Note the swelling of the shell before the casing splits. Radiographs by Frankford Arsenal.](Image)

These characteristics of the high-speed X-ray technique make this equipment uniquely suited to a study of ballistics.

Aside from some laboratory experiments which demonstrated the equipment’s possibilities, the first practical application of this technique was made by the Remington Arms Company. They were able to study the progress of a shot-gun charge down a gun barrel and to see the exact manner in which choking action took place while the charge was still in the barrel. Even after the charge left the barrel, light pictures could not be taken for a time because of the smoke and flame which surrounded the charge for some distance outside of the barrel. Among other things, they also studied the internal action of recoil and ejection mechanisms.

Fig. 7, taken at Frankford Arsenal, shows four stages in the penetration of a steel plate by a high-explosive 20-millimeter shell. For this particular series a different shell and plate were used for each frame and the radiograph was taken at greater and greater times after the initial impact. Successive shell penetrations are so consistent that it was possible to make a moving picture strip by combining many frames similar to those of Fig. 7, but taken with shorter time intervals between them. When seen on a screen the result appears to be a very slow-motion moving picture of a single shell penetrating a steel plate.

Fig. 8 shows radiographs taken at the Aberdeen Proving Center, before static detonation and also 49 microseconds after static detonation of a model bomb.

![Fig. 8—Radiograph showing the distribution pattern of a model bomb. Radiograph made at the Aberdeen Proving Center.](Image)

The same thing has been done with other types of bombs and shells, and, as can be seen, this technique results in very detailed information regarding the distribution of fragments.

It had been observed in both world wars that in some instances battle casualties suffered internal injuries which were very much more severe than external evidence would seem to indicate, and it was thought that this type of wounding was due to very-high-velocity fragments. During World War II, E. N. Harvey directed work at Princeton University on an Office of Scientific Research and Development contract...
in which this high-speed X-ray technique was applied to a study of the wounding effect of high-velocity fragments. Fig. 9 (a) is a radiograph of the leg of an anesthetized cat whose blood vessels have been injected with a material fairly opaque to X rays. Fig. 9 (b) is a high-speed radiograph taken approximately at the instant when the cavity caused by a small, very high-velocity steel sphere is at its maximum diameter. The steel sphere in this case is traveling perpendicular to the paper. Fig. 9 (c) is a radiograph of the cat’s leg taken after passage of the steel sphere, showing that the leg bone has been broken by the shock wave produced by a small sphere which penetrated the leg a considerable distance from the bone.

The applications cited above were considered to be representative of a few of the more interesting ones that have been released for publication. However, the most extensive use of the high-speed X-ray equipment to date has been in connection with the development of the atomic bomb, but no more may be said on this subject.

Oscillographic Analyzing Equipment

In order to understand more fully the exact nature of the field-emission discharge which is the basis of the high-speed X-ray tube’s operation, it was thought desirable to set up equipment which would make possible the simultaneous oscillographic recording of the tube anode voltage, anode current, and X-ray intensity, all with respect to time. Preliminary examination indicated that the attainment of this objective would necessitate the solution of several rather independent problems. However, to discuss the various lines of approach which were considered in arriving at the overall objective would require considerable space, so the results to date will be presented and discussed here.

Fig. 10 is a photograph showing two separate, doubly shielded rooms. The shielding of each room consists of solid sheet copper on the inside with galvanized sheet iron on the outside. The room on the left contains an experimental surge generator, part of which can be seen through the doorway, an oil tank in which high-speed X-ray tubes are placed for operation at up to 600 kilovolts, a 60-kilovolt power supply for the surge generator, a special triggering unit, current- and voltage-viewing resistances, and some miscellaneous control equipment. The small shielded room shown on the right houses the oscillographic recording equipment.

Fig. 11—The voltage-dividing units showing the type of shielding and co-axial wiring necessary to avoid the effect of external disturbances.

Fig. 11 is a photograph of the shielded voltage- and current-viewing resistors removed from the vicinity of the surge generator for clarity. The resistors which
rise out of the inverted copper pyramid are part of the high side of the voltage divider. The low side of the voltage divider is shielded by the cylindrical brass box on the left and is connected to the high side by a short length of coaxial cable. The cylindrical brass box on the right contains the current-viewing resistance. Fig. 12 is a close-up photograph showing how sixteen Globar resistors have been arranged in parallel to obtain low inductance and lower current density per resistor than could be obtained with fewer resistors. The two parallel coaxial cables, which can be seen leading out from between the two cylindrical boxes in Fig. 11, go to the vertical deflection plates of two of the synchroscopes.

Fig. 12—Inside view of voltage-dividing unit, showing arrangement of Globar resistors.

order to obtain sufficient deflection voltage for the cathode-ray tube.

Fig. 13 is a block diagram which shows how all of these components are arranged in the two shielded rooms. The procedure for obtaining simultaneous traces of voltage, current, and X-ray intensity is somewhat as follows: The surge-generator capacitors are first charged to the desired voltage. With the cathode-ray beams biased so that no light is visible on the screens and with the three synchroscopes set for single sweep operation, the camera shutters are opened and the microswitch firing control A is pressed. This results in simultaneous voltage trigger pulses being applied to the three synchroscopes from the trigger supply B. The trigger output from the current-viewing synchroscope is led by coaxial cable into the large shielded room and through a 10-microsecond delay line to a special triggering unit. This unit utilizes certain radar components to generate a very fast-rising, 30-kilovolt pulse which triggers the main surge generator. The delay line prevents interference feedback to the current synchroscope until after all traces have been

Fig. 13—A block circuit diagram of the experimental surge generator unit and associated oscillographic equipment.

The most unique problem encountered in this work was that of converting instantaneous X-ray intensity to deflection-plate voltage. This was accomplished by use of a fast-acting ionization chamber whose voltage output had to be amplified several hundred times in
Exact Design and Analysis of Double- and Triple-Tuned Band-Pass Amplifiers

MILTON DISHAL†, SENIOR MEMBER, I.R.E.

Summary—The purpose of this paper is to present a quick, complete, and exact method of design and analysis of double- and triple-tuned band-pass amplifiers.

The necessary small-percentage pass-band equations are derived giving the relationship between the circuit characteristics and the response characteristics. These circuit characteristics are: the resonant frequency \( f_0 \), coefficient of coupling \( K \), the circuit \( Q \), and the input and output capacitances \( C_{in} \) and \( C_{out} \). The response characteristics are: the percentage bandwidth between peaks \( \Delta f_0/f_0 \), the peak-to-valley ratio within the pass band \( V_p/V_v \), the peak-to-
"skirt" response ratio \( V_p/V_\delta \) at different skirt-to-peak bandwidth ratio points \( \Delta f/\Delta f_0 \) outside the pass band, the circuit gain at the peaks, and the phase shift \( \phi \) at any frequency.

These design equations, extended to the case of one to eight cascaded stages, are incorporated in two sets of conveniently used nomographs, one for double-tuned circuits and one for triple-tuned circuits. Specific examples of the use of these nomographs are given.

**Symbols**

\[ f_0 = \text{resonant frequency (see Section IV)} \]
\[ \omega_0 = \text{resonant radian frequency} \]

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† Federal Telecommunication Laboratories, Nutley, N. J.

\[ n = \text{decrement of a resonant circuit (see Section III)} \]
\[ Q = \text{reciprocal of decrement (see Section III)} \]
\[ K_c = \text{coefficient of capacitive coupling between resonant circuits (see Fig. 3)} \]
\[ K_T = \text{coefficient of inductive coupling between resonant circuits (see Fig. 3)} \]
\[ K_M = \text{coefficient of mutual inductive coupling between resonant circuits (see Fig. 3)} \]
\[ K = [K_c(\omega/\omega_0) - K_T(\omega/\omega_0)] \]
\[ C, L, M, R = \text{capacitance (farads), inductance (henrys), mutual inductance (henrys), resistance (ohms)} \]
\[ B = \text{susceptance} \]
\[ G = \text{conductance} \]
\[ F = (f/f_0 - f_0/f) = (\omega/\omega_0 - \omega_0/\omega) \]
\[ \theta = \text{phase angle between a resulting voltage and the driving current, or the phase angle between a resulting current and the driving voltage} \]
\[ \Delta f = \text{difference between two frequencies} \]
\[ V' = \text{response voltage} \]
\[ \beta = \text{a constant for double-tuned circuits (a function of peak-to-valley ratio)} \]
\[ N = \text{number of cascaded stages} \]
\[ \gamma = \text{a constant for triple-tuned circuits (a function of peak-to-valley ratio)} \]
\[ \Delta f_p = \text{bandwidth between response peaks} \]
\[ V_r = \text{voltage at peaks of the response} \]
\[ V_v = \text{voltage at the valley of the response} \]
\[ D = n_1/n_2 = Q_1/Q_2. \]

I. INTRODUCTION

To aid in the design and analysis of circuits which produce a band-pass response with respect to frequency, there has arisen a large body of literature under the two general headings of filter theory and coupled-circuit theory. However, it would be worth while to have collected in one place a method of design that will quickly and easily give answers to questions of the following type which might arise in the course of a thorough design of, say, a wide-band intermediate-frequency amplifier for receivers:

(a) For a given bandwidth and "flatness" of response, exactly how much more gain can be obtained if we use triple-tuned rather than double-tuned circuits?

(b) Can a certain skirt-selectivity specification be satisfied using only five double-tuned circuits? If so, what must the circuit constants be? What peak-to-valley ratio will there be in the pass band?

(c) How much more gain and how much greater skirt selectivity will be obtained if we accept a relatively poor response in the pass band by allowing a 1.3 peak-to-valley ratio in preference to a good 1.05 peak-to-valley ratio? What must the circuit constants be for both cases?

(d) Will more gain per stage be obtained if all the loading is done in one of the resonant circuits, or should the Q of all the resonant circuits be made equal?

In this paper, through the medium of two sets of three nomographs each, the writer hopes to provide in one place a ready means of obtaining exact answers (with a minimum of time and calculation) to the above and other questions for the case of double-tuned and triple-tuned band-pass circuits when small-percentage (20 per cent or less) bandwidths are used.

The concepts and constants used are those commonly associated with coupled-circuit theory. Filter-theory constants and concepts are always useful, and when many tuned circuits are coupled together it is practically necessary to use the filter-theory type of design. However, even for both double- and triple-tuned circuits, it is possible to obtain exact closed-form solutions for the circuit response (when band-pass percentages are approximately 20 per cent or less); and these solutions are more concisely stated in terms of coupled-circuit constants.

The circuit constants used are the resonant frequency

\[ \gamma = \text{a constant for triple-tuned circuits (a function of peak-to-valley ratio)} \]
\[ \Delta f_p = \text{bandwidth between response peaks} \]
\[ V_r = \text{voltage at peaks of the response} \]
\[ V_v = \text{voltage at the valley of the response} \]
\[ D = n_1/n_2 = Q_1/Q_2. \]

II. DESIGN AND ANALYSIS OF DOUBLE-TUNED CIRCUITS BY MEANS OF THE NOMOGRAPHS

From (18a), (19a), (21a), (23a), and (25a) of this paper, a set of nomographs have been prepared, and a family of curves have been prepared from the phase-shift equation (26). The use of these nomographs is best explained by a few specific examples.

Example 1

Knowing that the gain per stage is approximately

\[ \text{Gain} = G_m/4\pi \Delta f_p \sqrt{C_1 C_2} \]

and that \( C_1 = C_2 = 10 \mu \text{microfarads} \) and \( G_m = 5 \times 10^{-3} \) mho, it is decided that five stages are probably needed to obtain a certain desired gain. A ratio of peak gain to valley gain of 1.10 will be satisfactory. A bandwidth between peaks \( \Delta f_p \) of 2 megacycles is required; and to make the percentage bandwidth approximately 20 per cent or less, a midfrequency \( f_0 \) of 30 megacycles is chosen.

What loading resistances should be used to give the proper \( Q \) in the two tuned circuits? What exact gain per stage will be obtained? What must the mutual impedance be to give the proper coefficient of coupling? What will the bandwidth be 6 decibels down from the peaks? What will the bandwidth be 60 decibels down from the peaks?

Starting with Chart A, place a straight edge between point 5 on the "Number of Cascaded Stages (N)" column and point 1.10 on the "\( (V_r/V_v) \)" column. From the \( \lbrack Q/(f_0/\Delta f_p) \rbrack \) column, we find that the \( Q \) of each resonant circuit must be

\[ Q = 0.69 \frac{f_0}{\Delta f_p} \approx 10, \]

and from this same column the gain per stage will be

\[ \text{Gain} = 0.69 \frac{G_m}{4\pi \Delta f_p \sqrt{C_1 C_2}} \approx 14. \]

Knowing the necessary resonant-circuit \( Q \) and the reactance of the total shunt capacitances in the resonant circuits, the necessary resultant loading resistance is, of course, given simply by \( R = Q X_0 = 10 \times 500 \) ohms = 5000 ohms.
Since the coils used will usually have appreciable loss, they will effectively supply a shunt loading resistance of value $Q_L X_0$, where $Q_L$ is the $Q$ of the inductance and $X_0$ is the impedance of the shunt capacitance at the resonant frequency.

Thus, the resistance $R_+$ which must be added in parallel with the above effective resistance, due to a $Q_L$ of 50, for example, to produce the required resultant $Q$ is

$$R_+ = \left[ Q/\left(1 - \frac{Q}{Q_L}\right) \right] X_0 \approx 6250 \text{ ohms.}$$

From the "KQ" column of Chart A, the coefficient of coupling must be

$$K = \frac{1.22}{Q} = 0.122.$$
In the type of circuit chosen (see Figs. 1 and 2), the mutual reactance between the two resonant circuits is then found from the simple equation for the coefficient of coupling as given with each type of coupling in Fig. 3.

To consider skirt selectivity, use Charts B and C. On Chart B, place a straight edge between point 5 on the "Number of Cascaded Stages (N)" column, and 6 decibels (or 2) on the "(V_p/V)" column. Read 0.56 on the middle, or "Y" column. Now, going to Chart C, place the straight edge between 0.69 on the "[Q/(fo/Δf)]" column and 0.56 on the "Y" column and read from the middle column that

\[ \Delta f \approx 1.95 \Delta f_p = 3.9 \text{ megacycles}. \]

The bandwidth at the 60-decibel-down point is obtained in exactly the same way, i.e., on Chart B, place the straight edge between the point 5 on the "Number of Cascaded Stages (N)" column and 60 decibels (or 1000) on the "(V_p/V)" column. Read 3.6 on the "Y" column. Going to Chart C, place the straight edge between

Chart B—Double-tuned band-pass circuit design for factor Y.
point 0.69 on the \[ \frac{Q}{f_0/\Delta f_p} \] column and 3.6 on the "Y" column. Read from the "\( \frac{\Delta f}{\Delta f_p} \)" column that

\[ \Delta f_{60 \text{ decibels}} = 4.4 \Delta f_p = 8.8 \text{ megacycles}. \]

Any other points on the response curve are found in the same manner.

**Example II**

Knowing that the approximate gain per stage is

\[ \text{Gain} = \frac{G_m}{4\pi \Delta f_s \sqrt{C_1C_2}} \]

it is decided that only 3 stages are needed to give a certain desired gain. It is necessary that the skirt selectivity be such that the bandwidth 60 decibels down be only 5 times the bandwidth between the peaks, i.e., \( \Delta f_{60 \text{ decibels}}/\Delta f_p = 5 \). What must be the \( Q \) of each tuned circuit to obtain this skirt selectivity? What exact gain per stage will be obtained? What coefficient of coupling is required? What peak-to-valley
ratio must be accepted in order to obtain this selectivity.

Starting with Chart B, place a straight edge between point 3 in the "Number of Cascaded Stages \((N)\)" column and point 60 decibels on the \((V_p/V_v)\) column and read 9.6 from the "Y" column. Going to Chart C, place the straight edge between point 5 on the \((\Delta f/\Delta f_p)\) column and 9.6 on the "Y" column and read on the \([Q/(f_0/\Delta f_p)]\) column that the required \(Q\) is

\[Q = 1.1 \frac{f_0}{\Delta f_p}\]

Now, going to Chart A, place the straight edge between point 3 on the "Number of Cascaded Stages \((N)\)" column and 1.1 on the "\([Q/(f_0/\Delta f_p)]\)" column. The exact gain will be

\[\text{Gain} = 1.1 \frac{G_m}{\pi \Delta f_p \sqrt{C_1 C_2}}\]

and, from the "\(KQ\)" column, the required coefficient of coupling is

\[K = \frac{148}{Q}\]

From the \((V_p/V_v)\) column, the resulting peak-to-valley ratio will be

\[V_p/V_v = 1.27\]

The resonant frequency as 15 megacycles, what is the response curve?

The product of \(KQ\) is 1.7. Going to Chart A, set the straight edge between 1 on the "Number of Cascaded Stages \((N)\)" column and 1.7 on the "\(KQ\)" column. From the \((V_p/V_v)\) column, \(V_p/V_v = 1.15\). From the \([Q/(f_0/\Delta f_p)]\) column, the bandwidth between peaks will be

\[\Delta f_p = 1.38 \frac{f_0}{Q} = 1.04\]  megacycles.

To find the width of the skirts at different points, e.g., 10 times or 20 decibels down, go to Chart B. Place the straight edge between 1 on the "Number of Cascaded Stages \((N)\)" column and 20 decibels on the \((V_p/V_v)\) column and read 10 on the "Y" column. Going to Chart C, place the straight edge between 1.38 on the \([Q/(f_0/\Delta f_p)]\) column and 10 on the "Y" column and see that

\[\Delta f_{20\text{ decibels}} = 4.4\Delta f_p = 4.6\]  megacycles.

Any other points on the skirts are obtained in the same way.

**Example IV**

To find the phase shift at any point in the pass band, Chart D is used.

It should be noted that \(2(f - f_0)/\Delta f_p\) (which is the abscissa of the graph) is merely a way of writing \((\Delta f/\Delta f_p)\) to show more clearly that in the phase-shift equation

\[2(f - f_0)\]

**Example III**

The nomographs may be conveniently used for analysis of coupled circuits, as well as for design or synthesis.

Thus, given the \(Q\) of two resonant circuits as 20, the coefficient of coupling \(K\) between them as 0.085, and

\[26, \Delta f\] defines two frequencies equidistant from the resonant frequency. The abscissa is (+) for frequencies above the resonant frequency and is (−) for frequencies below the resonant frequency. (E.g., at the high-frequency peak \(f = f_{pA}\) and \(2(f - f_0)/\Delta f_p = +1\), and at the chart D.—Phase shift for a flat-top double-tuned circuit for different peak-to-valley ratios.
low-frequency peak \( f = f_p + \) and \( 2(f - f_0)/\Delta f_p = -1. \)

Note also that the ordinates give the phase shift per stage. If \( N \) cascaded identical stages are used, this phase shift is then multiplied by \( N \).

Finally, note that the peak-to-valley ratios for each curve are the ratios for a single stage.

Thus, if 6 cascaded stages are being used to produce a resultant peak-to-valley ratio of 1.05, each single stage must have a peak-to-valley ratio equal to the 6th root of 1.05, or 1.0083. For this case, curve 8 would therefore give the phase shift versus frequency per stage. This phase shift at each frequency is then multiplied by 6 to give the resultant phase-shift-versus-frequency curve.

III. CIRCUITS WHICH ARE ANALYZED

The basic circuit analyzed is the two-node network consisting of two resonant circuits coupled together both inductively and capacitively. This circuit and the response investigated are shown in Fig. 1.

By virtue of the exact equivalence of \( \pi \)'s, \( T \)'s, and transformers, the exact analysis of the basic circuit is immediately applicable to ten more circuits. These ten circuits are shown in Fig. 2 and the equations giving the values of the equivalent elements are given in Fig. 3. Lattice, bridged-\( T \), etc., equivalents may also be used.

By virtue of the concept of duality, \(^6\) the analysis of the basic two-node network is immediately applicable to the dual two-mesh network given in Fig. 4, where \( I \), the equivalent constant-current generator, and \( G \), \( C \), and \( L \), are substituted respectively for \( E \), \( R \), \( L \), and \( C \). Again, by virtue of the equivalence of \( \pi \)'s, \( T \)'s, and transformers, the analysis also applies to the ten additional circuits given in Fig. 5. Thus, a total of 22 band-pass circuits are effectively analyzed in this paper, plus any lattice, bridged-\( T \), etc., equivalents which the reader may desire to use.

The two-node circuit of Fig. 1 is picked as the circuit to be analyzed, rather than the dual two-mesh circuit of Fig. 4, because vacuum-tube amplifiers are effectively high-impedance generators, and for practically all high-frequency band-pass amplifier applications, high-impedance resonance is desired as obtained by the use of the circuits of Figs. 1 and 2.

If very-small-percentage pass bands are to be produced, and very slight inequality in the height of the two peaks can be tolerated, then all 22 of the circuits shown in Figs. 1, 2, 4, and 5 can be used as either high- or low-impedance circuits by means of the following reasoning. For the small-percentage band-pass case, it is convenient (and correct) to consider the band-pass characteristic as being produced in the following manner:

(a) Fundamentally, the configuration of only the lossless reactive components produces the band-pass response; the percentage bandwidth being fixed (to a first approximation) by the coefficient of coupling \( K \). Figs. 1 and 2 and Figs. 4 and 5 give the two-node and the two-mesh reactive networks which can produce a band-pass characteristic. (Consider the shunt resistors of Figs. 1 and 2 to be open-circuited and the series resistors of Figs. 4 and 5 to be short-circuited.)

(b) The peak-to-valley ratio is fixed to a first approximation by the required \( Q \) of the input and output resonant circuits. The correct resonant-circuit \( Q \) can be produced in three ways: (1) by placing a small resistance in series with the resonant circuit, \( Q = X_{in}/R_s \); (2) by placing a large resistance in parallel with the resonant circuit, \( Q = R_p/X_{in} \), or (3) by a combination of both series and parallel loading. For this case,

\[
Q = \frac{1}{(R_s/X_{in}) + (X_{in}/R_p)}
\]

(c) The driving force may be applied in two ways: either an infinite-impedance (i.e., zero conductance) constant-current generator may be placed in parallel with either the resonating inductance or the resonating capacitance (never across the mutual reactance); or a zero-impedance constant-voltage generator may be placed in series with either the resonating inductance or resonating capacitance (never in series with the mutual reactance).

