Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

(Including the WAVES AND ELECTRONS Section)

August, 1949

Volume 37 Number 8

Sylvania Electric Products Inc.

TOMORROW'S TUBES AND CIRCUITS

The performance characteristics of the traveling-wave tube, and its associated-circuit "plumbing," are under engineering study.

PROCEEDINGS OF THE I.R.E.

Waves and Electrons Section

Table of Contents Follows Page 32A

The Institute of Radio Engineers


UTC
for every application
While the catalogue line of UTC components covers a wide variety of applications, many people are not familiar with the full range of products produced by UTC. It is impossible to describe the thousands of special UTC designs as they become available. The illustrations below are intended to indicate some of the range in size of these special products.
A
This 100 cubic foot modulation transformer is for 50 Kw. broadcast service. Frequency response flat from 30 to 20,000 cycles.
B
The high Q toroid coil shown is 12" in diameter. It operates in a 50 Kw. circuit at supersonic frequency.
C
This sub-miniature (.18 cubic inch) output transformer is intended for hearing aid and other extreme compact service. While the dimensions are only 7/16" x 9/16" x 3/4", the fidelity is ample for voice frequency requirements.
D
This sub-miniature (.18 cubic inch) permalloy dust core toroid is available in a wide range of inductances, and for frequencies from 1,000 cycles to 50 Kc.
write for catalog PS-409
United Transformer Co.
150 VARICK STREET
NEW YORK 13, N. Y.
EXPORT DIVISION: 13 EAST 40th STREET, NEW YORK 16, N. Y.,
CABLES: "ARLAB"

SPRAGUE KOOLOHM

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RESISTORS

Tops for TV and other difficult applications!

WIRE-WOUND RESISTORS WITH LOW S.Q.!

S.Q.?? That means "Service Quotient." With dependable, tropicalized Koolohm resistors, it's practically nil!

That's why major television manufacturers specify Sprague Koolohms and avoid unnecessary and expensive service calls due to resistor failures.

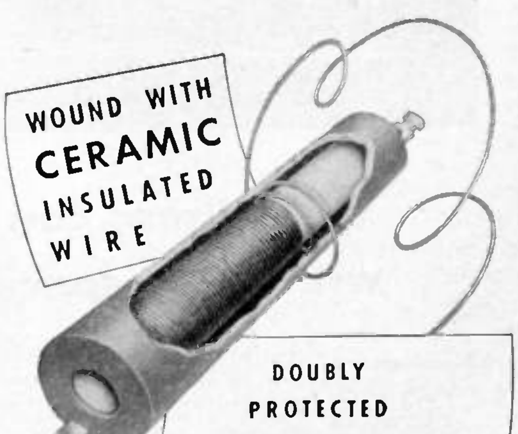
Koolohms far outperform and outlast ordinary wire-wound resistors, yet are smaller in size than ordinary units of the same wattage rating. Koolohms are wound with larger diameter wire for the same rating yet are available in far higher resistance values (for example, 70,000 vs 25,000 ohms at a full 10 watts). Koolohms are available in *truly non-inductive* windings when needed. Koolohms have exceptional resistance stability.

Koolohms have all these advantages because they are the *only* resistors wound with ceramic-insulated wire (an exclusive Sprague development) and are enclosed in glazed moisture-resistant ceramic outer shells. Mounted on a metal chassis, Koolohms will withstand a 10,000 volt breakdown test from winding to ground.

Koolohm resistors are ideal for television sets and other tough applications—and are equally good for all ordinary electronic and industrial control uses.

Best of all, premium-quality Koolohm resistors are *competitively priced.*

Sprague Catalog 100E tells The Complete Koolohm Story. Write for your copy today.



SPRAGUE ELECTRIC COMPANY

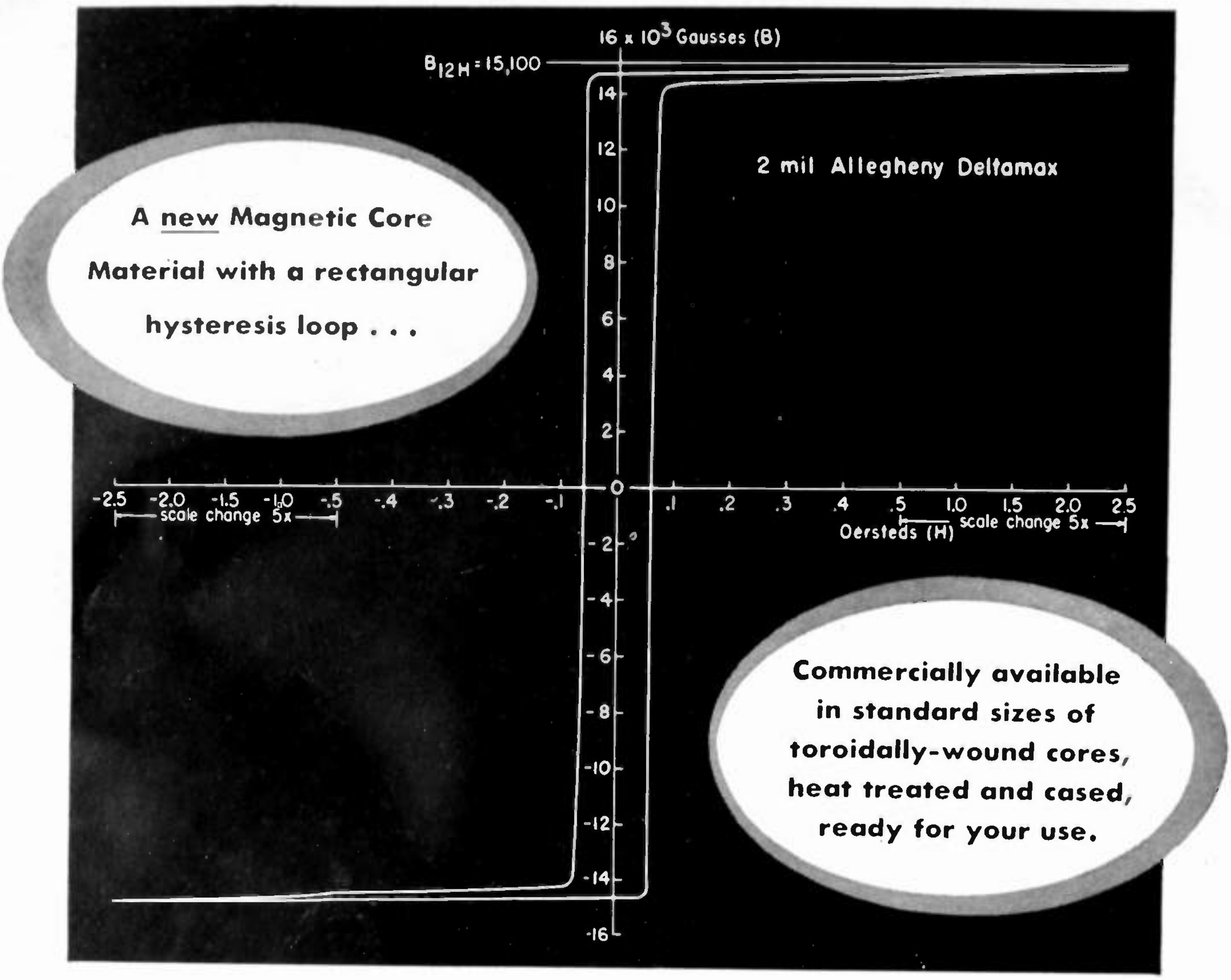
NORTH ADAMS, MASS.

PIONEERS IN ELECTRIC AND ELECTRONIC PROGRESS

PROCEEDINGS OF THE I.R.E. August, 1949, Vol. 37, No. 8. Published monthly in two sections by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price $2.25 per copy. Subscriptions: United States and Canada, $18.00 a year; foreign countries $19.00 a year. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

Table of contents will be found following page 32A

DELTAMAX -*now available!*



Where can YOU use a Magnetic Material with these specialized, dependable characteristics?

The properties of Deltamax are invaluable for many electronic applications, such as new and improved types of mechanical rectifiers, magnetic amplifiers, saturable reactors, peaking transformers, etc. This new magnetic material is available now as "packaged" units (cased cores ready for winding and final assembly) distributed by the Arnold organization. Every step in manufacture has been fully developed; designers can rely on complete consistency in each standard size of core.

Deltamax is the most recent extension of the family of special, high-quality electrical materials produced by Allegheny Ludlum, steel-makers to the electrical industry. It is an orientated 50% nickel-iron alloy, characterized by a rectangular hysteresis loop with sharply defined knees, combining high saturation with low coercivity.

● *Call on us for technical data.*

THE ARNOLD ENGINEERING COMPANY

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W&D 2379


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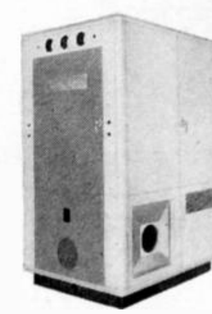
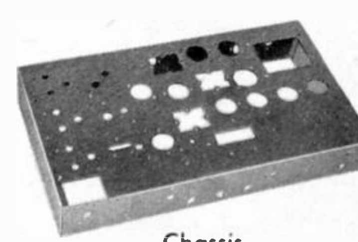
Desk Panel Cabinet Rack


Induction Heater Housing


Chassis


Marine Radio Housing


Cabinet

How Karp Makes Custom-Built Metal Cabinets and Boxes at Prices that Compete with those of Stock Items

The advantages and true economies of Karp custom-built cabinets, boxes, or housings over stock items are these:

- Your own exclusive design distinguishes and "styles" your product . . . gives it more market value.
- Flexibility of construction details speeds and simplifies your final assembly—saving you time and money.
- Our vast stock of dies can save you special die costs.
- Our 70,000 square feet of modern plant, with hundreds of craftsmen, means ample capacity for many types of work—simple or elaborate—at one time.
- Plant is fully equipped with every mechanical facility that aids economical production.
- Finishing is done in dustproof paint shop, with latest water-washed spray booths and gas-fired ovens mechanically and electronically controlled.
- We make no stock items or products of our own. Our plant, time and effort are 100% for our customers' work.
- Our engineering staff can help solve any possible design and production problems.
- It's results that count—and we give you the results you want.

Write for illustrated data book describing our facilities and showing the wide range of sheet metal fabrication we do.

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Pin point accuracy is

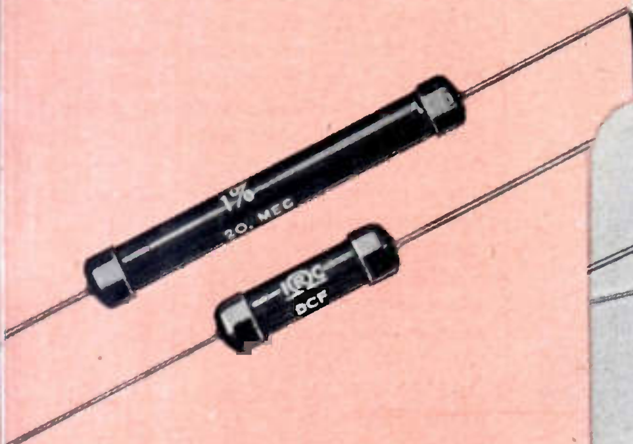
for resistors too!

And IRC provides it. Witness leading manufacturers who specify IRC resistors for advanced electronic circuits. In instrumentation and industrial applications, IRC resistors excel in every important characteristic.

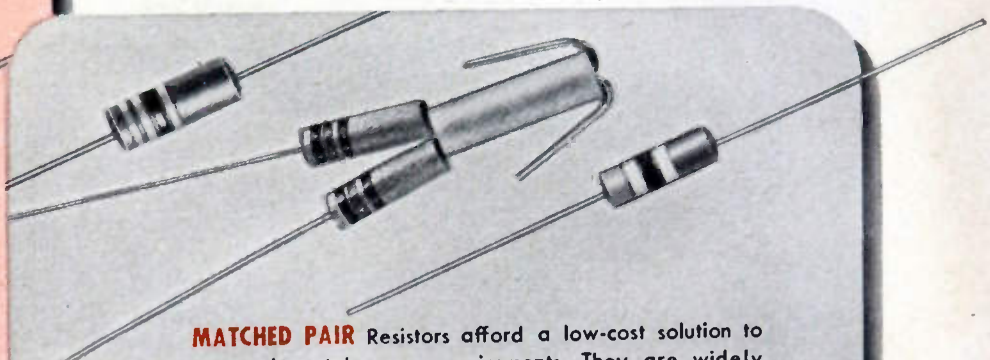
IN CRITICAL INSTRUMENTATION, IRC Precision Wire Wounds offer a fine balance of accuracy and dependability. Tolerances of 1% are standard, but ½%, ¼% and 1/10% are available. IRC Precisions also afford maximum temperature coefficient of .002% per °C. at no extra cost. And in addition, their design and construction assure stability—even where recurring surges are encountered. Labels are acetate. May we send you complete technical data? Just check the coupon.

essential

SEALED PRECISION Voltmeter Multipliers find many critical applications such as are encountered in marine service because of absolute dependability under the most severe humidity conditions. Type MF's are compact, rugged, stable, fully moisture proof and easy to install. They consist of individual wire wound precision resistors, mounted, interconnected and encased in glazed ceramic tubes—and these may be either inductive or non-inductive, for use on AC as well as DC. Send coupon for technical data bulletin.



ACCURACY AND ECONOMY in close tolerance applications make IRC Deposited Carbon PRECISTORS ideal for television and similar circuits. They are outstanding in their ability to provide dependable performance in circuits where the characteristics of carbon composition resistors are unsuitable and wire-wound precisions too expensive. Manufactured in two sizes, 200 ohms to 20 megohms in 1%, 2% and 5% tolerance. Coupon brings full details.



MATCHED PAIR Resistors afford a low-cost solution to many close tolerance requirements. They are widely used as dependable meter multipliers. Two insulated IRC resistors are matched in series or parallel to as close as 1% initial accuracy. Both JAN-R-11 approved Advanced BT resistors and low-range BW insulated wire wounds are available in Matched Pairs. Use the coupon to send for Bulletin B-3.

For fast, local service on standard IRC resistors, simply phone your IRC Distributor. IRC's Industrial Service Plan keeps him well supplied with the most popular types and ranges—enables him to give you prompt, round-the-corner delivery. We'll be glad to send you his name and address.

IRC

Wherever the Circuit Says -⁄\/\/\/-

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

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Please send me complete information on the items checked below:

- ☐ Precision Wire Wounds
- ☐ Deposited Carbon Precistors
- ☐ Wire Wound Controls
- ☐ Voltmeter Multipliers
- ☐ Matched Pair Resistors
- ☐ Name of Local IRC Distributor

NAME..

TITLE..

COMPANY..

ADDRESS..

....

INSTRUMENT SOCIETY HOLDS 4th CONFERENCE AND SHOW!

IRE Members from Region V Are Invited to St. Louis September 12-16, 1949

to attend this five day conference on all phases of instrumentation. Members will be registered free as IRE is a co-sponsoring society. Invitations will be mailed in advance to Region V members.

The beautiful Kiel Auditorium will be used for all sessions and the Exhibitions, including the National Telemetering Forum, Sept. 12 and 13, on telemetering activities relating to airborne and ground equipment, design, production and testing, and general theory; particularly as applicable to rocket instrumentation and similar work.

AIEE will conduct an Electronic Instrumentation Session, Sept. 15, and one on Electrical Instruments and Measurements, Sept. 16.

ASME will hold sessions Sept. 13 and 14 on Industrial Instruments and Regulators, one paper of which is "The Attenuation of Oscillatory Pressures in Instrument Lines," by Arthur S. Iberoll, Bureau of Standards.

The American Institute of Physics has scheduled papers on Scientific Instruments for Sept. 12 and 13, including several of importance on nucleonic instrumentation.

The Instrument Society sessions are extensive and cover the fields of Instrument Maintenance, Inspection and Gaging, Transportation Instruments and Instruments for Testing.

For a detailed program, write to William C. Copp, IRE Adv. Dept., 303 West 42nd St., New York 18, N.Y.

1949 IRE West Coast Convention and Pacific Electronic Exposition
Aug. 30 to Sept. 2, 1949
Civic Auditorium
San Francisco, Calif.

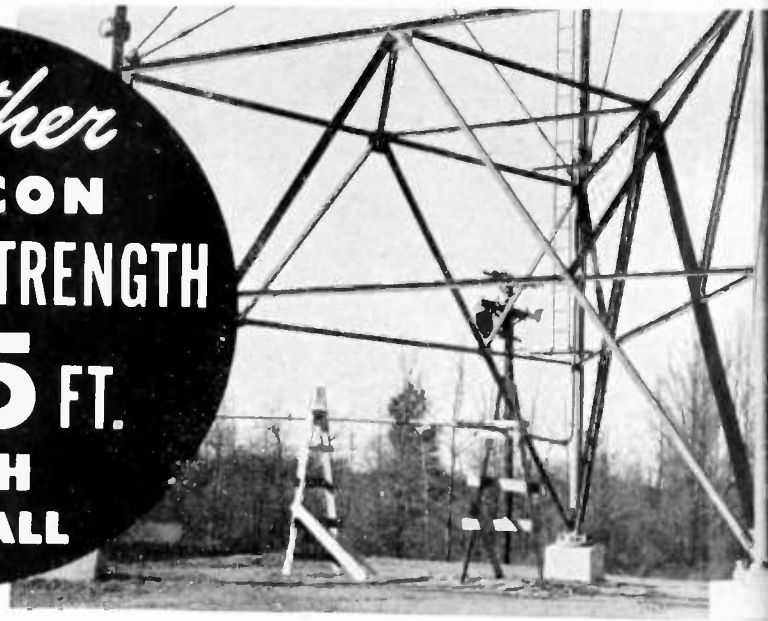
IRE Regional Meetings Accelerate Electronic Progress

Meeting FM and TV Needs for WRVB, Richmond, Virginia

● Rising high above the Tideland, this Truscon Self-Supporting Steel Radio Tower helps flash the cream of FM and TV entertainment to a great circle of Virginia audiences. The business end of this slender steel beauty mounts both an R.C.A. 2-section pylon FM antenna, and an R.C.A. 6-section TV antenna.

This outstanding installation emphasizes again the fact that every Truscon Steel Radio Tower is *fitted exactly to its specific location.* Truscon Radio Towers today are operating faithfully under world-wide extremes of weather . . . under almost every possible combination

Another
TRUSCON
TOWER OF STRENGTH
485 FT.
HIGH
OVERALL



of temperature, humidity and wind conditions . . . in mountains, deserts, plains and coastal areas.

Truscon engineers are ready now to put their vast experience at your service—ready to design and erect just the tower you need for AM, FM and TV broadcasting . . . tall or small, guyed or self-supporting, of uniform or tapered cross-section. Call the Truscon office nearest you or write our home office in Youngstown. There is no obligation.

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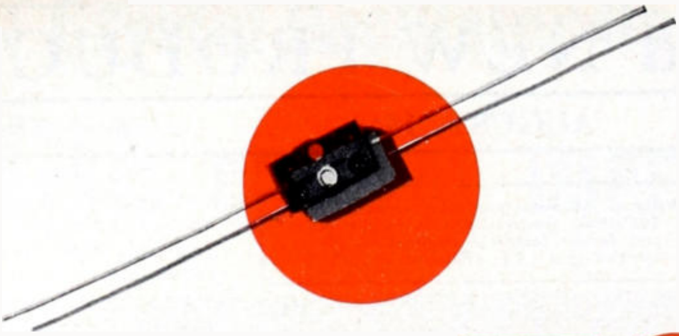
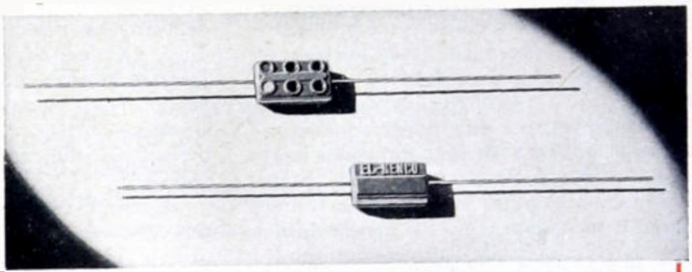
TRUSCON
SELF-SUPPORTING AND UNIFORM CROSS SECTION GUYED
TOWERS



EL-MENCO'S PLACE **in** YOUR PRODUCT

Wherever fixed mica dielectric capacitors are used, the first choice with men of experience is always El-Menco



Precision-made under rigid conditions, tested seven ways to meet strict Army-Navy standards, thoroughly impregnated and provided in water-sealed low-loss bakelite; these tiny capacitors protect and maintain your reputation for quality equipment. To insure performance-excellence, place *El-Menco* capacitors in *your* product. Results will prove El-Menco to be a wise choice.

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

CM 15 MINIATURE CAPACITOR

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

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CAPACITORS

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

NEWS and NEW PRODUCTS

AUGUST, 1949

50-Mc TV Oscilloscope

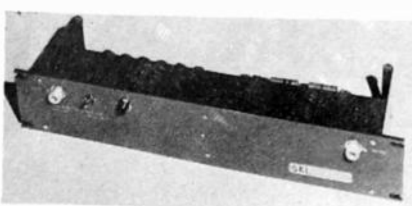
The FTL-32A a broad-band TV oscilloscope, which records wave forms having frequency components as high as 50 Mc and as low as 10 cps, is presented by **Federal Telecommunications Labs., Inc.,** 500 Washington Ave., Nutley 10, N. J.

This wide-band frequency response is obtained with sufficient amplification to provide deflection sensitivity of 0.1 peak-to-peak volts per inch. The horizontal amplifier has a bandwidth of 10 cps to 10 Mc.

Both repetitive and triggered sweeps are incorporated with time durations consistent with the 50 Mc bandwidth. Synchronization from an internal or external source is independent of the synchronization signal amplitude, provided a minimum of 0.1 volt is exceeded.

New Traveling-Wave Amplifier

A new addition to their line, Model 202 Wide Band Chain Amplifier, a traveling-wave type, is announced by **Spencer-Kennedy Labs., Inc.,** 186 Massachusetts Ave., Cambridge 39, Mass.



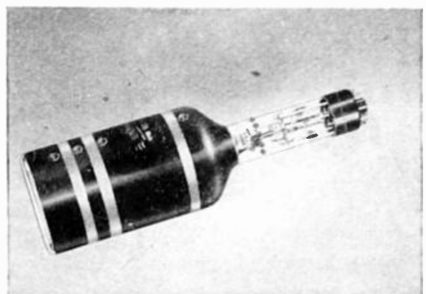
The model 202 has 2 stages of six 6AK5 tubes, gain of 20 db, bandwidth of 200 Mc. SWR is less than 1.5 db, and transmission characteristic is ±1.5 db from 100 kc to 200 Mc at an impedance level of 200 ohms.

The amplifier employs a traveling-wave circuit to achieve its wide bandwidth; thus it is adaptable for use with signal, pulse, and sweep generators, vacuum-tube voltmeters, and TV test equipment.

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

Sensitive Multiple-Intensifier Type CR Tube

A new multiple-intensifier type cathode-ray tube featuring a highly sensitive vertical-deflection system, and known as the Type 5XP-, is announced by **Allen B. Du Mont Laboratories, Inc., Instrument Div.,** 1000 Main Ave., Clifton, N. J.

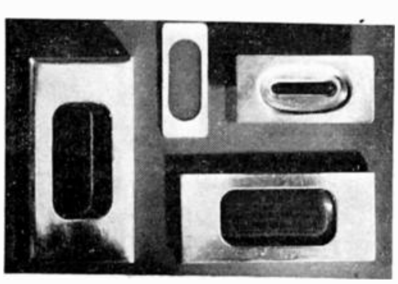


Potentials as low as 24 to 36 volts peak-to-peak are sufficient for 1" vertical deflection on the screen. Even though the Type 5XP employs high accelerating potentials, the manufacturer claims its deflection factor for plate-pair D_3-D_4 is but one-third of the deflection factor of similar tubes operating at low accelerating potentials.

A new type of deflection-plate structure is used in which the length of one pair of plates is considerably increased while the width is decreased. Because of this new deflection-plate design, the greater sensitivity of the tube is achieved with a plate-to-plate capacitance of only 1.7 $\mu\mu$f. A $\frac{7}{8}$" increase in the over-all length of the tube is also a contributing factor to the high deflection sensitivity.

Improved Glass-Metal Seal for Waveguide Windows

Glass waveguide windows, designed to permit silver soldering (without damage) to micro-waveguide systems operating at frequencies from 3,000 to 40,000 Mc, have been announced by **Electronics Div., Sylvania Electric Products Inc.,** 500 Fifth Ave., New York 18, N. Y.



Development of these resonant windows, in which glass stress is eliminated at relatively high temperature differentials required for soldering, makes available a wide range of window shapes and outside contours, for narrow- and wide-band transmissions. Power losses range from 0.02 to 0.1 db. For frequencies above 3,500 Mc, the windows will stand pressures up to 65 pounds per square inch absolute.

Continuous Loop Drive Tape Recorder Mechanism

Continual repetition of any recorded message, from 4 seconds to 10 minutes duration, is possible through the use of a continuous loop drive mechanism, available as optional equipment on any standard Twin-Trax Tape recorder, by **Twin-Trax Div., Amplifier Corp. of America,** 398-1 Broadway, New York 13, N. Y.

This mechanism is detachable to facilitate installation on previously purchased Twin-Trax models.

Two variations of the drive are available. The model CL-3 will record and play back information from 4 seconds to 3 minutes in duration. During operation, a sufficient length of tape is spooled on a stationary hub. The inside layer of tape is fed through a slot in the hub, past an idler, past the record-playback head, and subsequently engaged by the pulling capstan. The beginning of the tape is then joined to the end to form a continuous loop which will continue to repeat until it is manually or automatically shut off.

Model Cl-10 will reproduce information from 4 seconds to 10 minutes, and operates on the same principle, but utilizes a storage system with lower inherent friction which is accomplished with ball bearing rollers.

(Continued on page 24A)

THE FUTURE HOLDS GREAT PROMISE

Neither chance nor mere good fortune has brought this nation the finest telephone service in the world. The service Americans enjoy in such abundance is directly the product of their own imagination, enterprise and common sense.

The people of America have put billions of dollars of their savings into building their telephone system. They have learned more and more ways to use the telephone to advantage, and have continuously encouraged invention and initiative to find new paths toward new horizons.

They have made the rendering of telephone service a public trust; at the same time, they have given the telephone companies, under regulation, the freedom and resources they must have to do their job as well as possible.

In this climate of freedom and responsibility, the Bell System has provided service of steadily increasing value to more and more people. Our policy, often stated, is to give the best possible service at the lowest cost consistent with financial safety and fair treatment of employees. We are organized as we are in order to carry that policy out.

Bell Telephone Laboratories lead the world in improving communication devices and techniques.

Western Electric Company provides the Bell operating companies with telephone equipment of the highest quality at reasonable prices, and can always be counted on in emergencies to deliver the goods whenever and wherever needed.

The operating telephone companies and the parent company work together so that improvements in one place may spread quickly to others. Because all units of the System have the same service goals, great benefits flow to the public.

Similarly, the financial good health of the Bell System over a period of many years has been to the advantage of the public no less than the stockholders and employees.

It is equally essential and in the public interest that telephone rates and earnings now and in the future be adequate to continue to pay good wages, protect the billions of dollars of savings invested in the System, and attract the new capital needed to meet the service opportunities and responsibilities ahead.

There is a tremendous amount of work to be done in the near future and the System's technical and human resources to do it have never been better. Our physical equipment is the best in history, though still heavily loaded, and we have many new and improved facilities to incorporate in the plant. Employees are competent and courteous. The long-standing Bell System policy of making promotions from the ranks assures the continuing vigor of the organization.

With these assets, with the traditional spirit of service to get the message through, and with confidence that the American people understand the need for maintaining on a sound financial basis the essential public services performed by the Bell System, we look forward to providing a service better and more valuable in the future than at any time in the past. We pledge our utmost efforts to that end.

LEROY A. WILSON, *President*
American Telephone and Telegraph Company.
(From the 1948 Annual Report.)

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

Follow the Leaders to

Eimac TUBES

The Power for R-F

With the increasing demand for higher powers at higher frequencies the importance of close relationship between tube and circuit design has become paramount.

A large segment of the laboratory facilities at Eitel-McCullough is concerned with the development of basic new circuits closely correlated with vacuum tube development. The efforts of this group are receiving wide recognition for their outstanding accomplishments. These new circuits are being made available, as developed, to the industry enabling greater realization of a vacuum tube's potential abilities.

Evidence of these efforts is illustrated above . . . A 14-tube annular r-f generator. This compact equipment can provide 500 watts of CW power at 1000-Mc, and has operating possibilities as high as 2500-Mc. This is but one application of the basic annular circuit design developed by Eimac. The power-output in such a generator is directly proportional to the number of tubes used, and single tube efficiency is maintained.

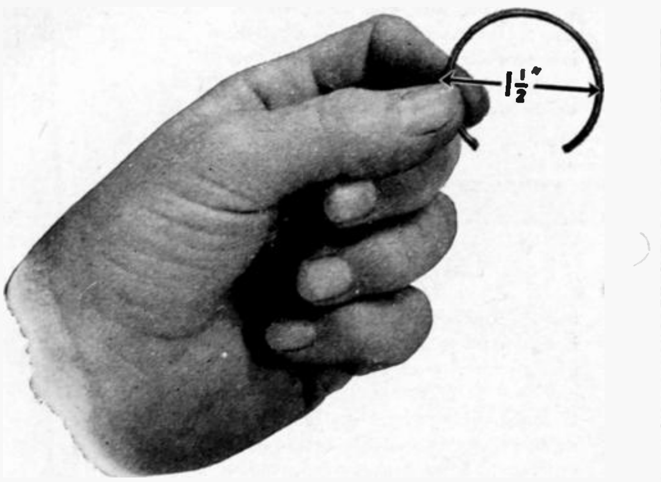
EITEL-McCULLOUGH INC.

728 SAN MATEO AVE., SAN BRUNO, CALIFORNIA

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

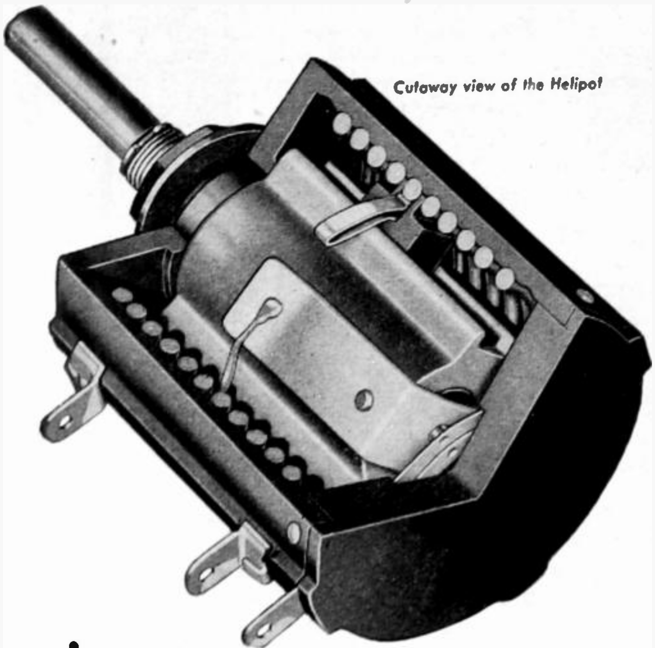
Compare!

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!



Cutaway view of the Helipot

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%, .05%, 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 is for Model A unit*

*** HELIPOTS ARE AVAILABLE IN MANY SIZES:**

MODEL A—5 watts, incorporating 10 helical turns and a slide wire length of 46 inches, case diameter 1¾", is available with resistance values from 10 ohms to 300,000 ohms.

MODEL B—10 watts, with 15 helical turns and 140" slide wire, case diameter 3¼", is available with resistance values from 50 ohms to 500,000 ohms.

MODEL C—2 watts, with 3 helical turns and 13½" slide wire, case diameter 1¾", available in resistances from 5 ohms to 50,000 ohms.

MODEL D—15 watts, with 25 helical turns and 234" slide wire, case diameter 3¼", available in resistances from 100 ohms to 75,000 ohms.

MODEL E—20 watts, with 40 helical turns and 373" slide wire, case diameter 3¼" is available with resistance values from 200 ohms to 1,000,000 ohms.

Other types and designs of Potentiometers available.

Send for the New Helipot Booklet!

THE Helipot CORPORATION, SOUTH PASEDENA 6, CALIFORNIA

For today's broadcast power needs . . .

RECTIFIER TUBES

★ **PACEMAKERS IN DESIGN**

★ **SERVICE-PROVED**

★ **AVAILABLE IN A WIDE RANGE OF CAPACITIES**

MODULATION, amplification, final output, all need d-c power . . . continuous, *dependable* if off-the-air periods are to be avoided. Your rectifier tubes are basic; good rectifier tubes make for good broadcasting. So buy General Electric—buy the best!

Design improvement is constant, with G-E rectifier types ever-new in their efficiency. For example, the new straight-side bulbs of Types GL-8008 and GL-673 give an increased temperature margin of safety; their slim contour also makes the tubes easier to handle, better to install.

Future AM-FM-TV power-requirement possibilities are matched by new G-E designs, such as the GL-5630 ignitron for a-c to d-c conversion. With this high-capacity tube it is possible to supply—economically, reliably—direct current in large amounts to broadcast transmitters.

If you build or design equipment, phone your nearby G-E electronics office for expert assistance in selecting the right G-E rectifier types. There are more than a dozen from which to choose. If a station operator needing tubes for replacement, your local G-E tube distributor will be glad to serve you promptly, efficiently, out of ample stocks on hand. *Electronics Department, General Electric Company, Schenectady 5, N. Y.*

GENERAL ELECTRIC

180-H5

FIRST AND GREATEST NAME IN ELECTRONICS

Type	Cathode voltage	Cathode current	Anode peak voltage	Anode peak current	Anode avg current
GL-866-A	2.5 v	5 amp	10,000 v	1 amp	0.25 amp
GL-8008	5 v	7.5 amp	10,000 v	5 amp	1.25 amp
GL-673	5 v	10 amp	15,000 v	6 amp	1.5 amp
GL-869-B	5 v	19 amp	20,000 v	10 amp (*20 amp)	2.5 amp (*5 amp)
GL-857-B	5 v	30 amp	22,000 v	40 amp	10 amp

(*Quadrature operation)


in communications...
IRELAND
NORWAY
SWEDEN
DENMARK
PORTUGAL
ICELAND
ARGENTINA
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Interpreted according to RMA standard M8-210

Heater Voltage (± 10%)	12.6	6.3	VOLTS
Maximum Heater-Cathode Voltage		90	VOLTS
Maximum Plate Voltage		300	VOLTS
Maximum Inverse Plate Voltage		1000	VOLTS
Maximum Plate Dissipation (each unit)		4.2	WATTS
Maximum Total Plate Dissipation (both units)		7.5	WATTS
Maximum Bulb Temperature (at any part of envelope)		220°	C
Maximum DC Grid Current (each unit)		6	MA.
Maximum External Grid Circuit Resistance (each unit)		1	MEG.

CHARACTERISTICS

Class A1 Amplifier—Each Unit

Heater Voltage	12.6	6.3		VOLTS
Heater Current	450	900		MA.
Plate Voltage	120	180	250	VOLTS
Grid Voltage	−2	−7	−12.5	VOLTS
Plate Current	36	23	16	MA.
Plate Resistance	1650	2750	4000	OHMS
Transconductance	11000	6400	4100	μMHOS
Amplification Factor	18	17.5	16.5	
Grid Voltage (approx.) For I_b—100 μA	−10	−15	−21	VOLTS

For more complete information about the 5687, write for these bulletins.

The Tung-Sol engineering which has produced the 5687 is constantly at work on a multitude of special electron tube developments for industry. Many exceptionally efficient general and special purpose tubes have resulted. Information about these and other types are available on request to Tung-Sol Commercial Engineering Department.

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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 14A)

Dc Microammeter and Magnetic Amplifier

A new microammeter, Type 100, designed to measure low dc is announced by **W. S. MacDonald Co., Inc.**, 33 University Rd., Cambridge 38, Mass.

The Type 100 has an input resistance of 50 ohms and a sensitivity of 1 ma full scale. The input may be overloaded ¼ ampere without causing damage. Unlike high sensitivity galvanometers, this instrument is not sensitive to position or vibration.

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A new addition to their line of test equipment, the Model 320 Signal Generator, is announced by **Electronic Instrument Co., Inc.**, 276 Newport St., Brooklyn 12, N. Y.

Designed for use in all phases of the radio industry, the Model 320 may be used for FM-AM alignment and to provide TV marker frequencies. The meter features a Hartley oscillator with a range of of 150 kc to 100 Mc, with fundamentals to 34 Mc. A Colpitts type audio oscillator supplies 400 cps sine wave voltage for modulation.

As are the majority of Eico products, this model is also available as a kit.

(Continued on page 47A)

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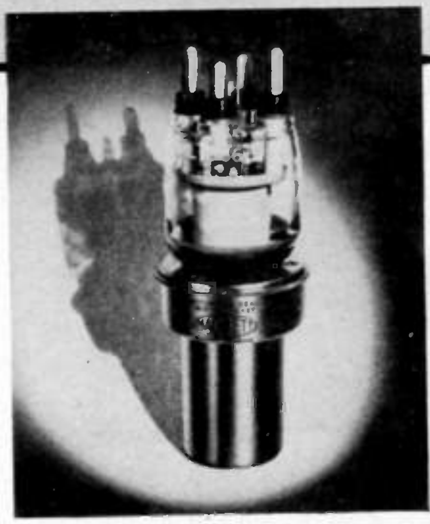

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* Adopted by Military Services, U. S. Government Agencies, and other large users as the standard replacement for 889RA, the 5667 is now their preferred tube type for 889RA sockets.

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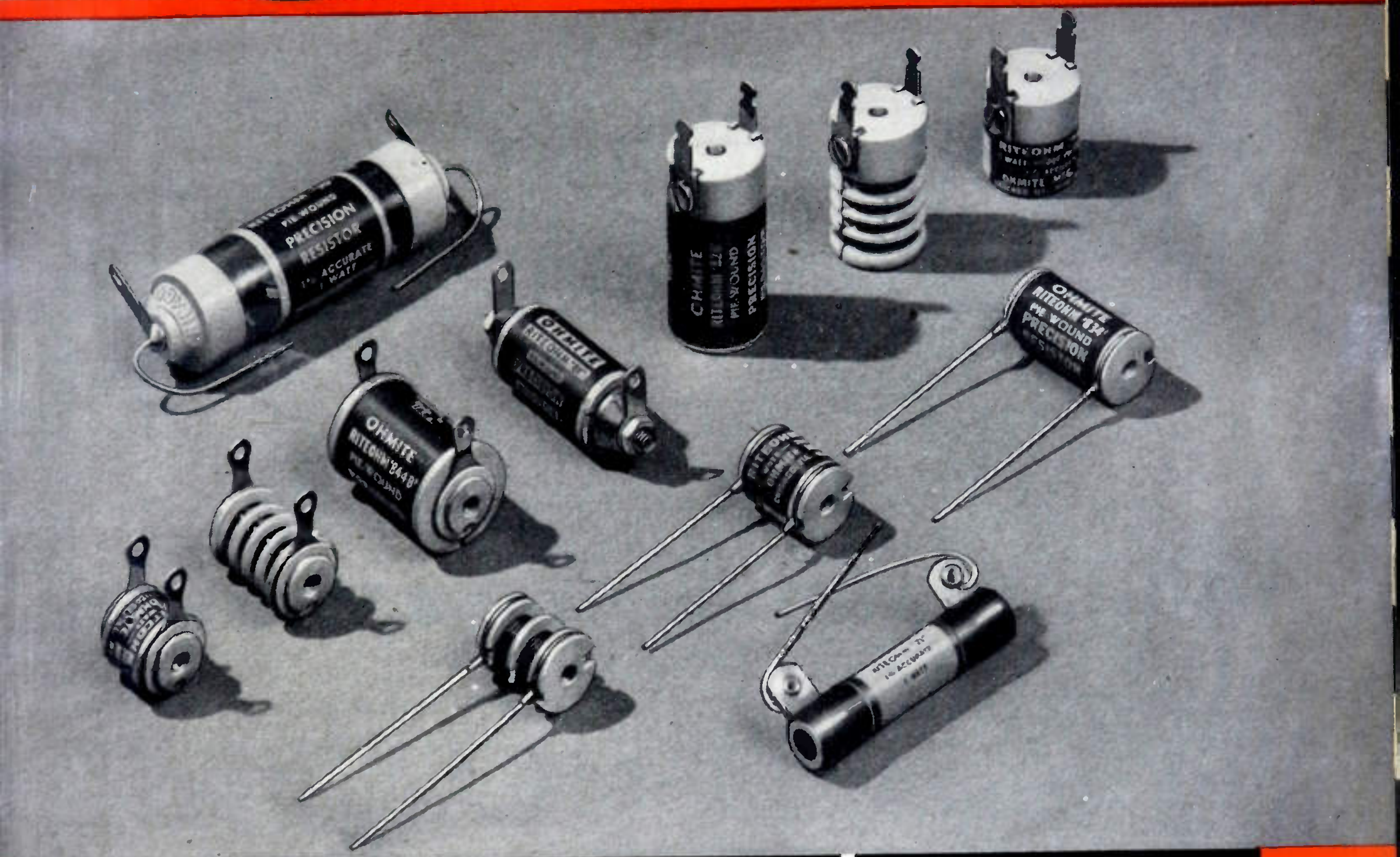

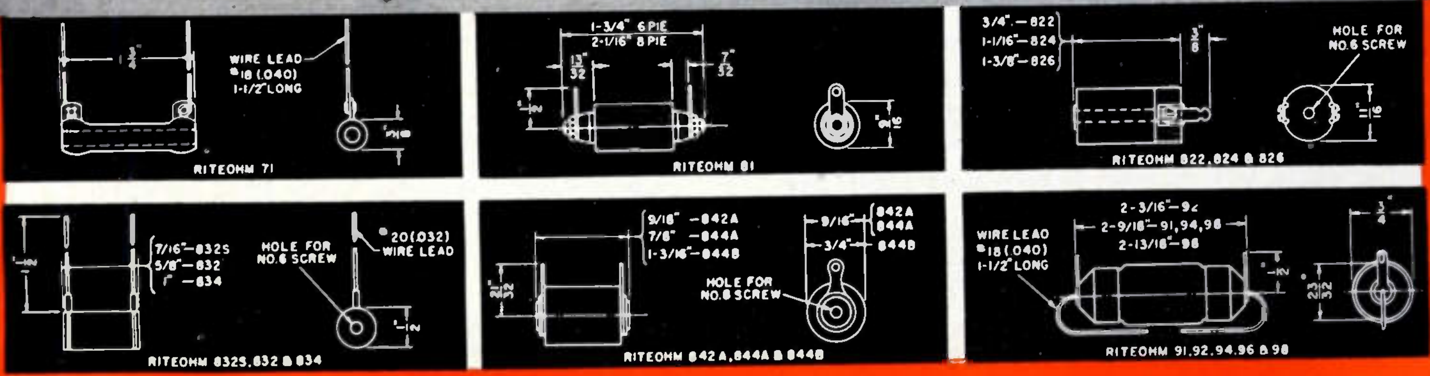

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PROCEEDINGS OF THE I.R.E.

(Including the WAVES AND ELECTRONS Section)

Published Monthly by

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VOLUME 37 — *August, 1949* — NUMBER 8

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

WAVES AND ELECTRONS SECTION

† Deceased

William L. Everitt

BOARD OF DIRECTORS, 1949–1951

William Litell Everitt was born on April 14, 1900, in Baltimore, Maryland. During World War I he served in the U. S. Marine Corps from 1918 to 1919. At the war's conclusion, he matriculated at Cornell University, where he taught electrical engineering from 1920 until he received the E.E. degree in 1922.

In that year he joined the North Electric Manufacturing Company of Galion, Ohio, as engineer in charge of the design and development of their relay automatic public switchboard exchanges. He left in 1924 to teach electrical engineering at the University of Michigan, transferring to Ohio State University in 1926—the year in which he received the M.A. from Michigan—to take charge of their communications engineering staff in the capacity of assistant professor. Meanwhile, during the summers from 1925 to 1930, he served with the department of development and research of the American Telephone and Telegraph Company.

Joining The Institute of Radio Engineers as an Associate Member in 1925, he became a Member in 1929, and also was elevated to the rank of associate professor at Ohio State in that year. In 1933 he received the Ph.D. degree and was promoted to a full professorship. At Ohio State he originated and directed the annual Broadcast Engineering Conference, in which the IRE participated. He became a Fellow of the IRE in 1938.

In 1940 Dr. Everitt was appointed a member of the Communications Section of the National Defense Research Committee. Two years later he took a leave of absence from the University to serve as director of operational research with the U. S. Army Signal Corps. He was appointed head of the University of Illinois' electrical engineering department *in absentia* in 1945, and at the war's conclusion he took up his duties there. Effective September 1, 1949, he will become Dean of the University of Illinois' engineering college.

The author and editor of numerous texts and articles on electrical engineering, Dr. Everitt has been a consultant for various broadcast stations and radio manufacturing companies. He was President of the IRE in 1945 and has been a member of a number of committees; currently he is on the Circuits Committee and the Board of Editors. Dr. Everitt is also a Fellow of the AIEE, a member of the National Council of Tau Beta Pi, and a member of Sigma Xi, Eta Kappa Nu, the Acoustical Society of America, the American Society for Engineering Education, and the American Association for the Advancement of Science.

The Valuable By-Product

V. K. ZWORYKIN

Scientists, in their systematic study of phenomena and their development of the covering theories, pursue a lengthy and sometimes tortuous path. Shall they always head unswervingly toward the distant goal or should they occasionally pursue, or at least point out, interesting side paths?

These matters are considered in the following guest editorial by an eminent research worker, the correctness and fruitfulness of whose methods have been amply proven. He is a vice-president and technical consultant of the RCA Laboratories, and a Fellow of the IRE and member of the Board of Editors.—*The Editor.*

"Select your goal carefully, and then, looking neither to left nor right, pursue it."

That is good advice. Yet, like most good advice, it can be overstressed. Much that is of great value has come from the exploration of bypaths in science and engineering. Certainly, few will dispute Röntgen's wisdom in ferreting out the cause of the lighting up of a fluorescent screen near his enclosed gas discharge tube, and studying the phenomenon with such thoroughness that he was able not only to announce to the world the existence of X rays, but also to give an extraordinarily complete description of their characteristics. Perhaps the ideal compromise between persistence in a chosen line of investigation and the exploration of an incidental find was practiced by Hertz in the discovery of the photoelectric effect. Having noted unexplained nonuniformities in the length of induced sparks, he performed a series of critical experiments with available equipment, which demonstrated that these arose from the action of ultraviolet light on the negative spark electrode. After publishing a careful description of these experiments for the guidance of later investigators, Hertz returned to his original objective—the demonstration of electric waves as predicted by Maxwell's theory—and thus helped to lay the basic groundwork for the radio industry.

There are many examples of the valuable by-product in our own experience in the development of television. Perhaps the first that comes to mind is the multiplier phototube which now plays such a vital role as detector in nuclear and cosmic-ray work, and has found innumerable other applications in industry and research. Our immediate interest in secondary-emission multiplication stemmed from the need of increasing the signal strength of television pickup tubes. Again, the electron microscope, which has extended the range of our vision by two orders of magnitude, naturally fitted into the orbit of our interests as the result of our intensive occupation with the electron-optical problems of television; the circuit experience gained in television development helped to create an instrument of compact dimension with a resolving power approaching the theoretical limit. The infrared-sensitive image tube, which came to play an important part as reconnaissance instrument during the war, was another early by-product of pick-up tube research. And now, a variety of storage tubes, the offspring of television research in tube construction, operating principles, and circuitry, promise to make material contributions to air safety through systems such as Teleran, to facilitate the study of high-speed phenomena, and to bring within reach the solution of mathematical problems of baffling complexity by modern electronic computing machines.

The list is not complete, nor is there any sign of a halt in the flow of new ideas and new devices that spring from television development. There can be little question that the same applies to any number of other lines of endeavor. It is with this in mind that I would change the advice given above to the following:

"Select your goal carefully, and then, looking both left and right, proceed on your way, marking out promising side paths as they meet your eyes. Thus, you may reach your goal sooner and help others, in turn, to reach their goals."

Path-Length Microwave Lenses*

WINSTON E. KOCK†, SENIOR MEMBER, IRE

Summary—**Lens antennas for microwave applications are described which produce a focusing effect by physically increasing the path lengths, compared to free space, of radio waves passing through the lens. This is accomplished by means of baffle plates which extend parallel to the magnetic vector, and which are either tilted or bent into serpentine shape so as to force the waves to travel the longer-inclined or serpentine path. The three-dimensional contour of the plate array is shaped to correspond to a convex lens. The advantages over previous metallic lenses are: broader band performance, greater simplicity, and less severe tolerances.**

Introduction

LENSES ARE useful microwave antennas because of their tolerance advantages. An amount of twist or warp of the completed antenna which would seriously degrade the gain and directional properties of a parabolic dish reflector antenna will generally have negligible effect upon the performance of an equivalent lens.

Two types of metallic microwave lenses developed by the Bell Telephone Laboratories have previously been described. The first comprised rows of conducting plates which acted as waveguides and achieved a focusing effect by virtue of the higher phase velocity of electromagnetic waves passing between the plates.[1]

The second type was a scaled-up version of the lattice structure of a true dielectric, whereby small conducting elements replaced the molecules of the dielectric, and the polarization which these elements engendered duplicated the polarization of the true dielectric.[2] This type of lens exhibited very broad frequency characteristics, so that it was effective over a much larger wavelength band than the previous waveguide lens. Because, however, the elements were small compared to the wavelength, large size lenses of this type contained large numbers of the conducting elements, and the construction became somewhat tedious. For example, the partially assembled 10-foot delay lens shown in Fig. 1 (strip type structure) requires several thousand strips to fill the 10 foot aperture. The lenses to be described in this paper retain the broad-band features of the delay type, but permit of simpler construction.

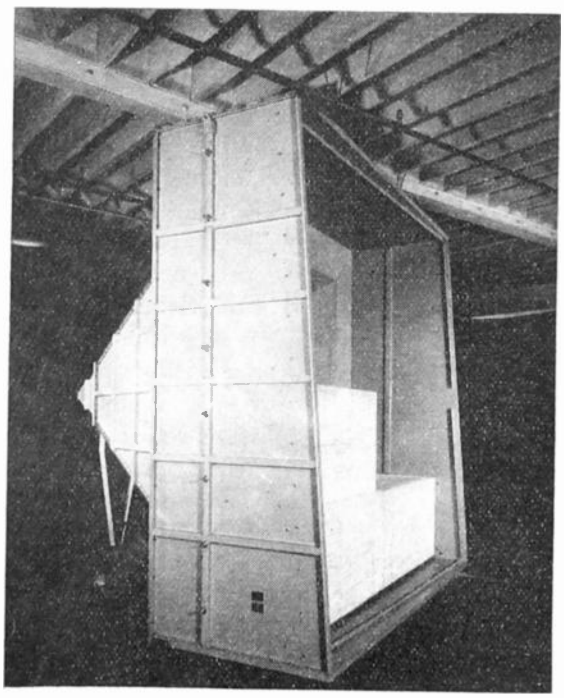

Fig. 1—A partially assembled 10-foot artificial dielectric shielded-lens antenna for microwave relay use.

Fundamental Principles

The principle of operation of the path-length lenses can be described in connection with Fig. 2. If parallel

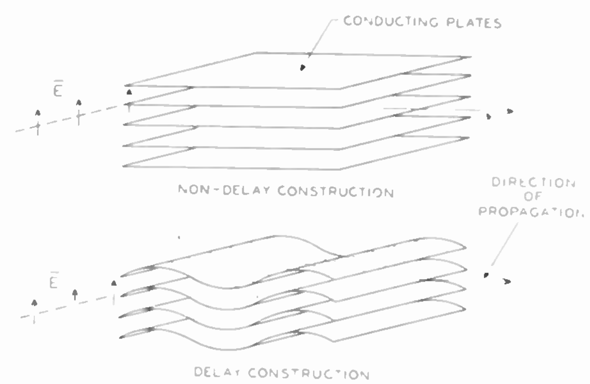

Fig. 2—Use of conducting plates or grids (shown dotted) to effect a wave delay.

conducting plates are presented to electromagnetic waves polarized perpendicularly to the plates, little effect will be produced on the progress of the waves, providing that the plates are flat and aligned along the direction of propagation, as shown in the top of the figure. If, however, the plates are bent into serpentine shape, as shown in the bottom of the figure, the sinuous path l, inside the plates, will be longer than that outside l_0, and delay will be produced. In order to avoid the generation of a second-order mode, the vertical spacing of the plates must be less than one-half wavelength, but for all wavelengths longer than this, a constant delay and thus a constant refractive index is obtained. If, instead of the serpentine construction, the plates are tilted so that they form an angle with the direction of propagation, a delay will also be produced, since the waves will be forced

* Decimal classification: R326.8×R310. Original manuscript received by the Institute, January 11, 1949.
† Bell Telephone Laboratories, Inc., Murray Hill, N. J.
[1] W. E. Kock, "Metal-lens antennas," Proc. I.R.E., vol. 34, pp. 828–837; November, 1946.

[2] W. E. Kock, "Metallic delay lenses," *Bell Sys. Tech. Jour.*, vol. 27, pp. 58–83; January, 1948.

to traverse the longer inclined path. Instead of solid plates, a grid structure can be used if desired, as shown by the dotted lines in the bottom of Fig. 2. The wires of the grid must be less than one-half wavelength apart in order to simulate the effect of a solid sheet.

Design Considerations

The design of microwave lenses using the path-length principle follows, in general, the design of any lens in which the material comprising the lens has an index of refraction greater than unity. For example, in Fig. 3, the contour $ACGHD$ defines the cross section of a slant plate lens looking end on at the plates, and the curved portion ACG is obtained by making the path lengths of all rays equal from the focus to the plane BEI. The added delay or increased path length of ray 2 is equal to the added length of the line CD, compared to its free-space path (the thickness of the lens from point C to the front surface). The effective index of refraction is thus equal to $1/\text{cosine}\ \theta$.

When the lens has a flat front surface, the curved side, which is towards the feed, turns out to be a hyperboloid of revolution, as in the usual case of dielectric lens. The equation of the generating hyperbola of this surface is[3]

$$(n^2 - 1)x^2 + 2fx(n - 1) - y^2 = 0, \qquad (1)$$

where n is the effective index of refraction ($1/\cos\theta$ for the slant plate and l/l_0 for the serpentine lens), f is the focal length, and the origin of the co-ordinate system is at the point C in Fig. 3.

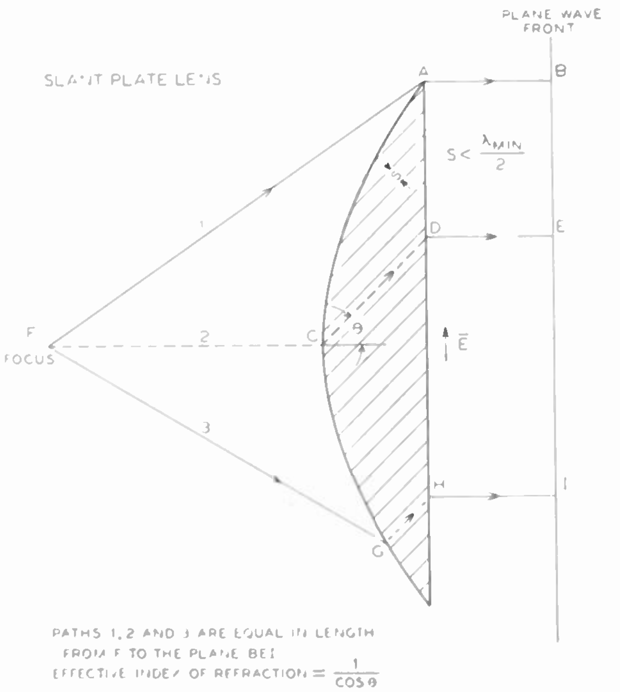

Fig. 3—Determinaton of lens contour for path-length lenses of the slant-plate variety.

For the lens reversed, so that rays starting from the focal point strike the flat surface, the contour of the outer curved surface must assume an asymmetrical shape in order that a plane wave emerge from it.

Construction Methods

Two types of lightweight grid structures are shown in Fig. 4. In both cases, the vertical spacing b and the horizontal spacing a of the grid wires must be less than one-half wavelength for proper operation, as explained earlier. The serpentine or slanted wires are affixed to polystyrene foam slabs, and the contour of the wire grid outline on each slab is a section parallel to the axis of the hyperboloid of revolution generated by (1). Polystyrene foam (Styrofoam, Dow), is light in weight and transparent to microwaves.

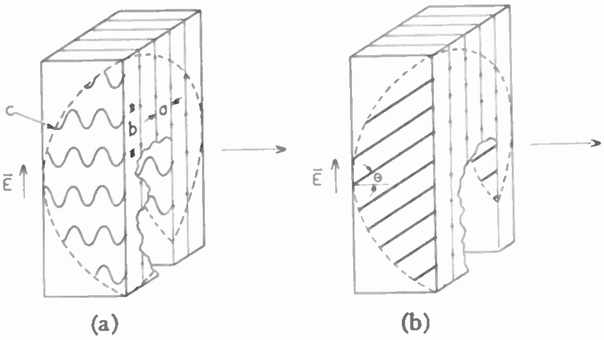

Fig. 4—Wire grid structures for path-length lenses. (a) Serpentine grid construction; (b) slant grid construction.

A photograph of a serpentine lens of the type sketched in Fig. 4(a) is shown in Fig. 5. This lens is 12 inches square and was designed for 3-cm microwaves. A two-foot square slanted wire lens of the type sketched in Fig. 4(b) is shown in Fig. 6. It was designed for use at 7-cm wavelengths. One of the foam slabs with grid wires attached is shown separately, and the remaining slabs have not yet been put into the frame.

Fig. 5—One-foot diameter serpentine wire lens for 3-cm wavelengths.

Fig. 7 shows the magnetic plane pattern of this lens when illuminated with a small feed horn. The feed horn

[3] See Fig. 17 in footnote reference 2.

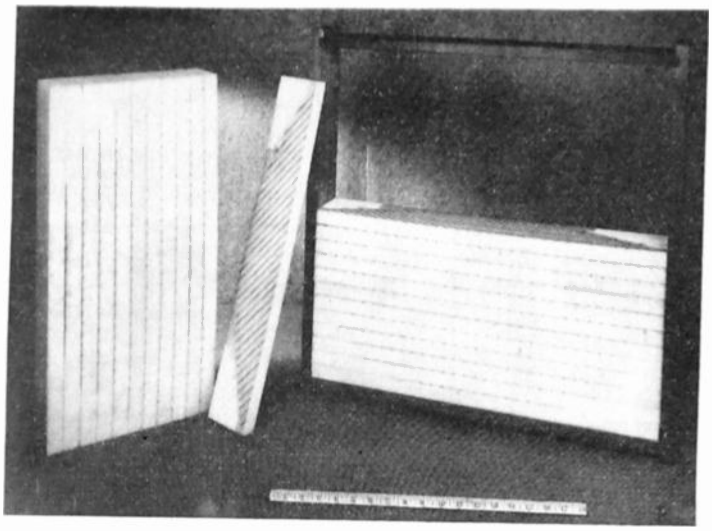

Fig. 6—A partially assembled slant wire lens for 7-cm operation.

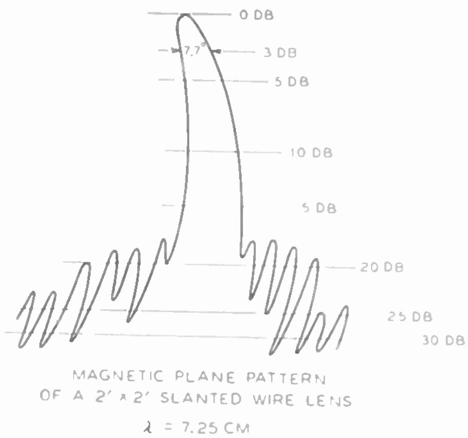

Fig. 7—Radiation pattern of the lens of Fig. 6.

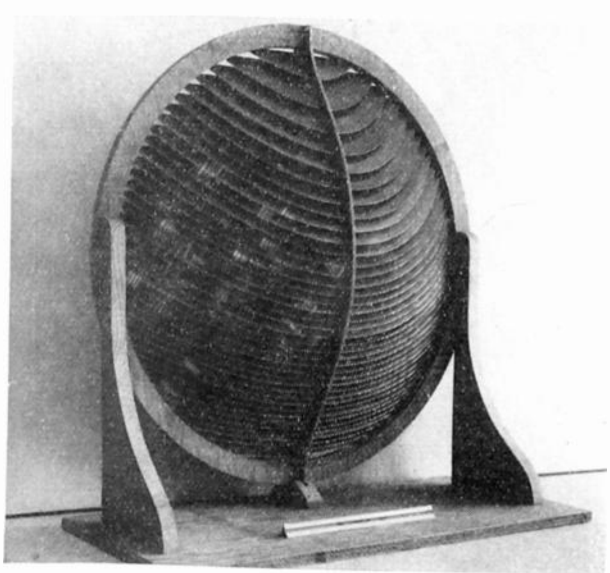

Fig. 8—Continuous sheet slant plate lens 30 inches in diameter.

had insufficient directivity to suppress the minor lobes as well as might be desired, but the symmetry and beam width of the pattern indicate satisfactory operation of the lens.

For slant plate lenses employing solid plates instead of grids, the profiles of the plates again are sections of the design hyperboloid, but they are now formed by planes making an angle θ with the axis of revolution. This is shown in Fig. 8, where a 30-inch diameter open-structure solid plate lens for use at microwavelengths from 2.5 cm upward is portrayed. In this lens, aluminum plates are supported at the edges and spaced by an insulating member along the central plane. It has a focal length (f in equation (1)) of 30 inches, and an angle of tilt of the plates of 48.2 degrees, corresponding to an index of refraction (n in equation (1)) of 1.5. Because of its open structure, air can pass readily between the plates.

A magnetic plane pattern of this lens is shown in Fig. 9, and it is seen that the minor lobes are fairly well sup-

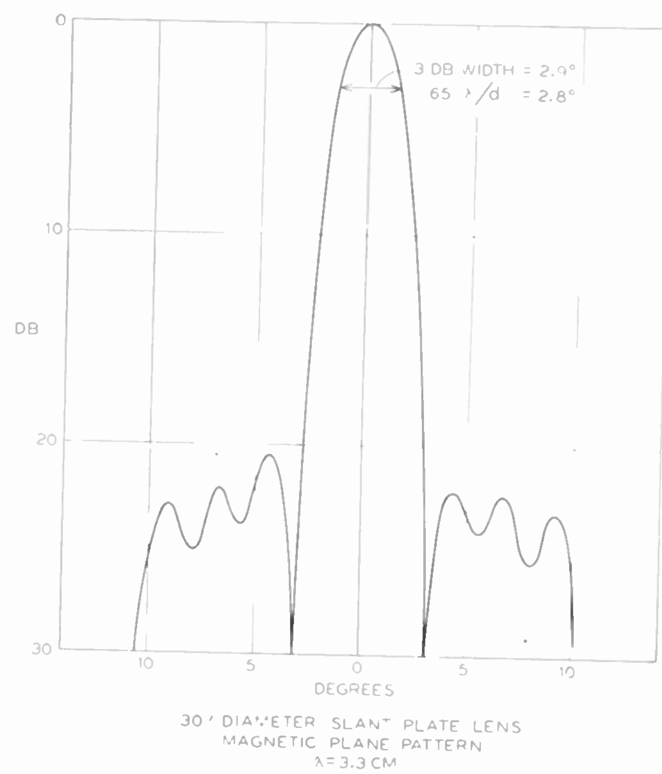

Fig. 9—Horizontal radiation pattern of the lens of Fig. 8.

pressed. A 3-inch diameter feed horn was used in this test and the wavelength was 3.3 cm. Because of the shift in energy distribution in the vertical plane as the waves pass through the tilted plates of the lens, the minor lobes in the electric plane are not as well suppressed as in the magnetic plane (Fig. 10).

The continuous plate lens of Fig. 8 could be made lighter in weight by replacing the aluminum plates with copper foil sheets of proper contour, and then employing foam slabs as spacers and supports between the plates. Currents flow along lines lying in vertical planes, so that the lens could be sectioned vertically for assembly purposes in the case of large structures.

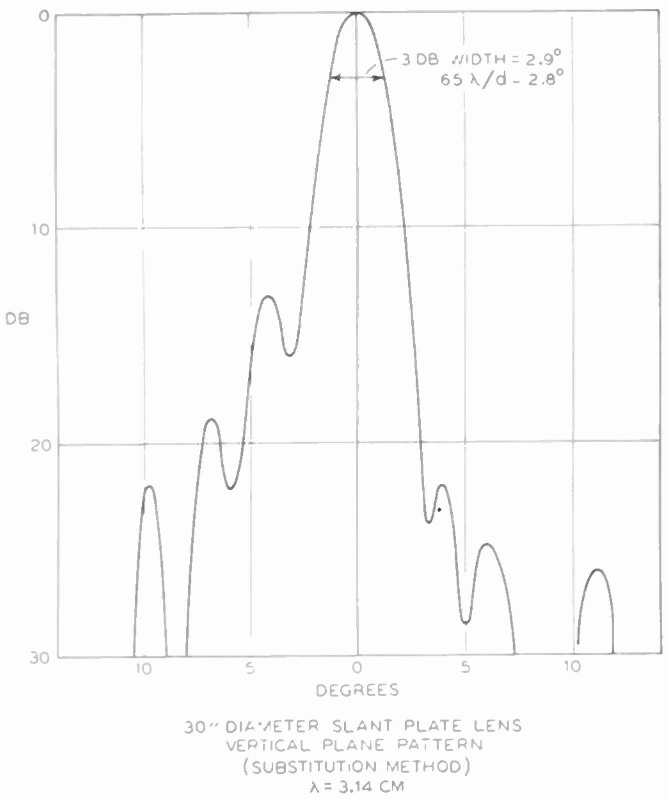

Fig. 10—Vertical radiation pattern of the lens of Fig. 8.

Comparisons with Earlier Types

Because the path-length lenses function identically for all wavelengths longer than the second mode wavelength, the index of refraction remains constant up to this wavelength limit. There is thus no dispersion or change of index of refraction with wavelength, as in the case of the earlier conducting element delay lens when operated too near the resonant frequency of the elements. If enclosed in a full horn shield, such a lens should, over its entire operating range, possess an effective area which is independent of frequency. It would thus constitute an extremely broad-band antenna.

Since the index of refraction of the slanted plate lens depends only upon the tilt of the plates and not upon the plate spacing as in the first waveguide lens, constructional tolerances of this lens are even less severe than in the earlier lenses. Thus the number of plates could be doubled, if desired, without altering its performance in the original operating band. Also, flatness of the plates is not important, so long as the second mode spacing is not exceeded. The simplicity of design is brought out by observing that only 60 flat sheets cut to proper profile are required to duplicate the effectiveness of the several thousand elements of the 10-foot lens of Fig. 1.

There are two disadvantages which the slant plate lens possesses. The first is that the lens plates must be held at the proper design tilt with respect to the feed horn (since the index of refraction depends upon plate tilt). This prohibits a lens tilt in the vertical plane unless so designed, and limits the scanning ability of a moving feed in the vertical plane. The usual lens scanning capabilities are retained in the horizontal plane, however. The second is that the energy distribution in the electric plane is unsymmetrical, and results in a poorer minor lobe suppression in this plane. This latter disadvantage may not be serious in applications such as microwave repeater work, where lobes in the vertical plane are not as objectionable as lobes in the horizontal plane.

Acknowledgment

The writer wishes to acknowledge the assistance of W. E. Legg in the construction and testing of the lenses discussed in this paper.

Mercury Delay Line Memory Using a Pulse Rate of Several Megacycles*

ISAAC L. AUERBACH†, MEMBER, IRE, J. PRESPER ECKERT, JR.‡, ASSOCIATE, IRE, ROBERT F. SHAW‡, SENIOR MEMBER, IRE, AND C. BRADFORD SHEPPARD‡, MEMBER, IRE

Summary—A mercury delay line memory system for electronic computers, capable of operating at pulse repetition rates of several megacycles per second, has been developed. The high repetition rate results in a saving in space and a reduction in access time.

Numerous improvements in techniques have made the high repetition rate possible. The use of the pulse envelope system of representing data has effectively doubled the possible pulse rate; the use of crystal gating circuits has made possible the control of signals at high pulse rates; and a multichannel memory using a single pool of mercury has simplified the mechanical construction, reduced the size, and made temperature control much easier.

The memory system described makes possible a significant increase in the over-all speed of an electronic computer.

* Decimal classification: 621.375.2×R117.19. Original manuscript received by the Institute, December 15, 1948; revised manuscript received, March 4, 1949.
† Formerly, Eckert-Mauchly Computer Corp.; now, Burroughs Adding Machine Co., Philadelphia, Pa.
‡ Eckert-Mauchly Computer Corp., Philadelphia, Pa.

I. Introduction

THE MAJOR design problem in any high-speed electronic computer of advanced design is that of a memory, or storage device. Such a device must be capable of storing several hundred or more numbers of, say, twelve decimal digits each. Furthermore, any one of these numbers must be available for use in the computing circuits within a time comparable to the time required for an elementary arithmetic operation. Although considerable research is being done on electrostatic memories, in which information is stored in the form of charges on a dielectric medium, the delay-line type of memory is more highly developed at the present time, and is being used in several computers. In a delay-line memory, information is stored in the form of groups of electrical or acoustical impulses or signals circulating in an electric delay line or medium suitable for transmission of acoustic waves.

A device capable of storing *n* pulses may also be considered as storing *n* binary digits. The presence or the absence of a pulse in each of the *n* successive positions provides the coding necessary to represent binary numbers. To represent a decimal digit, a minimum of four successive coded pulse positions is required.

Acoustic propagation has a lower velocity than electromagnetic propagation, so a proportionately greater amount of information can be stored in the acoustic type delay line than in the electrical delay line for a given delay bandwidth and attenuation. Where large storage capacity is required, the acoustical line is preferred. Because of the low velocity of propagation and the good impedance match between it and quartz, mercury has been selected as the acoustic delay medium.

II. History

One of the first applications of acoustic delay lines was in the Scophony television system,[1] where video signals were converted into acoustical wave patterns in a transparent medium. These signals were read out optically by making use of the Debye-Sears effect[2] which translates the strain pattern into a pattern of varying light intensity.

This type of acoustic line was a nonregenerative or "delay storage" device. Further work on acoustic lines was done by Shockley at Bell Telephone Laboratories, who also used them as simple time delay devices. In 1943, Eckert, then at the Moore School of Electrical Engineering at the University of Pennsylvania, carried on further development work under a contract from the Radiation Laboratory of the Massachusetts Institute of Technology. Eckert was the first to use mercury as the acoustic medium and the first to use nonreflective backings on the quartz crystal transducers. By this technique it was possible to eliminate multiple storage due to reflections, and it was no longer necessary to use amplitude discrimination at the receiving end, as had been necessary in Shockley's lines. It also became possible to store patterns of pulses with relatively close spacing, increasing the amount of information per unit length, due to the greater bandwidth which the nonreflective terminations provided. Once these things had been accomplished, regeneration circuits were introduced to convert the delay device into a long-time dynamic storage device.

Work at the Moore School on delay lines had to stop, however, when work began on the ENIAC, the first large-scale all-electronic digital computer. Further development in the field took place at the Radiation Laboratories.[3-5] These subsequent studies included both liquid and solid media acoustic lines for use in range measuring circuits and MTI equipment.