In practice, all equivalent generators have finite output impedances associated with them. Thus, the above steps, (b) and (c), are interrelated to the extent that the effect of the output impedance of the generator upon the resonant-circuit \( Q \) must be considered.

(d) The output voltage may be obtained across either the resonating inductance or the resonating capacitance in the output circuit. Of course, we must consider the effect of the resistive component of the load upon the \( Q \) of the output resonant circuit.

---

Fig. 2—Ten two-node circuits. The circuit of Fig. 1 is exactly equivalent to these circuits.

IV. Elements Which Are Resonated
It is important to know exactly what elements are resonated in the above circuits. The elements which are tuned to resonance in circuit I, Fig. 2, and all the two-node circuits are indicated by the following procedure:

Node 2 is shorted to ground and all the reactive elements remaining are resonated at the desired frequency; then Node 1 is shorted to ground (the short on Node 2 is removed) and all the remaining reactive elements are resonated to the above frequency. Thus, in circuit I,
Fig. 3—Coefficient of couplings used in the analysis and the π, T, and transformer exact equivalents, and approximations for small couplings.

$C_1$ plus $C_m$ is resonated with the resultant of $L_2$ and $L_n$ in parallel. Thus, for the circuit I in Fig. 2, we have

$$\omega_0^2 = \frac{1}{\left(\frac{L_2 L_n}{L_1 + L_n}\right)(C_1 + C_m)}$$

This method of defining the resonances also introduces a very practical method of aligning double- or
triple-tuned coupled resonant circuits. First, completely detune all but one of the resonant circuits without affecting the mutual impedance. This detuning effectively short-circuits the node to ground for all practical purposes, and may be accomplished simply by placing an additional capacitance across the resonant circuits whose value is approximately three or four times that of the capacitance in the circuit. Or, if iron-slug tuning is used, sufficient detuning can usually be accomplished merely by turning the slug to its extreme position. Second, feed a signal into the circuit at the desired resonant frequency and tune the remaining one circuit, which is not detuned, for maximum output. This procedure is then repeated until all the circuits have been resonated in the above manner.

Actually, for a certain distribution of the circuit constants, i.e., $Q_1 = Q_2$, there is a more convenient method of alignment which will be mentioned later.

In the dual two-mesh circuits, the elements to be resonated are indicated by the following procedure: Mesh 2 is open-circuited and all the reactances remaining in the circuit are resonated. Then, with Mesh 2 returned to its normal condition, Mesh 1 is open-circuited and all remaining reactive elements are resonated to the same frequency. Thus, for circuit A, Fig. 5, we have:

$$\omega_0^2 = \frac{1}{(L_1 + L_m) \left( \frac{C_1 C_m}{C_1 + C_m} \right)} = \frac{1}{(L_2 + L_m) \left( \frac{C_2 C_m}{C_2 + C_m} \right)}$$

(2)

V. Exact Response Equations

The node equations for the circuit shown in Fig. 5 are:

$$I = [G_1 + j(B_{1n} + B_{2m} - B_{1b} - B_{2n})] V_1$$

$$- j(B_{2n} - B_{1m}) V_2$$

$$0 = - j(B_{1m} - B_{2n}) V_1$$

$$+ [G_2 + j(B_{2n} + B_{1m} - B_{1b} - B_{2m})] V_2$$

As mentioned in Section 3, the solution of the above two equations for the response voltage $V_2$ contains the solution for all the 22 circuits shown in Figs. 1, 2, 4, and 5.

A great simplification is produced in the resulting equations for the circuits if the resonant frequency $f_0$, the coefficient of coupling $K$ between resonant circuits, and the decrement of each resonant circuit $n$ are introduced into the circuit equations. (The decrement is the reciprocal of the more commonly used $Q$.)

With the introduction of these constants, the equations can be expressed in terms of the three quantities only instead of in terms of the eight $L, C, R$ elements making up the circuit. Our mental picture of the circuit action is thus greatly simplified.

By solving (3) for the output voltage $V_2$ and introducing into the solution the three constants mentioned above, namely,

$$\omega_0^2 = \frac{1}{(L_1 L_m) \left( \frac{C_1 C_m}{C_1 + C_m} \right)} = \frac{1}{(L_2 L_m) \left( \frac{C_2 C_m}{C_2 + C_m} \right)}$$

(4)

$$n_1 = \frac{G_1}{\omega_0(C_1 + C_m)}$$

(5)

$$n_2 = \frac{G_2}{\omega_0(C_2 + C_m)}$$

(6)

$$K_e = \frac{C_m}{\sqrt{(C_1 + C_m)(C_2 + C_m)}}$$

(7)

$$K_L = \frac{\sqrt{L_1 L_2}}{\sqrt{(L_1 + L_m)(L_2 + L_m)}}$$

(8)

we obtain the exact solution for the magnitude of the response

$$V_2 = \frac{I}{\omega_0 \sqrt{(C_1 + C_m)(C_2 + C_m)}} K \sqrt{F^4 - 2 \left[ K^2 - \frac{n_1^2 + n_2^2}{2} \right] F^2 + (K^2 + n_1 n_2)^2}$$

(9)

and the phase of the output voltage with respect to the constant current source is

$$\tan \theta = \frac{\pm \left[ K^2 + n_1 n_2 - F^2 \right]}{\pm \left[(n_1 + n_2)F\right]}$$

(10)
Fig. 5—Ten two-mesh circuits. The circuit of Fig. 4 is exactly equivalent to these circuits.

where

\[ K = \left( K_e \frac{\omega}{\omega_0} - K_L \frac{\omega_0}{\omega} \right) \]

The sign to be used in the phase-shift equation (10) is the sign of the quantity

\[ F = \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right). \]
Thus, with capacitive coupling predominant, the top signs are used in numerator and denominator, and, with positive inductive coupling predominant, the bottom signs are used.

Examination of the numerator of (9) shows immediately one characteristic of the response. The numerator becomes zero and thus there is a null response at

$$\frac{\omega_{null}}{\omega} = \sqrt{\frac{K_L}{K_e}}. \quad (13)$$

With reference to circuit IC of Fig. 2, it should be mentioned that if the winding sense of the inductances is such that the mutual inductive coupling "aids" the capacitive coupling there is no null of response, for then the sign of $K_L$ in (11) is negative (−) and, therefore, the numerator never becomes zero.

We will now introduce into the above exact equations the approximations that produce the symmetrical and relatively simple small-percentage pass-band analysis.

VI. SMALL-PERCENTAGE PASS-BAND RESPONSE SHAPE

Because $K$ in (9) is a function of frequency, the exact response shape is not symmetrical either geometrically or arithmetically with respect to frequency. If, however, we limit ourselves to small-percentage pass bands where $\omega/\omega_0$ varies in value over the small range from, say, 0.9 to 1.1, then two important simplifications immediately result in the factors shown in (11) and (12).

Equation (11) becomes independent of frequency:

$$K = (K_e - K_L). \quad (11a)$$

(It must be realized that this approximation cannot be used in the region of the null given by (13).)

Equation (12) becomes

$$F = \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) = \frac{2(\omega - \omega_0)}{\omega \omega_0} = \frac{\Delta f}{\omega_0} \frac{2\omega}{2\omega} \Delta f \frac{f_0}{f_0} \Delta f. \quad (12a)$$

where $\Delta f$ is the frequency bandwidth between points equidistant from the resonant frequency $f_0$.

With the above limitation, (9) shows that, in the small-percentage pass-band case (where (11a) applies) the shape of the amplitude response curve is independent of the type of coupling used. The gain obtained with inductive coupling only is slightly greater than that obtained with capacitive coupling, for, as seen from (9), the capacitances which must be considered in figuring the gain are $(C_1 + C_m)$ and $(C_2 + C_m)$ ($C_m$ is the equivalent high-side capacitances of Fig. 5), and $C_m$ is zero for inductive coupling only.

The phase shift as given by (10) does differ for the two types of coupling. Since the top signs are used with capacitive coupling and the bottom signs with inductive coupling, we will have positive phase angles with capacitive coupling and negative phase angles with inductive coupling.

The frequency at which the response maximum and minimum occurs is given by differentiating (9) with respect to $F$ (i.e., $\Delta f/f_0$) and equating to zero. This results in

$$\left(\frac{\Delta f}{f_0}\right)^2 = K^2 - \frac{n_1^2 + n_2^2}{2} \quad (14)$$

and the location of the minimum is given by $\Delta f/f_0 = 0$.

The response at the peaks, obtained by substituting (14) in (9), is

$$V_{\text{peaks}} = \frac{I}{\omega_0 \sqrt{(C_1 + C_m)(C_1 + C_n)}} \frac{K}{\sqrt{K^2(n_1 + n_2)^2 + n_1^2n_2^2 - \left(\frac{n_1^2 + n_2^2}{2}\right)}}. \quad (15)$$

The response at the minimum or valley, which is at the resonant frequency, is obtained from (9) by setting $\Delta f/f_0 = 0$ and is

$$V_v = \frac{I}{\omega_0 \sqrt{(C_1 + C_m)(C_2 + C_m)}} \frac{K}{K^2 + n_1n_2}. \quad (16)$$

The peak-to-valley ratio is, therefore,

$$\left(\frac{V_v}{V_r}\right) = \frac{K^2 + n_1n_2}{\sqrt{K^2(n_1 + n_2)^2 + n_1^2n_2^2 - \left(\frac{n_1^2 + n_2^2}{2}\right)}}. \quad (17)$$

What we desire, in so far as design is concerned, is the values of the decrement $n$ (or $Q$) and the coefficient of coupling $K$ required to give a certain peak-to-valley ratio. By combining (14) and (17), we obtain

$$\frac{n_1 + n_2}{2} = \beta \left(\frac{\Delta f}{f_0}\right) \quad (18)$$

where

$$\beta = \sqrt{1 - \frac{1}{2^n \left[\left(\frac{V_r}{V_v}\right)_{2^n}^2 - 1\right]}}. \quad (19)$$

where the subscript 1 is to show that this is the peak-to-valley ratio of one double-tuned stage. Equation (18) is one of the desired design equations and shows that the required average of the decrements of the primary and secondary is fixed only by the peak-to-valley ratio desired and the percentage bandwidth.

The smaller we desire the peak-to-valley ratio to be (thus the flatter the response is in the pass band) the larger $\beta$ becomes and, therefore, the greater must be the average decrement, i.e., the lower must be the $Q$. From (18), we find that $\beta$ varies between the values of 1.75 to 0.42 as the peak-to-valley ratio varies respectively between the values of 1.01 to 1.50.

Now, making use of (14) and (18), we obtain for the required coefficient of coupling

$$K = \frac{\Delta f}{f_0} \sqrt{1 + \beta^2 \frac{2(1 + D^2)}{(1 + D)^2}} \quad (20)$$

where $\Delta f$ is the resonant frequency bandwidth between points equidistant from the resonant frequency $f_0$. $D$ is the double-tuning ratio (usually less than 0.1), and $f_0$ is the resonant frequency of the double-tuned stage.
where \( \beta \) is given by (22) and \( D \) is the ratio of the primary \( Q \) to the secondary \( Q \).

\[
D = \frac{n_2}{n_1} = \frac{Q_1}{Q_2}.
\]

Thus, we see that the coefficient of coupling required is fixed mainly by the percentage band pass desired and is also dependent (not to a great extent, however) on the ratio \( D \) of primary \( Q \) to secondary \( Q \). Equation (20) is the second of our desired design equations.

Dividing equation (20) by equation (18), we obtain

\[
\frac{K}{(n_1 + n_2)/2} = \frac{\sqrt{2(1 + D^2)}}{(1 + D)^2} + \frac{1}{\beta^2}.
\]

(21)

This is a very useful equation because it does not involve frequency. It shows that as soon as the peak-to-valley ratio (i.e., \( \beta \)) and the \( Q \) ratio are fixed, then the ratio of the coefficient of coupling \( K \) and the average decrement \((n_1 + n_2)/2\) is also fixed, and conversely, for a given circuit where the \( Q \) ratio and the ratio of the coefficient of coupling and the average decrement is fixed, the peak-to-valley ratio is fixed. It should be understood that the \( Q \) ratio \( D \) has an almost second-order effect; for the quantity \( 2(1 + D^2)/(1 + D)^2 \) is equal to unity when the \( Q \) ratio is unity, and approaches a maximum value of two when the \( Q \) ratio approaches either zero or infinity.

The next design equation desired is one that will give the output voltage or the gain of the circuit at the peaks of the response. By substituting the design conditions given by equations (18) and (20) in the equation giving the response at the peaks, which is (15), we obtain

\[
V_p = \frac{1}{\beta} \times \frac{I}{4\pi \Delta f_p \sqrt{(C_1 + C_m)(C_2 + C_m)}} \times \sqrt{2(1 + D^2)} + \frac{1}{\beta^2}.
\]

(22)

and for the usual case, where the constant-current generator of value \( I \) is a vacuum tube, \( I = G_m E_p \), and we have

\[
\text{Gain}_{\text{per stage}} = \frac{1}{\beta} \times \frac{G_m}{4\pi \Delta f_p \sqrt{(C_1 + C_m)(C_2 + C_m)}} \times \sqrt{2(1 + D^2)} + \frac{1}{\beta^2}.
\]

(23)

Design equation (23) brings out several points of interest with reference to the gain which is obtained with "flat-topped" band-pass circuits. We see that the gain depends directly on the \( G_m \) of the tube used and inversely on the numerical band width desired between peaks \( \Delta f_p \). The midfrequency has no effect on the gain (as long as the bandwidth \( \Delta f_p \) is a small percentage of the midfrequency \( f_0 \)). The gain is also inversely proportional to the square root of the product of the total capacitance across the input or output circuits that must be resonated. We see also that the gain is inversely proportional to the factor \( \beta \) which is given by (19) and which is a measure of the flatness of response in the flat top pass band. The flatter the pass band, the lower the gain obtainable. Finally, the gain depends on the square root of a quantity involving the ratio of primary \( Q \) to secondary \( Q \).

This square root has only an almost second-order effect on the gain. It is interesting, however, to see the effect of this \( Q \) ratio on the gain. If \( Q_1 \) equals \( Q_2 \), the factor under discussion becomes unity. If \( Q_1 \) is made infinite and all the loading is done on the primary side, we obtain

\[
\sqrt{\frac{1 + 2\beta^2}{1 + \beta^2}}.
\]

and, if \( Q_1 \) is made infinite and all the loading is done on the secondary side, we again obtain

\[
\sqrt{\frac{1 + 2\beta^2}{1 + \beta^2}}.
\]

(24)

In most practical designs, \( \beta \) will have a value close to unity; therefore, if all loading is done on one side of the band-pass circuit, approximately 25 per cent more gain per stage will be obtained, as compared to the case where the primary and secondary are equally loaded (i.e., \( Q_1 = Q_2 \)).

It may be mentioned here that practical considerations dealing with ease of circuit alignment, and "Miller effect" detuning, lead to the conclusion that in many cases it is better to make \( Q_1 = Q_2 \) and thus sacrifice the above 25 per cent additional gain per stage. These points will be discussed later.

The next desired design equation is concerned with the shape of the circuit response outside the pass band, i.e., the skirt selectivity. By combining (9), giving the response at any frequency, and (15), giving the response at the peaks, and (18) and (20) giving the required circuit constants, we obtain the ratio of peak response \( V_p \) to the response \( V \), at any band-width \( \Delta f \),

\[
\left( \frac{V_p}{V} \right)_1 = \sqrt{1 + \left[ \frac{(\Delta f/\Delta f_p)^2 - 1}{2\beta^2(1 + \beta^2)} \right]^2},
\]

and solving (24) for \( \Delta f/\Delta f_p \), we obtain

\[
\frac{\Delta f}{\Delta f_p} = \sqrt{1 + 2\beta^2(1 + \beta^2\sqrt{(V_p/V)^2 - 1}}
\]

(25)

where the subscript 1 is to show that the ratios are the voltage ratios for one double-tuned stage.

This is the last of our desired design equations and we see that the larger \( \beta \) is made (therefore, the flatter the response inside the pass band) the wider are the skirts at any skirt-response point; i.e., skirt selectivity becomes poorer as the pass-band response is improved. It should be noted from (24) or (25) that, for a given peak-to-valley ratio (i.e., a given \( \beta \)), the shape of the response curve is independent of the ratio of primary \( Q \) to secondary \( Q \).

The plus or minus sign in (25) should also be noted.
When the plus sign is used, we obtain the skirt bandwidths outside the response peaks, and when the minus sign is used, we obtain the bandwidths inside the peaks of the response curve.

To make analysis as complete as possible, the phase of the response voltage with respect to the driving current should also be given. By combining (10) for the phase shift with design equations (18) and (20), we obtain

\[
\tan \theta_{(\text{prestage})} = \pm \left[ 1 + 2B^2 - (\Delta f/\Delta f_p)^2 \right] \left[ 2B(\pm \Delta f/\Delta f_p) \right].
\]

In (26) the top sign is used in front of the numerator and denominator when \((K_1 - K_1)\) is plus, i.e., with a net capacitive coupling. (It should be remembered that these equations should not be applied to the region in the vicinity of the null given by (13).) The plus sign is used inside the bracket in the denominator for the frequencies above the resonant frequency and the minus sign is used for the frequencies below the resonant frequency.

From (26) we can see that, for inductive coupling, the phase shift at the midfrequency (i.e., \(\Delta f = 0\)) is \(-90\) degrees and at the low-frequency peak, the tan of the phase angle is \((-+/+)\beta\) and at the high-frequency peak, the tan of the phase angle is \((-/-)\beta\). Since, in many applications, satisfactory flatness in the pass band is given when \(\beta\) is approximately unity, we see that the phase shift at the low-frequency peak is usually approximately \(-45\) degrees and the high-frequency peak usually has a phase angle of approximately \(-135\) degrees.

With capacitive coupling, we see that the phase shift at the midfrequency is \(+90\) degrees; the tan of the phase angle at the low-frequency peak is \(+/-\) \(\beta\); the tan of the phase angle at the high-frequency peak is \(+/-\) \(\beta\), and for \(\beta\) equal approximately to unity the phase shift at the low-frequency peak is thus approximately \(+135\) degrees, and at the high-frequency peak it is approximately \(+45\) degrees.

It should be noted that for a given peak-to-valley ratio (i.e., a given \(\beta\)), the phase shift is independent of the \(Q\) ratio.

VII. SMALL-PERCENTAGE PASS-BAND DESIGN EQUATIONS WHEN \(Q_1 = Q_2\)

Design equations having even a small degree of complexity are, in many cases, not used by engineers. However, conveniently used graphical representations of the complex equations will usually be put to use.

In their most usual use, identical band-pass circuits are cascaded to produce intermediate-frequency-amplifier chains. Various applications may necessitate the use of from one to possibly eight cascaded stages. It would appear worthwhile to develop an exact, quickly used graphical method of designing the cascaded circuits so that they produce a specified response shape.

Since the number of cascaded stages used must be one of the design parameters, consideration of (18), (20), (21), (22), (25), and (26) shows that some form of family-of-curves representation or its equivalent is necessary.

We further note that (20) and (22) are complicated by the relatively second-order effect of the \(Q\) ratio which would necessitate an almost useless family of curves. Because of this complication, we will consider the case where \(Q_1 = Q_2\) in the graphical method of design; and the equations themselves can be used directly when \(Q_1\) does not equal \(Q_2\).

There are two important practical reasons why a design using \(Q_1 = Q_2\) should be used whenever possible. The first reason is concerned with the problem of aligning cascaded flat-topped band-pass circuits. The second reason is concerned with the detuning effect caused by the fact that the input and output capacitance of a pentode changes with gain-control setting due to plate-to-grid capacitance feedback (Miller effect), and space-charge effects.

With reference to the alignment of cascaded flat-topped circuits, if \(Q_1\) is made equal to \(Q_2\) the circuits can be aligned just as single-peaked or single-tuned circuits are aligned, i.e., by using a single-frequency signal generator (not a "sweeper"), and tuning for absolute maximum output. With double-peaked circuits, the signal generator is set at the frequency at which the low peak of the response is desired and all the circuits are tuned lower in frequency for maximum response. (Or the signal generator may be set at the frequency at which it is desired to have the high-frequency peak, and all the circuits are then tuned higher in frequency for maximum response.) It can be shown that if \(Q_1\) equals \(Q_2\), equal absolute maxima of response are obtained at the peaks only when both circuits are tuned to the same resonant frequency (as described in Section 3) and, conversely, when both circuits are tuned to the same resonant frequency, absolute maximum (and equal) response is obtained at both peaks (as long as there is no loss in the mutual reactance). It is this fact which is the basis of the method of alignment just described.

When \(Q_1\) does not equal \(Q_2\), tuning of the circuits to produce an absolute maximum of response at one frequency would necessitate the two circuits being tuned to different resonant frequencies and the two peaks are then of different amplitudes.

With reference to the second reason for making \(Q_1\) equal to \(Q_2\), it is desirable to have a response curve which is not affected when the gain (i.e., the \(G_n\) of the amplifier tube) is changed. Unfortunately, the change in input and output capacitance of a pentode with changing \(G_n\) (due to plate-to-grid capacitance feedback and space-charge effects) detunes the resonant circuits. However, it can be shown that with \(Q_1 = Q_2\) a slight detuning of the resonant circuits will have a practically negligible effect on the symmetry of the response curve. Thus, although the response curve as a whole will move slightly as the gain control is changed, the shape of the curve will remain sensibly constant when \(Q_1 = Q_2\).

When circuits are cascaded, the voltage responses at a given frequency are multiplied together to give the resultant voltage response. When the cascaded circuits are all identical it is obvious that to obtain the resultant
For the case of \( Q_1 = Q_n \), the design equations then become as follows:

\[
\beta = \sqrt{1/2} \left[ \frac{(V_p/V_r)^{1/2N}}{\sqrt{(V_p/V_r)^{2/N} - 1}} \right].
\]  

(19a)

\[
\frac{Q}{f_0/\Delta f_p} = 1
\]

(18a)

\[
KQ = \sqrt{1 + 1/\beta^2}
\]

(21a)

\[
\left( \frac{V_p}{V_r} \right) = \left\{ 1 + \left[ \frac{(\Delta f/\Delta f_p)^2 - 1}{2\beta^2 - 1 + 2\beta^2} \right]^{1/2} \right\}^{1/2}
\]

(24)

\[
\text{Gain per stage} = \frac{G_n/4\pi \Delta f_p/(C_1 + C_n)(C_2 + C_n)}{\beta}
\]

(23a)

\[
\frac{\Delta f}{\Delta f_p} = \sqrt{1 + \beta^2 + 1/\beta^2}\sqrt{(V_p/V_r)^{2/N} - 1}
\]

(25a)

\[
\tan \theta_{\text{resistance}} = \pm \left[ 1 + 2\beta^2 - (\Delta f/\Delta f_p)^2 \right]^{1/2}
\]

(26)

\[ \\Box ]}

\[ \text{VI. FORMATION OF TRIPLE-TUNED BAND-PASS CIRCUITS}^{4,7} \]

Any two of the circuit configurations shown in Figs. 1 and 2 may be connected in series to form a triple-tuned three-node band-pass circuit (this also means, of course, that one of the circuits shown can be used twice). Similarly, the circuit configurations of Figs. 4 and 5 can be used to form three-mesh band-pass circuits.

With respect to the calculation of the two equal coefficients of coupling which appear in the resulting triple-tuned circuit, maximum gain will be obtained if the following procedure is used: the middle resonant circuit formed when two of the node networks of Figs. 1 and 2 are connected in series should be considered to be formed from two identical resonant circuits in parallel (i.e., each one having twice the net inductance and one half the net capacitance). The input resonant circuit is then coupled to one of the above resonant circuits and the output circuit is coupled to the other resonant circuit.

The middle resonant circuit formed when two of the
case. It is difficult to obtain an inductance of this $Q$. However, if the midfrequency of the 400-kilocycle pass band were shifted down to 4 megacycles, the required $Q$ of the input and output circuits would then be about 10, and the necessary middle-circuit $Q$ would be at least 100. This $Q$ can be obtained without too much trouble. Thus, if triple-tuned band-pass circuits are to be used, it would be worth while choosing a 10 per cent, or even greater, bandwidth.

As in the double-tuned case, the high-impedance or node circuits will be considered to be used the most, and therefore the specific analysis will be made using three-node circuits having both inductive and capacitive coupling, as shown in Fig. 6. It should be clearly realized, however, that the resulting analysis applies exactly to all of the myriad triple-tuned networks which can be formed from the networks of Figs. 1 and 2 and 4 and 5.
IX. Exact Triple-Tuned Response Equation

The node equations which apply to the triple-tuned circuit of Fig. 6 are:

\[ I = \left\{ G_1 + j \left[ \frac{\omega(C_1 + C_m)}{\omega(L_1 L_{n_1}/L_1 + L_{n_1})} \right] \right\} V_1 - j \left( C_m - \frac{1}{\omega L_{n_1}} \right) V_2 + 0 \]

\[ 0 = -j \left( \frac{C_m}{\omega L_{n_1}} \right) V_1 + \left\{ G_2 + j \left[ \frac{\omega(C_2 + C_{m_1} + C_{m_2})}{\omega \left( \frac{L_{n_1} L_{n_2} L_{n_3}}{L_{n_1} L_{n_2} L_{n_3} + L_{n_1} L_{n_2}} \right) \right] \right\} V_2 - j \left( \frac{C_{m_2}}{\omega L_{n_2}} \right) V_3 \]

\[ 0 = 0 - j \left( \frac{C_{m_2}}{\omega L_{n_2}} \right) V_2 + \left\{ G_2 + j \left[ \frac{\omega(C_2 + C_{m_2})}{\omega \left( \frac{L_{n_1} L_{n_2} L_{n_3}}{L_{n_1} L_{n_2} L_{n_3} + L_{n_1} L_{n_2}} \right) \right] \right\} V_3. \]
Introducing the resonant frequency (as defined in Section IV) and the coefficient of coupling and the decrement, we obtain from (27) the complete, exact solution for the magnitude and phase of output voltage.

\[
\frac{V_3}{I/\omega_0 \sqrt{(C_1 + C_m)(C_2 + C_m)}} = \sqrt{\left(\frac{\Delta f_{\text{max}}}{f_0}\right)^2 - 2(K^2 - n^2)(\Delta f/f_0)^3} - 1/3(K^2 - n^2) = 1/3 \left(\frac{\Delta f_{\text{max}}}{f_0}\right)^2
\]

and from (30) the locations of the maxima and minima are given by

\[
\frac{\Delta f_{\text{max}}}{f_0} = 0 \quad \text{and} \quad (K^2 - n^2)
\]

\[
\frac{\Delta f_{\text{min}}}{f_0} = 1/3(K^2 - n^2) = 1/3 \left(\frac{\Delta f_{\text{max}}}{f_0}\right)^2
\]

\[
\frac{1}{2} K^2
\]

\[
\tan \theta = \frac{\pm \Delta f}{K^2 + n^2 - \left(\frac{\Delta f}{f_0}\right)^2}
\]

\[
\tan \theta = \frac{\left(\frac{\Delta f}{f_0}\right)}{K^2 n - 2n \left(\frac{\Delta f}{f_0}\right)^2}
\]

X. SMALL-PERCENTAGE BAND-PASS
DESIGN EQUATIONS

Applying the reasoning used in Section V of the double-tuned analysis, we will consider the small-percentage band-pass case, i.e., where \(\omega/\omega_0\) becomes only about 10 per cent greater or less than unity. We thus have the two great simplifications:

\[
K = (K_e - K_L) \quad \text{and} \quad \left(\frac{\omega}{\omega_0} - \frac{\omega}{\omega_0}\right) = \frac{\Delta f}{f_0}
\]

Setting the derivative with respect to

\[
F \left(\frac{\Delta f}{f_0}\right)
\]

of (28) equal to zero, we obtain for the location of the maxima (plus sign) and minima (minus sign)

\[
\Delta f_{\text{max}} = 0
\]

\[
\frac{\Delta f_{\text{max}}}{f_0} = 2/3 \left[K^2 - \frac{n_1^2 + n_2^2 + n_3^2}{2}\right] + 1/3 \sqrt{K^4 - K^2(n_1^2 + n_2^2 + n_3^2)(n_3 + 4n_2 + 3n_1 + 3n_2)} - (n_1^2n_2 + n_2^2n_3 + n_3^2n_1)
\]

We will obtain the design equations for the case where the Q of the input and output circuits are the same (\(n_1 = n_2 = n\)) and the middle resonant circuit Q is much greater than the Q of the input and output circuits (\(n_3 \ll n\)).

For this case, the general response (28) becomes the relatively simple equation

\[
\frac{V_3}{I/\omega_0 \sqrt{(C_1 + C_m)(C_2 + C_m)}} = \sqrt{\left(\frac{\Delta f_{\text{max}}}{f_0}\right)^2 - 2(K^2 - n^2)(\Delta f/f_0)^3} + (K^2 - n^2)^2(K^2/f_0^2) + K^2 n^2
\]

Introducing the location of the outside peaks (30a) into the peak-to-valley ratio (33), we can solve for the decrement which is required to produce a desired peak-to-valley ratio with a given percentage bandwidth between outside peaks. This is our first design equation:

\[
\frac{Q}{f_0/\Delta f_{\text{max}}} = \frac{1}{\gamma}
\]

\[
\frac{1}{2} K^2
\]
Chart III—Triple-tuned band-pass circuit design.

where

\[
\gamma = \frac{\left[ \frac{1 + (V_p/V_c)_1}{\sqrt{(V_p/V_c)^2 - 1}} \right]^{1/2} + \left[ \frac{1 - (V_p/V_c)_1}{\sqrt{(V_p/V_c)^2 - 1}} \right]^{1/2}}{\sqrt{3}} \tag{35}
\]

and, using (30a) we have as the equation giving the coefficient of coupling which is required in order to obtain a given peak-to-valley ratio with a given percentage bandwidth between peaks

\[
\frac{\Delta f}{\Delta f_P} = \sqrt{1 + \gamma^2} \tag{36}
\]

where \(\gamma\) is given by (35).

Multiplying (34) by (36), we obtain our second design equation:

\[
KQ = \sqrt{1 + \left(1/\gamma^2\right)} \tag{37}
\]
The next desired design equation is the one giving the skirt selectivity. Substituting the design equations (34) and (36) into the response equation (28a), we obtain the response at any point in terms of the peak-to-valley ratio (represented by \( \gamma \) of (27)) and the percentage bandwidth at the outside peaks. Dividing the result by the response at the peaks given by (24), we obtain the equation giving the skirt-response ratios in terms of the skirt bandwidth.

\[
\left( \frac{V_p}{V} \right)_t = \sqrt{1 + \left[ \left( \frac{\Delta f}{\Delta f_p} \right) \left( \frac{\Delta f_p}{\Delta f} \right) - 1 \right]^2} \quad (38)
\]

Solution of (38) for \( \Delta f/\Delta f_p \) gives

\[
\frac{\Delta f}{\Delta f_p} = \left[ d + \sqrt{d^2 - 4/27} \right]^{1/2} \quad (39)
\]

where

\[
d = \gamma(1 + \gamma^2)\sqrt{(V_p/V_t)^2 - 1}
\]

The above equations apply to a single triple-tuned stage. When \( N \) stages are cascaded and a resultant peak-to-valley ratio of \( V_p/V \) is desired, then the peak-to-valley ratio of each stage \( (V_p/V_s) \) must equal \( (V_p/V_t)^{1/N} \). Similar reasoning applies to the skirt-response ratio, so that \( (V_p/V_s) = (V_p/V_t)^{1/N} \).

Application of the above reasoning gives the following design equations for \( N \) cascaded triple-tuned circuits, where the input and output resonant circuits in each stage are of equal \( Q \) and the \( Q \) of the middle resonant circuit is much higher than that of the input and output circuits.

Let

\[
\gamma = -\frac{[1 + (V_p/V_s)^{1/N} - 1]^{1/3}}{\sqrt{3}} + \frac{[1 - (V_p/V_s)^{1/N} - 1]^{1/3}}{\sqrt{3}} \quad (35a)
\]

Then

\[
\frac{Q}{f_0/\Delta f_p} = \frac{1}{\gamma}
\]

\[
KQ = \sqrt{1 + 1/\gamma^2}
\]

\[
\frac{V_p}{V} = 1 + \left[ \left( \frac{\Delta f}{\Delta f_p} \right) \left( \frac{\Delta f_p}{\Delta f} \right) - 1 \right]^{N/2} \quad (38a)
\]

\[
\frac{\Delta f}{\Delta f_p} = \left[ d + \sqrt{d^2 - 4/27} \right]^{1/2} + \left[ d - \sqrt{d^2 - 4/27} \right]^{1/2} \quad (39a)
\]

where \( d = \gamma(1 + \gamma^2)\sqrt{(V_p/V_s)^2 - 1} \)

\[
\tan \theta = \frac{\pm \Delta f_p}{2\gamma^2 + 1 - \left( \frac{\Delta f}{\Delta f_p} \right)^2} \quad (41)
\]

\[
\frac{Gain_{stir}}{G_m/4\pi\Delta f_p\sqrt{(C_1 + C_m)(C_1 + C_m)}} = \frac{1}{\gamma} \quad (40a)
\]
\[
\tan \theta_{\text{per stage}} = \frac{\left( \pm \frac{\Delta f}{\Delta f_p} \right) \left[ 2\gamma^2 + 1 - \left( \frac{\Delta f}{\Delta f_p} \right)^2 \right]}{-\gamma \left[ \gamma^2 + 1 - 2 \left( \frac{\Delta f}{\Delta f_p} \right)^2 \right]}, \quad (41a)
\]

From (35a), (34), (36), (39a), and (40a), another set of nomographs has been prepared. From the phase-shift equation (41), a family of curves has been prepared.

The procedure for using these nomographs and curves is identical with the procedure given in Section 11 for the double-tuned nomographs. The reader should refer to the examples given in that section.

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**ACOUSTICS AND AUDIO FREQUENCIES**

S34.323  

S34.321.9.001  

S34.6:621.395.61  

S34.626.370.153  
Report of the Commission on Loudspeakers—D. E. Shorter. (B.B.C. Quart., vol. 1, pp. 54–57; October, 1946.) The frequency response curve taken after the interruption of the test note should mark resonances which are not present in the steady state. Tests were made on four types of cones with delays of 10 to 40 milliseconds and the response was determined by measuring the rate of decay of the sound. Total cancellation, gliding and other irritating effects were found to lie associated with the additional resonances. See also Wireless World, vol. 2, pp. 421–423; December, 1946.

S34.626.323  
Loudspeaker Transient Response—D. E. Shorter. (B.B.C. Quart., vol. 1, pp. 54–57; October, 1946.) The frequency response curve taken after the interruption of the test note should mark resonances which are not present in the steady state. Tests were made on four types of cones with delays of 10 to 40 milliseconds and the response was determined by measuring the rate of decay of the sound. Total cancellation, gliding and other irritating effects were found to lie associated with the additional resonances. See also Wireless World, vol. 2, pp. 421–423; December, 1946.

S34.626.328  
Large Electromagnetic Installations—J. Müller-Strooband. (Schweiz. Elektro., vol. 12, pp. 193–194; January 1947.) A general description of equipment manufactured by the Albawerk Zurick A-G. and particularly suitable for outdoor events, for erecting halls, works, etc. A special feature of the system is automatic control of the output volume by means of a voltage derived from a microphone which picks up the noise in the room or hall where the loudspeakers are situated.

S34.626.323  

S6.305.667  
Three-Band Variable Equalizer—L. D. Grignon. (Electronics, vol. 20, pp. 112–115; January, 1947.) Provides gain or attenuation adjustment in one-decibel steps independently on high-, and mid-frequency bands of the audio spectrum. Applications include recording, re-recording, sound system compensation, and broadcast equipment adjustment. For a fuller account see 11/10 of 1946.

**AERIALS AND TRANSMISSION LINES**

S6.392.537.291  
Electronic Amplifier Formed by a Guided Wave in a Medium of High Dielectric Constant—R. Wallauschek. (Comp. Rend. Acad. Sci. (Paris), vol. 224, pp. 191–193; January 20, 1947.) A wave of the longitudinal electric type (E) is propagated in a cylindrical guide in which almost the whole cross section is filled with a dielectric of high permittivity. The phase velocity of the wave is thus much less than that of light. Around the electric axis of the guide is an evacuated cylinder traversed in the direction of propagation of the wave by a beam of electrons of velocity very near the phase velocity of the wave. A solution is obtained of the problem of the interaction of wave and beam in the guide. Formulas are given for the progressive waves. Four possible reflected waves are found. Two of these exist for a small range of beam velocities around the phase velocity of the primitive wave; one has increasing and the other decreasing amplitude. These two waves have a phase velocity less than the velocity of the beam. The other two waves are of constant amplitude, one progressive and the other retrogressive, and their phase velocity is greater than the electron velocity. By combining these four waves, the limiting conditions at the ends of the guide can be satisfied. Cf. 1330 below (Lapeyre and Lapostolle).

S6.319.029.62+621.396.029.62  
Wide-Band Aerials and Transmission Lines for 20 to 85 Mc/Sec—F. E. Lutkin, R. J. J. Cary, and G. N. Harding. (Journ. I.R.E. (London), part 11/1, vol. 93, no. 5, pp. 546–548; 1948.) Systems covering the frequency bands 20 to 30, 40 to 50, and 50 to 85 megacycles are described which can deal with pulse transmissions of 600 kilowatts peak power. The aerial arrays are built up of full-wave center fed dipoles of wire-arc construction, with an input impedance of 600 ohms. Open wire transmission lines are used throughout. Wire mesh reflectors are used to obtain the horizontal diagram required; their effect on dipole impedance is discussed. An exponential transformer is described which simplifies the construction of arrays consisting of four full-wave dipoles, its function being to transform...
an impedance of 300 to 600 ohms. Compensating stubs may be used to increase the bandwidth of the wide-band aerials.

621.392.029.64 Waveguide Data—L. E. Sherbin. (Electronics, vol. 20, pp. 122-124; January, 1947.) Curves of the given attenuation and power-carrying capacity as a function of frequency for rectangular copper wave guides of various dimensions operating in the TE0m modes. Frequencies from 1600 to 44,000 megacycles are covered.

621.392.029.64 Propagation in Curved Guides—M. Jouglet. (Compt. Rend. Acad. Sci. (Paris), vol. 224, pp. 107-109; January 13, 1947.) The method of perturbing (349 of 1946 and 16 of February) is applied to the study of Ho. and Hm. waves in a perfectly conducting circular wave guide of radius R to determine those which reduce to Eo. and Em. waves when R increases indefinitely. The El. wave behaves normally, but the Hm. waves can only exist in a perfectly conducting curved guide if in combination with appropriate amplitudes in the ratio \(\sqrt{21}\), so that the transported energy is equally divided. The effect of curvature on the phase velocity of this type of wave is not to the second order, zero. This difference in comport for a cylindrical guide as long as the curvature is finite. In an actual guide, propagation of an \(H_m\) or \(E_{\lambda}\) wave by itself is impossible, provided that the curvature is sufficient, and tends towards zero with the resistivity of the wall.

621.392.029.64 Some Applications of the Principle of Variation of Wavelength in Wave Guides by the Internal Movement of Dielectric Sections—G. M. Dudgeon. (J.I.E.E., London, part II, 1946, vol. 93, no. 4, pp. 633-638; 1946.) This principle is used for loading adjustment of centimeter-wave magnetrons working in a complex load, for beam swinging in directive arrays without mechanical movement, and in switching systems.

621.392.029.64 Discussion on "Wave Guides" [I.E.E. Radio-location Convention]—J. O. H. Elton. (J. I.E.E., London, vol. 93, no. 4, pp. 778; 1946.) Points raised include the effect on performance of replacement of damaged parts, and the effect on standing-wave ratio of a small-frequency shift.

621.392.029.64:5384 Quasi-Stationary Field Theory and Its Application to Diaphragms and Junctions in Transmission Lines and Wave Guides—G. G. Macfarlane. (J. I.E.E., London, part III, 1946, vol. 93, no. 4, pp. 543-546.) Calculations are made of the shunt admittance of capacitive and inductive diaphragms in strip transmission lines and rectangular wave guides by combining quasistatic field theory and Babinet's principle for electromagnetism, a valid result is derived for the case when the diaphragm cross section is comparable with, or greater than, a wavelength.

621.392.029.64:621.317.336.5 Standing Wave Met-—Kullam. (See 1504.)


621.392.029.64:621.306.615.141.2 Problems and Practice in the Production of Guide-Wave Transmission Systems—L. W. Brown. (J. I.E.E., London, part II, 1946, vol. 93, no. 4, pp. 639-646; 1946.) The performance requirements of wave guides are considered in terms of reflections and standing waves. Selection or matching must be adopted, rather than random choice or interchange of sections, when the number of sections exceeds 2 or 3. Particular reference is made to the case of wave guides used with magnetron sources.

621.392.029.64:621.306.615.141.2 Attenuation Curves for 2:1 Rectangular, Square and Circular Wave Guides—E. O. Willoughby and E. M. Williams. (J. I.E.E., London, part II, 1946, vol. 93, no. 4, pp. 723-724; 1946.) Graphs are given which show the relationship between frequency and physical size for various values of attenuation in copper wave guides.

621.392.029.64:621.306.615.141.2 Calculation of Attenuation in Wave Guides—S. Kuhn. (J. I.E.E., London, part II, 1946, vol. 93, no. 4, pp. 663-678; 1946.) Tables and curves are derived which give the field equations and attenuation constants of rectangular wave guides in and circular wave guides and may be used to match with practice. Field equations are expressed in terms of field impedances and of the power transmitted by the wave by the products of "characteristic impedance" of energy. The attenuation constant caused by wall losses is tabulated for the case of an air-filled copper guide. The attenuation constant and phase constant are also tabulated for the case of an enclosed dielectric of less, i.e., for values of tan \(\delta\) below 0.1.

621.392.1:51283 Matrix Methods in Transmission-Line and Impedance Calculations—W. H. Watson. (J. I.E.E., London, vol. 93, no. 4, pp. 737-746; 1946.) An ordered expansion of methods applicable not only to calculations of the principal wave on a two-conductor transmission line, but also to all other wave processes capable of representation in terms of transmission lines. Some new results are also indicated.

621.392.2 The Progression Between a Progressive Wave and a Beam of Electrons of Velocity Near That of the Wave—A. Blanch-Laguette and P. Lapostolle. (Compt. Rend. Acad. Sci. (Paris), vol. 224, pp. 104-105; January 13, 1947.) For a thin line made up of discrete equal sections, the wave amplitude increases exponentially and the wave velocity is slightly lower than the beam velocity. For an infinite, uniform, continuous line, four waves are possible, only one of them increasing in amplitude. Such a line can be regarded as a model explaining qualitatively the phenomena of interaction in progressive-wave amplifiers.

621.392.41 Losses in Transmission Lines—A. W. Gent and J. W. Wallis. (J. I.E.E., London, part II, 1946, vol. 93, no. 3, pp. 539-563; 1946.) The lines considered are those such that the multiple of \(\pi/2\). If both the outer and inner conductors are tapered, there is an optimum taper which will give unity standing-wave ratio (s.w.r.) for \(\pi/2\) sections and a further advantage of different propagation from unity for other lengths. If it is desired to keep the diameter of one conductor the same on both sides of the junction, the method is to taper the two conductors in opposite ways for a half wavelength and then in the same way, thus producing either a bulge on the inner conductor or a hollow on the outer conductor. A summary of this paper is given in part II, vol. 93, no. 1, pp. 58-69; 1946.