During the final construction of the ENIAC, however, Eckert and Sheppard were able to return to the work on acoustic delay memories and eventually proposed this method for the EDVAC, a much improved electronic digital computer.[6]

The proper choice of a backing material for the quartz crystals is one of the most important design problems of the mercury delay line memory. It was during the time when work was being done on the EDVAC that Sheppard first used steatite as a crystal backing material. Until then the more practical choices were steel, air, and mercury. Steatite, it was found, gave a wider bandwidth than steel or air because there is a lower coefficient of reflection between it and quartz. The mercury-backed crystal had a higher capacitance than that with air backing, and further, this mounting subjected the crystals to breakage due to hydrostatic pressure of the mercury, and due to the method of mounting the thin crystals.

III. Theory

Delay line circulating memories have one common characteristic: an inevitable distortion of the circulating pulses. The pulses will be distorted in shape, and a timing shift will occur after each circulation. The shape distortion is due to attenuation, dispersion, phase distortion, and other less important practical considerations. If the shape distortion were not corrected, the pulse groups would lose their identity after a very few circulations. The timing shift arises through small errors produced in the temperature control system, changes in the length of the mercury column due to expansion and contraction, and the inaccuracy of construction which may give to different channels slightly different lengths. If the timing shift were not held within limits, it would be difficult to select a specific group of pulses from the several circulating in a single channel. Additional difficulties would arise in the arithmetic circuits, since timing synchronism is essential for proper operation. If the operation of the line is to be successful, both distortion and timing shift must be corrected.

[1] V. K. Zworykin and G. A. Morton, "Television," John Wiley and Sons, Inc., New York, N. Y., p. 254; 1940.
[2] See p. 251 of footnote reference 1.
[3] "A Theory of the Supersonic Delay Line," Rad. Lab. Report No. 733.
[4] "Multiple Reflection Delay Tank," Rad. Lab. Report No. 791.
[5] "On the Theory and Performance on Liquid Delay Lines," Rad. Lab. Report No. 792.
[6] "Progress Report on the Edvac," University of Pennsylvania, Moore School of Electrical Engineering, June 30, 1946.

The actual processes for correcting these variations may be thought of in two different ways. The terms "reshaping" and "retiming" imply that each pulse, as it emerges from the line, is sharpened to its original shape and shifted in time so that it coincides with some standard timing pulse. It is equally possible to think of the distortion correction process as being one in which each pulse, as it emerges from the line, is used to operate a gate circuit which, upon opening, allows a new standard timing pulse of the proper shape to enter the acoustic delay line. In this sense the old pulse containing some small error in time, and having been broadened by its passage through the memory channel and its associated amplifiers, is finally used to hold the gate open for the new timing pulse. Each recirculation therefore uses a new pulse gated into the line by the previous pulse. If no pulse is present in any given position, the gate circuit will suppress the standard timing signal.

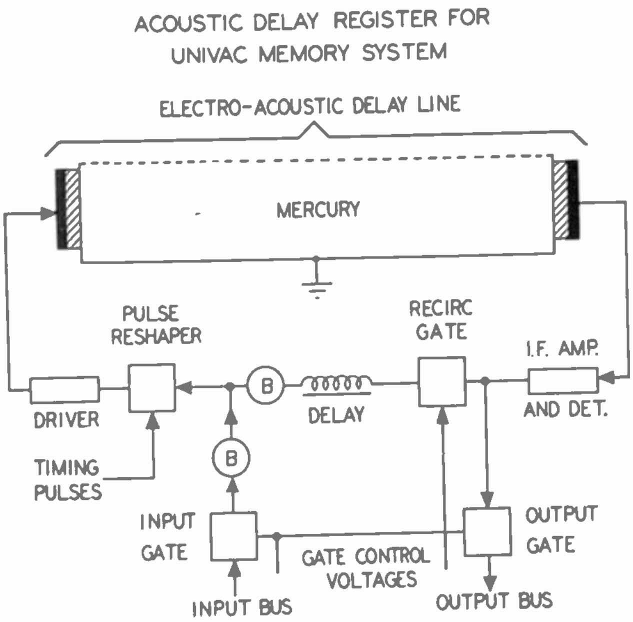

Fig. 1

Fig. 1 shows a block diagram of a typical memory channel. Pulses from the input bus enter the recirculation circuit through the input gate if an input gating signal is supplied. These pulses then gate timing pulses through the pulse reshaper or clock gate. The output of the clock gate feeds the driver, which is simply a power amplifier. Upon leaving the amplifier, the electrical pulse is applied to a quartz crystal transducer at one end of the mercury column, causing the transducer to produce an acoustic pulse in the mercury. The acoustic pulse then travels through the mercury with the velocity of sound in that medium. Because of the bandwidth limitation of the mercury tank, the acoustical pulse, after conversion into an electrical signal, has become a wave packet. After being amplified in the band-pass amplifier, the signal is rectified by the detector. The band-pass amplifier has a center frequency comparable to the crystal frequency and a bandwidth of several megacycles. The bandwidth limitation of the amplifier further broadens the wave packet. The output of the detector is a signal whose shape is that of the envelope of the wave packet. This signal, after passing through the recirculation gate, is used to gate a new timing pulse through the clock gate. The process is repeated indefinitely, unless the line is cleared by interrupting the recirculation path or a power failure occurs.

The two most important considerations of such a memory device as is here described, are the amount of information which it can store and the average length of time needed to remove or insert a given group of pulses. The waiting or latency time, often called the access time, directly affects the speed of operation of the computer. On the average, the access time will be one-half the recirculation time. It is as probable that a group of pulses will be desired just after they have entered the memory channel as it is that they will be desired just as they are about to emerge. In view of the access time consideration, several short channels are to be preferred over a few long ones.

The capacity of a delay line memory is determined by the length of the column and the repetition rate of the pulses circulating in the memory. Increasing either will increase the capacity, but it has already been pointed out that column length should be kept short in the interests of access time. Therefore, larger memory capacity should preferably be achieved through increased repetition rate. Considered in another way, a higher repetition rate will, for a given number of pulses per channel, reduce the access time because this number of pulses can be put in a shorter length of column. Any shortening of the access time makes it reasonable to increase the speed of performing arithmetic operations within the computer.

An important factor in determining the maximum repetition rate is the dispersion or phase distortion resulting largely from the effect of the walls of the tubes containing the mercury column. Such effects are greatly minimized by making the diameter of the column large compared to the crystal diameter and compared to the wavelength. Using crystal frequencies of about 15 Mc, this is not difficult, but still would result in a rather bulky memory assembly if the number of columns were large. It has been found, however, that a common pool of mercury can be used for a number of memory channels. The acoustical waves are so directional that each crystal affects only the one directly opposite it; and cross talk between channels can easily be held to two or three per cent with columns long enough to store over a thousand pulses per channel, particularly, if stainless steel tubes or "liners" such as those shown in Fig. 4 are used to isolate some channels from their neighbors.

The attenuation in the mercury has been found to be

$$\alpha = 0.0013f^2 + 0.9/fd^2,$$

where

α is the attenuation in db per hundred microseconds of delay

f is the frequency in megacycles

d is the column diameter in inches.

Having made the second term negligible by using a large column diameter, one is still faced with the problem of frequency discrimination. The bandwidth required depends only on the repetition rate of the pulses; by increasing the center frequency of the pass band, the variation of α over a given bandwidth can be decreased.

If effects of dispersion and attenuation are minimized, as described, the chief limitation on bandwidth is the electromechanical coupling characteristic of the crystals. The response is proportional to $\sin^2 \phi/\phi$, where ϕ is the ratio of impressed frequency to crystal frequency, and the distance between the half-power points $\Delta f = 0.66 f_c$, where f_c is the crystal frequency, assuming proper acoustic termination.[7] To find the required bandwidth for transmission of the wave packets, it will be assumed that the crystal frequency is made equal to about three times the maximum pulse repetition rate; it is then reasonable to use the Fourier transform of the envelope of the packet to obtain the frequency spectrum. Assuming an envelope having the shape of the probability curve, given by

$$\frac{e(t)}{e_{max}} = \epsilon^{-2.8(t/t_0)^2},$$

the envelope of the corresponding frequency spectrum is given by[8]

$$\frac{e(f)}{e_{max}} = \frac{1.1 t_0}{T} \epsilon^{-3.6(f-f_p)^2/f_r}$$

where t_0 is the width of the pulse packet at half voltage, T is the time between packets, and f_p is the frequency of the waves making up the packets. The distance between half-power points on the above frequency spectrum is $0.61 f_p$. The over-all bandwidth of the tank and amplifier should be a small amount greater than this to obtain the desired packet width at the detector, when the input of the tank is driven with a pulse having relatively rapid rise and slow fall.

Under the conditions just described, the detector output will drop to about 15 per cent of its maximum value between two pulses which are separated by a space, but will not drop at all (in fact, will rise slightly) between two adjacent pulses. Such a situation is somewhat unconventional in a pulse transmission system, but it will be recalled that the output pulses are not permitted to re-enter the line, but are only used to gate timing pulses. Therefore, if the latter are themselves sharp enough, it is only necessary for the output signal to be above or below the critical gating level by a safe amount in order to pass or reject a timing pulse. The gating signal can therefore, if necessary, retain its full amplitude for a number of pulse times if it represents an unbroken sequence of pulses, only dropping below the critical level when it is necessary to reject one or more pulses. Such a representation of pulse signals is known as a "pulse envelope" system, and can be shown to result in a reduction by a factor of two in the bandwidth requirements for transfer of a given amount of information per unit time.

IV. Construction

Further details of the mercury tanks are shown in Figs. 2, 3, and 4. Fig. 2 shows the construction of the

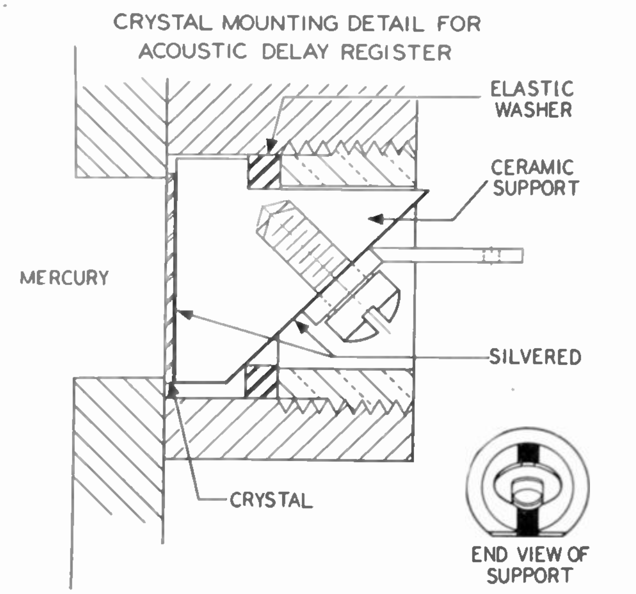

Fig. 2

crystal mounting. The ceramic support is silvered; one surface of the crystal is silvered, and the two are soldered together. The silvering is carried around the support to provide a connection to the back electrode; the mercury itself forms the front electrode. The support provides the proper acoustic termination for waves from the rear surface of the crystal. A threaded brass retaining ring and elastic washer hold the crystal support in the end of the mercury tank. Fig. 3 shows three crystals mounted on their steatite supports.

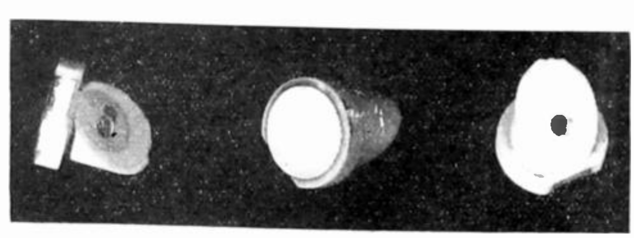

Fig. 3

Fig. 4 shows the various components of an 18-channel tank. The end cells into which the crystal mountings

[7] See sections, 2, 5, 3, and 5, respectively, of footnote references 3–5.

[8] G. A. Campbell and R. M. Foster, "Fourier Integrals for Practical Application," Bell System Monograph B-584; 1931.

fit are shown at bottom left and right, surrounded by mounted crystal, retaining bushings, and short coaxial connectors which are used between crystal element and recirculation chassis. The mounted crystals fit into the holes in the end cells and bear against retaining plates having holes slightly smaller than the crystals; these plates are shown just below the tank cylinder, with the liners used to reduce cross talk. Above the cylinder are tie rods which hold the assembly together, and at the right is one of the supports which center the assembled tank inside the outer jacket. The components at bottom center form an expansion chamber for the mercury. All metal parts in contact with mercury are made of a special stainless steel.

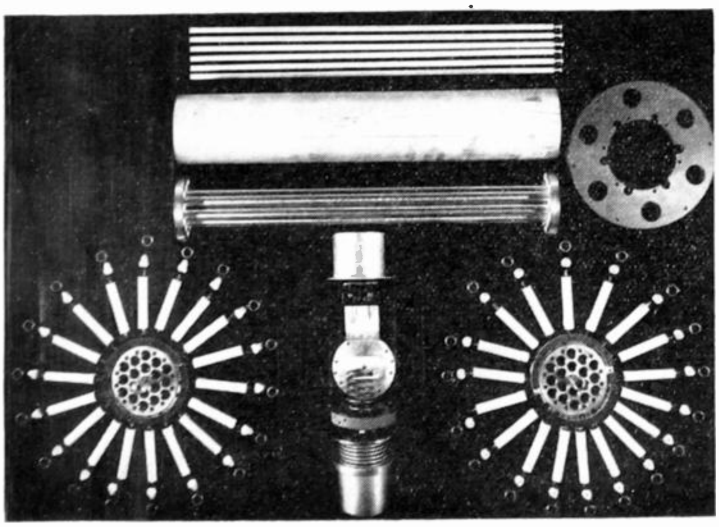

Fig. 4

The outer surface of the cylinder will be coated with insulating resin and wound with resistance wire, which serves as a heating element. The latter is part of a temperature control system which keeps the repetition rate of the pulses in synchronism with the clock or timing pulses. Here we note another advantage of the common mercury pool over a group of separate tanks: it is now easy to keep the temperature of all channels uniform. In the proposed designs of some previous computers, compensation for changes in mercury temperature was made by varying the clock frequency through a reactance tube. Now, because of the ease with which the mercury temperature can be controlled, it becomes possible to revise this procedure and use a crystal-controlled oscillator in the clock, which in turn controls the mercury temperature, and hence controls the circulation time of the tanks. This greatly simplified the problem of keeping all parts of the computer in synchronism, and in particular makes it easy to use a single clock to control two computers. The latter arrangement is valuable in cases where it is necessary to provide continuous checking of results by intercomparing the operation of the two computers.

Fig. 5 is a block diagram of the temperature control system. One channel in the multichannel tank is reserved as a temperature control channel. By means of a frequency divider, the timing pulse frequency is divided down to produce a series of pulses having a repetition rate corresponding to an integral submultiple of the total number of pulses stored in the tank. This submultiple is ordinarily made equal to the number of pulses in each number stored in a tank; for example, if each num-

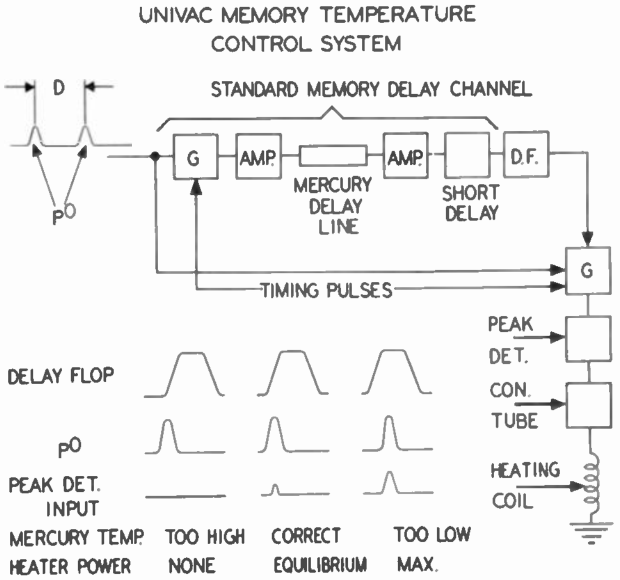

Fig. 5

ber requires 50 pulses to represent it, the basic pulse rate is divided by 50. Division is accomplished by means of an electric delay line with a single pulse circulating in it. This technique yields a large division ratio of precision equal to that obtained by a successive division, using a number of multivitiators or blocking oscillators in cascade. The pulses from the divider, which we may think of as corresponding to the last pulse in each number, are referred to as $p0$ pulses. They are supplied to the input gate of the temperature control channel and also, through the mercury line and a delay flop, to the output coincidence gate. Timing pulses are also applied to these gates to increase the accuracy of temperature control. This technique results in an over-all temperature control of $\pm 0.3°$ absolute.

If the capacity of a memory channel is, say, twenty 12-digit numbers, and the delay is accurately adjusted, then any $p0$ pulse entering the input will, after going through the line, coincide with the twentieth subsequent $p0$ pulse at the output gate. If the temperature is a little too high, however, the pulse through the line will arrive too late and fail to gate the twentieth following pulse. If the temperature is correct, the pulse from the line will have just reached half amplitude when the timing pulse, as gated by the twentieth following pulse, arrives, and the gate will produce an output pulse of intermediate amplitude. Finally, if the line is too cold, the pulse through the line will arrive too soon; ordinarily it would again fail to gate the twentieth following pulse, but if it is used to trigger a delay flop, the effect is to produce a pulse of maximum amplitude at the output of the gate. Thus the effect of the circuits shown is to produce an output signal whose amplitude is proportional to the amount of heat which should be supplied to the line to assure perfect synchronism. The gate output is fed to a peak detector to smooth out the pulse signal into a continuous signal which, after suitable amplification, controls the current to the heater coil around the tank.

A similar effect could have been achieved by using an average detector whose output is proportional to the amount of overlap of the $p0$ pulse and the signal from the line. Because of the low duty cycle, however, the detector output would be small and would require a considerable amount of amplification. The circuit used, giving an output whose amplitude varies in accordance with the relative position of a standard pulse and the relatively gradually sloping front of the signal from the line, in combination with the peak detector, gives a signal which requires little amplification.

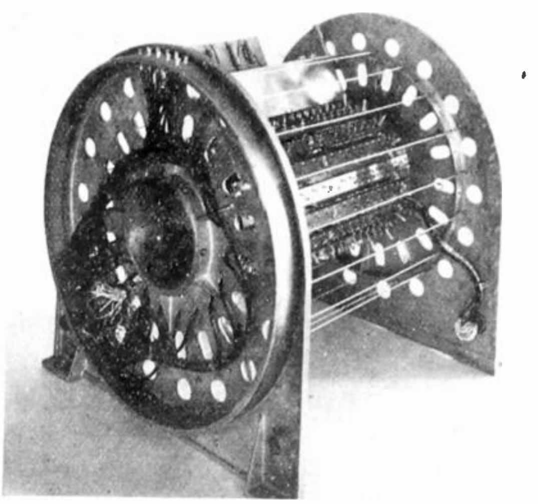

Fig. 6

Fig. 6 shows the outer cylinder mounted between its supporting end castings and surrounded by terminal boards to receive the plug-in recirculation chassis. The space between inner and outer cylinders is filled with santocel, a good thermal insulator. The radially disposed holes near the ends of the cylinder receive the short coaxial connectors shown in Fig. 4.

Fig. 7

In the plug-in recirculation chassis shown in Fig. 7, the four tubes at the left are the band-pass amplifiers, type 6AK5. These are followed by a detector using germanium crystals, and an AVC circuit. The latter feature is essential to compensate for changes in gain due to tube aging. A 50C5 amplifies the detector output signal and applies it to the output and recirculation gates; the recirculation and input gates feed another 50C5 which drives the clock gate. Timing pulses from the master oscillator are fed to the other input of this gate. Finally, the output of the clock gate is fed to a 50C5 driver which applies the reshaped and retimed signals to the crystal at the input end of the mercury column.

All gating is done by means of germanium diodes, which, because of their low forward impedance and small shunt capacity, are useful for gating operations at high pulse rates. The gating circuits are shown in Fig. 8.

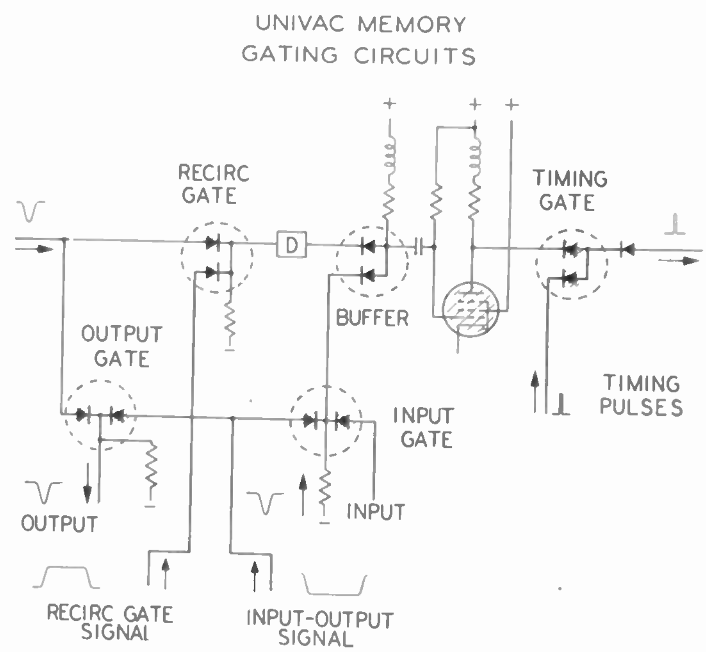

Fig. 8

Crystals which are normally conducting are shaded. The load resistors and the voltages to which they are connected are so chosen that the signal developed across the load resistor will not be sufficient to actuate the following circuit if only one of the crystals is conducting. However, it will actuate the following circuits if both crystals do not conduct. For example, an input signal will cut off current in the right-hand crystal of the input gate and cause somewhat less voltage drop across the load resistor, but the resulting signal will be of insufficient amplitude to cause the lower buffer crystal to conduct, unless a gate signal is simultaneously applied to the left-hand crystal of the input gate. The gates are not amplitude-sensitive in the usual sense, as input signals need only have sufficient amplitude to cut off their respective crystals.

Wave forms of the signals at various points in the system are shown in Fig. 9. The top curve shows the output of the clock gate as it is applied to the grid of the driver tube. Below is the voltage developed by the output crystal; it goes through about one and one-half oscillations as a result of bandwidth limitations in the tank.

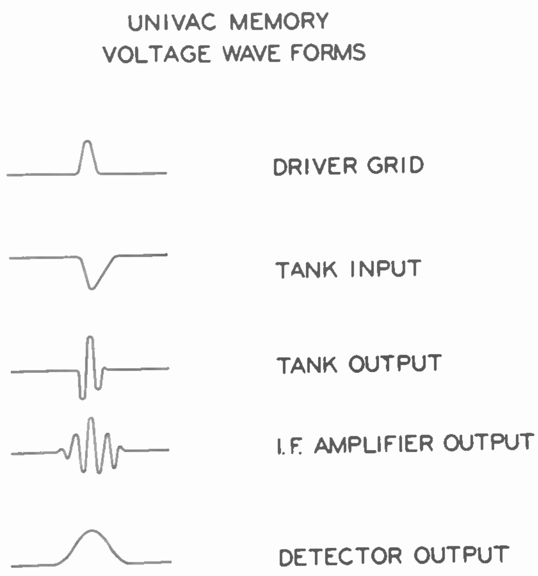

Fig. 9

The next curve is the output signal from the band-pass amplifier. In order to obtain adequate gain without too many stages, only sufficient bandwidth to pass a usable signal has been used; hence the amplifier introduces a noticeable amount of phase and amplitude distortion. The lower curve shows the detector output voltage; it is this signal which is amplified and used to gate a timing pulse, resulting in a signal of the form shown in the top curve in Fig. 9.

V. Conclusion

This paper has reviewed the results of developmental study of the basic acoustic delay line memory. As a result of this development, it was shown that it is possible to transmit and receive intelligence at the rate of 5,000,000 binary digits per second. This is a considerable extension over previous telegraphic keying rates and is thus of interest in its relation to information theory.

Of the various techniques employed to extend the useful frequency range of memory systems, the application of the pulse envelope system is probably the most significant in that it represents an important contribution to the concept of transmission of intelligence. Its importance lies in the fact that it provides a simple means of increasing the effective bandwidth of a transmission system without the necessity for a corresponding improvement in the characteristics of the components of the system. Another contribution was the application of germanium crystal gating circuits which permitted the effective control of signals at this high pulse repetition rate. The circuits developed in this work have found many other applications in the arithmetic and control circuits of computers.

The use of a single pool of mercury for multiple channels of information reduced the physical size of the whole memory system. As a result of the intimacy of metal contact between the channels, temperature gradients were greatly reduced, the temperature control problem was simplified, and synchronous operation of two computers for checking purposes facilitated.

The acoustic memory described represents one of the first practical applications of techniques which have extended the useful audio transmission range from its previous limits of some 150,000 cps to a new present value of 30 million cps or more.

This paper has reviewed the several improvements that have been made which helped to extend the useful acoustic transmission range.

CORRECTION

A drafting error in the paper, "A Digital Computer for Scientific Applications," by C. F. West and J. E. DeTurk, which appeared on pages 1452–1460 in the December, 1948, issue of the Proceedings of the I.R.E., has been brought to the attention of the editors by the authors. The error was made in Fig. 10, on page 1457, in which the signals feeding through 1N34 crystals to the grids of the indicated flip-flops should be transposed.

An Analysis of Magnetic Amplifiers with Feedback*

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D. C. BEAUMARIAGE†, STUDENT MEMBER, IRE

Summary—Two ways of obtaining feedback in magnetic amplifiers are studied. One, termed external feedback, employs a bridge rectifier and separate feedback coils, while the other, termed self feedback, uses two rectifiers so disposed that separate feedback windings are unnecessary. Commutation of the rectifiers, apparently not considered heretofore, is shown to be of major importance. The calculated currents are confirmed satisfactorily, both in shape and magnitude, by experimental checks. The two feedback methods are compared, and it is found that self feedback is advantageous.

Introduction

A SIMPLE magnetic amplifier consists of two identical single-phase transformers having one pair of similar windings in series with an ac source and a load, and the other pair in series with a dc source. A small direct current in the latter windings, through saturation of the cores, determines the inductive reactance presented on the ac side and so controls the load power. This and similar types of magnetic amplifiers are treated in a companion paper.[1] A more complex type achieves greater sensitivity through positive feedback or self-excitation schemes well known in practice.[2–6] The purpose here is to treat these feedback circuits more precisely than has been done before.

One way of obtaining feedback is shown in Fig. 5, which is derived from Fig. 1 of the companion paper[1] by adding a third set of windings N_3 to each core and connecting them to a bridge rectifier in series with the load. The mmf's produced in the feedback windings N_3 reinforce those produced in the control windings N_1, so that a given average saturation is achieved with less control current than would be needed without feedback. This arrangement is identified here by the term "external feedback."

A less obvious way of obtaining feedback is shown in Fig. 1, which is identical with Fig. 4 of the companion paper,[1] except for the addition of a rectifier in series with each of the two load coils. During one half cycle the load current links one core and during the next half cycle the other, thus producing direct components of mmf which aid those produced by the control current, and so gives feedback without additional windings. The term "self feedback" is used for this circuit.

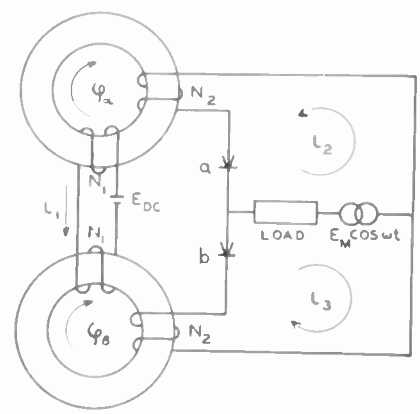

Fig. 1—Magnetic amplifier employing self-feedback.

The purpose here is to calculate the currents as functions of time under steady-state conditions for the self- and external-feedback circuits, and then to compare the two arrangements.

Method of Analysis

The analysis is similar to that in the companion paper.[1] The simplifying assumptions are that:

(a) The load is replaced by a short circuit, so that a sinusoidal voltage is impressed directly on the amplifier terminals;
(b) The control circuit is connected to a constant dc source having no impedance;
(c) Steady-state conditions exist;
(d) Eddy currents and hysteresis are negligible, and the B–H relationship is given by the normal magnetization curve of the core material;
(e) Leakage fluxes are negligible, and the flux density is uniform throughout each core;
(f) Coil resistances are very small, although not neglected entirely; and
(g) Rectifier reverse resistances are infinite, and forward resistances are very small and constant.

Briefly, the procedure is to apply Kirchhoff's voltage law to the electric network and solve for the magnetic fluxes in the cores; use the magnetization curve of the material to find the corresponding mmf's; apply Ampere's law to each core to relate mmf's and currents;

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† Carnegie Institute of Technology, Pittsburgh, Pa.
†† Formerly, Carnegie Institute of Technology.

[1] D. W. VerPlanck and M. Fishman, "An analysis of interlinked electric and magnetic networks with application to magnetic amplifiers," scheduled to appear in an early issue of the Proc. I.R.E.
[2] E. C. Wentz, "Direct-current controlled reactor," *Electric Jour.*, vol. 28, p. 561; October, 1931.
[3] T. Buchhold, "Über gleichstromvormagnetisierte Wechselstromdrosselspulen und deren Rückkopplung," *Arch. fur Elek.*, vol. 36, p. 221; month, 1942.
[4] T. Buchhold, "Zur Theorie des magnetischen Verstärkers," *Arch. fur Elek.*, vol. 37, p. 197; April, 1943.
[5] H. S. Kirschbaum and E. L. Harder, "A balanced amplifier using biased saturable core reactors," *Trans. AIEE, (Elec. Eng.)* vol. 66, pp. 273–278; 1947.
[6] A. Uno Lamm, "Some fundamentals of a theory of the transductor or magnetic amplifier," *Trans. AIEE, (Elec. Eng.)* vol. 66, pp. 1078–1085; 1947.

and, finally, solve for the currents. In doing this, it is necessary to account for commutation periods in which all rectifiers conduct simultaneously.

As it is the simpler of the two arrangements to be studied, the self-feedback circuit is treated first.

Self-Feedback Circuit

Referring to Fig. 1, and replacing the load by a short circuit, Kirchhoff's voltage law gives three equations represented in matrix form as

$$\begin{Vmatrix} E_{dc} \\ E_m \cos \omega t \\ E_m \cos \omega t \end{Vmatrix} = \begin{Vmatrix} N_1 & N_1 \\ N_2 & 0 \\ 0 & -N_2 \end{Vmatrix} \cdot \begin{Vmatrix} \dot{\phi}_\alpha \\ \dot{\phi}_\beta \end{Vmatrix} + \begin{Vmatrix} 2R_1 & 0 & 0 \\ 0 & R_2 + R_r & 0 \\ 0 & 0 & R_2 + R_r \end{Vmatrix} \cdot \begin{Vmatrix} i_1 \\ i_2 \\ i_3 \end{Vmatrix}. \qquad (1)$$

Mks units are used and the symbols are as shown in Fig. 1, with the addition that R_1 and R_2 are, respectively, the resistances of each of the single windings N_1 and N_2, and R_r is the rectifier resistance equal either to a small and constant value or infinity, depending on whether or not the rectifier is conducting.

When rectifier a alone is conducting, the ***positive period***, only the first two equations of (1) are useful. Similarly, in the ***negative period***, when b alone conducts, only the first and last of (1) apply. All three are useful in the *commutation period* when both a and b conduct.

The fluxes are found by simultaneous solution of the appropriate voltage equations, assuming that the resistance drops (and hence E_{dc}) are negligible. Flux densities B_α and B_β in the respective cores are then found by dividing the fluxes by the core area A. Each period must be treated separately but in each case the result is the same, namely

$$B_\alpha = B_m \sin \omega t + B_0 \qquad (2)$$

$$B_\beta = -B_m \sin \omega t + B_0 \qquad (3)$$

where $B_m = E_m/N_2\omega A$. B_0, the average flux density in each core, to be determined, is the same throughout the cycle because the linkage of the cores by closed circuits of finite resistance prevents sudden changes of flux at the transitions between periods; also, considering symmetry, it is clear that B_0 is the same for each core. For the commutation period, only two of the three voltage equations are needed to determine the flux densities; the third equation is useful later.

The field intensities are derived from the flux densities through graphical use of the magnetization curve, as in Fig. 5 of the companion paper.[1] In functional notation, the relationship is

$$H_\alpha = H(B_m \sin \omega t + B_0) \qquad (4)$$

$$H_\beta = H(-B_m \sin \omega t + B_0). \qquad (5)$$

The required mmf's are, respectively, LH_α and LH_β where L is the mean length of core. Then, applying Ampere's law, two equations result which in matrix form are:

$$\begin{Vmatrix} N_1 & N_2 & 0 \\ N_1 & 0 & -N_2 \end{Vmatrix} \cdot \begin{Vmatrix} i_1 \\ i_2 \\ i_3 \end{Vmatrix} = \begin{Vmatrix} LH_\alpha \\ LH_\beta \end{Vmatrix}. \qquad (6)$$

In the positive and negative periods, i_3 and i_2 are respectively equal to zero, and the remaining two currents are found directly as shown in Table I.

TABLE I
EXPRESSIONS FOR CURRENTS IN AMPLIFIER WITH SELF-FEEDBACK

Period	Currents
Positive	$i_1 = LH_\beta/N_1$ $i_2 = L(H_\alpha - H_\beta)/N_2$ $i_3 = 0$ $i_2 + i_3 = L(H_\alpha - H_\beta)/N_2$
Commutation	$i_1 = [k'L(H_\alpha + H_\beta)/2 + N_1I_1]/N_1(1+k')$ $i_2 = [k'L(H_\alpha - H_\beta)/2 + LH_\alpha - N_1I_1]/N_2(1+k')$ $i_3 = [k'L(H_\alpha - H_\beta)/2 - LH_\beta + N_1I_1]/N_2(1+k')$ $i_2 + i_3 = L(H_\alpha - H_\beta)/N_2$
Negative	$i_1 = LH_\alpha/N_1$ $i_2 = 0$ $i_3 = L(H_\alpha - H_\beta)/N_2$ $i_2 + i_3 = L(H_\alpha - H_\beta)/N_2$

In the commutation period all three currents have values, and so an additional equation is needed. This is obtained by eliminating the rates of change of flux from (1) but without neglecting the resistance drops. This equation and (6) are solved simultaneously, with the results shown in Table I. In these expressions,

$$I_1 = \frac{E_{dc}}{2R_1} = \frac{1}{2\pi}\int_0^{2\pi} i_1 d(\omega t), \qquad (7)$$

the average (dc) value of the control current, and

$$k' = \left(\frac{N_1}{N_2}\right)^2 \frac{R_2 + R_r}{R_1}. \qquad (8)$$

Thus, as in circuits with parallel-connected load coils,[1] resistances determine the currents when the mmf can be supplied in part by currents in either of two paths, even though the resistances may be negligible in other respects. Here the two paths are the control circuit and that through the two load coils and two rectifiers in series in the forward direction.

Although expressions for the currents have been derived for all periods of the cycle, the average flux density B_0, which enters all of the expressions, and the times of transition between periods have not been found. To determine these unknowns requires a graphical trial and error process, illustrated in Fig. 2, and consisting of the following steps:

(a) A value of B_0 is assumed.

(b) The centers of the commutation periods are located by plotting $N_2(i_2+i_3)$, which, as is indicated in Table I, is independent of transition times.

(c) As a first trial, commutation is assumed instantaneous and N_1i_1 is plotted using the expressions in Table I for the positive and negative periods (positive and negative referring to the sign of i_2+i_3) and continuing the plots to their intersections as shown by (1) in Fig. 2.

(d) The average control excitation N_1I_1, defined by (7), is found from the area under N_1i_1 as plotted in (c).

(e) The value of N_1I_1 from (d) is used to obtain a second approximation to N_1i_1 in the commutation period, (2) in Fig. 2.

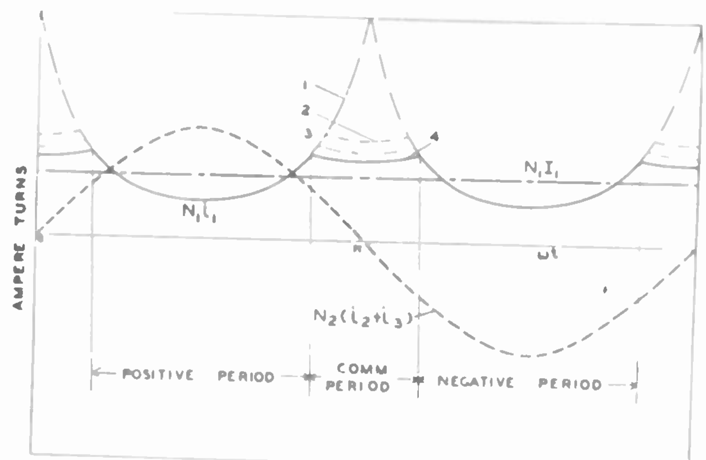

Fig. 2—Successive approximations to N_1i_1 and the limits of the commutation period. The first three approximations to N_1i_1 and the correct value are shown by 1, 2, 3 and 4, respectively.

$N_1 = 500$ turns
$N_2 = 4000$ turns
$E_m = 140\sqrt{2}$ volts

$\omega = 2\pi 60$ sec^{-1}
$L = 0.18$ meters
$A = 4.2 \times 10^{-4}$ (meters)^2

$R_1 = 35$ ohms
$R_2 = 190$ ohms
$R_r = 50$ ohms

This leads to a new value of N_1I_1, which is used to correct N_1i_1, and so on. The process converges rapidly, the fourth approximation to N_1I_1 in Fig. 2 being essentially correct. Thus the transition times are located, and the currents are completely determined for the assumed B_0. If the problem must be solved for a specified I_1, a succession of values of B_0 are tried until the one corresponding to the desired I_1 is found.

Experimental checks are given in Figs. 3 and 4, which are for the circuit parameters of Fig. 2 and for the magnetization curve shown in the companion paper.[1] Fig. 3 compares calculated currents (plotted as ampere-turns) with points measured from oscillograms for the same conditions as Fig. 2. Fig. 4 compares calculated and measured external characteristics over a range of N_1I_1 with other quantities fixed at the values of Figs. 2 and 3. The load current i_2+i_3 was measured with a rectifier-type ammeter, and calculated from areas under curves of instantaneous currents like that in Fig. 3.

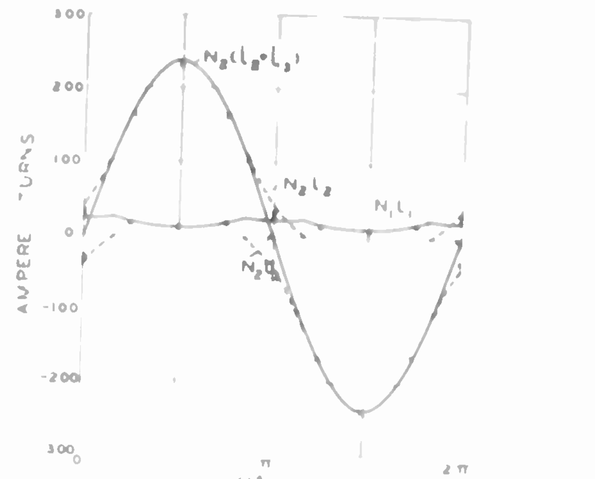

Fig. 3—Calculated instantaneous ampere turns for the circuit of Fig. 1 compared with points measured from oscillograms. Circuit quantities as in Fig. 2 and $N_1I_1 = 17$.

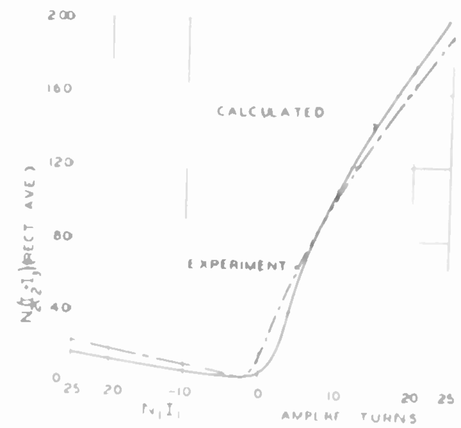

Fig. 4—Comparison of calculated and experimentally determined characteristic curves for the self feedback circuit, Fig. 1. Circuit quantities as given for Fig. 2 except that N_1I_1 is varied.

External Feedback Circuit

In the circuit with external feedback, Fig. 5, again one must consider three periods: the positive period,

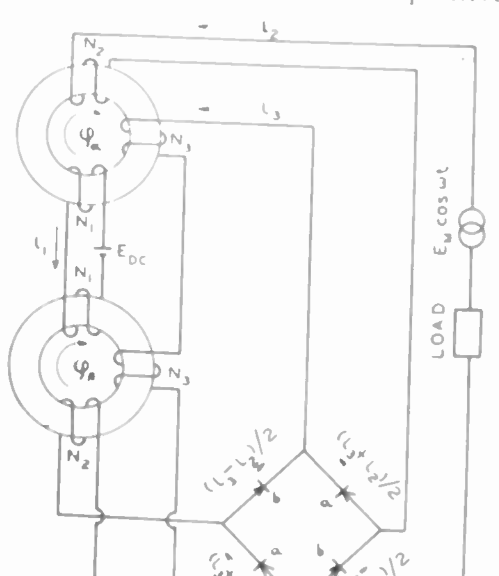

Fig. 5—Magnetic amplifier employing external feedback.

when rectifiers aa alone conduct; the negative, when bb conduct; and commutation, when all four conduct.

For the control circuit one voltage equation applies throughout, namely,

$$E_{dc} = N_1(\dot{\phi}_\alpha + \dot{\phi}_\beta) + 2R_1 i_1. \tag{9}$$

During the positive period, $i_3 = i_2$, and the voltage equation for the load circuit is

$$E_m \cos \omega t = (N_3 + N_2)\dot{\phi}_\alpha + (N_3 - N_2)\dot{\phi}^\beta + 2(R_2 + R_3 + R_r)i_2. \tag{10}$$

In the negative period, $i_3 = -i_2$, and the voltage equation may be obtained from (10) by changing N_3 to $-N_3$.

For the commutation period, (10) is replaced by two voltage equations: one for the loop through the generator, the coils N_2 and either pair of rectifiers ab,

$$E_m \cos \omega t = N_2(\dot{\phi}_\alpha - \dot{\phi}_\beta) + (2R_2 + R_r)i_2, \tag{11}$$

and the other for the loop through the coils N_3 and a pair of rectifiers ab,

$$0 = N_3(\dot{\phi}_\alpha + \dot{\phi}_\beta) + (2R_3 + R_r)i_3. \tag{12}$$

Simultaneous solution of the appropriate voltage equations for each of the periods results in expressions for the flux densities B_α and B_β in the respective cores identical with (2) and (3) for the self-feedback case, except that here $B_m = E_m/2N_2\omega A$. As in the previous case the field intensities are given by (4) and (5).

The mmf conditions in matrix form are:

$$\begin{Vmatrix} N_1 & N_2 & N_3 \\ N_1 & -N_2 & N_3 \end{Vmatrix} \cdot \begin{Vmatrix} i_1 \\ i_2 \\ i_3 \end{Vmatrix} = \begin{Vmatrix} LH_\alpha \\ LH_\beta \end{Vmatrix}. \tag{13}$$

For the positive and negative periods, i_3 is equal respectively to $+i_2$ and $-i_2$, and (13) may be solved for the currents. For the commutation period, it is necessary to get another equation from the voltage conditions, as in the self-feedback case. Doing this, the currents in all three periods are found readily, as shown in Table II. Here I_1 is defined exactly as in (7), and

$$k'' = \left(\frac{N_1}{N_3}\right)^2 \frac{2R_3 + R_r}{2R_1}. \tag{14}$$

TABLE II

EXPRESSIONS FOR CURRENTS IN AMPLIFIER WITH EXTERNAL FEEDBACK

Period	Currents
Positive	$i_1 = L[(1-N_3/N_2)H_\alpha + (1+N_3/N_2)H_\beta]/2N_1$ $i_2 = L[H_\alpha - H_\beta]/2N_2$ $i_3 = i_2$
Commutation	$i_1 = [k''L(H_\alpha + H_\beta)/2 + N_1 I_1]/N_1(1+k'')$ $i_2 = L[H_\alpha - H_\beta]/2N_2$ $i_3 = [L(H_\alpha + H_\beta)/2 - N_1 I_1]/N_3(1+k'')$
Negative	$i_1 = L[(1+N_3/N_2)H_\alpha + (1-N_3/N_2)H_\beta]/2N_1$ $i_2 = L[H_\alpha - H_\beta]/2N_2$ $i_3 = -i_2$

The limits of the commutation period and the relation between B_0 and I_1 are found by trial, as explained for the self-feedback case.

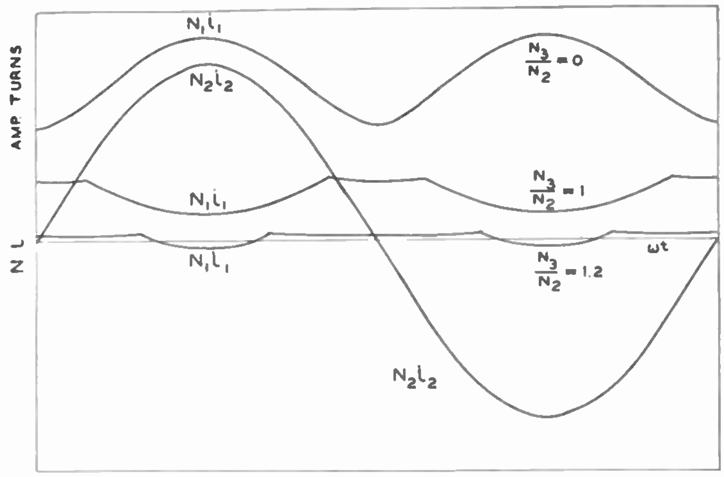

Fig. 6—Calculated excitation curves for the circuit of Fig. 5, showing the effect of varying the number of feedback turns while keeping the total amount of copper in feedback and control windings constant.

Fig. 6 and Table III show an application of the foregoing analysis of the magnetic amplifier with external feedback. The purpose is to show how varying the feedback turns affects the control excitation and power. The load windings N_2, the magnetic quantities B_m and B_0, the ac voltage, and the load current remain the same throughout. That is, conditions on the ac side are identical for the three cases shown. The total amount of copper in the control and feedback windings together is kept fixed by holding the quantity $(N_1^2/R_1 + N_3^2/R_3)$ constant, but the distribution of copper between these windings is varied. Forward resistances of the rectifiers are neglected in these calculations. The last column of Table III shows how greatly the power which must be supplied by the control source can be reduced in this particular case by using feedback. These results, however, must be viewed with some caution, because another effect of high feedback may be to cause the amplifier to become unstable. The present analysis is also useful in the study of stability.

TABLE III

EFFECT OF FEEDBACK TURN RATIO ON CONTROL POWER REQUIRED FOR A PARTICULAR CASE

N_3/N_2	N_1^2/R_1	N_3^2/R_3	$N_1 I_1$	$I_1^2 R_1 = (N_1 I_1)^2/(N_1^2/R_1)$
0	18,400	0	85	0.39
1.0	3,400	15,000	25	0.18
1.2	400	18,000	1.4	0.0049

COMPARISON OF SELF- AND EXTERNAL FEEDBACK

In comparing the self- and external-feedback circuits, Figs. 1 and 5, one must note that the latter is more flexible in that there may be various amounts of feedback, depending on the number of turns N_3. The comparison, therefore, is made on the basis that the feedback and load coils in the external-feedback circuit are equal; that is,

$$N_3 = N_2 \tag{15}$$

$$R_3 = R_2. \tag{16}$$

With this condition, Tables I and II show that the expressions for control currents in the two cases and for the load currents (i_2 and i_2+i_3, respectively) become identical if

$$k' = k'' \tag{17}$$

$$N_1' = N_1'' \tag{18}$$

$$N_2' = 2N_2'' \tag{19}$$

where single primes are used for self-feedback quantities and double primes for external feedback. The currents actually will be identical if L, H_α, and H_β are the same in the two cases. This identity will exist if core dimensions and material are identical, and impressed ac voltages and average values of control current I_1 are the same.

Having established these conditions for identity of magnetic conditions and of external currents in the two cases, it follows that

$$R_2' + R_r' = 2(2R_2'' + R_r''), \tag{20}$$

provided that

$$R_1' = R_1''. \tag{21}$$

With these relations between resistances and the other conditions of equivalence, it is found, using the expressions for the currents in Tables I and II, that the i^2R losses in the control windings are the same at every instant, and similarly in the load and feedback windings and the rectifiers taken together. Thus the performance of the two arrangements will be identical in all respects under the conditions set. Examination of the relations between resistances and numbers of turns shows, however, that external feedback requires twice as much copper in the load and feedback coils together as is required in the load coils of the self-feedback circuit.

Thus, for equivalent performance, the self-feedback arrangement, Fig. 1, can be made smaller than that employing external feedback, Fig. 5, and is therefore superior provided that the feedback turn ratio N_3/N_2 desired is unity. For greater feedback turn ratio, a combination of the arrangements is indicated;[6] rectifiers being provided as in Fig. 1 to give most of the feedback effect, and booster windings like N_3 in Fig. 5 being added to supply additional feedback as desired.

Conclusion

Two well-known methods for obtaining feedback in magnetic amplifiers have been analyzed quantitatively, with results that are checked closely by experiment. The analysis has been used to show quantitatively how feedback reduces the control excitation requirements and to compare the two feedback arrangements.

The analysis takes account of commutation, or simultaneous conduction of the rectifiers, a phenomenon which does not seem to have been considered so far, although it may have been observed.[3]

Electrical Network Analyzers for the Solution of Electromagnetic Field Problems*

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Part II
OPERATION

Summary—**This is the second part of a paper on the design, construction, testing, and operation of electrical networks capable of obtaining solutions of the wave equation in both two-dimensional, axially symmetric, cylindrical co-ordinates and rectangular co-ordinates. The circuit yields information on *TEM*, *TM*$_{om}(n)$ and *TE*$_{om}(n)$ modes of concentric lines, waveguides, and resonators. Field plots of electric- and magnetic-field configurations are directly available, and resonant frequencies of cavities, equivalent impedance, reactance characteristics, *Q*, and propagation characteristics can be determined.**

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Introduction

In Part I, the general theory, design, and construction of two network analyzers for solution of the wave equation in both two-dimensional, axial symmetric, cylindrical co-ordinates and rectangular co-ordinates were discussed. This section will deal only with the testing and general use of the networks. These network analyzers have been used at Stanford University for about two years, and have been found to be very helpful in the general solutions of field problems. The applicability of the network has been well established by comparing numerous results with known cases. The various boundary conditions that must be met and their apparent position in respect to the network are discussed. Techniques for testing and solving special problems, such as resonator characteristics, obtaining of field plots, and the measurement of shunt resistances and Q, are presented.

IV. Network Adjustment

A. Tuning the Cylindrical-Co-ordinate Network

The only tuning involved on the network is that of adjusting the trimmer capacitors. The inductances can be assumed to be correct from their original calibration. In tuning the capacitor, it was found that the most accurate method of adjusting the capacitors was a resonance-substitution method applied to each capacitor individually. Since the capacitance values vary with radius, this involves a number of different adjustments consuming considerable time. A careful adjustment of all the network capacitance values requires approximately two man days of work.

B. Tuning the Rectangular-Co-ordinate Network

Similarly with this network, the only adjustment is that of the capacitance values. The uniformity of his network makes these operations relatively simple.

Methods were developed for checking both continuity in coils and adjusting capacitor values which did not require that any coils be unplugged from their operating positions in the clips.

Adjustment of capacitance values was facilitated by the fact that the coils uniformly had a natural resonant frequency of 2 Mc. At this frequency, they exhibited a very high resistance compared to the proper value of capacitive reactance. Accordingly, a special coil was designed to be clipped across the capacitors and to be resonated with the proper value of capacitance at 2 Mc. Resonance was indicated with a Measurements Corporation Megacycle Meter. By this method, it was possible to adjust all the capacitances to a high degree of accuracy in approximately one-half man day.

V. Use of the Networks

A. Use of the Cylindrical Network

1. Determination of Cavity Resonant Frequencies. The determination of the resonant frequencies of cavity resonators is relatively simple. A scaled outline of the cavity is set up on the network by opening the circuit at the boundary corresponding to the cavity walls for TM_0 modes and shorting capacitors to ground for TE_0 modes. The network is then excited from a variable-frequency oscillator, and the frequency of oscillation varied until a resonance is obtained. Excitation of the network is achieved by simply connecting the ground plane and an arbitrary point of the network lattice to the oscillator through a high series resistance (100,000 ohms) simulating very loose coupling. Coupling may also be had inductively by a coupling loop in the vicinity of one of the network coils. It should be borne in mind that, for a given boundary condition, i.e., either open or short circuit there will be revealed resonances of the corresponding mode type only; i.e., TM_0 and TE_0, respectively. It should also be borne in mind that the corresponding edge of the field is different for the two types of boundary conditions. When the conductor is represented by an open circuit, the effective field boundary is very nearly one-half section beyond the capacitor at which the network is opened. For the short-circuit type of boundary, however, the effective edge of the field is at the short. These end corrections are the same by test for either radial or longitudinal boundaries with the cylindrical-co-ordinate network. A similar end correction applies to the rectangular-co-ordinate network.

2. Resonator Tuning Curves. Resonator tuning curves as one dimension of a cavity is varied are obtained by the method related in the previous section. It is only necessary to record and plot resonant frequency as a function of the corresponding cavity dimension. A curve of resonant wavelength versus cavity length for a concentric-line resonator is shown in Fig. 6. This figure exhibits several groups of curves for different modes and different ratios of radii in the concentric-line portion. The curves have been extended to allow the plunger to protrude into the end space of the resonator.

3. Cylindrical Cavity Field Plots. For cylindrical TEM and TM_0 modes, the modes of usual interest, complete magnetic- and electric-field plots are obtained from the junction voltage-to-ground data at network resonance. Since voltage to ground corresponds to magnetic field

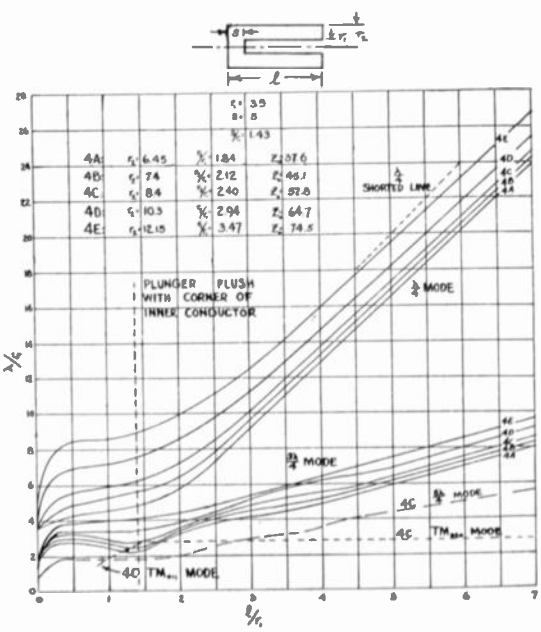

Fig. 6—Resonant wavelength of a loaded concentric-line resonator as a function of length.

multiplied by radius, it is only necessary to divide voltage values by the radius to get the relative magnetic field.

A type of electric-field plot is given directly by contours of constant network voltage to ground. Such contours have the direction of electric flux lines. Flux density is, however, given by line density of the voltage contours divided by radius. While this has not been a common convention for plotting electric flux fields, it should be pointed out that it is the only correct convention for cylindrical geometries which permits the use of continuous lines when such lines are not all parallel to the axis. In working with the network it has been found desirable to use the above convention exclusively because, in the first place, it is the only consistent convention which can be used for general field configurations, and, in the second place, such field plots are given directly by contours of constant voltage on the network.

The reason for the above relations is seen from equations (7) and (8) in Part I. Here it is seen that the electric flux density multiplied by radius is a vector whose magnitude is proportional to the gradient of rH_ϕ (hence of network voltage to ground), and whose direction is normal to the gradient of rH_ϕ (or network voltage). Since the direction of the electric flux is normal to the gradient of rH_ϕ (network voltage), which, in turn, is normal to the contours rH_ϕ (network voltage), then the electric flux has the direction of contours of constant rH_ϕ (network voltage). Since the gradient of rH_ϕ (network voltage) is proportional to electric field multiplied by radius, then the electric flux density is given by the density of lines of constant rH_ϕ (or network voltage) divided by the radius.

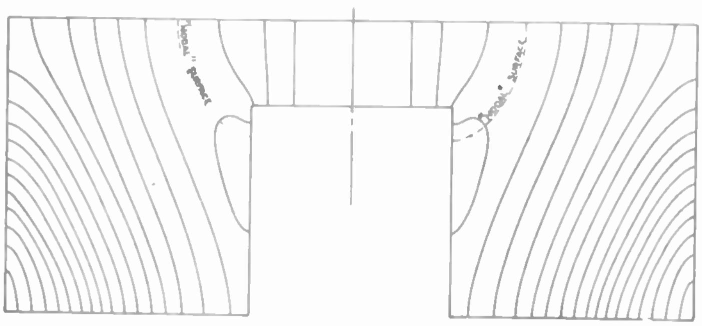

Fig. 7—Electric-field plot for a re-entrant cavity resonator operating on a three-quarter wave mode.

Shown in Fig. 7 is an electric-field plot of a three-quarter-wave resonance in a concentric-line resonator, drawn according to the convention given above. An interesting feature of higher-order modes, not ordinarily appreciated but revealed here, is the fact that, whereas with higher-order TEM and TM_0 modes there is always a nodal surface of magnetic field, there need not be a true nodal surface of electric field. There will, however, always be a pseudo-nodal surface of electric field which can be drawn through a point of zero electric field, which will always exist in such a way that no power is transmitted across this surface and such that, if the surface were replaced by a conductor, the two resonators thus created would have field configurations like portions of the orginal and would resonate at the same frequency as the entire cavity. The "nodal" surface of electric field having these properties is shown by a dashed line in Fig. 7.

4. Determination of Shunt Resistance and Q of Cylindrical Cavity Resonators. The cylindrical-co-ordinate network can be used to measure the shunt resistance and Q of cavity resonators directly. The details of the method of measurement and a discussion of some results obtained by its application will be given in a subsequent paper.[7] In this paper there will only be given a statement of the principles involved.

Since the network is the electrical analogue of the field problem, it is expected that its Q and input resistance should be related to those of a cavity resonator it represents. Normally, the network will correspond to a cavity with a nonuniform lossy dielectric, the series resistance of the inductances corresponding to the dielectric loss. To represent wall losses, it is necessary to shunt the terminal capacitances of the network with resistances which are large compared to the capacitor reactance and which increase linearly with radius.

In making measurements upon a network loaded with terminal resistors to represent wall losses, it is necessary to separate the effects of the power losses in the terminal resistors and in the coils themselves. For Q measurements, the Q of the network must be measured with and without the terminal resistors, thus determining the Q resulting from the terminal resistors alone. This value will be proportional to the cavity-resonator Q. The appropriate proportionality factor may be measured or deduced theoretically.

In making shunt-resistance measurements, use can be made of the fact that for TEM and TM_0 modes, the network quantities are the inverse of the field quantities; thus the electric field is represented by a network current. Accordingly, it is expected that the equivalent circuit of the network, as measured at the cavity gap, should be the inverse of the cavity-resonator equivalent circuit. Since the equivalent circuit of a cavity is given closely by a parallel combination of shunt resistance, lossless inductance, and capacitance, it is expected that the equivalent circuit of the network will be a series combination of resistance, inductance, and capacitance. This series impedance is measured by opening the network at the point where the radial gap currents flow into the axis circuit. It is further expected that the conductance of the series network equivalent will be proportional to the shunt resistance of the cavity. Here, again, it is necessary to separate components of shunt resistance due to wall losses and dielectric losses. This is done by measuring the network series resistance with

[7] F. W. Schott and K. R. Spangenberg, "A network analogue approach to the study of cavity resonator losses," presented, 1948 West Coast IRE Convention, Los Angeles, Calif., October 1, 1948.

and without the terminal resistors. The conductance of the difference of the two series resistances thus measured is proportional to the shunt resistance of the cavity.

B. Use of the Rectangular Network

1. Determination of Resonant Frequencies of Cavities. The method used here involves excitation and variation of frequency until electrical resonance is observed by the same method used on the cylindrical-co-ordinate network. Attention must again be given to proper boundaries and end corrections. As with the other network, the effective field edge is half a section beyond the capacitor at which the network is opened with open-circuit boundaries, and at the short with short-circuit boundaries.

2. Tuning Curves. These are obtained by the same method as is used with the cylindrical-co-ordinate network.

3. Field Plots. These can usually be obtained directly from network voltage measurements. In the case of the modes referred to here as *TM* modes, network voltage to ground is directly proportional to electric field strength. A plot of contours of constant voltage to ground constitutes a magnetic flux plot. With *TE* modes, voltage to ground gives strength of magnetic field, and contours of constant voltage constitute an electric-field plot.

4. Measurement of Shunt Resistance and Q. Shunt resistance and Q can be measured by the same general method indicated for cylindrical cavities. In the case of waveguide cavities operating on *TM* modes, however, the coil losses correspond to top and bottom surface losses. As a result, terminal resistors must be proportioned relative to the coil resistance to represent properly side-wall losses. Also, care must be taken in representing losses associated with gap loading, which, because of high current densities near the gap, will often be as large as all other losses combined.

5. Measurement of Ridge Waveguide Characteristics. Work on ridge waveguides has indicated a number of desirable characteristics.[8] These include low cutoff frequencies and increased separation of modes. Other characteristics, such as increased attenuation and lower power-transmission capabilities, are less favorable. The characteristics of ridge waveguides can readily be determined with the rectangular network. The network, being restricted to two-dimensional problems, can only give the field configuration at cutoff; however, with both the field configuration and cutoff frequency known, all of the characteristics of the waveguide at any frequency can be determined.[9]

The principal characteristics of interest in ridge waveguides are the cutoff frequencies of the various modes and the impedance of the fundamental mode. Cutoff frequencies are readily determined by simply setting up the waveguide in cross section with open boundaries, according to the *TE* analogy previously given. The network is then excited through a high series resistance to give the effect of loose coupling to a variable-frequency oscillator. The cutoff frequencies of the various *TE* modes of transmission are scaled by equation (12) in Part 1 from the frequencies at which the network resonates.

The characteristic impedance of a waveguide may be defined in terms of (a) equivalent current and voltage; (b) transmitted power and equivalent voltage; or (c) transmitted power and equivalent current.[10] The current-voltage definition is always the geometric mean of the other two. The spread of values is of the order of 2 to 1 in ordinary waveguide, and is expected to be relatively small in ridge waveguide because this approaches a strip transmission line for which the three above impedances are identical. The current-voltage definition is most useful in considering matching problems with two-conductor lines, and agrees most closely with the values measured through transducers from concentric line to waveguide. All of the impedance forms contain a factor which depends solely upon the geometry of the waveguide multiplied by a frequency-dependent term of the form $(1-\eta^2)^{1/2}$ where η is the ratio of the cutoff frequency of the mode considered to the actual frequency. This factor is infinite at the cutoff frequency and drops to unity at infinite frequency.

The current-voltage definition of impedance is

$$Z_{IV} = \frac{\int_b^t E_{y\max} dx}{\int_L^R H_z dx} \tag{21}$$

where the integral limits b, t, L, and R stand for bottom, top, left, and right, respectively, as shown in Fig. 8. All of the above quantities are simply related to the x

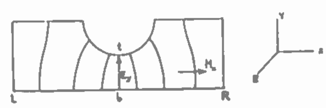

Fig. 8—Ridge-waveguide notation.

component of magnetic field of which the network analogue is voltage to ground. Specifically,

$$E_y = \frac{j\beta Z}{K^2} \frac{\partial H_z}{\partial x} \tag{22}$$

$$H_x = \frac{-j\beta}{K^2} \frac{\partial H_z}{\partial x} \tag{23}$$

where β is the phase factor, k is $2\pi/\lambda$, and $Z = (1-\eta^2)^{-1/2}$. With these substitutions, and replacing H_z by V, net-

[8] S. B. Cohn and P. Richards, "Very High Frequency Techniques," McGraw-Hill Book Co., Inc., New York, N. Y., 1947; vol. II, pp. 678–684.

[9] J. C. Slater, "Microwave Transmission," McGraw-Hill Book Co., Inc., New York, N. Y., 1942; p. 152.

[10] S. A. Schelkunoff, "Electromagnetic Theory," D. Van Nostrand Co., Inc., New York, N. Y., 1943; pp. 319–324.

work voltage to ground, the impedance expression becomes

$$Z_{VI} = 337 \frac{(1-\eta^2)^{-1/2} \int_b^t \frac{\partial V}{\partial x} dy}{V_R - V_L} \text{ ohms.} \quad (24)$$

From this it is seen that the impedance of the waveguide is given by 377 times the frequency factor, times the average transverse voltage gradient at the waveguide center, multiplied by the vertical dimension of the waveguide; all divided by twice the maximum voltage which appears on the edge of the network. Impedance values according to the other definitions are similarly obtained.

C. Solution of Radiation Problems

The networks may be used to solve antenna problems, give radiation patterns, surface currents, antenna impedances, junction transformations, etc. To simulate radiation problems, it is necessary to simulate the antenna contour and then terminate the edges of the network in such a way as to eliminate reflection of waves from it. Experiments with network terminations have shown that standing waves of the order of 1.10 or less are readily obtainable. This makes possible the determination of radiation patterns. Antenna impedances can be measured directly. Reactance transformation of junctions can be tested by simulating the surrounding of the antenna with reflecting spheres of different radii. For extensive work with antenna patterns, a special network containing a centrally located fine section with larger dimensions over-all in terms of wavelengths, and possibly having a polar co-ordinate system rather than a rectangular co-ordinate system, would be advantageous.

D. Accuracy and Speed of Network Determinations

The networks described here are not high-precision devices, but do give sufficient accuracy for engineering applications. Resonant frequencies may be determined to within 1 per cent accuracy for simple geometries or 2 per cent accuracy for complex geometries. Shunt resistance and Q determinations are accurate to within 5 per cent for complex geometries, but may be relied upon to be within 2 per cent accurate for simple geometries. This accuracy decreases somewhat as the electrical length of a section exceeds 25°.

A complete field plot of a resonator may be obtained in less than an hour. Resonant frequencies may be determined within a matter of minutes. Once equipment has been set up for shunt-resistance measurements, such determinations may also be made in a matter of minutes. In general, determinations are much more rapid than the equivalent calculation or model measurement.

VI. Experimental Verification of Performance

A. TEM Modes

A large number of resonant structures have been checked against the network to calibrate its operation. Checks on resonant frequency, field configuration, shunt resistance, and Q have been made. A few of these results are reported here to indicate the nature of agreement between the theory and the network results.

A basic reference test is the resonant concentric line operating on a TEM mode with its length being an integral number of half waves. Such a configuration, when set up, gives resonant frequencies which check within 1 to 2 per cent. The agreement on the field variation with line length is correspondingly good. Shown in Fig. 9 are plots of measured and theoretical

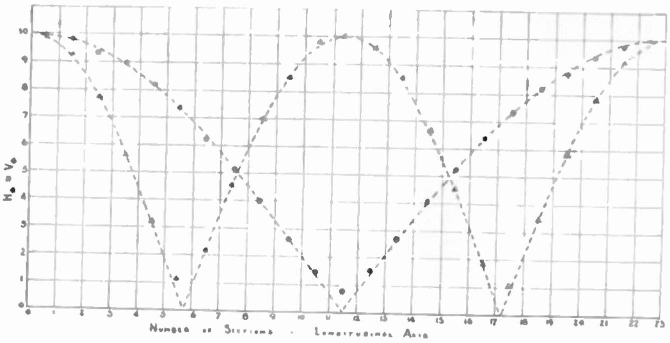

Fig. 9—Voltage distribution with length in a coaxial TEM mode (shorted coaxial line) 22 sections long, one-half section correction added at each end, (outer radius)/(inner radius) = 5. Test points:
- - - - - calculated
⊙ half-wave resonator
△ full-wave resonator.

voltage, analogue of rH_ϕ, versus length of the half-wave and full-wave resonance of a coaxial line. The agreement between the theoretical and measured cosine wave is good except in the vicinity of the nodes, where the true reading is somewhat obscured by harmonics and noise. This particular mode is convenient for checking the network adjustment on the cylindrical network. Since magnetic field varies inversely with radius, network voltage, which is proportional to rH_ϕ, will be constant with radius. It is very easy to check this condition with a vacuum-tube voltmeter, and any irregularities can usually be quickly traced. From resonant-frequencies measurements, the half-section correction for open-circuit boundaries can be verified.

B. Cylindrical TM_{0m0} Modes

The cylindrical-resonator TM_{0m0} modes are formed by radial waves which have no variations in intensity in the axial direction. The corresponding resonant frequencies are inversely proportional to radius. Tests with a large number of radii showed that measured network frequencies agreed with theoretical values within about 1 per cent when the radial-field edge correction was considered to be half a network section. For these modes, the radial electric-field variation is expected to be a zero-order Bessel function of the first kind. The magnetic field is expected to vary as a first-order Bessel function of the first kind. Figs. 10 and 11 show that these expectations were fulfilled quite closely.

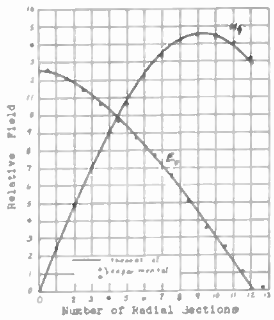

Fig. 10—Radial field variation in the TM_{010} mode.

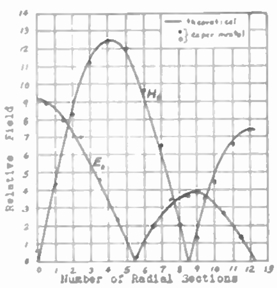

Fig. 11—Radial field variation in the TM_{020} mode.

C. TM_{011} Mode

The cases checked above have shown that the cylindrical-co-ordinate network behaves properly for purely radial or longitudinal waves. The question naturally arises as to whether it will properly represent a combination of these. The simplest mode which exhibits both radial and axial field variations is the TM_{011} mode. This mode is found with the same boundary conditions as the TM_{010} mode, it being necessary only to raise the frequency of the excitation voltage to the proper value. The excitation frequency in a typical experimental test was 136.8 kc, as against a theoretical value of 135.3 kc.

D. Net-Point Computations

In order to check the performance of the network on resonator structures presenting more complex geometries than the simple forms shown above, it is necessary to have recourse to rather involved theoretical approximations or resort to net-point computations. Net-point computations were preferred in checking network results because they are adaptable to arbitrary geometrical configurations, and also because the network gave the resonant frequency with sufficient accuracy that the computations were greatly simplified.