621.396.611.029.5 Theory of Mode Separation in a Coaxial Wave Guide—D. J. Sutro. (Proc. I.R.E. and Waves and Electrons, vol. 34, pp. 960-962; December, 1946.) An analysis of the separation of the first and third modes in a coaxial wave guide. It is shown that the difference between the two resonant lengths of one line which gives these modes increases with the difference between the products of the terminating interelectrode capacitance and the characteristic impedance for the two lines.


621.396.67 Slot Aerials and Their Relation to Composite Wire Aerials (Babinet's Principle)—H. C. Booker. (J. I.E.E., London, part II, 1946, vol. 93, no. 4, pp. 620-626; 1946.) Babinet's principle of complementary screens is applied to resonant slots in thin plane conducting screens. A \(\lambda/2\) slot, with an input impedance of about 145 ohms, is shown to be similar to that of a \(\lambda/2\) dipole but with the directions of vibration of electric and magnetic fields interchanged. Resonant slots may be used to form linear or broadband arrays, to produce polarized waves and for band-pass filters or any similar device used in conjunction with wire aerials. With the aid of Babinet's principle, several of these problems can be reduced to problems whose solutions are already known.

621.396.67 Slot Feeders and Slot Aerials—C. E. G. Bailey. (J. I. E.E., London, part II, 1945, vol. 93, no. 2, pp. 485-499; December, 1945.) A discussion of the properties of combinations of parallel strips, slots, and gapformed by cutting narrow strips out of infinite plane conducting sheets. Equivalence and inversion theorems are established for slots and strips and for interconnected slots and wires. These theorems are used to give an approximate solution for the radiation field of a half-wave slot.

621.396.67 The R.C.A. Antenna—An Instrument Useful in the Design of Directional Antenna Systems—G. H. Brown and W. C. Morgan, J. E. E. and Waves and Electrons, vol. 34, pp. 960-962; December, 1946.) An entirely electrical instrument for deriving the radiation pattern of an aerial or characteristics of a rectilinear array which would give a desired radiation pattern. The application is mainly to broadcast arrays with up to several hundred elements, each element being characterized by its position and phase and magnitude of its current. These parameters are separately controllable by means of variable capacitances. Radiation patterns are presented as point-spectrums, on cathode-ray tube in either polar or rectangular co-ordinates.

621.396.67:621.317.336.5 The Measured Impedance of Cylindrical
Dipole—D. D. King. (Jour. Appl. Phys., vol. 17, pp. 844–852; October, 1946.) The im-
pedance characteristics of a half-dipole and image plane were measured by a resonance
method which is suitable for high
damping a standing-wave-ratio method sup-
plemented the resonance data. The results are
shown graphically and are in good agreement
with those of Brown and Woodward (2267 of
1945).

621.366.57:621.397.5

Line-of-Sight Aerials [Antennes de Vision]
—R. Tabard. (Télé. Franc., nos. 9 and 11,
pp. 22–23 and 19, 24; January and March,
1946.) Discusses aerials of the Hertz half-wave
and medium wave types and corre-
spending feeder systems.

621.366.670.02:621.396.029.62

The Use of a Common Aerial for Radar
Transmission and Reception of 200 Mcs—
C. J. Banwell. (Jour. I.E.E. (London),
part II A, vol. 93, no. 3, pp. 567–574; 1946.)
An account of the development of a com-

621.366.670.029.62:621.396.029.62

The Design and Positioning of Aircraft
Radar Aerials for Metric Wavelengths—B.
Russell. (Jour. I.E.E. (London), part II A,
vol. 93, no. 3, pp. 567–574; 1946.) A discus-

tion of the various problems involved in connection
with aircraft aerials for all-round looking, hom-

621.366.670.029.62:621.396.029.62

The Use of Spherical Reflectors as Micro-
Wave Scanning Aerials—J. Ashmead and A. B.
Pippard. (Jour. I.E.E. (London), part II A,
vol. 93, no. 3, pp. 627–632; 1946.) The cond-

621.366.670.029.64

Dielectric Housings for Centimetre-Wave
Antennae—J. B. Birs. (Jour. I.E.E. (Lon-
don), part II A, vol. 93, no. 4, pp. 647–657;
1946.) Expressions are derived for the re-
fection and transmission coefficients of plane waves when incident normally or obliquely on two
parallel semi-infinite dielectric discon-
tinuities. The dielectric constant of a variety of dielectric housings (radomes) at 10 and 3 centimeters are
discussed, with reference to the effect on aerial
gain, impedance, polar diagram, and polariza-

621.366.671.014.1

and Dielectric Absorption and Aerial Short-
ening of the Transmitter-Dipole—J. Müller-
Strehlow. (Bull. Schweiz. Elektrotech. Ver.,
vol. 37, no. 24, pp. 710–714; 1946, reprint, in

German, with French summary.) The theory of
sustained oscillations is applied to a trans-
mitting dipole aerial (a) neglecting and (b)
taking account of aerial losses. Formulas
described in detail.

621.366.671.029.64:621.317.34

The Influence of Re-Radiation on Mea-
surements of Standing-Wave Ratio in an
(Jour. I.E.E. (London), part II A, vol. 93,
no. 4, pp. 720–722; 1946.) A modification of
Purcell's method is given, taking account of
re-radiation which frequently introduces con-
siderable error. In the direct method of meas-
uring power gain, re-radiation errors are prac-
tically negligible.

621.366.676.529.13

Cavity Aircraft Antennae—H. Kees and F.
January, 1947.) Description, with polar dia-

621.366.677

Elimination of Errors from Crossed-Dipole
Direction-Finding Systems—R. A. Smith and
C. Holt Smith. (Jour. I.E.E. (London),
part II A, vol. 93, no. 3, pp. 573–587; 1946.) The
main sources of error in the early types of
'chain home for low-flying aircraft' apparatus are
discussed. The two most serious sources of error
are the feeder and reflector systems. In later
apparatus, aerials were designed to match the transmission lines, reflector systems to
give adequate sense discrimination, reliable
operation and minimum errors, and transmis-
sion systems to give maximum stability and
energy transfer with minimum errors. Diffi-
culties at the receiving end were overcome by
the inclusion of an intermediate transmission of
a lattice-type phase-shifting network followed by
an impedance matching unit. Very careful
bonding of the transmission lines was found
necessary. Formulas for calculating the vari-
ous procedures are outlined. A summary of this paper is
given in part III A, vol. 93, no. 1, pp. 55–
60; 1946.

621.366.677

Theoretical Treatment of Short Yagi
Aerials—W. Walkinslaw. (Jour. I.E.E. (Lon-
don), part II A, vol. 93, no. 3, pp. 598–614; 1946.)
Design data are given for four arrays with
driven half-wave dipoles and various
arrangements of parasitic radiators. Polar
diagrams and variation of radiation resistance
curves, as a function of the self-reactance of
the radiators, are given for each array system.

621.366.677

The Gain of an Idealized Yagi Array—D. G.
Reid. (Jour. I.E.E. (London), part II A,
vol. 93, no. 3, pp. 566–566; 1946.) An expression
is derived for the gain of an impressed with
a doublet, of an end-fire array having an infinite
number of infinitely closely spaced elements,
the currents in successive elements being
suitable for practice but progressively and uni-
formly retarded in phase. Curves show the
variation of gain with the over-all length of the
array for a number of fixed wavelengths.
The overlap of these curves enables the
maximum gain for a given length of array and
the corresponding value of the phase velocity
to be found.

621.366.677

The Gain of a New Parabolic Beam
p. 40; April, 1943.) The modification consists in
increasing the length of each half-wave
section in a single array by equal small admittances at opposite ends of the wave-guide spacing
are discussed.

621.366.677.029.5

(Proc. I.R.E., vol. 31, pp. 40–41; January,
1947.) Constructional details of a four-
element, close-spaced aerial array for use on
frequencies of 28 megacycles and above,
the beam in each half-wave section can be
very easily adjusted.

621.366.677.029.6

A New Broadband Aerials with Applica-
tions to 200 Mc/s Ground Radiolocation
Systems—D. Taylor and C. J. Westcott. (Jour.
I.E.E. (London), part III A, vol. 93, no. 4,
pp. 693–699; 1946.) A method of determining
the length of each half-wave
section can be very easily adjusted.

621.366.677.029.62

On Some Problems of the Theory of
Highly-Directional Antenna Arrays—S. Tset-
baum. (Jour. Phys. (London), vol. 10, no. 3,
p. 282–292; 1946.) The properties of broad-
side arrays of elementary dipoles excited in
phase are considered for the case in which
curvature of the wavefront is important, i.e.,
when the problem is concerned with the mutual
coupling of a single dipole and a broadside
array. It is shown that for a plane array, there is
an optimum form for maximum transmis-
tion between dipole and array. Spherical broadside
arrays, in which individual dipoles are
centered on the surface of a sphere having the single
dipole at its center, are considered. In this case
the determination of energy takes place and there
are no optimum dimensions. Formulas for radi-
ation resistance are given.

621.366.677:621.392.029.64

The Design of a Wave-Guide-Fed Array
of Slots to give a Specified Radiation Pattern—
A. L. Cullen and H. K. Goward. (Jour. I.E.E.
(London), part III A, vol. 93, no. 4, pp. 683–
692; 1946.) A method of obtaining any specified
aperture distribution from a wave-guide-fed array of
slots. The method of obtaining such a specified radiation pattern can then be solved by
using the well-known Fourier transforma-
tion. Experimental results confirm the theory.

621.366.677:621.392.029.64

A Wide-Band Linear Array Aerial—L. H.
Dawson and N. M. Rust. (Jour. I.E.E. (Lon-
don), part II A, vol. 93, no. 4, pp. 693–699;
1946.) This 10-centimeter array consists of
half-wave dipole elements spaced 4a/4 apart in
a 3 by 1-inch wave guide, with alternate ele-
ments of metal sheet from the center line. For
an 8-foot array the beam width is about 2.5
degrees for half power in the horizontal plane,
with the largest side lobe 1 percent of the main
beam. When used with a cylinder of parabolic
reflector, the array produces a beam width of
about 7 degrees in the vertical plane. The beam
shape remains almost constant over a 10 percent
frequency shift. The effects of the mutual
coupling, change of line length, and of
characteristic admittance in a feeder loaded by
equal small admittances at equal powers spacing
are discussed.
of a mobile ground-controlled-interception system are given and also of a fixed ground-controlled-interception station, including a specially designed rotary capacitance switch capable of handling peak powers up to 150 kilowatts.

**CIRCUITS AND CIRCUIT ELEMENTS**

621.302.5:621.302.9

The Application of Riemann's Number Sphere and Its Projection to A. C. Engineering—F. Steinor. (Tele-Tech, vol. 1, pp. 23-26; October, 1946.) Stereographic projection of the number sphere leads to a better representation of resistances than that using Gauss's map; transfer from resistance to conductance merely requires a rotation through 180 degrees. A parallel projection method is developed for obtaining the terminating resistances or the terminal voltages of two quadrupoles in series.

**General Theory of the Autotransformer—P. Thévenin.** (Radio en France, no. 2, pp. 37-38; 1947.) The quadrupole equivalent of a auto-transformer and associated amplifiers, for use in short-circuit conditions, is derived. The T scheme is found particularly suitable for determining short-circuit voltages.

621.317.715

Contact Modulated Amplifier—Perkin-Elmer Corporation. (Jou. Math. Phys., vol. 25, pp. 261-278; January, 1947.) Methods are given for determining quantitatively the extent to which message and noise can be separated and for the design of a filter to effect this separation. The problem of simultaneous filtering is considered. The root-mean-square approach used is an approximation to and a simplification of the transcendental case developed by N. Wiener.

621.317.835

On the Stability Conditions of Oscillating Systems—Couffignal. (Sect 1497.)
enables the modes of vibration of a cavity made up from one with a plane side and its mirror image in that plane side, to be deduced from those of the former. The equivalence between the two of Maxwell’s equations and a principle of minimum energy analogous to the principle of least action enables the natural frequencies and the fundamental fields of a cavity to be studied by the methods of the calculus of variations. In this way the variations in natural frequencies caused by small deformations of the cavity wall can be determined. The principle of orthogonal trajectories can also be applied.

621.396.611.4.029.64 1379 Resonance Cavities used for Ultra-Short Waves—A. Briot. (Trés. Fr. Acad., no. 8, pp. 7–10; December, 1945.) A mathematical treatment, based on Maxwell’s equations, of the general case of cavity resonance, with application to the determination of the natural resonance frequencies of cylindrical cavities.

621.396.615 1380 Complete Theory of Valve Oscillators—D. Guitton. (Rev. Sci. (Paris), vol. 84, pp. 165–168; August, 1946.) The grid and anode currents are assumed to be sinusoidal. An approximate formula for the wavelength is derived mathematically better with results of experimental theory than that of classical theory which neglects the grid current.

621.396.615 1381 A Stabilized Modulated Oscillator—A. E. Hayes, Jr. (Radio News, vol. 37, pp. 32–33; January, 1947.) Spurious frequency-modulation in an amplitude-modulated oscillator is eliminated by a secondary circuit producing frequency-modulation which is controlled and switched to oppose it.

621.396.615.14 1382 Rf Oscillators for U.H.F. Transmission—Goette. (See.)

621.396.615.17:621.317.755 1383 New Timebase Circuit for Cathode-Ray Oscillographs—(Elektroch. Zeit., vol. 65, pp. 138–139; April 20, 1944.) An auxiliary circuit, with properly chosen time constant, enables the capacitor connected across the time-deflection plates of the cathode-ray oscilloscope to be charged with sensibly uniform current. A modification of the arrangement gives a linear single sweep.

621.396.619.11/13 1384 Theory of the Frequency Discriminator—P. Göttinger. (Bull. Schweiz. Elektrotech. Ver., vol. 37, pp. 531–534; September 7, 1946, reprint, in German, with French summary.) A general theory is developed. The conditions for linearity are considered from a new point of view. The main problem investigated is the modulation of frequency-modulation waves, with particular reference to the transformation of frequency modulation into alternating modulation with the least possible distortion.

621.396.619.23 1385 Overmodulation without Sideband Splitter—Villard. (See.)

621.396.645:621.43.019.8 1386 Constant-Gain Knack Pickup Amplifier—Krems and Dallaire. (See.)

621.396.662.3 1387 Introduction of the Idea of the Coupling Quadrupole—L. Bé. (Radio en France, no. 2, pp. 29–31; 1947.) A coupling quadrupole is any quadrupole with zero open-circuit impedance and infinite short-circuit impedance. It is the basis of nearly all the simplifications possible in systems of quadrupoles of the general type.

621.396.662.3 1388 General Theory of Decimal Attenuators—

621.396.662.3 1389 Components—M. Chauvière. (Radio en France, no. 6, pp. 47–49; 1947.) A general critical discussion of present design trends in various countries for tubes, circuit components, tuning units, and loudspeakers.

621.396.692.029.3 1390 Notes on the Construction of Attenuator Resistors—F. Tourney. (Radio en France, no. 2, pp. 32–33; 1947.) Economical methods of construction for all audio frequencies are described.

621.396.694.011.3/4 1391 Study of Electronic Reaction-Variation Devices—W. Maxel. (Ténis la Radio, vol. 14, pp. 34–48; March, 1946.) Reaction modification of a multihole tube circuits are described whose behavior for reactance variation depends solely on a proper choice of the values of the resistors and capacitors involved. The operation of such circuits is explained with the aid of graphs and the results obtained with particular circuits are discussed.

GENERAL PHYSICS

53.081 1392 On Units [units] and Dimensions—H. B. Dorgelo and J. A. Schouten. (Proc. Acad. Sci. (Amsterdam), vol. 49, pp. 123–131 and 282–291; February and March, 1946.) Discussion of the analogy of absolute and metric quantities, of the relative merits of rationalized and unrationaized, and of the dimensional and non-dimensional tables. Tables are given of units, dimensions, and the field equations using the centimeter-gram-second and Gauss and Gaussian systems.


530.145.65 1395 An Interpretation of L. de Broglie’s Equations for the Photon—R. Murdai. Physical Interpretation of L. de Broglie’s Equations for the Photon—R. Murdai. (Compl. Rend. Acad. Sci. (Paris), vol. 222, pp. 1030–1032 and 1075–1076; April 29, and May 6, 1946.) Relativistic theory of systems of articles shows that the photon can be considered as a system of two Dirac corpuscles whose relative motion round their common center of gravity is evanescent. Regarding the photon as a Dirac corpuscle of positive energy combined with a lumina in the series of states of negative energy, the relativistic wave mechanics of systems of particles completely justifies L. de Broglie’s photon theory.
Energy Distribution of Electrons in High Frequency Gas Discharges—T. Holstein. (Phys. Rev., vol. 70, pp. 367-384; September 1-15, 1946.) An equation for the energy distribution is obtained and the limitations of its application are defined. Methods of solution and examples are given.


Magnetic Field in Permeable Media and Definition of Magnetic Moment—H. Dieseldorff. (Elektrotech. Z., vol. 65, pp. 119-122; April 6, 1947.)

Induced Magnetization and Magnetic Moments—E. Brylinski. (Compt. Rend. Acad. Sci. (Paris), vol. 222, pp. 1035-1037; April 29, 1946.) Consideration of the effects produced in a uniform field, when a magnetic field is applied, shows that the moment of the couple exerted by a uniform field \( H \) on a plane closed current is the algebraic sum of the moment caused by the field \( H \) and that due to the field \( \vec{\mu} \), which results from the reaction of the material, the magnetic moment remaining independent of the material. On the assumption that the magnetic moment is due to moving electric charges, it is concluded that the volume integral definition of magnetic moment, which leads to a contradiction, should be abandoned.

Applications of the Riesz Potential to the Theory of the Electromagnetic Field and the Meson Field—N. E. Frenberg. (Proc. Roy. Soc. A, vol. 188, pp. 18-31; December 31, 1946.) Riesz' method "yields simple deductions of concrete results... and also the results recently obtained by Dirac regarding the proper energy and proper momentum of an electron." Bhabha's analogous theorem of the neutral meson field is also treated.

The Distribution of Currents Induced in a Plane Wave on the Surface of a Conductor—V. Fock. (Jour. Phys. (U.S.S.R.), vol. 10, no. 2, pp. 130-136; 1946.) If the wavelength is small compared with the dimensions and radius of curvature of the conductor, the current distribution near the geometrical shadow can be expressed in terms of an universal function, which is tabulated.


In general the dielectric constants and the permeabilities alone determine whether Maxwell's equations for this problem have a unique solution or no solution; the conductivities are only involved in certain limiting cases.


Comparison of the dimensions and radii of curvature of the conductor, the current distribution near the geometrical shadow can be expressed in terms of a universal function, which is tabulated.


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Abstracts and References


579.511.535
The Mechanism of Ionospheric Ionization: Part 2—R. v.d.R. Woolley. (Proc. Roy. Soc. A., vol. 187, pp. 403-415; December 13, 1943. "The mechanisms for the production of electrons in the three regions of the ionosphere are discussed with special reference to the question whether it is possible to account for the observed electron densities without supposing that the sun emits far more energy in the remote ultraviolet spectrum than would be emitted by a black body at 6000°K. The consequences of the mechanisms made by metastable states of atoms and molecules are examined. It is concluded that the observed electron densities may be accounted for without requiring high energy in the ultraviolet if the effective recombinant coefficient in the F2 region is 10^-4. The F2 region is supposed formed by the ionization of atomic oxygen, and the E region by the ionization of molecular oxygen. The electrons forming the F1 region are supposed to be provided by metastable N or O by "NO. For part 1 see 416 of March."

551.510.535:521.396.11
The Role of the Ionosphere in the Propagation of Radio Waves—R. Jouann. (Bull. Soc. France Éléc., vol. 6, pp. 346-354; June-July, 1946.) A survey of present knowledge. Regular daily observations on the ionosphere have been carried out at the National Radio Laboratory, Bagneux, since April, 1946. The program envisions the extension of the observations to all the hours of the day and night and the progressive provision of further recording stations in France overseas.

514.594
A New Theory of Atmospheric Electricity—D. S. Kohari and L. S. Lhotzy. (Sci. Culture, vol. 13, pp. 261-263; December, 1946.) A discussion of Frenckel's theory (2186 of 1946) showing that it offers a qualitative explanation of some puzzling phenomena. It may have wider application to dust storms in the electrification of powders injected into gases.

515.594:551.574

614.825:551.594.221

LOCATION AND AIDS TO NAVIGATION

621.396.677

621.396.0:623.454.25
Radio Frequency-Trade Development—H. Man and Brunetti. (See 1535.)

621.396.93

621.396.027.0:621.396.664
Radio Controlled Boats—A. F. Hopkins, Jr., and F. H. Scaborough. (Electronics, vol. 20, pp. 84-86; January, 1947.) A simple very-high-frequency remote control selective relay system, originally used for rapid blackout of unattended boats. Its application is being extended to faghorns and similar navigational aids only required occasionally.

621.396.063:621.38.001.8
Analyzing Present Position of Electronic Aides for Airplanes—G. Siele. (Telc-Tech, vol. 6, pp. 34-41, 133; January, 1947.) A discussion of the merits of the problems encountered in flight together with recommended electronic solutions which have been offered the PICAO delegates. See also 112 of February and 428 of March.

621.396.96

621.396.95

621.396.05:518.5
The Ballistic Computer—Juley. (See 1494.)

MATERIALS AND SUBSIDIARY TECHNIQUES

535.37:535.61-15

538.221.029.64
Theory of the Dispersion of Magnetic Permeability in Ferromagnetic Materials at Microwave Frequencies—C. Kittel. (Phys. Rev., vol. 70, pp. 281-290; September 1-15, 1946.) An explanation of the experimental facts is proposed, based on a consideration of the equations of motion of a domain boundary in an applied magnetic field for frequencies such that the skin depth of the magnetic field is smaller than the thickness of the domain. A criticism is given of theories of ferromagnetic resonance.