Basically, the net-point computation consists of finding a difference equation which corresponds to the differential equation pertinent to the problem; assuming values of field and then applying the difference equation to obtain successive improvements on the originally assumed values of field.[11-12]

In making net-point computations of field in resonators with rotational symmetry, it is convenient to use the parameter rH_ϕ, which is the analogue of network voltage to ground, V. This is convenient because, as has been shown, contours of constant rH_ϕ or V constitute an electric-field plot. Likewise, values of rH_ϕ or V divided by radius give magnetic field. The boundary conditions on rH_ϕ or V are also simple in that the normal gradient of rH_ϕ is zero at any conducting surface. The differential equation for V in two-dimensional, cylindrical co-ordinates with rotational symmetry has the form

$$\frac{\partial^2 V}{\partial z^2} + \frac{\partial^2 V}{\partial r^2} - \frac{1}{r}\frac{\partial V}{\partial r} + k^2 V = 0 \tag{25}$$

where k is 2π divided by wavelength. Note that this equation differs properly from the scalar wave equation in that the sign of the first derivative term in r is negative. The corresponding net-point or difference equation in terms of the network notation of Fig. 12 is

$$V_0 = \frac{V_1 + V_2 + V_3 + V_4 - \dfrac{(V_4 - V_3)}{2I}}{4 - h^2k^2} \tag{26}$$

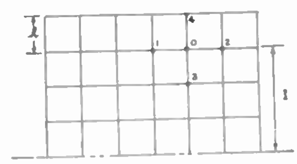

Fig. 12—Notation for net-point computations in cylindrical co-ordinates.

where h is the net-point spacing in wavelengths and I is the integral number of lattice sections from the axis. This equation relates the voltage in the center of a square to the voltage at four symmetrically disposed points arranged as shown in Fig. 12. If values of voltage are assumed at all points in a field or are taken from actual network measurements, then application of (26) will smooth out the values and give values which are more correct than those originally given.

Shown in Table VI are values of network voltage as taken directly from the network (above), compared with voltage values refined by a successive application of (25). It is seen that the errors in the field as presented in the network are relatively slight.

For cases in which the resonant frequency is not known, it is necessary to reiterate between values of field and frequency. However, the dependence of the field values upon the frequency is low. Hence if the resonant frequency is known within a few per cent, it is

[11] G. H. Shortley and R. Weller, "Numerical solution of Laplace's equation," *Jour. Appl. Phys.*, vol. 9, pp. 334–348; May, 1948.

[12] H. Motz, "Calculation of the electromagnetic field, frequency, and circuit parameters of high-frequency resonator cavities," *Jour. IEE* (London), vol. 93, part III, pp. 335–338; September, 1946.

possible to proceed to refine field values without correcting frequency.

In some applications, it is desired to use a different form of the difference equation given above. Where extensive computations are to be made, it is convenient to have a form where the center potential of a square is given by the sum of the four surrounding values divided by a constant dependent upon the radius. Such a form results if there is used the variable

$$U = V/\sqrt{r}. \tag{27}$$

Another useful transformation is

$$W = V/r^2. \tag{28}$$

The paramenter W has the advantage that it will tend to be constant in a resonator gap on the axis. Since V is proportional to rH_ϕ, and H_ϕ tends to increase linearly with radius in a gap, V/r^2 will tend to be constant.

It should be emphasized that the tests reported here and many others indicate a good agreement between the analyzer results and theory in cases where the results may be simply predicted and also in cases where the theoretical results are not simply predicted. As a result, it is felt that the analyzer is a useful tool which can be trusted in cases where the boundaries are too complicated to admit of only experimental or numerical determination.

VII. Conclusions

Networks have been constructed which give the solution of the wave equation in two-dimensional rectangular and cylindrical co-ordinates. Resonant frequencies of cavities, field plots, equivalent reactance of obstacles, shunt resistance, and Q are obtainable with little effort. The devices are simple to operate, and results can be obtained with considerable speed. The networks contain no vacuum tubes or moving parts and are simple and inexpensive to build. Accuracy, though not at the precision level, is sufficient for most engineering applications.

Two-dimensional networks are believed to be very practical. It is not thought that three-dimensional networks are practical.

The educational value of the networks is considerable. They possess a form of "intelligence," in that they often reveal the presence of unsuspected resonance modes. In addition, their use reveals a large number of important properties of fields which are little known. They can be operated by people not highly skilled in field theory.

VIII. Acknowledgment

The assistance of E. A. Goddard, P. Tissot, J. F. Honey, R. Tanner, and M. B. Adams with various phases of the work is gratefully acknowledged.

TABLE VI

	A	B	C	D	E	F	G	H	I	J	K
11–12	120 121	160 100	235 235	345 346	475 475	625 625	780 780	930 930	1090 1090	1190 1190	1250 1250
10–11	72.0 68.3	105 103.1	175 174.5	280 280	400 400	540 540	630 630	840 840	980 980	1070 1060	1120 1120
9–10	22.5 23.2	10.0 2.2	57.0 63.8	157 155	265 266	390 390	515 515	640 640	755 755	820 820	890 890
8–9	140 141	113 119.1	62.0 75.0	13.0 4.13	105 37.3	210 202.5	320 314	420 428	515 523	590 606	690 645
7–8	260 255	240 236.5	220 200	148 153.8	80.0 84.0	11.5 4.3	103 96.4	138 209	280 275	340 332	370 382
6–7	335 338	330 323	310 310	280 283.5	250 242	187 187.5	117 114.3	44.0 38.5	23.5 27.4	80.0 77.7	105 103.8
5–6	365 369	370 371	370 371	370 363	362 330	330 330	282 285	230 231	180 178.5	140 138.4	120 116.3
4–5	360 353	365 363	385 375	420 422	440 435	415 428	400 399	360 359	315 318	280 285	265 268
3–4	285 287	295 300	335*	425 429	470 470	470 470	450 453	415 421	380 385	355 355	340 339
2–3	187 189.9	191 195.5	202 205								
1–2	92.0 94.1	94.0 95.7	98.0 98.2								
0–1	24.0 25.4	25.0 25.6	25.0 25.9								
0	0*	0*	0*								

* This value kept constant.

Design of Optimum Transient Response Amplifiers*

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Summary—This paper describes a method based entirely upon transient considerations for the design of amplifiers with optimum transient response in the "least squared" sense. This method is applicable to video amplifiers, symmetrical band-pass amplifiers, and dissymmetrical band-pass amplifiers used with low-level modulation, and has provisions for taking into account noise acceptance and adjacent channel rejection. Examples are given of the results obtained with this method.

INTRODUCTION

Although all communication amplifiers are used in conjunction with signals of an essentially transient character, amplifier circuit design and synthesis has for the most part concentrated on the problem of obtaining predetermined steady-state characteristics. Methods of solving this steady-state problem have been brought to a high degree of perfection by Cauer,[1] Guillemin,[2] and Lee,[3] amongst others.

Recent developments, however, first of television, later of radar, and finally of pulse communication systems, showed that there are important cases where the performance of a system cannot be simply related to its steady-state characteristics, but rather is best expressed by its transient response.

Existing design methods for transient amplifiers are derived from steady-state theories, particularly those relating to filters. Butterworth,[4] Landon,[5] Wallman,[6] Bedford,[7] and Kallman, Spencer, and Singer[8] all studied amplifier transient response as related to steady-state characteristics. The shortcomings of their procedures are well understood, however, and have been pointed out by Kallmann.[8]

This paper describes a design method, based entirely upon transient analysis, which leads to amplifiers with an optimum transient response. This method is derived from operational analysis using Laplace transforms.

The procedure by which the transient response of an amplifier to some particular signal is calculated with Laplace transform technique involves the determination of the "operational transfer impedance," $AG(s)$, of the circuit under consideration. If a signal of Laplace transform $F_1(s)$ is applied to the amplifier, the output will have the transform

$$AF_2(s) = AG(s)F_1(s). \tag{1}$$

(The amplifier gain A, at low frequencies, is extracted as a coefficient for convenience.) Considerable difficulty is generally encountered when it is desired to calculate the actual amplifier response $f_2(t)$, the inverse transform of $F_2(s)$ in (1). This process requires the extraction of the poles of $G_2(s)$, an algebraic process only possible in general with known numerical values and of great difficulty when several poles are involved. It is apparent first that a simpler method is necessary for calculating the transient response. Such a method is described in part I.

I. THE LAGUERRE SERIES

The problem is that of determining $f(t)$ from a specified $F_2(s)$. This will be simplified first by using a function $F_3(t)$, defined as

$$f_3(t) = f_2(t) - C$$
$$C = f_2(\infty) = \lim_{s \to 0} sF_2(s) \tag{2}$$

or a function differing from $f_2(t)$ by the steady-state response $f_2(\infty)\cdot f_3(t)$ can be calculated as

$$f_3(t) = \sum_{n=0}^{\infty} \alpha_n B_n(t), \tag{3}$$

in which the terms α_n are the coefficients of an expansion in a Taylor series of

$$\frac{2}{1-x} F_3\left(\frac{1+x}{1-x}\right) = \sum_{n=0}^{\infty} \alpha_n x^n$$

and the terms $B_n(t)$ are "Laguerre functions"[9] defined by

$$B_n(t) = e^{-t} \sum_{p=0}^{n} \frac{(-2)^p}{p!} \binom{n}{p} t^p. \tag{4}$$

The series (2) is convergent for all values of t. Since

$$\frac{2}{1-x} F_3\left(\frac{1+x}{1-x}\right)$$

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‡ Carnegie Institute of Technology, Pittsburgh, Pa.

[1] W. Cauer, "Siebschaltungen," published by V.D.I., Verlag G.M.B.H., Berlin; 1931.

[2] E. A. Guillemin, "Communication Networks," John Wiley and Sons, Inc., New York, N. Y., 1931.

[3] Y. W. Lee, "Synthesis of electric networks by means of the fourier transform of Laguerre's functions," *Jour. Math. Phys.*, vol. 11, pp. 83–113; 1932.

[4] S. Butterworth, "On the theory of filter-amplifiers," *Exp. Wireless and Wireless Eng.*, vol. 7, pp. 536–541; October, 1940.

[5] V. D. Landon, "Cascade amplifiers with maximal flatness," *RCA Rev.*, vol. 5, pp. 347–362; January, 1941.

[6] H. Wallman, "Stagger Tuned I.F. Amplifiers," MIT Radiation Lab. Report 524; February, 1944.

[7] A. V. Bedford and G. L. Fredendall, "Transient response of multi stage video frequency amplifiers," Proc. I.R.E., vol. 27, pp. 277–289; April, 1939.

[8] H. E. Kallman, R. E. Spencer, and C. P. Singer, "Transient response," Proc. I.R.E., vol. 34, pp. 138–144; March, 1946.

[9] The Laguerre functions are related to the Laguerre polynomials, for which see "Orthogonal Polynomials", Szegö, Gabor, Colloquium Publications, American Mathematical Society, vol. 23; 1939.

is a rational fraction in x, α_n can be obtained by division of two polynomials. Furthermore, the $B_n(t)$ form a close orthogonal set from $t=0$ to $t=\infty$, that is

$$\int_0^\infty B_n B_m dt = \begin{cases} 0 & \text{if} \quad n \neq m \\ \frac{1}{2} & \text{if} \quad n = m \end{cases} \tag{5}$$

and every function $L(t)$ such that

$$\int_0^\infty L(t) B_n(t) dt = 0 \quad \text{for all } n \tag{6}$$

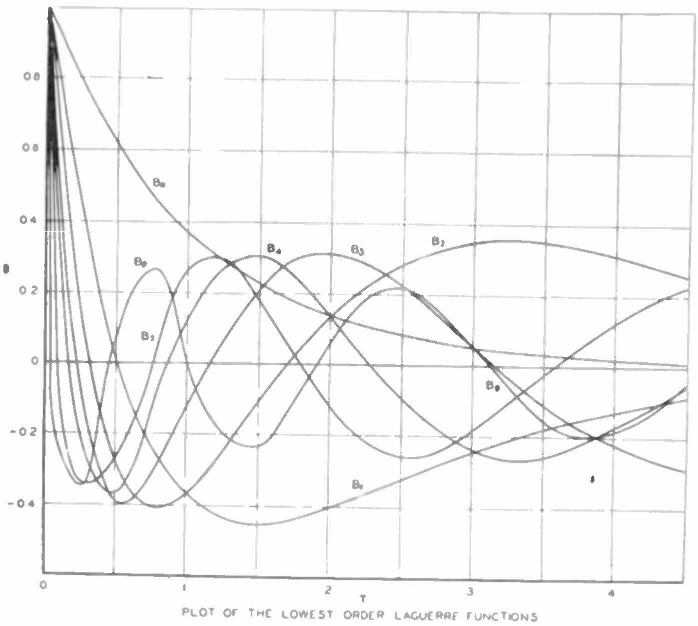

Fig. 1—Plot of the first nine Laguerre functions.

is identically zero almost everywhere. The first ten $B_n(t)$ have been computed and are given in Table I. Their form is shown in Fig. 1. Calculation of transient response is often simplified by the use of these $B_n(t)$ and (2) and (3); the principal value of these functions, however, is that they result in a simple means for determining the effect of individual parameter variations on the over-all transient response. This subject will be treated in part III. Part II deals with the definition of transient distortion.

II. Definition of Transient Distortion

The distorting or disturbing effects of principal interest in communication systems are non-linear distortion, random noise, adjacent channel interference, and transient distortion. When defining the first three it is customary to relate the actual response $f_q(t)$ to a specified signal to the output $f_p(t)$ of a "distortionless" system. The actual distortion is defined as the time average of

$$|f_p(t) - f_q(t)|^2$$

or, if the input is of finite duration, as

$$\int_{-\infty}^{\infty} |f_p(t) - f_q(t)|^2 dt.$$

This definition has given satisfactory results when applied to most types of distortion and it seems logical to extend it to transient distortion. It has, in fact, previously been so used.[10] Following this concept, transient distortion will be defined as follows: Assume a signal $f_1(t)$ to be fed into the amplifier at time $t=0$. The output will be $A\, f_2(t)$. An ideal amplifier would yield an output $A\, f_1(t-T_0)$ in which T_0 represents a delay time. The "transient distortion" will be defined as

$$I = \int_0^\infty |f_2(t) - f_1(t - T_0)|^2 dt. \tag{7}$$

This integral is convergent for all $f_1(t)$ with a finite energy and for all bounded input functions[11] which reach a final value in a finite time (e.g., a step function), provided the amplifier has no low frequency distortion. The low-frequency distortion problem in video amplifiers, however, is fairly distinct from that of high-frequency distortion, so that it is possible to define an "equivalent low-pass amplifier" without low-frequency distortion and work with it exclusively.

TABLE I

Values of the First Nine Laguerre Functions

t	$B_0(t)$	$B_1(t)$	$B_2(t)$	$B_3(t)$	$B_4(t)$	$B_5(t)$	$B_6(t)$	$B_7(t)$	$B_8(t)$	$B_9(t)$	$B_{10}(t)$
0.25	0.7788	0.3894	0.0974	−0.1136	−0.2576	−0.3470	−0.3926	−0.4026	−0.3881	−0.3527	−0.3033
0.50	0.6065	0	−0.3033	−0.4043	−0.3791	−0.2830	−0.1558	−0.0243	0.0935	0.1883	0.2551
0.75	0.4724	−0.2362	−0.4133	−0.3248	−0.1366	0.0550	0.2008	0.2828	0.3016	0.2679	0.1974
1.00	0.3679	−0.3679	−0.3679	−0.1226	0.1226	0.2698	0.3025	0.2441	0.1320	−0.0391	−0.1537
1.25	0.2865	−0.4298	−0.2507	0.0776	0.2753	0.2958	0.1897	0.0309	−0.1176	−0.2171	−0.2529
1.50	0.2231	−0.4463	−0.1117	0.2331	0.3068	0.1897	−0.0028	−0.1666	−0.2473	−0.2368	−0.1564
1.75	0.1738	−0.4344	0.0217	0.3005	0.2467	0.0310	−0.1668	−0.2507	−0.2160	−0.1013	0.0306
2.00	0.1353	−0.4060	0.1353	0.3158	0.1353	−0.1174	−0.2511	−0.2206	−0.0847	0.0737	0.1866
2.25	0.1054	−0.3689	0.2240	0.2833	0.0091	−0.2185	−0.2443	−0.1093	0.0701	0.1991	0.2286
2.50	0.0821	−0.3283	0.2873	0.2189	−0.1060	−0.2599	−0.1715	0.0267	0.1836	0.2206	0.1441
2.75	0.0639	−0.2877	0.3276	0.1372	−0.1943	−0.2457	−0.0634	0.1427	0.2250	0.1578	0.0145
3.00	0.0498	−0.2490	0.3185	0.0498	−0.2490	−0.1892	0.0498	0.2120	0.1952	0.0498	−0.1107
3.25	0.0388	−0.2132	0.3538	−0.0347	−0.2693	−0.1060	0.1467	0.2295	0.1194	−0.0682	−0.1875
3.50	0.0302	−0.1812	0.3473	−0.1107	−0.2605	−0.0156	0.2067	0.1905	0.0097	−0.1586	−0.1990
3.75	0.0235	−0.1529	0.3323	−0.1750	−0.2273	0.0718	0.2313	0.1202	−0.0897	−0.2015	−0.1510
4.00	0.0183	−0.1282	0.3114	−0.2259	−0.1771	0.1295	0.2202	0.0327	−0.1640	−0.1932	−0.0651

[10] W. W. Hansen, "Transient response of wide-band amplifiers," *Proc. Nat. Elect. Conference*, vol. 1, pp. 544–553; October, 1944.

[11] If the input is a delta (δ) impulse, the definition of transient distortion must be modified slightly to yield a convergent result. A suitable definition is

$$I = \int_0^\infty [f_2^2(t) - 2\delta(t - T_0) f_2(t)] dt = \int_0^\infty f_2^2(t) dt - 2 f_2(T_0).$$

In this study, in accordance with common practice in video amplifiers, $f_1(t)$, will be taken as a step function defined by

$$f_1(t) = 0 \quad \text{for} \quad t < 0$$
$$f_1(t) = 1 \quad \text{for} \quad t > 0.$$

The method can, if desired, readily be extended to other signals. The delay T_0 in (7) must be chosen to minimize I. For the step function

$$f_2(T_0) = \tfrac{1}{2}.$$

III. Calculation and Minimization of I

The transient distortion integral I can very easily be evaluated when the Laguerre series expansion of $f_3(t)$ is used. In fact, if we write

$$f_1(t - T_0) = \sum_{n=0}^{\infty} \gamma_n B_n(t) + 1,$$

then, using $f_3(t)$, from (2),

$$I = \sum_{n=0}^{\infty} (\alpha_n - \gamma_n)^2. \tag{8}$$

This formula permits the evaluation of the integral I without calculating $f_2(t)$ explicitly, providing values of $\gamma_n(T_0)$ are available. Table II gives these values for the case in which $f_1(t)$ is a step function.[12]

pendent parameters $\xi_1, \xi_2, \xi_3 \cdots \xi_n$. The integral I may be minimized when

$$\xi_1 = C_1,\ \xi_2 = C_2, \cdots, \xi_n = C_n,$$

if

$$\frac{\partial I}{\partial \xi_1}(C_1, C_2, \cdots, T_0) = 0$$
$$\frac{\partial I}{\partial \xi_2}(C_1, C_2, \cdots, T_0) = 0, \text{ etc.} \tag{9}$$

The series (3) and (4) are absolutely convergent, and (8) may be differentiated term by term. The optimizing equations corresponding to (9) are:

$$\sum_{n=0}^{\infty} (\alpha_n - \gamma_n) \frac{d\alpha_n}{d\xi_1} = 0. \tag{10}$$

Since the Laplace transform of the derivative of a function of time with respect to an independent parameter is the derivative of the Laplace transform of this function with respect to the parameter,

$$-\frac{\partial}{\partial \xi_1} \frac{2}{1+x} G\left(\frac{1+x}{1-x}\right) = \sum_{n=0}^{\infty} \frac{\partial \alpha_n}{\partial \xi_1} x^n. \tag{11}$$

This yields $\partial\alpha_n/\partial\xi_1$ just as (3) yields α_n. The system (11) solves the problem mathematically, but actually an approximate method is desirable for its solution. We as-

TABLE II

Values of $\gamma_p(T_0)$ for the Case of the Step Input Function

T_0	$\gamma_0(T_0)$	$\gamma_1(T_0)$	$\gamma_2(T_0)$	$\gamma_3(T_0)$	$\gamma_4(T_0)$	$\gamma_5(T_0)$	$\gamma_6(T_0)$	$\gamma_7(T_0)$	$\gamma_8(T_0)$	$\gamma_9(T_0)$	$\gamma_{10}(T_0)$
0.25	0.442	0.336	0.248	0.174	0.114	0.065	0.026	−0.006	−0.023	−0.048	−0.051
0.50	0.787	0.426	0.181	0.021	−0.072	−0.120	−0.134	−0.129	−0.107	−0.083	−0.051
0.75	1.055	0.362	−0.008	−0.171	−0.203	−0.180	−0.112	−0.052	0.015	0.053	0.088
1.00	1.264	0.207	−0.207	−0.283	−0.207	−0.087	0.024	0.091	0.133	0.198	0.137
1.25	1.427	0.006	−0.364	−0.293	−0.103	0.062	0.151	0.167	0.130	0.069	0.003
1.50	1.557	−0.215	−0.454	−0.215	0.048	0.186	0.199	0.129	0.053	−0.074	−0.097
1.75	1.672	−0.476	−0.436	−0.121	0.229	0.201	0.195	−0.027	−0.052	−0.167	−0.096
2.00	1.729	−0.647	−0.436	0.079	0.278	0.227	0.041	−0.102	−0.170	−0.147	−0.079
2.25	1.789	−0.841	−0.345	0.227	0.322	0.133	−0.082	−0.188	−0.170	−0.088	−0.029
2.50	1.836	−1.015	−0.216	0.353	0.297	0.011	−0.188	−0.209	−0.105	0.031	0.122
2.75	1.872	−1.189	−0.022	0.403	0.260	−0.158	−0.207	−0.205	0.041	0.094	0.193
3.00	1.900	−1.303	0.108	0.490	0.108	−0.228	−0.250	−0.074	0.108	0.183	0.172
3.25	1.922	−1.418	0.284	0.493	−0.023	−0.303	−0.202	0.037	0.184	0.192	0.047
3.50	1.940	−1.517	0.460	0.456	−0.157	−0.333	−0.111	0.144	0.218	0.119	−0.038
3.75	1.953	−1.600	0.630	0.385	−0.280	−0.318	−0.001	0.223	0.197	0.047	−0.168
4.00	1.963	−1.670	0.791	0.283	−0.381	−0.232	0.051	0.324	0.069	−0.011	−0.245

It is necessary, however, to evaluate T_0 (to avoid an explicit solution) and a simplified method of calculation is given subsequently. Equation (8) not only permits the calculation of I but also provides a means for choosing circuit parameters to render I a minimum. In general, all the coefficients of $G(s)$ cannot be chosen arbitrarily. Some relations between them are introduced by amplifier gain requirements and by unavoidable minimum values of interstage shunt capacitances. It is convenient to express the coefficients of $G(s)$ as functions of inde-

[12] If $f_1(t)$ is a delta impulse, $\gamma_n = B_n(T_0)$.

sume that approximate values $C_{10}, C_{20}, \cdots, C_{n0}$ can be found for $C_1, C_2, \cdots, C_n$ so that $\Delta_1, \Delta_2, \cdots$, defined by

$$\Delta_1 = C_1 - C_{10},$$
$$\Delta_2 = C_2 - C_{20},$$

are small quantities.

The equations (10) can be expanded in a series of powers of $\partial\alpha_n/\partial\xi_1$ and, neglecting all of order above the first, this series is

$$\sum_{n=0}^{N} (\alpha_n - \gamma_n) \frac{\partial \alpha_n}{\partial \xi_1}$$

$$+ \Delta_1 \left\{ \sum_{n=0}^{N} \left(\frac{\partial \alpha_n}{\partial \xi_1} \right)^2 + (\alpha_n - \gamma_n) \frac{\partial^2 \alpha_n}{\partial \xi_1{}^2} \right\}$$

$$+ \Delta_2 \left\{ \sum_{n=0}^{N} \frac{\partial \alpha_n}{\partial \xi_1} \cdot \frac{\partial \alpha_n}{\partial \xi_2} + (\alpha_n - \gamma_n)^2 \frac{\partial^2 \alpha_n}{\partial \xi_1 \partial \xi_2} \right\} = 0, \quad (12)$$

and other similar equations. This system (12) is linear and readily solved. The number N in this system corresponds to the last series terms considered to be of importance. Since numerous approximations are involved in (12), the values of $\xi_1, \xi_2, \cdots$, obtained will differ from $C_1, C_2, \cdots$. However, the amount by which ξ_1, differs from C_1, etc., is not so significant as the amount by which the transient distortion I of the calculated circuit differs from that of the best possible circuit. The approximations can be shown to introduce the following errors in I:

1. An error ϵ_1, due to the neglect of terms of order above the Nth:

$$\epsilon_1 < 2 \left| \sum_{n=0}^{\infty} 2\alpha_n \gamma_n - \alpha_n{}^2 \right| .$$

2. An error ϵ_2, due to taking for T_0 an approximate value,

$$\epsilon_2 < 0.01I \quad \text{if} \quad 0.47 < f_2(T_0) < 0.53$$

$$< 0.001I \quad \text{if} \quad 0.49 < f_2(T_0) < 0.51.$$

3. An error ϵ_3, due to the linear approximation to equation (10). This error cannot be ascertained easily, but is of the third order in $\Delta_1, \Delta_2, \cdots$, so that it is very small if $\Delta_1, \Delta_2, \cdots$ are small.

There remains the problem of determining $C_{10}, C_{20}, \cdots$. Any approximate design methods may be utilized. Marcuwitz[13] has proposed a method in which the first terms in the Laplace transform transfer impedance $G(s)$ in powers of s are equated to those of the series expansion of e^{-sT_0}. If all the terms could be equated a perfect transient response could result and further, if circuits of increasing complexity are considered, so that more and more terms can be equated, results of uniformly increasing quality are obtained. Although this method does not necessarily lead to an approximately optimum circuit in any given case with a finite number of terms, it does lead consistently to usable designs as a starting point for our optimization procedures. The Marcuwitz method may be replaced by any other approximate design data.

As an illustrative application consider the circuit of Fig. 2. This interstage coupling network has been studied by Kallmann, Spencer, and Singer[8] for equal input and output capacitances. We shall determine the optimum values of the parameters subject to choice with the sum of input and output capacitances fixed, but with an additional choice as to the division of this total capacitance between input and output circuits; that is, with

[13] N. Marcuwitz, "Distortionless Correction of Video Networks," M.E.E. Thesis, Polytechnic Institute of Brooklyn, June, 1941.

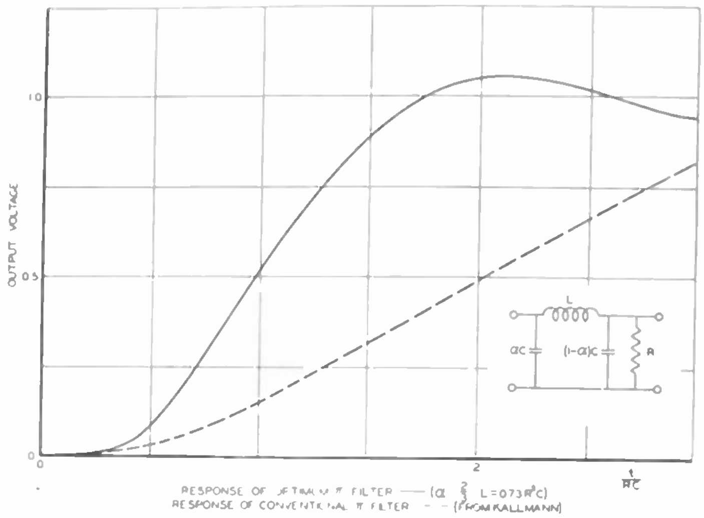

Fig. 2—Pi-filter interstage coupling network. Responses of the conventional type and the optimum type are compared.

fixed values of C and R (R from the gain selected) and variable L and α_c. The time unit is chosen as $t/(RC)$ so that a general solution is obtained. The notations $\xi_1 = \alpha_c LC$ and $\xi_2 = \alpha_c(1 - \alpha_c)$ will be used.

The Laplace transform of this circuit is

$$G_{(s)} = \frac{1}{p(s)} = \frac{1}{1 + s + \xi_1 s^2 + \xi_2 s^3},$$

and the transform of the response to a step function is

$$G_1(s) = \frac{1}{s} G(s) = \frac{1}{sp(s)}.$$

The derivatives of $G_1(s)$ with respect to ξ_1 and ξ_2 are

$$\frac{\partial G_1}{\partial \xi_1} = \frac{s}{p^2(s)} \qquad \frac{\partial G_1}{\partial \xi_2} = \frac{s^2}{p^2(s)}$$

$$\frac{\partial^2 G_1}{\partial \xi_1{}^2} = \frac{-2s^3}{p^3(s)} \qquad \frac{\partial^2 G_1}{\partial \xi_1 \partial \xi_2} = \frac{-2s^4}{p^3(s)} .$$

As the starting point approximation, use the circuit given by Kallmann, Spencer, and Singer[8] in which $\xi_1 = 0.45$ and $\xi_2 = 0.10$. Substituting $S = (1+x)/(1-x)$ in the above expressions, with these values of ξ_1 and ξ_2

$$\frac{2}{1-x} G_1\left(\frac{1+x}{1-x}\right) = \frac{3.1 - 3.6x + 1.5x^2}{2.55 - 3.25x + 1.85x^2 - 0.35x^3}$$

$$\frac{2}{1-x} \frac{\partial G_1}{\partial \xi_1} = (1 - x)M(x),$$

$$\frac{2}{1-x} \frac{\partial G_1}{\partial \xi_2} = (1 + x)M(x)$$

$$\frac{2}{1-x} \frac{\partial^2 G_1}{\partial \xi_1{}^2} = (1 - x)^2 N(x),$$

$$\frac{2}{1-x} \frac{\partial^2 G_1}{\partial \xi_1 \partial \xi_2} = (1 - x^2) N(x)$$

$$\frac{2}{1-x} \frac{\partial^2 G_1}{\partial \xi_2} = (1 + x^2) N(x),$$

in which

$$\sum_0^\infty m_p x^p = M(x) = \frac{1 - 2x + 2x^3 - x^4}{3.25(1 - 2.55x + 3.07x^2 - 2.12x^3 + 0.87x^4 - 0.2x^5 + 0.02x^6)}$$

$$\sum_0^\infty n_p x^p = N(x) = \frac{1 - 3x^2 + 3x^4 - x^6}{4.14(1 - 3.82x + 7.04x^2 + 6.15x^4 - 3.27x^5 + 1.2x^6 - 0.29x^7 + 0.04x^8)}.$$

We can now arrange in Table III the coefficients of the expansion of the above expressions in powers of x. Of course

$$\frac{\partial \alpha_p}{\partial \xi_1} = m_p - m_{p-1}, \qquad \frac{\partial \alpha_p}{\partial \xi_2} = m_p + m_{p+1},$$

$$\frac{\partial^2 \alpha_p}{\partial \xi_1^2} = n_p - 2n_{p-1} + n_{p-2}, \qquad \frac{\partial^2 \alpha_p}{\partial \xi_1 \partial \xi_2} = n_p - n_{p-2},$$

$$\frac{\partial^2 \alpha_p}{\partial \xi_2^2} = n_p + 2n_{p-1} + n_{p+2}.$$

IV. Extensions of the Method to Band-Pass Amplifiers and to the Treatment of Noise and Adjacent-Channel Rejection

The design method described so far applies only to video amplifiers. It can be extended to symmetrical band-pass amplifiers, or asymmetrical amplifiers with low-level modulation, however, by making use of the band-pass, low-pass analogy.[15,16] The design method here involves finding the low-pass equivalent of the band-pass amplifier under consideration, obtaining the optimum design for the video equivalent and reconverting to the band-pass equivalent.

TABLE III

p	γ_p	α_p	$\gamma_p - \alpha_p$	$3.25m_p$	$\frac{\partial\alpha_p}{\partial\xi_1}$	$4.19n_p$	$\frac{\partial\alpha_p}{\partial\xi_2}$	$\frac{\partial^2\alpha_p}{\partial\xi_1\partial_2}$	$\frac{\partial^2\alpha_p}{\partial\xi_1^2}$	$\frac{\partial^2\alpha_p}{\partial\xi_2^2}$
0	1.264	1.216	0.048	1.	0.308	−1.	0.308	−0.242	−0.242	−0.242
1	0.207	0.137	0.070	0.548	−0.139	− 3.82	0.475	−0.923	−0.437	−1.397
2	−0.207	−0.197	0.010	−1.678	−0.638	− 4.55	−0.347	−0.857	0.502	−3.166
3	−0.283	−0.184	−0.099	−1.839	−0.049	1.57	−1.076	1.302	1.644	−2.741
4	−0.207	−0.073	−0.134	−0.240	0.491	10.85	−0.638	3.72	0.758	2.014
5	−0.087	0.015	−0.102	1.202	0.442	14.39	0.295	3.12	−1.354	8.88
6	0.024	0.047	−0.023	1.458	0.079	9.12	0.816	−0.415	−2.162	11.82
7	−0.031	0.039	0.052	0.776	−0.209	− 1.26	0.685	−0.378	−1.200	7.30
8	0.133	0.018	0.115	−0.079	−0.262	−11.21	0.214	−4.88	0.103	−1.11
9	0.198	0.001	0.197	−0.559	−0.147	−18.22	−0.196	−4.67	0.706	−10.12

The coefficients of the equations (22) calculated from the elements of Table III are $1.08\Delta\xi_1 - 2.36\Delta\xi_2 = 0.169$ and $-2.36\Delta\xi_1 + 1.08\Delta\xi_2 = -0.233$, giving as a final result $\xi_1 = 0.36$ and $\xi_2 = 0.21$ or $\alpha_c = 2/3$ and $L = 0.73\ R^2C$. For this optimum circuit, $I = 0.0956$. The circuit response is given in Fig. 2 and compared with the conventional pi section derived from filter considerations.[14] The order of magnitude of the error terms is

$$\epsilon_1 \simeq 0.001, \quad \epsilon_2 \simeq 0.0009, \quad \text{and} \quad \epsilon_3 \simeq 0.0008.$$

In the event that the error terms should be too large, it would be necessary to use new starting values of ξ_1 and ξ_2 and to repeat the procedure. The results of this particular calculation indicate a pronounced improvement from using this video coupling network when the driving tube output capacitance exceeds the driven tube input capacitance by a substantial amount; the optimum ratio of two to one might readily be realized with some tube types.

[14] H. A. Wheeler, "Wide-band amplifiers for television," Proc. I.R.E., vol. 27, pp. 429-437; July, 1939.

There are many cases in which the transient response of an amplifier is not a sole performance criterion. Other factors of interest, such as noise acceptance and adjacent channel rejection, can be taken into account by a modification of the basic method.

The "noise acceptance" of an amplifier may be defined by:

$$M = \int_0^\infty A^2(f)df,$$

in which $A(f)$ is the amplifier amplitude response at a frequency f. It can also be expressed as

$$M = \int_0^\infty [f_2'(t)]^2 dt,$$

where $f_2'(t)$ is the response to a δ function. M is proportional to the power of the noise present at the output of the amplifier, when its input is connected to a random (or "white") noise source.

[15] V. D. Landon, "The band-pass low-pass analogy," Proc. I.R.E., vol. 24, pp. 1583–1584; December, 1936.

[16] P. R. Aigrain, B. R. Teare, Jr., and E. M. Williams, "Generalized theory of the band-pass low-pass analogy," to appear in Proc. I.R.E.

It would be almost hopelessly complex to try to choose the noise acceptance, and then minimize the transient distortion; indeed, the noise acceptance is given by a complex function of the circuit parameters. A preferable procedure is to minimize an expression of the form:

$$U = I + aM,$$

where a is some constant determined by the relative importance assigned to the transient response and to the noise acceptance. This procedure is particularly simple if one uses as test signal a delta-function. Corresponding to (7):

$$U = (1 + a)\int_0^{\infty} f_2'(t)dt - 2f_2'(T_0).$$

The calculations are similar to those involving transient response only, and the resulting equations are the same except that γ_n is replaced by $B_n/1+a$.

Adjacent channel rejection is of interest particularly in band-pass amplifiers in which it may be specified that some particular adjacent frequency must be strongly attenuated. As in the case of noise acceptance, it would be impractical to introduce this condition as a relation between the circuit parameters. It is, however, possible to handle this problem by minimizing the sum of the transient distortion and of the amplifier response to the undesired frequency. The amplifier input signal may be treated as the sum of a step function, for example, and a damped sine wave of frequency equal to the undesired frequency and with damping and amplitude dependent on the desired rejection. The frequency spectrum of such a signal appears in Fig. 3. Due to this additional

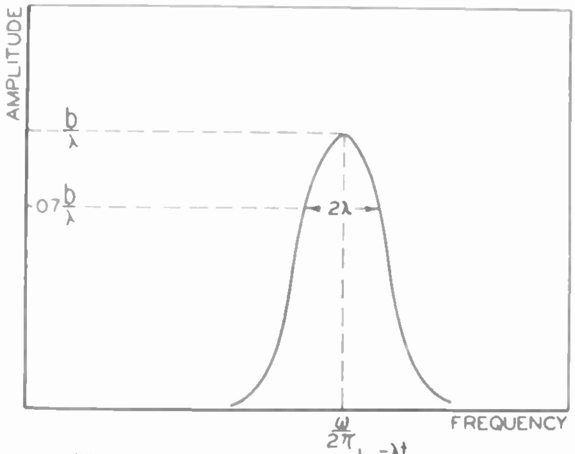

Fig. 3—Frequency spectrum of $be^{-\lambda t}\cos 2\pi ft$, the function used in minimizing adjacent channel rejection.

input signal, I as calculated from (1) will be increased by an amount which will be least if the desired frequency is properly rejected. If the damping and amplitude are small, the main part of the transient will be practically unaffected and the increase of I is given approximately by:

$$I_2 = I + \frac{A_f^2 b^2}{4\lambda}$$

A_f = response of amplifier at the rejected frequency

if the added function is $be^{-\lambda t}\cos 2\pi ft$. This procedure is thus equivalent to minimizing the sum of the transient distortion and the amplifier response at the undesired frequency.

V. Cascade Amplifiers

The method of optimizing transient response is, of course, applicable to cascade amplifiers. It is of considerable interest to determine whether all stages should be identical for optimum transient response. If one calls $\xi_{11}, \xi_{12}, \xi_{21}, \cdots, \xi_{r}$, ... the adjustable parameters referring to the 1st, 2nd, etc., circuit of the cascade, it is evident that I can depend on these parameters only through symmetrical combinations of $\xi_{11}, \xi_{12}, \cdots$, and $\xi_{21}, \xi_{22}, \cdots$. But any symmetrical function H of these is such that

$$\frac{\partial H}{\partial \xi_{1k}}\left[\sum_k \xi_{1k} = \text{constant}\right] = 0 \quad \text{if} \quad \xi_{11} = \xi_{12} = \xi_{13},$$

i.e., any symmetrical function of a number of variables does not change in value when all the variables are approximately equal, if any two variables are changed by a small amount, the sum of all variables staying constant.

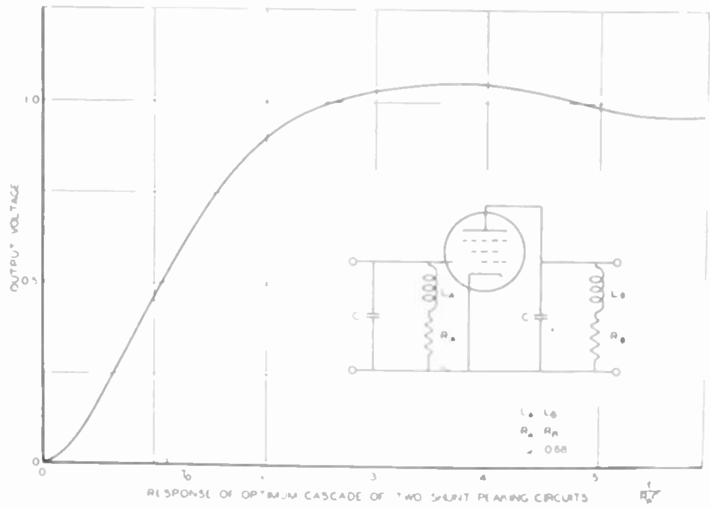

Fig. 4—Optimum cascade of two identical shunt peaked circuits and its response.

Thus

$$\frac{\partial I}{\partial \xi_{1k}}\left[\sum_k \xi_{1k} = \text{constant}\right] = 0 \quad \text{for} \quad \xi_{11} = \xi_{12} = \xi_{13}$$

since I depends only on functions such as H.

The design for which $\xi_{11} = \xi_{12} = 0 \cdots$ is thus optimum, provided that

$$\frac{\partial I}{\partial \xi_{11}} = 0, \quad \frac{\partial I}{\partial \xi_{21}} = 0.$$

since the other optimizing relations will follow immediately from these. The consequence of this can be expressed in other words by the theorem "For optimum transient response in a video amplifier cascade,[17] all sections of the cascade should be made identical."

[17] These conclusions also apply to the video equivalent, when such exists, of a band-pass amplifier. This includes the case of stagger-tuned intermediate-frequency amplifiers, for instance, in which there are an even number of stages, each pair being symmetrically tuned with respect to carrier frequency.

It is not possible to prove, but there are reasons to believe, that the optimum design for a section in a cascade is not very different from the optimum design of this section used alone. Indeed the amplitude as well as the phase distortion of each section will be cumulative, and thus it appears that the only way to improve the response of the amplifier as a whole is to make each section as good as possible. The results of a calculation of the optimum design for two shunt-peaking circuits are given in Fig. 4 and support this view.

Admittance of the 1B25 Microwave Switching Tube*

RALPH W. ENGSTROM† AND ARNOLD R. MOORE‡

Summary—**The operation of an RCA-1B25 neon-argon-filled glow tube as a switching element in a transmission-line network at ultra-high frequencies has been investigated. A test method has been developed and theory formulated for evaluating the behavior of the tube in terms of an equivalent susceptance and conductance. Experimental measurements have been made which show the variation of the admittance characteristics as a function of gas pressure and of the direct current.**

THE 1B25 IS a neon-argon-filled diode which fits into a coaxial transmission line and is used to switch a small ultra-high-frequency signal. Admittance characteristics of the tube were studied because of their importance in tube replacement.

A block diagram of the test equipment used to measure the conductance and susceptance of the tubes under test is given in Fig. 1. The 1B25 tube terminates the 150-ohm tunable coaxial line, as indicated. The method of determining the admittance of the 1B25 in operation requires measurement of the voltages at junction k (Fig. 1) with the switching tube unfired and fired as a function of the electrical angular distance of the 1B25 from the junction. The ratio of these voltages is termed the switching ratio

$$SR = |V_{ku}/V_{kf}|, \tag{1}$$

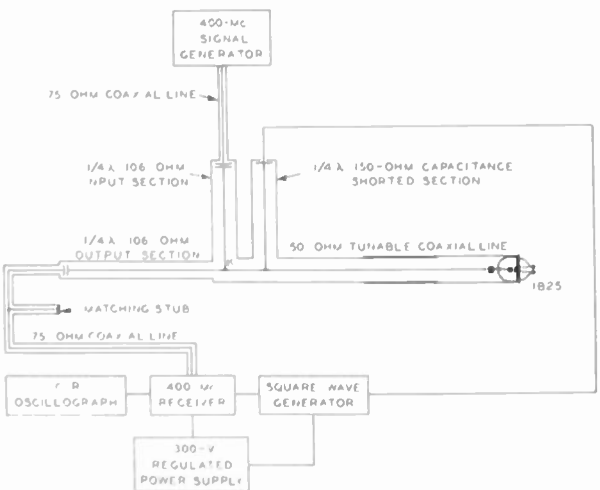

Fig. 1—Block diagram of experimental equipment.

wherein subscripts u and f refer to the unfired and fired states, respectively. This switching ratio was evaluated because of its applicability to operating conditions in the direction-finding equipment in which the tube finds its principal use.

The tube under test is fired at audio frequencies by a square wave of 250 peak volts. Included in the circuit is a frequency-doubler arrangement which delivers a square wave at twice the fundamental frequency. This double-frequency wave is applied to disable the receiver during part of each fired and unfired period. The switching ratio may then be measured as the relative height of two square-wave pips on the oscillograph.

The relative admittance of the 1B25 may be defined as

$$Y_u/Y_0 = jb_u + g_u, \text{ for the unfired state,} \tag{2}$$

and

$$Y_f/Y_0 = jb_f + g_f, \text{ for the fired state.} \tag{3}$$

Y_0 is the surge admittance of the line, Y_u and Y_f are the unfired and fired admittances of the 1B25, and the b's and g's represent corresponding relative susceptances and conductances.

The tunable coaxial line terminated by the 1B25 may be considered to have an electrical length of θ degrees

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from the junction k. When an analysis is made of the junction voltages for the fired and unfired condition, the following expression for switching ratio may be derived:

$$(SR)^2 = \frac{\left[3 + \dfrac{(g_f + 1)^2 + 2g_f + b_f^2}{(\cos\theta - b_f \sin\theta)^2 + (g_f \sin\theta)^2}\right]}{\left[3 + \dfrac{(g_u + 1)^2 + 2g_u + b_u^2}{(\cos\theta - b_u \sin\theta)^2 + (g_u \sin\theta)^2}\right]}. \qquad (4)$$

This working expression gives the ratio squared of the magnitudes of the received signal for the two conditions of the tube as a function of the length of the tunable line and of a set of constants which represent the lumped relative admittance of the switching tube during nonconducting and conducting periods.

The results of various possible switching-tube admittances, plotted from (4), are illustrated in Fig. 2. The parameters chosen and the equivalent circuits representing the types of "short" introduced by the firing of the 1B25 are indicated in the accompanying legend. Very distinct characteristics classify the possible types of tube behavior.

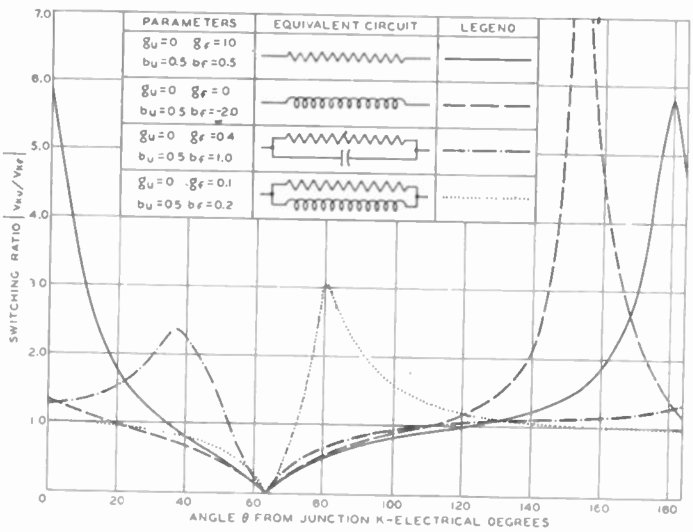

Fig. 2—Variation of switching ratio with length of tunable line, based on (4) for different possibilities of fired-tube admittance.

The inverse problem of determining the admittance characteristics from the experimentally measurable SR function required a graphical solution. Fig. 3 shows loci of $SR(\theta_{\text{min. fired}})$ and $\theta_{\text{min. fired}}$. The latter parameter was chosen as most convenient for the graphical analysis; it represents the electrical angular length of the tunable line for the condition that the junction voltage be a minimum during the fired condition. The switching ratio for $\theta_{\text{min. fired}}$ is in general nearly equal to the maximum switching ratio. In Fig. 3, g_u is assumed to be zero, as may be expected in a properly insulated tube. The value of b_u is 0.36, which corresponds to a $\theta_{\text{min. fired}}$ of 70.3°, or a capacitive reactance of 417 ohms at 400 Mc, and is the value to which all tubes are adjusted during manufacture. The intersection of loci $\theta_{\text{min. fired}}$ and $SR(\theta_{\text{min. fired}})$ defines b_f and g_f.

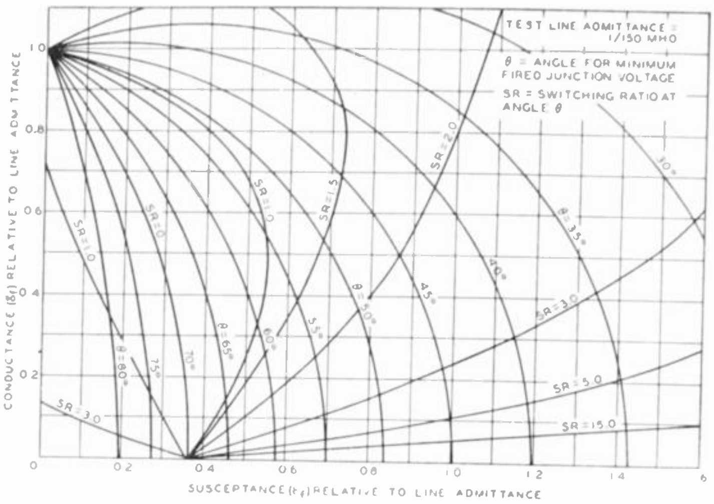

Fig. 3—Chart for obtaining conductance and susceptance from $\theta_{\text{min. fired}}$ and $SR(\theta_{\text{min. fired}})$.

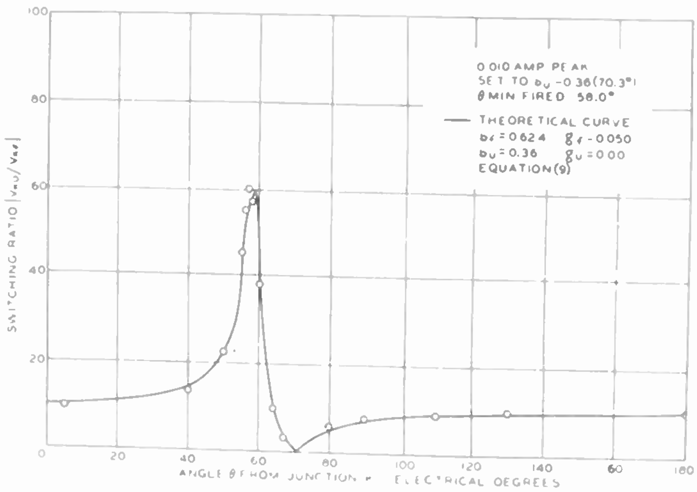

Fig. 4—Variation of switching ratio with length of tunable line for a typical 1B25, illustrating the accuracy with which the tube's behavior may be identified with admittance parameters.

Experimental points for an $SR(\theta)$ curve are shown in Fig. 4. The solid line shows the theoretical SR curve according to (4) with the parameters b_f and g_f as obtained from $\theta_{\text{min. fired}}$ and $SR(\theta_{\text{min. fired}})$ and the graph of Fig. 3. The fired gap admittance is, thus, equivalent to a capacitive susceptance in parallel with a conductance. All of the 1B25 tubes measured exhibited a similar admittance at 400 Mc within the range of attainable currents.

A more fundamental method of showing the behavior of a 1B25 tube is to plot susceptance and conductance of the ionized gas as a function of the peak discharge current. It is convenient to define the relative susceptance and conductance of the ionized gas:

$$b = b_f - b_u$$
$$g = g_f - g_u. \qquad (5)$$

Figs. 5 and 6 show values of b, g, and the ratio of b/g as a function of peak current through the tube. Gas pressure in these two cases was 100 and 195 mm of mercury of a mixture of neon and $1\frac{1}{4}$ per cent argon. The standard 1B25 tube is filled to a pressure of 150 mm of mercury. Large variations of these characteristics were found, but

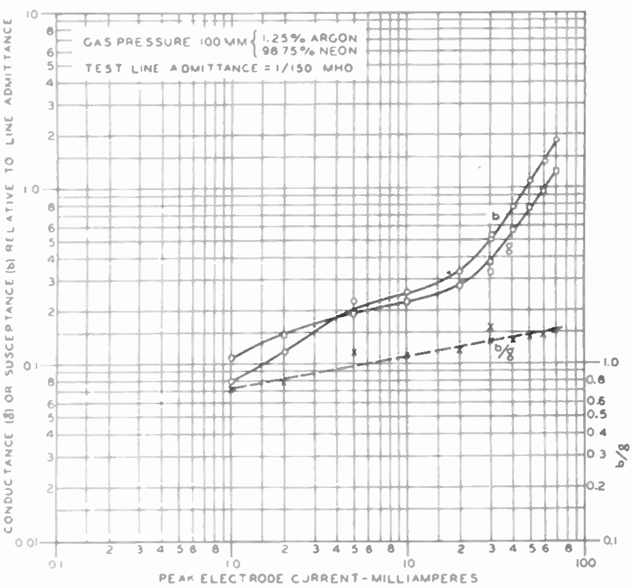

Fig. 5—Variation of conductance and susceptance at 400 Mc as a function of the direct current through a 1B25 tube having low gas pressure.

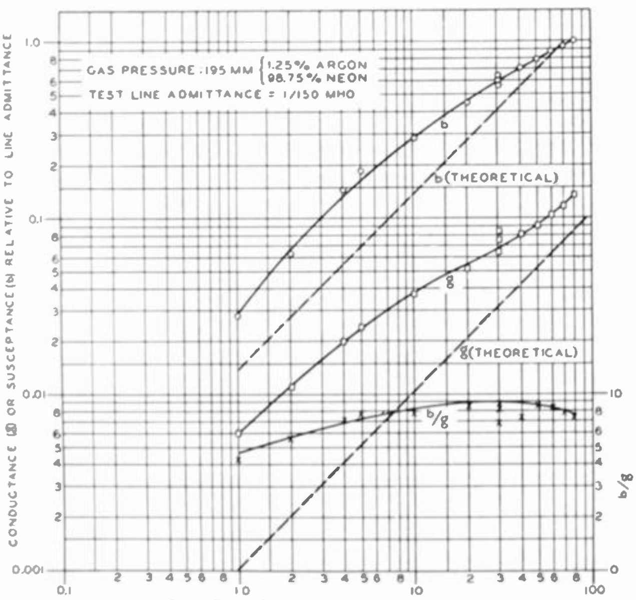

Fig. 6—Variation of conductance and susceptance at 400 Mc as a function of the direct current through a 1B25 tube having high gas pressure.

susceptance and conductance always increased with the direct current. The admittance of a tube usually changes during operation, particularly at the outset; the plotted values were measured, therefore, after the tube had been aged at least 30 minutes. The differences between operation at 100 and 195 mm of mercury, as illustrated in Figs. 5 and 6, are quite typical of the general shift in characteristics with gas pressure. Conductance (particularly) and susceptance both increase with decrease in pressure.

In Fig. 6 theoretical curves also are shown for b and g variation with current. Since the theoretical computation is lengthy and necessarily approximate, it has been omitted. The calculated behavior agrees with experiment only to the extent of indicating trends.

A large switching ratio was desirable in the antenna-switching application. This ratio could be obtained at high gas-discharge currents, but practical limitations on tube life limited the current to about 40 ma. Operation at pressures lower than 150 mm was excluded because of excessive sputtering and eventual short-circuiting of the electrodes. An attempt was made to increase the switching efficiency by mechanically restricting the area of glow between the two electrodes. A ceramic collar was built to enclose the sides of the cathode as illustrated in the sketch of Fig. 7. This collar restricted the cathode-glow region to the flat surface facing the anode. A more uniform distribution of cathode glow resulted, which tended to prevent erratic shifts in the switching ratio.

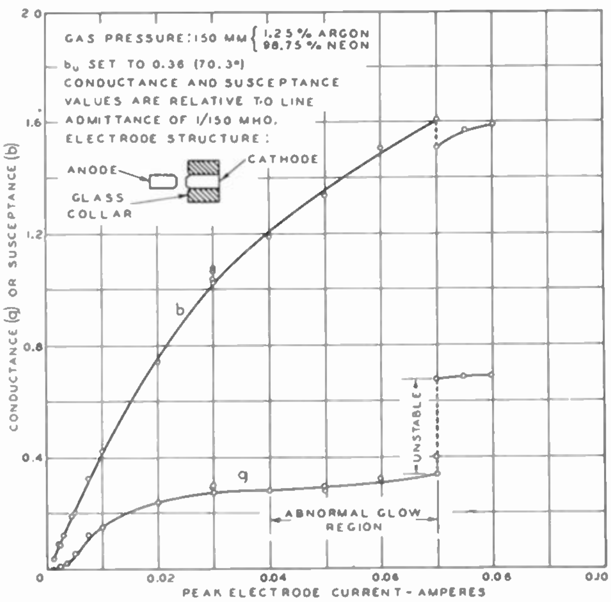

Fig. 7—Variation of the 400-Mc conductance and susceptance with current for a restricted discharge.

The voltage drop was increased, however, due to the abnormal glow condition. Conductance and susceptance values for the restricted discharge are plotted in Fig. 7. At high current values beyond the indicated unstable point, the cathode glow jumped around the collar and returned to a condition of unrestricted discharge. It may be observed that restriction of the discharge increases the susceptance and decreases the conductance of the ionized region.

Bridged Reactance-Resistance Networks[*]

G. R. HARRIS†

Summary—**Consideration is given to the various possible six-arm, six-element *RC* bridged networks of which the parallel-tee is the previously known example. It is found that six symmetrical *RC* structures exist having the infinite attenuation property of the parallel-tee and other properties differing from those of this structure. The duality of certain pairs of such structures is demonstrated.**

Scope of This Paper

THIS PAPER is limited to the development of certain networks having six arms only, each arm comprising a single resistance or reactance of common sign. The chief property of interest of these networks is the infinite attenuation attained at some real frequency determined by the network constants.

The designation "bridged" for this class of networks is descriptive of less than half the number found, but is adopted by analogy with the familiar bridged-tee configuration, which is at the same time a pi in series with a single shunt arm.

The Problem

A familiar example of the above-defined class is the parallel-tee configuration of resistances and reactances, usually capacitive. This structure has been employed for a variety of purposes: as a power-supply filter,[1] a radio-frequency bridge,[2] a frequency-determining element of an oscillator,[3] a selective amplifier,[4] low- and high-pass filters,[5] a linear frequency discriminator,[6] a motor control circuit,[7] and others.

In view of this fairly extensive application, extending over a period of ten years or more it is of interest to inquire whether the parallel-tee configuration is the only three-terminal configuration of resistances and capacitances (or inductances) capable of affording infinite suppression at a selected frequency. It will be shown that the parallel-tee is not in fact unique, but is a member of a class of six six-arm structures having this property.

Development of Equivalent Networks

The parallel-tee configuration, including its two terminating impedances, can be drawn in generalized form as the eight-arm mesh of Fig. 1. This mesh can be considered either as a configuration of impedances or admittances; it happens that the equation stating the

(a) (b)

Fig. 1

conditions for infinite attenuation of the parallel-tee falls in a more elegant form when expressed in admittances. For half of the networks to be developed, however, the reverse is true, and thus it is convenient to begin with both forms of the mesh in Fig. 1. It will be noted that the subscripts of the chord impedances of Fig. 1(a) are those of the branch admittances of Fig. 1(b), and vice versa. This inversion brings the equations for dual networks into parallel form and halves the number of equations to be solved.

If the opposite branches Y_1 and Y_5 of Fig. 1(b) are made the terminating impedances, the parallel-tee configuration of Fig. 2 appears. If, however, opposite chords are made terminating impedances, as Z_1 and Z_5 of Fig. 1(a), the configuration of Fig. 3 is obtained. These two networks are dual, as the series and shunt arms of either one appear as the correspondingly numbered shunt and series arms, respectively, of the other. The configuration of Fig. 3 might well be called a series-pi network by analogy with the parallel-tee of Fig. 2.

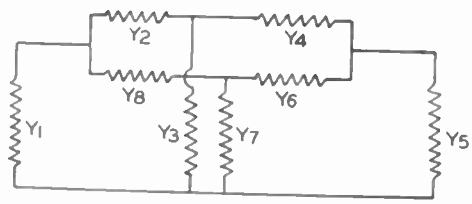

Fig. 2

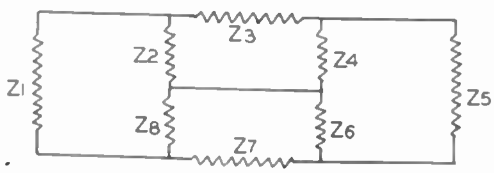

Fig. 3

Two other symmetrical configurations may be obtained by making adjacent chords of Fig. 1(b) and adjacent branches of Fig. 1(a) the terminating impedances. If Y_2 and Y_8 are so chosen in the first of these cases, the configuration of Fig. 4 results. This may be called a shunt-ladder network, since it is in fact a ladder

* Decimal classification: R143. Original manuscript received by the Institute, October 29, 1948; revised manuscript received, February 23, 1949.

† Jones & Laughlin Steel Corporation, Pittsburgh, Pa.

[1] H. W. Augustadt, United States Patent No. 2,106,785; 1938.

[2] W. N. Tuttle, "Bridged-T and parallel-T null circuits for measurements at radio frequencies," Proc. I.R.E., vol. 28, pp. 23–30; January, 1940.

[3] W. G. Shepherd and R. O. Wise, "Variable-frequency bridge-type frequency-stabilized oscillators," Proc. I.R.E., vol. 31, pp. 256–269; June, 1943.

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[5] G. J. Thiessen, "R-C filter circuits," *Jour. Acous. Soc. Amer.*, vol. 16, pp. 275–279; April, 1945.

[6] J. R. Tillman, "Linear frequency discriminator," *Wireless Eng.*, vol. 23, pp. 281–286; October, 1946.

[7] D. S. Bond, United States Patent No. 2,429,257; 1947.

in series with a single shunt arm. If Z_2 and Z_8 are chosen as terminating impedances, the configuration of Fig. 5 appears, which may be called a bridged-ladder. The networks of Figs. 4 and 5 are dual.

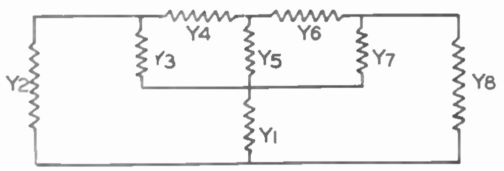

Fig. 4

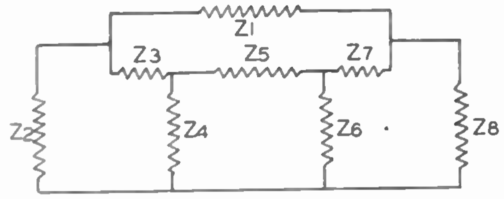

Fig. 5

Conditions for Balance

All four of the networks developed above, if considered as symmetrical about a vertical center line, have, of course, lattice equivalents, which may be easily derived through Bartlett's bisection theorem. It is, therefore, convenient and not inaccurate to speak of the conditions for infinite attenuation as the balance conditions.

Balance conditions may be determined by setting up the general circuit equations in terms of E, I, and either Y or Z, and equating either E or I to zero. As the network of Fig. 3 is drawn as a configuration of impedances, it will be convenient to determine its balance conditions from its mesh equations. The order of the meshes is obvious from Fig. 1(a) when E, the applied voltage, is considered to be in series with Z_1. The mesh currents are taken as flowing clockwise in each mesh.

$$
\begin{array}{llllll}
E &= I_1(Z_1+Z_2+Z_3) & - I_2Z_2 & 0 & & - I_4Z_8 \\
0 &= -I_1Z_2 & + I_2(Z_2+Z_3+Z_4) & - I_3Z_4 & & 0 \\
0 &= 0 & - I_2Z_4 & + I_3(Z_4+Z_5+Z_6) & & - I_4Z_6 \\
0 &= -I_1Z_8 & 0 & - I_3Z_6 & & + I_4(Z_6+Z_7+Z_8)
\end{array}
$$

$$
I_3 = E\,\frac{\begin{vmatrix} -Z_2 & Z_2+Z_3+Z_4 & 0 \\ 0 & -Z_4 & -Z_6 \\ -Z_8 & 0 & Z_6+Z_7+Z_8 \end{vmatrix}}{\Delta}
$$

where Δ is the full determinant. The requirement for balance is that $I_3=0$. If this substitution is made in the above expression, the minor is easily expanded to:

$$
0 = Z_2Z_4(Z_6+Z_7+Z_8) + Z_6Z_8(Z_2+Z_3+Z_4) \qquad (1)
$$

which expresses the conditions of balance for the network of Fig. 3. If the Z's in (1) are replaced by Y's of the same subscripts, the resulting equation in Y's expresses the conditions of balance for the parallel-tee network of Fig. 2.

The balance equation for the networks of Figs. 4 and 5 is derived in the same manner, and is found to be, for the shunt-ladder configuration of Fig. 4:

$$
\begin{aligned}
0 = {} & Y_3Y_7(Y_4+Y_5+Y_6) + Y_5(Y_3Y_6+Y_4Y_7) \\
& + Y_4Y_6(Y_1+Y_3+Y_5+Y_7). \qquad (2)
\end{aligned}
$$

This equation in Z's applies to the bridged-ladder configuration of Fig. 5.

The graphic method described by Tellegen[8] may be used to demonstrate the duality of these two pairs of configurations. If the configuration of Fig. 4, for example, is drawn in solid lines as in Fig. 6; a node placed in each of the four meshes of this configuration, as well as

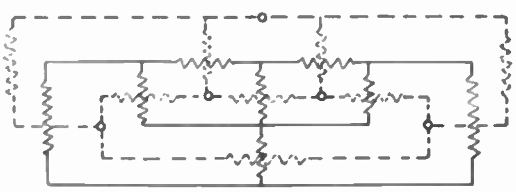

Fig. 6

one outside corresponding to the mesh circumscribing the configuration; and branches drawn from node to node so that each node is the junction of as many branches as its corresponding mesh has chords, then the dotted-line configuration of Fig. 6 is obtained. The dotted-line configuration is the dual of the solid-line configuration, and is seen to be the configuration of Fig. 5 previously developed.

[8] B. D. H. Tellegen, "Geometrical configurations and duality of electrical networks," *Phil. Tech. Rev.*, vol. 5, pp. 324–330; November, 1940.

Physically Realizable Structures: Parallel-tee and Series-Pi

As has been mentioned, this paper is confined to six-arm structures of six elements only. Since RC networks are of more general application than RL networks, the latter will not be discussed, although it is obvious that for every RC structure considered, a corresponding RL structure exists.

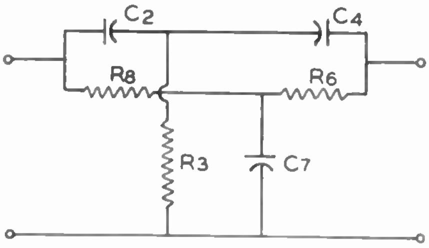

Fig. 7

The RC parallel-tee structure corresponding to Fig. 2 is that of Fig. 7. Its balance conditions are expressed in general terms by the Y form of (1) above

$$0 = Y_2Y_4(Y_6 + Y_7 + Y_8) + Y_6Y_8(Y_2 + Y_3 + Y_4).$$

Substituting g_3, g_6, g_8, jb_2, jb_4, and jb_7 for the corresponding Y's gives:

$$0 = -\ b_2b_4(g_6 + g_8 + jb_7) + g_6g_8(g_3 + j[b_2 + b_4].)$$

The real and imaginary terms may be separated and equated to zero in the usual manner:

Real: $$g_3g_6g_8 = b_2b_4(g_6 + g_8) \qquad (3)$$

Imaginary: $$b_2b_4b_7 = (b_2 + b_4)g_6g_8. \qquad (4)$$

At this point it is convenient to replace the g's and b's by the R's and C's of Fig. 7:

$$w^2C_2C_4\left(\frac{1}{R_6} + \frac{1}{R_8}\right) = \frac{1}{R_3R_6R_8}$$

$$w^3C_2C_4C_7 = \frac{w(C_2 + C_4)}{R_6R_8}$$

$$w^2 = \frac{1}{C_2C_4R_3(R_6 + R_8)} \qquad (5)$$

$$w^2 = \frac{C_2 + C_4}{C_2C_4C_7R_6R_8} \qquad (6)$$

$$f = \frac{1}{2\pi\sqrt{C_2C_4R_3(R_6 + R_8)}} \qquad (7)$$

$$f = \frac{1}{2\pi}\sqrt{\frac{C_2 + C_4}{C_2C_4C_7R_6R_8}}. \qquad (8)$$

If (5) is equated to (6),

$$(C_2 + C_4)C_2C_4R_3(R_6 + R_8) = C_2C_4C_7R_6R_8$$

$$\frac{C_2 + C_4}{C_7} = \frac{R_6R_8}{R_3(R_6 + R_8)}. \qquad (9)$$

The balance conditions for the structure of Fig. 7 may thus be expressed by two equations, either (7) or (8), and (9), the latter being independent of frequency.

The RC structure corresponding to the configuration of Fig. 3 is that of Fig. 8, the dual of Fig. 7. It is, of course, unnecessary to derive its balance equations, as they may be obtained from (5) through (9) previously derived merely by replacing all C's with R's of corresponding subscripts, and vice versa. These corresponding equations are:

Imaginary: $$w^2 = R_2R_4C_3(C_6 + C_8) \qquad (10)$$

Real: $$w^2 = \frac{R_2 + R_4}{R_2R_4R_7C_6C_8} \qquad (11)$$

$$= \frac{1}{2\pi\sqrt{R_2R_4C_3(C_6 + C_8)}} \qquad (12)$$

$$f = \frac{1}{2\pi}\sqrt{\frac{R_2 + R_4}{R_2R_4R_7C_6C_8}} \qquad (13)$$

$$\frac{R_2 + R_4}{R_7} = \frac{C_6C_8}{C_3(C_6 + C_8)}. \qquad (14)$$

Either of (12) or (13) together with (14) expresses the balance conditions for the structure of Fig. 8.

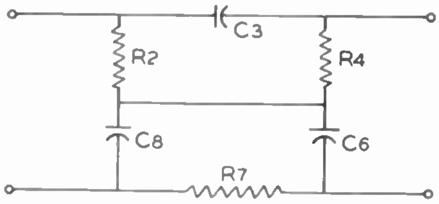

Fig. 8

Physically Realizable Structures: Shunt-Ladder and Bridged-Ladder

The parallel-tee RC structure discussed above was previously known, and the series-pi is its dual. The problem of deriving real RC structures for the networks of Figs. 4 and 5 however, is less easy of approach. Analogy with the previously developed structures and regard for symmetry suggest that if the three shunt arms, Y_3, Y_5, and Y_7 of Fig. 4, are made capacitances, the remaining arms must be resistances and the structure of Fig. 9 is thus obtained. Its balance equations

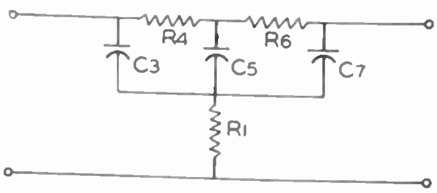

Fig. 9

may be derived from (2) by substituting the conductances and susceptances of Fig. 9 for the generalized admittances:

$$0 = -\ b_3b_7(g_4 + g_6 + jb_5) + jb_5(g_6jb_3 + g_4jb_7)$$
$$+\ g_4g_6(g_1 + j[b_3 + b_5 + b_7]).$$

The real and imaginary terms of the above may be separated and equated to zero:

Real: $$g_1g_4g_6 - b_3b_7(g_4 + g_6) - b_5(g_6b_3 + g_4b_7) = 0$$

Imaginary: $$b_3b_5b_7 - g_4g_6(b_3 + b_5 + b_7) = 0.$$

In terms of the element designations of Fig. 9 these equations become:

$$\text{Real:}\quad 1 = R_1w^2(R_4C_3[C_7 + C_5] + R_6C_7[C_3 + C_5]) \tag{15}$$

$$\text{Imaginary:}\quad w^2 = \frac{C_3 + C_5 + C_7}{R_4R_6C_3C_5C_7} \tag{16}$$

or

$$f = \frac{1}{2\pi}\sqrt{\frac{C_3 + C_5 + C_7}{R_4R_6C_3C_5C_7}}. \tag{17}$$

If the value for w^2 of (16) is substituted in (15) the resulting equation is:

$$\frac{R_4R_6C_3C_5C_7}{R_1(C_3+C_5+C_7)} = R_4C_3(C_7+C_5)+R_6C_7(C_3+C_5). \tag{18}$$

The balance conditions for the structure of Fig. 9 may thus be expressed by (17) and (18), the first giving the frequency at which balance occurs, and the second a relation which the network elements must satisfy.