546.287

546.467.854.82

549.514.51:548.24
Artificial Electrical Twinning in Quartz Crystals—J. J. Vormer. (Tijdschr. Ned. Radiogrool., vol. 11, pp. 215-219; November, 1946. In Dutch with English summary.) Electrical twinning can easily be produced below 573 degrees centigrade at well-defined places in quartz and GT-cut quartz plates. Such twinning may occur at points where leads have been soldered to the metal coatings. In some cases this type of twinning can be corrected by suitable heat treatment, but attempts to correct crystal electrically-twinned quartz met with little success.

520.193.21:669.721
Metals: Magnesium: Corrosion Resistance under Accelerated Atmospheric Conditions—A. R. Rogers, D. A. Tetu, and H. Livingstone. (Metal Ind., vol. 70, pp. 9-10; January 3, 1947.) Magnesium and its alloys offer good resistance to corrosion except in marine atmospheres, while protective coatings of paint are desirable.

621.032.53:533.5
Glass-to-Metal Seals—G. D. Redston and J. E. Stanworth. (Jour. Soc. Glass Tech., vol. 29, pp. 48-76; April, 1945.) Stresses-optical and photoelastic measurements on standard sandwich seals over a wide temperature range are discussed. Axial stresses for head seals at room temperature were determined by the method of Hill and Burger.

621.032.53:533.5
Glass-to-Metal Seals, with Particular Reference to Current Lead-In Seals in Vacuum Devices—R. W. Douglas. (Jour. Soc. Glass Tech., vol. 29, pp. 92-110; April, 1945.) A procedure for avoiding extreme stresses in manufacture or operation is described.
Electrical and Electronics—(Electrician, vol. 137, pp. 1279–1280; November 8, 1946.) A short account of the various features to be incorporated in the A.C.E. (automatic computing engine) designed by the National Physical Laboratory.

621.317.35
The Cinematic Analyzer—E. Aebischer. (Radio Craft, vol. 18, pp. 18–19, 66; December, 1946.) This frequency counter resembles a receiver and is used to tune to 410 kilocycles, a converter stage, an oscillator stage, and an output stage which are synchronized with the supply to the voltage amplifier. Applied frequencies between 410 and 510 kilocycles thus appear as peaks on the screen, the amplitude of the peaks indicating the signal amplitude. In addition to this direct application to intermediate-frequency stages, the apparatus can also be used for the study of receiver radio-frequency amplifiers, selectivity curves, and audio-frequency stages.

621.317.35
Wave and Pulse Counter—R. Blitzler. (Radio Craft, vol. 18, pp. 25, 49; December, 1946.) The incoming wave is amplified, clipped, and reduced to a frequency which is applied to the grid of a thyratron.

621.317.205.60
Apparatus for Measurement of Centimetre Waves: Q-Meter and Wattmeter—A. G. Clavier and R. Cabessa. (Onde Électr., vol. 44–49, November, 1946.) In the Q-meter, a waveguide grid tube, with anode voltage varied periodically (e.g., at 210 cycles), is used to create a variable electromagnetic field; this in turn excites the resonant cavity, the Q of which is to be measured. A crystal detector is used to measure the field in the resonant cavity, the detector current being applied after amplification to the cathode of a cathode-ray oscilloscope; the horizontal sweep is derived from the voltage used for modulation. The resonance curve thus obtained enables Q to be found. Q = ΔW/ΔV, where ΔV is the mean frequency variation, corresponding to the peak of the resonance curve, and ΔW is the width of the curve where the amplitude is half the maximum. A double-beat method is also used.

The apparatus described enables measurements of Q to be made from 2000 to about 50,000, with an accuracy of the order of 10 per cent. It has also been used for measuring, at wavelengths of 8 centimeters, the attenuation constants of coaxial cables or of circular guides with Eo or Ho waves. The wattmeter comprises (a) a measurement line of characteristic impedance 90 ohms and of adjustable length, which is inserted between the ultra-high-frequency source and the apparatus to be measured; (b) a horizontal deflecting deflecting deflection line, which is a rack-mounted measurement equipment. The length of the line and the position of the bolometer are adjustable to obtain a system of static stable points. Powsers from 50 milliwatts to 150 watts can be measured with the apparatus on wavelengths between 3 centimeters and 5 centimeters with an accuracy of about ±1 per cent.

621.317.33:621.383.5.032.2
1501
Inter-Electrode Capacitances of Triode Valves and Their Dependence on the Operating Currents of the Cathode-Ray Oscillograph—R. Kragh. (Indian Jour. Phys., vol. 20, pp. 81–99; June, 1946.) Inter-electrode capacitances are measured by a double-heap method and their variation with filament and anode current for three different anode voltages with no grid bias are studied. Results for eight commercial tubes are given and discussed.

621.317.33:621.383.6.072
1502
The Measured Impedance of Cylindrical Dipoles—King (See 1338.).

621.317.33:621.392.029.64
1503
Standing-Wave Meter—H. E. Kallman (Electronics, vol. 20, pp. 90–96; January, 1947.) A detailed description of an automatic standing-wave meter of moderate accuracy designed for quick adjustments of microwave equipment. Power from a constant source is injected into a U-shaped wave guide. Power reflections caused by mismatch are measured by a device which gives a direct meter indication of standing-wave ratio.

621.317.35

621.317.7.082.78

621.317.7.082.78.085.31
1511
Electronic Position Pickup—D. W. Moore, Jr. (Electronics, vol. 20, pp. 100–101; January, 1947.) An earth-inductor pickup for measuring a magnetic field in some permanent magnet mounted on the pointer whose position is to be transmitted.

621.317.7.082.78.6.1504
1512
Contact Multiplexed Amplifier—Perkins-Elei-mer Corporation. (Rev. Sci. Instr., vol. 15, p. 569, 1944.) The frequency characteristic of the amplifier described in 2629 of 1946 (Liston, Quinn, Sargent, and Scott). It is designed to replace sensitive galvanometers; various applications are indicated.

621.317.7.085.39
1513

621.317.7.087.25
1514
A Very High Impedance R. M. Veitmeter for Beam Testing—D. C. Gall and F. C. Widdas. (Jour. Sci. Instr., vol. 23, p. 287; December, 1946.) A tube amplifier, with high-input impedance, which “has no voltage primary, has high input impedance, giving in effect a very high impedance to the alternating-current voltmeter which it operates. The ratio of input to output is linear.”

621.317.7.087.25.015
1515
Alternating Current Potentiometer—S. Heywood (Eriac Rev., vol. 23, no. 4, pp. 297–307; 1946.) The required alternating-current quantities are transformed thermally into direct voltages, which are measured with direct potentiometers. Circuit diagrams are given for power, current, and voltage measurement.

621.317.7.087.3
1516
A Visual Null Indicator for Impedance Bridge Measurements at Radio Frequencies—P. J. Boeing and W. H. Whitehead. (Rev. Sci. Instr., vol. 17, pp. 537–539; December, 1946.) Greater accuracy in balance is obtained by applying the detector amplifier output to one pair of points of a cathode-ray tube, and the modulated output of the radio-frequency oscillator to the other pair. A straight-line trace indicates balance. Using this indicator, a bridge of 916A bridge, impedance measurements to an accuracy of 0.2 ohm can be made in the presence of interference levels more than 40 decibels above 1 microvolt.

621.317.7.092.66
1517
Television Synchronizing Signal Generating Units—Part I—A. E. Backhouse. (Telech, vol. 6, pp. 50–54; 144; January, 1947.) Describes picture and synchronizing test equip-
ment for studio, laboratory, and receiver-produc-
tion lines.

621.317.794 1518
The Organic Thermistor Bolometer—C. D.
A, pp. 93-102; November, 1946.) Tests of a bolom-
eter using a cellulose film painted with aquadag
show it to be much inferior to the inorganic
thermistor developed at the Bell Telephone
Laboratories, particularly as regards drift and
speed of response.

621.306.614.5:538.560.4 1519
Theory of a Microwave Spectroscope—W.
September 1-15, 1946.) A discussion of (a) the
measurement of the exponential decay of the
radiation in an untuned echo box between pulses
of radio-frequency power, and (b) the steady-
state response of a large untuned cavity.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

518.5 1520
Calculations and Electronics—(See 1943.)

537.553.7 1521
Imperfections of Shape in Electron-Optical
Instruments—F. Berstein (Comp. Rend. Acad.
Sci. (Paris), vol. 224, pp. 106-107; January 13,
1947.)

539.16.08 1522
Design of a Proportional Counter for
vol. 17, pp. 533-536; December, 1946.)

539.16.08 1523
A Universal Radiation Measurement Ap-
paratus, Its Description, Operation and Possi-
ble Applications—Reiter. (Rev. Instrument.
Kde, vol. 64, pp. 103-121; April-June, 1944.)
An instrument for use with counters, having a
special stopcock arrangement suitable
not only for single inaudible counts but for
reading the relative values of large radiation
quantities from galvanometer deflections. Various
applications include: the recording of the intensity character-istics of X-rays and of all
kinds of corpuscular radiation.

550.837:621.39 1524
The Development of Electrical and Radio
Methods of Geophysical Prospecting—V.
139-150; April 10-24, 1946.) A survey of present-
day direct-current audio- and high-frequency
methods, with applications to ore, coal, and
oil prospecting; water detection; investigation of building sites; and lightning protection.

621.317.39:531.768 1525
64, pp. 30-46; January-March, 1944.)

621.317.708.784.058.31 1526
Electronic Position Pickup—More. (See 1111.)

621.317.704.535.61-15 1527
Fast a Superconducting Bolometer—D. 1.
September, 1946.) For detection of infra-
red signals. A ribbon of BiN is used at an
operating temperature of about 15 degrees
Kelvin which can be maintained with the aid of
liquid hydrogen and nitrogen for several hours.
The primary response time is about 5X10-4 seconds at about 3000 cycles while the noise
level is 5X10-4 microvolts. The approxi-
mates has a secondary re-response time of 5X10-8
seconds at about 140 cycles.

621.318.572 1528
The German-Guiller-Muller Counters—Graves. (See 1364.)
1. Such as butterfly wings, onion skin, etc., interposed between the lead and a sheet of photographic paper.


3. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.


6. The Propagation of Electromagnetic Waves in Two or More Successive Media and the Diffraction of These Waves Referred to the Study of Cauchy's Problems—Rohin. (See 1413.)

7. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

8. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

9. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

10. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

11. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

12. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

13. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

14. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

15. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

16. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

17. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

18. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

19. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.

20. R. P. Krebs and T. Dallas. (Electronics, vol. 20, pp. 87-89, January, 1947.) Cathode-follower input, special feedback circuit, and simplified phase-inverter stage provide a flat frequency response from 5 to 20,000 cycles with a gain of 100,000 for portrayal of knock patterns of internal combustion engines on a cathode-ray oscilloscope.
Vertical and Horizontal Deflection in Television Systems—J. A. Widlaman. (Télégr. Franç. no. 8, pp. 2-4; December, 1945.) An abec is given from which the two definitions can be found for any system of television, given the size of lines and of images and the cut-off frequency.

Considerations on the Bedford Velocity-Modulation Television System—P. J. Froulon. (Télégr. Franç. no. 8, p. 11; December, 1945.) For original paper of Bedford and Puckee see 1944 abstracts, p. 506.

Two Systems of Color Television—D. G. Fink. (Electronics, vol. 21, pp. 72-77; January, 1947.) A general discussion on the relative merits of the sequential and simultaneous systems. Both provide good fidelity of color transmission without flicker if bandwidth, frame frequency, etc., are suitably chosen. As simultaneous color transmission can be reproduced in black-and-white by existing receivers, the transition to color television will be easier on this system. For examples of each system see 2051 and 2363 of 1946 and 1240 of May.


TV [Television] on Modulated Light-Beam—(Télé-Tech, vol. 6, pp. 96, 98, January, 1947.) Describes the transmission of video signals over short distances by means of a light-beam carrier system employing a cathode-ray tube as the modulated light source and a multiplier-type photocell receiver.

Line-of-Sight Aerials [Antennes de Vision]—Tabard. (See 1339.)

Television Transmission Centre Paris—H. Delaby. (Télégr. Franç., no. 11, pp. 3-4; March, 1946.) A general account of the present Eiffel Tower equipment, with proposed extensions to relay stations.


The Main Types of Faults in a Television Receiver—R. Aeghen. (Télégr. Franç., no. 9 and pp. 4-5, 21 and 6; January and February, 1946.)

Large Television Screens and Their Evolution—Hendalryver. (Télégr. Franç., no. 11, pp. 17-18, 27; March, 1946.) Outlines the development of large-surface multiscan screens.


Carrier-Difference Reception of Television Sound—R. B. Dome. (Electronics, vol. 20, pp. 102-105; January, 1947.) Use of a common intermediate-frequency amplifier for video and sound signals eliminates the effects of local oscillator hum and frequency drift. The over-all intermediate-frequency bandwidth is greater than the carrier frequency bandwidth, and by use of the absorption trap circuits, the sound intermediate-frequency level is kept below the minimum level expected from the picture carrier. The two circuits are in parallel separated. A receiver incorporating this system is reliable in performance, and costs are reduced.

Television Synchronizing Signal Generating Units: Part I—Baxter. (See 1517.)

TRANSMISSION

361.316.726.078.3: 361.396.613.14

Carrier Stability in Frequency-Modulated Transmitters—G. Guanella. (Brown Bosler Mitt., vol. 33, pp. 193-197; August, 1946.) Several automatic control methods are briefly described, with block diagrams. (a) An alternating voltage of lower frequency is derived by heterodyning, while the control voltage is obtained for control through a frequency discriminator. (b) Modulation of a stable two-phase voltage by a fraction of the carrier voltage gives a rotating field which is used for re-orienting. (c) Modulation of a two-phase voltage, with differentiating and heterodyning, gives a control voltage proportional to frequency error. (d) The carrier frequency is compared directly with a suitable harmonic of a reference oscillator.

361.394.61

New 10-kW Transmitter for Telegraphy—E. Guery and M. Fabre. (Brown Bosler Mitt., vol. 33, pp. 175-178; August, 1946.) The frequency range is 410-500 kilocycles with stability to one part in 107. The transmitter is designed for 3-phase 230/380 volts 50-cycle supply and is capable of 450 words per minute. A transmitter attachment permits use for telephony.

361.396.61: 361.395.61

Moderated i-kW Transmitter—E. Meili. (Brown Bosler Mitt., vol. 33, pp. 172-174; August, 1946.) A description, with illustrations, of a short-wave transmitter for wavelengths of 12.8 to 90 meters and adaptable for either telephony or television. A Franklin oscillator with quartz control gives frequency stability of +7.10^-5.

361.61: 361.394.619.13


361.61: 361.029.54/58

The R.A.F. T. 1154 Transmitter—(Short Wave Mag., vol. 4, pp. 3, 119-123; October, 1946.) A brief description of a 1000 watt telegraph and telephone aircraft transmitter with three frequency ranges: 200 to 500 kilocycles, 3.0 to 5.5 megacycles, and 5.2 to 10 megacycles.

361.61: 361.029.58

A 7-1/4-m/M Transmitter—(Short Wave Mag., vol. 4, pp. 811-813; October, 1946.) General description and performance of a 70-watt amateur transmitter. See also Short Wave Mag., vol. 4, pp. 312-313; July, 1946.

58: 361.029.58

Five-Band 25-Watt Transmitter—B. Rendel. (Short Wave Mag., vol. 4, pp. 694-695; January, 1947.) A combination of crystal oscillator and power amplifier which can be used either as a low-power transmitter or as an efficient exciter in a power radio-frequency amplifier. Data are included for operation on wavebands from 1.7 to 28 megacycles.

361.396.610.295.63

Decimetre-Wave Transmitter Giving 50-W Aerial Power—R. Schweizer. (Brown Bosler Mitt., vol. 33, p. 227; August, 1946.) A general description, with photographs but with no operational details of a transmitter, with no stabilized anode and heater voltages, suitable for field strength measurements.

361.61: 361.029.64

Projectors of Centimetre Waves—H. Gutter. (Onde Élec., vol. 26, pp. 459-466; December, 1946.) Calculation of the radiation from projectors is based on Hughes’ principle. Three types of projectors are discussed: electro-magnetic lenses, reflectors, and dielectric aerials as used by the Germans during the war.

361.615.14

Ring Oscillators for U.H.F. Transmission—T. Gootèe. (Radio News, vol. 37, pp. 48-50, 121; January, 1947.) An even number of 4 or more tritides of the same type are arranged in a ring and tuned by resonant lines to obtain greatly increased power output, 16 triodes giving about 8 times that of a pair in push-pull.

361.619.13: 361.619.712

N.B. September Transmitters—Bell Lab. Rev., vol. 25, pp. 20-23; January, 1947.) Three transmitters are illustrated and briefly described, rated at 1, 3, and 10 kilowatts respectively. The last consists of the 1-kilowatt, 2125, together with power amplifiers. A frequency synchronization system is employed to control the carrier frequency.

361.619.23

Overmodulation without Sideband Splitter—O. G. Hard Jr. (Electronics, vol. 20, pp. 90-95; January, 1947.) Full circuit details of a balanced modulator for incorporation in an amplitude-modulation phone transmitter to produce amplitude modulation in excess of 100 per cent without causing adjacent channel interference.

VACUUM TUBES AND THERMONIQUES

537.291+538.691

The Paths of Ions and Electrons in Non-Uniform Crossed Electric and Magnetic Fields—N. D. Coggeshall, (Phys. Rev., vol. 70, pp. 176-190; September 1-15, 1946.) The force equations for a charged particle moving in such fields can be integrated by a very simple procedure under certain conditions. These conditions are satisfied when motion takes place in a median plane symmetrically situated relative to magnetic pole faces and electrostatic electrodes. Numerical integration can be used when the analytical difficulties are too great, or when the fields are only known empirically. A summary was noted in 3882 of 1945.

537.291: 536.615.142

Theory of Small Signal Bunching in a Parallel Electron Beam of Rectangular Cross Section—E. Eeimberg and D. Feldman. (Jour. Appl. Phys., vol. 17, pp. 1025-1037; December, 1946.) The non-uniform distribution of the bunch density along a field which operates the bunching process so that a kinetic solution may be validly applied only for a limited length of the drift space. An accurate solution of the bunch density requires the integration of the dynamical and field equations which, for ‘small-signal’ conditions, reduce to a linear homogeneous system. These conditions are satisfied when the fields are non-sonoidal or nonsonoidal according to the motion produces or does not produce a high-frequency charge density within the beam. This general problem is solved by a suitable linear combination of both solutions, a large part of the solution for practical conditions being of the nonsonoidal type.

536.383: 536.616.822

Calculated Frequency Spectrum of the Shot
Noise from a Photo-Multiplier Tube—R. D. Sard. (Jour. Appl. Phys., vol. 17, pp. 768-777, October, 1946.) A general expression for the power spectrum of the shot noise produced by a secondary-emission multiplier tube is applied to the Radio Corporation of America 931 family. It is deduced that the noise intensity should be constant from zero up to about 100 megacycles, then fall off appreciably between 100 and 1000 megacycles, and become very weak at higher frequencies.

621.385

Beam Production in Radial Beam Tubes, Beam Splitter Tubes, and Other Low Voltage Electronic Devices—A. M. Sklinski. (Rev. Mod. Phys., vol. 18, pp. 379-383; July, 1946.) In the magnetic-focus radial-beam tube, the beam is focused entirely by an external magnetic field of between 50 and 250 gauss and in the power tube by the action of the grid wires. Other tubes operating at 300 volts or less include the "magic eye" tuning indicator and the orthicon pickup tube in television systems.

621.385-621.372:518.6

Electrostatic Field Plotting—Balchowycz. (See 1493.)

621.385:217.7

Portable Precision Valve Tester—Haas. (See 1509.)

621.385.1:321.01:1.2

Impedance of Gas-filled Tubes Traversed by a High-Frequency Discharge—P. Menage. (Compt. Rend. Acad. Sci. [Paris], vol. 219, pp. 55-56; July 10, 1944.) Measurements were made with the tubes placed axially in inductance coils tuned to the frequency in use. Two types of discharge were found. In one the resistance is sensibly independent of the exciting field, the inductance is a decreasing function of the field. The imaginary part of the impedance may have either sign and is equivalent to a capacitance for gas and to an inductance, decreasing with increasing field, for metallic vapors.

621.385.18:029:64

Physical Processes in the Recovery of TR Tubes—H. Marganis, F. L. McMillan, Jr., L. H. Dearlery, C. S. Pearall, and C. C. Montgomery: (Phys. Rev., vol. 70, pp. 380-387; September 15, 1946.) The techniques are described for the measurement of the time of diminution of ions on termination of the discharge in TR tubes. The capture of electrons by gas molecules is found to be the principal factor in recovery.

621.385.3

Development of a Water-Cooled Transmitting Triode of 50-kW Anode Dissipation—F. Jenny. (Electrical World, vol. 33, pp. 211-214; August, 1946.) The cathode consists of 12 taut-glass wires, arranged for singles, threes or sixes, phase alternating-current heating. The anode is of special electrolytic copper. Details of construction, test procedure, and results are described.

621.385.3:029:63

A High-Power Triode for 600 Megacycles—S. Frankel, J. J. Gruber, and J. P. Wattenberg. (Proc. I.R.E. and WAVEs and Elec. Trans., vol. 34, pp. 989-991; December, 1946.) A maximum of 126 kilowatts peak power is developed by 25 kilowatts peak pulse power at 600 megacycles and of a water-cooled version for generating continuous wave power up to 500 kilowatts at the same frequency.

621.385.3:032:2:621.317:33

Inter-Electrode Capacitances of Triode Valves and Their Dependence on the Operating Condition—Mitra and Khastgir. (See 1501.)
Regulated Power Supplies

**MODEL 106-PA**
Characteristics:
- D.C. Voltage Range: 200-300V, 140 Ma.
- A.C. Fil. Power: (2) 6.3V, 5 amps.
- Ripple Content: 1/10 of 1%
- A.C. Input: 115V, 50/60 cycles
- Size: 5" x 19" x 9" deep

Output remains constant within 1%, even though line voltage varies between 95—130 volts. Price $225 (f. o. b. Cambridge, Mass.)

**MODEL 207-PA**
Characteristics:
- D.C. Voltage Range: 0-3500V, 1 amp. positive or negative grounding
- A.C. Input: 220V, 50/60 cycles (Varicor Control)
- Overload Relay: Adj. 0.6—1 amp.
- Size: 26½"x32"x36" deep

Meters on front panel indicate line voltage, output voltage, and output current. Power supply is mounted on casters for portability. Access doors provided with interlock safety switches.

**MODEL 206-PA**
Characteristics:
- Ripple Content: .05 of 1%
- A.C. Input: 115V, 50/60 cycles
- Size: 12½"x19"x13" deep
- Interlocking relay protection at all voltages insures safe operation. Time delay for high voltage circuit applications prevents tube damage. Price $490 (f. o. b. Cambridge, Mass.)

**MODEL 306-PA**
Characteristics:
- D.C. Voltage Range: 300-750V, 30 Ma.
- Ripple Content: 300-750V, 0.01%
- A.C. Input: 750-1800V, 0.1%
- Size: 17½"x19"x13" deep
- Regulation control is provided for adjustment to perfect load regulation, or to provide over-regulation, if desired. Safety devices are incorporated to protect operating personnel. Meters indicate line voltage, output voltage, and output current.

For Every Purpose

Harvey Radio Laboratories, Inc.
456 Concord Avenue • Cambridge 38, Massachusetts
BROADCASTERS—
Simplify
YOUR PROGRAM SWITCHING

EVEN your most complicated program switching operations are reduced to the simple operation of one key—when you use Western Electric's new Relay Type Program Dispatching System. It speeds up the switching involved in serving several destinations with rapidly interchanged studio, line and transcribed programs, auditions and announcements—yet reduces operating errors.

Check these features against your operating requirements:

1. Provides simple, fool-proof method of pre-setting the next scheduled program condition—leisurely—while the present program is "on the air."

2. Operation of a single key, instantaneously switches from the program "on the air" to the pre-set condition.

3. This one-key switching operation can be controlled from either the Master Panel or any selected control booth.

4. During light load periods, control of selected lines may be extended to any studio control booth.

5. "On Air" and pre-set circuit conditions—including point of release control—are positively indicated by lamps at all control points.

6. Any or all programs may be interrupted instantly for "flash booth" announcements without upsetting the existing studio circuit conditions.

7. System may be engineered and furnished to meet your individual operating requirements—regardless of number of program sources or outgoing lines.

For further details, call your local Graybar Broadcast Representative or write Graybar Electric Co., 420 Lexington Avenue, New York 17, N.Y.

Western Electric
QUALITY COUNTS

---

Six-line Master Control Panel for Western Electric Relay Type Program Dispatching System.

Below—Flash Booth Indicator Panel (at left) and Control Signal Indicator Panel (at right).

PROCEEDINGS OF THE I.R.B.
June, 1947
These are the famous Andrew semi-flexible coaxial cables in 1/4 and 3/8 inch diameters (shown in actual size). Because of their better construction and design they are used throughout the world by thousands of broadcast, police, government, and military radio stations as the most efficient device for connecting antenna to transmitter or receiver.
Offer low-cost Magnetic Recording

...Design for Brush Paper Tape

No matter what type of magnetic recorder you design, the low cost, excellent fidelity and uniformity of Brush Paper Tape make it your best all-round recording medium. With this new development by the pioneer and leader in the field of magnetic recording you can bring magnetic recording to the great mass market of all America! Brush Paper Tape will be furnished you either in bulk in varying widths or 1225 ft. ¼-inch wide on a metal reel (standard item).

Look at these advantages of Brush Paper Tape...

- Easy to handle
- Excellent high frequency reproduction at slow speed
- Permanent... excellent reproduction for several thousand play-backs
- Greater dynamic range
- Minimum wear on heads
- Easily erased

Other Brush developments in magnetic recording components include Plated Wire and vastly improved Tape and Wire Recording Heads and Cartridges.

Write today for further information

The Brush Development Co.
3405 Perkins Ave. • Cleveland 14, Ohio

(Continued from page 35A)

PHILADELPHIA

PORTLAND


*The Geiger Counter and its Industrial Applications,* by S. G. Forbes, Graduate Student in Physics, Oregon State College; April 12, 1947.

*Sacramento


SAN DIEGO
*The Latest Developments in Mobile Communication Systems,* by S. Freedman, United States Navy Electronics Laboratory; April 1, 1947.

*Continuation of Discussion of Audio-Distortion Measurements by the Two-Tone Intermodulation Method,* by J. R. McLaughry, United States Navy Electronics Laboratory; April 18, 1947.

SYRACUSE

TORONTO

*Ontario Provincial Police Radio System,* by J. E. Reid, University of Toronto; April 28, 1947.

WASHINGTON

SUB-SECTIONS

FORT WAIN
*Sea Power In The Pacific,* (Sound Film); January 13, 1947.


*Some New Developments in the Electronics Field,* by E. D. Cook, General Electric Company; April 14, 1947; Installation of Officers; April 14, 1947.

HAMILTON
*Industrial Electrochemical Rectifiers,* by J. T. Thwaites, Canadian Westinghouse Company; March 10, 1947.

*German Radio and Electrical Equipment,* by C. G. Lloyd, Canadian General Electric Company; April 14, 1947.

PRINCETON
*Recent Developments in Electromagnetic Recording,* by S. J. Begun, Brush Development Company; April 11, 1947.

PROCEEDINGS OF THE I.R.E.
June, 1947
For over a quarter century the name Meissner has stood for the finest in electronic equipment. Founded in 1922 by William O. Meissner, (famous for his outstandingly successful inventions in communications and electronics) this company has long specialized in the development and manufacture of fine coil equipment for every application... As a result of this vast background of electronic research and experience, Meissner Coils have become the accepted standard among those who demand high quality performance. Precision-made, designed to the most exacting requirements, these superior components are backed by a twenty-five year reputation for quality and uniformity in manufacture.

A complete line, including Air Core I. Fs.
Iron Core Plastic I. Fs. and standard I. Fs.

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Proceedings of the I.R.E.  June, 1947 37A
The Sorensen system of AGelectronic voltage regulation provides quick, accurate response to even the smallest voltage change with a minimum wave distortion and a regulation accuracy of \( \frac{1}{2} \) of 1%.

This same electronic regulation system has been incorporated into the Nobatron, providing a source of regulated DC voltage at currents and stabilities that, in the past, was available only with batteries.

This new source of stabilized DC voltage is obtainable in six standard models operating on a 95-125 AC source of 50 to 60 cycles.

Among the more important uses for Nobatrons are DC ammeter calibration in experimental and quality control laboratories, testing of components in the automotive and aircraft industries in battery-operated relays and in other applications where it is desirable to replace a battery to guarantee continuous regulated power supply.

### GENERAL AC REGULATOR SPECIFICATIONS

<table>
<thead>
<tr>
<th>Input Voltage Range (–1 model)</th>
<th>95-125</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>(–2 model)</td>
</tr>
<tr>
<td>Output Voltage Range (–1 model)</td>
<td>110-120</td>
</tr>
<tr>
<td></td>
<td>(–2 model)</td>
</tr>
<tr>
<td>Load Range</td>
<td>25-30,000 V. A.</td>
</tr>
<tr>
<td>Regulation Accuracy</td>
<td>( \frac{1}{2} ) of 1%</td>
</tr>
<tr>
<td>Harmonic Distortion</td>
<td>5% Max. (2% in &quot;S&quot; Models)</td>
</tr>
<tr>
<td>Input Frequency Range</td>
<td>50-70 cycles</td>
</tr>
<tr>
<td>Inductive Power Factor Range</td>
<td>Down to 0.7 P.F.</td>
</tr>
</tbody>
</table>

For standard voltage regulation, Sorensen Model 500 is a proven leader in its field—compact, accurate and dependable. This model typifies the Sorensen line of AC and Nobatron all-purpose voltage regulators. Let a Sorensen engineer help you with your next voltage regulation problem.

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**SORENSEN & COMPANY, INC.**

375 FAIRFIELD AVENUE • STAMFORD, CONNECTICUT
NEW miniature capacitors

for use with Miniature Tubes

Oil-impregnated paper-dielectric capacitors molded in phenolic

To meet requirements for miniature components for use in hearing aids, pocket radio receivers, airborne radio apparatus, and other devices in which economy of space is a primary factor:

<table>
<thead>
<tr>
<th>Type No.</th>
<th>Capacitance Mfd.</th>
<th>Case Size — Inches</th>
<th>Wire Size</th>
</tr>
</thead>
<tbody>
<tr>
<td>HAC-001</td>
<td>0.001</td>
<td>9/16 x 5/16 x 29/64</td>
<td>3/32</td>
</tr>
<tr>
<td>HAC-005</td>
<td>0.005</td>
<td>9/16 x 5/16 x 29/64</td>
<td>3/32</td>
</tr>
<tr>
<td>HAC-01</td>
<td>0.01</td>
<td>9/16 x 5/16 x 29/64</td>
<td>3/32</td>
</tr>
<tr>
<td>APC-05</td>
<td>0.05</td>
<td>9/16 x 5/16 x 29/64</td>
<td>3/32</td>
</tr>
</tbody>
</table>

Specifications

- Impregnated: mineral oil.
- Case: molded of mica-filled phenolic; sealed to withstand 90% relative humidity.
- Terminal Leads: solid, tinned copper.
- Operating Temperature: -55°C to +65°C; the .001 and .005 Mfd. ratings can be furnished for service up to 85°C at slight additional cost.
- Working Voltage: 75 volts d-c.
- Capacitance Tolerance: +60%, -20%.
Another Astatic FIRST for improving the clarity and beauty of Phonograph Reproduction

MODEL "QT" PHONOGRAPH CARTRIDGE

- With surface noise and needle talk VASTLY reduced by the revolutionary type needle mounting and design of this new cartridge, the proverbial mouse would lose his reputation for quietness by comparison. Increased vertical as well as lateral compliance of the replaceable needle used in the "QT" Cartridge has resulted in a great reduction in acoustic noises, which, together with an extremely low order of distortion, insures clearer, cleaner and therefore more enjoyable "quiet talk" phonograph reproduction. The "QT" Cartridge is being extensively used in new equipment installations. Two models are available, "QT-M" with precious metal-tipped stylus and "QT-J" with jewel point.

CHARACTERISTICS

Cartridge Models "QT-M" and "QT-J" have the following specifications: Minimum Needle Pressure, 1-1/4 oz.; output voltage .75, average at 1,000 c.p.s. on Audiophone 78-1 frequency test record; cutoff frequency, 5,000 c.p.s.; terminals, pin type.
NOW! a new standard of performance in cutting heads

THE PRESTO 1-D

The new Presto 1-D Cutting Head offers: wide range, low distortion, high sensitivity and stability through a temperature range of 60°-95° F. The Presto 1-D Cutting Head is a precision instrument made entirely of precisely machined parts, expertly assembled and carefully calibrated. These factors, plus its sound basic engineering design, produce a cutter unequaled in performance by any other mechanically damped magnetic device.

Note from the light pattern below: The correct location of the cross-over point at 500 cycles, the 6 db per octave slope below this point, and flat response above 500 cycles, which is free from resonant peaks. The range of the cutter is 50-10,000 cycles. The Presto 1-D is damped with "Prestoflex" which is impervious to temperature changes between 60 and 95 degrees Fahrenheit.
You Can Get TRUE FACTS ON PERFORMANCE with these TESTING UNITS

**ACME VOLTROL**

The Acme Voltrol provides a full range stepless control from 0 to 135 volts. Its regulation is accurate to within 4/10 volt adjustment. Unlike resistance regulators, the output voltage is practically independent of the load. Voltrol is the ideal testing instrument for predetermining the performance of any electrical device or product under voltage fluctuation conditions. Available in portable model (illustrated) and panel mounting types. Write for Bulletin 150.

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An entirely new kind of testing unit that provides for actual checking of circuits at approved standard testing voltages and in addition indicates grounds, shorts or opens. 100% leakage type transformer limits current under short circuit conditions, thereby preventing needless destruction to materials at point of breakdown.

Instead of simply indicating the resistance value of the insulation, which serves no practical purpose, the Acme Insulation Tester permits the application of high voltages to positively prove the safety qualifications of the electrical device or apparatus under test. The Acme Insulation Breakdown Tester may be adjusted to supply voltages of double the rated voltage plus 1000 in accordance with Underwriters' Laboratories testing recommendations.

**ACME ELECTRIC CORPORATION**

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(Continued from page 40A)

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Dubl., L., 5631 S. Belmont St., Portland 15, Ore.
Duffus, R. A., Jr., Research Laboratory, Stromberg-Carlson Co., Rochester, N. Y.
East, W. L., Jr., AA & GM Branch, T&S, Box 598, Ft. Bliss, Tex.
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(Continued on page 44A)
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P. R. MALLORY & CO., Inc., INDIANAPOLIS 6, INDIANA
...covers ALL television and FM frequencies

THE Andrew Co., pioneer specialist in the manufacture of a complete line of antenna equipment, continues its forward pace with the introduction of this new DI-FAN receiving antenna.

The DI-FAN antenna provides excellent reception on all television and FM channels. It thus supersedes ordinary dipole antennas or dipole-reflector arrays which work well over only one or two television channels.

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- Impedance of DI-FAN matched to impedance of transmission line, preventing ghost images.
- Designed for use with 500 ohm transmission lines, conforming to RMA standards for FM and TV receivers.
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- Light in weight but strong and durable. High strength aluminum alloy elements. Supporting members of heavily plated steel.
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- Mounting supports available for either chimney or roof installations.

This graph illustrates the superiority of the Andrew DI-FAN over an ordinary folded dipole.

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(Continued from page 42A)

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Lundy, W. R., 91 Woodland Ave., Summit, N. J.
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O'Conor, C. J., 6237 Marie St., Cincinnati 74, Ohio
Ozone, K., 9404 N. Wilton, Chicago 13, Ill.
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(Continued on page 46A)

PROCEEDINGS OF THE I.R.E. June, 1947
WHEN you design a small portable radio around the new high-energy "Eveready" No. 950 batteries...and the "Eveready" "Mini-Max" No. 467 "B" battery—you can keep the receiver small and compact without sacrificing battery life.

The "Eveready" No. 467 "B" battery, because of its exclusive space-saving flat-cell design, gives longer life in radios than any other "B" battery of equal size. The "Eveready" No. 950 battery, nationally famous before the war, has been redesigned and offers vastly more energy than ever before...without any increase in size or in price. These batteries are available everywhere.

Some Typical Filament Circuits

Here are four typical circuits that demonstrate how one or more "Eveready" No. 950 flashlight batteries can be connected to heat tube filaments. Many other combinations are, of course, possible and practicable.

For further information on these and other "Eveready" radio batteries, write to National Carbon Company, Inc., for Battery Engineering Bulletin No. 1.
Radio equipment used in United Air Lines Mainliners tested at maintenance base.

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UNITED AIR LINES selects

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for Group Training of its
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2. Enabling its staff to perform duties more efficiently and in less time.
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... and rated at 70°C Ambient Temperature

Fixed resistors are usually rated at ambient temperatures of 40°C. But Bradley unit fixed resistors are rated at 70°C.

At this high temperature, Bradley units ... in 3/8-watt, 1-watt, and 2-watt ratings ... operate at full rating for 1000 hours with a resistance change of less than 5 per cent. All three sizes are offered in standard R. M. A. values from 10 ohms to 22 megohms, inclusive.

Bradley unit resistors require no wax impregnation ... yet they pass salt water immersion test. The solid molded construction assures high mechanical strength and permanent electrical characteristics. War time uses proved that Bradley units withstand wide variations in temperature and humidity. Send for resistor data sheets. Allen-Bradley Co., 114 W. Greenfield Avenue, Milwaukee 4, Wisconsin.
FEED-THRU AND STAND-OFF CERAMIC CAPACITORS

are made to the same high performance standards as conventional CN and CI type Hi-Q capacitors. Engineers who have thoroughly investigated Hi-Q capacitor performance are unanimous in their approval. We invite inquiries for samples to meet the exact needs of your applications.

- OTHER Hi-Q COMPONENTS

WIRE WOUND RESISTORS

CERAMIC CAPACITORS

S. I. TYPE Duarez Coated

C. N. TYPE

C. I. TYPE

CHOKE COILS

Hi-Q ELECTRICAL REACTANCE CORPORATION
FRANKLINVILLE, N. Y.
The famous Model 80 Even Speed Alliance Phonomotor operating on 110 or 220 volts is made for 40, 50 or 60 cycles, 16 watts input, 78 RPM. It has no gears—runs at an even speed—has a smooth, quiet, positive friction-rim drive. Amply proportioned bearings with large oil reservoirs assure long life. A slip-type fan gives cool operation—avoids any possible injury.

The Alliance Model K Phonomotor, a 25 cycle companion to the Model 80, operates on 110 volts, 25 cycles at 12 watt input. Motor and idler plate on Alliance phonomotors are all shock mounted to the cabinet mounting plate, to minimize vibration.

The trend is to make things move!

Designs will call for more action—movement! Flexible product performance needs power sources which are compact, light weight! Alliance Powr-Pakt Motors rated from less than 1/400th on up to 1-20th h.p. will fit those "point-of-action" places! Alliance Motors are mass produced at low cost—engineered for small load jobs! For vital component power links to actuate controls...to make things move...plan to use them!

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PHYSICISTS
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Equipment for photographing single electrical transients of very short duration. Includes demountable cathode ray tube with mechanical and molecular pumps, film drum with high speed drive motor, 50 KV power supply, Norinder relay and timing circuits. Range markers down to 2 microseconds. Phenomena of 1 to 20,000 microseconds duration may be recorded. Maximum film drum speed 7,000 RPM. Unused.

Write for further information

BOX 472
THE INSTITUTE OF
RADIO ENGINEERS
1 East 79th St., N.Y. 21, N.Y.

Five positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
1 East 79th St., New York 21, N.Y.

RADIO ENGINEER

Graduate radio engineer for research and development work on high frequency antennas and transmission line. Firm is a progressive subsidiary corporation of one of the nation's largest radio manufacturers, and is located in the middle west. Salary to be commensurate with qualifications of accepted person. Box 454.

PATENTS

Patent Department of large industrial corporation requires experienced patent attorneys or agents. At least one with electronic background, for expanding research effort. Minimum supervision. Office in New York suburb, southwest Connecticut. Box 455.

DEVELOPMENT ENGINEERS

Brooklyn engineering and manufacturing company, established in naval fire control and precision instrument work, requires qualified engineers for servomechanism and related development program. Positions require mature, responsible engineers with analytic as well as laboratory background, and a few years' experience in fundamental development work. Facility in applied mathematics and advanced circuit development requisite. Starting salary commensurate with experience, and subsequent rewards commensurate with accomplishment. Box 456.

ENGINEERS—PHYSICISTS

Engineers experienced with the operation and maintenance of the SCR-584 radar. Also graduate physicists experienced with optical instruments and electronic timing. Unusually attractive advantages. For application forms write P.O. Box 661, Ventura, California.

INSTRUCTOR IN ELECTRICAL ENGINEERING

Instructors and professors to teach electrical engineering at prominent university in metropolitan area. Undergraduate and graduate courses in all fields and excellent research facilities. Professors must have advanced degrees. Salaries $3,000 to $7,500 depending on qualifications. Write to Box 459 with full details of education and experience.

PHYSICISTS

Arsenal has positions open for physicists in Civil Service. Salaries from $2,644 to $3,905 per annum. Apply to: Frankford Arsenal, Philadelphia 37, Pennsylvania.

The Naval Air Material Center has urgent need for engineers qualified under U. S. Civil Service Commission standards in the fields of radio and radar. Regular work consists of five eight hour days. Employees accrue vacation and sick leave. Permanent employees are also eligible for Civil Service Retirement. Salaries range from $2,644.50 to $8,179.00. Write to Naval Air Material Center, Industrial Relations Department, Bldg. 75, U. S. Naval Base Station, Philadelphia 12, Pa., for particulars concerning the filing of applications, types of positions and appointments.

INSTRUCTOR IN ELECTRICAL ENGINEERING

Electrical, graduate with Master's degree and some teaching experience; to teach undergraduate courses in circuits and machinery. Opportunity to do graduate work. Salary $3,000 for 9 months. Location, southwest. Box 456.

ASSISTANT PROFESSOR OF ELECTRICAL ENGINEERING

Electrical graduate with Master's degree and several years' teaching experience; to teach AC and DC circuits, AC and DC machinery, and other power courses. Opportunity to do graduate work. Salary $3,600 for 9 months. Location, midwest. Box 466.

PHYSICISTS AND ELECTRICAL ENGINEERS

For vacuum tube research. Apply by letter stating qualifications to Director of Research, National Union Radio Corporation, 350 Scotland Road, Orange, New Jersey.

FACTORY ENGINEER

We have an opening in our factory engineering division for an outstanding man. This position requires experience background of at least 5 years' engineering work on factory problems relating to receiving tube manufacture. An engineering degree would be helpful but the primary requirements of the position are the experience and the ability to successfully solve every day problems encountered in the manufacture of receiving tubes. Apply by letter to PERSONNEL DEPT. National Union Radio Corporation, Lansdale, Pa.

ELECTRICAL ENGINEERING TEACHERS

Instructors, assistant and associate professors to teach electrical engineering at state university in the southeast. Salaries $2800 to $4500 for 9 months depending upon qualifications. Write full details of education and experience. Box 467.

RF-1F—TRANSFORMER ENGINEER

Radio engineer with theoretical and practical knowledge and experience in design of RF-1F transformers. Familiar with modern practice and requirements in FM and television npact. Excellent opportunity with established growing company. Write giving full details. Box 468.

(Continued on page 574)

PROCEEDINGS OF THE I.R.E. June, 1947
WANTED: MEN WHO CAN FILL THESE JOBS!

**FEDERAL'S NEW PLANT**, at Clifton, N. J., is the last word in modern design, modern equipment, and modern methods for precision manufacture of tele-communication and electronic equipment. Remember, too, that Federal is an associate of the International Telephone and Telegraph Corporation—one of the oldest and most security-founded organizations in the industry. A job with Federal is a job with an assured future!