Although the networks of Figs. 7 and 8 are the only possible symmetrical RC structures corresponding to the configurations of Figs. 2 and 3, the structure of Fig. 9 is not a unique embodiment of the configuration of Fig. 4. If arms Y_3, Y_5, and Y_7 are made resistances and the remaining arms capacitances, the structure of Fig. 10 is

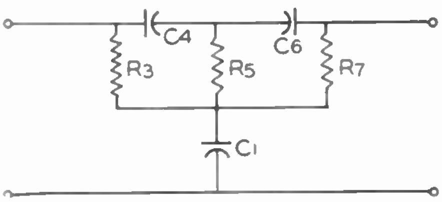

Fig. 10

obtained. Its balance equations, obtained as before, are:

$$\text{Real:}\quad 1 = w^2C_4C_6(R_5R_7 + R_3R_7 + R_3R_5) \tag{19}$$

$$\text{Imaginary:}\quad w^2C_1C_4C_6 = \frac{C_4(R_3 + R_5) + C_6(R_5 + R_7)}{R_3R_5R_7}. \tag{20}$$

From these are derived:

$$f = \frac{1}{2\pi\sqrt{C_4C_6(R_5R_7 + R_3R_7 + R_3R_5)}} \tag{21}$$

and

$$\frac{C_4(R_3+R_5)+C_6(R_5+R_7)}{C_1} = \frac{R_3R_5R_7}{R_5R_7+R_3R_7+R_3R_5}. \tag{22}$$

The bridged-ladder structures dual to the shunt-ladder structures of Figs. 9 and 10 may now be drawn immediately as Figs. 11 and 12, respectively, and their balance equations written down by inspection of those previously obtained. These equations for the structure of Fig. 11 are:

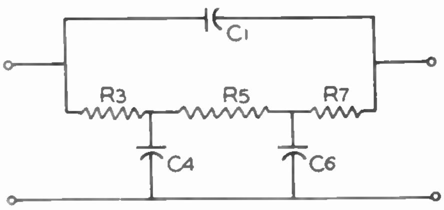

Fig. 11

$$f = \frac{1}{2\pi}\sqrt{\frac{R_3 + R_5 + R_7}{C_4C_6R_3R\ R_7}} \tag{23}$$

$$\frac{C_4C_6R_3R_5R_7}{C_1(R_3+R_5+R_7)} = C_4R_3(R_7+R_5)+C_6R_7(R_3+R_5). \tag{24}$$

The equations of balance for the structure of Fig. 12 are:

$$f = \frac{1}{2\pi\sqrt{R_4R_6(C_5C_7 + C_3C_7 + C_3C_5)}} \tag{25}$$

$$\frac{R_4(C_3 + C_5) + R_6(C_5 + C_7)}{R_1} = \frac{C_3C_5C_7}{C_5C_7 + C_3C_7 + C_3C_5}. \tag{26}$$

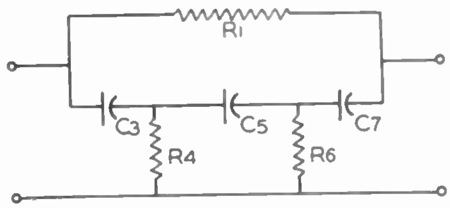

Fig. 12

Properties of Simplified Structures

When elements of the structures of Figs. 7–12, inclusive, are duplicated or triplicated, the balance equations reduce to more manageable forms, and, incidentally, reveal network properties which are masked by the general equations (7) through (26). If in the structures of Figs. 7 and 8 the elements having subscripts 2 and 6 are set equal to those of subscripts 4 and 8, respectively, the equations of balance become:

For Fig. 7:

$$f = \frac{1}{2\pi\sqrt{2}C_2\sqrt{R_3R_6}} \tag{27}$$

or

$$f = \frac{\sqrt{2}}{2\pi R_6\sqrt{C_2C_7}} \tag{28}$$

and

$$\frac{4C_2}{C_7} = \frac{R_6}{R_3}. \tag{29}$$

For Figure 8:

$$f = \frac{1}{2\pi\sqrt{2}R_2\sqrt{C_3C_6}} \tag{30}$$

or

$$f = \frac{\sqrt{2}}{2\pi C_6\sqrt{R_2R_7}} \tag{31}$$

and

$$\frac{4R_2}{R_7} = \frac{C_6}{C_3}. \tag{32}$$

As the properties of the parallel-tee have been discussed rather extensively and those of the series-pi are identical, no further consideration will be given them here.

If in the structures of Figs. 9–12, inclusive, elements of subscripts 3 and 4 are set equal to those of subscripts 7 and 6, respectively, the equations of balance become:

For Fig. 9:

$$f = \frac{\sqrt{1 + 2\frac{C_3}{C_5}}}{2\pi C_3 R_4} \tag{33}$$

$$\frac{R_4}{R_1} = \frac{4C_3}{C_5} + \frac{2C_5}{C_3} + 6 \tag{34}$$

For Fig. 10:

$$f = \frac{1}{2\pi R_3 C_4 \sqrt{1 + 2\frac{R_5}{R_3}}} \tag{35}$$

$$\frac{C_1}{C_4} = \frac{4R_5}{R_3} + \frac{2R_3}{R_5} + 6. \tag{36}$$

For Fig. 11:

$$f = \frac{\sqrt{1 + \frac{2R_3}{R_5}}}{2\pi R_3 C_4} \tag{37}$$

$$\frac{C_4}{C_1} = \frac{4R_3}{R_5} + \frac{2R_5}{R_3} + 6. \tag{38}$$

For Fig. 12:

$$f = \frac{1}{2\pi C_3 R_4 \sqrt{1 + 2\frac{C_5}{C_3}}} \tag{39}$$

$$\frac{R_1}{R_4} = \frac{4C_5}{C_3} + \frac{2C_3}{C_5} + 6. \tag{40}$$

These equations are significantly different from the corresponding equations (27) through (32). The frequency-determining equations here contain a term which is a ratio of two like elements; this makes it possible to multiply or divide within certain limits the frequency determined by the duplicated elements. Consider, for example, (33) and (34), which apply to Fig. 9. If $C_3/C_5 = 1$, these become:

$$f = \frac{\sqrt{3}}{2\pi C_3 R_4} \tag{41}$$

$$\frac{R_4}{R_1} = 12. \tag{42}$$

Suppose, however, that $C_3/C_5 = 2$. Then

$$f = \frac{\sqrt{5}}{2\pi C_3 R_4} \tag{43}$$

$$\frac{R_4}{R_1} = 15. \tag{44}$$

A more interesting property appears from further consideration of these equations. Suppose $C_3/C_5 = 1/2$. Then

$$f = \frac{\sqrt{2}}{2\pi C_3 R_4} \tag{45}$$

$$\frac{R_4}{R_1} = 12. \tag{46}$$

For given values of C_3, R_4, and R_1 there are two independent frequencies of balance, (41) and (45), corresponding to two particular values of C_5. This double frequency of balance results from the form of (34) and corresponding equations for the other structures of this group.

In (34) for example, let $R_4/R_3 = P$, and $C_3/C_5 = Q$. The equation then becomes:

$$P = 4Q + \frac{2}{Q} + 6. \tag{47}$$

It has been shown just previously that $P = 12$ for two different values of Q. The minimum values of P and Q are obtained in the usual manner by differentiating P with respect to Q and setting the derivative equal to zero.

$$\frac{dP}{dQ} = 4 - \frac{2}{Q^2} = 0$$

$$Q^2 = \frac{1}{2}$$

$$Q_{min} = \frac{1}{\sqrt{2}} \quad \text{or} \quad 0.707 \tag{48}$$

$$P_{min} = \frac{4}{\sqrt{2}} + 2\sqrt{2} + 6 = 11.656. \tag{49}$$

Values of P less than this cannot be chosen if real results are to be obtained from (33) through (40). A graph of (47) is shown in Fig. 13, the Q scale being logarithmic. Drawn in this manner, the curve is a parabola.

The shunt-ladder and bridged-ladder structures differ from the parallel-tee and series-pi in still another way. It is possible to make either all R's or all C's the same value in both the latter two structures, but not in former structures. In the shunt-ladder and bridged-ladder configurations, the three resistances or capacitances occupying similar arms may be made equal, but not the elements of the other three arms. In Fig. 9, for example, all C's can be made equal, but not the R's; in

Fig. 10 the R's can be made equal, but not the C's. This follows from (49) which shows that the minimum value of P is a little less than 12.

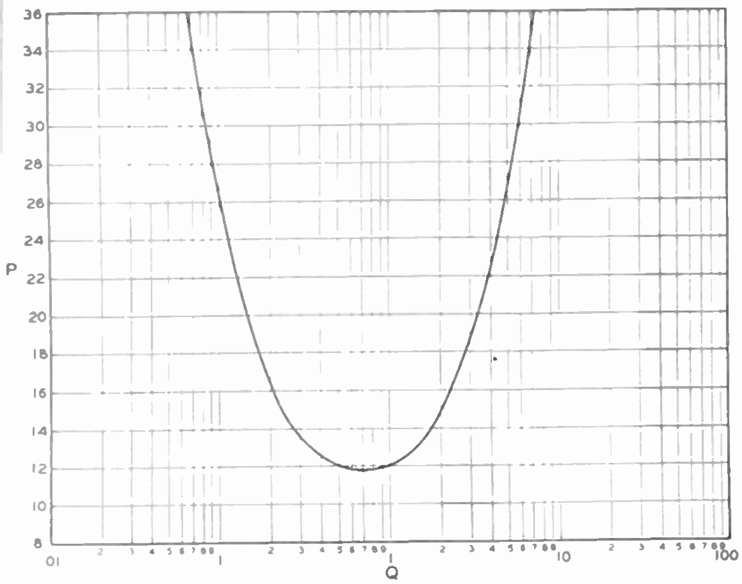

Fig. 13

Although this paper will not discuss particular applications of the structures considered, some practical advantages of the shunt-ladder and bridged-ladder structures will be briefly mentioned. Where networks balancing at low frequencies are desired, the structure of Fig. 12 offers economies in the size of the capacitances required. Where a high frequency balance is wanted, the structure of Fig. 9 is more suitable, as its capacitance values may be made appreciably larger for a given balance frequency than those of the others. The structure of Fig. 9 is useful if the three capacitances are to be varied, as they have a common terminal. This terminal need not be too far above ground potential, since the ratio of the series to the shunt resistances may be made fairly large.

Additional Networks

The configurations of Figs. 2 through 5 are not all those which may be developed from the mesh of Fig. 1. Two other such exist, but as they are not symmetrical and do not appear to possess any properties not previously known, they will be mentioned only briefly. One is obtained from making an adjoining branch and chord the terminating impedances; the resulting three-terminal configuration cannot be embodied in an RC structure which balances at any real frequency. The other is obtained from making an opposite branch and chord the terminating impedances; the resulting four-terminal configuration is that of the well-known Anderson Bridge. It is possible to balance this structure when only two of its six arms are capacitances, but in this respect it merely equals the Wien RC Bridge. Each of these additional configurations is its own dual.

Ionospheric Virtual Height Measurements at 100 Kilocycles*

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***Summary*—A simple high-power ionosphere sounding equipment for use at low frequencies is described. Its application to the measurement of virtual height at vertical incidence at a carrier frequency of 100 kc is reported. Results of nearly a year of intermittent nighttime measurements showed a remarkably large variation in virtual height which ranged from about 34 km to as high as 106 km. Some evidence was found to indicate that frequently at night the reflecting region consists of clouds or patches of ionization, rather than the more nearly uniform ionization characteristic of the regular layers at high frequencies. A rotation of the polarization of the reflected signal with respect to that of the transmitted signal was observed.**

I. Introduction

RECENTLY, considerable interest has arisen in the behavior of the ionosphere at low and very low frequencies. The development of low-frequency precision aids to navigation, such as lf Loran, has necessitated a detailed knowledge of the variations in the electrical properties of the ionospheric reflecting region at these frequencies. Of particular interest is the virtual height[1] of reflection of low frequency sky-wave signals, for upon the virtual height depends the time delay and arrival angle of a sky wave. It is the purpose of this paper to report the initial results of virtual height measurements made at a carrier frequency of 100 kc, and to describe the special equipment developed for this purpose.

Much of the knowledge of the electrical and physical properties of the upper atmosphere has been obtained directly or indirectly by radio methods applied mainly at frequencies above 500 kc. The foremost of these is the group retardation or "pulse" method of Breit and Tuve which gives the virtual height of reflection of a short pulse at a given carrier frequency. Because the heights

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[1] "Virtual height" is defined as the height at which a signal would be reflected, if the ionosphere were replaced by a perfectly reflecting surface.

at which propagation is influenced by the presence of ionization tend to increase with the frequency, the characteristics of the lowest regions of the ionosphere can best be investigated by radio methods only at low frequencies. Thus it would seem that the study of the ionospheric reflections of low-frequency waves offers an excellent means of increasing our knowledge of the lower ionosphere.

A limited amount of data on the propagation of sky waves at very low frequencies have been obtained by the continuous-wave interference method[2] and by the observation of signals from lightning discharges.[3] In the continuous-wave technique, variations in effective height are determined from changes in the relative phase of ground and reflected waves arriving at a distant receiver. Polarization and field intensity can also be determined by this method. One of its limitations is the difficulty in interpreting the results when more than one reflected wave is present, which is the usual situation during nighttime. Another difficulty arises in attempting to apply the method at vertical incidence, where one must somehow attenuate the ground wave by several orders of magnitude.

The other method makes use of the fact that a lightning discharge radiates a pulse-type signal which is short enough to permit the resolution of multiple reflections between earth and ionosphere. The virtual height of the reflecting layer and the distance of the source are found from a measurement of the relative times of arrival of three or more pulses which have been reflected from the ionosphere a known number of times. This method is fundamentally restricted to very low frequencies, since the principal frequency components in a typical lightning discharge lie below about twenty kc. Various practical limitations prevent the fullest utilization of these natural radio signals. These include the difficulty of determining the location and exact character of the discharge. Furthermore, there is evidence to indicate that abnormal ionization may be produced above a thunderstorm. If this be the case, then vertical-incidence measurements of the normal ionosphere would be impossible when using this method.

[2] J. E. Best, J. A. Ratcliffe, and M. V. Wilkes, "Experimental investigations of very long waves reflected from the ionosphere," *Proc. Roy. Soc.*, vol. 156, pp. 614–633; September, 1936.

[3] B. F. J. Schonland, et al., "The wave form of atmospherics at night," *Proc. Roy. Soc.*, vol. 176, pp. 180–202; October, 1940.

II. Apparatus

General Description

The transmitted pulse consists of an exponentially damped oscillation which is produced about once every second by directly exciting a large "inverted-L" type of antenna. This pulse travels by way of the ground and the ionosphere to the receiver, located 2.7 km away. A block diagram of the system is shown in Fig. 1. The ground and principal sky waves are radiated by the 30-meter high vertical part and the 440-meter long

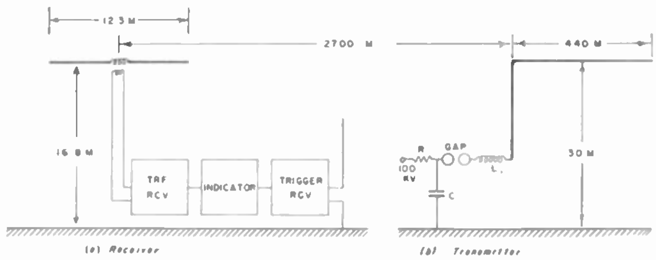


Fig. 1—Block diagram.

horizontal part of the antenna, respectively. Since the antenna is less than a quarter-wave long at 100 kc, a loading inductance (L_1) is required to tune the antenna to resonance. Energy for the pulse is obtained by slowly charging a relatively large capacitor C, through a high resistance R, from a dc source of high voltage ranging from 100 to 200 kv. The charged capacitor is connected in series with the antenna by means of a variable-length

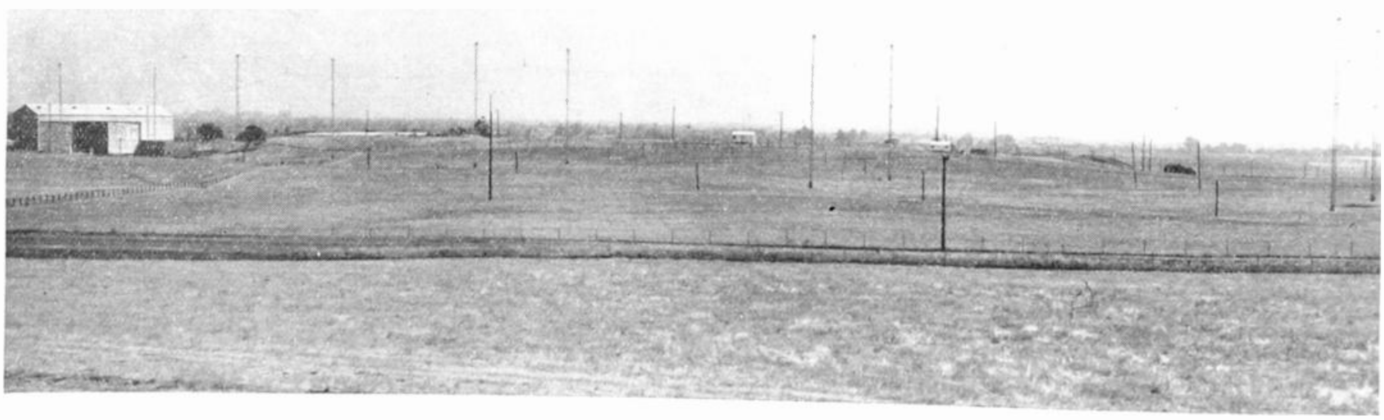


Fig. 2—Transmitter site.

sphere gap. One of the spheres is mounted on an arm which rotates at a rate of about 1 rps. An instant before the moving sphere reaches the position nearest the fixed sphere, the gap breaks down and the energy stored in the capacitor flows into the inductance and the antenna which may be regarded simply as a transmission line. The repeated reflections of diminishing amplitude provide an exponentially damped oscillation whose period depends primarily on the line length and the size of L_1, and whose time constant depends primarily on the resistance of the ground connection. To reduce the pulse length, additional resistance may be placed in series with the antenna. With typical circuit constants, no external resistance, and an applied voltage of 100 kv, the peak current in the pulse is 200 amperes, and the time constant is 150 μs. A photograph of the transmitting antenna and the Ryan High Voltage Laboratory is shown in Fig. 2.

The principal receiving problem is obtaining an adequate signal-to-noise ratio. In spite of the large current flowing in the transmitting antenna, the power radiated from the horizontal part is only of the order of 1 kw because of the extremely low radiation resistance. At low frequencies and at low and medium latitudes, atmospheric static is usually the limiting factor in receiver sensitivity. A substantial reduction in the pickup of such static is effected by use of a balanced horizontal dipole which is relatively insensitive to signals arriving at low vertical angles. To permit the "nulling out" of ground waves and to allow the polarization of the downcoming waves to be investigated, the dipole is universally mounted. Azimuth and tilt are controlled at the receiver by two sets of selsyns. By proper orientation of the dipole in the horizontal plane, its response to vertically polarized signals can be reduced by 60 db below that with the dipole vertical. The dipole is shown in the photograph of the receiving site in Fig. 3.

The pulses picked up by the dipole are amplified by a simple broadband TRF receiver of straightforward design. The output of the receiver is fed to a five-inch oscilloscope whose time base is triggered slightly in advance of the arrival of the ground pulse at the indicator by the strong ground pulse which is received on a broadly tuned vertical antenna. The time delay of the echoes with respect to the ground pulse is measured by means of timing markers which are derived from a 30-kc ringing oscillator. The ringing oscillator is triggered by the oscilloscope gate voltage. A photographic record of each trace is made with a motor-driven single-shot 16-mm movie camera, which is controlled by the gate voltage from the oscilloscope in such a way that the film moves ahead one frame each time the sweep is triggered. The time is indicated on each frame by a seconds-counter situated in the upper right-hand corner directly in front of the cathode-ray tube.

Fig. 3—Receiver site.

Power Supply

With the relatively low repetition rate of one pulse every one and one-half seconds, the average demand from the power supply is small, being about twenty watts under typical operating conditions. A maximum voltage of about 200 kv is obtained from nine cascaded half-wave rectifiers which are supplied from a 23-kv, 60-cycle transformer. A simplified schematic of the supply is shown in Fig. 4. Each stage consists of two 27,000-volt, 0.125-μf capacitors in series with a half-wave high-vacuum rectifier, type 250R. To avoid the problem of filament transformer insulation, the filament power is obtained from a 1-kw 450-kc oscillator which supplies

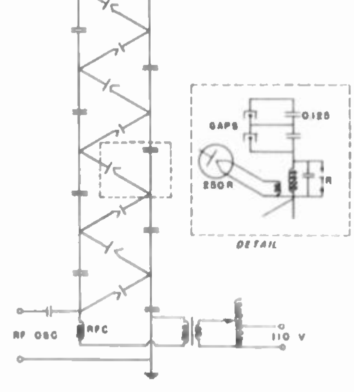


Fig. 4—Power supply schematic.

a current of one ampere to the autotransformer primaries which are connected in series through the capacitors. The secondaries supply the required filament current of ten amperes to each of the nine rectifiers.

Each autotransformer is tuned approximately to resonance by means of a shunt capacitance. The effect of variations in primary current from stage to stage is compensated by individually adjusting the tuning of each stage. The measurement of filament power is readily made with the aid of an optical pyrometer which is first calibrated, using a filament carrying a known effective current.

Over-voltage protection is provided by spark gaps connected across each capacitor, and "transmit-receive" gas tubes connected across each auto transformer.

Antenna Current

The current which flows in the antenna can be determined by considering the circuit to be made up of an inductance in series with a transmission line and a source of voltage. It is assumed that the capacitance C, which appears in the block diagram of Fig. 1, is so large compared with the antenna capacitance that the variation in voltage across C during the charging of the line is small and can be neglected to a good approximation.

The method of solution most adaptable to the present problem is to consider the total current to be the superposition of a series of sinusoidal components. Their amplitudes and frequencies may be determined by solving the transmission line differential equations in the presence of the given initial and boundary conditions by the method of the Laplace Transformation.[4] From such an analysis one finds the current at a distance x from the sending end of the line to be

$$I = \frac{4E}{\pi z_0} \sum_{n=1}^{\infty} \sin 2\pi f_n t \frac{f_0\left[\cos \frac{\pi f_n}{2lf_0} x - \frac{2\pi f_n L_1}{z_0} \sin \frac{\pi f_n}{2lf_0} x\right]}{f_n\left[1 + 4f_0 \frac{L_1}{z_0} + 4\pi^2 f_n^2 \left(\frac{L_1}{z_0}\right)^2\right]} \quad (1)$$

and the frequency of the nth mode to be given by

$$\tan \frac{\pi}{2} \frac{f_n}{f_0} = \frac{z_0}{2\pi f_n L_1} \quad (2)$$

where

E = voltage of source capacitor
z_0 = characteristic impedance of line
f_n = frequency of the nth mode of oscillation
$f_0 = c/4l$ = natural resonant frequency of line
c = velocity of light
l = length of line
L_1 = loading inductance.

[4] H. S. Carslaw and J. C. Jaeger, "Operational Methods in Applied Mathematics," Oxford University Press, 1941.

Equation (1) is useful in obtaining the amplitude of the fundamental component of current from which the field intensity of the radiated wave may be calculated. Equation (2), which is best solved graphically, gives the loading inductance required to tune the line to the desired frequency.

Field Intensity

The complete expression for the radiation components of the field produced by an "Γ" antenna is rather involved;[5] however, for vertical incidence work, the field intensity of interest is that directly above the antenna. This may be computed readily by assuming a sinusoidal current distribution on the antenna, and adding the direct and ground-reflected components produced by the horizontal part. This leads to the following expression for the magnitude of the field intensity in volts per meter at the receiving antenna assuming unity reflection coefficient at the ionosphere:

$$E_r = \frac{30I_0}{d} \frac{1 - \cos \frac{2\pi b}{\lambda}}{\sin 2\pi \frac{a+b}{\lambda}} \left|1 - r_T/\underline{\alpha_T}\right| \left|1 - r_R/\underline{\alpha_R}\right|, \quad (3)$$

where

I_0 = base current in amperes
d = total path distance in meters
b = length of transmitting antenna flat top in meters
a = height of transmitting antenna above ground
λ = operating wavelength in meters
r_T, r_R = magnitudes of ground reflection coefficients at transmitter and receiver, respectively (normal incidence, horizontal polarization)

$$\alpha_T = \rho_T + \frac{4\pi a}{\lambda}$$

$$\alpha_R = \rho_R + \frac{4\pi h}{\lambda}$$

ρ_T, ρ_R = phase angles of ground reflection coefficients at transmitter and receiver, respectively (normal incidence, horizontal polar)
h = height of receiving antenna above ground.

As an illustration of the magnitude of the field intensity to be expected, it is found that in a typical case in

[5] G. W. Pierce, "Electric Oscillations and Electric Waves," McGraw-Hill Book Co., Inc., New York, N. Y., First Edition, 1920, Chap. IX.

which $I_0=302$ amperes and $d=180$ km (round trip distance to ionosphere), the peak field intensity at the receiving dipole calculated from (3) is 340 μv per meter. It is of interest to note that, for an antenna which is elevated only a small fraction of a wavelength above ground, as is the case here, the upward field is increased appreciably as the soil conductivity is reduced.

The practical measurement of the peak field intensity of the reflected pulses may be carried out by calibrating the receiver in terms of the total field present at the antenna. This calibration is effected by using the ground pulse as a standard field[6] in which the dipole is oriented at a known angle, with respect to the electric vector of the ground wave.

Under favorable conditions, a loop antenna may be used directly to measure the strength of the downcoming echoes. Usually, however, the response of a loop to noise and interfering signals is too great to permit such measurement.

Receiver

The receiver used in these experiments is a simple TRF type having three tuned circuits (not counting the tuned antenna). The dipole transmission line is coupled to the receiver through an electrostatically shielded transformer, the over-all bandwidth of which is about 25 kc.

The pulse response of the receiver is best described by the oscillogram of Fig. 5, which shows the form of the 75-μs ground pulse as it appeared on the indicator. The negative pips are timing markers spaced $33\frac{1}{3}$ μs apart. The pulse rise time is seen to be about 30 μs and the duration measured to one-third peak amplitude is about 150 μs, or twice the input time constant. The wideband of the detector and following circuits permits some ripple of twice the carrier frequency to appear on the output pulse. This ripple is actually quite useful, since it serves as a time scale for measuring pulse rise time more accurately than is possible with the 5-km markers.

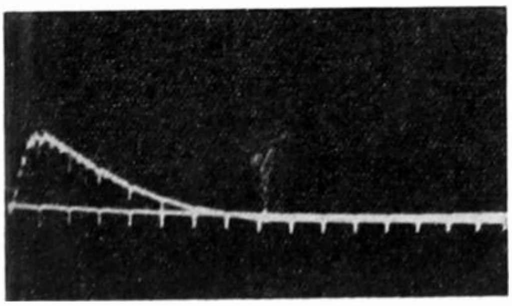


Fig. 5—Ground pulse, $T=75$ μs, 5-km markers.

[6] The field intensity of the ground wave may be measured approximately with a loop-type intensity meter in which the induced pulse is matched by a similar inserted pulse of known peak amplitude. The approximation is that the effective height of the loop is assumed the same for the sideband components of the pulse as for the carrier, although it actually varies directly with frequency for a small loop.

III. Results

Virtual Heights

Apparatus of the type described in previous sections has been used to measure the virtual heights of the night-time E layer of the ionosphere at a carrier frequency of 100 kc. These measurements, made on several occasions from the fall of 1947 to the summer of 1948, show certain interesting effects which will be described.

A rather unusual oscillogram was selected to illustrate the measurement technique, and is shown in Fig. 6. It is unusual in that it is easy to explain. The effective beginning of the ground pulse coincides with the start of the trace and the timing markers. The limiter in the receiver flattened the top of the ground pulse and thus prevented the otherwise excessively strong ground pulse from carrying the timing markers out of the picture. The virtual height or time delay of the first reflection is measured between corresponding parts of the ground pulse and the reflection. The simplest procedure is to extrapolate to the base line the linear part of the leading edge of the reflection and measure the time delay from the beginning of the sweep. Accordingly, the time delay of the first reflection is found by counting the number of 30-kc timing markers, which is 20.9. This corresponds to a time delay of 698 μs. To find the total time required for transmission over the almost vertical path (horizontal reflecting layer assumed), we add the time required for transmission of the ground pulse from transmitter to receiver. Since this distance is 2.70 km, the correction is 9.0 μs, assuming ground-wave propagation at the velocity of light. Thus the total time delay was 707 μs. Although the equivalent path in this case is triangular, the base of the triangle is short, compared with the height of the reflecting region. Assuming the triangle to be isosceles, the half angle at the apex is in this case $\sin^{-1} 9/707 = 44$ minutes of arc. It is thus clear that the difference between the hypotenuse and the altitude of the triangular path may be neglected. The virtual height of the one-hop reflection is accordingly found by multiplying the velocity of light by one-half the measured time delay, which gives a value of 106 km. Fig. 6 also shows a two-hop reflection; i.e., a pulse that has made two roundtrips between the earth and ionosphere. Its virtual height, including the base line

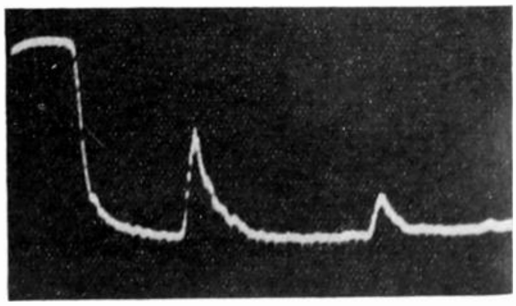


Fig. 6—Single reflection, 2250: 07 PST, July 19, 1948, $T=75$ μs, 5-km markers.

correction, is 210 km. One-half the height of the two-hop reflection should equal the height of the one-hop reflection. In this case, the difference is 1 km, which is the over-all error of measurement in the case of well-defined reflections such as those of Fig. 6. It should be observed that the reflection points, or areas, for two-hop reflection do not coincide with those for one-hop reflection. Although the separation is only 0.7 km in a horizontally stratified layer, it would be greater if the layer were concave downward. When the region is not horizontally stratified, the propagation characteristics of the two-hop reflection in general will not be the same as those of the one-hop reflection. A tilted ground plane will likewise affect the paths followed by a multihop signal. These factors must be considered in attempting to account for occasional differences in the virtual heights obtained from a one-hop reflection and various multihop reflections.

The rise time of both the one-hop and two-hop reflections is about 30 μs, which is the same as the rise time of the ground pulse, while the shape of the tail of the one-hop pulse closely approximates that of the ground pulse in Fig. 5. The slight fore-shortening of the two-hop pulse is believed to result principally from the nonlinear response of the diode detector at low signal levels.

Virtual heights determined in the manner outlined above have been obtained from several thousand oscillograms taken on various nights from October 13, 1947, to July 1, 1948. In cases of splitting (discussed below) the first pulse was used. The median virtual height for each minute during which data were taken is plotted in Fig. 7. The time is indicated in minutes after the Pacific Standard Time of the beginning of each run. Perhaps the most noticeable feature of the data is the irregular character of the variations in height, which ranges from 84 to 106 km. During some periods the height remained practically constant for as long as ten minutes. For example, on the January 28 run, the height remained at 96 km ±1 km from 2108 to 2119 PST. At other times, however, the height varied widely. Thus on October 13, 1947, the height varied between 106 and 93 km in a ten-minute interval. The values which occurred most frequently in the entire group of data were 91 and 96 km, while the average height was 95.2 km.

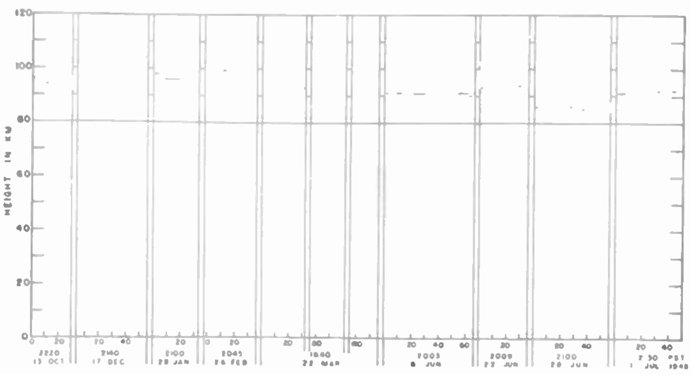


Fig. 7—Nighttime virtual heights at 100 kc.

An interesting variation in virtual height occurred on February 26, 1948. Referring to Fig. 7, one can see that the variation tends to be cyclic with a period of about ten minutes. Possibly it is associated in some way with air currents in the upper atmosphere.

The virtual heights reported here are somewhat higher than those measured by the continuous-wave and lightning-discharge methods described briefly in the introduction. Thus Best, Ratcliffe, and Wilkes[2] found nighttime virtual heights of about 90 km at a frequency of 16 kc using continuous waves over an oblique-incidence path. Others, studying the signals from lightning discharges, found heights ranging from about 60 to 90 km,[3,7] the higher values being obtained at night. The difference may be explained by the fact that lowering the frequency and increasing the angle of incidence reduces the ionization density required for reflection, and hence lowers the virtual height in the case of a single layer.

Field Intensity

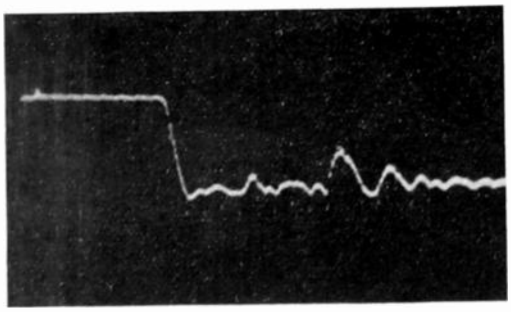
Absolute field intensities were not measured except towards the end of the period, with which this paper is concerned. The field intensities of the one-hop and two-hop reflections appearing on the oscillogram of Fig. 6 were determined by the method outlined in section II, and were found to be 33 and 14 μv/m, respectively, at the dipole receiving antenna. The calculated field intensity assuming unity reflection coefficient at the ionosphere is 340 μv/m. If we assume a plane-stratified ionosphere, we can determine the reflection coefficient of the layer by multiplying the ratio of the amplitude of the two-hop signal to the amplitude of the one-hop signal by two and dividing by the magnitude of the ground reflection coefficient. If we take the ground reflection coefficient to be 0.97, the ionosphere reflection coefficient in the above case is found to be 0.88. This reduces the computed field intensity to 300 μv/m, which is about nine times the measured value. A possible explanation of the large discrepancy between measured and calculated field intensities will be given later on in the discussion of the polarization effect.

Echo Splitting

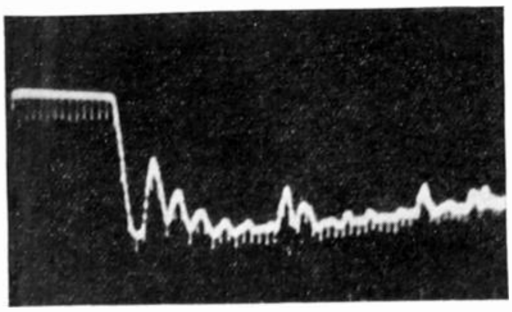
In contrast to the oscillogram of Fig. 6 is the more usual type in which the reflected pulses appear to be split into two or more principal components. An example of the effect is shown in the oscillogram of Fig. 8(a). The first reflected pulse appeared at a virtual height of 91 kilometers. It was followed by a second, weaker pulse at 105 kilometers. A second example of the effect is seen in Fig. 8(b). In this case the pulse time constant was

[7] T. H. Laby, et al., "Reflections of atmospherics by the ionosphere," *Nature*, vol. 142, p. 353; August 21, 1938.

150 μs, twice that of the pulse of Fig. 8(a). The first reflection appeared at 90 km. Its peak was followed by at least three additional components at 15-km intervals. A two-hop group of reflections appeared at about 178 km and shows two distinct peaks separated by about 12 km. A single three-hop reflection appeared at 270 km. Further examples of splitting are afforded by the oscillograms of Figs. 9(a), (b), and (c). In these cases the pulse time constant was again 150 μs, but the receiver bandwidth was roughly seventy per cent of that used to obtain the data of Figs. 8(a) and (b). The timing marker spacing was 10 km, except for the first space which was 11 km. It will be observed that the rise time of the first component in each group was about 45 μs, as compared with 30 μs for the previous cases. The chief difference between these oscillograms and those of Figs. 8(a) and (b) is the relative lack of separation of the different components in each group. This overlapping was accompanied by considerable variability in the shape of the envelope in the space of a few seconds. Thus, in Fig. 9(a), two distinct peaks appeared on the one-hop reflection, and a slight indication of two peaks appeared on the two-hop reflection. In Fig. 9(b), three seconds later, the second peak in the first group had become only a bulge, but the two-hop group exhibited three noticeable peaks, the first of which was about one-half the amplitude of its counterpart in Fig. 9(a). A rather unusual condition is illustrated in Fig. 9(c), in which the one-hop group exhibited three relatively weak peaks, while the second group showed only a single peak of normal amplitude. The separations of these peaks varied between about 10 and 15 km on most of the records that have been examined.

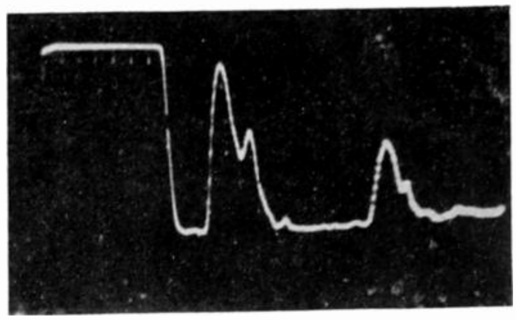


(a)

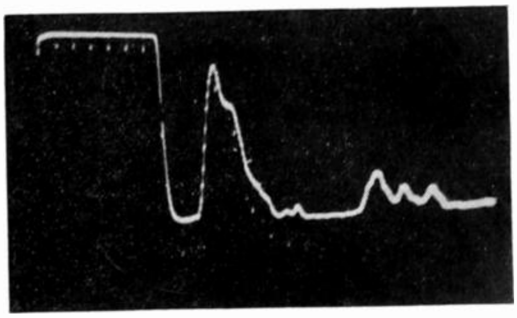


(b)

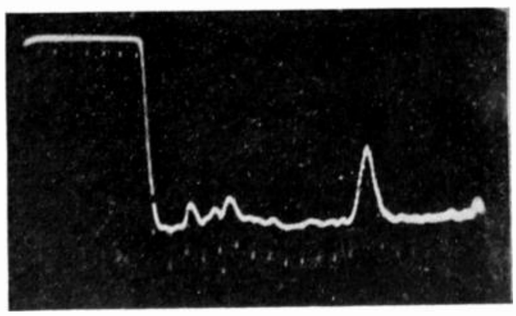
Fig. 8(a)—Split reflection, 2106: 49 PST, June 8, 1948, $T=75$ μs, 5-km markers. (b) Split reflection, 2132: 19 PST, June 8, 1948, $T=150$ μs, 5-km markers.



(a)



(b)



(c)

Fig. 9(a)—Split reflection, 2155: 22 PST, March 22, 1948, $T=150$ μs, 10-km markers. (b) Split reflection, 2155: 25 PST, March 22, 1948, $T=150$ μs, 10-km markers. (c). Split reflection, 2250: 36 PST, March 22, 1948, $T=150$ μs, 10-km markers.

The variation in the shape of the envelope can be explained by assuming the presence of several component pulses adding together with random relative phases to produce the observed result. Why there should be more than one reflected pulse is more difficult to explain. However, the relative constancy of the separations of most of the multicomponent reflections (of which Fig. 8(b) is a good example) suggests the possibility that the lower part of the E region actually consists of a number of patchy, nonuniform layers, each of which contains "holes" large enough to permit some energy to penetrate (at nearly vertical incidence) to patches of ionization in the next higher layer. Another explanation is that the reflections of longer delay come from patches of ionization which are at about the same height, but which are not directly overhead. Further experiments aimed at resolving this uncertainty are planned.

Polarization

One of the most interesting and consistent effects observed during the course of these experiments has been the rotation of the plane of polarization of the downcoming pulses, with respect to the plane of polarization of the transmitted pulse. The polarization of the signal transmitted vertically upwards will, of course, be in the direction of the horizontal part of the transmitting antenna, which is oriented fifteen degrees west of true north. At the receiver, located 2.7 km away in a direction 45 degrees west of north from the transmitter, maximum signal strength is obtained with the dipole antenna oriented about ninety degrees from true north. Furthermore, the signal strength with the dipole at right angles to this position is usually zero or a small fraction of the peak intensity, indicating that the polarization is linear, or very nearly so.

One possible explanation of the observed rotation is that the component of signal at right angles to the direction of maximum signal strength is not reflected, or is highly absorbed, while that in the direction of maximum signal strength is reflected with relatively small loss. In order to bring the calculated signal strength into agreement with that measured from Fig. 6, it is necessary for the component of current producing the observed signal to be one-ninth of the total current. Thus the direction of the polarization of the received signal would have to be at an angle of six degrees ($\sin^{-1} 1/9$) with respect to a normal to the transmitting antenna. There are, of course, two such directions separated by twelve degrees. One corresponds to the direction sixty-nine degrees west of south and the other to the direction eight-one degrees west of south.

The variation of field intensity with dipole azimuth was not obtained at the time the oscillogram of Fig. 6 was made. However, approximate measurements of the direction of polarization made at other times indicate that the average direction is east-west with an estimated variation of plus or minus ten degrees. The required direction of eight-one degrees lies within this range, and hence a check is provided on the explanation given above, within the limits of error of measurement, which, in the case of the polarization directions, are appreciable.

In terms of the magneto-ionic theory, one would expect the propagation to be of the "ordinary" component requiring considerably less ionization density for reflection than that required by the "extra-ordinary" component. Even with a relatively sharp ionization gradient, some difference in time delay between the two components would be expected. It would thus be possible for the polarization of the sum of the two reflected components to become more or less elliptical, with the major axis shifted with respect to the direction of the incident plane-polarized wave.

The detailed correlation of the experimental results with the theory will be postponed until more data are available.

IV. Conclusions

This experimental investigation has shown that the minimum height of the lower region of the nighttime E layer is far from constant, varying all the way from 84 to 106 km during the course of nearly a year of intermittent measurements. If low-frequency transmission problems are to be solved on the basis of virtual height data, as is done at high frequencies, then measurements similar to those described in this report must be made at many different locations and at several frequencies to determine what variations in height may be expected.

Some evidence has been found to indicate that frequently at night the region of reflection of 100 kc waves consists of clouds or patches of ionization, rather than the more nearly uniform ionization characteristic of the regular layers at higher frequencies. If this be the case, the relation between oblique and vertical-incidence transmission (sometimes called the "equivalence theorem"), which holds for a horizontally stratified layer, cannot necessarily be expected to hold at low frequencies at night. It is therefore highly desirable that studies be made to determine the nature of the relation at low frequencies.

The reflection of 100-kc waves is apparently accompanied by changes in polarization. Since the details of this effect have not been elucidated, it is important that further experiments be conducted for this purpose, both at vertical incidence and at oblique incidence.

V. Acknowledgments

The author wishes to acknowledge the co-operation and financial support of the Central Radio Propagation Laboratory of the National Bureau of Standards in carrying out these measurements.

The author is grateful for the generous aid received during the course of the research from members of the staff of the department of electrical engineering, Stanford University. Particular mention s ould be made of Joseph S. Carroll and William G. Hoover, whose help in the design and construction of the high-voltage equipment was invaluable. The author is likewise indebted to Frederick E. Terman, Dean of the School of Engineering; Hugh H. Skilling, Executive Head of the department of electrical engineering; and Karl Spangenberg, for their helpful suggestions. Thanks are due to Laurence A. Manning and Oswald G. Villard, Jr., for their many stimulating comments. Mention should also be made of the enthusiastic participation of the several members of the project staff who were engaged in various phases of the work.

The Demodulation of a Frequency-Modulated Carrier and Random Noise by a Discriminator*

NELSON M. BLACHMAN†

WITH THE AID of an idealized representation of the discriminator, Middleton[1,2] has shown that a frequency-modulated carrier remains intelligible for smaller carrier strengths when narrow-band frequency modulation is used and the limiter is omitted than with a limiter and/or wide-band frequency modulation. It is, therefore, of some interest to treat the demodulation process as it *actually* occurs, in the absence of a limiter. Thus, the discriminator has been taken to consist of two selective circuits, both fed by the output of the intermediate-frequency amplifier, but peaked at different frequencies, feeding rectifiers whose outputs are subtracted. Each half of the device is treated in the manner used by Rice[3-5] to determine the result of passing random noise and a sine wave through a rectifier; however, there is a correlation between the noise voltages fed to the two rectifiers.

The signal output and the spectral distribution of the noise output are obtained first for quadratic rectification, then in the general case, which is specialized then to linear rectification. The results are applied to a case of rectangular if noise spectrum, and the signal-to-noise ratio is determined for the cases of narrow-band and wide-band FM.

These results are found to be very much like those for the idealized representation of the discriminator;[6] all are compared, along with amplitude modulation,[7-10] in Table I. The optimum signal-to-noise ratio for FM without a limiter is found to obtain with narrow-band FM when the discriminator is designed for no wider a band than necessary; this optimum signal-to-noise ratio differs very little from that for amplitude modulation.

TABLE I

COMPARISON OF MEAN-SQUARE SIGNAL-TO-NOISE RATIOS FOR FM WITHOUT A LIMITER AND AM

System	Demodulator	Weak Carrier ($x \ll 1$)		Strong Carrier ($x \gg 1$)	
		Narrow Band ($f_a \gg F$)	Wide Band ($f_a \ll F$)	Narrow Band ($f_a \gg F$)	Wide Band ($f_a \ll F$)
FM	Quadratic Discriminator	R^2x^2	$\frac{1}{2}(F/f_a)R^2x^2$	$\frac{2}{3}Rx$	$\frac{1}{2}(F/f_a)Rx$
	Linear Discriminator	$2R^2x^2$	$(F/f_a)R^2x^2$	Rx	$2(F/f_a)Rx$
	Idealized Discriminator	$\frac{1}{4}\pi x^2$	—	x	$2(F/f_a)x$
AM	Quadratic Detector	$2x^2$	$2(F/f_a)x^2$	$\frac{2}{3}x$	$\frac{2}{3}(F/f_a)x$
	Linear Detector	$1.83x^2$	$1.83(F/f_a)x^2$	x	$(F/f_a)x$

$2F$ = if bandwidth, f_a = audio upper cutoff frequency, x = carrier-to-noise ratio $\propto 1/F$, deviation ratio $= 3-\frac{1}{2}F/f_a$ for FM, modulation index = 100 per cent for AM, $1 < R < 4/3$.

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† Formerly, Cruft Laboratory, Harvard University, Cambridge, Mass.; now, Brookhaven National Laboratory, Upton, L. I., N. Y.

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Discussion on

"Application of Velocity-Modulation Tubes for Reception at U.H.F. and S.H.F."*

M. J. O. STRUTT AND A. VAN DER ZIEL

E. Barlow:[1] Strutt and van der Ziel have given an expression for the minimum noise figure obtainable with a particular adjustment of a three-resonator klystron amplifier. They have, using their notation,

$$N_{\min} = \frac{eI_0F^2}{2kT} R_s\left[\frac{S_{23}^2}{S_{13}^2 + S_{12}^2S_{23}^2R_S^2}\right]. \qquad (14)$$

With the assumptions used by Strutt and van der Ziel, this equation should read, instead,

$$N_{\min} = \frac{eI_0F^2}{2kT} \frac{R_s(S_{12} - S_{13})^2}{S_{13}^2 + S_{12}^2S_{23}^2R_s^2}.$$

The two equations give the same result if the transconductances are calculated from simple bunching theory (a condition not implicit in the derivation of (14)) since, then,

$$S_{13} = S_{12} + S_{23},$$

but if any space-charge debunching is taken into account they are no longer equivalent. This opens up the possibility of getting a zero noise figure (as far as shot noise is concerned). Suppose simple debunching theory be considered, so that the transconductances take the form

$$S_{12} = \frac{1}{2}\frac{I_0}{V_0}\frac{\omega}{v_0} \sin\frac{(hx_{12})}{h}, \text{ etc.,}$$

where h is the debunching wave number. It is now no longer true that

$$S_{13} = S_{12} + S_{23},$$

and if, for example,

$$\sin (hx_{13}) = \sin (hx_{12})$$

and

$$x_{12} \neq x_{13},$$

a zero noise figure is indicated by the correct form of (14). For example, if

$$hx_{13} = 2\pi/3,$$

$$hx_{12} = \pi/3,$$

$$hx_{23} = \pi/3,$$

the conditions are satisfied.

A fundamental assumption of Strutt and van der Ziel's paper is that there is complete coherence of the noise fluctuations along the electron beam. It is of interest to investigate how closely this assumption is met in practice. In the example given by the authors, an improvement of 100/1 was obtained by noise cancellation. In order to achieve this performance a "coherence" of 99 per cent is required, and, of course, even more stringent requirements are imposed if better noise cancellation than this is desired. To formulate the coherence requirements more exactly, a correlation coefficient can be introduced. If the noise currents at two points along the electron beam are designated as i_1 and i_2 such that

$$\overline{i_1} = \overline{i_2} = 0$$

and

$$\overline{i_1^2} = \overline{i_2^2},$$

then

$$\overline{(i_1 - i_2)^2} = 2\overline{i^2}(1 - \rho_{12})$$

where ρ_{12} is the correlation coefficient,

$$\rho_{12} = \overline{i_1 i_2}/\overline{i^2}.$$

This means that, for 99 per cent cancellation of noise currents,

$$\frac{\overline{(i_1 - i_2)^2}}{\overline{i^2}} = 2(1 - \rho_{12}) = 0.01,$$

and a value of $\rho_{12} = 0.995$ is required for the author's example.

Some of the factors which tend to lower the noise correlation in klystron amplifiers are the following:

1. Space-charge debunching
2. Variation of beam coupling coefficient across the electron beam
3. Partition noise
4. The thermal velocity spread of the electrons.

Debunching of the electron beam due to space-charge forces occurs to some extent in all klystron designs, and should be taken into account. If this effect varies across the electron beam, the current-density fluctuations will lose coherence, since those adjustments of the klystron gain, resonator impedance, etc., necessary to cancel fluctuations on the edges of the electron beam will not cancel those near the center, and vice versa. The debunching of the beam depends on beam current and voltage, beam diameter and drift-tube diameter, and on the distance traversed down the beam axis. If the

* M. J. O. Strutt and A. van der Ziel, "Application of velocity-modulation tubes for reception at u.h.f. and s.h.f.," Proc. I.R.E., vol 36, pp. 19–24; January, 1948.

[1] Formerly, Sperry Gyroscope Company, Great Neck, L. I., N. Y.; now, Rand Corporation, Santa Monica, Calif.

beam is of large section relative to $\lambda v/c$ (the "electron wavelength) the debunching near the center of the beam can be characterized by

$$\overline{i_2^2} = \overline{i_1^2}(\cos hx_{12})^2$$

where i_1^2 is the mean square noise current at the entrance to the klystron drift tube, $\overline{i_2^2}$ is the mean square noise current at a distance x_{12} down the electron beam, and h is the "debunching wave number" given by

$$h = \frac{173 I_0^{1/2}}{a V_0^{3/4}}$$

where

$$a = \text{beam radius.}$$

At the edge of the beam the image charges along the drift tube wall almost neutralize the debunching forces, so that, in this case, roughly

$$\overline{i_2^2} \cong \overline{i_1^2};$$

thus the correlation between the total noise currents at positions 1 and 2 can be roughly estimated as

$$\rho_{12} \cong 1 - \left(\frac{1 - \cos hx_{12}}{4}\right)^2.$$

For the amplifier cited by the authors, we can take

$$x_{12} = 10 \text{ cm}$$

$$h = \frac{173\sqrt{0.01}}{a(1000)^{3/4}} = \frac{1}{10a} \ (a \text{ in cm}).$$

A reasonable value for a would be 0.5 cm. This yields a value of

$$\rho_{12} = 1 - \left(\frac{1 - (\cos 2)}{4}\right) \cong 0.88.$$

Since, as was pointed out, a correlation coefficient of not less than 0.995 was needed to secure the performance required by the authors, the introduction of debunching theory brings the efficacy of the proposed klystron design into question. Of course, the preceding analysis of the effects of debunching is approximate since the debunching in the center of the beam is somewhat less than that indicated, while there will be some debunching along the edge of the beam, especially if the beam diameter is much less than the drift-tube diameter. Of course, the beam must not be made too small or the drift tube too large, or the debunching will be excessive, the beam coupling coefficients to the resonators will be low, and the effective resonator shunt resistances will be low. This points up the conclusions to a study made of low-noise-klystron amplifier design at the Sperry Gyroscope Company. Almost anything done to improve the coherence of the noise along the beam acts to lower the klystron gain or increase the noise figure before compensation, so that very little performance increase is obtained. A study has been made of debunching, and rigorous results including the effects of finite beam diameter and the presence of the drift tube wall have been reached by E. Feenberg. From these results the lack of coherence due to debunching can be estimated, and it turns out to be less serious than the previous rough estimate, but sufficient to make noise cancellation schemes such as that proposed by Strutt and van der Ziel of questionable use.

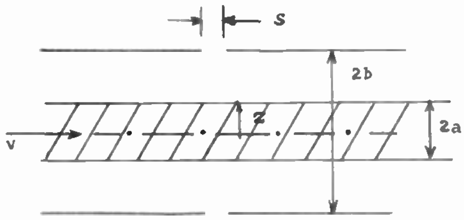

Fig. 1

An analogous effect which must be taken into account is the variation of the beam coupling coefficient across the electron beam. If gridless resonators are used, this coefficient varies as shown in Fig. 1.

$$\mu(z) = \frac{\sin \frac{\omega s}{2v}}{\frac{\omega s}{2v}} \left[\frac{J_0(jk'z)}{J_0(jk'b)}\right]$$

where

$$k' = \frac{2\pi c}{\lambda v}.$$

If a differential current element dI is considered and the equations of the klystron amplifier applied, a relation will be found for the minimum noise figure, and in that relation will appear

$$\mu_1(z_1), \qquad \mu_2(z_2)$$

and

$$\mu_3(z_3),$$

the beam coupling coefficients the current element experiences in traversing the three gaps. In order to be able to cancel the noise fluctuations of all these filaments with any one klystron adjustment, it is necessary that the ratios

$$\frac{\mu_1(z_1)}{\mu_2(z_2)} \quad \text{and} \quad \frac{\mu_1(z_1)}{\mu_3(z_3)}$$

be the same for all filaments within roughly 2 per cent in the example cited. This puts very stringent requirements on the electron beam and the resonator-gap design. Transverse thermal velocities have been estimated to be sufficient to invalidate this requirement for a typical tube. Any irregularities in the electron-gun structure or space-charge beam spreading, particularly with nonuniform current distribution across the electron beam, will be deleterious. If fine grids are used, the beam coupling coefficient will be more nearly uniform than with gridless gaps, but will still vary enough to cause trouble and will introduce unwanted partition noise.

There are two distinct effects which are associated with the interception of part of the beam current by the klystron grids.

First, there is "partition noise" in the sense that, if the original beam had shot noise which was reduced by space-charge smoothing, the introduction of a grid intercepting current would partially nullify this smoothing, so that the rms fluctuations after the grid would be greater than those before it. This additional noise would be called partition noise. If the original beam has full shot effect, the partition noise in this sense is negligible.

Second, even though the rms noise output is not increased by the addition of a grid, there will be an element of noncoherence introduced between the noise fluctuations before and after the grid if current is intercepted by it. It is this latter effect which is important in studying the low-noise-klystron problem. Consider a three-resonator klystron using grids as shown in Fig.2. Any current interception on grids 2, 3, 4, or 5 or on the

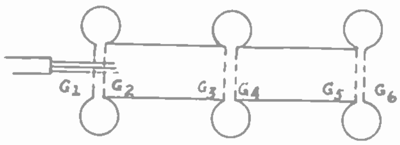

Fig. 2

drift-tube walls will reduce the coherence between the noise currents in the first and third resonators. Assuming full shot-effect fluctuations on the electron beam, the total intercepted current must be $\frac{1}{2}$ per cent or less to keep ρ_{12} above 0.995. Even if there is no current intercepted by the drift-tube walls, the interception per grid in the above example must be less than $\frac{1}{8}$ per cent, while the best figures obtained for fine grids are more like 2 or 3 per cent. This would yield a value of ρ_{12} of roughly 0.90. The authors' example was a klystron having a noise figure, before compensation, of 4,000. With the above figures for grid interception, this could not be reduced below about 400 or 26 db. Strutt and van der Ziel mention that intercepted noise currents affect their conclusions, but state that "if F^2 is near unity . . . no appreciable increase of mean square output current fluctuations will result from these partition fluctuations." This is not the important point; it is the lack of coherence introduced by partition currents which must be considered, as has been pointed out above. The authors further state that "special methods of compensation may be applied, reducing these partition fluctuations to a relatively negligible fraction of over-all mean-square output fluctuations." When these special methods are investigated, it turns out that there is some question as to whether they can be applied to velocity-modulation tubes. They are concerned with elimination of the wrong effect of "partition noise." Consider the case shown in Fig. 3. Let the dc cathode current be

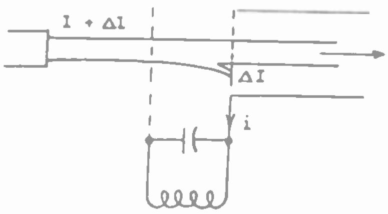

Fig. 3

$I+\Delta I$, and let the current I pass through the grids, while the component ΔI is intercepted by grid No. 2. If the noise current i represented only the noise fluctuations of the intercepted current ΔI, it might perhaps be of some use for compensation, but of course the whole idea of velocity-modulation tubes at microwave frequencies is that the current I is closely coupled to the tuned circuit and i represents the fluctuations in both I and ΔI, and, hence, the noise fluctuations in ΔI cannot be isolated. If this were not so, the beam coupling coefficient would be identically zero, and grids Nos. 1 and 2 might as well be removed. A further point is that, even if the noise fluctuations in ΔI could be isolated, there is a question as to where a compensating voltage should be applied. If the compensating voltage is applied to grid No. 1, a density modulation results, but that in turn introduces an additional noise current present in both resonator No. 1 and resonator No. 2, so that the coherence has not been improved. If the compensating voltage is applied to grid No. 2, a velocity modulation is produced, and one is then in the position of trying to compensate a density fluctuation with a velocity fluctuation.

The thermal distribution of forward velocity of the electrons leaving the cathode will cause some loss of correlation, as pointed out by the authors. This effect can be evaluated in terms of the correlation coefficient. Assuming full shot effect, and neglecting space-charge forces in the drift tube, the correlation coefficient is given by

$$\rho_{12} = \frac{1}{\sqrt{1 + \left(\frac{1}{2} k' x_{12} \frac{kT_c}{eV_0}\right)^2}} .$$

This expression includes both the effects of thermal debunching of density fluctuations and bunching of initial velocity fluctuations. For correlations near unity, this can be approximated by

$$\rho_{12} \cong 1 - \frac{1}{2}\left(\frac{1}{2} k' x_{12} \frac{kT_c}{eV_0}\right)^2 ,$$

and for

$$k'x_{12} = 100,$$

$$\frac{kT_c}{eV_0} = 10^{-4},$$

$$\rho_{12} = 1 - 1.2 \times 10^{-5}.$$

For these values, the thermal velocity spread causes negligible effect. This agrees with the discussion of Strutt and van der Ziel on this point. If a klystron amplifier is designed for low-noise performance by lowering the beam voltage and increasing the drift-tube length, however, it is quite possible that the thermal velocity spread will have to be taken into account. In one such design, a value of $\rho_{12}=0.95$ was found.

In conclusion, it is felt that any scheme such as that suggested by the authors employing noise cancellation demands a careful investigation of noise coherence, and that in particular limitations imposed by (1) space-charge debunching, (2) variations of beam coupling coefficient, and (3) partition noise must be carefully investigated. The experimental investigation of noise cancellation made by the writer and referred to by Strutt and van der Ziel showed correlation coefficients considerably below unity (0.6 to 0.9), and it is felt that this result is in line with the foregoing discussion, since space-charge debunching and partition noise both existed in the klystron studied.

Max J. Strutt[2] and A. van der Ziel[3]: Mr. Barlow[1] has discussed our paper on noise figures of the three-resonator klystron, and has given an interesting contribution to the question of whether the current fluctuations along the beam are completely coherent.

In our paper we showed that reasonably low noise figures could be obtained with the three-resonator klystron, if the following assumptions are valid: (1) simple bunching theory can be applied, and (2) complete coherence exists for the current fluctuations along the beam. The second condition was stated explicitly, and it follows from our calculations that the first condition was introduced implicitly. We shall now show that, if the first condition is invalid, the second condition is violated also.

Simple bunching theory is not sufficient, if space-charge debunching, transit-time debunching, or variation of beam coupling along the beam must be taken into account. As shown by Barlow, and as shown for transit-time debunching in our paper, the fluctuations along the beam are not completely coherent in these cases, so that the second assumption also does not hold. As our equations are valid only if the second condition is satisfied, it is useless to correct our equation (14) as was done by Barlow. For, if simple bunching theory can be applied, the corrected equation is equal to our equation, whereas both equations do not hold if the first condition, and hence the second one, are not satisfied; therefore, the conclusion of zero noise figure mentioned by Barlow is of doubtful value.

[2] Swiss Federal Institute of Technology, Zurich, Switzerland.
[3] Physics Department, University of British Columbia, Vancouver, B.C., Canada.

In one of our examples we derived a theoretical improvement of the noise figure by a factor of 100, and although Mr. Barlow does not deny that considerable improvement can be obtained, he has considerable doubt whether an improvement by a factor of 100 is possible. Although we consider Mr. Barlow's discussion very valuable, it still does not give a sufficiently clear indication, of what will be the lowest limit that can be obtained for the noise figure.

In general, we agree with the way in which Mr. Barlow discusses the influence of space-charge debunching and of variation of beam coupling. We think, however, that he overestimates the influence of partition noise. Mr. Barlow mentions that the partition noise due to the current flowing to the grids G_2, G_3, G_4, and G_5 (and to the wall of the drift tube) has to be taken into account. We shall now show, that only the currents to the grids G_2 and G_3 contribute appreciably to the noise figure.

In order to do so, we turn to equation (19) of our paper. After (13) and (16), this holds under the conditions that

$$(S_{23}R_s)^2 \gg 1; \qquad C_2 = 0.$$

Under these conditions, we have, for C_1,

$$\frac{1}{\omega C_1} = \frac{1}{S_{12}}; \quad \text{or, since} \quad Z_1 = (j\omega C_1)^{-1},$$

$$1 - jS_{12}Z_1 = 0.$$

We obtain (19) also if we introduce this relation into (10). For, bearing in mind that $S_{13}=S_{12}+S_{23}$, we have, after (10) and (16),

$$TN_{\min} = N_0 \left| \frac{1 - \dfrac{S_{13}}{S_{12}}}{jS_{23}Z_2} \right|^2 = N_0 \frac{1}{S_{12}^2 R_s^2},$$

which is the same as equation (19).

This gives the following simple picture of our noise-reducing circuit: The noise figure of a normal klystron amplifier is due to the noise voltage induced in the input circuit by the beam current. In order to improve the noise factor, this voltage must be reduced. This is done with our precircuit, and, after (5), the noise voltage across the input circuit is just equal to zero if the condition $(1-jS_{12}Z_1)=0$ is satisfied. If $S_{12}\neq S_{23}$, the condition for minimum noise is slightly different from the condition of zero noise voltage across the input circuit, but at any rate the major part of the noise reduction is still obtained.

Therefore, it is not very important if some current flows to the grids G_1, G_4, G_5, or G_6, because that hardly increases the noise voltage across the input circuit, especially if the beam current is saturated. The currents flowing to the electrodes G_2 and G_3 are the only ones which have to be taken into account, because these currents give rise to partition noise which is uncorrelated to the noise voltage induced in the precircuit, and hence this partition noise will induce a noise voltage in the input circuit, which cannot be compensated by the precircuit. The best solution would be to cut down the currents to the electrodes 2 and 3 as much as possible.

In discussing partition noise in velocity-modulated tubes at microwave frequencies, two facts have to be borne in mind:

1. The beam current is closely coupled to the tuned circuits.
2. The beam current only contains partition noise due to the current flowing to a grid, *after* the beam has passed that grid.

If, for example, I_1 is the beam current before grid G_2 is passed and I_2 the beam current after G_2 is passed, then the noise current flowing through the circuit between grids 1 and 2 does not contain the partition noise generated by the current to G_2, at any rate not if the circuit is of the cavity type. This point is not made sufficiently clear by Mr. Barlow. It is difficult indeed to design a circuit which compensates the partition noise due to the currents flowing to G_2 and G_3, the best way is to cut down these currents to a low value.

In conclusion, we think it to be still worth while to investigate what the lowest possible limit will be for the noise factor of the above circuit, although we admit that Mr. Barlow's measurements give some indication that this lowest limit might still be too high.

But even if this were the case, the circuit might still be applied to the traveling wave tube. Measurements given by Koempfner indicate that the noise figure of a traveling wave tube is much lower than for a klystron amplifier. For the traveling-wave tube, the noise will be chiefly due to the noise induced by the electron beam into the beginning of the helix. By using a precircuit, the noise figure might easily be decreased to such a low value that it becomes lower than the noise of a crystal mixer. One of the advantages of the traveling wave tube, the very wide band amplification, would have to be sacrificed, in that case; but the low noise figure would still make it valuable. It would seem worth while to try the circuit described in our paper in this application as well.

Contributors to the Proceedings of the I.R.E.

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Mr. Beaumariage is a member of Tau Beta Pi, Eta Kappa Nu, Sigma Xi, and the American Institute of Electrical Engineers.

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Contributors to the Proceedings of the I.R.E.

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J. P. Eckert, Jr.

Following this, he was engaged with P. S. Grumaer in the design and construction of a device for measuring the concentration of naphthalene vapor by means of ultraviolet light, for the Barrett Division of the Allied Chemical and Dye Company. During this time, Mr. Eckert was also working with John C. Sims, Jr., on the development of equipment designed to measure fatigue limits in metals for the Southwark Division of the Baldwin Locomotive Works. With the completion of his work for Baldwin's, Mr. Eckert undertook the designing and construction of a device for measuring the strength of very small magnetic fields. He was also engaged in the solving of a number of radar problems, and worked on various devices for measuring targets to an accuracy of two yards out of a hundred thousand yards. He spent some time on problems involving the Differential Analyzer at the Moore School, and aided in the improvement of torque amplifiers (servomechanisms) and curve followers for the Analyzer.

In June, 1943, Mr. Eckert received the M.S.E.E. degree from the Moore School, after which he became chief engineer for the ENIAC project. In 1945 Mr. Eckert held the same position for the design and construction of the EDVAC, the successor to the ENIAC. The following fall, Mr. Eckert and Dr. Mauchly formed the Electronic Control Company, now known as the Eckert-Mauchly Computer Corporation.

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Ralph W. Engstrom

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Max Fishman (S'42–A'49) was born in Haverhill, Mass., on April 30, 1921. He received the B.S. degree in 1942 and the M.S. degree in 1943 from Carnegie Institute of Technology, both in electrical engineering. He entered the U. S. Navy in 1944 and served until August, 1946. While in the Navy he was an instructor at the U. S. Naval Academy at Annapolis, after returning from a tour of sea duty on the USS *Gardiners Bay*, AVP-39. Leaving active duty with the rank of lieutenant (junior grade), he returned to Carnegie Institute in September, 1946, and received the D.Sc. in electrical engineering in 1948. Dr. Fishman is a member of Sigma Xi and the American Institute of Electrical Engineers.

Max Fishman

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G. R. Harris was born in Pueblo, Colo., on March 31, 1907. He was graduated from Carnegie Institute of Technology in 1927 with the B.S. degree in electrical engineering, after which he was employed by the Bell Telephone Laboratories, Inc. In 1932 he became associated with Jones and Laughlin Steel Corporation, Pittsburgh, Pa., serving in various technical capacities. At present he is a patent attorney for that corporation, having been graduated from Duquesne University Law School in 1941.

G. R. Harris

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Robert A. Helliwell (S'41–A'45) was born in Red Wing, Minn., on September 2, 1920. He received the degrees of A.B. in 1942 A.M. in 1943, E.E. in 1944, and Ph.D. in 1948, all from Stanford University.

R. A. Helliwell

Since 1942, Dr. Helliwell has been on the staff of the department of electrical engineering of Stanford University where he now holds the position of acting assistant professor. In addition to teaching, he has been engaged in government-sponsored ionosphere research carried out in co-operation with the National Bureau of Standards. During the war he worked on a study of the correlation of direction-finder errors with ionospheric conditions.

Dr. Helliwell is a member of Phi Beta Kappa, Sigma Xi, Tau Beta Pi, and Commission 3 of the URSI.

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Arnold R. Moore was born in New York, N. Y., on January 14, 1923. He received the B.S. degree in chemistry in 1942 from the Polytechnic Institute of Brooklyn. From 1942 to 1945, and again during the summer of 1941, he was a research and development engineer at the Lancaster, Pa., Laboratory of the RCA Victor Division, Radio Corporation of America.

Arnold R. Moore

In 1945 Mr. Moore enrolled at Cornell University to work toward the Ph.D. degree in physics. While at Cornell he has been successively a teaching assistant, research assistant, and RCA Fellow in electronics. He is a member of Phi Kappa Phi, Sigma Xi, and the American Physical Society.

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For photographs and biographies of K. Spangenberg, G. Walters, and F. Schott, see page 780 of the July, 1949, issue of the Proceedings of the I.R.E.

Contributors to the Proceedings of the I.R.E.

Winston E. Kock (SM'45) was born at Cincinnati, Ohio, on December 5, 1909. He received the E.E. and M.S. degrees at the University of Cincinnati in 1932 and 1933, respectively, followed by the Ph.D. degree in 1934 from the University of Berlin. After a year as Teaching Fellow at the University of Cincinnati, he continued graduate study at The Institute for Advanced Study at Princeton, and at the Indian Institute of Science in Bangalore, India. Following several years as Director of Electronic Research at the Baldwin Piano Company, he joined the Radio Research Department of the Bell Telephone Laboratories at Holmdel, N. J., in 1942, where he first engaged in microwave radar antenna research, and later conducted research on lens antennas for radio relay circuits. In January, 1948, he entered the transmission research department of the Bell Laboratories at Murray Hill, N. J., as research engineer in charge of acoustics.