**THE FEDERAL TELECOMMUNICATION LABORATORIES**, at Nutley, N. J., represent the most modern laboratory and research facilities available anywhere. This, the American unit of IT&T's world-wide research and engineering organization, also represents the most advanced thinking in the field—pioneering that will shape the future of the radio and electronic industries. An affiliation with this organization offers great opportunities for the right men!

Federal now has openings for a few top-grade engineers who seek an unusual opportunity

**IF YOU ARE** an electronic or communication engineer with a really outstanding background—both academic and practical—this may be just the opportunity you've been looking for.

Federal now has a limited number of excellent jobs available for engineers with superior ability—men who want permanent positions with a company known the world over for its far-sighted research and development work in all fields of tele-communications and electronics.

Development engineers with 3 to 15 years experience in high-power and low-power transmitter design; engineers with 3 to 15 years experience in instrument landing of aircraft, mobile transmitters and receivers or wire transmission; telephone engineers with 3 to 15 years experience in circuits and equipment.

If you can meet these qualifications and want a job with an assured future, send complete resume giving educational background, job experience, age and salary requirements, to Federal Telephone and Radio Corporation, Clifton, New Jersey, attention of J. A. Abbott, Personnel Manager. All information will be kept in strict confidence.

---

Federal Telephone and Radio Corporation


PROCEEDINGS OF THE I.R.E. June, 1947
PARA-FLUX REPRODUCERS
WITH NEW MODEL EL-2 EQUALIZER
for realistic reproduction of transcriptions

The New Model EL-2 EQUALIZER has all components enclosed in one compact housing. This built-in feature replaces the old-style two-piece equalizer, and also eliminates heavy cable. The newly designed Equalizer, in one complete package, embodies a double housing which gives double shielding against hum pickup. Combines the switch mechanism as well as impedance matching and correct equalization for following switch positions:

**VERTICAL NO. 1** — Linear output from 40 to beyond 11,000 C.P.S.

**VERTICAL NO. 2** — Linear output from 40 to 1500 C.P.S. with roll-off to —10 D.B. at 10,000 C.P.S.

**LATERAL NO. 1** — Linear output from 40 to beyond 11,000 C.P.S. for N.A.B. pre-emphasis.

**LATERAL NO. 2** — Linear output from 40 to beyond 11,000 C.P.S. for orthophonic pre-emphasis.

**LATERAL NO. 3** — Linear output from 40 to 3500 C.P.S. with roll-off to —10 D.B. at 10,000 C.P.S. for shellac recording.

Output impedance: 30, 250, and 500/600 ohms.

Equalizer requires only single %" dia. hole for mounting. Accommodates any panel thickness from 1/16" to 1/32".

The PARA-FLUX REPRODUCER, with interchangeable heads for Vertical, Lateral or Universal, uses only one arm and Equalizer. All possess the same impedance matching to the Equalizer. High output level affords an important advantage in broadcasting as to value of signal level to background noise. Each head is fitted with a selected, hard African diamond stylus, polished and finished to tolerance of 1/10,000 of an inch. Universal and Vertical: 2 mil radius. Lateral: 2.5 radius. "Hair-line" indicator on head and precise stylus construction make accurate cueing possible and permit "back-tracking" without damage to record or reproducer.

Reproducer is sturdy built, embodies up-to-the-minute features, including convenient finger lift which prevents Reproducer from slipping when lifted off record.

More than 1,000 PARA-FLUX REPRODUCERS are now on the air over FM and AM stations. Also widely used by recording studios, wired program services, sound distribution systems, and for high fidelity home sets.

AVAILABE THROUGH AUTHORIZED JOBBERs
Descriptive, illustrated Bulletin PR51, upon request.

RADIO-MUSIC CORPORATION
EAST PORT CHESTER CONN.
Machlett Laboratories now makes available, for early delivery, two new tubes—the ML-5604 for forced air cooling and the ML-5619 for water cooling—both specifically designed to withstand the rigorous and non-uniform operation inherent in industrial heating applications. In the development of every feature of these tubes, such conditions as widely varied loads, severe vibration, heavy irregular physical shocks and operation by personnel untrained in electronics, have been given full consideration.

A. Heavy wall high conductivity copper anode—specially processed.
B. One piece high conductivity copper grid and filament support terminals... for maximum strength, minimum lead resistance and elimination of electrode distortion.
C. Improved filament spring design. Minimizes bowing and increases filament life.
D. Chemically cleaned, vacuum fired internal parts for longer life and stable operation.
E. Stronger self-supporting grid for uniform electron control.
F. Rugged Kovar grid and filament seals.
G. Rigidly supported grid and filament assemblies. Glass surfaces completely shielded against electron bombardment and radiant filament energy.
H. Glass contour provides long leakage path and more efficient cooling.
I. Rugged Kovar plate seal located in air stream.
J. Gold plated contact surfaces. Insure permanent low contact resistance.

These completely new tubes are an outstanding contribution to industrial electronics. They may, of course, also be used for communications purposes. For further information, write Machlett Laboratories, Incorporated, Springdale, Conn.

ML-5604
TRIODE R.F. HEATING OSCILLATOR AND POWER AMPLIFIER

<table>
<thead>
<tr>
<th>Component</th>
<th>Characteristics</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filament Voltage</td>
<td>11.0 a.c. Volts</td>
</tr>
<tr>
<td>Current</td>
<td>150 Amps</td>
</tr>
<tr>
<td>D.C. Grid Voltage</td>
<td>-2000 max. Volts</td>
</tr>
<tr>
<td>D.C. Plate Current</td>
<td>7.75 max. Amps</td>
</tr>
<tr>
<td>D.C. Grid Current</td>
<td>40 max. Amps</td>
</tr>
<tr>
<td>Plate Input</td>
<td>27.5 max. KW</td>
</tr>
<tr>
<td>Plate Dissipation</td>
<td>10 max. KW</td>
</tr>
<tr>
<td>Plate Dissipation</td>
<td></td>
</tr>
<tr>
<td>Rating</td>
<td></td>
</tr>
<tr>
<td>Maximum ratings</td>
<td>Absolute values</td>
</tr>
<tr>
<td>Mon. Frequency</td>
<td>5000 max. Volts</td>
</tr>
<tr>
<td>Max. Grid Voltage</td>
<td>27.5 max. kW</td>
</tr>
<tr>
<td>Max. Plate Voltage</td>
<td>25.000 max.</td>
</tr>
<tr>
<td>Max. Plate Temperature</td>
<td>220°C</td>
</tr>
</tbody>
</table>

AMPLIFICATION FACTOR: 18.5
DIRECT INTERELECTRODE CAPACITANCES:
- Plate to Grid: 25 mmfd
- Plate to Filament: 125 mmfd
- Grid to Filament: 30 mmfd

COOLING: Minimum air flow through radiator 270 c.f.m. at 105°C back pressure.
Minimum air flow of 15 c.f.m. from 3" nozzle on center of dish.
TRUSCON RADIO TOWERS
Made Strong
Stay Strong
under the most difficult service requirements

- WSPR, Springfield, Mass., owns and operates two Truscon Radio Towers. Recently, it was necessary to move one of the towers 50 feet to a new foundation. The Truscon tower was left intact...even the tower lights were left in position...the whole job of moving, as shown by the sequence of photos here, was accomplished in two days...and the tower was put back into service immediately.

This is a typical example of Truscon Radio Tower ruggedness—the result of good engineering, good materials and good construction. Truscon can engineer any type of tower you desire...gued or self-supporting, either tapered or uniform cross-section...tall or small...AM or FM. Truscon engineering consultation is yours without obligation.

Write or phone our home office at Youngstown, Ohio, or any of our numerous and conveniently located district sales offices.

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

Manufacturers of a Complete Line of
Self-Supporting Radio Towers...
Uniform Cross-Section Guyed Radio
Towers...Copper Mesh Ground
Screen...Steel Building Products.
ECONOMICAL—K-TRANS cost less to purchase—less to use.

EFFICIENT—K-TRANS will duplicate or exceed the performance of your present i.f. transformers.

STABLE—Permeability tuning, magnetic shielding of windings, silver mica condensers combine to give a stability never before obtainable in a standard commercial i.f. Transformer.

VERSATILE—Four models of K-TRAN meet all 455 KC requirements. Also available for 262 K.C., and 10.7 M.C. for F.M. receivers.
S.S. WHITE FLEXIBLE SHAFTS

Satisfy every need for

SMOOTH,

SENSITIVE REMOTE CONTROL

OVER LONG DISTANCES

S.S. White remote control type flexible shafts are specially engineered and built for remote control applications. Their torsional deflection under load is very small and is the same for either direction of rotation.

With these S.S. White remote control flexible shafts you can get smooth, easy operation with any required degree of sensitivity over long distances as well as short. For the full story,

WRITE FOR 260-PAGE HANDBOOK—FREE

It gives complete facts and technical data about flexible shafts and their application. A copy will be sent free if you write for it direct to us on your business letterhead and mention your position.

S.S. WHITE INDUSTRIAL DIVISION
THE S.S. WHITE DENTAL MFG. CO.

DEPT. G 10 EAST 40TH ST., NEW YORK 16, N. Y.

Flexible Shafts • Flexible Shaft Tools • Aircraft Accessories
Small Cutting and Drilling Tools • Special-Formula Accessories
Mold Revisions • Plastic Specialties • Contract Plastic Molding

One of America's AAAA Industrial Enterprises

Positions Wanted

(Continued from page 54A)

JUNIOR ENGINEER


SALES ENGINEER

M.S. in E.E., Boston College, 1940. B.S., Boston College, 1938. Age 30. Two years' teacher, public schools, college physics. Fellowship, Boston College. Desires position in technical sales or application engineering field. Has had experience as research physicist, production engineer, assistant sales engineer. Prefers New England location. Box 85W.

ENGINEER

B.E.E., Rensselaer, Army officer, Harvard and M.I.T. training, teaching electronics at night, college level, 3 years' experience with LORAN, Countermeasures equipment, and servos, 1 year experience in the coke industry. Desires association with the electronic industry in Boston, Mass. Box 86W.

JUNIOR ENGINEER

Completing Junior Engineering, two non-electrical courses for B.E.E. degree. N.Y.U. Desires work in industrial electronics or sales engineering in metropolitan area. Age 29. Details on request. Box 87W.

ELECTRONICS ENGINEER

Experienced receiver and transmitter design and production, F.M., A.M.; B.S. London. Age 31. 1934-37, RX designer large British radio firm; 1937-39 Air Ministry production of airborne RX and TX equipment; 1925-45 R.A.F. pilot 3,500 hours, transatlantic Captain; 1945 to date, technical director of large broadcasting station designing PM equipment, and color television. Speak French, well traveled. Desire position Connecticut or New York area. Consider representation or sales engineering. Box 88W.

JUNIOR ENGINEER

Age 23. Married, 2 children. Experienced in design and all other phases of aeronautical navigation, radio communications, and broadcasting, radar, GCA, etc. 1st Class phone license. Now in charge of a CAA communications station. Prefer west coast. Résumé on request. Dan W. Crockett, MTIC, UMIAT Radio, c/o Box 1310, Fairbanks, Alaska.

INSTRUCTOR

B.S.E.E. '43 Washington University, M.E.E. Cornell due in June. Bowdoin M.I.T. radar school; Experience: some radio industry, some teaching, Navy Elec. Off.; Tau Beta Pi, Sigma Xi, Phi, Ham; Married; Midwest preferred. Box 90W.

RADIO ELECTRONICS ENGINEER

Desires Paris position in September, 1947. Worcester Polytechnic Institute, Sigma Xi; 8½ months Navy Radar Bowdoin, M.I.T., A.M. Harvard. Experience: 2½ years RadLab. M.I.T., ½ years Airborne Radar Navy, 1 month microwave, 9 months patent work, 3 months Optic Research Lab. Box 103W.

(Continued on page 58A)
FOR YOUR CONVENIENCE . . .

PANEL INSTRUMENTS NOW AVAILABLE
Right Off the Shelf!

To supply you with instruments at the time you need them, General Electric is accumulating a stock of 3½-inch, round and rectangular panel instruments in all the popular ratings. No waiting for delayed shipments . . . just place your order and they’re on the way to you.

NEW DISCOUNT BENEFITS
Also, more favorable prices, made possible by new discount benefits, are now available to you when ordering these standard types and ratings.
   Included in these stocks of compact, high-quality G-E instruments are ammeters, voltmeters, milliammeters, and microammeters . . . instruments for applications where AVAILABILITY counts.

INSTRUMENTS MADE TO ORDER
   In addition, General Electric is equipped to solve your individual instrument problems. Requests are welcomed for special, made-to-order instruments to be incorporated in your product where standard models cannot be used. For further information contact your General Electric representative or write to Apparatus Department, General Electric Company, Schenectady 5, N. Y.

VU METERS are now normally stocked in the following styles: non-illuminated, no mask, “A” scale; illuminated, no mask, “A” or “B” scale. Black covers are standard; gray covers can be furnished.
Specify "GREENOHMS" if you want the toughest things in power resistors. For Greenohms are those green-colored cement-coated power resistors found in the finest receivers, amplifiers, transmitters and other electronic assemblies.

Greenohms have proved that "they can take it" day after day, year in and year out. Handle heavy overloads without flinching. The exclusive cold-setting cement coating means that wire winding is unimpaired in fabrication. Withstands high temperatures and sudden cooling, on-off operation, without cracking, flaking, peeling. No tougher power resistors are made.

★ Write for DATA...

Bulletin 113 sent on request. Contains all necessary engineering data on standard and special Greenohms to meet your particular requirements. Let us quote on your needs.
The WILCOX 96C Transmitter is used throughout the world by the Army Air Force Communications System, and by foreign and domestic air-carriers. It has earned the respect of operators and engineers because:

1. **SIMULTANEOUS CHANNEL OPERATION** on several frequencies brings new flexibility and operational ease; increases by 3 times the volume of traffic normally handled.

2. **INSTANTLY REMOVABLE COMPONENTS** make maintenance easy. The transmitter slides from its cabinet like a desk drawer. Plugs, receptacles, and clips make all components easily removable for servicing while adjacent channels continue to operate.

3. **RESEARCH AND ENGINEERING** combined with modern production techniques have produced a transmitter capable of sustained operation through all conditions of temperature, humidity, and weather.

Write for Free Catalog... DEPENDABLE COMMUNICATION
**Individually Calibrated Scale**

**OUTPUT:** Continuously variable, .1 millivolt to 2.2 volts.
**OUTPUT IMPEDANCE:** 5 ohms to .2 volt, rising to 15 ohms at 2.2 volts.

**MODULATION:** From zero to 100%, 100 cycles, provision for external modulation. Built-in, low distortion modulating amplifier.

**POWER SUPPLY:** 117 volts, 60 cycles, AC.

**DIMENSIONS:** 11" high, 20" long, 10" deep, overall.

**WEIGHT:** Approximately 50 lbs.

Catalog on request

---

**MEASUREMENTS CORPORATION**
**SOONTON**
**NEW JERSEY**

---

**Sub-Miniature Inductors**

Supplementing the Cambridge Thermionic Corporation, 445 Concord Ave., Cambridge 38, Mass., line of slug-tuned inductors, new sub-miniature inductors have been added. Known as Type LSM, they are particularly useful where a variable inductance is desired and space is limited. Their height when mounted is only 1/8". These units are available in a group of windings which cover normal inductance ranges and special windings may be obtained to meet other specifications.

---

**UHF Double Triodes**

Two new double triodes providing single-ended operation for cascade amplifiers operating at frequencies up to 400 megacycles are being manufactured by the Radio Tube Division of Sylvania Electric Products, Inc., 500 Fifth Avenue, New York 18, N. Y. These high mutual-coupling tubes have independent elements with the exception of heaters permitting an appreciable saving in space and number of tubes required.

---

Circuit applications include grounded-grid, cathode follower, and push-pull amplifiers and converters in FM and television bands where low equivalent noise resistance, obtained with triode converters, is desirable. Type 7F8 is supplied with a 6.3 volt, 0.300 ampere heater. Similar ratings for Type 14F8 are 12.6 volts and 0.150 ampere. Heaters may be operated from an alternating or direct current source.

---

(Continued from page 48A)
Everywhere, when quality is important, AUDIODISCS are referred over all other recording blanks combined.

This universal acceptance by recording engineers in radio, motion pictures, commercial recording studios, and in the production of phonograph records, is the natural result of the consistent high quality of these fine recording discs.

For AUDIODISCS are manufactured by a patented precision-machine process which assures uniform results, and AUDIODISC recording lacquer is produced in our own plant from a formula developed by our research engineers. The manufacturing process is thus fully controlled from raw materials to the finished disc.

Praise of AUDIODISCS comes from everywhere, not only from all fields of recording, but from every type of climate. In arctic cold or the heat and humidity of the tropics, AUDIODISCS are consistently dependable.

There is an AUDIODISC designed for every recording need. See your local distributor or write:

Audiodevices, Inc., 444 Madison Avenue, New York 22, N.Y.

Export Department: Rocke International Corp., 13 E. 40th Street, New York 16, N.Y.

Audiodevices are manufactured in the U.S.A. under exclusive license from PYRAL, S.A.R.L., Paris.


For Recording Quality

EVERYWHERE...it's audiodiscs

they speak for themselves audiodiscs
No size limitations—custom built for specific applications
These large B & W coils are popular for tank circuits, antenna matching networks and similar applications where rugged dependability must be combined with design adaptability to meet individual conditions. Custom built units, based on standard B & W designs are available in either fixed, tapped or continuously variable types and with either fixed link or fixed variable link in any combination. Send details of your application for recommendation and quotation.

BARKER & WILLIAMSON, Inc.
237 Fairfield Ave., Upper Darby, Pa.

TRANSMITTING AND SPECIAL PURPOSE TUBES

NEWARK NOW AGENTS OF WAR ASSETS ADMINISTRATION

Newark has been appointed agents of the War Assets Administration for transmitting and special purpose electronic tubes.

HUGE STOCKS! WIDE SELECTION!
This means that you can now get prompt Newark service on the previously hard-to-get tubes, priced at a fraction of their original cost. Make Newark your headquarters for tubes — whether it's for experimental work or production runs.

ACTING AS AGENTS FOR WAR ASSETS ADMINISTRATION UNDER CONTRACT WAS(p) 7-167

Antenna Switch

A new type of antenna-switching relay is being manufactured by the Advance Electric & Relay Company, 1260 W. 2nd St., Los Angeles 26, Calif. To facilitate switching of two-wire open lines, twin relays can be spaced the same distance apart as the transmission line. Any spacing down to two inches is possible thus minimizing discontinuities in line spacing and line impedance.
"A HEAD OF THE TIMES"

All recessed head screws and bolts have definite advantages over the older, slotted head type but only Reed & Prince recessed heads can be driven in any size from the smallest to the largest—WITH ONE DRIVER!

We make Hand Drivers and Bits for power Drivers with long, short and special shafts, but the POINT is always the same! Buy Reed & Prince.

REED & PRINCE
MANUFACTURING COMPANY

CHICAGO, ILL.

WORCESTER, MASS.
At last—

THE RADAR ENGINEER'S HANDBOOK

—presenting up-to-date data on theory and practices of radar technology
—telling how to design equipment
—describing typical radar systems

Here is a comprehensive, handy reference guide to the practical and engineering aspects of radar. It is specifically designed to acquaint engineers and technical workers in radio and electronics with new techniques, and with special applications of old techniques used in radio detection and ranging of objects. Completely covering radar theory and practice, this handbook gives you the fundamentals essential to understanding the practical and effective employment of radar apparatus, describes various radar systems developed and used during the war, and explains in technical detail, the design of specific radar equipment.

Just Out

RADAR ENGINEERING

By Donald C. Fink

Editor, Electronics: Formerly Staff Member, Radiation Laboratory, M.I.T.; and Expert Consultant, Office of the Secretary of War.

644 pages, 471 illustrations, 6x9, $7.00

This is the first complete book on radar—bringing together in convenient form, full, authoritative data on the many individual developments to date in this field. It enables the engineer to understand quickly and easily the underlying theory of all branches of radar, and to judge critically the use of that theory in the design of radar equipment. It supplies many design formulas which may be applied not only in the fields of radar but in related fields of ultra-high-frequency and super-high-frequency communication, in navigation aids, etc. Theoretical material includes data on pulse generation and transmission, wave-guides, reflections of radio energy, etc. Practical aspects covered deal with components, circuits, and structures used in radar equipment. The various types of equipment described are those employing wide ranges of frequencies: 200, 600, 3,000, and 10,000 megacycles.

News—New Products

Discharge Capacitors

A new series of discharge capacitors for pulsed-lighting applications, such as speedflash photography and high-intensity flashing signals, and beacons, has been announced by Solar Manufacturing Corporation, 285 Madison Avenue, New York 17, N. Y. All units are specially designed for energy-storage service with heavy internal leads to carry high discharge current. The manufacturer states that Type QLX series of discharge capacitors have a low inherent inductance, and are of special construction to minimize discharge stresses.

(Continued from page 66A)

RCA HB-3 Tube Handbook

Now in 3 Binders—6 Volumes

No other tube handbook provides as much up-to-the-minute technical data on tube types as the RCA HB-3 Handbook, which has been a standard technical reference book for over 15 years. Indexed contents include general data, characteristic curves, socket connections, outline drawings, price lists, preferred-type lists, etc., for the complete line of RCA tubes.

New Sheets Mailed Regularly. The U. S. subscription price of $10.00 brings you the complete Handbook in three binders, plus a supplementary sheet containing new or revised data as issued during the year. Annual service fee thereafter is $2.00. (These prices apply only in the U. S. and its possessions.)

Subscribe Now. Insure early delivery. Mail your remittance today to: RCA, Commercial Engineering, Section W-52F, Harrison, N. J.
You need a dependable, well engineered transmitter for point-to-point, ship-to-shore, or ground-to-plane commercial radio communication. The Collins 231D or the Collins 16F is the answer. These transmitters have proved themselves thoroughly reliable and efficient in all climates, and under difficult operating conditions.