Winston E. Kock

Dr. Kock is a Fellow of the American Physical Society, a Fellow of the Acoustical Society, and a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

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Robert F. Shaw (S'35–A'43–M'45–SM'47) was born at Warsaw, Ohio, on July 10, 1915. He received the A.B. degree in physics from Princeton University in 1937, after which he did graduate work in electrical engineering at the Moore School of the University of Pennsylvania. He taught radio theory for the U.S. Navy in 1942 and 1943, and in the latter year he returned to the Moore School as a research engineer connected with the development of the ENIAC (electronic numerical integrator and computer). He remained there until completion of the project, and carried out preliminary logical design work on the EDVAC (electronic discrete variable computer).

Robert F. Shaw

In 1946, Mr. Shaw joined the staff of the Electronic Computer Project at the Institute for Advanced Study, and later that year became a member of the Electronic Control Company, now known as the Eckert-Mauchly Computer Corporation. His work has been concerned mainly with logical design of computers and associated equipment.

Mr. Shaw is a member of the American Association for the Advancement of Science, the Association for Computing Machinery, the Franklin Institute, and Sigma Xi.

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C. Bradford Sheppard (S'40–A'44–M'45) was born in Camden, N. J., on October 25, 1918. He received the B.S. degree in electrical engineering in 1941 from the University of Pennsylvania, at which time he was awarded a fellowship at that University. He obtained the master's degree in electrical engineering in 1942, during which year he did part-time teaching in EMSDT courses, and worked on a Government-sponsored project on antenna field patterns at the University of Pennsylvania. From 1942 to 1946 he was employed by Hazeltine Electronics Corp., first in their antenna design group, and later on test equipment for the Mark V IFF System. In 1946 he returned to the Moore School as a research associate to work on the EDVAC electronic computer.

C. B. Sheppard

Mr. Sheppard left the University of Pennsylvania in 1947 to join the then newly formed Electronic Control Company, currently known as the Eckert-Mauchly Computer Corp. Here he has worked on acoustic delay and electrostatic storage memory devices, and other electronic computing equipment. He is a member of the American Physical Society, the Association for Computing Machinery, and the American Association for the Advancement of Science.

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D. W. Ver Planck was born at Swampscott, Mass., on January 29, 1906. He received the B.S. and M.S. degrees in electrical engineering in 1928 and 1929, respectively, from the Massachusetts Institute of Technology. From 1929 until 1936 he was employed by the General Electric Company. In 1936 he became assistant professor of electrical engineering at Yale University, where he earned the degree of doctor of engineering in 1940, his research being on spark discharges in air. From 1940 until 1942, Dr. Ver Planck was engaged in the development of degaussing of ships at the Naval Ordnance Laboratory in Washington, D. C. From 1942 until 1946 he was an officer in the Naval Reserve in charge of degaussing in the Navy Bureau of Ordnance. He became professor of electrical engineering at Carnegie Institute of Technology in 1946, and continued in that capacity until until 1947, when he was made George Tallman Ladd Professor of Engineering and head of the department of mechanical engineering.

D. W. Ver Planck

Dr. Ver Planck is a member of the American Institute of Electrical Engineers, the American Society of Mechanical Engineers, the American Society for Engineering Education, and Sigma Xi. For his work in the Navy he was awarded the Navy Distinguished Civilian Service Award and the Navy Commendation Ribbon, and was also made an Honorary Member of the Order of British Empire (Military Division).

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Everard M. Williams (S'36–A'41–SM'44) was born in New Haven, Conn., in 1915. He received the B.E. degree in 1936 and the Ph.D. degree in 1939, both from Yale University. During the summer of 1937 he was employed by the General Electric Company, and during the academic year 1938 and 1939, he was the recipient of a Charles A. Coffin Fellowship from this company. From 1939 to 1942, he was an instructor in electrical engineering at the Pennsylvania State College. From 1942 to 1945, he served as chief engineer of the development branch, Special Projects Laboratory, Radio and Radar Subdivision, ATSC, Wright Field, Ohio. He subsequently received the President's Certificate of Merit for outstanding service in this post.

E. M. Williams

In 1945, Dr. Williams was appointed associate professor of electrical engineering at the Carnegie Institute of Technology, in Pittsburgh, Pa., where he is now a professor. He is also serving as expert consultant to the Research and Development Board of the National Military Establishment.

Dr. Williams was the recipient of the 1946 Eta Kappa Nu award. During 1947–1948 he served as Chairman of the IRE Pittsburgh Section.

Institute News and Radio Notes

Technical Committee Notes

The **Standards Committee** met on June 9, under the chairmanship of J. G. Brainerd. The following Standards material was acted upon: (a) Additional definitions submitted by the **Electron Tubes and Solid State Devices Committee** were approved; (b) The proposed standards of the **Piezoelectric Crystals Committee** were approved and will be published shortly; (c) Material submitted by the **Radio Receivers Committee** "Tests for Effects of Mistuning and Downward Modulation" was approved; and (d) The Standards Committee approved its subcommittee material on the numbering and indexing of standards. Technical committees and subcommittees will also be identified by numbers under the proposed system. A subcommittee of this committee is studying the question of the most economical and thorough method of circulating IRE Standards to assure the widest distribution. The Standards Committee is reviewing the scope of twenty-three technical committees. . . . The **Antennas and Wave Guide Committee** held a meeting on May 17, Chairman L. C. Van Atta presiding. Dr. Van Atta recommended the formation of two subcommittees to expedite the work of the main committee, in the preparation of the definitions of terms. L. J. Chu was appointed Chairman of the Subcommittee on Antennas, and A. G. Fox was made Chairman of the Subcommittee on Wave Guides. . . . A meeting of the **Electron Tubes and Solid State Devices Committee** was held on June 2 at which it was decided to submit Power Output High Vacuum Tube Definitions and Small High Vacuum Tube Definitions material to the Standards Committee for approval. . . . **The Navigation Aids Committee**, meeting on June 7, is working on definitions of Pulse Systems, under Henri Busignies, Chairman. The following tasks have been assigned: classification of electronic methods of navigation systems; presentation of accuracy data; concept of reading time to radio navigation systems; and coverage of air navigation systems. . . . **The Piezoelectric Crystals Committee** met on May 9. Work on "Proposed Standards on Piezoelectric Crystals" having been completed this material has been submitted to the Standards Committee for approval. . . **On June 8 the Video Techniques Committee** held a meeting, at which the following subcommittees reported on the status of their work: Subcommittee V-1, Definitions and Symbols of Video Terms, R. H. Daugherty, Jr., Chairman; Subcommittee V-2, Utilization, Including Video Recording; Methods of Measurement and Test, R. L. Garman, Chairman; Subcommittee V-3, Video Systems and Components, Methods of Measurement and Test, W. J. Poch, Chairman; Subcommittee V-4, Video Signal Transmission, Methods of Measurement and Test, L. W. Morrison, Chairman. . . . **Professional Group Chairmen and Sponsors** met on May 9 at Syracuse, N. Y., where reports were given from three of the six organized Groups, there being no representative present from the Circuit Theory, Audio, or Vehicular and Railroad Radio Communications Groups. Planning reports were presented by sponsors of potential Groups. A progress report on the Quality Control Group was made by Jerome Steen and R. F. Rollman, and Virgil Graham reported for the potential Electron Tube Group. The name of the Broadcast Engineers Group was changed to the IRE Professional Group for Broadcast Transmission Systems, which will include AM, FM, television, facsimile, and the like. O. W. Towner is the Chairman of this Group. . . . On June 7 the **Joint Symposium Planning Group of the IRE/AIEE** held its initial meeting at IRE headquarters to formulate the plans for the 1949 Nucleonics Symposium, tentatively scheduled for October 31, and November 1 and 2, at the Hotel Commodore, in New York City. Harner Selvidge was appointed Chairman of the Joint Symposium Planning Group, representing the IRE. Other members are R. L. Butenhoff, IRE; W. A. Geohegan, AIEE; and G. W. Dunlap, AIEE. Plans were discussed on the basis of a three-day symposium. The tentative program is as follows: First Day, Electronic Aids to Medicine; Second Day, Nucleonics and Biological Tolerances; Third Day, Nucleonics and Industry. Final plans will be announced at a later date.

Newfoundland Added to Montreal Section

Because Newfoundland has become a part of the Dominion of Canada, the territory of Newfoundland and the IRE members residing therein have been added to the Montreal Section of The Institute of Radio Engineers.

Calendar of COMING EVENTS

Summer Seminar, Emporium Section, IRE, Emporium, Pa., August 19–20

AIEE Pacific General Meeting, San Francisco, Calif., August 23–26

1949 IRE West Coast Convention, San Francisco, Calif., August 30–September 2

1949 National Electronics Conference, Chicago, Ill., September 26–28

National Radio Exhibition, Olympia, London, England, September 28 to October 28

AIEE Midwest General Meeting, Cincinnati, Ohio, October 17–21

Radio Fall Meeting, Syracuse, N. Y., October 31, November 1–2

1949 Nucleonics Symposium, New York City, October 31, November 1–2

IRE–URSI Fall Meeting, Washington, D. C., October 31, November 1–2

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

Dayton Section Television Lecture Series

To make available recent engineering knowledge of television, the Dayton Section of the IRE last winter conducted a program of television lectures, beginning on January 31, and continuing every Monday until completion, except for the week of the National IRE Convention. The subjects and lecturers for the series were as follows: "Picture Quality," by Madison Cawein, Television Consultant; "Pickup and Transmitting Equipment," by Howard Lepple, Crosley Broadcasting Corp.; "Interconnecting Facilities," by L. W. Morrison, Bell Telephone Laboratories; "Receiving Antenna and Distribution Systems," by Andrew Alford, Consultant; "Receivers," by H. R. Shaw, Colonial Radio Corp.; and "Future Trends," by Joshua Sieger, Freed Radio Corp.

IRE Mentioned in Report on "Scientific Manpower"

The Annual Report on "Scientific Manpower" submitted by the Secretary of the Army to the Secretary of Defense, and covering the period July 1, 1947 to November, 30, 1948, mentions six professional societies as having agreed to participate in a plan whereby their committees will give advice and assistance to the Department of the Army in the fields unique to the societies which the committees represent. Of these six professional societies, The Institute of Radio Engineers is the only one representing electronics or electrical engineering.

IRE to Hold Tenth Annual Seminar at Emporium

The tenth annual summer seminar of the Emporium Section of The Institute of Radio Engineers will be held on August 19 and 20, Roger W. Slinkman, Chairman of the Emporium Section, announced recently. Speakers on Friday, August 19, will include H. G. Clavier, who will discuss carrier power requirements for long-distance communications by microwaves; and Allen A. Barco, whose subject will be problems of television deflection and high voltage supplies. On Saturday, August 20, Michael Landis will speak on modern orthicon camera chain in television; and Gerrard Mountjoy and Henry Appel will discuss design features for a new television receiver.

Following the technical sessions on Saturday morning, the seminar committee will sponsor a picnic at Erskine's Grove to which the public will be invited. George Brunner is Chairman of the Seminar Committee.

Joint Technical Advisory Committee

Summary of Activity— October, 1948 to May, 1949

On October 1, 1948, the Chairman of JTAC, Philip F. Siling, wrote to the Chairman of the FCC, Wayne Coy, offering the assistance of JTAC in obtaining further information and data pertinent to the use of the uhf spectrum for television, or other pressing problems. Mr. Coy's reply, dated October 28, referred to the agenda for the FCC Engineering Conference on VHF Television and FM to be held November 30, 1948, and requested the assistance of JTAC concerning vhf propagation, antennas, co-channel and adjacent-channel interference protection ratios, signal levels required for satisfactory service, and power generation capabilities of transmitters.

In response to this request the JTAC prepared, and offered as evidence at the Engineering Conference, Volume II of the JTAC Proceedings, entitled "Allocation Standards for VHF Television and FM Broadcasting," dated December, 1948. In this report JTAC stated the industry findings on apparatus capabilities but declined to approve or disapprove the FCC T.I.D. reports on vhf propagation, inasmuch as time was not available for the IRE Wave Propagation Committee to review the substantiating data, the methods of using these data, and to give JTAC its expert opinion. The T.I.D. reports were referred by the FCC to the FCC Ad Hoc Committee (on VHF Propagation for Television and FM Broadcasting). JTAC offered, at the Conference, to review the findings of the Ad Hoc Committee when available.

The following Table, taken from JTAC Proceedings Volume II, summarizes the allocation standards recommended by JTAC to the FCC.

RECOMMENDED ALLOCATION STANDARDS

Service	Signal Strength	Co-channel protection	1st Adjacent channel protection	2nd Adjacent channel protection
Suburban —Rural TV	500 μv/m (see note)	40 db 25 db*	0 db** 6 db***	— —
Urban TV	5000 μv/m	40 db	0 db**	—
Rural FM	25 μv/m	10 db	−16 db	−40 db
Urban FM	—	10 db	−16 db	−40 db

* Tentative value when co-channel carriers are phase-synchronized.
** Upper adjacent channel.
*** Lower adjacent channel.

In Volume II, the JTAC took note of the technique of phase-synchronization of television picture transmitters, recommending that this technique be taken into account in allocations planning and that further study of it be undertaken. An improvement in co-channel interference of 15 db was mentioned as the best estimate (class-C reliability) then available on the benefit of carrier synchronization.

On December 16, Mr. Siling informed Mr. Coy of the desire of JTAC to assist the commission further in these or other matters. In reply Mr. Coy wrote JTAC under date of December 28, 1948, suggesting the following studies be made:

(1) A study of the number of uhf television channels needed to supplement the vhf channels to provide a sound, competitive national allocation.

(2) A study of the place of color television in the uhf spectrum.

(3) Further study of phase-synchronization of television picture carriers, including possible application to uhf monochrome and color transmissions.

(4) Further study of receiver characteristics, particularly adjacent-channel selectivity, oscillator radiation, choice of a standard intermediate frequency and its effect on channel assignments.

(5) Re-examination of uhf transmitter power-output capabilities.

(6) Preparation of a plan for obtaining more extensive knowledge of propagation on frequencies between 475 and 890 Mc.

On February 18, JTAC replied in a letter addressed to Mr. Coy, expressing the following conclusions:

> JTAC declined to define the number of television stations required in each metropolitan center for national competitive system, on the ground that this was not a question within its province. However, JTAC indicated the range of values which this number might take, based on limitations of channel space within the 475–890-Mc band. The minimum number of stations in each metropolitan center was taken as 2 (since 1 station suffers no competition) and the maximum was taken as 5 (since 6 or more could not be accommodated in the space in the uhf spectrum reserved for television broadcasting). The following table, taken from the letter, shows the number of uhf channels required for a minimum number of stations of 2, 3, 4 and 5 per metropolitan centers, under the condition that the co-channel separation between stations is 150 miles for synchronized and 210 miles for non-synchronized stations. The adjacent-channel separations are 75 miles and 105 miles, respectively.

Allocation Plan No. 1
(Employing phase synchronization on VHF only)

2 Stas. Min.	*3 Stas. Min.*	*4 Stas. Min.*	*5 Stas. Min.*
23	35	54	65
(46)	(34)	(15)	(4)

Allocation Plan No. 2
(Employing phase synchronization on VHF and UHF)

2 Stas. Min.	*3 Stas. Min.*	*4 Stas Min.*	*5 Stas. Min.*
18	27	40	52
52)	(42)	(29)	(17)

There are 69 6-Mc channels available between 475 and 890 Mc.

The figures in parentheses indicate the number of 6-Mc uhf channels which would remain for other systems of television. JTAC expressed the opinion that it was not practical to assign vhf channels only to primary cities and uhf channels only to secondary cities.

Regarding color television, JTAC urged that experimentation with color television be actively encouraged, and recommended that adequate space be set aside above 900 Mc and below 6,000 Mc for this purpose. It was recommended that no bandwidth restriction be placed on experimental color television transmission for the present, until the relative importance of color and detail in the total value of a television picture can be determined.

Regarding phase-synchronization of picture carriers, the 15-db figure of improvement was reiterated and its reliability raised from class C to class B.

The opinion was expressed that, in theory at least, this technique could be applied to uhf transmitters, but the figure of improvement could not be stated pending actual tests.

The other items listed in Mr. Coy's letter of December 28 were stated as being under study. It was recommended that a program of uhf propagation studies be set up, under government auspices if possible.

On February 17, 1949, E. W. Allen, Jr. of the FCC wrote to Mr. Siling pointing out an apparent inconsistency in the field strengths for satisfactory service cited in JTAC Proceedings Volumes I and II. Under date of April 18, JTAC replied, stating that the field strength cited in Volume I were too low by a factor of 2, since the attenuation due to the vestigial sideband characteristic of the receiver had not been taken into account. The field strength figures in Volume II were reaffirmed.

Following the receipt of Mr. Coy's letter of December 28, JTAC circulated a questionnaire to the IRE and RMA Television Systems Committees, concerning the suitability of the vhf television system standards for use in the uhf service. The replies to this questionnaire were summarized and transmitted to Mr. Coy under date of April 4. A large majority of the replies stated that the vhf standards were applicable without change, while a small number stated that certain tolerances on carrier frequencies and levels should be examined and changes introduced for vhf as well as uhf transmissions. The opinion was also expressed by several of the respondents that a portion of the uhf spectrum should be reserved for experimentation with color and high-definition monochrome systems.

Under date of April 4 a letter was sent by JTAC to Mr. Coy commenting on the possible use of frequency modulation for the picture transmissions. Previous experiments with this type of transmission were mentioned, and the difficulties encountered from multipath transmission were described. JTAC suggested that any plan to consider using frequency modulation for picture transmissions should be preceded by a comprehensive field test under normal conditions of multipath transmission. JTAC questioned whether, in view of the present state of knowledge on this subject, the time, effort and expense of such a field test would be justified.

Under date of March 30, the JTAC Secretary, L. G. Cumming, transmitted to Mr. Coy copies of replies to a questionnaire circulated by JTAC among manufacturers of transmitting tubes and equipment, indicating the amount of power which might be developed at various frequencies, at that present time, and what increases might be expected in the future.

On March 22, Mr. Coy sent a letter to JTAC raising further questions relating to his letter of December 28 and JTAC's reply of February 18. The questions included:

(1) The preferable arrangement and position of uhf channels within the 474–890 meter band, i.e. as a contiguous block, or interspersed with other assignments.

(2) The justification for assuming a smaller adjacent channel separation for synchronized operation than for nonsynchronized operation.

(3) The effect of terrain on propagation.

(4) The absolute values of signal-to-interference ratios which give satisfactory service, with and without synchronization.

(5) Characteristics of receivers, particularly adjacent-channel interference ratio, acceptable signal-to-rms-noise ratio and its relation to signal-to-interference ratio, and noise measurements of commercially available receivers.

Under date of May 22, JTAC replied to Mr. Coy, recommending that uhf channels be assigned in a single block beginning at the lower end of the 475- to 890-Mc band. The difficulty of providing against adjacent channel interference was admitted. It was stated that considerable improvement in adjacent channel selectivity could be achieved in receivers, at some increase in cost. JTAC offered to secure the absolute values of signal to interference ratios requested, but recommended against any delay in proceeding with the allocation while these data were being collected. The off-set system (carriers separated about 10,500 cps) of reducing co-channel interference was mentioned as offering a degree of improvement somewhat superior to that of phase synchronization, with less apparatus complexity. JTAC offered to keep the FCC advised of progress with this system. Noise figures of 12 to 13 db were reaffirmed as typical of properly adjusted receivers of the present date, but an additional tolerance of 3 to 5 db was suggested to take account of production variations. The choice of a standard intermediate frequency, with concomitant arrangement of channels to reduce oscillator radiation and image interference, was endorsed, and it was stated that the RMA Television Receiver Committee was being asked to recommend a standard value.

Professional Groups Given Nominating Privilege

Professional Groups now have the opportunity of nominating the Chairman of the subcommittee of the Papers Procurement Committee which is concerned with that Group's field. They may also nominate interested and competent men in their fields as members of the Papers Review Committee and the Board of Editors, all such nominations being subject to acceptance by the Chairman of the Committee concerned, the Editor, and the Executive Committee.

JTAC Officers Announced

The Boards of Directors of the IRE and RMA have appointed Donald G. Fink to serve as Chairman of the Joint Technical Advisory Committee during the period from July 1, 1949, to June 30, 1950; and J. V. L. Hogan to serve as Vice-Chairman for the same period. The JTAC announced that L. G. Cumming will continue to serve as Secretary.

EJC Plans File of Key Engineering Personnel

To provide source material for a Who's Who in engineering research, development, and other scientific operations for use by the National Military Establishment, the Engineers Joint Council, acting through The American Society of Mechanical Engineers, has accepted the task of providing the Office of Naval Research with the names, addresses, ages, and details of professional and scientific qualifications of 100,000 key engineers in all branches of American engineering.

The source file of key engineering personnel thus obtained will provide a valuable tool for solving a variety of technical personnel problems. Its use will diminish disturbances of the national economy, organization of industry, and the personal welfare of engineers, and provide a means by which national resources of technical personnel can be ascertained. The file will also point up weak spots which should be strengthened by education, training, and other means. As a national asset the body of facts will be available to private industrial, educational, and professional society planning groups, and for other legitimate purposes.

A four-page questionnaire will soon be mailed to 100,000 engineers holding the grade of member or higher in 18 national professional engineering societies. The returns will be collected by the ASME and turned over to the Office of Naval Research of the National Military Establishment for classification.

The project is the result of a conference held in Washington, D. C. last fall, attended by EJC representatives and many other engineering agencies, at which was discussed the need for a list of 25,000 key engineers working in research, development, and other scientific projects who could be called in on a full or part-time basis to work on the broad scientific programs of the National Military Establishment. The task of collecting personal and professional data fell to the EJC as the largest joint agency of the engineering profession. The Engineers Joint Council points out that this will not be just another questionnaire, but one sent to engineers selected from the upper echelon of the profession. The data sought are not intended for general government use, but will go directly to the engineering agencies of the National Military Establishment. As the questionnaire will provide the key to opportunity to professional and patriotic service, engineers selected to receive it are urged to give it serious attention, and to answer all questions fully. The Institute of Radio Engineers is co-operating in this project, and asks its membership to assist in every possible manner.

Television to be Government Monopoly in Australia

Australia will have television in two years, Prime Minister J. B. Chifley announced recently. Run as a government monopoly, television stations will be built in six Australian capital cities—Brisbane, Sydney, Melbourne, Adelaide, Perth, and Hobart. It was also disclosed that each picture will be produced by 625 closely spaced horizontal lines, and compared with 405 lines used by Britain, and 525 in the United States. It was stated that the use of a greater number of lines should ensure a better image than is now available under either the British or American system.

Press Switching Center Aids Foreign Council Meeting

A new press switching center at New York and special direct cable facilities that have been set up in Paris are working together to provide fast and efficient communications to the government and press representatives attending the Council of Foreign Ministers meeting, it was announced recently. One of the special facilities is an exclusive Western Union office, established adjacent to the American delegation conference room in the Hotel Crillon, which flashes thousands of words to the new press switching center in New York. The stories arrive in New York in the form of perforated tape, and are sent through automatic transmitters over direct circuits to press services and newspapers throughout the United States.

National Radiotelephone Network Formed

The formation of the first national network of independent radiotelephone stations for mobile service to the general public was announced recently in a statement filed with the Federal Communications Commission in behalf of the National Mobile Radio System. The new system will offer a practical and unprecedentedly low-cost means of communication between occupants of automobiles, trucks, buses, and other vehicles with offices or homes hundreds of miles distant. The network also has important potentialities as an auxiliary communications system in event of national emergency. By August, the facilities will be in operation between Boston and Washington, and will gradually be extended to encompass all member stations.

The system differs from the mobile radio service offered by telephone companies in that brief messages or conversation are relayed back and forth through the intermediary of a station operator, who receives them by telephone from one end and relays them by radio to the other, or vice versa. Although the new system started with stations on the Eastern seaboard, others in the South and West have joined or have indicated desire to affiliate. It is anticipated that by 1950 at least 100 stations will be active in the network.

Radiocommunication Institute Established in Argentina

The Instituto de Radiocomunicaciones, established last fall at the National University of Tucuman in Argentina, recently announced the appointment of Daniel Eduardo Frias (M'48) as organizer and director. In September, 1948, the Board of Trustees approved curricula for the training of telecommunication engineers, radio technicians, radio fitters, and radio operators. The following departments have also been formed: ultra-high frequency technique, supersonic and hydrophonic waves, air navigation radio, television, radio broadcasting, applied electronics with tubes used in radio communication, and a measuring laboratory. The schedule of studies for the telecommunications courses were adapted especially for the National University of Tucuman by Dr. Frias.

Early in 1949 various measuring equipment and instruments were purchased, and a full broadcasting station was installed by arrangement of Dr. Horacio Descole, President of the University.

First NAB Program Directors' Clinic Held in Chicago

The first National Association of Broadcasters Program Directors' Clinic was held in Chicago on June 27, 28, and 29, on the downtown campus of Northwestern University. All meetings were devoted exclusively to specific, practical discussions for improving program structures at local stations, and aiding program directors to utilize available services to better advantage. Justin Miller, NAB president, spoke to the representative program delegates, stressing the fact that "programs mean audience and audience means income and economic stability."

Industrial Engineering Notes[1]

FCC Actions

The FCC has adopted new rules permitting **private sale of radio broadcasting stations** without public advertising or competitive bidding. The Commission dropped the regulation it has had since 1945 requiring all proposed transfers of all types of broadcasting facilities to be advertised locally for a 60-day period. Under the new procedure, an application for transfer of a radio station will simply be filed with the FCC so that the legal and financial qualifications of the purchaser may be examined. . . . A **total of 2,792 AM, FM and television broadcasting stations** were on the air at the end of May, the FCC reported. . . . The FCC, in an action possibly affecting the use of uhf frequencies by television broadcasting services, today amended its order of May, 1948, relating to the utilization of frequencies in the 475–890 Mc band for television to include **consideration of the allocation of the 470-400 Mc band to multichannel broad-band common carrier mobile radio operation** in lieu of television broadcasting.

National Bureau of Standards Announces New Developments

The National Bureau of Standards recently announced that magnetic fluids, which were originally used in the NBS electromagnetic fluid clutch, have several unique features that make possible important applications of iron-oil mixtures. Studies of the properties of the mixtures reveal that magnetic fluids may be employed to good advantage in a number of ways, including use as variable electrical resistors.

An electronic gating instrument has been designed and constructed at the National Bureau of Standards for the "accurate determination of the deadtime and recovery characteristics of Geiger-Muller counters." Complete information on the new development is available as Research Paper RP1965 at ten cents per copy from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C.

Preferred Parts List Explained by ASESA

To assist further in promoting the electronic component standardization program, the Armed Services Electro Standards Agency has prepared a "Preferred Parts List" for use as a guide in selecting components commonly used in the development and maintenance of electronic equipment. This list contains the electronic components for which preferred form factors, values, and tolerances have been selected. Consideration has been given to cost and production factors; these are reflected in the list by the fact that components included have been screened for the most reasonable tolerances used in the maximum of engineering applications and for those which will not appreciably increase the cost of purchase when bought in large quantities.

Television News

The FCC recently released the long-awaited report on radio propagation effects in the frequency range between 50 and 250 Mc, as compiled by a small group of government and industry engineers since the FCC's television engineering conference last December.

Documents submitted by the committee included the report and four reference volumes pertaining to its findings. The main document (Vol. I) contains the following committee information and recommendations:

A. Evaluation of the Random Variations in Field Intensity from Median Levels Due to Local Terrain and Buildings.

B. Method of Combining the Effects of the Spatial and Time Variations of the Desired Signal and One or More Interfering Signals.

Postponement of the lifting of the present TV "freeze" until late fall to allow time for a new hearing and other proceedings, including consideration of 6-Mc color on both vhf and uhf channels, was announced by the Federal Communications Commission.

Radio Set Production Declines as Television Rises

The weekly rate of television receiver production in April was the highest yet attained by the industry, according to an RMA tabulation showing that RMA member-companies manufactured 166,536 TV sets during that month.

Radio receiver production in April dropped to new low levels. Radio sets reported for the month by RMA members totaled 506,409, of which 37,563 were FM and FM-AM types. AM set production of 468,906 was the lowest since January, 1946, when the industry was reconverting from military to civilian manufacturing.

For the first time RMA members reported the number of TV receivers with FM reception facilities. These sets totaled 47,264 or about 28 per cent of the TV receivers produced.

FM and FM-AM set production in April was about 62 per cent under the weekly average for the first quarter of 1949.

Canadian Radio News

Approximately 3,750,000 radio receivers were in use in Canada at the end of 1948, and 575,000 sets were produced during the year. Canadian sales of radio receivers rose in numbers in January, 1949, but declined in price. A total of 40,794 units valued at \$3,308,270 were sold, compared with 39,046, sets valued at \$3,720,102 in January, 1948.

[1] The data on which these NOTES are based were selected, by permission, from the Radio Manufacturers Association's "Industry Reports," issues of February 18, 25, and March 4, 11, and 18.

Sales of radio receiving sets by Canadian manufacturers in February totaled 44,268 units valued at $3,328,642, compared with 35,833 sets valued at $3,230,740 in the corresponding month of 1948, according to a report received recently by the U. S. Department of Commerce.

Canadian imports of sets in February amounted to 1,567 units valued at $81,551, compared with 1,379 units valued at $58,529 in January. Exports in February totaled 2,898 sets valued at $106,143, against 2,124 units valued at $74,920 in January.

Receiving tube production in February totaled 309,728 tubes valued at $163,574, compared with 263,582 units valued at $133,612 in January. Canadian manufacturers imported 54,049 tubes valued at $57,916 in February, as against 128,050 tubes valued at $85,885 in January. Imports of tube parts were valued at $40,372 in February, against $30,986 in January.

Radio and Television Abroad

Local assembly of radio receivers was started in **Greece** during March, 1948, and 7,000 sets were completed during the year.

Production of radio receivers in **Austria** in 1948 totaled 96,437 sets, as compared with 21,246 in 1947 and 127,472 in 1937 . . . . At the end of 1948 there were 475,000 radio receivers in Hungary, 188,000 of them in **Budapest** . . . . On December 31, 1948, there were 42 radio broadcasting stations in **Panama**, compared with 26 at the close of 1947.

Licensed radio receivers in **Denmark** on March 31 totaled 1,177,608 and represented an increase of 4.43 per cent over the 1,127,677 licensed sets on the same date in 1948. . . . In **Brazil** the value of imports of radio sets, parts, and tubes in 1948 dropped to 230,128,000 cruzeires from 454,733,000 cruzeires in 1947, according to a report received by the U. S. Department of Commerce. U. S. manufactured equipment accounted for 86.5 per cent of the total in 1947 and 76.6 per cent in 1948. Holland supplied 10.5 per cent of the Brazilian radio imports in 1948 and 3.7 per cent in 1947. The United Kingdom accounted for 8.0 per cent of the Brazilian total in 1948 and 4.9 per cent in 1947. . . . Radio receiver production in the Bizonal area of **Germany** totaled 407,472 in 1948, according to statistics received by the U. S. Department of Commerce. Production of receiver and amplifier tubes and valves totaled 3,111,885 units in 1948. Production of radio sets in **Austria** totaled 96,437 in 1948, compared with 21,246 in 1947 and 127,472 in 1937. . . . Production of radio receivers in **Japan** in 1948 averaged 67,749 sets monthly, according to statistics received recently by the U. S. Department of Commerce. This compares with monthly averages of 64,870 receivers and 43,221 sets in 1947 and 1946, respectively. Transmitter production averaged 168 monthly in 1948, compared with 227 in 1947 and 97 in 1946. . . . A bill was recently introduced in the legislature of **Cuba** to "regulate industries" including the manufacture and assembly of phonographs, radio and television transmitting and receiving apparatus. The measure would grant for a period of ten years exemption from consular fees, duty, taxes, charges, and surcharges on imports by these industries of parts and pieces used in manufacture or assembly. . . . Television is still in an experimental stage in **Sweden** and experts agree it will be from two to three years before programs for the public are introduced, according to a report on television prepared by Valdemars Kreiebergs, of the American Embassy at Stockholm. "Televising of programs to the public will not be introduced before Sweden and a number of other countries of the European Continent, including Denmark, Norway, Holland, Belgium and Switzerland, have agreed on standardization, i.e., the number of lines to a surface unit in the transmitted picture," the report points out. Different types of foreign receivers have been tested but it has not yet been decided which type will be adopted for Sweden. A modified U. S. system, a receiver using 625 lines instead of the 525 lines used in this country, is "seriously considered." Current television research in Sweden is being carried on under the supervision of the Television Research Board. The Board was established in November, 1947, at a conference of representatives of interested Swedish agencies and firms. Several press releases from the Television Board mentioned the building and installation of a television transmitter in the Stockholm Institute of Technology. Its power is 1 kilowatt, and at present it operates with a carrier frequency of 80 Mc. Later one will test operation on frequencies of 60 and 200 Mc.

RMA Elects New Officers

At its "Silver Anniversary" convention held on May 16–19, at the Stevens Hotel, Chicago, the RMA elected and appointed the following officers:

President

R. C. Cosgrove, Exec. Vice Pres., Avco Mfg. Corp., Cincinnati, Ohio (new-4th term)

Executive Vice-President and Secretary

Bond Geddes (A'41), 1317 F Street, N.W., Washington, D. C. (re-elected)

Vice-Presidents

G. M. Gardner, Chairman, Wells-Gardner & Co., Chicago, Ill. (re-elected)

R. E. Carlson, Vice-Pres., Tung-Sol Lamp Works, Inc., Newark, N. J. (re-elected)

W. J. Barkley (M'29–SM'43), Exec. Vice-Pres., Collins Radio Co., Cedar Rapids, Ia. (re-elected)

A. D. Plamondon, Jr. (SM'46), Pres., Indiana Steel Products Co., Chicago, Ill. (re-elected)

A. Liberman (SM'49), Pres., Talk-A-Phone Co., Chicago, Ill. (new)

Treasurer

Leslie F. Muter, The Muter Co., Chicago, Ill. (re-elected)

Director of Engineering Department

W. R. G. Baker (A'19–F'28), Vice-Pres., General Electric Co., Syracuse, N. Y. (reappointed)

Books

Vacuum Tube Amplifiers, edited by George E. Valley, Jr., and Henry Wallman

Published (1948) by the McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 18, N. Y. 733 pages +9-page index +xvii pages. 186 figures. $6\frac{1}{4} \times 9\frac{1}{4}$. $10.00.

This book is volume eighteen of the twenty-eight-volume Radiation Laboratory Series instituted to record wartime contributions to the art. As one can readily surmise, it has been necessary to fill in wartime contributions with considerable background material, in order to give the volume continuity. It is believed that the book would have been more useful and valuable, had the background material been carefully annotated with footnote references.

The preface states that:

> The amplifiers discussed in this volume are designed to have extreme values in one of several of the pertinent characteristics: bandwidth, sensitivity, linearity, constancy of gain over long periods of time, etc.

This represents a good summary of the material included, to which should be added that the manner in which the material is treated ranges from profound and precise mathematics to handbook material, such as JAN characteristics and tolerances. It is, therefore, difficult to recommend the book to specific groups; rather, it will probably be found that certain parts of the book will be valuable to widely divergent interests. Since there is no other book of comparable scope now available, those actively engaged in amplifier development and design will find it very useful. The material included will doubtlessly also be valuable to college professors in preparing classroom lectures. Frequent tabulations, summaries, and circuit diagrams are of considerable utility.

Chapter Four, "Synchronous and Staggered Single-Tuned High-Frequency Bandpass Amplifiers," Chapter Five, "Double-Tuned Circuits," and Chapter Fourteen, "Measurements of Noise Figure," contain material of sufficient interest and importance to warrant special mention. Dc restorer circuits are treated only briefly. The chapter on dc amplifiers does not include consideration of flicker effect on stability, and a nine-page index hardly seems adequate for a book of this size and scope.

L. J. Giacoletto
RCA Laboratories
Princeton, N. J.

Cosmic Ray Physics, by D. J. X. Montgomery

Published (1949) by the Princeton University Press, Princeton, N. J., 357 pages+12-page index +viii pages. 123 figures. 6×9. $5.00.

There is perhaps no field of research which has undergone as rapid and revolutionary a development during the past two decades as has the study of cosmic rays. During this short period the original concept that the primary cosmic radiation consisted of photons changed first to an electron hypothesis, then to a positive proton theory (with occasional suggestions of negative protons), and more recently to the proton plus heavier charged particles concept. To write a book on a subject which is undergoing such rapid changes is a hazardous undertaking; the book might well be out of date before it is published.

Since "Cosmic Ray Physics" by D. J. X. Montgomery is the most recent book published on cosmic rays to the reviewer's knowledge (at the time of writing), it is also the most up to date. In general, the author has handled the broad scope of the subject very well by covering both the early work, as well as most of the important recent work (up to the publication date of September, 1948), including a chapter on the very recent findings on charged particles heavier than protons in the cosmic radiation.

The basis for the material in the book is a series of lectures given by Professor Marcel Schein of the University of Chicago at Princeton University in the spring of 1948. Professor Schein is one of the outstanding authorities in the field of cosmic rays, having worked with A. H. Compton at Chicago for many years, and, during the past ten years, having himself made many notable contributions in this field. The original notes prepared from Professor Schein's lectures have been considerably expanded, and much material has been added, including some contributions by Drs. Neils Arley and Shuichi Kusaka, both outstanding theoretical physicists in the field of cosmic rays.

The book is written from the experimental point of view, and only such theory as is necessary to the fundamental concepts and the immediate experiments is included. Several chapters are devoted to the experimental equipment and techniques used in cosmic-ray work. A chapter on the intensity of cosmic radiation discusses the absolute intensity as well as the geomagnetic effect and its relation to the directional intensity. The simple Stoermer theory of the motion of a charged particle in the field of a magnetic dipole (the earth) is given in an Appendix.

Mesons in the cosmic radiation are discussed very completely under the heading of "Hard Component," and the types of meson—their charge, mass, lifetime, spin, etc.—are discussed in great detail. The chapter on the soft components, written by Dr. Arley, reviews the subject of the electron and photon components without recourse to the very successful but somewhat complex cascade theory for which Dr. Arley was partly responsible.

The book is well organized and is profusely illustrated with excellent reproductions of cloud chamber and photographic emulsion tracks, as well as many curves and line drawings. One of the Appendices contains the very useful range-energy and range momentum curves for protons, mesons, and electrons, which were prepared by Princeton as part of an Office of Naval Research project.

"Cosmic Ray Physics" should prove valuable both to the individual working in other fields who has a good physics background and a casual interest in cosmic rays, as well as a reference text for the worker in the field. It should also serve as an excellent textbook for the study of cosmic rays.

E. H. Krause
Naval Research Laboratory
Washington 20, D. C.

Radio at Ultra-High Frequencies, Volume II

Published (1948) by the *RCA Review*, RCA Laboratories, Princeton, N. J., 485 pages+x pages. 350 figures. 6×9. $2.50.

This is a well-organized reference work for the engineer engaged in development and research on systems and components in the ultra-high and higher frequency portions of the spectrum. Papers of major importance by RCA authors, which were published in various technical journals during the years 1940 through 1947, and covering in general the frequency range 300–3000 Mc, are here grouped in sections covering antennas and transmission lines, propagation, reception, radio relays, microwaves, measurements components, and navigational aids. A complete bibliography of the ultra-high-frequency field is appended in addition to summaries of all papers appearing in Volume I of "Radio at Ultra High Frequencies."

For those engineers and scientists unfamiliar with ultra-high-frequency techniques, components, and systems, this book provides under one cover material which previously was available only after much laborious library research.

J. L. Heins
Sperry Gyroscope Co.
Great Neck, L. I., N. Y.

The Radio Amateur's Handbook

Published (1949) by the American Radio Relay League, West Hartford, Conn., 605 pages+10-page index. 1,651 illustrations. 6½×9½. $2.00.

This hardy perennial in the field of technical literature, now in its twenty sixth edition, remains a mine of practical circuitry for the radio beginner, the active amateur, and the professional engineer alike. In no sense a substitute for a university course in electrical engineering, it nevertheless presents a surprising amount of fundamental radio theory in an easily absorbable form, with a strong emphasis on physical concepts, rather than mathematical derivation and exposition. From the opening chapter reviewing briefly the history, scope, and purpose of the amateur radio movement to the final tabulation of vacuum-tube data, the book hews closely to its avowed purpose of presenting within one cover the basic information needed to enter the field of amateur radio, and to remain abreast of those developments in the field of communications most readily adapted to the amateur service.

Following two chapters devoted to a compressed but lucid exposition of electrical laws and vacuum tube principles, chapters four to ten are given over to applications of principles in practical equipment. Actual component values for typical circuits are given, together with photographs clarifying mechanical layout, and a considerable amount of the engineering "know-how" needed in practical design work is conveyed in this sugar-coated way. The succeeding five chapters are devoted to vhf and uhf principles and practice, with considerable use of circuitry brought to high development during the war, such as broadband rf amplifiers, triode mixers, etc. Some principles of klystrons, magnetrons, waveguides, and cavity resonators are also expounded, but the coverage of this field is spotty, and little practical equipment for microwave work is described.

Chapter 16 is concerned with measuring equipment, mainly conventional frequency meters, crystal calibrators, grid dip oscillators, and similar devices. The next seven chapters are given over to actual assembly and operation of a complete amateur station, and a dissertation on the organization and function of the American Radio Relay League. Two chapters on miscellaneous and vacuum-tube data of considerable reference value complete the text, and there is a fairly comprehensive index. A "catalogue" section, in which many of the more prominent component manufacturers list such of their products as are well adapted to amateur service, completes the volume.

It is perhaps imprudent to point out small deficiencies in a technical publication with an all-time circulation figure over the 2,000,000 mark, yet it may be mentioned in passing that treatment of low noise rf input circuits, double superheterodynes, superselective if circuits, and single sideband suppressed carrier techniques is rather inadequate, to say the least, and the question of television interference, currently the most vexing problem facing the active amateur, has been merely skimmed. With these few qualifications, the book is an excellent successor to previous volumes in the series, and practically a "must" for those participating in the amateur hobby.

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Radio Fundamentals, by Arthur L. Albert

Published (1948) by the McGraw-Hill Book Co., 330 W. 42 St., New York 18, N. Y. 583 pages+11-page index+vii pages. 320 figures. 6½×9½. $4.50.

In his preface, the author indicates that this new book is designed primarily for beginning students, for radio technicians, and for radio amateurs, rather than for advanced radio engineers. Within the limitations he prescribed for himself, he has done well, although portions of his material may be beyond the comprehension of the group for whom he is writing. Although the important aspects of radio have all been dealt with in a manner that should be understood by most readers, the use of complex algebra in those cases where mathematics is used seems to be a sophistication incommensurate with the avowed level of the text.

Chapter One, "Fundamentals of Acoustics," presents an interesting summary of architectural acoustics as applied to problems of studio design, and also a general treatment of microphones. The next three chapters, entitled "Electrical Fundamentals," "Series and Parallel Resonant Groups," and "Power Transference and Impedance Matching" respectively, serve as a review of the basic electrical theories underlying most radio work. The other chapters deal with transmission lines, cables, and networks; vacuum tubes; rectifiers; voltage and power amplifiers; oscillators; modulation and demodulation; radio transmitters and receivers; and antennas and radio transmission. The use of numerous sketches to illustrate the operation of the important modulators and demodulators makes the chapter on that subject particularly noteworthy. A large number of illustrated examples is given throughout the text, serving to indicate how problems that arise in radio are analyzed and solved. In some cases, however, the analytic basis for the calculations has not been included.

A rather serious weakness of the text is the looseness in the discussion of the equivalent plate circuit representation of a vacuum tube, and also the subsequent discussion of the voltage gain of an amplifier. Equation 80 on page 273 gives the gain of a single stage amplifier as positive, thereby ignoring the 180-degree phase shift through the amplifier. All subsequent gain formulas suffer in the same way. Thus one cannot tell from equation 115 whether the output from a cathode-follower is in phase or out of phase with the input. While this may be a minor point when discussing the operation of a single stage, the relative phase is frequently more important than the gain. Several obvious errors have eluded the proofreaders. For example, the photograph on page 113 is upside down. Also, on page 129 it is stated that *G*, the shunt leakage of a transmission line, is given in ohms per unit length, instead of mhos.

On the whole, however, this book is one of the better books on the beginner's level, and should find an important place in the literature of radio.

Samuel Seely
Syracuse University
Syracuse 10, N. Y

IRE People

Philip J. Senn (A'43–M'46), an active member of the Dayton Section of the IRE, died in an automobile accident while on his way to attend the Cincinnati Spring Technical Conference on Television.

Born in Chicago, Ill., on June 17, 1916, Mr. Senn studied at the Lewis Institute and also took courses in special military communications and radar. In 1942 he joined the staff of the Signal Corps Aircraft Radio Laboratory at Wright Field, Dayton, Ohio, transferring the following year to the Signal Corps Aircraft Radio Engineering School in Dayton as an instructor. Since April, 1946, he had been associated with the USAF Air Matériel Command, Aircraft Radiation Laboratory, as a radio project engineer on the research and development of guided missile radio control equipment.

Arthur V. Loughren (A'24–M'39–SM'43–F'44) has been elected vice-president in charge of research of the Hazeltine Electronics Corp.'s Board of Directors. **Orville M. Dunning** (A'31–SM'44) was made vice-president in charge of engineering; and **James F. Willenbecher** (SM'44) was chosen vice-president in charge of manufacturing.

Mr. Loughren has been on the Hazeltine staff since 1936, during which time he has made many contributions to the theory and practice of television. Previously he was associated with the General Electric Co. as a consultant on vacuum-tube problems, and supervisor of broadcast receiver development, and later with the RCA Manufacturing Company, where he was in charge of testing and inspection.

Mr. Dunning was chief engineer for the Sonora Phonograph Co.; then became manager of the Telediphone department of the Thomas A. Edison Co., and, in 1940, chief engineer of the Gray Manufacturing Co. Joining Hazeltine in 1941, he has been responsible for the administrative and technical supervision of the company's engineering work under military contracts.

Before he joined Hazeltine in 1942, Mr. Willenbecher served on the staff of the RCA Manufacturing Co., and also of the Philco Corp. Prior to this advancement, he was manager of the production division.

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J. E. Shepherd (A'36–SM'44–F'48) was recently elected the 1949–1950 president of the Technical Societies Council of New York, Inc. He is the fourth president of the council, which was organized in 1946 and has a membership of seventeen engineering, scientific, and technical societies in the metropolitan area, representing more than 25,000 members. Dr. Shepherd, a director of the IRE, has been active in the New York Section activities, serving as Chairman in 1947–1948, and Secretary in 1944–1945. He was given the Fellow award in 1948 "for his contributions to the development of airborn radar armament and for his active participation and leadership in the functions of the Institute."

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Louis G. Pacent (A'12–M'15–F'47), president and technical director of the Pacent Engineering Corp. in New York, N. Y., has been appointed consulting engineer by Plessey International, Ltd., in Ilford, Essex, England.

Mr Pacent was born on June 23, 1893, in New York, N. Y., and he received the E.E. degree from Pratt Institute in 1916. From 1915 to 1917 he was advertising manager for the IRE, and he has also been a member of numerous Institute Committees. In 1947 he received the War Department Certificate of Appreciation for his services to the Signal Corps

A Fellow of the Society of Motion Picture Engineers, the Radio Club of America, and the AIEE, Mr. Pacent is also a member of the last organization's Board of Examiners, and a member of the Acoustical Society of America and the Engineers Club of New York.

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Allen B. DuMont (M'30–F'31), president of the Allen B. DuMont Laboratories, Inc., received the honorary doctor of science degree, conferred by the Brooklyn Polytechnic Institute on June 15, 1949.

Ralph E. Hantzch (M'48), superintendent of manufacturing of exploration and production instruments and equipment at the Shell Oil Co. in Houston, Tex., died early this year.

Mr. Hantzch was born in Milwaukee, Wis., on April 2, 1896. In 1918 he was employed as an instructor at the Signal Corps School of the University of Wisconsin, following which he served as a student assistant in physics and radio for the next two years. From 1920 to 1921 he was an instructor in physics and elementary mathematics at the University of Wisconsin, receiving the B.S. degree in electrical engineering at the end of that period.

Following his graduation, Mr. Hantzch joined the Bell Telephone Laboratories as a designing engineer. He left in 1927 to become chief engineer at the Best Manufacturing Co. In 1930 he resigned to work as a private consulting engineer on theater acoustical apparatus. Six years later he became a development engineer for the T. A. Edison Co. He joined the Shell Co. in 1938.

Samuel Lubkin (SM'46) has been appointed consultant to the machine development laboratory, a division of the National Bureau of Standards' applied mathematics laboratories. There he will advise on the logical, mathematical, and engineering aspects of electronic computers, particularly the development of the NBS Interim Computer and a full-scale computer for one of the Army agencies.

Beginning his career in 1929 as a design engineer and test supervisor for the Otis Elevator Co., Dr. Lubkin left ten years later, and the following year became director of the Philadelphia Signal Depot's inspection laboratory, acting as general consultant on technical matters to the Signal Corps Inspection Agency and the Signal Corps Procurement Agency. In 1946 he was appointed engineer-in-charge of computing machines at the Aberdeen Proving Ground, leaving to become engineer-in-charge of the digital computer section of the Reeves Instrument Corp. the following year. At the same time, he was lent as consultant to the University of Pennsylvania for research on the EDVAC and ENIAC. Dr. Lubkin left the Reeves Instrument Corp. in 1948.

Inventor of a number of control devices and author of many technical papers, Dr. Lubkin has conducted extensive research in electrical, electronic, and mechanical engineering, and applied mathematics. His work has included experimental and mathematical investigations on stresses and vibrations in mechanical structures, analyses of transients in motor-speed control arrangements, programming and design of electronic computers, and instrumentation.

Wesley Tate Guest (A'32–SM'47), formerly a colonel in the Army Communications Service Division of the Signal Corps, was promoted recently to the rank of brigadier general.

General Guest was born on March 18, 1900, and received the engineering degree from Cornell University in 1921. The following year he attended the Army Signal School and in 1928 and 1929 he did graduate work at Yale University, receiving the M.S. degree. In 1946 he was at the National War College.

Wilson A. Maisel (A'46), head of the electronics department of the Frank H. Parks Co. in Portland, Ore., died early this year.

Mr. Maisel was born on November 5, 1918, in Deming, N. M., and was educated at the Stayton High School in Stayton, Ore., graduating in 1936. After working as a technician in a radio service shop, in 1940 he became assistant to the superintendent of the Instrument Laboratories, Inc., of Seattle, Wash., where he was in direct charge of the design, manufacture, and servicing of marine and industrial electronic instruments.

William Webster Hansen (A'39–F'47), radar pioneer, died recently of a heart attack after a long illness.

Born in Fresno, Calif., in May 1909, Dr. Hansen was graduated from Leland Stanford University at the age of twenty, and he became a faculty member the next year, remaining there until his death. Early in his teaching career he taught several advanced physics courses which he himself had not taken as a student. Later he included them in his studies for the doctorate, which he received in 1932.

In 1937 Dr. Hansen began work on a device to prevent airplanes from flying into mountains, and from this research came the klystron, one of the most important elements of radar. Afterward Dr. Hansen and his co-workers developed other pieces of radar equipment, including the rhumbatron, which is also used in atom-smashers.

In 1944 Dr. Hansen was awarded the IRE's Morris Liebmann Memorial Award for the "application of electromagnetic theory to radiation antennas, resonators, and electron bunching, and for the development of practical equipment and measurement techniques in the microwave field." During the war he worked an eight-hour shift in the Massachusetts Institute of Technology's radiation laboratory; then commuted to the Sperry Gyroscope Co.'s laboratory on Long Island for another eight-hour stretch. It was as a result of this wartime schedule that he became ill. He entered a hospital last fall on his return from attendance at an international atomic physics conference at the University of Birmingham in England.

At the time of his death he was professor of physics at Stanford and director of the Univerity's microwave laboratory. The billion-volt linear accelerator and atom smasher now under construction on the Stanford campus was planned and designed by him. He was a member of the National Academy of Science.

Irving G. Wolff (A'27–F'42), director of the Radio Tube Research Laboratory at the RCA Laboratories, received the Distinguished Public Service Award of the Navy Department, the highest honor bestowed on a civilian by the Navy, in recognition of his achievements in electronics and radar.

Born in July, 1894, in New York City, Dr. Wolff received the B.S. degree in physics from Dartmouth College in 1916 and the Ph.D. from Cornell University in 1923. During the interim he taught at Iowa State College in 1919 and 1920, and at Cornell from 1920 until he received the doctorate. In that year he did research on polarization capacity for the Heckscher Research Council, joining RCA in 1924 as a member of the technical and test department. From 1930 to 1941 he was with the research division of the RCA Manufacturing Co. in Camden, N. J., and since then he has been with the Laboratories.

Dr. Wolff's specialization is in problems dealing with microwaves, radar, and aviation. He has been a member of a number of Institute Committees and represented the IRE on the American Standards Association's Sectional Committee on Acoustical Measurements and Terminology. A fellow of the American Association for the Achievement of Science and of the Acoustical Society of America, he also is a member of the American Physical Society.

Alfred C. Viebranz (A'48), formerly government sales representative for the electronics division of Sylvania Electric Products Inc., at Boston, Mass., has been appointed their special representative at Washington, D. C. In addition he will act as a technical consultant in all phases of government relations. Prior to joining Sylvania, he was in the submarine service as a lieutenant in the U. S. Naval Reserves. During the war he was chief engineer and executive officer on the *U.S.S. Haddo*, and was awarded two silver stars and a bronze star for combat duty in the South Pacific. Later he served as electronics officer on the *U.S.S. Sarda*.

A native of New Rochelle, N. Y., Mr. Viebranz received the B.S. degree in physics from St. Lawrence University in 1942, while serving the United Press as a staff correspondent. Later he attended the postgraduate school of the United States Naval Academy, and was graduated as a communications engineer.

The College of Wooster conferred the honorary degree of doctor of science on **Victor J. Andrew** (A'25–M'30–SM'43) at its seventy-ninth annual convention on June 13, 1949. Dr. Andrew was graduated from the College in 1926, and received the Ph.D. degree from the University of Chicago in 1932, in recognition of work there in physics. He is now chairman of the Board of Directors of the Andrew Corp., which he established in 1936.

Ambrose C. Kibler (A'44), formerly chief engineer for radio stations WNEX and WNEX-FM, in Macon, Ga., was fatally injured in an automobile accident on May 16, 1949. Born in South Carolina in 1915, he received the A.B. degree from Newberry College in 1934. He became a radio technician by studying in his spare time, and was an instructor at the radio training school maintained by the Signal Corps, WRASC, in Warner Robins, Ga., from 1942 until the end of the war. Mr. Kibler's title was civilian radio training administrator.

William E. Osborne (A'41) has been designated chief of the electronic and guidance division of the Marquardt Aircraft Co. This division has been an operating unit under Mr. Osborne since 1947, and has just recently been moved to Van Nuys, Calif.

Mr. Osborne has been engaged in electronic and radar work since 1925, when he received the E.E. degree from Queens College, in Melbourne, Australia. Now a U. S. citizen, he was a British Navy captain during 1939 and 1940, and subsequently served as a radar liaison officer to both the United States and British Governments. In 1945 he was associated with Gilfillan Brothers, Inc., as principal radar design engineer until 1947, when he organized the division of which he is now chief.

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J. W. Head (A'44–M'48) was awarded the honorary LL.D. degree by Piedmont College, Demorest, Ga., at commencement exercises on June 5, 1949. The degree is being conferred for distinguished service as an educator, engineer, and practical scientist and is the highest honor the school can bestow. Mr. Head is president and founder of the Electronics Institute, Inc., of Detroit, Ill., and also directs the activities of Industrial Electronics, Inc., a professional consultant organization.

A graduate of Oglethorpe University in 1935, Mr. Head later became associated with Lee DeForest, the "father of radio." He was recently elected president of the Detroit Section of the Instrument Society of America, and holds membership in the Detroit Television Roundtable and the Engineering Society of Detroit.

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Roger E. Robertson (A'48) has been appointed to the staff of the National Bureau of Standards, where he will conduct engineering research for the guided missile projects of the electronics division.

A native of Jaffrey, N. H., Mr. Robertson won the freshman competitive scholarship to the Massachusetts Institute of Technology, and was the recipient of further undergraduate and graduate scholarships. While studying for the bachelor of science and master of science degrees in electrical engineering, both of which he received in 1942, he worked on co-operative assignments at the General Electric Co. Upon his graduation he joined General Electric on a full-time basis, leaving in 1946 to become an electronics engineer at the Bell Telephone Corp., where he directed and supervised the electronic phases of a flight-test program for guided missiles.

Mr. Robertson has worked on the design of fire-control radar and other electrical and electronic systems. He is a member of Eta Kappa Nu.

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Elmer W. Engstrom (A'25–M'38–F'40) vice-president of RCA in charge of research was given the honorary doctor of science degree by New York University on June 15, 1949.

George C. Schleter (A'38–VA'39) has been appointed to the staff of the National Bureau of Standards, where he will conduct an engineering development program on guided missiles, including missile systems and components, in the Bureau's electronics laboratories.

Mr. Schleter was born in Seymour, Ind., and attended Purdue University, from which he received the B.S. degree in 1922. He spent the next three years as a member of the faculty of New York University, from which he received the master of science degree in 1925. In that year he was appointed head of the mathematics and physics departments of Broaddus College in Philippi, W. Va., and he left in 1928 to become a civilian physicist on the staff of the U. S. Army Air Corps at Wright Field.

From 1931 to 1934 Mr. Schleter worked on the development of blind landing systems at the National Bureau of Standards; then he engaged in microwave research at the Naval Research Laboratory, as well as developing antennas for ASB radar, radio transmitters, and confidential microwave research. Early this year he rejoined the National Bureau of Standards.

Mr. Schleter is a member of Eta Kappa Nu.

❖

Everard M. Williams (S'36–A'41–SM'44), of the Carnegie Institute of Technology, has just been promoted to a full professorship at that institution.

Born in 1916, Dr. Williams studied at Yale University, receiving the doctorate in 1939. Shortly afterward he became an instructor at Pennsylvania State College, leaving in 1942 to become chief engineer of a secret radio weapons laboratory at Wright Field, Dayton, Ohio.

At the end of the war, in 1945, Dr. Williams joined the faculty of the Carnegie Institute. There he helped in the development of the new Synchro-Cyclotron, of which he designed the oscillator and deflector. In 1946 he was voted America's outstanding young electrical engineer for 1946 by the Eta Kappa Nu Assn. He was also awarded the President's Certificate of Merit for his war work.

❖

Paul Ware (A'17–SM'44) has been appointed head of the DuMont Laboratories' new electronics parts division.

From 1907 to 1912 Mr. Ware served as a wireless operator for Atlantic Deforest, American Deforest, United Wireless, and Marconi. He received the M.E. degree from the Stevens Institute of Technology in 1917 and subsequently enlisted as a sergeant in the U. S. Army during World War I. Rising to the rank of second lieutenant, he was responsible for the 75-meter break-in trench telegraph set known as SCR77, on which he had previously worked with Professor Hazeltine at Stevens.

From 1922 to 1925 Mr. Ware was engaged in operating one of the pioneer radio receiver enterprises which developed, manufactured, and sold over 200,000 receivers known as the "Ware Neutrodyne." He developed and patented a duplex radio communications technique embodying the break-in feature and many other improvements important at the time, including the first completely-shielded receiver and the double-oppositely-wound loop.

From 1925 to 1935 Mr. Ware was consultant for Splitdorf-Bethlehem, Sonora Phonograph, R. E. Thompson, and also for the Monmouth Memorial Hospital, and engaged in the design and manufacture of radio sets for various companies. In 1935 he joined P. R. Mallory and Co. as consulting engineer and remained there until 1939, when he transferred to DuMont.

In 1942 and 1943 Mr. Ware served as president of the Radio Club of America.

❖

Leslie J. Woods (M'35–SM'43) has been appointed vice-president and director of research and engineering to head all engineering and research activities of the Philco Corp.

Mr. Woods joined Philco in 1925, after serving in the British Army during the first World War, and playing an important role in the development of British communications in the Middle East. In 1928, when Philco began research work in television, Mr. Woods was named first television engineer, and served in this capacity for two years. From 1930 to 1938 he occupied positions of increasing responsibility in the company's engineering department, designing and developing both vacuum tubes and radio receiving sets. During this period he took charge of automobile radio engineering for the corporation, and in 1941 was made manager of the car manufacturers' division, with headquarters in Detroit.

On the outbreak of war he was transferred to Washington, where, in 1942 he became vice-president and general manager of the National Radio Corp., at that time a Philco subsidiary, to assist that company in expanding its organization to meet the greatly increased wartime demands of the army and navy. Following the war, he returned to Philco as manager of the company's industrial division, which takes care of automobile and aircraft radio, as well as advanced radar equipment for the armed forces, and in 1948 he was elected vice-president of the division.

❖

Daniel R. Donovan (A'44), formerly vice-president and sales manager of the Callite Tungsten Corp., has been appointed sales manager on elmet and fine wire products, for the North American Philips Co., Inc., in Lewiston, Me.

❖

George L. Downs (A'44) has been placed in charge of the transformer operation of Ratheon Manufacturing Co., at Waltham, Mass. Mr. Downs was formerly head of Raytheon's production test department, and methods and cost estimating department. A native of Texas, he is an amateur radio operator, his call letters being WICT.

Cover Sheet for Technical Memoranda—A Technique in Information Exchange*

R. C. MATHES†, SENIOR MEMBER, IRE

Summary—**An account is given of the origin and use of a routine for disseminating new technical information and ideas rapidly and flexibly through the large engineering and research organization of Bell Telephone Laboratories. The essential philosophy back of this routine is that technical memoranda have the status only of the individual engineer's personal authority, and hence may be circulated directly and widely across departmental lines for information and discussion. Departmental policy in relation to such memoranda is established by covering letters where such action is necessary. The objectives approached and benefits accruing from this procedure are briefly discussed.**

Introduction

ONE OF THE routines to which a new technical employee in Bell Telephone Laboratories is soon introduced is that of the "Memorandum for File," together with its inseparable companion, "Cover Sheet for Technical Memoranda." Such an individual may be tempted to regard this latter as a piece of bureaucratic red tape designed to supply work for stenographers and filing clerks, but in time he learns that it has definite relations to the organic needs of our type of work.

Starting in a small way about twenty years ago in a small research group working on carrier transmission, this cover sheet has had its use spread in turn to the department, the general department, and finally to all of our technical general departments. We will not trace out the little variations in its form and use from then to now, as a description of its present status will suffice to bring out the essential philosophy in back of its initiation and survival.

Without attempting to delve deeply into the nature of our Laboratories' organization or its technical work, there are two aspects of it which appear apposite to our present discussion. First, there is a vertical division of effort between such major functions as Research, Apparatus, and Systems. Each of these, in turn, is divided into other functional divisions, such as Physical Research, Chemical Research, Acoustic Research, Transmission Research, etc., which in turn are broken down into still more specialized compartments. It is in these specialized compartments that we strive for new knowledge, new types of apparatus, and new systems for application. The second aspect to our method of working is a horizontal one. Closely related lines of work must continually draw on each other for supporting information and specific projects, such as a new machine switching system or a microwave transmission system will need to be in touch with the work of a wide range of specialized workers. Within this simplified picture of the duality in methods of operation, it can be seen that there are numerous areas of work which need to be co-ordinated, continuously and flexibly. The technique to be described is one means to this end.

Objectives

From the above brief introduction to the nature of the problem, a little consideration will enable us to set up a group of objectives whose attainment we believe to be furthered by the use of the cover sheet.

1. Rapid dissemination of information to those most probably interested
2. The encouragement of direct contact between individual engineers
3. The avoidance of extensive duplication of effort.
4. The avoidance of vested interests in or undue isolation of fields of work
5. Speeding up reviews of past work

The two media which we have for the spreading of ideas are the oral and the written word. From the standpoint of speed and fluidity of interchange of ideas the former is the more effective, while the latter has its advantages from the standpoints of record and study. In the Laboratories, the use of the oral word is fostered by the medium of conferences in groups and between individuals with all the advantages attendant upon personal acquaintanceship. However in a large organization one can never be quite sure that all necessary or desirable personal contacts are established, and time alone precludes trying to establish all possible ones. This situation is rendered more serious by the fact that many, if not most, of the best engineers and research men are of the subjective type,[1] and do not venture far in establishing personal contacts.

The Mechanism

It thus becomes important to have a method for using the written word in the most flexible and direct fashion for furthering the objectives outlined above. The cover sheet, a typical example of which is shown in Fig. 1, has been evolved to supply this need. The first line is the subject and the case number under which expenditures for the work are authorized. Below this on the left is a routing list. If the memorandum is deemed to contain patentable material, the No. 1 copy, signed by the author, is forwarded to the Patent Department, generally with a covering letter containing additional discussion. The second, or yellow copy, also signed, goes to the Case File for permanent record. The third copy generally goes up the line of supervision to the department files. It is the remainder of the routing list which accomplishes our object of rapid information dissemination. Here the author lists the names of individual engineers in his own or other departments to whom he believes the subject is of direct interest. There may be only a few or there may be several dozen. If he believes that a certain group is or ought to be interested, but he does not know the individuals concerned or their responsibilities, he may route a copy to the group head. The engineer may confer with his immediate supervisor as to the general coverage of the list. It may cut across not only all department lines, but also all ranks from a general department head to individual engineer.

To the right of the page is a routine MM identification number, the date and the author's name. MM signifies "Memorandum for File." The identification number consists simply of the last two digits of the year, the department number, and the serial number of the memorandum for that department for that year. Under the author's name there is included a filing subject or type of work title.

The objective of rapid dissemination of ideas and knowledge is attained by virtue of the way in which the routing list cuts across all departmental lines. No time is lost in going five or six steps up and a similar number down with the inevitable delays due to pressure of other work, questions raised, etc. The objective of encouraging personal contact is attained because an individual of the subjective type (who may rarely speak up in a group conference) will, upon reading a report of special interest to him, reach for the telephone to make an appointment for a personal conference. His object may be to seek further information or to argue a divergent viewpoint, but either way a new friendship or improvement in mutual interests generally results.

Below the routing and filing information on the cover sheet there is a brief abstract stating the substance of the memorandum. This is of especial interest to supervisors and department heads. First, in relation to objective No. 1 they can review the routing list to see if every one in their group who should have a copy of that material is on the list. Second, it enables them to keep a running check on the third and fourth objectives having to do with interdepartmental division of effort and interdepartmental correlation.

The fifth objective is furthered if a file is kept by subject matter. We have done this in the past by keeping a file of cover sheets only arranged by subject or class of work. At present this type of file is being kept on 3×5 inch cards. An engineer reviewing or reopening a line of work thus has available a good brief guide to the more important memoranda to read. He also gets a good steer as to which individuals he should seek out first for personal contact.

A further check on the adequacy of dis-

* Decimal classification R009. Original manuscript received by the Institute, March 3, 1949.
This paper is an expansion of discussion prepared in relation to the paper by Allen H. Schooley, U. S. Naval Research Laboratory, entitled, "Information exchange as a management tool in a large research organization," presented at the National Electronics Conference, Chicago, Ill., on November 5, 1948, and published in Proc. I.R.E., vol. 37, pp. 429–432; April 1949.
† Bell Telephone Laboratories, Inc., Murray Hill, N. J.

[1] C. E. Broadley and M. Broadley, "Know Your Real Abilities," Chaps. 6 and 7. McGraw-Hill Book Co., New York, N. Y. 1948

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BELL TELEPHONE LABORATORIES INCORPORATED

COVER SHEET FOR TECHNICAL MEMORANDA

SUBJECT: Terminology for Semiconductor Triodes - Committee Recommendations - Case 38139-8

MM-48-130-10
DATE May 28, 1948
AUTHOR L.A.Meacham
C.O.Mallinckrodt
H.L.Barney

Surface States - Terminology

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ABSTRACT

Recommendations are made for an equivalent circuit representation, and terminology relating to semiconductor triodes.

Fig. 1

tribution is maintained by the monthly circulation, to a certain level of supervision, of a list of titles and authors of all technical memoranda for file. Several sets of circulation file copies are maintained for use on a reference basis to take care of requests resulting from this circulation list. This freedom of circulation is only rarely restricted at certain stages of special jobs and, of course, in the case of work being done on classified government projects. In such work the routing list includes only authorized individuals, and the MM title is withheld from the monthly circulation list.

Status of Technical Memoranda

It may be objected that this free circulation of memoranda might induce confusion between the work of departments. This is avoided by the status accorded these memoranda. They are regarded as the technical expositions of his work by the individual engineer and carry only his personal authority. Thus, a supervisor may approve[2] them for file and circulation, even when not in full accord with some of the ideas expressed. They are only made a part of interdepartmental policy and action when transmitted with covering letters signed by supervision and to the degree set forth in each letter.

From this independent status given to the technical memoranda for file, there derive benefits quite aside from the original objectives of dissemination of information. These memoranda become, for the individual engineer, a medium of publication to a large group of his associates, not only of the results of work done, but also of his judgment of the significance of the work and of his plans and hopes for the future. Sympathetic supervision can thus encourage the development of the individual through constructive criticism of this medium of expression. The high standard attained in the writing of many of these memoranda is testified to by the fact that, not infrequently, they become, with only slight changes to adapt them to a new and wider audience, the manuscripts for publications in the technical press.

[2] There is some feeling that even this degree of approval is not necessary.

Radioactive Standards and Methods of Testing Instruments Used in the Measurement of Radioactivity*

L. F. CURTISS†

Summary—**A summary of the program for standardization of radioisotopes at the National Bureau of Standards is given in this paper, with some of the reasons for need of absolute measurements of radioisotopes in particular uses. This is accompanied by an outline of some of the methods used to ascertain characteristics of commercial instruments used in the measurement of radioactivity, with particular emphasis on those methods used at the National Bureau of Standards for this purpose.**

Introduction

THE MEASUREMENT of radioisotopes has assumed new importance in recent years. This can be attributed to the increased production of artificially produced isotopes, which, in turn, has led to a greater diversity of uses. The expansion in the requirements for measurements has developed much faster than the development of equipment for the purpose, both in quantity and in quality. Instruments commonly used in measurements of radioactive radiations rarely have been of the type which could be reliably calibrated quantitatively with the expectation that such calibrations could be relied upon over long periods of time. As more sensitive and complex instruments have been developed, the instability has increased. Therefore, many difficulties are encountered, which are chiefly of an instrumental nature, where absolute measurements of activity are required. While this situation may be said to result indirectly from the increased amounts of various radioisotopes now available, it arises chiefly from new uses which have been found for these isotopes, where absolute measurements are essential.

Before radioisotopes were available in quantities to encourage practical uses, they were mainly the subject of investigations regarding their structure, modes of disintegration, and related basic information. Measurement in this field can be accomplished without any knowledge of the absolute amount of the isotope under study. Each investigator sets up his own arbitrary scale of units, based on instruments and techniques usually available only to himself. The only comparison of his results with those of others, using different quantities and methods, is made in the conclusions deduced from the measurements. In many tracer experiments

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† National Bureau of Standards, Washington, D. C.

a similar situation exists. An arbitrary scale of activities set up for a particular experiment suffices to yield the desired quantitative results.

There is one radioisotope which has required quantitative measurements for many years. This radioisotope is radium. Accurate standards[1] of radium were set up internationally in 1913. A survey of the principal reasons for this procedure throws considerable light on the nature of the present demand for absolute measurements for artificially produced radioisotopes. The more important of these are that (1) radium has become an article of commerce; (2) it is used principally in the treatment of disease; (3) certain physical investigations required absolute measurements of this isotope. An example of (3) is the determination of the disintegration rate of radium.

The same situation regarding radium is now true of many of the artificially produced radioisotopes, and the need for their absolute measurement has the same general basis. This has created a demand for standards of a large number of radioisotopes in forms suitable for calibration of a wide variety of instruments. The more recently developed instruments for radioactive measurements are more sensitive than the earlier instruments. They also exhibit a greater degree of instability, and have more complex components which may fail to function properly. Therefore, the usual procedure in measurements of radioactivity—that of frequently checking the response of an instrument by using a standard—has become more necessary with present-day instruments. Most of these instruments are linear in their response only over limited ranges. This necessitates a series of standards of different intensities to reduce errors which might be caused by this lack of linearity in the instruments.

Accurate standards will not, by themselves, solve the problem of accurate absolute measurement. It is obviously equally important that the instruments function reliably within their limitations, and that these limitations be clearly recognized. As the instruments have increased in complexity, as compared with the gold-leaf, or Lauritsen, electroscope, with which many important measurements have been made, the number of tests which must be made to ascertain proper operation has increased. The testing of instruments is thus a new requirement in radioactive measurements. Until quite recently, practically all equipment used in radioactive measurements was constructed in the laboratory where it was to be used, usually by the investigator himself. Thus, the construction and operational characteristics were entirely familiar to the user. This situation no longer obtains. Measuring equipment is now being produced commercially. Numerous companies are engaged in this production, with no two companies making a given instrument in the same way. The prospective user must, therefore, before he uses a particular instrument, ascertain whether it operates properly and determine its peculiar limitations. This must be accomplished either by tests which he makes himself or which are made for him. There is a sufficient diversity in instruments made by a single manufacturer to necessitate individual tests for each instrument. In many situations these tests must be made for the user, who frequently may not have the equipment and training to make them for himself. Thus we see that standards of radioactivity, and tests of instruments with which these standards are used, are both essential to absolute measurements of radioactivity.

It is the purpose of this paper to describe some of the standards of radioisotopes already available; to outline the program for producing additional items in this list, and to discuss some of the precautions required in the correct use of these standards. This is followed by a description of some of the tests required for a few types of instruments, including the equipment used for this purpose.

Standards of Radioactivity

A radioactive standard is conveniently defined as a specimen of a given radioisotope for which the amount of the isotope present has been determined quantitatively. An illustration is provided by the National Radium Standard which is deposited at the National Bureau of Standards. In 1913 Madame Pierre Curie prepared some very pure radium salts which were used to determine the atomic weight of radium. Several hundred milligrams of these salts were used. After the work of determining the atomic weight was completed, this material was divided into a number of portions, each portion carefully weighed and sealed in a glass tube. These were then primary standards of radium, since the number of milligrams of the isotope radium in each tube had been accurately determined. This method of preparing a standard could be used for radium since the amount of radium, in equilibrium with its decay products, which produces a conveniently useful amount of radiation is of the order of several milligrams. This is a direct result of the slow rate of disintegration of radium.

A little consideration shows at once that the gravimetric method of determining the amount of radioisotopes will fail for those isotopes which disintegrate more than 100 times as fast as radium. We might consider Co^{60} as an example. A sample of radium containing 10 mg of the isotope will have 2.66×10^{19} atoms of the isotope. The number of atoms disintegrating per second will be this number multiplied by λ, the decay constant for radium. This constant has the value 1.36×10^{-11} per second. Hence, the disintegration rate per second for 10 milligrams of radium is $2.66\times10^{19}\times1.36\times10^{-11}=3.61\times10^{8}$. We will now compute the weight of an amount of Co^{60} which has the same rate of disintegration. The decay constant for Co^{60}, assuming a half-period of 5.3 years, has the value of 4.15×10^{-9} per second. The number of atoms required is, therefore,

$$\frac{3.61 \times 10^{8}}{4.15 \times 10^{-9}} = 8.70 \times 10^{16} \text{ atoms.}$$

Since there are 10^{22} atoms in a gram of cobalt, the weight of this amount is

$$\frac{8.7 \times 10^{16}}{10^{22}} = 8.7 \times 10^{-6} \text{ grams.}$$

A microchemical balance of highest sensitivity, used under the most exacting conditions, would be required to determine this weight to an accuracy of a few per cent.

Co^{60} has a half period of 5.3 years. Many radioisotopes in use today have half-periods of a few days. The gravimetric method is out of the question for them. This leads to a consideration of radioactive methods of determining the amount of a radioisotope. We already have an example of this in the members of the decay chain of radium. For example, polonium has a half-period of approximately 140 days. Although it cannot be weighed, its rate of disintegration can be determined by measuring the rate at which alpha particles are emitted. The amount of polonium in equilibrium with 1 gram of radium, in analogy to the case of radon, is defined as 1 curie of this isotope. The rate of disintegration of 1 curie is the same as for 1 gram of radium; that is, 3.61×10^{10} disintegrations per second. Hence, the number of millicuries in a preparation of polonium can be ascertained by measurements of the radioactivity in terms of rate of disintegration. If necessary, the number of atoms present can be computed by dividing this rate by the disintegration constant.

In principle, this method of measurement may be extended to any radioisotope. However, the details become more complicated when applied to those isotopes which do not disintegrate by the emissoin of alpha particles. Those isotopes which emit beta rays or positrons, or which disintegrate by electron capture, frequently have complicated disintegration schemes which do not permit a simple computation of the disintegration rate from measurements which can be made of the radiations emitted.

Even in the simple case of an isotope which disintegrates only by emitting beta particles of a single maximum energy, the nature of the continuous spectrum of nuclear beta rays introduces a difficulty. These beta rays are emitted with all energies from zero to the maximum. Any convenient detector for measuring the rate at which they are emitted will fail to detect those below a certain minimum energy. This energy depends on the characteristics of the detector and the amount of absorber between the source and the active volume in which the particles are detected. Furthermore, these particles are emitted with an equal probability in all directions. This requires an accurate determination of the solid angle subtended by the detector window at the source. Proceeding to the more complicated modes of disintegration, where two or more beta rays of different maximum energies are emitted, each followed by one or more gamma-ray transitions before the stable, nonradioactive state is reached, other difficulties appear.

One of the most serious is that some of the gamma rays may be internally converted. This process consists in the ejection of a shell-electron from the atom by a gamma ray from the nucleus. Therefore, in place of a gamma ray, an electron is emitted with an energy equal to that of the gamma ray decreased by the binding energy of the electron in its shell. These conversion electrons thus have energies which are usually comparable with those of individual beta rays. In ordinary measurements with counters, they can-

[1] See definition in section on Standards.

not be distinguished from beta rays. This leads to erroneous determinations of disintegration rates, unless an accurate correction is made for the conversion electrons.

It is obvious that disintegration rates cannot be determined unless all details of the disintegration scheme are available. Fortunately, the number of radioisotopes for which disintegration schemes are known is steadily increasing. At present, however, they represent a small fraction of the known isotopes which might have practical use.

The general situation is such that the average worker who needs to measure radioisotopes in absolute terms might well be discouraged. His difficulties would be removed if he had standards for each isotope to be measured. The measurements would then consist in mounting the sample to be measured in any convenient way with respect to the detector, and observing the effect. A standard of suitable strength mounted in the same way would give the data to compute the disintegration rate of the sample. This procedure transfers many of the complications in regard to the measurement of disintegration rates to the laboratory which prepares the standards.

The short half-lives of many of the artificially produced radioisotopes do not permit the preparation of permanent or semipermanent standards. These can be prepared for such isotopes as Co^{60} and Na^{22}, where the half-life is of the order of years. For those isotopes having half-periods measured in days, resort must be made to standardizing expendable samples which are distributed by a standardizing laboratory at sufficiently frequent intervals to maintain the calibration of the measuring equipment.

Calibration of Standards

Various methods are available for the calibration of standards. We will discuss typical examples of the more important of these to illustrate the principles.

One of the most reliable methods of determining absolute disintegration rates is the coincidence method. Perhaps an explanation of the meaning of the term "coincidence" in this connection is justified. By referring to Fig. 1, which shows the disintegration scheme of Co^{60}, we note that the disintegration occurs by the emission of a single beta ray which transforms the nucleus into Ni^{60}, and that the nucleus of the newly formed nickel atom emits two gamma rays before it becomes stable. The nucleus is then said to be in its ground state, which means that it has no internal energy in excess of that possessed by a stable nonradioactive nucleus. Actually, there must be a finite interval of time between the emission of the beta ray and the subsequent emission of the gamma rays, since we know that the shell-electrons in the newly formed nickel atom have had time to arrange themselves in the levels appropriate for this atom, which differ distinctly from the levels of the cobalt atom. However, this interval of time is too short to measure by any conventional methods, and is probably of the order of 10^{-12} second. Consequently, for all practical purposes the gamma rays instantaneously follow the beta ray with which they are associated. This means that, under favorable conditions, it should be possible to

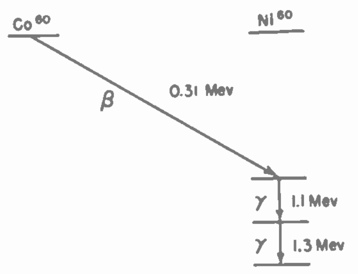

Fig. 1—Disintegration scheme for Co^{60}. In this disintegration a single beta-ray spectrum is emitted with a maximum energy of 0.3 Mev. Each newly formed nucleus of Ni^{60} emits two gamma rays, immediately bringing the nucleus to the ground state.

detect a gamma ray which is emitted simultaneously, within the limits of the resolving power of usual arrangements for detection, with a beta ray. This supposition has been confirmed. The determination of the relative number of these simultaneous events is called a coincidence measurement. In addition to beta-gamma coincidences, it is obvious that gamma-gamma coincidences occur in the disintegration of Co^{60} and, in principle at least, it should be possible to observe them. The disintegration of Co^{60} follows what is called a simple disintegration scheme, since it involves only one beta ray. In more complex schemes the details can frequently be unravelled by a study of the coincidence rates between the various radiations emitted. The coincidence method can only be used for those isotopes emitting beta rays followed by one or more gamma rays. It is most reliable when no measurable amount of the gamma rays suffers internal conversion.

Consider an arrangement in which a beta-ray counter and a gamma-ray counter, respectively, detect beta particles and photons from the same source. These counters are connected to electronic equipment which will record the photons and beta particles independently. It also records the coincidences between the two counters. Fig. 2 shows a block diagram of the equipment. The observations are corrected for counter backgrounds, chance coincidences, and cosmic-ray coincidences. Let N be the actual disintegration rate of the source, A the rate of the gamma counter, and B and rate of the

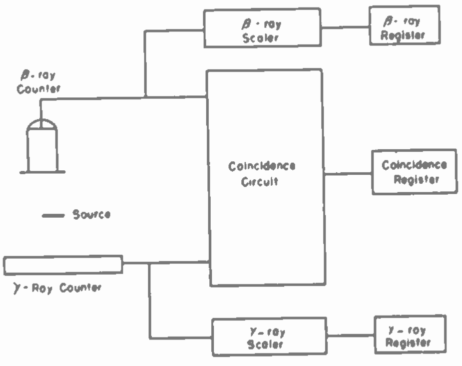

Fig. 2—Block diagram of a coincidence circuit for determining β-γ coincidences.

beta counter; E_γ represents the over-all efficiency of the gamma-ray counter, and E_β that of the beta-ray counter. For present purposes we need not know the actual value of these efficiencies, which are defined as the fraction of the total number of particles or photons emitted which are detected by the respective counters.

Then we have

$$A = NE_\gamma \quad \text{or} \quad E_\gamma = \frac{A}{N}$$

$$B = NE_\beta \qquad E_\beta = \frac{B}{N}$$

If C is the coincidence rate,

$$C = NE_\gamma E_\beta = N\frac{A}{N}\frac{B}{N},$$

whence

$$N = \frac{C}{\frac{A}{N}\frac{B}{N}} = \frac{N^2C}{AB}$$

or

$$N = \frac{AB}{C}.$$

Thus the absolute disintegration rate is obtained by dividing the product of the rates of the rates of the two counters by the coincidence rate. The efficiencies do not appear in the final result.

For an isotope which emits only beta rays, the absolute disintegration rate can be obtained by counting all particles in a definite geometry under conditions where all absorption and scattering effects have been reduced to a negligible amount. The simplest geometry is that which includes all particles in the 4π solid angle about the source. In practice, this is a little difficult to obtain, since the source must be supported. The effect of the support can be reduced by making it very thin. One satisfactory way to make measurements of this type is to use a large evacuated chamber with the source mounted on a very thin support at the center. The rate at which the source and support acquire electric charge is then a measure of the disintegration rate. This requires moderately strong sources for accurate results. A modification of this principle of using a definite geometry is to set up a series of apertures which define a small solid angle, which can be computed from the dimensions. The region in which the beta rays travel to the detector is evacuated. If the detector has a window of a thickness less than 0.5 mg/cm² and the maximum energy of the particles is of the order of 1 Mev, only a small fraction of the particles emitted in the defined angle will miss detection as the result of absorption. This method is not very satisfactory for beta rays whose maximum energy is much less than 1 Mev.

The problem of calibrating standards in terms of disintegration rates becomes more difficult when we deal with radioisotopes which emit beta rays in two or more continuous spectra with different maximum energies, each followed by several gamma rays. A very careful study of the disintegration scheme is required in order to devise a coincidence arrangement which will give measurements which can be interpreted in terms of disintegration rates. This problem is further complicated when some of the gamma rays undergo internal conversion. The conversion electrons usually will be of sufficient energy to be detected in the beta-ray counter. The relative number of these conversion electrons must be determined, and a correction made for their contribution to the particles recorded by the beta-ray counter used.

The problem of producing standards for radiosotopes has become much more complicated than that of preparing standards of radium, as far as physical measurements are concerned. The chemical balance has been replaced by instruments for determining the rate of disintegration of the sample undergoing standardization by measurements on the radiation emitted. Since many radioisotopes do not have half-periods which permit the preparation of permanent standards, the production of standards is a continuous process which must be repeated periodically.

The demand for radioactive standards is naturally greatest for those radioisotopes which are used in relatively large amounts, under conditions where absolute measurements are required. A tentative list of such radioisotopes is given below.

	Half-Period
C^{14}	about 6000 years
Na^{24}	14.8 hours
P^{32}	14.3 days
S^{35}	87.1 days
Ca^{45}	180 days
Fe^{59}	42.5 days
Co^{60}	5.3 years
I^{131}	8.0 days

All of the isotopes in the above list can be produced in a pile. Thus they can be made available in large amounts, which accounts in part for their extensive use. If radioisotopes are produced in comparable quantities in cyclotrons, there are a number of other radioisotopes which may become widely used. The most important of these, at present, is Na^{22} (3.0 years).

C^{14} and I^{131} are used extensively for biological and medical investigations, and standards of these isotopes are therefore needed. The National Bureau of Standards has a program for the development of standards for C^{14}. At present it appears that these standards will be most useful if prepared as calibrated expendable samples, rather than as permanent standards. Although the long half-life would permit permanent standards, it is difficult to prepare samples of compounds containing radioactive carbon in which the ratio of the active to the inactive isotope remains constant. For example, solid deposits of barium carbonate might be used as standards. In use these would be exposed to the atmosphere, and it is known that such samples exchange CO_2 with the atmosphere in the presence of moisture. Therefore, they would gradually lose C^{14} atoms with use, even if stored in a dry atmosphere when not in use. The problem of providing specific forms of standards for all types of possible uses becomes so complicated that some simplification is required. Some investigators need to use the standard in the form of CO_2, and others need to prepare their working standards to conform to the requirements of a specific type of measuring equipment. For the present, it is proposed to furnish these standards in the form of calibrated solution of sodium carbonate sealed in glass ampoules. The calibration will give the disintegration rate per milliliter, so that definite amounts of this solution can be removed and converted into a working standard of the desired form.

In the case of I^{131}, the short half-period precludes the preparation of permanent standards, and calibrated expendable samples must be used. Although I^{131} emits both beta rays and gamma rays, and therefore should be susceptible of calibration by coincidence methods, the disintegration scheme is complicated by the presence of two beta rays and a number of gamma rays, some of which undergo internal conversion to a measurable degree. When the details of this scheme are completed, a satisfactory coincidence method for calibration can be made available. Present information allows calibration to ± 5 per cent.

The National Bureau of Standards has made several distributions of identical samples of I^{131} over the past two years. The reported results reveal a disagreement in measurements approaching a factor of 2. These disagreements are to be attributed, in part, to the lack of suitable standards with which to compare the activity of samples of I^{131}. It has been indicated that the use of a standard of one isotope to calibrate a quantity of a different isotope may lead to serious errors. Therefore, if one attempts to measure I^{131} with a RaD+E standard which has a very different beta-ray spectrum from that of iodine, a considerable number of corrections and precautions are required to obtain results that are not in error by more than 20 per cent.

It is, therefore, obvious that many of the possible sources of error can be eliminated, if calibrated samples of I^{131} are used for the measurements. Using such samples, the results should be free from serious errors if the unknown and calibrated samples are compared under identical conditions. If this requirement is complied with, no corrections need to be applied to the data used in computing the disintegration rate of the unknown sample. This reveals the need for a distribution of calibrated samples of this isotope at intervals of the order of every few months. Work is in progress to prepare and distribute such samples.

P^{32} emits only beta rays in a single spectrum with a maximum energy of approximately 1.7 Mev. This isotope therefore cannot be calibrated by coincidence measurements. The most convenient method of calibration is counting the beta particles in definite solid angle with corrections for absorption and scattering. Those using P^{32} can make a fairly accurate measurement of quantities deposited by evaporation, using the RaD+E standards prepared by the National Bureau of Standards. These standards contain a measured amount of RaD. Within a month of perparation, RaE grows into equilibrium with the RaD, so that the disintegration rate of the RaE is known. Fig. 3 shows one of the standards and two blank disks on which standards are deposited. The RaD is deposited on the palladium face of a silver disk 1/16 inch thick and 1 inch in diameter, faced with a layer of palladium

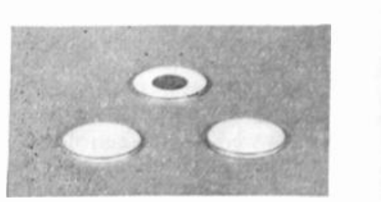

Fig. 3—RaD(Pb^{210})+RaE(Bi^{210}) standards. Two of the disks are blank and the third has a circular deposit of lead, part of which is Pb^{210}. Approximately one month after preparation, RaE is in equilibrium with the RaD, after which the activity decays with the half-period of RaD (22.2 years). Blank silver disks are used to mount deposits for comparison of activity with the standards.

0.002 inch thick. The deposit is circular and approximately 12 mm in diameter. It is located centrally on the face of the disk. This disk provides saturation back-scattering of the beta particles. If the phosphorus is deposited on a similar disk the back-scattering correction is reduced to a few per cent. These RaD+E standards can be used for moderately accurate measurements of isotopes which disintegrate only by beta ray or positron emission when the spectrum is simple and has a maximum energy near that of the RaE spectrum. Appropriate absorption corrections must be applied in each case. The absorption correction is determined experimentally at the time of comparison by extrapolating the beta-ray absorption curve to zero thickness of absorber.

It should be noted that these standards emit soft beta rays from RaD, and alpha particles from the polonium which is growing in them. Therefore, they must always be used with a filter which excludes these radiations from the counter. A layer of aluminum 0.001 inch thick is adequate for this purpose.

Fig. 4 shows a series of radium gamma-ray standards consisting of 5 milliliters of solution sealed in Pyrex ampoules. Ranging from 0.1 micrograms to 100 micrograms, they are intended for gamma-ray measurements of radium. Since the solution does not completely fill the ampoule, radon collects in the air space above the solution, producing an uneven distribution of activity. This disadvantage can be overcome by shaking the solution vigorously several minutes prior to use.

Radium in amounts less than 10^{-8} grams is most accurately determined by measuring the production of radon from the radium salt in solution. The standards required for this measurement consist of a solution containing a known amount of radium. Fig. 5 shows two standards and a blank solution. The radium is contained in 100 milliliters of solution, one standard having a total radium content of 10^{-9} grams, the other 10^{-11} grams. Nearly all water, no matter how carefully distilled, contains measurable amounts of radium. Therefore, ordinary distilled water cannot be used to rinse the flasks containing the standards when transferring the solutions to the apparatus in which it is to be used. To do so might involve a risk of adding several per cent to the amount of radium in the standard. This difficulty is removed by the use of

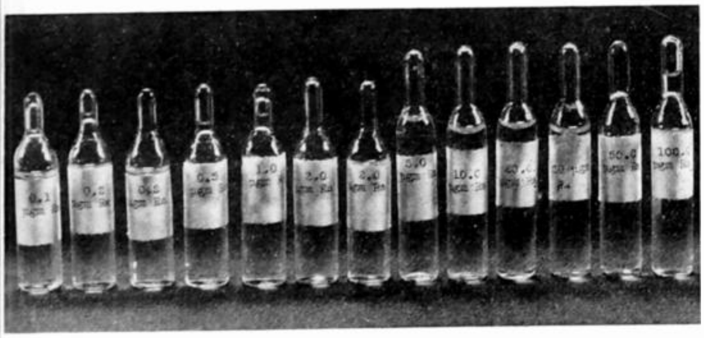

Fig. 4—Gamma-ray standards. Microgram standards of radium. The standards cover a range from 0.1 to 100 microgramsof radium, and are used for gamma-ray measurements of radium.

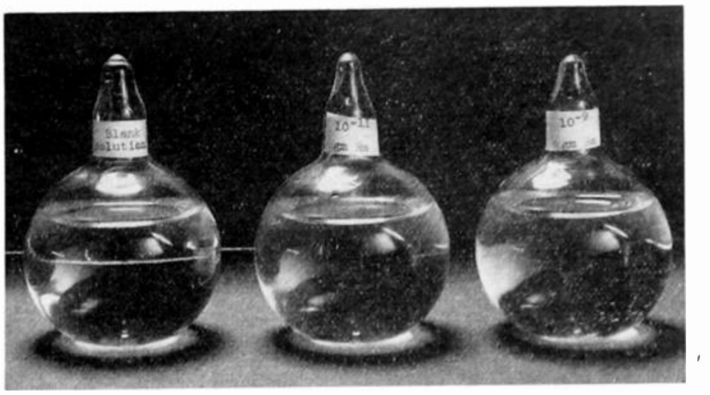

Fig. 5—Radon standards. These flasks contain 10^{-11} and 10^{-9} grams of radium in 100 milliliters of solution. The blank solution has a measured low radium content, and is used for rinsing when transferring the standard solution to the apparatus in which it is to be used.

"blank" solutions sealed in flasks similar to the standard. The amount of radium in these solutions has been measured and found to be 0.25×10^{-12} grams of radium. This amount is insignificant compared with 10^{-9} grams of radium in the larger standard, but it must be taken into account when used with the 10^{-11} gram standard.

The only other isotope for which the National Bureau of Standards has prepared calibrated standards at present is Co^{60}. This isotope lends itself readily to the coincidence method of calibration. The current standards are gamma-ray standards, since they consist of a 5-milliliter solution sealed in a glass ampoule. The solution from which these standards were prepared was measured in three different laboratories by the coincidence technique with agreement in the results of the order of 1 per cent. They are available in two sizes; 1.5 rutherfords and 0.15 rutherford. Typical ampoules are shown in Fig. 6.

The rutherford (rd) is a unit recommended by the National Bureau of Standards for expressing quantities of radioisotopes other than members of the radium family. It is defined as the amount of a radioisotope undergoing 10^6 disintegrations per second. It is, therefore, a decimal unit which

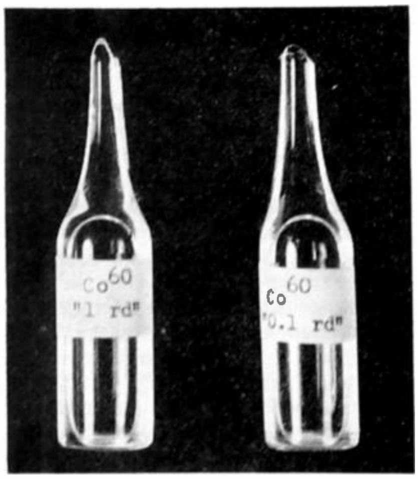

Fig. 6—Co^{60} standards. The cobalt solution is sealed in a glass ampoule which then serves as a gamma ray standard for measurement of solutions of Co^{60}.

facilitates computations. It is also of a convenient magnitude for expressing quantities of a radioisotope usually handled and measured in the laboratory. For example, a sample which can be conveniently measured by an end-window Geiger-Müller counter is of the order of 0.1 to 0.4 millirutherford. Tracer quantities are of the order of 1 rutherford. Therapeutic doses are of the order of 100 rutherfords. In addition to covenience there is another reason for the introduction of a unit which is uniquely defined in terms of disintegration rates. This is the widespread misuse of the curie. It has become fairly common to use the curie to refer to quantities of radioisotopes when the disintegration rate has not been measured. It is even used frequently when disintegration rates cannot be measured. It has been pointed out that a complete knowledge of the disintegration scheme is necessary before a disintegration rate can be measured by radioactive methods. Therefore, when we see quantities of isotopes expressed in curies for which disintegration schemes are either partially or wholly unknown, we can only conjecture as to the quantities intended. Generally, such statements have no meaning and only lead to confusion. Finally, there is still considerable uncertainty regarding the value of the curie when used as a unit disintegration rate. Logically, from its original definition as "the quantity of radon in equilibrium with 1 gram of radium" it has the value of the disintegration rate of 1 gram of radium. Values for this figure varying from 3.4 to 3.8×10^{10} have been reported by competent investigators. The most recent value is 3.61×10^{10}. The value 3.4×10^{10} is still in current use. An attempt to smooth over this situation has been made by adopting 3.700 $\times 10^{10}$ as an arbitrary figure. However, recently Madame Joliot-Curie has proposed an "international curie" to have the arbitrary value 3.600×10^{10}. The result of this situation is that it is incumbent on each author of a paper to state the value of the curie which he uses. The rutherford eliminates this difficulty, since it is not associated with any natural constant, the value of which may fluctuate with the precision of measurement, or the arbitrary selection of different values by different authors.

Testing of Instruments for Radioactive Measurements

Instruments used in the measurement of radioactivity may vary from the simple gold leaf or quartz-fiber electroscope to the complicated assembly of electronic equipment required to operate a Geiger-Müller counter and record the pulses from it. Devices which use ionization chambers as the sensitive element also frequently employ electronic circuits to detect the required direct-current potentials produced in the measurement of the ionization currents. In the more complex assemblies, over-all checks may reveal either satisfactory or unsatisfactory operation, but in cases where performance falls short of that expected or required, detailed tests must be made of each component. In any case, a laboratory testing equipment of this type must be prepared to test the various components, since they are frequently purchased separately from different manufacturers.

The nature of the problems encountered can be presented by a discussion of a few typical cases. A good example is the equipment which is required for the operation of a Geiger-Müller counter when the actual number of pulses are to be recorded. The block diagram in Fig. 7 shows the essential components of this equipment. They may be arranged in separate units as indicated in the figure, or these components may all be assembled in a single unit. Although the trend is toward consolidation, there are commercial firms which manufacture scalers only, others which make mechanical registers, and still

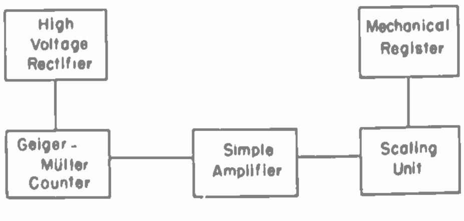

Fig. 7—Block diagram showing essential components of Geiger-Müller counting equipment.

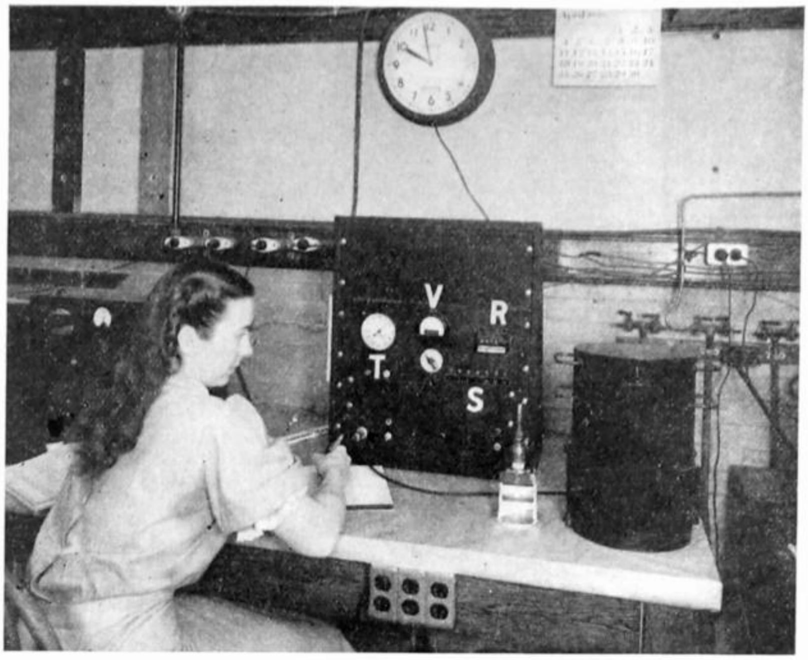

Fig. 8—Geiger-counter scaler assembly with all components in a single chassis. *T*—electric timer; *V*—high voltage voltmeter; *M*—mechanical register; *S*—interpolation lamps of scaler.

others that make only electric timers. Also, there are a number of firms which make Geiger-Müller tubes but none of the accessory equipment. The nature of the tests required on each component will be given. A scaler in which all components are mounted in a single chassis is shown in Fig. 8.

We will start with the detector of radiation, the Geiger-Müller tube. These tubes fall into general classifications which differ in structural form, depending on whether they are designed to detect and record gamma rays or beta rays. Although a counter designed to detect beta rays is also sensitive to gamma rays, its efficiency for detecting beta rays is of the order of 50 to 100 times that for gamma rays, so that the effect of gamma rays is usually neglected. Gamma-ray counters are usually totally insensitive to beta rays of the usual energies. The basic principles on which these two types of counters operate are the same. Consequently, some tests are common to both types. One of the principal tests which is common to the two types is related to the characteristic curve of the Geiger-Müller counter, or plateau, and pulse height.

The plateau curve shows the pulse rate of a counter plotted against the voltage applied to the counter for a steady intensity of radiation on the counter. A typical plateau is shown in Fig. 9. The voltage *T* at which the counter begins to respond to the radiation is referred to as the starting voltage of the counter. Above this voltage the counting rate rises rapidly for a short interval of voltage, and then at *A* levels off in a gradually sloping line to a point *B*, where further increase in voltage produces another rapid rise in counting rate. The portion *A* to *B* of the curve is called the "plateau." The desirable features for this characteristic are that the value of *T* should remain fixed for a given counter; that the slope of the curve from *A* to *B* should be small (of the order of 1 or 2 per cent per hundred volts), and that the length of the plateau *AB* should extend over 200 to 300 volts In addition, the counter is required to give a single clear-cut pulsef or each response to radiation. Multiple pulses are objectionable since they reveal that the counter is not functioning as a true Geiger-Müller counter, and because they either increase the dead time of the counter or falsify the measurements.

The dead time of a Geiger-Müller counter is the time after the production of a pulse during which the counter is in a state where it cannot produce a pulse, even when radiation has produced ions in the sensitive volume. This dead time is a direct result of the quenching mechanism of the counter which causes a discharge in the counter to cease abruptly, resulting in a definite electrical pulse. This time varies from about 10^{-3} to 10^{-4} second for current types of counters. It is desirable that the dead time be as short as is consistent with proper operation of the counter, since the length of the dead time determines the number of lost counts at a given steady average counting rate. If the counter is to be operated at rates where the lost counts are a measurable percentage of the total counts, the length of the dead time must be known to make a computed correction to the observations, in order to obtain the true number of counts.

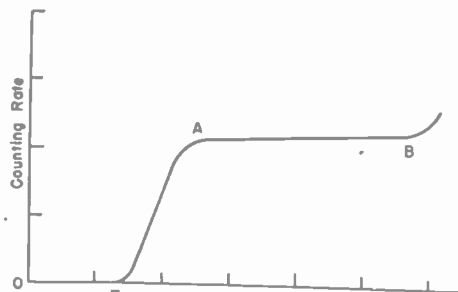

Fig. 9—Plateau of the Geiger-Müller counter. *T*—starting voltage; *A–B*—working region.

From considerations similar to those outlined above, we arrive at the following tests which are required of all types of Geiger-Müller counters for their use in quantitative measurements.

1. *Determination of Background Counting Rate:* The actual background counting rate of a counter may be considered as made up of three factors: (a) the intrinsic background rate; (b) response to cosmic rays; and (c) the response to local distributed radioactivity. The actual rate will depend on the condition under which the counter is used. Response to cosmic rays and local radioactivity can be controlled to a considerable extent by shielding. They are, therefore, under the control of the user to a certain degree. He has no control over the intrinsic background rate, which is produced by radioactive radiations originating within the counter. This rate can be controlled only at the time the counter is constructed by selecting materials with low natural radioactivity, and by using careful techniques to avoid the introduction of radioactive contamination into the counter. The background rate controls the lower limit of the net detectable counts from a source, which for many applications is important. Therefore, the intrinsic background rate should be low.

The test for intrinsic background rate is the obvious procedure of placing the counter inside a heavy lead shield to screen off local radioactive radiations and to reduce the response to cosmic rays. In this way the lowest background rate which can be expected under laboratory conditions is obtained. If the intrinsic background is excessive, this test reveals it immediately. It is obvious that the intrinsic background depends on the area of the interior surfaces of the counter. Beta-ray counters with a tube of about 1 inch diameter and 2½ inches long may have an over-all background when shielded with 2 inches of lead, of from 0.3 to 0.5 counts per second.

2. *Determination of Plateaus for Geiger-Müller Counters:* The procedure for obtaining the plateau of any type of counter is essentially the same. The measurements are made with the counter exposed to a constant source, which gives a convenient counting rate. A scaler and mechanical register are used to record the counts. Before measurements can be started, it is necessary that the high voltage be applied to the counter and adjusted to the value at which the counter begins to count, called the starting voltage. This voltage is sometimes called the threshold voltage, but it seems better to reserve the latter term for the voltage at which all pulses first become of uniform height in the counter. The threshold thus defines the beginning of the true Geiger-Müller region. The starting voltage must be applied to the counter for a period of 15 to 20 minutes to allow electric charges on various insulating parts to assume steady values. This is true of both the beta-ray and gamma-ray types of

counters. During this period the starting voltage is subject to slow changes. When this starting voltage has become constant, it is safe to assume that static charges have reached an equilibrium value.

The procedure for making the observations consists in raising the high voltage by approximately equal arbitrary steps, of the order of 25 volts. At each voltage the time required to record a fixed number of counts is observed. The number of counts depends on the statistical accuracy required. For a standard deviation of 1 per cent, a total of 10,000 counts must be recorded for each point. In recording the data for the plateau it is essential to watch for a sudden increase in the counting rate carefully as the voltage is increased. Most counters in use today are readily ruined if voltages above those corresponding to the plateau are applied. The resulting excessive rate quickly destroys the organic vapor used for quenching.

The usefulness of a counter depends on the length of the plateau and its slope. The slope may be defined as the percentage change in counting rate per 100 volts change in the high voltage. A counter can be expected to have a slope of the order of 2 or 3 per cent per hundred volts over a range of at least 200 volts.

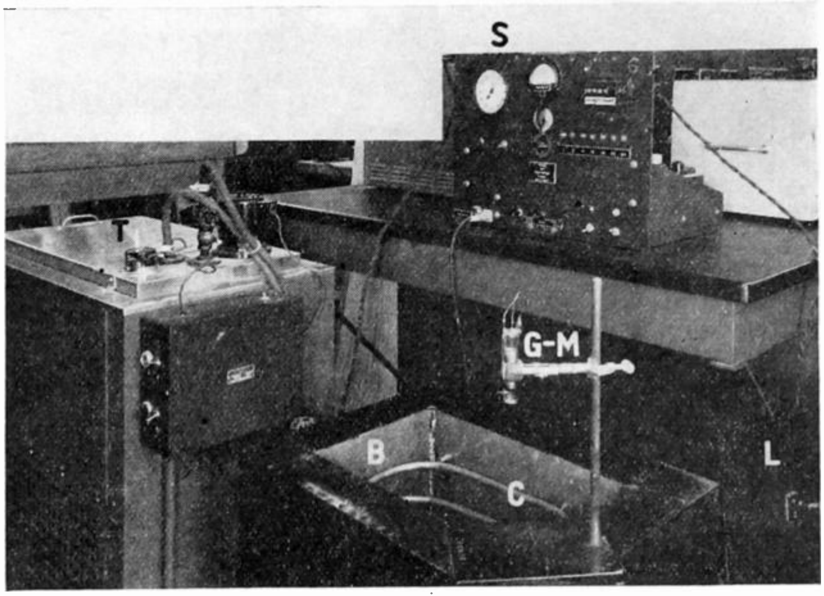

Fig. 10—Chamber with controlled temperature for determining the effect of temperature on Geiger-Müller counters. *T*—temperature control unit; *B*—temperature chamber; *C*—copper coil; *G-M*—Geiger-Müller counter ready to be lowered into chamber; *S*—scaler; *L*—lid of chamber.

When plateaus must be plotted for large numbers of counters, it may be convenient to do this automatically. A device for the purpose has been described by Crits and Newlin.[2] A rate meter with a recording microammeter is used to draw a curve of the counting rate versus the time. The voltage applied to the counter is automatically increased by one step every 15 or 20 seconds by means of an automatic timer. A rate meter connected to the counter actuates a recorder which draws a curve consisting of a series of steps. The steps are large at the beginning of the characteristic, small in the region of the plateau, and larger above the plateau. To protect the counter from excessive voltages, a relay is arranged to discontinue the operation when the rate meter exceeds a predetermined reading. This equipment also could be arranged to proceed from one counter to another and draw the plateaus for a considerable number of counters without attention from the operator.

3. *Effects of Changes of Temperature on Counters:* The starting voltage of most Geiger-Müller counters depends on the temperature. This is particularly true of self-quenching counters which use an organic vapor as a quenching agent. A shift of the starting voltage tends to move the whole plateau curve to the right or left, with respect to the voltage applied to the counter. If the temperature effect is large, a change of temperature of a few degrees during operation may throw the counter entirely off of the plateau. In an case, it will affect the counting rate to an extent dependent on the temperature coefficient, the temperature change, and the slope of the plateau. Therefore, it is important to know the effect of temperature on the counting properties of the counter.

For determination of the effects of temperature, it is convenient to mount the counter in a roomy box, well-insulated thermally. To control the temperature, a coil of copper tubing is mounted inside the box. The coil consists of a large number of turns mounted near the walls and extending over the greater part of the interior surface. The air inside the box is circulated rapidly by a fan, to aid in decreasing the time required for temperature equilibrium. Liquid at a controlled temperature is pumped through the copper coil. For most applications, tests conducted between 0°C and 30°C give adequate information. When operating at the lower temperatures it is essential to have a drying agent, such as silica gel, inside the thermal chamber to prevent condensation of water on the counter and leads. The temperature of the interior of the box is measured by several thermometers extending through the wall. No observations are made until these thermometers show a steady and uniform temperature. A photograph of this chamber is shown in Fig. 10.

To determine the temperature coefficient of the starting voltage, a constant source is mounted in a fixed position, chosen to give a counting rate of approximately 50 counts per second at the middle of the plateau. With the interior at the desired constant temperature, the data required for plotting the plateau are taken. This gives not only the effect of the temperature on the starting voltage, but reveals any effects on the lengths and slope of the plateau. With some filling mixtures, such effects are noticeable. For those counters where the principal effect of change of temperature is to alter the starting voltage, a temperature coefficient of 4 or 5 volts per °C is usually considered acceptable, although counters which have a lower coefficient have been encountered.

4. *Dead-Time Measurements:* As has been indicated, a knowledge of the dead time of a counter is often useful. Wherever a counter is operated at a counting rate such that the loss of counts due to dead time is appreciable, it is necessary to make a correction for this loss. Actually, the resolving time of the counter depends on the sensitivity of the equipment used to detect the pulses in the counter. The difference between the two is usually small, and the resolving time can never be less than the dead time. The relation between them is illustrated in Fig. 11.

The simplest observation of dead time is made according to the method developed by Stever,[3] using an oscilloscope provided with an external synchronization terminal and a driven sweep. The center wire is coupled through a capacitor of adequate voltage rating to the y-axis terminals of the oscilloscope. The amplified pulses from the counter are simultaneously applied to the external synchronization terminal. This results in the triggering of the sweep for each pulse. A source is placed near the counter to give between 100 and 200 counts per second. The pattern on the scope is that shown in Fig. 11.

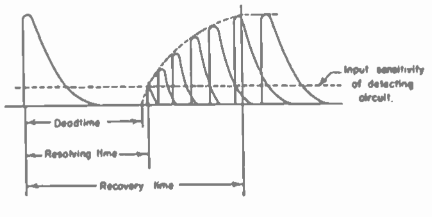

Fig. 11—Diagram showing recovery of a Geiger-Müller counter, subsequent to an initial pulse. In the interval labeled "dead time" no other pulse appears. The recovery time is determined in part by the size of the pulse which can be detected by the indicating circuit. The recovery time is determined by the time required for the pulses to regain their original size.

[2] G. J. Crits and G. C. Newlin, "Recording type automatic Geiger tube calibrator," U. S. Atomic Energy Commission Report 2025; 1948.

[3] H. G. Stever, "The discharge mechanism of fast G-M counters from the deadtime experiment," *Phys. Rev.*, vol. 61, pp. 38–52; January, 1942.

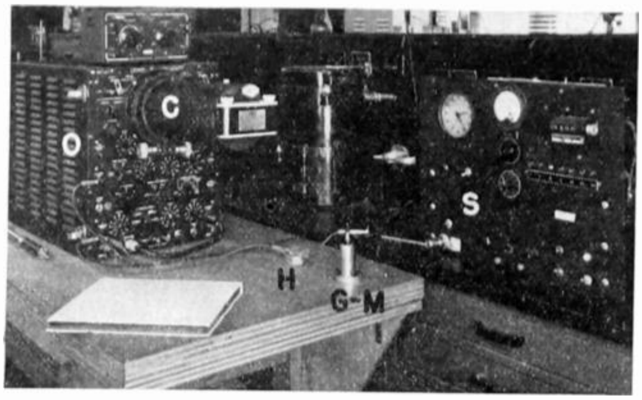

Fig. 12—Oscilloscope equipped with camera for photographing dead-time patterns. O—oscilloscope; C—camera; H—high-voltage capacitor; S—scaler; G-M—counter under test.

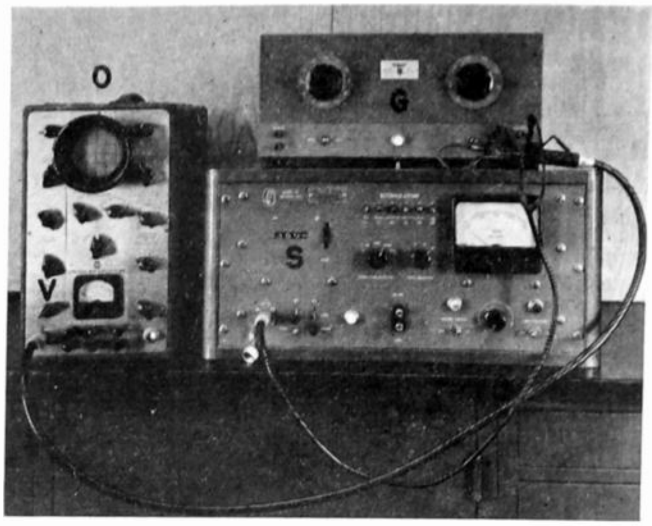

Fig. 13—Determination of input sensitivity of a scaler. O—oscilloscope; V—voltmeter; G—square-wave generator; S—scaler under test.

The duration of the sweep should be about 1000 microseconds with time markers each 10 or 100 microseconds. The measurement consists in photographing the pattern on the screen and measuring the time intervals on the negative. Fig. 12 shows equipment for photographing this pattern.

For more accurate measurements, a method of electronic gating may be used. This refinement is not always necessary in the routine tests of counters, but it is helpful in investigations of factors which affect the dead time. By detection of small changes in dead time, it reveals the direction in which improvement has been achieved. This information may result in further modifications which give still greater improvements. A suitable method for using electronic gating is described in a forthcoming issue of the *Journal of Research of the National Bureau of Standards.*[4]

5. *Measurement of Pulse Size:* The sizes of the pulse in volts from Geiger-Müller counters vary from counter to counter, depending on the size of the electrodes, the filling gas, and other constructional features.

It is desirable to know the pulse size of a counter at some representative voltage, since this determines whether the counter can be used with a given amplifying and detecting equipment. The difference in voltage between the operating and threshold voltages is defined as the overvoltage. The size of the pulses in a given counter increases with overvoltage. Therefore, some definite part of the plateau must be selected in comparing pulse sizes in counters. A knowledge of this pulse size is needed to determine whether a counter can be used with a given amplifying and detecting equipment. In addition, the pulse size is a function of the capacitance across the counter. Therefore, in making comparative measurements it is essential that observations be made for a definite fixed value of this capacitance, which includes a number of elements in an actual experimental arrangement. These include the cable and coupling capacitors and those in the instrument used for observations. It is usual to select either 50 or 100 $\mu\mu f$ for this capacitance, and adjust the connecting system to bring it to this value. Observations are usually made by applying the pulses to the y-axis amplifier of an oscilloscope for which the value of the deflections in millimeters per volt have been ascertained. A complete set of values for a number of arbitrary voltages covering the entire plateau is taken. For purposes of comparison, the size of the pulse at the center of the plateau is a convenient parameter. The equipment for these measurements is shown in Fig. 13.

6. *Photosensitivity:* Finally, it should be mentioned that it is desirable to ascertain whether counters are photosensitive. Such a test should be applied to all counters constructed in a way which permits light to strike the interior electrodes. When a counter is photosensitive, its response to light is usually manyfold greater than to radioactive radiations under usual conditions of operation. However, this is not always true and photosensitivity is not always recognized immediately. It may then lead to erroneous conclusions regarding the functioning of the counter or its associated equipment, and is always a possible source of error in measurements. This defect of counters is readily detected by enclosing the counter in a light-tight shield provided with a small aperture through which light may be directed. A counter which is photosensitive must either be used in a light-tight enclosure or completely coated with opaque varnish. Since photosensitivity occasionally develops with use, it is a good precaution to screen all counters from light unless they are constructed in such a way that light does not reach the interior.

The above tests are appropriate for any type of Geiger-Müller counter. There are a few tests, in addition, that may be applied to beta-ray counters only, and, similarly, one or two which relate to gamma-ray counters. For beta-ray counters, perhaps the most important is a measurement of the thickness of the window. In quantitative measurement with beta-ray counters it is important to know the thickness of the window to permit application of corrections for absorption of beta particles in the window. This factor can be obtained most accurately and conveniently from measurements made before the window is mounted on the counter. This information is not always given by manufacturers of commercial tubes. Therefore, it is advantageous to measure the thickness on a finished counter.

One method of measuring the thickness of mica windows on end-window counters which has been applied with good results involves the use of polonium alpha particles. This method can be applied to windows in the range of 1 to 4 mg/cm², which includes those used on most commercial counters. The procedure is to mount the counter in a rigidly fixed position, with the window end down. Below this a polonium source is mounted in such a way that its distance from the window can be determined accurately. The source can be moved vertically by means of a micrometer screw with a vernier scale calibrated in millimeters. A photograph of the arrangement is shown in Fig. 14. The counter is operated in the proportional region, where it only detects alpha particles, and pulses are recorded by use of a linear amplifier and scaler. The observations consist in plotting the counting rate against the distance of the source from the window. A typical curve is shown in Fig. 15. The steep part of this curve is extrapolated to the x-axis, which gives the extrapolated range of the polonium alpha particles minus the window thickness. The extrapolated range of polonium alpha particles is 3.897 cm at 15°C, 76 cm pressure.[5] The observations are corrected for temperature and pressure and subtracted from the above value for the range which gives the thickness of the window in centimeters of air. This value is converted to mg/cm² to give the thickness in the units commonly used. This method has been applied to windows of known thickness and gives excellent agreement.

Optical methods can be used for measuring thin, transparent films, as has been pointed

[4] L. Costrell, "Accurate determination of the dead-time and recovery characteristics of Geiger Muller counters," *Jour. Res. Nat. Bur. Stand.*, vol. 42, pp. 241–251; March, 1949.

[5] M. G. Halloway and M. S. Livingston, "Range and specific ionization of alpha particles," *Phys. Rev.*, vol. 54, pp. 18–37; January, 1938.

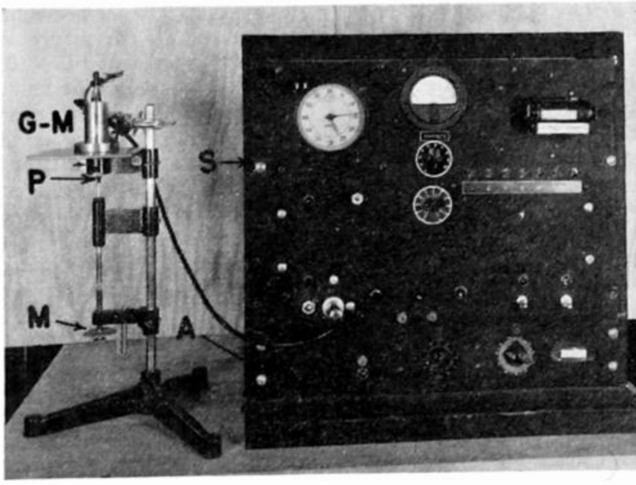

Fig. 14—Equipments for measuring thickness of mica windows in counters using α particles from polonium. *G–M*—mica window counter; *P*—polonium source; *M*—micrometer screw; *A*—linear amplifier; *S*—scaler.

out by Wood.[6] Recently, Brown and Willoughby[7] described a practical adaptation of the procedure. The window is illuminated by an ordinary incandescent lamp at an angle of incidence of 10°. The reflected light, when viewed by a small direct-vision spectroscope, shows a continuous spectrum crossed by interference fringes. The number of fringes between two lines in a comparison spectrum are counted. The comparison spectrum can be obtained from a fluorescent light, and it is convenient to take the 577-mμ yellow line and the 436-mμ blue line of mercury as reference points. An experimental curve gives the relation between the number of milligrams per square centimeter and the number of fringes between the reference lines. W = weight in mg/cm²; ρ is measured density of the mica = 2.85; μ is the index of refraction = 1.58; γ is the angle of incidence;

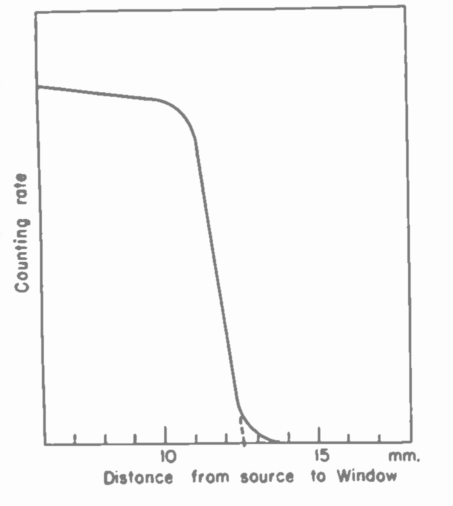

Fig. 15.—Graph showing the number of alpha particles detected as a function of the distance of the polonium source from the window of the counter.

[6] R. W. Wood, "Physical Optics," p. 192.

[7] F. W. Brown, III, and A. B. Willoughby, "Optical method of determining thickness of Geiger tube windows," *Rev. Sci. Instr.*, vol. 19, pp. 820–821; November, 1948.

$\lambda_1 = 577$ mμ; $\lambda_2 = 436$ mμ; and n is the number of fringes. The formula for the curve reduces to

$$W = 0.163n \frac{\text{mg}}{\text{cm}^2}$$

for an angle of incidence of 10°, where W = thickness in mg/cm², and n is the number of fringes.

For gamma-ray counters it is useful to know the relative efficiency of a counter for gamma radiation of various energies. This can be determined by coincidence measurements between the pulses from the gamma-ray counter and a beta-ray counter when both are exposed to the same preparation of a radioisotope emitting both beta and gamma rays, provided the energy of the gamma rays is known and the disintegration scheme is reasonably simple. Strictly, this method requires that the isotope emit a single gamma ray, and relatively few radioisotopes have such a simple disintegration scheme. However, it is possible to find several which emit two or three gamma rays with energies near enough together to give a reasonable average for the energy of the group. Among these are Na^{22}, Na^{24}, Co^{60}, and Au^{198}. The procedure consists in observing the counting rates of the two counters separately and also the coincidences between them, using a Rossi-type coincidence circuit, illustrated in Fig. 2. The efficiency of the gamma-ray counter in the particular arrangement used is given from the following considerations:

Let A = counting rate of gamma counters
B = counting rate of beta counters
C = coincidence rate
N = disintegration rate of source, which need not be known.

If E_γ is the efficiency of the gamma counter and E_β that of the beta counter,

$$A = E_\gamma N, \quad B = E_\beta N \quad \text{and} \quad C = E_\gamma E_\beta N.$$

Therefore,

$$E_\gamma = \frac{C}{E_\beta N} = \frac{C}{B},$$

or the efficiency of the gamma counter is the ratio of the coincidence rate to the counting rate of the beta-ray counter. This experiment can be repeated, maintaining a fixed geometrical arrangement of counters and source, using radioisotopes emitting gamma rays of various energies to give the desired information regarding the variation of sensitivity of the counter with the energy of the gamma rays.

7. *Test of Scalers:* A scaler for use with counters usually consists of a number of stages, each containing two electron tubes, connected in a circuit so that only one can be in the conducting state. This arrangement permits each stage to transmit one-half the pulses which enter it. Therefore, the scaling ratio is 2^n where n is the number of stages. Decade scalers consist of a modification of the scale-of-two connections in stages so that every tenth pulse is transmitted by a stage. Although there is usually some type of electron-tube amplification between the counter and the scaler, in practice this amplifier is included with the scaler assembly. This permits operation of the counter by a direct connection to the scaler unit. Tests of the scaler unit therefore automatically include the amplifier.

Scalers are usually tested to determine the minimum pulse voltage to which the scaler will respond as well as the maximum pulse voltage at which it fails to operate satisfactorily. The minimum pulse on which a scaler will operate satisfactorily may be determined by use of a square-wave generator of variable amplitude synchronized with the 60-cps power line. The square pulses are passed through a differentiating network to convert them into pulses with a steep rise and an exponential decay with a time constant of approximately 50 microseconds. The alternator of the square-wave generator is set for a minimum voltage output, and the pulse size increased gradually until the scaler responds accurately to the 60-cps rate. This pulse size is measured on a cathode-ray oscilloscope, and is the minimum input sensitivity of the scaler. The pulses may then be increased in voltage until the scaler begins to show erratic operation, which gives the maximum voltage of pulse for the scaler.

The input sensitivity may also be checked under actual operating conditions, using pulses from a Geiger-Müller counter. With a counter connected to the scaler, the voltage on the counter is set at approximately 100 volts above the threshold. Assuming that the scaler then functions correctly, which can be checked by observing counting rates with sources of known ratios of intensity, the voltage on the counter is reduced until the scaler begins to fail to respond accurately. If this occurs while the counter is still operating in the Geiger-Müller range, the pulses will still be of uniform height. This height can be measured with a cathode-ray oscilloscope, as before.

The scaling accuracy is most simply checked by using a generator synchronized with the 60-cps power line and timed by a electric clock driven by the same supply. The pulse size is set at about 1½ times the minimum input sensitivity, and the clock and scaler started simultaneously. At the end of an hour the elapsed time is compared with the recorded counts. If it is desired to check the accuracy of scaling at other frequencies, a stable oscillator which can be set at a number of frequencies is substituted for the 60-cps synchronization, and the observations are repeated. Sometimes a scaler is found which will scale correctly over a small range of input pulses, but will fail for larger pulses. It is usual, therefore, to increase the pulse size to several times the lower limit of input sensitivity in steps, and determine the scaling accuracy at each step.

In the test of scalers it is important to check all control switches, particularly the "start and stop" switch, to ascertain whether the scaler starts reliably when this switch is thrown to the "start" position and whether this operation introduces spurious counts into the scaling stages. Commercial scalers have been encountered which do not always start scaling when this switch is thrown to the starting position. The introduction of spurious counts by the switch is also a defect encountered occasionally.

It is desirable to check the operation of scalers over extended periods of time, since failures due to overloading components become evident in this way. Continuous operation for twenty-four hours followed by the usual tests reveals defects of this type, and also reveal whether sufficient ventilation has been provided to prevent overheating.

Most scalers operate satisfactorily over a range of from 90 to 130 volts on the power-line input. In a few instances scalers have been encountered which become erratic if the line voltage fluctuates ±10 volts. To detect whether this characteristic is present, scalers are tested with the supply voltage at 90 volts and at 130 volts, as well as at 110 volts.

The high-voltage rectifier is tested for regulation. Since counters usually have plateaus with slopes of the order of 5 per cent per 100 volts, it is important that the high voltage remain stable within better than 1 per cent for the fluctuations in power-line voltages ordinarily encountered. A rough test of the stability can be made by using a voltmeter of appropriate range and varying the supply voltage over a range of 90 to 130 volts. If changes in the voltmeter readings are readily discernible, this usually means that the high voltage is insufficiently regulated. Where it is desirable to determine the actual degree of regulation, a null-type difference voltmeter circuit is used. This permits measurements of changes of less than 1 volt in the high voltage.

The above tests are also applied to ascertain the stability of the high voltage with time for a constant regulated supply voltage. To obtain a satisfactory test of this kind, it is necessary to make observations on the high voltage over short intervals of time, say every minute, for 10 minutes, and over longer intervals of the order of every 10 minutes for an hour. Slow changes are detected by observations every hour over a period of several hours.

If the high-voltage rectifier is supplied with a voltmeter, which is desirable, it is necessary to check the accuracy of this voltmeter. This meter, to be of real use, must read standard volts to within a few per cent. It is usually relied upon to adjust the voltage for operation of a new counter. If it is off by more than a few per cent, there is danger of applying an excessive voltage to the counter before the operating range for the counter has been ascertained in terms of the inaccurate meter. It is equally important that the errors be uniformly distributed over the scale. The voltmeter can be checked at a number of readings by use of a calibrated voltmeter of high resistance connected in parallel with it.

The only component of the scaling circuit assembly remaining to be discussed is the mechanical register. Fortunately, these devices are ruggedly built, and are not required to operate at very high recording rates. When high counting rates are used, additional scaling stages are usually introduced to relieve the register of excessive rates. It is rare that a register of the decade type fails to operate properly, and little attention is given to them in testing unless obvious defects become apparent.

8. *Testing of Rate Meters:* Equipment for the operation of Geiger-Muller counters using the principle of the rate meter for visual indication is less commonly used for quantitative measurements than is the scaler type of instrument. This situation arises from the fact that the indicating meter is subject to fluctuations inherent in the emission of radioactive radiations, and is therefore difficult to read accurately. These fluctuations become pronounced in the more sensitive models with a short time constant to give more nearly instantaneous readings of intensity. Rate meters are usually employed in counter units for survey purposes where high accuracy is not expected. They also lend themselves readily to continuous recording on moving-chart recorders. Rather simple tests are required on these instruments, such as determinations of the stability of the high-voltage supply, variations of meter indication with power-line voltage, and a test of the linearity of the meter scale with the intensity of the radiation incident on the counter. Usually, commercial models are supplied with a counter tube, so that tests are made on the complete assembly. Occasionally, rate-meter dials are calibrated in counts per unit of time. The accuracy of these scales can be checked by use of a stable oscillator with adjustable frequency to produce suitable pulses for activating the circuit.

9. *Health Survey Meters:* A considerable number of instruments manufactured for the measurement of radioactivity consist of devices using ionization chambers to measure gamma radiation in roentgens. The visual indication is achieved either by quartz-fiber electroscopes, or by electronic amplification of the ionization current. The accuracy required of these meters is of the order of ±15 per cent.

The most important test of survey meters consists in checking the accuracy of the scale of the instrument, which usually is marked to read milliroentgens per hour. Those using quartz fibers require observation of the rate of deflection of the fiber. In some cases a timer, such as a flashing light, is built into the instrument.

The calibration of these survey meters is conveniently checked by use of the gamma radiation from standardized preparations of radium. One milligram of radium enclosed in 0.5 mm of platinum gives, with sufficient approximation, 0.85 mr per hour at a distance of 1 meter. Therefore, preparations of radium of the order of 10 to 100 mg are convenient in checking the calibration of the usual survey meter. The inverse square law can be assumed to hold for distances of the order of 25 to 200 cm, if care is taken to reduce the effects of scattering of gamma radiation by near-by objects. Measurements of distance of the source are made from the center of the ionization chamber. More accurate tests are obtained if a series of radium preparations are available, so that the indications of the instrument at differeng intensities of the radiation are made in a fixed geometry.

The variation of the sensitivity with temperature is made at two temperatures, about 20°C apart. Temperatures in the neigborhood of 0°C and 20°C are usually chosen for this test.

Survey meters should give readings in terms of roentgens, which means that they should indicate 1 roentgen for that intensity of radiation which produces the amount of ionization in air required by the definition of the roentgen. Unless the ionization chamber has been properly designed to produce the same relative ionization as is obtained in an air-wall chamber for gamma radiations of various energies, the chamber will not give true readings for all energies of gamma rays. Such chambers are sometimes called "energy-dependent." The degree of deviation with energy in these chambers can be checked by comparing the readings with those of an instrument which has been checked over a range of energies when both are exposed successively to radiations of differing energies. In practice, this is most conveniently accomplished by using as sources a series of radioisotopes which emit gamma rays of different energy. This tests the instrument under conditions approximately those under which it will be used.

Some health survey meters are of the integrating type; that is, they indicate the total amount of radiation in milliroentgens to which they have been exposed. The most important of these is the pocket chamber. This is a small ionization chamber provided with best quality of electrical insulation. In use, the central electrode is charged to a definite potential, and the chamber is issued to an individual whose work exposes him to gamma radiation. At the end of the day the chamber is returned to the laboratory, and the residual potential measured. The charging device contains a quartz-fiber electroscope which measures the initial charging potential, and is also used to measure the residual potential. The scale of this instrument is calibrated in roentgens.

Pocket chambers are tested for natural leakage of charge due to imperfect insulation. In a good chamber a negligible fraction of the initial charge will leak off in twenty-four hours when protected from exposure to radiation. Tests are also made of the loss of charge due to vibration or shock. A good chamber may be dropped on the floor from height of several feet and show no change in potential. Finally, the accuracy of the calibration is checked by comparing the readings obtained from the charging electrometer with the radiation computed from exposures of measured duration at various distances from standardized preparations of radium.

The Columbia Long-Playing Microgroove Recording System*

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Summary—The Columbia LP (long-playing) microgroove recording system was developed to fill the need for music reproduction which would avoid interruptions not intended by the composer, and which would be of excellent quality at a reasonable cost. This all-important factor of cost and the public's familiarity with the handling of phonograph records made it desirable to solve the task on the basis of records, rather than tape or wire.

Standard 78-rpm records were originally designed to generate sound mechanically by direct transfer of energy from the groove of the record to the vibrating diaphragm. Because the entire acoustical energy had to be extracted from the grooves, these had to be quite rugged, and remained so up until now.

The new Columbia recording system was an inevitable outcome of the use of electrical amplification between the groove and the loudspeaker. Today, practically no mechanical energy needs to be extracted from the groove, and thus, for the first time, it has been possible to develop much finer grooves, permitting longer playing time and distortion-free reproduction.

Introduction

The Columbia LP[1] (long-playing) microgroove recording system was developed to provide uninterrupted music reproduction of better quality at a reasonable cost. This all-important factor of cost, together with the public's familiarity with the handling of phonograph records, made it desirable to solve the problem on the basis of disk records, rather than by tape or wire.

Standard 78-rpm records were originally designed to generate sound mechanically by direct transfer of energy from the groove of the record to the vibrating diaphragm. Because the entire acoustical energy had to be extracted from the grooves, these had to be quite rugged, and remained so until now. The weight of the diaphragm and armature, together with the acoustical reproducer, resulted in high vertical needle forces, requiring large stylus diameters. The latter made necessary the use of the high-speed turntable (78 rpm) in order to allow for a sufficient frequency range and a minimum of distortion. Today, practically no mechanical energy need be extracted from the groove, thus allowing a far greater latitude in the design of an improved system.

The fundamental principles of the Columbia LP records are the slow rotational speed of 33⅓ rpm, representing a time-saving factor of roughly 2.3; and finer grooves, up to 300 per inch.

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† Columbia Broadcasting System, Inc., New York, N. Y.
‡ Columbia Records, Inc., New York, N. Y.
[1] Trade-mark of Columbia Records, Inc., New York, N. Y.

Playing Time

It was found that the average playing time for classical works is approximately 36 minutes. Nevertheless, standards were developed to accommodate as much as 50 minutes of playing time on a 12-inch record.

The following analysis establishes the optimum relationship between the inside and outside groove diameters for maximum playing time.

The variables used are:

d = inside groove diameter in inches
D = outside groove diameter in inches
R = revolutions per minute
N = total number of grooves
T = playing time in minutes.

The constants used are:

n = grooves per inch
$V_{\min}$ = minimum permissible linear groove velocity in inches per second.

The playing time T is equal to the total number of grooves N divided by the revolutions per minute R of the record.

$$T = \frac{N}{R}. \tag{1}$$

N is a function of the number of grooves per inch and of the radial distance utilized:

$$N = n\left(\frac{D-d}{2}\right). \tag{2}$$

R is determined by the minimum recorded radius and the minimum permissible linear groove velocity:

$$R = \frac{60V_{\min}}{\pi d}. \tag{3}$$

Substituting (2) and (3) in (1),

$$T = \frac{n\left(\frac{D-d}{2}\right)}{\frac{60V_{\min}}{\pi d}} = \frac{\pi n}{120V_{\min}}(Dd - d^2). \tag{4}$$

Maximizing T with respect to d,

$$\frac{\delta T}{\delta d} = \frac{\pi n}{120V_{\min}}(D - 2d) = 0$$

and

$$d = \frac{D}{2}. \tag{5}$$

In other words, maximum playing time for a given inner-groove linear velocity is obtained when the inner diameter of the record equals one-half the outer diameter.

From (4), with $d = D/2$

$$T = \frac{\pi n D^2}{480 V_{\min}}$$

and

$$n = \frac{480 V_{\min} T}{\pi D^2}. \tag{6}$$

Thus, when $T = 20$ minutes, $V_{\min} = 10$ inches per second and $D = 11.5$ inches; the number of grooves per inch n will be 230. As the nearest practical value, 224 grooves per inch are used for all records up to 20 minutes per side, or 40 minutes total playing time. This requires an inner diameter of 5.5 inches, with a corresponding velocity of 9.6 inches per second. On record sides requiring more than 20 minutes of playing time, 260 grooves per inch are used.

Tracing Distortion

Increasing the length of playing time by an appreciable factor was only one of the goals to be accomplished. With the advent of FM and professional tape recorders, improved performance over standard 78-rpm and even transcription records became a necessity.

In the following, an analysis will be made of the performance of Columbia's LP records in relation to commercial 78-rpm records, and also to transcription-type recording. The chief factors considered will be those which are influenced by the linear speed and by the amplitude excursions of the reproducing stylus. It has been found in disk recording in general that the limitations imposed by cutter performance, such as clearance angle of cutting stylus, acceleration of the stylus, etc., are usually less limiting than the tracing distortion, which is a function of the minimum radius of curvature of the traced waves and the effective radius of the reproducing stylus.[2]

First, the minimum radius of curvature corresponding to maximum deviation amplitude for various types of records (78-rpm, transcription, and LP) will be determined; at the same time, it will be postulated that the radius of curvature is not to be less than the effective radius of the reproducing stylus. This does not imply that such a condition ensures distortion-free reproduction, since, even under this condition, the center of the reproducing stylus does not trace a sine wave. Distortion increases very rapidly when the effective stylus radius exceeds the radius of curvature.[2] The purpose of this analysis, however, is to show the relative performances of 78-rpm, transcription, and LP records, rather than to define certain requirements already well-known in the recording field.

Arbitrarily, the condition where the two radii (namely, minimum radius of curvature of wave and effective radius of reproducing stylus) are equal, will be named the limiting condition, and the corresponding frequency will be termed the limiting frequency.

In general, the radius of curvature for a function $y = f(x)$ can be expressed as follows:

$$r = \frac{\left[1 + \left(\frac{dy}{dx}\right)^2\right]^{3/2}}{\frac{d^2y}{dx^2}}. \tag{7}$$

Applying this to a sine wave, $y = D \sin 2\pi x/\lambda$, it can be shown that r has a minimum for $x = \lambda/4$. Thus,

$$r_{\min} = \frac{\lambda^2}{4\pi^2 D}. \tag{8}$$

The wavelength λ as traced on the record is equal to the linear velocity V of the groove at that point, divided by the frequency f of the energy delivered to the cutting stylus. Therefore, the frequency at which the minimum radius of curvature will equal the effective reproducing-stylus tip radius for the deviation D can be called limiting frequency, and will be

$$f_l = \frac{V}{2\pi\sqrt{r_{\text{eff}} D}}. \tag{9}$$

Actual values of f_l can now be determined for 78-rpm records, transcription disks, and LP records for the maximum possible deviation ($D_{\max}$) when using the innermost grooves. The critical frequency values are of greatest interest for those grooves nearest the center because of their low linear velocity.

For 78-rpm records at a 4-inch diameter, the linear velocity $V = 16.3$ inches per second. The tip radius is usually 3 mils, and the effective radius $r_{\text{eff}} = 0.003 \cos \alpha/2$, where α is the total groove angle. Thus, since $\alpha = 90°$, $r_{\text{eff}} = 0.0021$ inch. The peak displacement $D_{\max} = 0.002$ inch; substituting these values in (9),

$$f_{l(78)} = \frac{16.3}{2\pi\sqrt{(0.0021)\cdot(0.002)}} = 1{,}270 \text{ cps}. \tag{10}$$

The limiting frequency for transcription records is

$$f_{l(TR)} = \frac{14}{2\pi\sqrt{(0.0017)\cdot(0.0011)}} = 1{,}620 \text{ cps}. \tag{11}$$

[2] W. S. Bachman, "Phonograph dynamics," *Electronic Ind.*, vol. 4, pp. 86–89; July, 1945.

Finally, the innermost grooves of LP records have a linear velocity of the order of 9.6 inches per second; the reproducing-stylus tip radius is 0.001 inch, and the effective radius equals 0.0007 inch. The maximum groove deviation is 0.0009 inch. From these values, it follows that

$$f_{l(LP)} = \frac{9.6}{2\pi\sqrt{(0.0007)\cdot(0.0009)}} = 1{,}940 \text{ cps.} \qquad (12)$$

Thus, the limiting frequency for LP records is higher than for either transcription records or standard 78-rpm records.

With regard to these frequency values, an obvious conclusion would be that none of the three recording systems under scrutiny is capable of reproducing frequencies above the values corresponding to the limiting frequency; in the case of the LP record, this would be 1,940 cycles. It is necessary to keep in mind, however, that the various values for f were established using the maximum groove displacement D_{max}.

Since second-harmonic distortion is automatically eliminated by virtue of the complete symmetry of the distortion produced in the traced waves, another way of comparing tracing distortion between 78-rpm commercial records, transcription records, and the LP microgroove records is to utilize an expression for the rms value of third-harmonic lateral tracing distortion[3] T.

$$T = \frac{6\pi^4\cdot r^2\cdot D^2\cdot f^4}{V^4}\cos^2\frac{\alpha}{2}, \qquad (13)$$

where

r = tip radius of the reproducing stylus in inches
D = peak amplitude of the recorded wave in inches
f = frequency of the recorded wave in cps
V = linear groove velocity in inches per second
α = groove angle.