Any one of eleven frequencies between 2.0 mc and 18.1 mc is available at the flip of a dial, with all circuits tuned and ready to operate. The widely acclaimed Collins Autotune system is utilized to shift the frequency quickly and accurately.

Compressor circuits are incorporated to raise the average modulation level during voice or MCW transmission. CW transmission is also available, with keying speeds of 60 wpm on MCW and 200 wpm on CW. Both transmitters can be adapted for frequency shift keying.

For dependable, trouble-free radio communication use either the 231D or the 16F. They are built for that purpose. Write today for free illustrated bulletins giving detailed information.
UNIVERSAL PLATE
SUPPLY TRANSFORMERS

Professional Series

CUSTOM BUILT—MASS PRODUCED

Electroseal Construction ▶
High Overload Capacity ▶
Extreme Compactness ▶
Complete Reliability ▶
Unlimited Life ▶
Fully Guaranteed ▶

Electro Transformers embody the distinctive "Electroseal" construction and are designed to combine maximum performance with reduced weight and size. They are adaptable to a wide variety of uses and offer efficient, economical performance. Quality materials, excellent workmanship and twenty years of research are combined to give you the particular transformer for your specific job.

ELECTRO ENGINEERING WORKS
6021 College Avenue, Oakland 11, California

for every photo tube application

There is a High Sensitivity

RAULAND VISITRON

There is a Rauland VISITRON available for every phototube application . . . for industrial electronic control devices, sound-on-film, research and development. For high sensitivity, uniformity and dependability, Rauland VISITRONS are the preferred phototubes.

Specify VISITRON . . . and be sure!

Send for valuable Rauland VISITRON Catalog—it's yours for the asking.

THE RAULAND CORPORATION
CHICAGO 41, ILLINOIS

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)

High Current Igniton

A new 400-ampere sealed igniton for high-power rectifier service has been made available by the Tube Division of General Electric Company's Electronics Department, Syracuse, N.Y. The GL-507, largest sealed igniton in its class, with an average current rating of 400 amperes, will be primarily used in power rectifiers for mining, electrochemical, transportation, and steel industries.

Transmission Measuring Set

A mercury-pool tube of permanently-sealed steel construction, the new igniton has two igniters, only one of which is used at a time. The tube is over two feet tall and weighs about 100 pounds. According to General Electric tube engineers its design provides the control characteristics of the thyatron, the versatility of the half-wave tube in circuit application work, and the very high emission capacity of the mercury pool.

FINCH TELECOMMUNICATIONS
INCORPORATED
SALES OFFICE
10 EAST 40TH STREET, NEW YORK
FACTORIES: PASSAIC, N. J.
**Specify**

**MYCALEX**

**LOW LOSS INSULATION**

Where high mechanical and electrical specifications must be met.

**MYCALEX 410**

(MOLDED MYCALEX)

makes a positive seal with metals . . . resists arcing, moisture and high temperatures.

27 years of leadership in solving the most exacting high frequency insulating problems.

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"Owners of 'MYCALEX' Patents"

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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 66A)

**Miniature Phototube**

A capsule-size phototube, Type RCA-1P42, manufactured by the Tube Department of the Radio Corporation of America, Camden, N. J., is one of the smallest phototubes ever offered commercially. About the size of a .22 calibre long rifle cartridge it has a maximum diameter of only 1/4" and an overall length just under 1 13/32". It is activated by light entering through a tiny window at its larger end.

Comparing favorably with larger phototubes in sensitivity, this new tube is expected to find many applications, particularly in devices and machines where the size of former phototubes has been a problem.

(Continued on page 66A)

---

**AN IMPORTANT TECH LAB DEVELOPMENT**

**New Type 1250**

**R. F. SWITCH**

High r. f. current carrying capacity 50 amps. max. intermittent load; 30 amps. steady load. Low loss factor. Sturdy mechanical design . . . Mycalex insulation. Furnished in any number of decks.

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Manufacturers of Precision Electrical Resistance Instruments

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**Features:** Compensated for ambient temperature changes from -40° to 110° F. Hermetically sealed; not affected by altitude, moisture or other climate changes. Explosion-proof. Octal radio base. Compact, light, rugged, inexpensive. Circuits available: SPST Normally Open; SPST Normally Closed.

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**Microphones by Turner**

"The Password to Sound Performance"

**The Turner Company**

909 17th Street N. E.
Cedar Rapids, Iowa

**News—New Products**

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(Continued from page 674)

**Sequence Relay**

Carter powers the MOTOROLA 162 MC F.M. Transmitter

Reliable power assures the reliable operation of the new Motorola FM Mobile 162 megacycle Transmitter...Instantaneous power from the CARTER Generator shown on the chassis above. The ONLY Dynamotor that delivers full power in 3/10 second and over 100,000 service-free transmissions. Write for latest catalog.

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**Delay Relays**

**Provide Delays Ranging From 1 To 120 Seconds**

**Ampereite Regulators**

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**The Password to Sound Performance**

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909 17th Street N. E.
Cedar Rapids, Iowa

**News—New Products**

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- For Microameters
- For Delicate Relays
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PROCEEDINGS OF THE I.R.E.
June, 1947
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Radiograms "Via RCA" to and from overseas points now are processed by automatic machines which speed your messages through such gateway cities as New York, London, San Francisco and Manila, without delay.

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FROM 0.1 MICROHENRY TO 1 HENRY

The Type 667-A Inductance Bridge is widely used in the laboratory and in production for precision measurements of the inductance of coils. It employs a standard bridge circuit with a number of unique design features which make it direct-reading, with high accuracy and convenience.

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- ERRORS from sliding-zero balance have been eliminated.
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PRICE: Type 667-A Inductance Bridge $400.00
(Accessories needed are oscillator, amplifier and head telephones)

NOTE — A few in stock for immediate shipment — ORDER NOW!

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The 226 series is recommended for high-frequencies and features optional variable pitch winding for wide band coverage. The 224 is wound with 3/8" or 1/2" copper tubing and offers the highest ratings of the types shown.

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- Transmitting Capacitors
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(Continued from page 66A)

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Rubin, S., 1825 W. Fifth St., Brooklyn 23, N. Y.
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Scheuer, B. C., 4148 Grove St., St. Louis 7, Mo.
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Schwartz, P., 1883 E. 12 St., Brooklyn 29, N. Y.
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Sharpless, T. K., 629 Walnut Lane, Haverford, Pa.
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Simonds, R. E., 104-30-187 St., Hollis, N. Y.
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Sokolinski, E. A., 300 Riverside Dr., New York 25, N. Y.
Southwell, J. D., Central Fire Station, Beaumont, Tex.
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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.
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New Enterprises

- Radio Engineering Company, 8 State Street, New York, N. Y., will specialize in high-quality receiver design.

Interesting Abstracts


HV Coupling Capacitors

Designed to withstand 10 test impulses of 95 kilovolts, new high-voltage coupling condensers manufactured by the Sprague Electric Company, 189 Beaver St., North Adams, Mass., are expected to be the practical solution to the long-standing problem of coupling telephone equipment to existing 7200-volt alternating current distribution lines. These units have a capacity of 0.002 microfarad and are rated for 8700-volt 60-cycle operation.

Solderless Coaxial Fittings

Solderless fittings for rigid coaxial transmission lines, available in both bronze and aluminum in all standard sizes, which are gas-tight and have excellent electrical characteristics are being manufactured by Raybouild Coupling Company, Meadville, Pa. The manufacturer states that the new couplings will sustain sufficient end-pull to support the line, and their solderless feature simplifies field installations.

(Continued on page 71A)
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Ohmite offers a complete line of dependable resistors wound on a ceramic tube and protected by vitreous enamel. Ratings from 10 to 200 watts. Available in the fixed type for general use, and in the "Dividolum" type with adjustable lugs for use as a multi-tap resistor or voltage divider.
Membership
(Continued from page 68A)

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Wade, E. 5009—45 St., Woodside, L. I., N. Y.
Walcott, H. R., Jr., RDF I, Allendale, N. J.
Walter, C. W., P., 427 Lincoln Ave., Rutherford,
N. J.
Ward, B. R., 2720 Grand Concourse, New York 58,
N. Y.
Warner, R. E., 514 N. E. Seventis St., Abilene, Kan.
Washubeski, E., 131 Hudson Ave., Red Bank, N. J.
Watson, R. E., 5020 Eringer Pk., Philadelphia 44,
Pa.
Weinstein, A., 3925 S. St., S. E., Washington, D. C.
Welch, H. W., Jr., Department of Engineering Re-
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THE JOURNAL OF RADIO RESEARCH AND PROGRESS FOR ENGINEERS & PHYSICISTS

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THE JOURNAL FOR MANUFACTURERS AND TECHNICIANS OF ALL GRADES

WIRELESS WORLD is Britain's leading technical magazine in the general field of radio, television and electronics. Founded over 35 years ago, it has consistently provided a complete and accurate survey of the newest British technique in design and manufacture. Articles of a high standard cover every phase of the radio and allied industries, with reviews of equipment, broadcast receivers and components. Theoretical articles deal with design data and circuits for every application.

TOD TECHNICAL BOOKS: "Television Receiving Equipment," by W. T. Cocking, M.I.E.E., editor of "Wireless", an up-to-date and informative work, 13 shillings (£2.60); "Basic Mathematics for Radio Students," by F. M. de B.Sc., D.I.C., A.C.G.I., a particularly helpful new book, 11 shillings (£2.20); available from the address above.
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 684.)

Plant Expansions

• • • At Wasco, Minn., E. F. Johnson Company, to manufacture a line of Indicator Lights which was recently purchased from the Gotthard Manufacturing Company.

• • • At Canonsburg, Pa., by Radio Corporation of America, totaling 115,000 square feet, for the production of phonograph records.

(Continued on page 764.)
Hallicrafters' famous radio equipment, sold and distributed around the world before the war and used with superb effectiveness in every theater during the war is once again on the move. Watch for latest details of the Gatti Hallicrafters mobile radio equipped expedition to the Mountains of the Moon in deepest Africa—a new and exciting test for the ingenuity of hams and the performance of Hallicrafters equipment.

**Model SX-42**  Described by hams who have operated it as "the first real postwar receiver." One of the finest CW receivers yet developed. Greatest continuous frequency coverage of any communications receiver—from 540 kc to 110 Mc, in six bands. FM-AM-CW. 15 tubes. Matching speakers available. **$275.00**

**Model S-40A**  Function, beauty, unusual radio performance and reasonable price are all combined in this fine receiver. Overall frequency range from 540 kc to 43 Mc, in four bands. Nine tubes. Built-in dynamic speaker. Many circuit refinements never before available in medium price class. **$89.50**

**Model S-38**  Overall frequency range from 540 kc to 32 Mc, in four bands. Self contained speaker. Compact and rugged, high performance at a low price. Makes an ideal standby receiver for hams. CW pitch control is adjustable from front panel. Automatic noise limiter. .............................. **$47.50**

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THE HALLCRAFTERS CO., MANUFACTURERS OF RADIO AND ELECTRONIC EQUIPMENT, CHICAGO 16, U.S.A.

[Additional text not legible]
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 744)

**UHF Signal Generator**

A particularly valuable laboratory standard for determining gain or alignment, obtaining antenna data or measuring standing-wave ratios, for reading single-stage or conversion gain, signal-to-noise ratios, circuit Q, or transmission-line characteristics, within its frequency range of 500 to 1350 megacycles, is being manufactured by the Hewlett-Packard Company, 395 Page Mill Road, Palo Alto, Calif.

This instrument, known as the 610A UHF Signal Generator, is described as unusually stable, and over its frequency range will supply accurately known voltages ranging from 0.1 microvolt to 0.1 volt. The radio frequency may be continuous, amplitude-modulated, pulsed, or square-wave-modulated. Pulse length can be readily controlled between 2 and 50 microseconds, and pulse rate is variable from 60 to 3000 times per second. A simplified direct-reading control is incorporated and operating charts are not necessary.

**Mercury Vapor Rectifiers**

Further completing their line of electron tubes, Eitel-McCullough, Inc., 1018 San Mateo Ave., San Bruno, Calif., announces the availability of Types 866A and 872A mercury vapor rectifiers. These new low-priced rectifiers are directly interchangeable with Types 866A/866 and 872A/872 of other manufacture.

Type 866A/866 operates with 2.5 filament volts, peak inverse voltage as high as 10,000 volts, and a maximum average plate current of 0.25 amperes. The 872A/872 has a five-volt filament and carries a maximum peak inverse voltage rating of 10,000 volts and a maximum average current rating of 1.25 amperes.

**Hand Microphone**

A newly developed hand microphone has been added to the line of microphones manufactured by The Turner Company, 909 17th Street, N.E., Cedar Rapids, Iowa. Designated as Model 20X, it features a “Metalwall” crystal which withstands humidity conditions not tolerated by the ordinary crystal. Factory tests reveal excellent response characteristics for a low-priced unit, which is light in weight and natural to hold.

**Lightweight Towers**

Because of persistent demands from amateur radio operators for a lightweight tower for directional beam antennas, the Fabricated Light Metals Company, 42 W. 15 Street, New York 11, N.Y., has introduced a new line of reasonably-priced aluminum towers which can be assembled by one man using only a common wrench. Other features of these 10’, 20’ or 30’ self-supporting towers are that they require only small ground space, are excellent for roof-top mounting, and are easily insulated when desired. Maintenance is negligible, it is stated, requiring no painting in normal installations.

**Spot Frequency Generator**

The Electronic Manufacturing Company, 714 Race St., Harrisburg, Pa., announces the production of its Spot Frequency Generator, Model #200. Engineered for use by the average service man, it contains 12 pre-set frequencies chosen to cover adequately 95% of the sets in use. Stability is assured by an electron-coupled circuit and low leakage is effected through use of double shielding. The manufacturer further points out that the unit attenuates to less than one microvolt.
Here’s the Helipot Principle that is Revolutionizing Potentiometer Control in Today’s Electronic Circuits

**CONVENTIONAL POTENTIOMETERS** have a coil diameter of approximately 1½” and provide only 4” (about 300”) of potentiometer slide wire control.

**THE BECKMAN HELIPOT** has the same coil diameter, yet gives up to 46” (3600”) of potentiometer slide wire control—nearly TWELVE times as much!

Some of the multiple Helipot advantages

EXTENSIVELY used on precision electronic equipment during the war, the Helipot is now being widely adopted by manufacturers of quality electronic equipment to increase the accuracy, convenience and utility of their instruments. The Helipot permits much finer adjustment of circuits and greater accuracy in resistance control. It permits simplifying controls and eliminating extra knobs. Its low-torque characteristics (only one inch-ounce starting torque*, running torque even less) make the Helipot ideal for power-driven operations, Servo mechanisms, etc.

And one of the most important Helipot advantages is its unusually accurate linearity. The Helipot tolerance for deviations from true linearity is normally held to within ± 0.5%, while precision units are available with tolerances held to 0.1%, 0.5%, and even less—an accuracy heretofore obtainable only in costly and delicate laboratory apparatus.

The Helipot is available in a wide range of types and resistances to meet the requirements of many applications, and its versatile design permits ready adaptation of a variety of special features, as may be called for in meeting new problems of resistance control. Let us study your potentiometer-rheostat problem and make recommendations on the application of Helipot advantages to your equipment. No obligation of course. Write today.

*Data above is for the standard Type A unit.

THE Helipot CORPORATION, 1011 MISSION STREET, SOUTH PASADENA 6, CALIFORNIA

PROCEEDINGS OF THE I.R.E. May, 1947
### Five Standard Slug-Tuned LS3 Coils Cover 1/2 to 184 mc

For strip amplifier work, the compact (1/4" high when mounted) LS3 Coil is ideal. Also for Filters, Oscillators, Wave-Traps or any purpose where an adjustable inductance is desired.

**Five Standard Windings:**

| 1 | 5 | 10 | 30 | 60 | megacycle coils cover inductance ranges between 750 and 0.065 microharys.

**CTC LS3 Coils** are easy to assemble, one 1/4" hole is all you need. Each unit is durably varnished and supplied with required mounting hardware.

**SPECIAL COILS**

CTC will custom-engineer and produce coils of almost any size and style of winding...to the most particular manufacturer's specifications.

**Consult CTC for Three-Way Component Service**

Custom Engineering... Standardized Designs... Guaranteed Materials and Workmanship

**Cambridge Thermionic Corporation**

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**PROCEEDINGS OF THE I.R.E. May, 1947**
The IMC Engineer ... on your staff

but not on your payroll....

OFFERS EXPERIENCED ELECTRICAL INSULATION ASSISTANCE...

If your problem is increased cost and assembly time, here's a ready answer. Call on your nearest IMC Engineer. He represents a complete line of insulating materials, and he's backed up with a wealth of experience that can help solve your problems. He can be a real help because he represents many manufacturers, and therefore can offer you the right product to best meet your needs. Ask him to...

1—Assist you in the selection of the best insulating material for your job.
2—Familiarize you with the proper method of application.
3—Suggest ways to eliminate waste.
4—Help increase your production.

Television—a Season Pass to Baseball!

Every home game—day or night—played by the New York Giants, Yankees and Brooklyn Dodgers will be seen over television this season!

Owning a television receiver in the New York area will be like having a season pass for all three ball clubs. And in other cities, preparations for the future telecasting of baseball are being made.

When more than one home game is on the air, baseball fans can switch from one to the other—see the most exciting moments of each through television!

Those who own RCA Victor television receivers will enjoy brighter, clearer, steadier pictures through the exclusive RCA Victor Eye-Witness picture synchronizer that “locks” the receiver in tune with the sending station.

To witness baseball or any other event in the ever-growing range of television programs—you’ll want the receiver that bears the most famous name in television today—RCA Victor.

When you buy an RCA Victor television receiver or radio, or Victrola radio-phonograph, or an RCA Victor record or a radio tube, you know you are getting one of the finest products of its kind science has achieved.

Radio Corporation of America, RCA Building, Radio City, New York 20. Listen to the RCA Victor Show, Sundays, 2:00 P.M., Eastern Standard Time over the NBC Network.

Several television cameras cover the baseball diamond to bring you a close-up of the action wherever it occurs. Here is a supersensitive RCA Image Orthicon television camera used by NBC's New York station WNBT in televising home games of the New York Giants.

Television gives you a choice seat at the game.
A mighty long time to answer one question, you say? Well, not when you've answered it 250,000 times... with a different answer each time! This is just another way of saying that C-D engineers have designed and built over a quarter-million different types of capacitors to meet manufacturers' specific requirements. And they've saved millions of dollars in manufacturers' assembly costs.

For many applications there is just one capacitor that will do the job right—for other applications several types may be equally suitable. Whether your problem calls for special capacitor design... or a standard type unit... C-D's specialized experience in the field of capacitors will be able to meet your needs quickly and efficiently.

Consult with our engineers. Catalog of standard types will be mailed at your request. Cornell-Dubilier Electric Corporation, Dept. MS, South Plainfield, New Jersey. Other large plants in New Bedford, Worcester and Brookline, Mass., and Providence, R.I.

MICA • DYKANOL • PAPER • ELECTROLYTIC CAPACITORS

CAPACITOR #1 One of the Type MC spark suppressors for use on heavy-duty vehicles. Capacitor unit is hermetically sealed, oil filled and impregnated.

CAPACITOR #2 This is an oil impregnated paper capacitor for by-pass applications. Available in a wide variety of capacities and voltage ratings to fit many applications where a sealed unit is desirable.

CAPACITOR #3 This low capacity, high voltage capacitor unit was designed especially for FM and television applications. Hermetically sealed and provided with glass insulated terminal.

CAPACITOR #4 A dual capacity unit, with one terminal connected to the case for operation on 220 V.A.C. Ideal for noise suppression, it is constructed for simple installation on flat or round surfaces.
To meet an insistent demand, we have concentrated some of our production facilities on these two popular wavemeters and have managed to accumulate a moderate stock of each.

Both are of the well-known absorption type with accuracies sufficient for a wide variety of routine frequency checks on the ranges of coils, oscillators, receivers, transmitters, etc.

The Type 566-A Wavemeter consists of a variable air condenser, five plug-in coils and an incandescent lamp for resonance indication. The Type 758-A Wavemeter covers its entire frequency range with a single-turn loop the inductance of which is varied simultaneously with the capacitance. An incandescent lamp is used for resonance indication. For low power uses (less than about 2 watts), resonance indication is obtained by the reaction of the wavemeters on the plate or grid currents of the oscillator.

These wavemeters are compact, rugged, inexpensive and direct reading in terms of our primary standard of frequency.

**TYPE 566-A WAVEMETER**

- **FREQUENCY RANGE**: 0.5 to 150 Mc
- **COILS**: Five plug-in type, all supplied. When not in use coils can be plugged into a rack on side of the instrument case
- **DIAL CALIBRATION**: Direct reading in frequency
- **ACCURACY**: ±2%, 0.5 to 16 Mc; ±3%, 16 Mc to 150 Mc
- **RESONANCE INDICATOR**: Incandescent lamp
- **ACCESSORIES SUPPLIED**: Two spare indicator lamps
- **DIMENSIONS**: 4 3/4 x 3 1/8 x 5 3/4 inches, over-all
- **WEIGHT**: 3 pounds
- **PRICE**: Type 566-A WAVEMETER—$60.00

**TYPE 758-A WAVEMETER**

- **FREQUENCY RANGE**: 55 to 400 Mc
- **COILS**: A single turn loop, inductance and capacitance are varied simultaneously
- **DIAL CALIBRATION**: Direct reading in frequency
- **ACCURACY**: ±2%
- **RESONANCE INDICATOR**: Incandescent lamp
- **TEMPERATURE AND HUMIDITY**: Over ranges normally encountered, accuracy is independent of both
- **DIMENSIONS**: 5 x 5 x 4 1/4 inches, over-all
- **WEIGHT**: 1 pound, 12 ounces
- **PRICE**: Type 758-A WAVEMETER—$35.00

**ORDER NOW—DELIVERY PROBABLY FROM STOCK**