Since the purpose is to determine the relative distortion of various disk-recording systems, a graph has been prepared (Fig. 1) in which, based on (13), the distortion of LP records, commercial transcription records, the recently announced RCA 7-inch record, and conventional 78-rpm records has been determined in relation to the tracing distortion on the inside of the standard 78-rpm disk. The ordinate gives the inverse of this distortion ratio, so that, the higher the ordinate value, the better the theoretical quality of the record. For instance, the LP record on the outside (zero minutes playing time) shows 85 times less distortion than the inside of the conventional 78-rpm record. After 21 minutes playing time, the LP record shows roughly 6.2 times less distortion. The RCA 7-inch record shows 7.4 times less distortion. With respect to actual demonstrable differences among these various types of records, it can be seen that the tracing distortion on the inside of commercial transcription records (after 15 minutes playing) is roughly one-third of that found on the inside of the 78-rpm record. It is questionable whether additional reduction in tracing distortion can be demonstrated with anything short of professional laboratory equipment when using regular program material.

[3] W. D. Lewis and F. V. Hunt, "A theory of tracing distortion in sound reproduction from phonograph records," *Jour. Acous. Soc. Amer.*, vol. 12, p. 353, 1941.

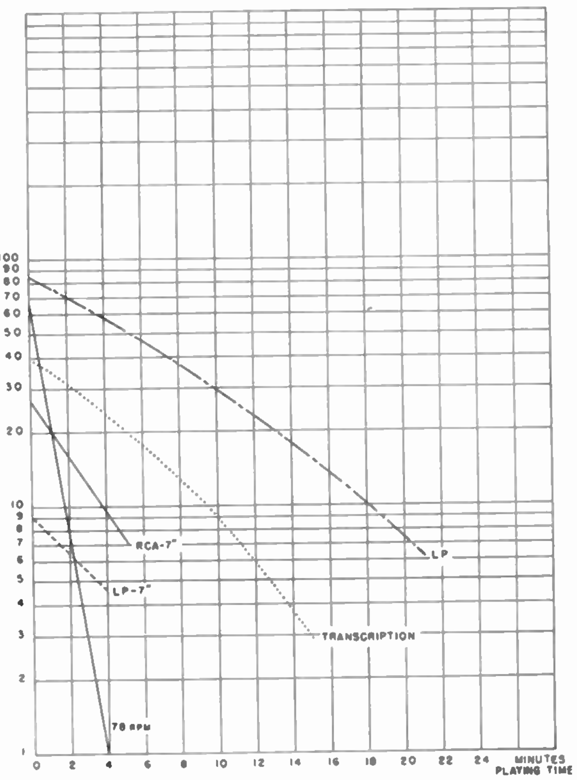

Fig. 1—Performance characteristics of various types of records.

Equation (9) may be written with $r_{min} = r_{eff}$.

$$D_{usable} = \frac{V^2}{4\pi^2 f^2 r_{eff}}. \qquad (14)$$

This corresponds to the usable deviation that may be employed in recording a frequency f while still retaining the original condition that the stylus radius does not exceed the minimum radius of curvature of the recorded wave. From (14),

$$\frac{D_{usable}}{D_{max}} = \frac{\dfrac{V^2}{4\pi^2 f^2 r_{eff}}}{\dfrac{V^2}{4\pi^2 f_l^2 r_{eff}}} = \left(\frac{f_l}{f}\right)^2. \qquad (15)$$

In Fig. 2, D_{usable}/D_{max} is plotted in percentage versus frequency for 78-rpm records, for transcription records, and for LP records. The ordinate indicates percentage usable maximum groove displacement where 100 per cent equals the maximum possible values used in (10),

(11) and (12). Thus, 100 per cent modulation equals 0.002 inch (10) for 78-rpm records; 0.0011 inch (11) for transcription records; and 0.0009 inch (12) for LP records. As indicated previously, these conditions are those prevailing at the lowest linear velocity (inside grooves) for each particular type of record.

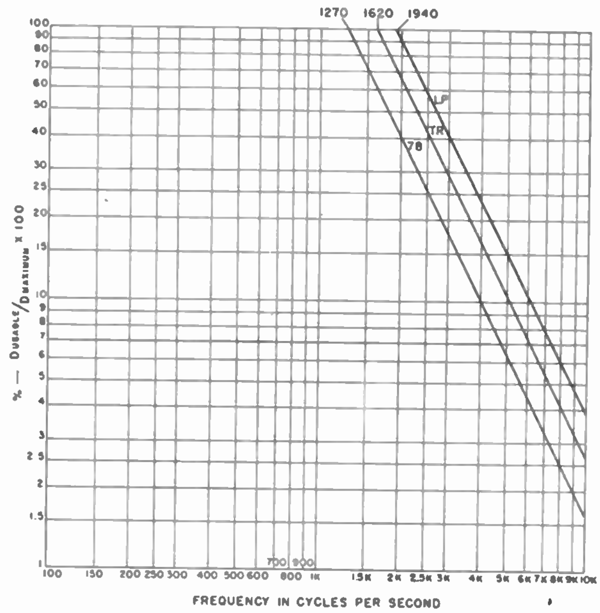

Fig. 2—Percentage maximum deviation versus frequency.

The LP recording characteristic (velocity versus frequency) is shown by curve 1 in Fig. 3. Curve 2 in Fig. 3 shows the NAB velocity recording characteristic, which deviates from the LP characteristic in the low bass portion only. The purpose for the LP bass lift is to reduce rumble and hum pickup, which would otherwise be more pronounced with the smaller deviations.

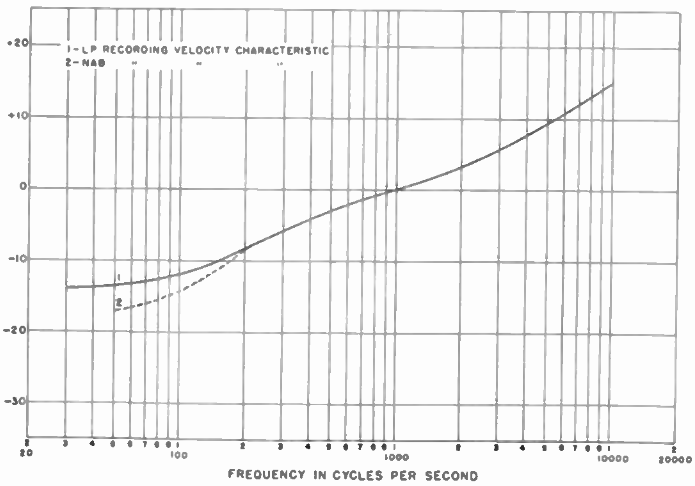

Fig. 3—LP and NAB recording characteristics.

Hand in hand with the LP record, suitable reproducing equipment has been developed in the Columbia Laboratories. Existing pickups, such as used for 78-rpm records, obviously could not be employed. First, as has been mentioned previously, the tip radius for LP records should be 0.001 inch ($\pm$ 10 per cent). Second, the weight requirements are quite different. Extensive tests have shown that, for a stylus of 0.001-inch radius, as used with the LP records, a downward stylus force not exceeding 6 grams is desirable. An LP stylus employing a 6-gram force exerts no more pressure than the average 78-rpm pickup.

The necessity of developing a single cartridge capable of tracking both LP and 78-rpm records with only 6 grams stylus force was also realized. As a result, two cartridges with a new nonresonant type of arm were developed; a single cartridge playing LP records only, and another cartridge having two styli and a single crystal capable of playing either LP or 78-rpm records. Both cartridges track with a downward force of only 5 to 6 grams and have approximately 0.5 volt rms output. These cartridges, with an associated filter, have a substantially flat response over the entire recorded frequency range.

Since the included groove angle of both LP and 78-rpm records is approximately 90 degrees, the vertical force F_v tending to force the stylus out of the groove is equal to the total lateral force on the stylus F_L. This force has a component F_s which overcomes the stiffness of the stylus and tends to force the stylus away from its position of equilibrium, and an opposing component F_a, which accelerates the stylus toward its position of equilibrium. At low frequencies, where most tracking problems were encountered, F_a, which decreases with the square of the recorded frequency, becomes negligible compared to F_s, and can, therefore, usually be neglected.

Then,

$$F_v = F_L = F_s \text{ (at low frequencies).} \tag{16}$$

Now

$$F_s = \frac{D_{\max}}{C} \tag{17}$$

where C is the compliance measured at the point of the stylus.

From (16) and (17),

$$F_v = \frac{D_{\max}}{C} \quad \text{or} \quad C = \frac{D_{\max}}{F_v}. \tag{18}$$

Substituting the $D_{\max}$ and F_v used in the LP system in (18),

$$C = \frac{(0.0009)\cdot(2.54)}{(6)\cdot(980)} = (0.39)\cdot 10^{-6} \text{ cm/dyne.} \tag{19}$$

Substituting the corresponding values for the 78-rpm side of the dual cartridge,

$$C = \frac{(0.002) \cdot (2.54)}{(6) \cdot (980)} = (0.87) \cdot 10^{-6} \text{ cm/dyne}. \quad (20)$$

This higher value of compliance was obtained by employing a stylus wire of a smaller diameter. Actual compliance values in these cartridges are more than double the values given above, which represent the lowest ones required theoretically.

A number of laboratory player attachments were built using the newly developed pickups and employing standard two-pole motors, modified to give wow- and rumble-free reproduction. The players and the records were field-tested in homes over a considerable period of time. Fig. 4 shows the LP attachment designed by Columbia to play both 78-rpm and LP records using the nonresonant arm and dual cartridge described previously. Fig. 5 shows the Columbia player attachment engineered by Philco.

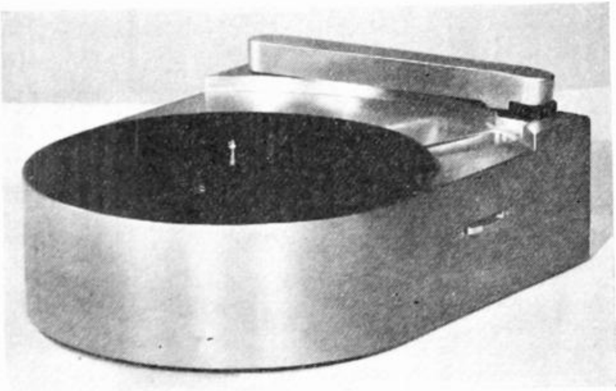

Fig. 4—Columbia's experimental dual cartridge-dual speed player.

Fig. 5—Philco LP record player.

ACKNOWLEDGMENTS

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Direct Voltage Performance Test for Capacitor Paper*

HAROLD A. SAUER† AND DAVID A. McLEAN†

Summary—**Performance of capacitors on accelerated life test may vary over a wide range depending upon the capacitor paper used. Indeed, at present a life test appears to be the only practical means for evaluating capacitor paper, since, within the limits observed in commercial material, the chemical and physical tests usually made do not correlate with life. Lack of correlation is ascribed to obscure physical factors which have not yet been identified.**

Generally, several weeks are required to evaluate a paper by life tests of the usual severity. Unfortunately, the duration of these tests is too long for quality control of paper.

The desire for a life test which requires no more than a day or two for evaluation led to the development of a rapid dc test. The philosophy of rapid life testing is based upon the experimental evidence that the process of deterioration under selected temperature and voltage conditions is principally of a chemical nature, and also upon the well-known fact that rates of chemical reaction increase exponentially with temperature.

Life tests on two-layer capacitors conducted at 130°C provide an acceleration in deterioration many fold more than that obtained in the lower-temperature life tests, and correlate well with these tests.

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† Bell Telephone Laboratories, Murray Hill, N. J.

and

$$d = \frac{D}{2}. \tag{5}$$

In other words, maximum playing time for a given inner-groove linear velocity is obtained when the inner diameter of the record equals one-half the outer diameter.

From (4), with $d = D/2$

$$T = \frac{\pi n D^2}{480 V_{min}}$$

and

$$n = \frac{480 V_{min} T}{\pi D^2}. \tag{6}$$

Thus, when $T = 20$ minutes, $V_{min} = 10$ inches per second and $D = 11.5$ inches; the number of grooves per inch n will be 230. As the nearest practical value, 224 grooves per inch are used for all records up to 20 minutes per side, or 40 minutes total playing time. This requires an inner diameter of 5.5 inches, with a corresponding velocity of 9.6 inches per second. On record sides requiring more than 20 minutes of playing time, 260 grooves per inch are used.

Tracing Distortion

Increasing the length of playing time by an appreciable factor was only one of the goals to be accomplished. With the advent of FM and professional tape recorders, improved performance over standard 78-rpm and even transcription records became a necessity.

In the following, an analysis will be made of the performance of Columbia's LP records in relation to commercial 78-rpm records, and also to transcription-type recording. The chief factors considered will be those which are influenced by the linear speed and by the amplitude excursions of the reproducing stylus. It has been found in disk recording in general that the limitations imposed by cutter performance, such as clearance angle of cutting stylus, acceleration of the stylus, etc., are usually less limiting than the tracing distortion, which is a function of the minimum radius of curvature of the traced waves and the effective radius of the reproducing stylus.[2]

First, the minimum radius of curvature corresponding to maximum deviation amplitude for various types of records (78-rpm, transcription, and LP) will be determined; at the same time, it will be postulated that the radius of curvature is not to be less than the effective radius of the reproducing stylus. This does not imply that such a condition ensures distortion-free reproduction, since, even under this condition, the center of the reproducing stylus does not trace a sine wave. Distortion increases very rapidly when the effective stylus radius exceeds the radius of curvature.[2] The purpose of this analysis, however, is to show the relative performances of 78-rpm, transcription, and LP records, rather than to define certain requirements already well-known in the recording field.

Arbitrarily, the condition where the two radii (namely, minimum radius of curvature of wave and effective radius of reproducing stylus) are equal, will be named the limiting condition, and the corresponding frequency will be termed the limiting frequency.

In general, the radius of curvature for a function $y = f(x)$ can be expressed as follows:

$$r = \frac{\left[1 + \left(\frac{dy}{dx}\right)^2\right]^{3/2}}{\frac{d^2y}{dx^2}}. \tag{7}$$

Applying this to a sine wave, $y = D \sin 2\pi x/\lambda$, it can be shown that r has a minimum for $x = \lambda/4$. Thus,

$$r_{min} = \frac{\lambda^2}{4\pi^2 D}. \tag{8}$$

The wavelength λ as traced on the record is equal to the linear velocity V of the groove at that point, divided by the frequency f of the energy delivered to the cutting stylus. Therefore, the frequency at which the minimum radius of curvature will equal the effective reproducing-stylus tip radius for the deviation D can be called limiting frequency, and will be

$$f_l = \frac{V}{2\pi\sqrt{r_{eff} D}}. \tag{9}$$

Actual values of f_l can now be determined for 78-rpm records, transcription disks, and LP records for the maximum possible deviation (D_{max}) when using the innermost grooves. The critical frequency values are of greatest interest for those grooves nearest the center because of their low linear velocity.

For 78-rpm records at a 4-inch diameter, the linear velocity $V = 16.3$ inches per second. The tip radius is usually 3 mils, and the effective radius $r_{eff} = 0.003 \cos \alpha/2$, where α is the total groove angle. Thus, since $\alpha = 90°$, $r_{eff} = 0.0021$ inch. The peak displacement $D_{max} = 0.002$ inch; substituting these values in (9),

$$f_{l(78)} = \frac{16.3}{2\pi\sqrt{(0.0021)\cdot(0.002)}} = 1{,}270 \text{ cps.} \tag{10}$$

The limiting frequency for transcription records is

$$f_{l(TR)} = \frac{14}{2\pi\sqrt{(0.0017)\cdot(0.0011)}} = 1{,}620 \text{ cps.} \tag{11}$$

[2] W. S. Bachman, "Phonograph dynamics," *Electronic Ind.*, vol. 4, pp. 86–89; July, 1945.

Finally, the innermost grooves of LP records have a linear velocity of the order of 9.6 inches per second; the reproducing-stylus tip radius is 0.001 inch, and the effective radius equals 0.0007 inch. The maximum groove deviation is 0.0009 inch. From these values, it follows that

$$f_{l(LP)} = \frac{9.6}{2\pi\sqrt{(0.0007)\cdot(0.0009)}} = 1{,}940 \text{ cps.} \qquad (12)$$

Thus, the limiting frequency for LP records is higher than for either transcription records or standard 78-rpm records.

With regard to these frequency values, an obvious conclusion would be that none of the three recording systems under scrutiny is capable of reproducing frequencies above the values corresponding to the limiting frequency; in the case of the LP record, this would be 1,940 cycles. It is necessary to keep in mind, however, that the various values for f were established using the maximum groove displacement $D_{\max}$.

Since second-harmonic distortion is automatically eliminated by virtue of the complete symmetry of the distortion produced in the traced waves, another way of comparing tracing distortion between 78-rpm commercial records, transcription records, and the LP microgroove records is to utilize an expression for the rms value of third-harmonic lateral tracing distortion[3] T.

$$T = \frac{6\pi^4\cdot r^2\cdot D^2\cdot f^4}{V^4}\cos^2\frac{\alpha}{2}, \qquad (13)$$

where

r = tip radius of the reproducing stylus in inches
D = peak amplitude of the recorded wave in inches
f = frequency of the recorded wave in cps
V = linear groove velocity in inches per second
α = groove angle.

Since the purpose is to determine the relative distortion of various disk-recording systems, a graph has been prepared (Fig. 1) in which, based on (13), the distortion of LP records, commercial transcription records, the recently announced RCA 7-inch record, and conventional 78-rpm records has been determined in relation to the tracing distortion on the inside of the standard 78-rpm disk. The ordinate gives the inverse of this distortion ratio, so that, the higher the ordinate value, the better the theoretical quality of the record. For instance, the LP record on the outside (zero minutes playing time) shows 85 times less distortion than the inside of the conventional 78-rpm record. After 21 minutes playing time, the LP record shows roughly 6.2 times less distortion. The RCA 7-inch record shows 7.4 times less distortion. With respect to actual demonstrable differences among these various types of records, it can be seen that the tracing distortion on the inside of commercial transcription records (after 15 minutes playing) is roughly one-third of that found on the inside of the 78-rpm record. It is questionable whether additional reduction in tracing distortion can be demonstrated with anything short of professional laboratory equipment when using regular program material.

[3] W. D. Lewis and F. V. Hunt, "A theory of tracing distortion in sound reproduction from phonograph records," *Jour. Acous. Soc. Amer.*, vol. 12, p. 353, 1941.

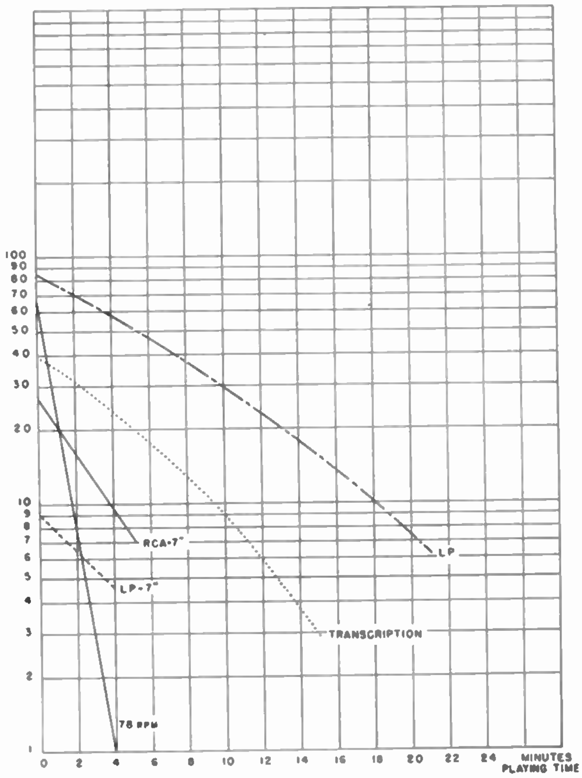

Fig. 1—Performance characteristics of various types of records.

Equation (9) may be written with $r_{\min} = r_{\text{eff}}$.

$$D_{\text{usable}} = \frac{V^2}{4\pi^2 f^2 r_{\text{eff}}}. \qquad (14)$$

This corresponds to the usable deviation that may be employed in recording a frequency f while still retaining the original condition that the stylus radius does not exceed the minimum radius of curvature of the recorded wave. From (14),

$$\frac{D_{\text{usable}}}{D_{\max}} = \frac{\dfrac{V^2}{4\pi^2 f^2 r_{\text{eff}}}}{\dfrac{V^2}{4\pi^2 f_l^2 r_{\text{eff}}}} = \left(\frac{f_l}{f}\right)^2. \qquad (15)$$

In Fig. 2, $D_{\text{usable}}/D_{\max}$ is plotted in percentage versus frequency for 78-rpm records, for transcription records, and for LP records. The ordinate indicates percentage usable maximum groove displacement where 100 per cent equals the maximum possible values used in (10),

(11) and (12). Thus, 100 per cent modulation equals 0.002 inch (10) for 78-rpm records; 0.0011 inch (11) for transcription records; and 0.0009 inch (12) for LP records. As indicated previously, these conditions are those prevailing at the lowest linear velocity (inside grooves) for each particular type of record.

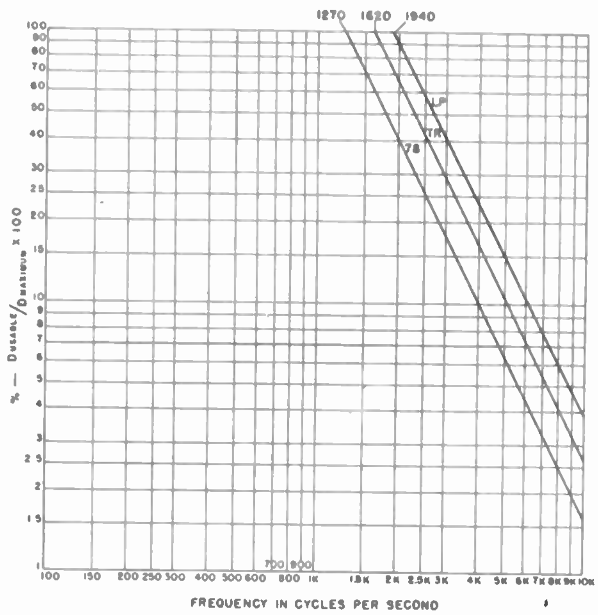

Fig. 2—Percentage maximum deviation versus frequency.

The LP recording characteristic (velocity versus frequency) is shown by curve 1 in Fig. 3. Curve 2 in Fig. 3 shows the NAB velocity recording characteristic, which deviates from the LP characteristic in the low bass portion only. The purpose for the LP bass lift is to reduce rumble and hum pickup, which would otherwise be more pronounced with the smaller deviations.

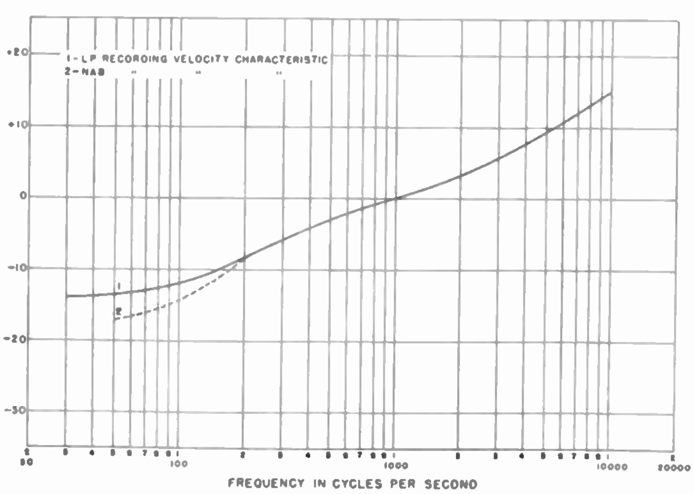

Fig. 3—LP and NAB recording characteristics.

Hand in hand with the LP record, suitable reproducing equipment has been developed in the Columbia Laboratories. Existing pickups, such as used for 78-rpm records, obviously could not be employed. First, as has been mentioned previously, the tip radius for LP records should be 0.001 inch ($\pm$ 10 per cent). Second, the weight requirements are quite different. Extensive tests have shown that, for a stylus of 0.001-inch radius, as used with the LP records, a downward stylus force not exceeding 6 grams is desirable. An LP stylus employing a 6-gram force exerts no more pressure than the average 78-rpm pickup.

The necessity of developing a single cartridge capable of tracking both LP and 78-rpm records with only 6 grams stylus force was also realized. As a result, two cartridges with a new nonresonant type of arm were developed; a single cartridge playing LP records only, and another cartridge having two styli and a single crystal capable of playing either LP or 78-rpm records. Both cartridges track with a downward force of only 5 to 6 grams and have approximately 0.5 volt rms output. These cartridges, with an associated filter, have a substantially flat response over the entire recorded frequency range.

Since the included groove angle of both LP and 78-rpm records is approximately 90 degrees, the vertical force F_v tending to force the stylus out of the groove is equal to the total lateral force on the stylus F_L. This force has a component F_s which overcomes the stiffness of the stylus and tends to force the stylus away from its position of equilibrium, and an opposing component F_a, which accelerates the stylus toward its position of equilibrium. At low frequencies, where most tracking problems were encountered, F_a, which decreases with the square of the recorded frequency, becomes negligible compared to F_s, and can, therefore, usually be neglected.

Then,

$$F_v = F_L = F_s \text{ (at low frequencies).} \qquad (16)$$

Now

$$F_s = \frac{D_{\max}}{C} \qquad (17)$$

where C is the compliance measured at the point of the stylus.

From (16) and (17),

$$F_v = \frac{D_{\max}}{C} \quad \text{or} \quad C = \frac{D_{\max}}{F_v}. \qquad (18)$$

Substituting the $D_{\max}$ and F_v used in the LP system in (18),

$$C = \frac{(0.0009)\cdot(2.54)}{(6)\cdot(980)} = (0.39)\cdot 10^{-6} \text{ cm/dyne.} \qquad (19)$$

Substituting the corresponding values for the 78-rpm side of the dual cartridge,

$$C = \frac{(0.002)\cdot(2.54)}{(6)\cdot(980)} = (0.87)\cdot 10^{-6}\ \text{cm/dyne}. \quad (20)$$

This higher value of compliance was obtained by employing a stylus wire of a smaller diameter. Actual compliance values in these cartridges are more than double the values given above, which represent the lowest ones required theoretically.

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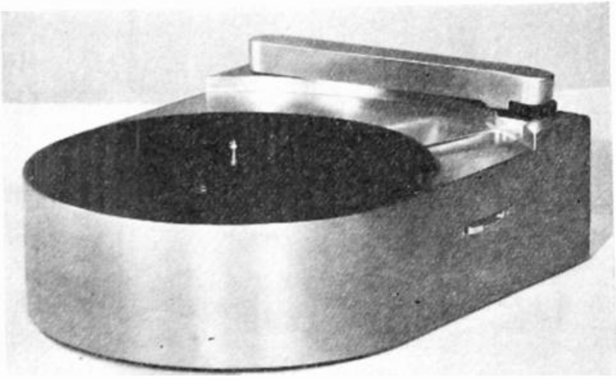

Fig. 4—Columbia's experimental dual cartridge-dual speed player.

Columbia to play both 78-rpm and LP records using the nonresonant arm and dual cartridge described previously. Fig. 5 shows the Columbia player attachment engineered by Philco.

Fig. 5—Philco LP record player.

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HAROLD A. SAUER† AND DAVID A. McLEAN†

Summary—Performance of capacitors on accelerated life test may vary over a wide range depending upon the capacitor paper used. Indeed, at present a life test appears to be the only practical means for evaluating capacitor paper, since, within the limits observed in commercial material, the chemical and physical tests usually made do not correlate with life. Lack of correlation is ascribed to obscure physical factors which have not yet been identified.

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* Decimal classification: R281×R215.13. Original manuscript received by the Institute, June 7, 1948; revised manuscript received, November 24, 1948.

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Introduction

Paper capacitors, like other components of electronic apparatus, must continue to face demands for higher-temperature and higher-voltage operation, for more dependable service, and for greater compactness. Enormous improvements in dependability and general quality level of paper capacitors have been made over the past several years. Nevertheless, where dependability and long service life are paramount considerations, designs must still be based on the minimum expected life, a value which may be only a few per cent of the maximum expected life.

Our work has indicated that large variation in the durability of paper capacitors subjected to direct voltage results from uncontrolled variables in the paper. Many chemical and physical tests have been proposed for determining the durability of capacitor paper. Our experience with such tests indicates that they may suffice to eliminate the extremely poor papers but are valueless for choosing between, e.g., a moderately poor paper and an excellent paper. In Table I, it is seen that papers of widely different accelerated life test performance exhibit chemical and physical properties which are either identical or which fail to correlate with the life. In our opinion, the reason for this is that the life characteristics of paper are very sensitive to variations in its microstructure which none of the usual chemical and physical tests detects. Pending discovery of tests which will detect such variations, it is necessary to use life tests to determine the voltage performance of paper.

The goal which was set was that of an accelerated life test which could be made in an elapsed time similar to the chemical and physical tests, i.e., a day or two. The result has been a testing procedure requiring about a day for preparation of samples and about the same time for performing the life test.

It is generally agreed that increasing the voltage to accelerate life testing gives unreliable results at voltages much above two and one-half times the rated voltage. Owing to the chemical nature of the deterioration process, high acceleration may be gained through increasing the test temperature.[1]

Equipment

The method developed consists of subjecting two-layer capacitor test units to a direct voltage life test carried out at 130°C and 560 volts per mil. The equipment consists principally of a variable voltage rectifier producing potentials up to 2,000 volts, an oven, and a measuring and control circuit all housed in a steel cabinet. Fig. 1 is a photograph of the apparatus displaying the general arrangement of the assembly and locations of the important components. Each capacitor on test (Fig. 2) is in circuit with one pole of a double-pole breaker. Facilities are provided for testing ten capacitors at a time. No units are substituted for failed

Fig. 1—Direct-voltage performance test set.

specimens. Failure is marked by a steep rise in the current in the capacitor circuit. When this current reaches about 0.025 ampere the trip circuit is energized, remov-

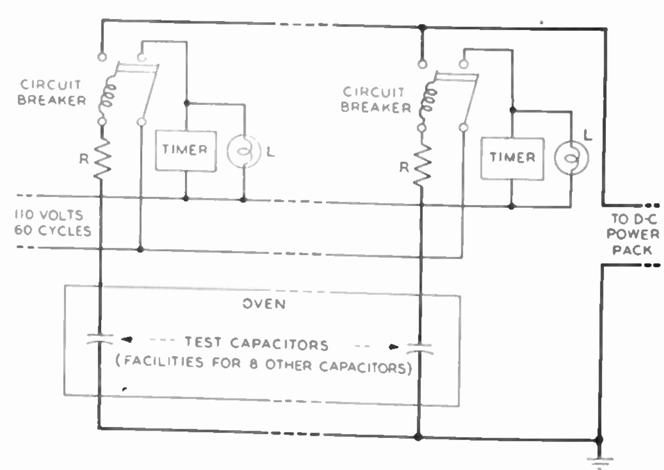

Fig. 2—Test circuit schematic.

ing the voltage from the failed specimen and opening the 110V-60 cps timing circuit associated with the other

[1] D. A. McLean, L. Egerton, G. T. Kohman, and M. Brotherton, "Paper dielectrics containing chlorinated impregnants. Deterioration in dc fields," *Ind. and Eng. Chem.*, vol. 34, pp. 101; January, 1942.

pole of the breaker. The timer is a motor-driven counter which records the time to one-tenth of an hour. The lamp circuit in parallel with the timing circuit aids in determining the status of the test.

Description of the Test

If a life test on capacitors is to be used to evaluate capacitor paper, all other variables must be controlled. This means faithful reproduction of conditions of winding, drying, and impregnation, together with constancy and purity of the other constituents; namely, the impregnant and the electrode metal.

In this investigation, capacitor units are wound with two layers[2] of paper $2\frac{3}{4}$ inches wide and with 0.00025 inch thick aluminum foil electrodes to provide a capacitance, when impregnated, of approximately 1.0 μf for 0.0004 inch thick paper. For thicknesses less than 0.0004 inch, the windings are constructed to provide a test on the same area of paper as contained in a 1-μf 0.0004 inch unit. For thicker papers, in order to avoid excessive physical size, less area is employed.

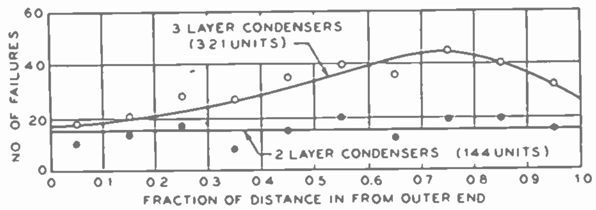

Fig. 3—Distribution of points of failure within test capacitor.

The choice of units of these dimensions resulted from extensive auxiliary investigation in connection with the necessity of determining whether or not overheating develops within the unit during testing under these conditions, causing premature failure. Statistically, the weakest point, and hence the point of failure, in the paper sheet should occur with equal probability on any surface element of the sheet. Post-mortem examination of a large number of specimens has shown that the two-layer units of the above specification exhibit a random distribution of failure along the length of the sheet, indicating the absence of local internal heating. On the other hand, when tested at the much higher voltages required to produce failure in a short time, local overheating occurred in three-layer units of the same physical dimensions as evidenced by an observed peak in the distribution curve, Fig. 3. Another factor which has favored the choice of two-layer units is that such a test has been found to be more discriminating than one on three-layer units, producing a greater difference in life from sample to sample. Another major reason is that the bulk of capacitors for telephone use is of two-layer construction.

In the processing and testing, apparatus which is clean and is inert toward the impregnating compound is essential. Materials such as stainless steel, aluminum, nickel- or tin-plated steel, and glass are preferred for test jigs and containers.

The units in Fig. 4 are held under spring pressure between aluminum or stainless plates during all processing and testing. This provides a mechanical pressure of about 12 pounds per square inch. The processing is conveniently considered in three steps: air-baking, vacuum-drying, and impregnating.

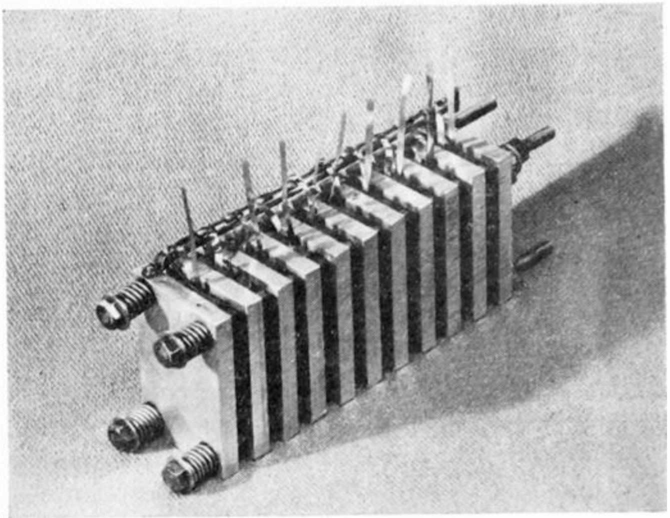

Fig. 4—Capacitor units assembled in clamping jig.

The purpose of the air-bake is to produce limited oxidation of the paper to improve its electrical properties.[3] Also, during the air-bake, the bulk of the moisture is removed. The air-bake is carried out at atmospheric pressure and 130°C for 16 to 18 hours. This temperature is also maintained throughout the following steps.

In the vacuum drying, which follows the air-baking immediately, the units are held under vacuum (<1 mm Hg) for 4.5 hours. During the first 3.5 hours, the vacuum is released momentarily every 30 minutes by admitting dry air. This aids in removing residual moisture. During the last hour the vacuum is applied continuously. The resulting residual moisture content is about 0.05 per cent or less based on the dry weight of the paper tested.

The units are immersed in hot impregnant (130°C) for two hours directly following the drying and without breaking the vacuum. During the first hour, vacuum is maintained. The last hour is at atmospheric pressure.

[2] Two sheets of paper, one superimposed upon the other, are sandwiched between aluminum foils.

[3] D. A. McLean, "Paper capacitors containing chlorinated impregnants. Benefits of controlled oxidation of the paper," *Ind. and Eng. Chem.*, vol. 39, p. 1457; November, 1947.

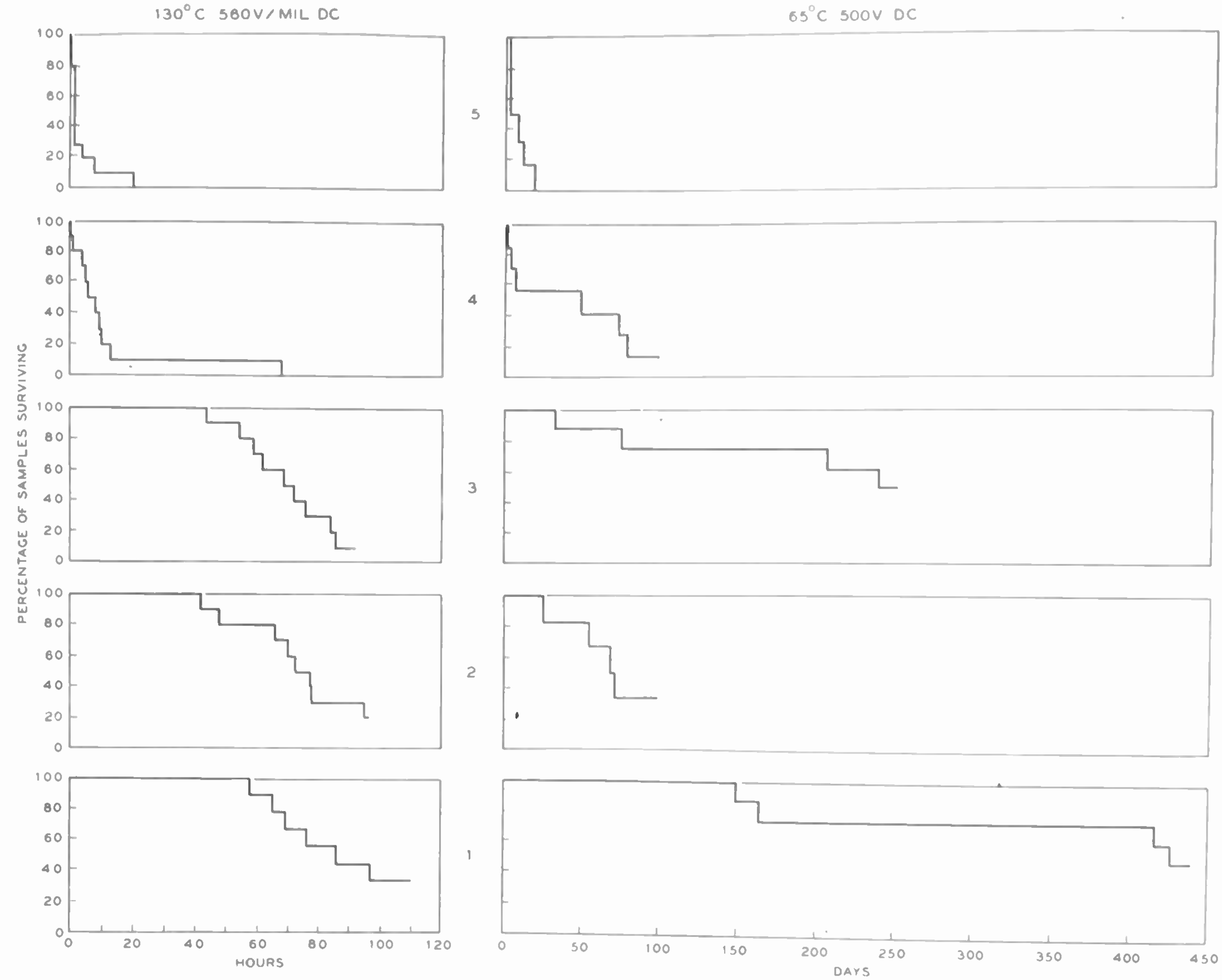

Fig. 5—Comparison of two-layer, 0.4-mil kraft capacitor papers listed in Table I.

The impregnant used in these tests consists of chlorinated naphthalene (Halowax 1001) stabilized with 0.5 per cent of anthraquinone.[4]

A set of test conditions could probably be found which would involve neither stabilization nor air-bake. However, the process described here closely simulates that employed in the manufacture of telephone capacitors.

The test jig with impregnated units in place is then rapidly transferred to the test container and covered with the same compound used in impregnation. Electrical connections are made to the container terminals. Solder connections have been avoided in order to eliminate the possibility of solder flux contamination of the impregnant. The cover is attached and the container placed in the oven. After equilibration at test temperature, the capacitance of each unit is measured as a circuit check, followed by a short time direct-voltage dielectric strength test performed by applying a voltage about 50 per cent in excess of the life test voltage for eliminating mechanically defective samples. The direct-current leakage at 100V is then recorded before the voltage is raised to the life test potential.

[4] D. A. McLean and L. Egerton, "Paper capacitors containing chlorinated impregnants. Stabilization by anthraguinone," *Ind. and Eng. Chem.*, vol. 37, p. 73; January, 1945.

Discussion of Life Test Results

The 130°C life test results on papers referred to in Table I are compared in Fig. 5 with tests conducted at 65°C and 500 volts. In the step-wise plot the time axis in the 130°C life test is hours; in the slower life tests it is days. The scales of the time axes are in the ratio of 48:1 and it will be observed that the 130°C life test provides additional acceleration in deterioration of the order of the time scale ratio.

The time to fifty per cent failure for both the 130°C and 65°C tests has been recorded in Table I as an arbitrary yardstick of comparison. The five papers show the same relative order on both tests with the exception of reversal of the second and third place. The order (1 to 5) as shown is on the basis of the results of the 130°C tests.

Satisfactory duplicability on check runs, shown on Fig. 6, is readily obtained. The important aspects of duplicability are the assurance that the equipment is performing satisfactorily, and that test samples are being prepared in a consistent manner. Early failures arise from various causes; partly, from mechanical defects in the wound unit, but principally from variability of the paper. Apart from this, it will be observed that the course of the life chart results in comparable total integrated life for duplicate runs, the criterion upon which duplicability is based.

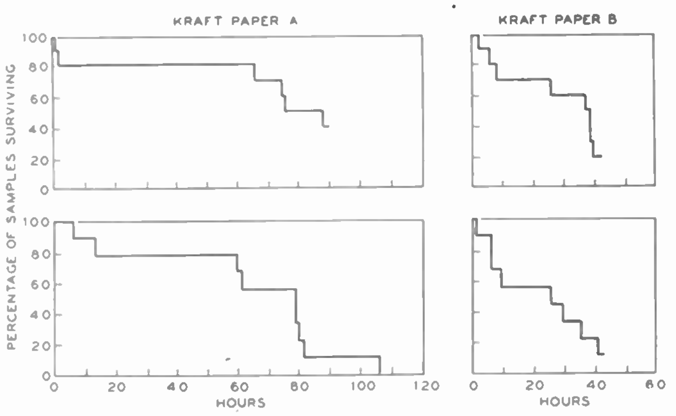

Fig. 6—Duplicability of results by the 130°C life test 560 volts/mil dc on same paper and same quantity of units

The spread of results that can be expected of commercial capacitor paper can be judged approximately from Fig. 5. The setting up of criteria for acceptable paper is beyond the scope of this paper. However, it appears that in a period of twenty-four to forty-eight hours the general quality level of the lot being tested can be judged, provided that the limited number of test units which it is practical to test is representative of the lot. Sampling, required by statistical considerations, to accomplish this end to a high degree has not been determined.

It appears that this test will be of value in research on capacitors, both in improving the quality of the paper and in isolating the factors which govern the life of capacitors.

For purposes of illustration, data have been given on only 0.0004 inch papers. Extensive investigation indicates that similar relationships hold for other thicknesses of paper.

Acknowledgment

The authors are indebted to J. R. Weeks of these Laboratories for numerous helpful suggestions, and for independent corroborative evidence using equipment of the same design.

TABLE I

Chemical and Physical Properties of 0.0004-Inch Kraft Capacitor Papers

Paper Sample	M-86	M-77	M-89	M-95	M-87
	(1)*	(2)*	(3)*	(4)*	(5)*
Chemical Properties:					
Conductivity of water extract (micromhos/cm)	4.6	7.5	5.0	2.7	5.5
Chloride Content (%)	0.002	0.002	0.003	0.002	0.003
Ash (%)	0.28	0.37	0.36	0.28	0.33
Reaction of water extract (meq/gm)	0.002 Alk	0.003 Alk	0.0001 Alk	0.0002 Acid	0.011 Alk
Water soluble (%)	0.23	0.30	0.44	0.39	0.30
Alcohol soluble (%)	0.06	0.18	0.14	0.16	0.10
Benzene soluble (%)	0.06	0.10	0.04	0.05	0.04
*p*H	7.5	8.4			
Permanganate number	11.5	12.7	12.7	13.3	12.3
Pentosans	6.9	7.2	7.9	8.0	7.9
Physical Properties:					
Porosity (Average)	1.6	1.8	0.4	0.4	2.6
Porosity (Max. in 15 values)	3.7	3.9	0.8	0.6	7.9
Apparent Density (Gm/cc)	0.94	0.94	0.98	0.99	0.94
Conducting Particles (per sq. ft.)	2.4	2.6	0.6	5.9	3.4
Life (2-layer units—Halowax plus ½% anthraquinone):					
Days to 50% Failure (65°C.–625V/mil dc)	427	72	240	50	8
Hours to 50% Failure (130°C.–560V/mil dc)	86	73	69	8	1

* Order as shown on Fig. 5.

Some Aspects of Cathode-Follower Design at Radio Frequencies*

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Summary—**Pertinent high-frequency design data including the circuit gain, the gain phase angle, the input impedance in resistive and reactive components, and the maximum allowable input signal voltage may be determined quickly from simple design charts derived by approximations which are applicable over a wide range of frequency and circuit parameters. The only circuit quantities required for use of the charts are the magnitude and phase angle of the cathode load impedance, the internal tube capacitances, and the grid-to-plate transconductance of the tube at the operating point.**

A discussion of the effects of the grid-to-cathode capacitance on circuit operation at high frequencies is presented, including a critical analysis of the source impedance. It is shown that the source impedance may be greatly affected by the impedance of the grid driving source unless remedial measures are taken. Various circuit changes which reduce or eliminate the undesirable effects of the grid-to-cathode capacitance are offered with discussion and analyses.

I. Gain Function for a Complex Load Impedance

Referring to Fig. 1, if the load impedance is a resistance R_k, the gain of the cathode follower is the ratio of e_0 to E_g and has been given by many investigators[1,2] as

$$A = \frac{\mu}{1+\mu} \cdot \frac{R_k}{R_k + \dfrac{R_p}{1+\mu}} \tag{1}$$

where

A = gain of stage

μ = amplification factor of tube

R_p = plate resistance of tube

R_k = cathode load resistance.

Equation (1) is easily derived by application of the equivalent plate-circuit theorem.

It is apparent in (1) that if μ is large compared to unity considerable simplification is possible, resulting in

$$A = \frac{1}{1 + \dfrac{1}{R_k G_m}}. \tag{2}$$

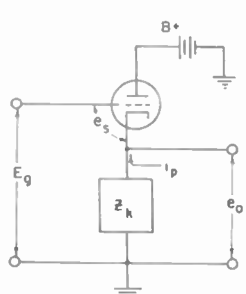

Fig. 1—The gain of the cathode follower is the ratio of e_0 to E_g when R_k is the load impedance.

If the cathode load impedance is complex, R_k in equation (2) becomes $Z_k = R_k \pm jX_k$. Rationalizing and simplifying the resulting expression, the magnitude of the gain A and its phase angle α (which is the phase angle between e_0 and E_g in Fig. 1) can be shown to be given by the following expressions

$$|A| = \frac{G_m|Z_k|}{\sqrt{1 + 2G_m|Z_k|\cos\theta + (G_m|Z_k|)^2}} = \frac{\sin\theta}{\sin\alpha} \cdot \frac{1}{\dfrac{1}{G_m|Z_k|} + G_m|Z_k| + 2\cos\theta} \tag{3}$$

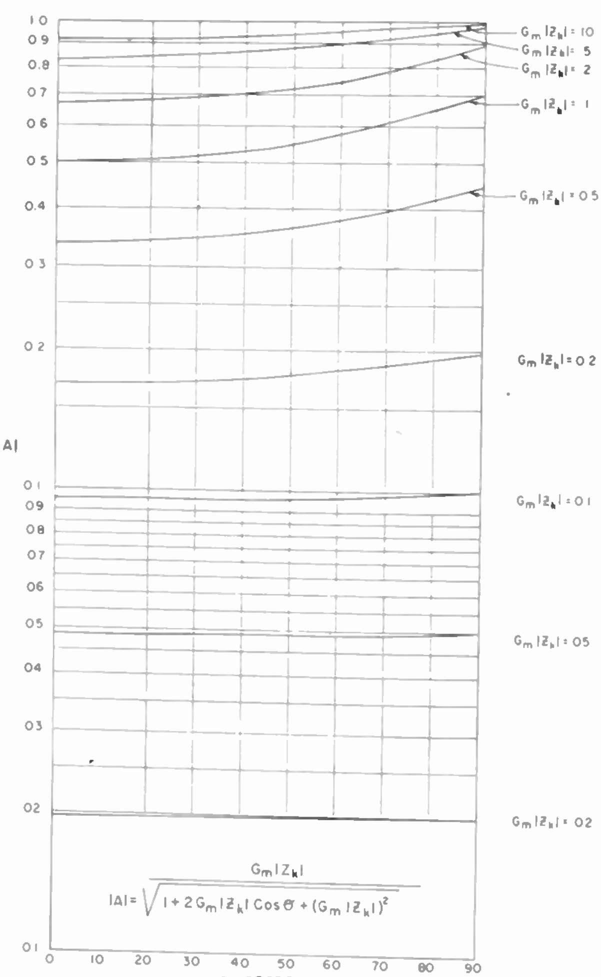

Fig. 2—Magnitude of complex gain $A/\underline{\alpha}$ as a function of cathode load phase angle $\underline{\theta}$.

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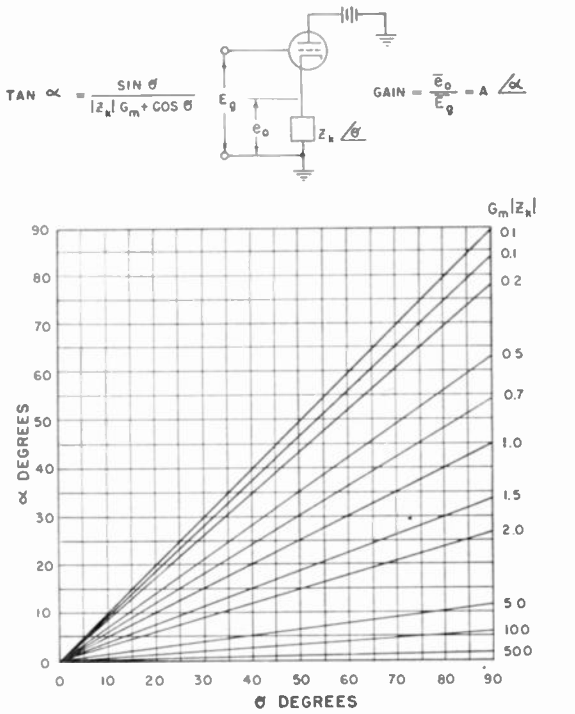

Fig. 3—Relation of cathode load phase angle to angle of complex gain function for various values of parameter $G_m|Z_k|$.

where

g_m = grid-to-plate transconductance

θ = phase anglé of cathode load impedance

$\theta = \arctan X_k/R_k$

$|A|$ = magnitude of complex gain.

The phase angle of the complex gain is

$$\alpha = \arctan \frac{\sin \theta}{G_m|Z_k| + \cos \theta}. \tag{4}$$

Equations (3) and (4) are plotted as Figs. 2 and 3 with the product $g_m|Z_k|$ as a parameter. These charts permit very quick and easy evaluation of the magnitude and phase angle of the stage gain for any load impedance. At low frequencies where the load impedance is essentially resistive, it may be considered that $\theta = 0$ and the left ordinate of Fig. 2 will give the results obtained from previously published nomographs[3] for these conditions. α is then zero, of course.

II. Effects of Grid-to-Cathode Capacitance

At high frequencies, the grid-to-cathode capacitance of commonly used tubes acts as a coupling impedance of serious proportions between input and output circuits. Several first-order results of this coupling are observed. First, the input impedance (dynamic impedance between grid and ground) which is normally equivalent to the grid-to-plate capacitance shunted by a portion of the grid-to-cathode capacitance and an extremely high resistance, is lowered significantly. Depending upon the phase angle of the cathode load impedance, the resistive component may become negative in sign, although the reactive component is always capacitive. With capacitive cathode loads of suitable proportions, the negative resistance seen looking into the grid circuit may become so low that oscillation in some mode is virtually a certainty. The input capacitance may be significantly increased. Using the symbols of Fig. 1, the input current may be written:

$$i_g = \frac{\bar{E}_g - \bar{e}_0}{-jX_{gk}} \tag{5}$$

$$i_p \gg i_g$$

where i_g = current flowing to the grid. By previous definition,

$$e_0 = E_g A \underline{/\alpha}. \tag{6}$$

Then

$$Z_g = \frac{\bar{E}_g}{i_g} = \frac{-jX_{gk}}{1 - A\underline{/\alpha}}. \tag{7}$$

In deriving (7), nothing has been said about the current flowing into the grid-to-plate capacitance. This component of the impedance is a static quantity unaffected by the tube operation (since the plate is assumed to be grounded for rf) and may be added in later.

For convenience, (7) may be rationalized to separate the resistive and reactive components. Performing this operation yields

$$Z_g = |X_{gk}|\left[\frac{\mp|A|\sin\alpha - j(1 - |A|\cos\alpha)}{1 - 2|A|\cos\alpha + |A|^2}\right]. \tag{8}$$

The $\mp$ sign corresponds to that of the cathode-load phase angle θ.

Equation (8) expresses the new components of input impedance due to coupling of the grid-to-cathode capacitance as a series circuit of capacitance and resistance across the input terminals. It is more convenient to express this new impedance in terms of an equivalent parallel circuit of resistance and capacitance. This is easily done with the standard circuit transformations, yielding

$$R_{\text{par}} = \frac{X_{gk}}{|A|\sin\alpha} = \frac{X_{gk}\sqrt{1 + 2G_m|Z_k|\cos\theta + (G_m|Z_k|)^2}}{A\sin\theta}. \tag{9}$$

$$X_{\text{par}} = \frac{X_{gk}}{1 - |A|\cos\alpha} = \frac{X_{gk}}{1 - |A|\dfrac{G_m|Z_k| + \cos\theta}{\sqrt{1 + 2G_m|Z_k|\cos\theta + (G_m|Z_k|)^2}}}. \tag{10}$$

$$C_{\text{par}} = C_{gk}(1 - |A|\cos\alpha) = C_{gk}\left[1 - |A|\frac{G_m|Z_k| + \cos\theta}{\sqrt{1 + 2G_m|Z_k|\cos\theta + (G_m|Z_k|)^2}}\right] \tag{11}$$

where

R_{par} = equivalent shunt resistance appearing across input terminals

[3] M. B. Kline, "Cathode follower nomograph for pentodes," *Electronics*, vol. 20, p. 136; June, 1947.

X_{par} = equivalent shunt reactance appearing across input terminals

C_{par} = equivalent shunt capacitance appearing across input terminals.

Equations (9), (10), and (11) are normalized with respect to X_{gk} and C_{gk} and plotted in Figs. 4, 5, and 6. After first finding $|A|$ and α from the curves previously presented, the input impedance components may be very quickly found with sufficiently good accuracy for most engineering purposes.

Equation (9) shows that for frequencies and tube capacitances such that X_{gk} is a few hundred ohms, the cathode follower will make a vigorous oscillator if the cathode load is capacitive so that R_{par} has a negative sign. This phase of operation has been well covered in several of the references.

The second very important result of the grid-cathode coupling is its influence upon the effective source impedance of the cathode follower.

The source impedance may be defined as that impedance seen looking into the output terminals toward the tube as in Fig. 7. Here a shunt circuit of three elements is seen represented by the cathode load impedance Z_k, the tube plate resistance R_p in series with the equivalent generator, and the cathode-grid reactance in series with whatever impedance appears between grid and ground represented by the symbol Z_{gg}. It is important to note that Z_{gg} may be anything from a very low value to a very high value depending on the nature of the driving source and stray reactances.

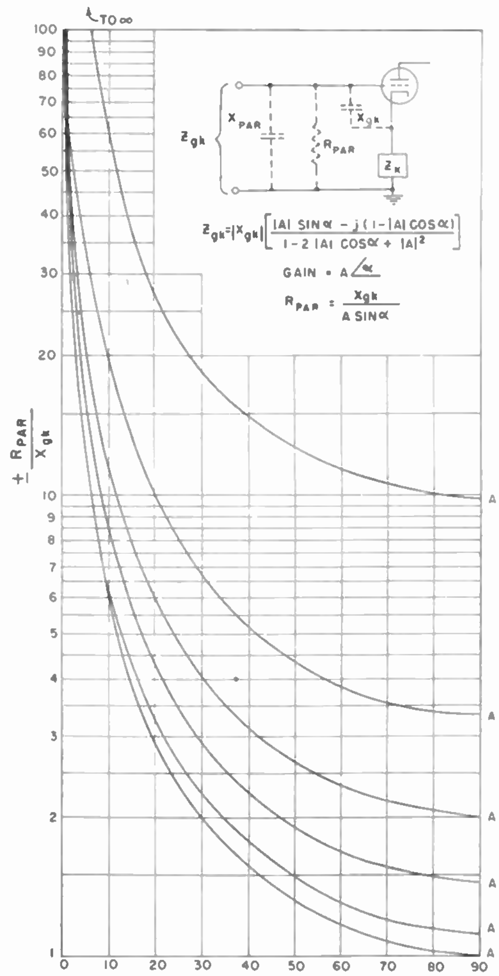

Fig. 4—Equivalent shunt resistance of cathode-follower input impedance.

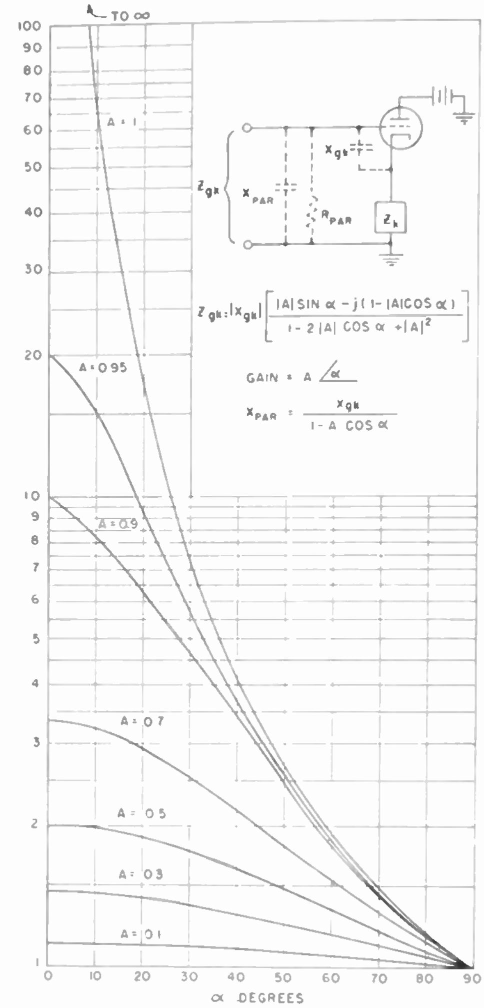

Fig. 5—Equivalent shunt reactance of cathode-follower input impedance.

The circuit of Fig. 7 may be solved by Kirchhoff's laws from the equations

$$i_t Z_k - i_p Z_k - i_g Z_k = E_t, \quad (12)$$

$$i_p R_p = E_t + \mu e_s, \quad (13)$$

$$i_g (Z_{gg} - jX_{gk}) = E_t, \quad (14)$$

where E_t = test voltage applied across output terminals.

For convenience, a new quantity β will be defined which is given by

$$\beta = \frac{-jX_{gk}}{Z_{gg} - jX_{gk}}. \quad (15)$$

β represents the portion of cathode-ground voltage E_t which appears between grid and cathode e_s due to static voltage division by the impedances from grid to cathode and from grid to ground.

Simultaneous solution of (12), (13), and (14) and insertion of (15) gives the source impedance as

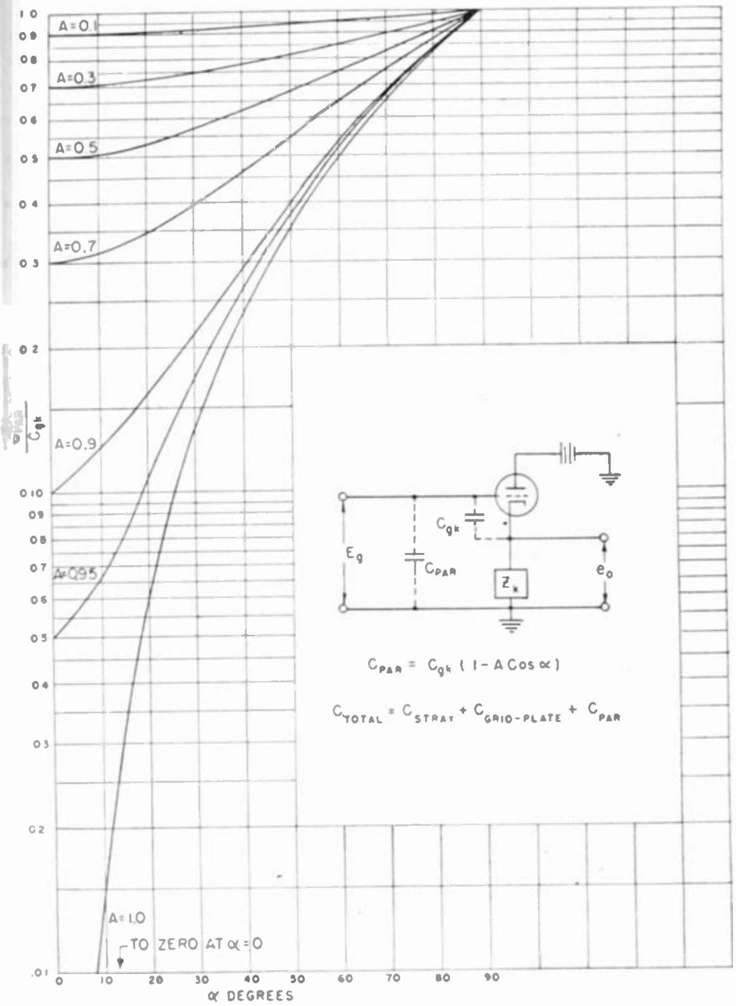

Fig. 6—Ratio of apparent input capacitance to grid cathode capacitance as a function of cathode follower complex gain $A/\underline{\alpha}$.

$$Z_s = \frac{1}{\dfrac{1}{Z_k} + \dfrac{1}{Z_{gg} - jX_{gk}} + \dfrac{1+\mu\beta}{R_p}}. \tag{16}$$

No approximations are involved in this equation except the assumption that all the voltage across Z_k is due to the plate current; i.e., $i_p \gg i_g$.

Inspection of the circuit reveals that the first 2 denominator terms are static quantities unaffected by tube operation. The term $(1+\mu\beta)/R_p$ corresponds to the dynamic component of the source impedance which is of particular interest. If $\beta = 1$, this term is seen to have a value so high with most tubes that the other denominator terms may be neglected and the source impedance written in the form usually given for low frequencies;

$$Z_g \Big|_{\text{low frequency}} = \frac{R_p}{1+\mu} \cong \frac{1}{G_m}. \tag{17}$$

Equation (15) shows that, if the grid-to-ground impedance is comparable to the grid-to-cathode impedance, (17) is no longer a good approximation. This is only likely to happen in the case where a tuned circuit is connected between grid and ground so that the stray reactances are tuned out, becoming part of a tuned circuit at parallel resonance. In this case, however, the grid-to-ground impedance may assume values so large

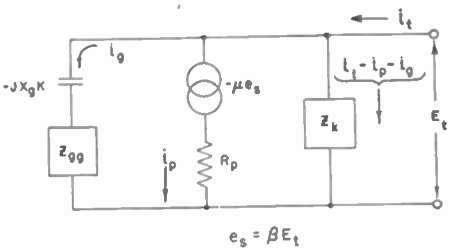

Fig. 7—Shunt circuit represented by the cathode load impedance Z_k and the cathode-to-grid reactance in series with impedance Z_{gg}.

compared to X_{gk} that β approaches zero and the term $(1+\mu\beta)/R_p$ approaches $1/R_p$. The source impedance then becomes quite large and the expected advantages of the cathode follower fail to materialize.

From the foregoing, it is clear that, if a cathode follower is driven by a tuned grid circuit at frequencies of the order of megacycles, the operation is likely to be unsatisfactory. The output will be lower than expected due to the increased source impedance and oscillation is likely at either the tuned circuit frequency, the input leads frequency, or both. This presupposes that the cathode load impedance will be capacitive, a probable result of the cathode-to-heater capacitance and other strays.

III. Remedial Measures

If operation with a tuned grid circuit is required, several methods are available for nullifying, or at least reducing, the effects of grid-to-cathode coupling.

(a) The oscillation may be suppressed by inserting a resistor of suitable size in series with the grid to damp the circuit. Depending on the frequency, this resistor will load the tuned circuit in a manner to be analyzed later. This method has definite limitations.

(b) A coil may be placed from grid to cathode, with a suitable dc blocking capacitor to resonate the grid-to-cathode capacitance (Fig. 8). This increases the value of X_{gk} by a very large amount and is effective in lowering the source impedance and suppressing oscillation at the design frequency. However, oscillation at frequencies determined by the grid-circuit leads may still occur since Z_{gk} is high at only one frequency. This method of operation is subject to the serious disadvantage that the coil must be changed for each change of operating frequency.

It is, of course, possible to make the blocking capacitor C_B variable so that the total reactance may be ad-

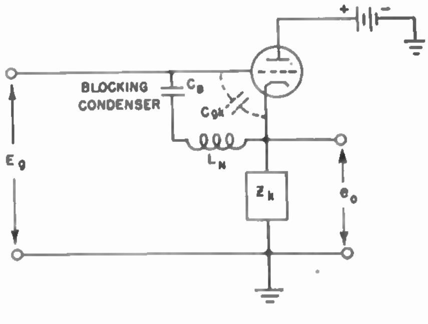

Fig. 8—Blocking capacitor C_B in series with L_N to resonate the grid-cathode capacitance C_{gk}.

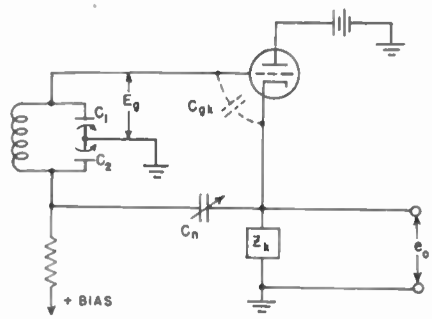

Fig. 9—Bridge neutralization using grid-to-cathode capacitance as one arm.

justed to the correct value at each operating frequency.

(c) The analogy of the foregoing to the familiar "coil neutralization" used in fixed-frequency broadcast and other transmitters leads to the circuit of Fig. 9 using the grid-to-cathode capacitance as one arm of a bridge circuit. Variations of this circuit are those of Figs. 10 and 11.

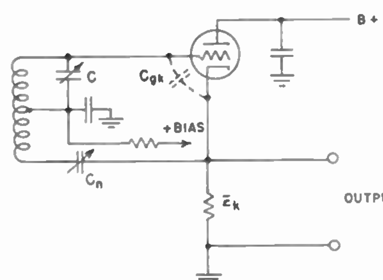

Fig. 10—Alternate method of coil neutralization from that used in Fig. 9.

To analyze the effects of these circuits upon the source impedance, refer to (16). It is seen that the only variable is β and that if it can be kept close to unity the source impedance will remain low. Remembering that β was defined as the ratio of grid-cathode to grid-ground voltages, we may draw the equivalent circuit of the cathode follower with bridge neutralization, look into the output terminals toward the tube, and deduce the source impedance. The equivalent circuit is given in Fig. 12. It is unnecessary to consider Z_k in the equations since it is across the bridge and does not disturb the ratio of the arms. Referring to Fig. 13, the equations to be solved are

$$Z_{gk}i_2 + Z_1 i_3 = e_{in} \tag{18}$$

$$Z_n i_1 - (Z_n + Z_L + Z_{gk})i_2 + Z_L i_3 = 0 \tag{19}$$

$$Z_2 i_1 + Z_L i_2 - (Z_2 + Z_L + Z_1)i_3 = 0. \tag{20}$$

Simultaneous solution of the above yields the currents

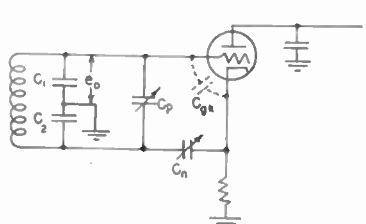

Fig. 11—Another method of coil neutralization than that used in Figs. 9 and 10.

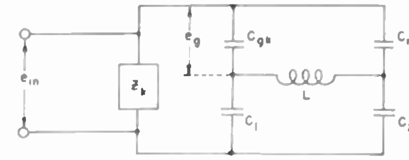

Fig. 12—Equivalent circuit of the cathode follower with bridge neutralization.

which may then be used to find the desired ratio of voltages. This is

$$\beta = \frac{e_o}{e_{in}} = \frac{C_1 + C_2 + j\omega Z_L C_1(C_n + C_2)}{C_1 + C_2 + C_{gk} + C_n + j\omega Z_L(C_1 + C_{gk})(C_n + C_2)}. \tag{21}$$

The conditions for balance of the bridge are that the ratio C_{gk} to C_n should be the same as the ratio of impedances on each side of the tank-circuit ground tap. This is expressed by

$$\frac{C_1}{C_2} = \frac{C_{gk}}{C_n}. \tag{22}$$

If (22) is substituted in (21) there results

$$\beta = \frac{C_1}{C_1 + C_{gk}}. \tag{23}$$

This simple relation is seen to contain no frequency term. Since the tuning capacitor C_1 is nearly always much larger than the grid-to-cathode capacitance C_{gk}, β is nearly always close to unity.

If the circuit of Fig. 11 is used, the output voltage may be increased by making $C_2 = nC_1$ and $C_N = nC_{gk}$, where $n > 1$. Equation (23) still holds.

Theoretically, for perfect balance to exist it is necessary that the stray reactances on both sides of the tank circuit be balanced, and that if a resistor is used in series with the grid another be placed in series with the neutralizing capacitor. In practice, sufficiently good results are usually obtained without these precautions. With any of the three neutralizing circuits shown (Figs. 9, 10, or 11) measurements indicate that the source impedance is constant to about 10 per cent when the grid circuit is tuned through resonance. Even better operation can be obtained at one frequency by careful adjustment.

Circuits in which the *coil* center tap is grounded for rf should be avoided as they will tend to oscillate at the resonant frequency of the circuit between grid and ground, usually 20 or 30 Mc. Parasitics in the split capacitor circuits shown (Figs. 9 and 11) usually occur at frequencies determined by the lead path from the grid through the upper half of the split capacitor to ground.

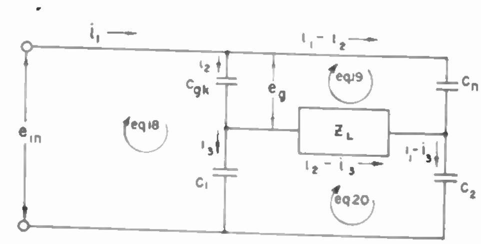

Fig. 13—Circuit analysis of the currents which are used to find the desired ratio of voltages.

This path is normally much shorter so that the oscillation has a frequency of several hundred Mc and is much less vigorous. In all split-capacitor circuits tested in the frequency range from 2 to 30 Mc. a grid resistor of 47 ohms was sufficient for complete stability, and a value of 10 ohms was sufficient in many cases.

It is important to keep this resistor as small as possible, since it may have a serious effect on the input circuit Q at the higher frequencies. This is due to the fact that the grid circuit voltage causes a current to flow through a path composed of the grid-cathode capacitance, the neutralizing capacitance, and the grid resistor. The loss due to this current results in a lowering of the Q of the input circuit. The amount of loss is directly proportional to frequency squared since the current is proportional to frequency.

By simple transformations, it is possible to postulate an equivalent resistance in series with the tank circuit producing the same loss. The working Q of the circuit can then be expressed in terms of the unloaded Q, the circuit parameters, and the frequency. The relation is

$$Q_{eff} = \frac{\omega L}{\dfrac{\omega L}{Q} + (R_g + R_n)\left(\dfrac{n^2 C_{gk}^2}{[C(1+n) + nC_{gk}]^2}\right)} \quad (R_g \ll X_{gk}) \tag{24}$$

where

ωL = reactance of tank coil
Q = unloaded Q of tank coil
R_g = grid resistor
R_N = resistor in series with neutralizing capacitor if used
C = total effective tuning capacitance
Q_{eff} = effective Q of input circuit
$n = C_n/C_{gk}$.

With commonly used circuit values, a grid resistor of 47 ohms will lower the circuit Q by one half at 20 Mc.

IV. Permissible Input Voltage

One aspect of cathode follower operation not usually stressed in the literature is the effect of a complex cathode load on the magnitude of signal which may be applied to the input circuit.

Consider the case where the cathode load is a complex impedance. Then the input and output voltages have a phase angle α as discussed before, and it is found that the grid-to-cathode voltage, which is the vector difference between the input and output voltages, must of necessity become comparable to those voltages when α assumes moderate values. The magnitude of this effect depends on the value of the gain $|A|$ as well as the angle α, being greatest when $|A|$ is large. It is easily evaluated by setting the the output voltage equal to the input voltage times the complex gain. Thus

$$e_0 = E_g A\underline{/\alpha}. \tag{25}$$

The grid-cathode voltage e_s is given by

$$\bar{e}_s = \bar{E}_g - \bar{e}_0 = E_g(1 - A\underline{/\alpha}). \tag{26}$$

Dividing (26) by (25) gives

$$\frac{e_s}{e_0} = \frac{1}{A\underline{/\alpha}} - 1. \tag{27}$$

Equation (27) is plotted in Fig. 14 for various values of α. Inspection of Fig. 14 shows that for moderate gains of the order of 0.8, the grid-to-cathode voltage becomes equal to the output voltage at $\alpha = 45°$ so that the signal which may be handled under class A conditions is quite limited. The ratio of grid-to-cathode voltage to input (grid-to-ground) voltage is found by dividing the above result by $|A|$.

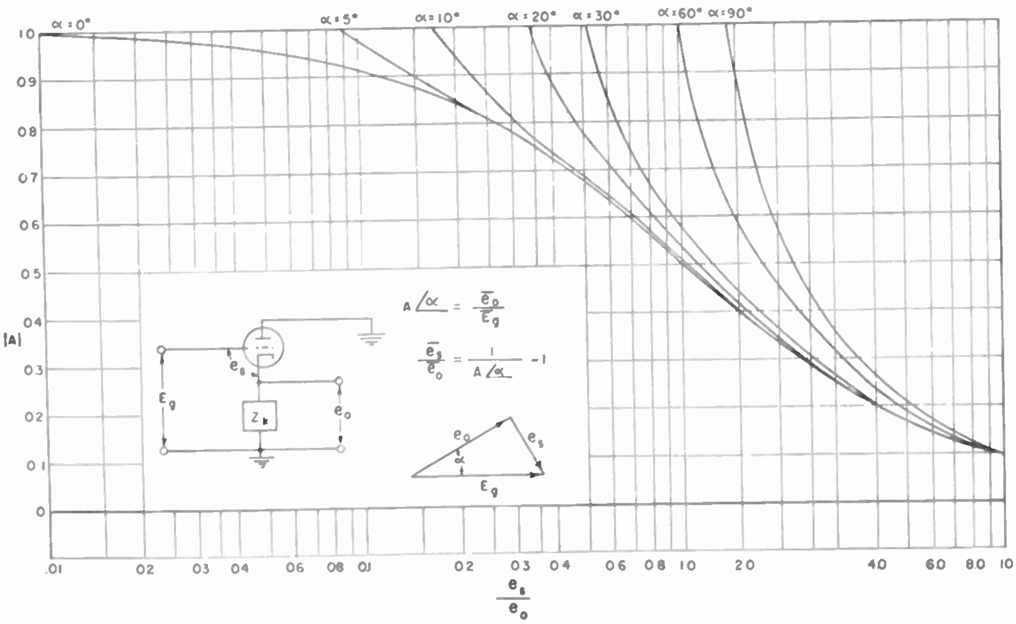

Fig. 14—Ratio of e_s/e_o for various values of complex gain factor $A\underline{/\alpha}$.

Noise from Current-Carrying Resistors 20 to 500 Kc*

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Summary—Measurements have been made of the effect, first studied by Bernamont that when a direct current is flowing through certain types of resistors, the noise voltage generated at their terminals is considerably in excess of the thermal value. The effect has been investigated for solid carbon-composition, "metallized," palladium film, and "cracked-carbon" resistors, for resistance values from 1,000 to 30,000 Ω, currents from 1 to 10 milliamperes, and at frequencies between 20 and 500 kc.

For some individual resistors, especially of the solid carbon-composition type, the noise voltage measured at any frequency in the range stated (with direct current flowing) showed much larger fluctuations than those characteristic of thermal noise, with a detecting system having a response up to a few cps. The fluctuations frequently reached peak amplitudes of several times the mean level of the noise. Several records of the fluctuations, taken on a recording meter, are shown. The fluctuations were particularly large when the current was first applied to any resistor, while its resistance changed to a stable value at a higher temperature, suggesting an analogy to the Barkhausen effect in the magnetization of ferromagnetics. No simple correlation of the occurrence of the fluctuations with other factors has been found.

I. Introduction

THAT NOISE in excess of thermal noise is generated by certain types of resistors when a steady current flows through them was first observed in detail by Bernamont in 1934,[1–4] although a brief reference to what was probably the same phenomenon was given by Hull and Williams in 1925,[5] and a similar effect for a mixture of granular carbon particles was mentioned by Kawamoto in 1919[6] and by Frederick in 1931.[7]

Bernamont found the spectral density of this extra noise to be often many times that of thermal noise for the same resistor, and to show a marked decrease with increase of frequency. Since Bernamont, many investigations, both theoretical and experimental, have been made of this type of noise.[8–17] The experimental results have commonly shown that the spectral density of this noise is proportional to some power of the current close to 2 (values from 1.5 to 2.7 are reported) and for frequencies below about 10 kc varies inversely as the first power of the frequency. At higher frequencies, the spectral density decreases more rapidly.

No name has so far been attached to this additional component of noise from current-carrying resistors. We will refer to it as current noise throughout this paper.

This paper presents more comprehensive data than has so far been reported on current noise from several types of solid carbon-composition and "metallized" (i.e., carbon-composition film) resistors in common use, showing variation with resistor type, resistance value, dc current, and frequency. Measurements have been made with one-half, 1-, and 2-watt resistors, having resistance values between 1,000 and 30,000Ω, carrying steady currents of 1 to 10 milliamperes, the noise output being measured at frequencies between 20 and 500 kc.

There is also described in some detail another noise phenomenon, to which only one slight reference has previously been reported,[3] a case of current-noise voltages whose fluctuations about their mean value far exceed the fluctuations characteristic of thermal or shot-effect noise, for the same mean value of noise and otherwise identical conditions of measurement.

II. Apparatus and Method of Measurement

Fig. 1 shows a block diagram of the equipment used in the noise measurements. The resistor under test is mounted (closely shielded) in unit A, which contains

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[15] L. I. Schiff, "Noise in radar crystal detectors," (Abstract), *Phys. Rev.*, vol. 69, p. 682; June 1 and 15, 1946.

[16] E. J. Harris, W. Abson, and W. L. Roberts, "Semi-conductor noise at low frequencies," T.R.E. Report T 2051 (unclassified), November 22, 1946.

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the necessary power supply, filters, chokes, etc., to allow passing measured dc currents up to several milliamperes through the resistor, while maintaining a suitably high shunt impedance across it at the frequency of measurement. The unit also contains a noise diode (British type CV172), with auxiliary circuits, which is used as a calibration-noise source. Its load is a noninductive wire-wound resistance.

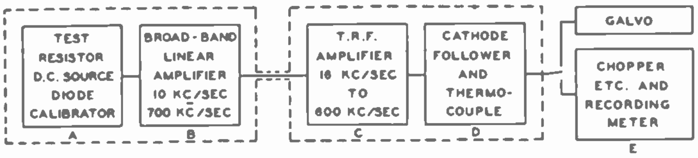

Fig. 1—Schematic diagram of the apparatus used in the noise measurements.

The noise output of unit A is amplified by unit B, which is a broad-band linear amplifier with cathode-follower output to unit C. The maximum gain of unit B is about 33 db over the frequency range 10 to 700 kc.

Unit C is a modified low-frequency radio receiver type TE236B, built by RCA Victor Co., Ltd., Canada, for the Royal Navy. This tunes over the frequency range 15 to 600 kc. The 3 tuned amplifier stages result in an effective noise bandwidth that varies from about 1 to 4 kc over the tuning range of the amplifier, having the lower value at the lowest frequency. The calibration method used in the measurements makes it unnecessary to know this bandwidth.

The output of unit C goes to a cathode follower in unit D, the output of which is capacitively coupled to the heater of a vacuum thermocouple. The output of the thermocouple is filtered with a low-pass filter. For mean-square noise-voltage measurements, it is then connected directly to a galvanometer. To obtain a record of the fluctuations in the filtered thermocouple output, it is fed to unit E, where it is chopped at 60 cps by a "high-speed" relay. The chopped signal is amplified, rectified, filtered, and fed to a recording milliammeter of maximum range 5 milliamperes. While the input to the thermocouple is at the frequency to which unit C is tuned, the highest frequency in its output voltage is limited by its thermal time constant, which is of the order of 0.1 seconds. The time constant of the recording meter is somewhat higher than this, but less than 0.5 seconds. The filters used have low attenuation at frequencies below 10 cps. The result is a recording system whose response begins to fall off at a frequency of about 1 cps and is small at frequencies above 10 cps.

III. Experimental Measurements

1. Calibration Technique

For any test resistor in unit A, if the filtered output voltage of the thermocouple in unit D has only "normal" fluctuations, this voltage is a quasi-linear function of the mean-square noise voltage at the input of the thermocouple, which noise voltage is made up of contributions from shot effect, flicker effect, thermal noise, and current noise. However, mean-square noise voltages from various sources are directly additive in producing the resultant mean-square noise voltage, provided the voltage amplification of the apparatus involved is linear for the range of amplitudes covered.

If, therefore, the test resistor terminals in unit A are shorted, the thermocouple output voltage will then be a measure of the noise generated in the amplifier itself, which noise will remain as a constant mean-square component when additional noise is added. When a test resistor is connected, the thermocouple output will increase because of the thermal noise contributed by the resistor.

If now a small dc current is caused to flow through the test resistor, it is found in general (but not with wire-wound resistors), that the thermocouple output rises again, the increase in this case being due to a current-noise component. The amount of this current-noise contribution can be measured by connecting the diode noise generator in place of the test resistor and adjusting its plate current to give the same two thermocouple outputs that were recorded for the test resistor with and without the dc current flowing. From the well-known calculable noise properties of a saturated diode, the spectral density of the current noise, expressed for example as mean-square volts per 1 cps noise bandwidth, can be found from the values of the diode plate current and the diode load resistor. The gain, frequency, and bandwidth of the intervening amplifiers are not involved in the calculation.

This is the method of measurement and calibration that has been used throughout the present work. It means that values given for current-noise spectral density for any resistor are for that component of noise alone.

2. Mean-Square Current-Noise Voltages

The current-noise output of a typical one-half watt "metallized" resistor having a dc resistance of 11,100Ω is shown in Fig. 2 as a function of current and frequency. At any frequency, the variation is as the square of the current, within experimental error. The variation with frequency at any current, however, does not follow a simple law over the frequency range of the measurements. At frequencies below 50 kc, the very limited data suggest that the noise spectral density varies as an inverse power of the frequency somewhat less than unity. At higher frequencies, an inverse power of about 1.6 is indicated.

It is evident that the frequency-variation law is independent of current, and that the current-variation law is independent of frequency.

Frequency-variation data similar to that of Fig. 2 for 6 different types of resistors of approximately equal dc resistance and all carrying the same value of constant

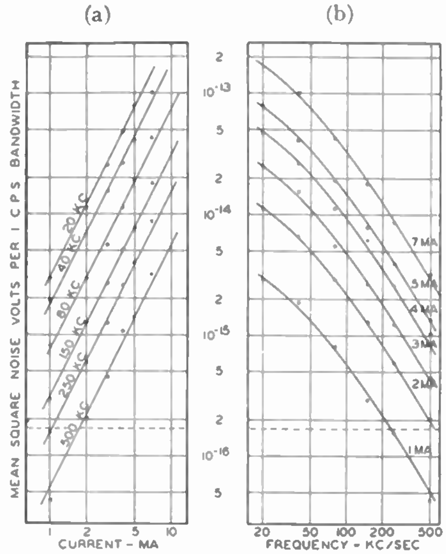

Fig. 2—Spectral density of current noise for a one-half watt "metallized" resistor, dc resistance 11,100 Ω, resistor *A* of Table I. (a) as a function of current at various constant frequencies; the full lines have a slope of 2; the broken line is the theoretical spectral density for thermal noise at room temperature. (b) as a function of frequency at various constant currents; the curves are of identical shape with vertical displacement; the negative slope at low frequencies is slightly less than 1, and at high frequencies is about 1.6.

dc current are given in Fig. 3, which shows that the physical nature of a resistor determines very markedly the magnitude of its current-noise output. The frequency-variation law, however, is apparently identical for the different resistors.

Measurements on a large number of each of 3 types of resistors, and covering a range of resistance values, are plotted in Fig. 4. The frequency and the dc current are constant throughout. This shows that apparently identical resistors (see the points for 20 1-watt "metallized" resistors having dc resistances between 11,000 and 12,000Ω) have current-noise outputs varying by a factor of 20. It is, therefore, not possible to infer with any accuracy a resistance-variation law for current noise. The full lines shown correspond to a square law of variation, which does appear to fit the data at least better than a line corresponding to a first-power law, which would be parallel to the dotted thermal-noise line.

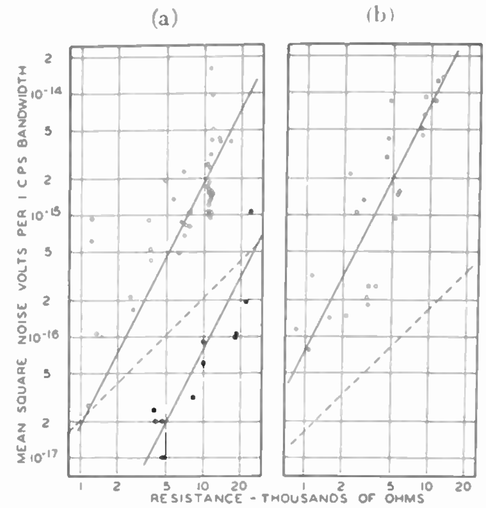

Fig. 4—Spectral density of current noise for 79 resistors of 3 types, as a function of dc resistance value. The current is 5 milliamperes and the frequency of measurement 150 kc in every case. (a) the open circles are for 1 watt "metallized" resistors (including resistor *B* of Table I); the black circles are for one-half watt carbon-composition resistors (including resistor *D* of Table I). (b) one-half watt "metallized" resistors (including resistor *A* of Table I). The full lines have a slope of 2. The broken lines show the corresponding theoretical thermal noise level and have a slope of 1.

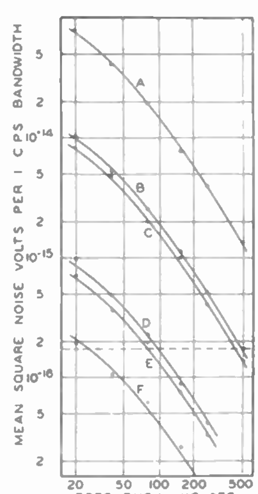

Fig. 3—Spectral density of current noise for 6 different types of resistor (see Table I below) as a function of frequency. The current in each case is constant at 5 milliamperes. The range of dc resistance is from 9,800 to 11,100 Ω.

TABLE I

Resistor	DC Resistance Ω	Wattage Rating	Type
A	11,000	½	"metallized"
B	11,000	1	"metallized"
C	10,300	2	"metallized"
D	9,800	½	carbon composition (Ohmite Little Devil)
E	10,300	1	"cracked carbon"
F	10,200	1	palladium film
G	11,400	½	carbon composition

3. Fluctuations in Current-Noise Voltages

While making the current-noise measurements described in the last section, it was found that, for some resistors and currents, the galvanometer indicating the thermocouple output fluctuated wildly. This was in marked contrast to its fluctuations when the noise input to the amplifier was either thermal noise or diode shot-effect noise, in which cases its peak fluctuations did not exceed about 5 per cent of its mean reading, a figure which is in reasonable quantitative agreement with the theoretical conclusions of Rice[18] who has calculated what the ratio of standard deviation to mean value should be for random noise energy, as a function of bandwidth and averaging time for any detector. While his calculations are for noise with a spectral density in-

[18] S. O. Rice, "Filtered thermal noise—fluctuation of energy as a function of interval length," *Jour. Acous. Soc. Amer.*, vol. 14, pp. 216–227; April, 1943.

dependent of frequency, the variation of spectral density of current noise over the bandwidths used in the present work is sufficiently small that his conclusions should apply.

Unit E of the apparatus, described in section II, was designed to permit comparison of these low-frequency fluctuations in the current-noise output of different resistors, by making graphical records of the thermocouple output voltage as a function of time.

Fig. 5 shows portions of typical records of these fluctuations.

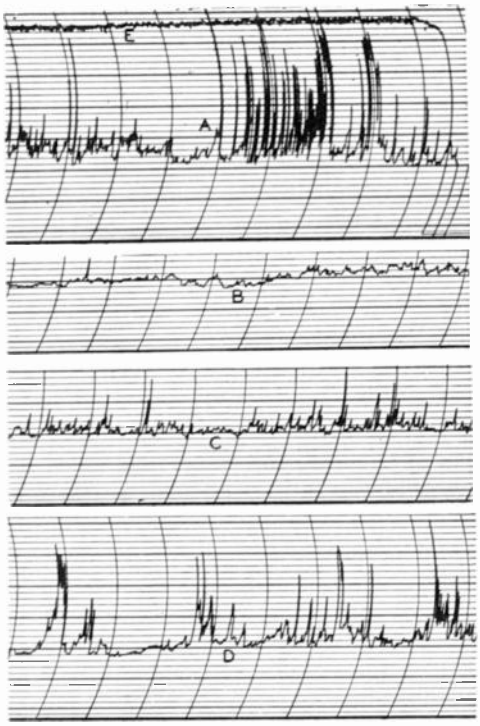

Fig. 5—A, B, C, and D are 35-minute sections of a continuous 16-hour record of fluctuating current noise from a one-half watt carbon-composition resistor (resistor G of Table I). The time scale is from right to left. The current is constant at 5 milliamperes and the frequency of measurement 150 kc throughout. The amplifier gain is the same for all four traces, and for trace E.
Trace A—the initial 35 minutes of record after connecting the resistor;
Trace B—a section showing very low fluctuation;
Trace C—a section showing average fluctuation; about one half the 16-hour record was similar to this;
Trace D—a section showing large fluctuations; 4 or 5 sections of this type occurred during the 16-hour record.
Trace E is for comparison and shows nonfluctuating current noise from resistor A of Table I. The vertical scale of the charts, while not linear, is indicated roughly by the location of trace E, which from Fig. 2 corresponds to a noise spectral density of 8×10^{-15} mean-square volts per 1 cps bandwidth.

Trace E is a representative section from a continuous 8-hour record of the noise output (mainly current noise) from a one-half watt "metallized" resistor of dc resistance 11,100Ω (resistor A of Figs. 2 and 3 and Table I), carrying a current of 5 milliamperes. The measurement frequency was 150 kc. The fluctuations in this record are of the same magnitude as occur in records of shot-effect noise or thermal noise having the same mean level. They may, therefore, be regarded as "normal" fluctuations.

Traces A, B, C, and D are sections from a single continuous 16-hour record of the noise output from a one-half watt solid carbon-composition resistor, dc resistance 11,400Ω (resistor G of Table I). The amplifier gain, the dc current, and the measurement frequency are the same as for trace E.

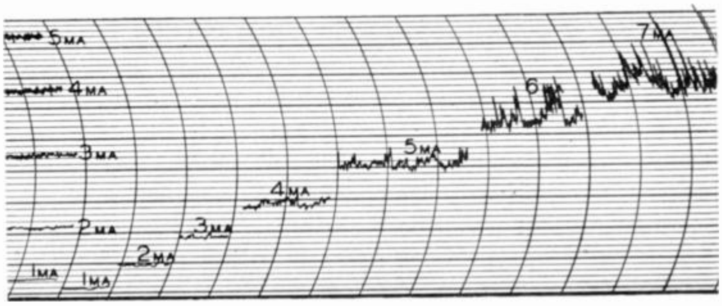

Fig. 6—Fluctuations of current noise as a function of current. The traces at the left are for resistor A of Table I, those at the right for resistor G. The frequency is 150 kc.

Fig. 6 shows records, each of a few minutes duration, of the noise from the same two resistors for various values of dc current. All the conditions for the 5-milliampere record from resistor G are the same (except amplifier gain) as in Fig. 5. The fluctuations in Fig. 6 appear by comparison to be below average for this particular resistor.

A large number of traces taken with these and other resistors and at frequencies from 20 to 500 kc, all other factors remaining constant, show no change in the nature of the fluctuation phenomenon with amplifier frequency for any one resistor. Quantitative data have not been obtained, nor has it been possible to determine whether or not the abnormal fluctuations are simultaneous at different frequencies.

IV. Relation of Experimental Results to Previously Published Work

Attempts to develop theoretical expressions for the spectral density of current noise as a function of current and frequency have most commonly arrived at an equation of the form

$$A(f, I) = \text{constant}\cdot I^2\cdot\frac{1}{1+\omega^2\tau^2}$$

where A is the spectral density (for example in mean-square volts per unit bandwidth) at frequency f with current I flowing, ω is $2\pi f$ and τ is some characteristic time of the postulated noise mechanism. Various physical significances have been attached to τ, such as the free life of the conduction electrons between release and capture,[14] a time constant of fluctuations in the total number of current carriers,[4] and the time of influence of stray atoms on boundary layers.[16]

If τ is single valued, this form of equation implies a spectral density independent of frequency at low frequencies, and having an inverse square variation at high frequencies. Some small intermediate range of frequencies would show an inverse first-power relationship.

Experimentally, no one seems ever to have observed low-frequency spectral density variation at any inverse power of the frequency appreciably different from unity, even at frequencies as low as 0.3 cps.[16] The present work does not include data at sufficiently low frequencies for comparison with these results, but is in agreement with the original observations of Bernamont[3] at frequencies above 25 kc in finding current-noise spectral density to vary as an inverse power of the frequency greater than 1 and approaching 2. None of the other papers listed shows data for this frequency range.

A formula of the type mentioned does not fit the results of the present work as given in Figs. 2 and 3 if τ is assumed to be single valued. It is obvious that for some distribution of values of τ such an equation could be made to fit any curve of this general shape.

The theories of current noise which have specifically included resistance value as a variable have predicted variation of mean-square noise voltage as the square of the resistance value.[16] The results shown in Fig. 4 are not incompatible with this.

The fluctuating current noise described in section III, (3) above seems not to have been described previously except for 2 passing comments in Bernamont's major publication[3] about experimental situations where his galvanometer deflections showed "mean-square fluctuations as much as 5 per cent of the mean deflection."

V. Conclusions

The results given in Figs. 3 to 6, and a great deal of additional data, lead to the following conclusions as to the relative merits of different types of resistors with respect to mean-square current noise and abnormally fluctuating current noise:

(a) The single palladium-film resistor available showed considerably less current noise than any other single resistor of comparable resistance value tested, except noninductive wire-wound units.

(b) For the same dc resistance, power rating, and conditions of measurement, the best solid carbon-composition resistors showed markedly less current noise than the best "metallized" resistors; on the other hand, many makes of solid carbon-composition resistors showed more noise than the noisiest equivalent "metallized" units.

(c) For equal power dissipation and similar measurement conditions, resistors of a given make having low power rating generally showed more current noise than those with a higher power rating; resistors A, B, and C in Fig. 3 are all the "metallized" type, made by 1 manufacturer, and all are dissipating the same amount of power (within 10 per cent), but have respectively power ratings of one-half, 1, and 2 watts. Whether this is a temperature phenomenon or a volume effect (as suggested by Brillouin[8] for thin metal films) has not been determined.

(d) Without exception, all records of current noise from solid carbon-composition resistors have shown abnormal fluctuations. About half the "metallized" resistors tested have shown only normal fluctuations, comparable to those characteristic of thermal and shot-effect noise. Many solid carbon-composition resistors have shown fluctuations much greater than the greatest obtained from " metallized" resistors.

(e) A high mean level of current noise is not necessarily accompanied by fluctuations. Trace E of Fig. 5, for example, shows much greater mean level of current noise than the other 4 traces, but much less fluctuation. Fig. 6 shows, as might be expected, that for any one resistor, the amplitude of the fluctuations in the current noise does increase as the mean noise level increases (resistor G of Table I). Traces of current noise without abnormal fluctuations (resistor A of Table I) are shown for comparison. Rough calculations suggest that the amplitude of the fluctuations increases as about the fourth power of the dc current carried by the resistor.

It can be said that, in general, large fluctuations occur only when the mean level of the current noise is high, but that there are frequent exceptions, especially with solid carbon-composition resistors.

In cases where the current noise is less than the thermal noise, as for resistor F in Fig. 3, the normal fluctuations of the thermal noise conceal any abnormal current-noise fluctuations that might exist.

Trace A of Fig. 5 shows a frequently observed phenomenon that is probably significant. In many records, the fluctuations observed during the first few minutes after connecting a resistor were much greater than in the rest of the record. This suggests that while the resistance is changing to its final equilibrium value at a temperature above room temperature, the change may occur in discontinuous steps, each step producing a pulse of noise voltage because of the current flowing and the nature of the coupling circuits. An analogy with the Barkhausen effect in the magnetization of iron is suggested. The magnitude of the fluctuations observed must mean internal rearrangements in the resistor material involving units of a very large number of atoms.

In conclusion, it is suggested that, in many practical situations, the selection of a resistor for either low mean current-noise level, or for low amplitude of current-noise fluctuations, may lead to improved operating conditions. It seems likely that current noise may often be the major component of the total noise in such equipment as broad-band amplifiers at audio and low radio frequencies when carbon-composition type resistors are used in coupling circuits.

VI. Acknowledgments

This work has been supported, for the purchase of equipment and the payment of part salary to one of us (R. H. C., Jr.) by a grant from the Defence Research Board, of the Department of National Defence, Ottawa, Ont., Canada. We wish to express our appreciation for this assistance and for permission to publish the results.

An Analysis of the Sensing Method of Automatic Frequency Control for Microwave Oscillators*

EUGENE F. GRANT†, ASSOCIATE, IRE

Summary—An analysis is made of the type of automatic-frequency-control circuit that uses a simple cavity resonator for the stable element and either frequency modulation of the controlled oscillator or modulation of the cavity resonance frequency to obtain effective discriminator curves which give a null output for a frequency equal to the average cavity resonance frequency. An analysis is made of the complete automatic-frequency-control loop gain, including the transmission through the cavity as a function of the various parameters and the frequency-modulation swing. A discussion is presented of the best method of decreasing the pulling of the oscillator frequency by the cavity, and the pulling of the cavity frequency by a load which has a variable susceptance.

Glossary

A_1 = the power attenuation between the oscillator and the cavity.
A_2 = The power attenuation between the cavity and the detector
a, b, c, d = the constants characterizing the general four-terminal network
$C(W, V, n)$ = the fourier component coefficient of the frequency modulation present in the detected output
δ = per unit frequency change
$D(W, V, n)$ = the second fourier component coefficient of the frequency-modulation frequency present in the detected output
$E_c(t)$ = the detected output voltage as a function of time
E_{c1} = the fundamental component of the sensing frequency in $E_c(t)$
E_{c2} = the second-harmonic component of the sensing frequency in $E_c(t)$
E_1 = the voltage across the input terminals of a four-terminal network
E_2 = the voltage across the output terminals of four-terminal network
f = the instantaneous frequency of the oscillator
f_0 = the resonant frequency of the cavity
ΔF = the total frequency modulation or cavity resonant frequency swing
Δf = the center frequency deviation of the oscillator from f_0
f_m = the frequency-modulation sensing frequency
G = the loop-gain, measured by breaking the loop at any point
G_a = the ratio of the output of the phase discriminator to the input of the amplifier
G_f = the ratio of the frequency deviation of the oscillator to the control-element displacement. (This may be either a voltage or a mechanical displacement depending on the control method used.)
I_1 = the current flowing into a four-terminal network
I_2 = the current flowing out of a four-terminal network
K_n = detector constant; $E_c(t) = K_n P_c^{n/2}$
n = the detector-law exponent; n is equal to unity for a linear detector, and to 2 for a square-law detector
P_{osc} = the power output of the oscillator
P_c = power incident on crystal detector
Q_1 = the input Q of the cavity
Q_2 = the output Q of the cavity
Q_0 = the unloaded Q of the cavity
Q_L = the loaded Q of the cavity
Q_{osc} = the loaded Q of the oscillator

$$\frac{\partial C(0, V, n)}{\partial W} = \left.\frac{\partial C(W, V, p)}{\partial W}\right| W = 0$$

T = power relative to incident power transmitted through a cavity
$\theta = 2\pi f_m t$
$V = Q_L \Delta F / f_0$, which is the frequency-modulating swing in terms of the cavity 3-db bandwidth
$W = 2Q_L \Delta f / f_0$, which is the frequency deviation in terms of one-half the cavity 3-db bandwidth
Y_{in}, Y_{out} = admittance relative to the admittance of matched load.

Introduction

There are several methods of accomplishing automatic frequency control of oscillators operating in the microwave region. The five methods which are commonly used at present are:

1. The Pound dc discriminator.[1,2]
2. The Pound ac discriminator.[2,3]

* Decimal classification: R355.6×R119.39. Original manuscript received by the Institute, November 17, 1947; revised manuscript received August 19, 1948.
† Electronic Research Laboratories, Air Matériel Command, Cambridge 39, Mass.

[1] R. V. Pound, "Frequency stabilization of microwave oscillators," Proc. I.R.E., vol. 35, pp. 1405–1415; December, 1947.
[2] C. G. Montgomery, "Technique of Microwave Measurements, Radiation Laboratory Series No. 11," McGraw-Hill Book Co., Inc., New York, N. Y., 1947.
[3] W. G. Tuller, W. C. Galloway, and F. P. Zaffarano, "Recent developments in frequency stabilization of microwave oscillators," Proc. I.R.E., vol. 36, pp. 794–800; June, 1948.

3. The use of absorption lines.[4]
4. The method of frequency modulation.
5. The method of resonant-circuit sensing.[5]

Some of the design details of the methods (4) and (5) are to be discussed in this paper. As in all control systems, the general method of operation is as follows: The frequency of the oscillator is compared with that of the stable element (in this case a resonant circuit or a cavity resonator), with the result that a voltage (or a displacement) proportional to the amount of deviation is produced. This voltage is then applied to a frequency-controlling element on the oscillator in such a direction as to return the oscillator frequency to the frequency of the stable element. In the microwave region, a transmission cavity can be made to have the property that the power transmission is a maximum at a particular frequency, but decreases very rapidly as the frequency is deviated from that of maximum transmission. Such cavities can be made to have very high Q's and a relatively stable resonant frequency. A cavity cannot be used directly as the stable element, since the transmission is an even function of the frequency deviation from resonance. Therefore, no information can be obtained from the output, whether the deviation is plus or minus. One method of realizing a voltage which shall be positive or negative dependent upon the direction of the deviation of the oscillator frequency is to apply a small amount of frequency modulation to the oscillator frequency. It will be observed in Fig. 3 that the output of the cavity will be amplitude modulated by an amount which is dependent upon the relative amount of frequency modulation and the average oscillator frequency. In particular, it will be observed that the phase of the resulting amplitude modulation reverses with respect to the phase of the frequency modulation when the oscillator frequency passes from above the cavity resonance frequency to below it. This fact is made use of by providing an amplitude-modulation detector (usually a silicon crystal detector for the microwave region) at the output of the cavity, and a suitable amplifier to operate a phase detector. The output of the phase detector is then not only proportional to the magnitude of the deviation of the average oscillator frequency from the resonant frequency of the cavity, but agrees with the algebraic sign of the deviation. Fig. 1 is a complete schematic of such a system of control. The sensing frequency is defined as the rate of the frequency modulation on the oscillator.

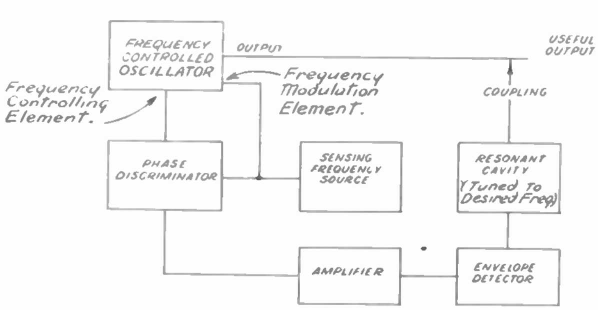

Fig. 1—Schematic of the frequency-modulating method of automatic frequency control.

From the point of view of the over-all system behavior, it makes little difference whether the oscillator is frequency-modulated or the reasonance frequency of the cavity is caused to vary about a mean frequency at the sensing frequency rate. The operation is essentially the same. This is the method of resonant-circuit sensing. See the schematic of this system in Fig. 2. Some over-all design considerations will dictate which method is preferable. There are several methods available for causing the resonant frequency of a resonator or cavity to vary. (Since, in general, this paper will be concerned with a microwave system where cavities are generally employed as the resonant system, the word cavity will be used to indicate a resonator. However, the arguments and the analysis may be generally applicable to a system which operates at a frequency where lumped components may be used.)

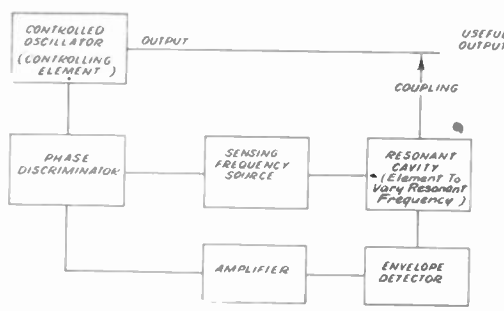

Fig. 2—Schematic of the resonant-circuit sensing method of automatic frequency control.

1. The cavity may be equipped with a diaphragm or a plate which is displaced harmonically at the sensing frequency, causing a variation of a suitable dimension of the cavity.
2. The cavity could be constructed in such a manner that it could be resonant mechanically to the sensing frequency and driven in a suitable manner; a convenient means would be a piezoelectric crystal, a magnetostriction bar, or even an electromechanical drive. The changing dimensions would effect a periodic change in the resonant frequency of the cavity.
3. A variable reactance could be coupled to the cavity. A nonlinear element, such as a silicon or germanium crystal, may be used if placed in a suitable network.

In order to simplify the ensuing discussion and arguments, only the method of frequency modulation will be discussed. However, from the standpoint of analysis,

[4] W. V. Smith, Jose L. Garcia de Quevedo, R. L. Carter, and W. S. Bennet, "Frequency stabilization of oscillators by spectrum lines," *Jour. Appl. Phys.*, vol. 18, p. 112; January, 1947.

[5] G. G. Gerlach, "A microwave relay communications system," *RCA Rev.*, vol. 7, pp. 560–600; December, 1946.

the systems are practically identical. It is only a question of the reference for the time scale. Hence, to apply this discussion to a resonant circuit sensing system, it is only necessary to realize that the frequency swing of the cavity from its resonant point is identical to the frequency-modulation swing of the oscillator.

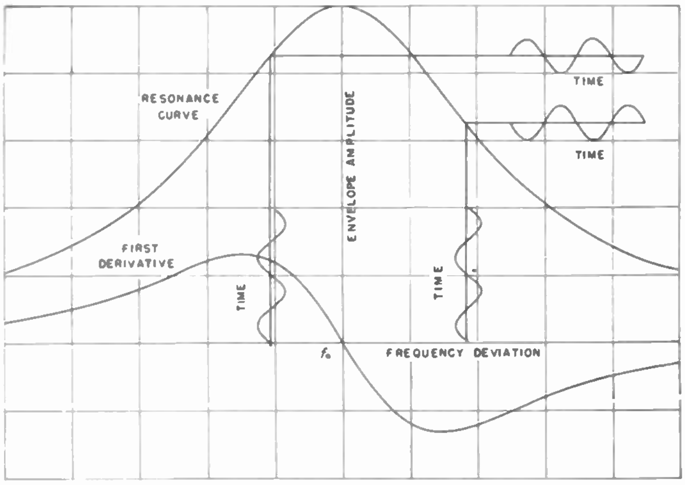

Fig. 3—Cavity characteristics.

Fundamental to the design of a system which uses a cavity in this manner for the stable element, is the relationship between the frequency-modulation swing and the sensitivity of the output of the phase detector in terms of the various circuit parameters. An important element of the system is the amplitude-modulation detector. Usually, this detector will be a crystal rectifier, which has a law relating the input amplitude to the output amplitude not easily described. However, for low-level operation the behavior will approach a square-law relationship, and for high-level operation, the behavior will approach a linear relationship. The analysis is carried out for the square-law detector and the linear-law detector. For the case where the operation is not described by either of these two laws, a heuristic approximation will have to be made on the basis of experience and the analyzed cases. It is more convenient to discuss the behavior of a cavity in terms of its Q's and resonant frequency f_0 than in terms of its equivalent lumped-constant parameters. The Q's and f_0 are more adapted to direct measurement in the physical system.

An automatic frequency-control system using the method of frequency modulation was developed by the author for use as the beacon-frequency automatic-frequency-control system for the AN/APS-6 airborne radar.

General

The various parameters of the control-system design will be taken up in term. It is assumed that the frequency at which the system is to operate has been decided. It remains, then, to decide the various other cavity parameters, the loop gain of the system, and the padding that must be inserted at various locations. A cavity must be chosen which has a resonant frequency of the desired stability. The normal frequency drift of the oscillator is ascertained to determine the amount of loop gain necessary to control it within the necessary limits. Since the system is a normal negative feedback system, the frequency drift of the controlled oscillator, relative to the resonant frequency of the cavity, is equal to the frequency drift of the oscillator uncontrolled divided by one plus the loop gain. As will be seen later, the loop gain may be increased by increasing the unloaded Q of the chosen cavity, other factors remaining the same. The unloaded Q of a copper cavity is a function of the resonant frequency chosen and of its physical size. That is, a large E-mode cavity will have a much higher unloaded Q than a small re-entrant H-mode cavity for the same frequency. The physical size of the equipment may be a deciding factor in the unloaded Q obtainable.[6] For economy in the design of a control system, each component should be used to its full capability. The loop gain of the automatic-frequency-control circuit is controlled to a very large extent by the cavity parameters, and it would be wise to adjust those parameters such that the effective gain is the maximum obtainable commensurate with the other factors which must be considered.

The analysis of the loop gain in a frequency-control system with frequency modulation will be based upon three assumptions.

1. The sensing frequency and the frequency modulation swing is sufficiently low that the cavity may be considered to be in steady state for each instantaneous frequency. This will be true if the sensing frequency is small compared to the 3-db bandwidth of the cavity and if the maximum rate of change of the instantaneous frequency is less than the square of the 3-db bandwidth of the cavity. (The bandwidth must be expressed in cycles per second.)[7]

2. The relationship between the envelope amplitude applied to the detector and the output voltage can be represented by either a square-law relationship or a linear relationship.

3. The phase discriminator output is linearly related to the amplitude of the fundamental component of the detector output.

Analysis of Effective Gain Through the Cavity

The cynosure for the analysis is the cavity behavior. It is convenient to treat the high-frequency energy in terms of the incident power rather than in terms of its voltage. The reason for this is that the incident power is a relatively easy quantity to measure. Further, for work

[6] T. Moreno (editor), "Microwave Design Transmission Data," Sperry Gyroscope Co., Manhattan Bridge Plaza, Brooklyn 1, N. Y., 1944; chap. X.

[7] Balth van der Pol, "The fundamentals of frequency modulation," *Jour. IEE* (London), vol. 93, pp. 153–158; May, 1946.

of this kind, it is not necessary to know absolute impedance or admittance values, but only the relative values in the circuit. The Q's of the cavity are defined or measured under the same conditions that it will have in the operating system. Some methods of measuring the various Q's of a cavity have been described.[2,8]

It can be shown the instantaneous power transmitted through a cavity for a unit incident power is (see Appendix I)

$$T = \frac{4Q_L^2}{Q_1Q_2} \frac{1}{1 + \left[\frac{2Q_L(f - f_0)}{f_0}\right]^2}. \tag{1}$$

The instantaneous frequency of the oscillator may be expressed in terms of the sensing frequency f_m, the total frequency swing ΔF, and the deviation of the center frequency from the cavity resonant frequency and time. It is seen that

$$f = f_0 + \Delta f + \frac{\Delta F}{2} \sin 2\pi f_m t. \tag{2}$$

Since the high frequency is most easily described in terms of incident power, it is convenient to express the detector law in terms of the relationship between its incident power and the voltage output. Since it is expressed this way, it is very easy to obtain the coefficient in the expression for the limiting cases experimentally. Let the power from the cavity be fed through an attenuator to the crystal detector, which then will be considered to have an instantaneous output voltage related to the input power as follows:

$$E_c(t) = K_n P_c^{n/2}. \tag{3}$$

The combination of the previous expression and the effect of an attenuator between the cavity and the crystal detector and the effect of the decoupling between the oscillator and the cavity results in

$$E_c(t) = K_n(2Q_L)^n \left[\frac{P_{osc}A_1A_2}{Q_1Q_2}\right]^{n/2} \left[1 + \left(\frac{2Q_L\Delta f}{f_0} + \frac{Q_L\Delta F}{f_0} \sin 2\pi f_m t\right)^2\right]^{-n/2}. \tag{4}$$

The above expression can be much simplified by the use of some dimensionless constants. Substituting W, V, and θ for the expressions $2Q_L\Delta f/f_0$, $Q_L\Delta F/f_0$, and $2\pi f_m t$, respectively, results in the following:

$$E_c(t) = K_n(2Q_L)^n \cdot \left[\frac{P_{osc}A_1A_2}{Q_1Q_2}\right]^{n/2} \cdot [1 + (W + V \sin \theta)^2]^{-n/2}. \tag{5}$$

[8] W. Altar, "Q-circles—a means of analysis of resonant microwave systems," Proc. I.R.E., vol. 35, pp. 355–361; April, 1947.

By Fourier's theorem, the fundamental component of the sensing frequency in the detected output is

$$E_{c1} = \frac{1}{\pi} \int_{-\pi}^{\pi} E_c(t) \sin \theta d\theta. \tag{6}$$

Combining (5) and (6), it follows that

$$E_{c1} = K_n(2Q_L)^n \left[\frac{P_{osc}A_1A_2}{Q_1Q_2}\right]^{n/2} \frac{1}{\pi} \int_{-\pi}^{\pi} [1 + (W + V \sin \theta)^2]^{-n/2} \sin \theta d\theta. \tag{7}$$

$$C(W, V, n) = \frac{1}{\pi} \int_{-\pi}^{\pi} [1 + (W + V \sin \theta)^2]^{-n/2} \sin \theta d\theta. \tag{8}$$

$C(W, V, 1)$ and $C(W, V, 2)$ are evaluated in Appendix II. Figs. 4 and 5 show the plots of these functions with

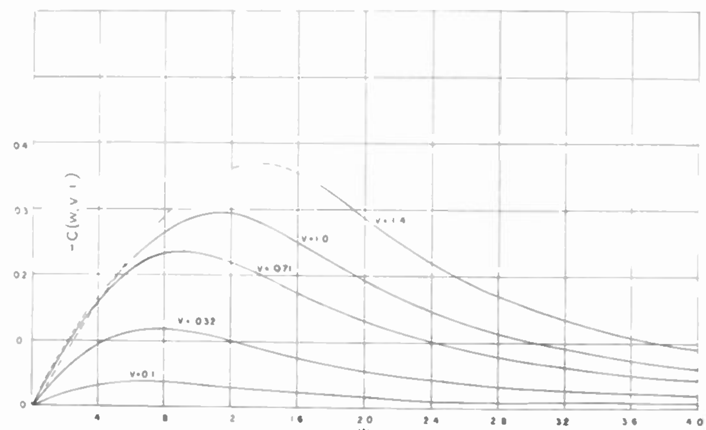

Fig. 4—Effective discriminator curves for a cavity and linear detector, W being proportional to the oscillator frequency deviation, while $-C(W, V, 1)$ is the relative amplitude of the detector output at the sensing frequency.

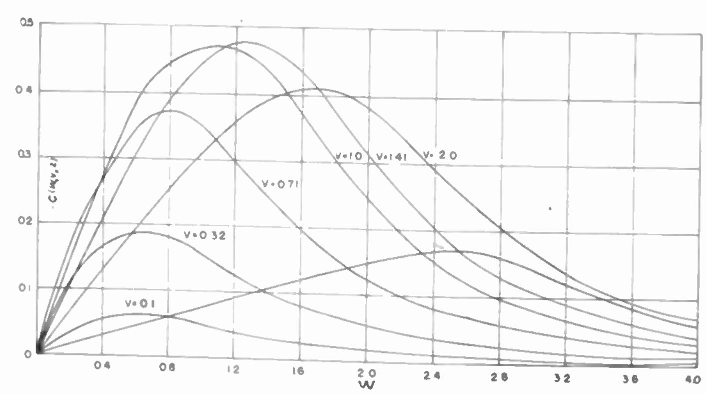

Fig. 5—Effective discriminator curves for a cavity and a square-law detector, W being proportional to the oscillator frequency, while $-C(W, V, 2)$ is the relative amplitude of the detector output at the sensing frequency.

various values of V as a parameter, $C(W, V, n)$ is an odd function of W, so it is only necessary to plot the function for positive values of W.

Since, in general, the operating point of the control system is near the W origin, the function $C(W, V, n)$ may be approximated by the first term in its Taylor expansion

$$C(W, V, n) \cong \frac{\partial C(0, V, n)}{\partial W} W, \quad W \ll 1. \tag{9}$$

The response of the output to a frequency variation on the input is approximately linear, and the system may be treated as a linear system. The expression $\partial C(0, V, n)/\partial W$ may be computed as a function of V. (See Appendix III.) These results are plotted in Fig. 6.

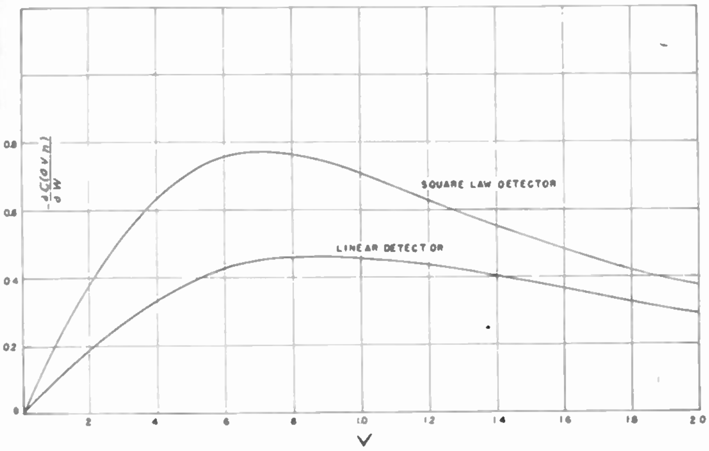

Fig. 6—Slope of the effective discriminator curves at resonance.

These curves indicate that the function $\partial C(0, V, n)/\partial W$ goes through a maximum as a function of V. For n equal to unity, the maximum occurs at V equal to 0.88, and for n equal to 2 the maximum occurs at V equal to 0.707. This indicates that in the system under design, the amount of frequency modulation of the oscillator (or the amount of frequency swing of the resonator) can be adjusted to provide a maximum gain through the cavity system.

The relationship between a small frequency deviation in the frequency of the power incident on the cavity and the fundamental component of the sensing frequency can now be completely stated and interpreted. From (7), (8), and (9), by substitution

$$E_{c1} \cong K_n(2Q_L)^n \left[\frac{P_{osc}A_1A_2}{Q_1Q_2}\right]^{n/2} \frac{\partial C(0, V, n)}{\partial W} W, \qquad (10)$$

but

$$W = 2Q_L\Delta f/f_0$$

hence

$$E_{c1} \cong K_n(2Q_L)^{n+1} \left[\frac{P_{osc}A_1A_2}{Q_1Q_2}\right]^{n/2} \frac{\partial C(0, V, n)}{\partial W} \frac{\Delta f}{f_0}. \qquad (11)$$

It has been pointed out in a preceding paragraph that $\delta C(0, V, n)/\delta W$ may be maximized by the proper choice of V for the detector law in effect. In addition, the expression for E_{c1} may also be maximized with the proper choice of Q relationships. As was pointed out in general section, the unloaded Q is usually determined by the other physical factors in the system; so it remains to discover if there is a proper choice of the ratios of the window or loaded Q such that the ratio of E_{c1} to Δf is a maximum. Differentiating the expression

$$\frac{Q_L^{n+1}}{(Q_1Q_2)^{n/2}} \qquad (12)$$

partially with respect to Q_1 and Q_2, equating to zero, and solving, it is seen that for maximum E_{c1} as a function of Q_1 and Q_2,

$$\text{or} \quad \left.\begin{aligned} Q_1 = Q_2 &= 2Q_0 \\ Q_0 &= 2Q_L \end{aligned}\right\} \quad \text{for} \quad n = 1, \qquad (13)$$

and

$$\text{or} \quad \left.\begin{aligned} Q_1 = Q_2 &= Q_0 \\ Q_0 &\equiv 3Q_L \end{aligned}\right\} \quad \text{for} \quad n = 2. \qquad (14)$$

While it is very interesting that such a maximum occurs, it frequently happens in the design of a system that is not desirable to take advantage of it for reasons to be considered.

Discussion of Oscillator Stability and Frequency Pulling

In coupling resonant cavities to microwave oscillators a certain amount of care must be exercised to prevent two undesirable effects:

1. If the cavity is coupled too strongly to the oscillator, the system may exhibit frequency discontinuities or jumps and refuse to oscillate at the cavity resonance frequency.
2. If the system is the type that uses the cavity-resonance frequency sensing, the varying susceptance presented to the oscillator through the coupling may cause a slight frequency and/or amplitude modulation of the oscillator at the sensing rate.

An excellent discussion of the first point appears in the literature.[1,9] Ford and Korman derived a formula which relates the frequency pulling of an oscillator to the change in load susceptance

$$\delta a = \frac{(\text{susceptance relative to matched admittance})}{2Q_{osc}}. \qquad (15)$$

The load susceptance must, of course, be referred to the oscillator terminals. In the case of a microwave oscillator, the terminals may be defined as the point in the transmission system where the frequency of the oscillator is independent of the load conductance. The susceptance at the cavity terminals may be calculated from the formula (see Appendix I):

$$Y_{in} = \frac{Q_1}{Q_2}\cdot Y_{out} + \frac{Q_1}{Q_0} + j\,\frac{2Q_1\Delta f}{f_0}. \qquad (16)$$

It is observed that there is a pure susceptance change with frequency (either resonance or applied), but with a reflection through a transmission system, this susceptance variation with frequency may manifest itself as conductance variation as well as a susceptance variation at the oscillator terminals. As may be seen with the

[9] J. R. Ford and N. I. Korman, "Stability and frequency pulling of loaded unstabilized oscillators," Proc. I.R.E., vol. 34, pp. 794–799; October, 1946.

and of a Smith chart,[10] the proper length of transmission line may favor either conductance variation or a susceptance variation. If the system under consideration uses a frequency varied resonator, the reflected susceptance variation will cause a frequency modulation of the oscillator, but the reflected conductance variation will cause an amplitude modulation of the oscillator at the sensing frequency and possibly at its harmonics. An engineering decision will have to be made as to the amount of these modulations that can be tolerated if this system is to be satisfactory. In the event that the coupling to the cavity must be reduced to bring these effects to a tolerable minimum, it may be necessary to sacrifice loop gain by the insertion of attenuation or other means to achieve the desired coupling. This point will be considered again in the discussion of the loop-gain equation.

In a similar manner to the pulling of the oscillator frequency by a susceptance presented to its terminals, the center frequency of the cavity may be pulled by a susceptance connected to its terminals. (This is one method of varying the resonance frequency.) Since, however, the cavity is to be used as the stable element of the control system, its frequency pulling must be considered, for many of the crystal detectors for microwaves have very broad tolerance on input admittance. Equation (19) may be used to calculate the pulling of the cavity by a susceptance of the detector if Q_{osc} is replaced by Q_2.

Derivation of Loop-Gain Limitations

A loop gain may be assigned to the complete loop as would be done in the case of an ordinary feedback amplifier. This is possible since E_{ct} is linearly related to the oscillator frequency deviation (11). All that is necessary to determine the loop gain is to multiply all of the individual gains together, including the various constants that relate the controlling sensitivity of the oscillator. (This last quantity may be influenced by a frequency sensitive load upon the oscillator.[1,9]

Hence, from Fig. 1 and (11) it follows that

$$G = \frac{G_f}{f_0} G_2 K_n (2Q_l)^{n-1} \left[\frac{P_{osc} A_1 A_2}{Q_1 Q_2} \right]^{n/2} \frac{\partial C(0, V, n)}{\partial W}. \qquad (17)$$

The following equation results for the gain optimized with respect to the cavity Q's and the amount of frequency modulation.

$$G = (0.23) \frac{G_f}{f_0} G_2 K_n (A_1 P_{osc})^{n/2} Q_0. \qquad (18)$$

The ratio of the frequency stabilities before and after the loop is closed can now be determined. The deviation in the oscillator frequency from the resonance frequency of the cavity after closing the loop is equal to the deviation before closing the loop divided by the quantity $(1+G)$.

In general, it may be stated that the relation between a small output admittance variation and input admittance variation is as follows for a matched coupling system

$$\left|\frac{\Delta Y_{in}}{Y_{in}}\right| = \left|\frac{\Delta Y_{out}}{Y_{out}}\right| \frac{1}{(\text{Power attenuation})} \qquad (19)$$

where Y_{in} and Y_{out} are the input and output admittances respectively. This expression applies to the case where the cavity is decoupled from the oscillator output by means of a directional coupler or a "magic tee" as well as by a conventional attenuator. For proof of this expression, see Appendix V. It is seen, then, that the input admittance variation varies inversely with the power attenuation. In deciding the admittance variation as a function of frequency due to the cavity and its effect upon the oscillator, it is observed from (19) that the admittance at the input terminals of the cavity will be very large at resonance and infinite at frequencies far from resonance. This is due to the fact that Y_2 is nearly unity and the ratio of the input Q to the unloaded Q will be chosen to be greater than unity. Hence, when used, the cavity will be placed an odd multiple of quarter-wavelengths from the oscillator terminals (if used in a series system, it will be placed an integral number of half-wavelengths from the oscillator terminals). Therefore, as seen at the oscillator terminals, a small variation of admittance will be observed as the frequency is varied through the cavity resonance frequency. The magnitude of this variation is seen to vary inversely with the input termination Q, since the input admittance to a quarter-wave section of transmission line is the reciprocal of the output admittance. (In the series case, impedance is the reciprocal of admittance.) Therefore, from the point of view of decreasing the oscillator pulling, is it better to increase the attenuation due to the coupling between the cavity or to increase the input termination Q of the cavity? The resonance frequency of the cavity is also sensitive to admittance variation presented to its terminals and ordinary mismatches that occur in an operating system may be responsible for some frequency variation; hence, from that point of view the termination Q of the cavity should be as high as possible. In looking at (17), it is noticed that the loop gain is influenced in one term by the ratio of attenuation to Q_1, hence, this term is independent of the decoupling for a constant amount of pulling on the oscillator. However, since the gain is proportional to a power of Q_l and Q_l varies in the same direction as Q_1 there is benefit to be reaped by increasing Q_1 over that of increasing the attenuation. These same arguments will apply to the decoupling of the cavity from the admittance variations of the crystal detector which might occur due to temperature variations or other effects. It follows, therefore, that for the reasons considered, it is preferable to decouple the cavity by the use of large input and output Q's with no additional attenuators.

[10] P. H. Smith, "A transmission line calculator," *Electronics*, vol. 12, pp. 29-33; January, 1939.

Derivation of the Second-Harmonic Output of the Detector

It may be important for design consideration to know the magnitude of the second harmonic of the sensing frequency contained in the detector output. In some cases, this component may have to be reduced to the proper amplitude by suitable filtering if the amplifier and associated circuits are to operate properly. In operation, the magnitude of the second-harmonic component may be many times that of the fundamental component resulting in the saturation of the amplifier. By Fourier's theorem,

$$E_{c2} = \frac{1}{\pi}\int_{-\pi}^{\pi} E_c(t) \cos 2\theta d\theta \qquad (20)$$

$$= K_n(2Q_L)^n \left[\frac{P_{osc} A_1 A_2}{Q_1 Q_L}\right]^{n/2} D(W, V, n) \qquad (21)$$

where

$$D(W, V, n) = \frac{1}{\pi}\int_{-\pi}^{\pi} \frac{\cos 2\theta d\theta}{[1 + (W + V \sin\theta)^2]^{n/2}}. \qquad (22)$$

$D(0, V, n)$ is evaluated in Appendix IV. Fig. 7 shows it as a function of V for n equal to 1 and 2. The values of

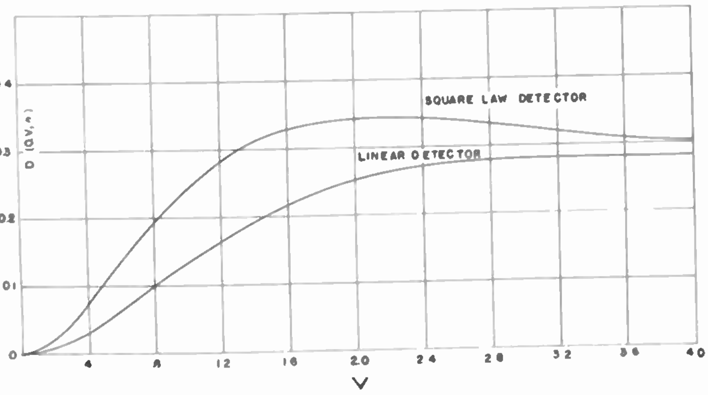

Fig. 7—Quantity of second harmonic of the sensing frequency in the detected output.

$D(W, V, n)$ are not important as a function of W since the operating point is near W equal zero. The value of the second-harmonic component is maximum at W equal zero.

Appendix I

If it is assumed that the cavity has only a single mode within the desired range of operation and has a bandwidth (to the 3-db points) which is small compared to its center frequency, then a number of simplifying assumptions can be made to expedite the analysis of its behavior. For the purposes of this paper, the cavity may be described sufficiently by the constants Q_1, Q_2, Q_0, and f_0. The restrictions and assumptions of the lumped-circuit model which will be assumed are described.[9] This paper is generally concerned with cavities that have only one window or loop, but the arguments are applicable to transmission cavities as well. Fig. 8 is a schematic of the model cavity. Use is made of a similar model in describing the behavior of cavities used as wavemeters;[11] the expressions used have a slightly different notation. The following are the definitions of the Q's:

$$Q_0 = 2\pi f_0 \frac{\text{(Maximum energy stored in the cavity)}}{\text{(Power dissipated in the walls of the cavity)}}; \qquad (23)$$

$$Q_1 = 2\pi f_0 \frac{\text{(Maximim energy stored in the cavity)}}{\text{(Power transmitted back through the input window)}}; \qquad (24)$$

$$Q_2 = 2\pi f_0 \frac{\text{(Maximum energy stored in the cavity)}}{\text{(Power transmitted through output window)}}; \qquad (25)$$

$$Q_L = 2\pi f_0 \frac{\text{(Maximum energy stored in the cavity)}}{\text{(Total power lost to the cavity)}}; \qquad (26)$$

$$\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_1} + \frac{1}{Q_2}. \qquad (27)$$

Since the actual values of the admittance of the load and the source to which the cavity is connected are not known, the effective shunt resistance cannot be known. However, for the purposes of this paper, it is not necessary to know this; it is only necessary to know the admittance relative to the value for which the proper Q was determined.

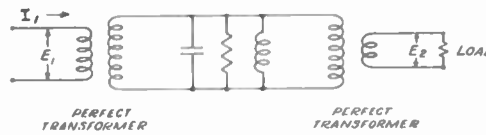

Fig. 8—Equivalent circuit of a cavity.

Since the cavity will, in general, be rather loosely coupled to the oscillator, the most useful way of describing the power transmission will be in terms of the ratio of incident power to the output power. Taking account of the reflected power at the input terminals of the cavity, by conventional circuit analysis[12] there results[13]

[11] See chap. 5 of footnote reference 2.

[12] E. A. Guillemin, "Communications Networks," John Wiley and Sons, New York, N. Y., 1931 and 1935.

[13] These expressions may be obtained directly from page 65 and page 291 of footnote reference 10 with the corresponding change in notation.

$$Y_1 = \frac{Q_1}{Q_2} Y_2 + \frac{Q_1}{Q_0} + j2Q_1 \frac{\Delta f}{f_0} \tag{28}$$

and

$$T = \frac{(2Q_L)^2}{Q_1 Q_L} \frac{1}{1 + \dfrac{(2Q_L \Delta f)^2}{f_0^2}}. \tag{29}$$

Appendix II

For n equal to unity, the following integral is to be evaluated

$$C(W, V, 1) = \frac{1}{\pi} \int_{-\pi}^{\pi} \frac{\sin \theta d\theta}{\sqrt{1 + (W + V \sin \theta)^2}}. \tag{30}$$

From a practical point of view, it is most expedient to evaluate this by a series expansion about V equal to zero. Expanding the denominator by a Taylor series, then, it is seen that

$$C(W, V, 1) = \frac{1}{\pi} \int_{-\pi}^{\pi} \sum_{n=0}^{\infty} \frac{d^m}{dW^m} \frac{V^m \sin^{m+1} \theta d\theta}{n! \sqrt{1 + W^2}}. \tag{31}$$

Exchanging the order of integration and summation and integrating,[14]

$$C(W, V, 1) = 2 \sum_{m=0}^{\infty} \frac{1 \cdot 3 \cdots (2m+1)}{2 \cdot 4 \cdots (2m+2)} \frac{d^{2m+1}}{dW^{2m+1}} \frac{1}{\sqrt{1 + W^2}} \frac{V^{2m+1}}{(2m+1)!}. \tag{32}$$

Performing the differentiating and factoring gives the required series

$$C(W, V, 1) = \frac{-WV}{(1 + W^2)3/2} \left[1 + \frac{3(2W^2 - 3)}{8(1 + W^2)^2} V^2 + \cdots + \frac{5(8W^4 - 40W^2 + 15)}{64(1 + W^2)^4} V^4 + \frac{35(16W^6 - 168W^4 + 210W^2 - 35)}{81,920(1 + W^2)^6} V^6 \cdots \right]. \tag{33}$$

This series is convergent for V^2 less than $1 + W^2$ since the radius of the circle of convergence of the series for $1/1 + (W+z)^2$ in the Z-plane is $\sqrt{1 + W^2}$, when expanded about $Z = 0$.

For n equal to 2, the following integral must be evaluated

$$C(W, V, 2) = \frac{1}{\pi} \int_{-\pi}^{\pi} \frac{\sin \theta d\theta}{1 + (W + V \sin \theta)^2}. \tag{34}$$

By expanding the integrand in partial fraction, it is seen that[15]

$$C(W, V, 2) = \frac{1}{2V\pi j} \int_{-\pi}^{\pi} \left[\frac{W + j}{W + j + V \sin \theta} - \frac{W - j}{W - j + V \sin \theta} \right] d\theta. \tag{35}$$

$$C(W, V, 2) = \frac{1}{Vj} \left[\frac{1}{\sqrt{1 - \left(\dfrac{V}{W + j} \right)^2}} - \frac{1}{\sqrt{1 - \left(\dfrac{V}{W - j} \right)^2}} \right]. \tag{36}$$

By a few algebraic simplifications, it is found that

$$C(W, V, 2) = - \frac{2\sqrt{W^2 + 1}}{V \sqrt[4]{(W^2 + 1)^2 - 2V^2(W^2 - 1) + V^4}} \cdot \sin \tfrac{1}{2} \arctan \frac{2V^2 W}{(W^2 + 1)^2 - V^2(W^2 - 1)}. \tag{37}$$

Appendix III

The functions representing the slopes of the effective discriminator curves at the origin are to be evaluated. The slope of the curves for the linear detector may be evaluated rather easily in terms of complete elliptic integrals from the original integral.

$$\frac{\partial C(0, V, 1)}{\partial W} = - \frac{V}{\pi} \int_{-\pi}^{\pi} \frac{\sin^2 \theta d\theta}{(1 + V^2 \sin^2 \theta)^{3/2}}. \tag{38}$$

Substituting $1 - \cos^2 \theta$ for $\sin^2 \theta$ and $\theta = \pi/2 - \phi$, it follows that

$$\frac{\partial C(0, V, 1)}{\partial W} = \frac{-4V}{\pi(1 + V^2)^{3/2}} \int_0^{\pi/2} \frac{(1 - \sin^2 \phi) d\phi}{\left[1 - \dfrac{V^2}{1 + V^2} \sin^2 \phi \right]^{3/2}}. \tag{39}$$

Consulting de Haan,[16] and making a few algebraic simplifications, it is seen that

[14] Mathematical Tables from "Handbook of Chemistry and Physics," Chemical Rubber Publishing Co., Cleveland, Ohio, 5th edition, 1936; no. 337, p. 271.

[15] See no. 349, p. 272, of footnote reference 14.

[16] D. Bierens de Haan, "Nouvelles Tables D'integrales Definies," G. E. Strechert and Co., New York, N. Y.; 1939; p. 91.

$$\frac{\partial C(0, V, 1)}{\partial W} = \frac{-4}{\pi V\sqrt{1+V^2}}\left[K\left(\frac{V}{\sqrt{1+V^2}}\right) - E\left(\frac{V}{\sqrt{1+V^2}}\right)\right] \quad (40)$$

where K and E are elliptic integrals.[17]

The slopes of the effective discriminator curves for the square-law detector may be found most easily by differentiating the function directly, resulting in

$$\frac{\partial C(0, V, 2)}{\partial W} = \frac{-2V}{(1+V^2)^{3/2}}. \quad (41)$$

Appendix IV

The integral representing the amount of second harmonic of the sensing frequency may be worked out at W equal to 0. For the linear detector,

$$D(0, V, 1) = \frac{1}{\pi}\int_{-\pi}^{\pi}\frac{\cos 2\theta d\theta}{\sqrt{1+V^2\sin^2\theta}}. \quad (42)$$

Substituting $1-\cos^2\theta$ for $\sin^2\theta$, $\pi/2-\phi$ for θ, $1-2\sin^2\phi$ for $\cos 2\theta$, and shifting the integration intervals, it is seen that

$$D(0, V, 1) = \frac{-4}{\pi\sqrt{1+V^2}}\int_0^{\pi/2}\frac{(1-2\sin^2\phi)d\phi}{\sqrt{1-\dfrac{V^2}{1+V^2}\sin^2\phi}}. \quad (43)$$

It is found that[18]

$$D(0, V, 1) = \frac{-4}{\pi\sqrt{1+V^2}}\left[K\left(\frac{V}{\sqrt{1+V^2}}\right) - 2D\left(\frac{V}{\sqrt{1+V^2}}\right)\right]. \quad (44)$$

For the square-law detector,

$$D(0, V, 2) = \frac{1}{\pi}\int_{-\pi}^{\pi}\frac{\cos 2\theta d\theta}{1+V^2\sin^2\theta}. \quad (45)$$

Substituting $\frac{1}{2}(1-\cos 2\theta)$ for $\sin^2\theta$, $\phi=\theta/2$, and shifting the interval of integration, it is seen that

$$D(0, V, 2) = \frac{2}{\pi(2+V^2)}\int_0^{2\pi}\frac{\cos\phi d\phi}{1-\dfrac{V^2}{2+V^2}\cos\phi}d\phi. \quad (46)$$

Making some algebraic simplifications,[19] it is found that

$$D(0, V, 2) = \frac{4}{V^2}\left[\frac{2+V^2}{2\sqrt{1+V^2}} - 1\right]. \quad (47)$$

[17] E. Jahnke and F. Emde, "Tables of Functions with Formulae and Curves," Dover Publications, New York, N. Y., 1943.
[18] See p. 73 of footnote reference 17.
[19] See p. 272, no. 349, of footnote reference 14.

Appendix V

The expression for the relation between the variation in input admittance and the variation in output admittance for a matched four-terminal network may be derived easily using matrix methods.[12] Consider the general four-terminal network characterized by the four constants a, b, c, and d. Not all are independent, for a bilateral system with a consistent set of units $ad-bc$ must equal unity. Consider the network operating into and admittance Y_{out} with an associated input admittance Y_{in}. Let E_1 and I_1 be the input voltage and current to the network, and let E_2 and I_2 be the output voltage and current. It follows, then, that

$$\begin{vmatrix} E_1 \\ I_1 \end{vmatrix} = \begin{vmatrix} a & b \\ c & d \end{vmatrix} \times \begin{vmatrix} 1 & 0 \\ Y_{out} & 1 \end{vmatrix}\begin{vmatrix} E_2 \\ 0 \end{vmatrix}.$$

Hence,

$$E_1 = (a + bY_2)E_2$$

and

$$I_1 = (c + dY_2)E_2.$$

By dividing I_1 by E_1, it follows that

$$Y_{in} = \frac{c + dY_{out}}{A + BY_{out}}.$$

By differentiating with respect to Y_{out}, it follows that

$$\Delta Y_{in} \cong \frac{\Delta Y_{out}}{(a + bY_{out})^2}.$$

The power attenutaion for the network can be derived for the relationship between E_1 and E_2 if Y_{in} and Y_{out} be considered as pure conductances. If they are not, then the power input and output expression must contain only the real parts of Y_{in} and Y_{out}. To include this refinement generalizes the final expression, but it does not increase its usefulness. Hence,

$$(\text{Power attenuation}) = \left|\frac{E_2^2 \quad Y_{out}}{E_1^2 \quad Y_{in}}\right| = |a + bY_{out}|^2\frac{Y_{out}}{Y_{in}}.$$

It then follows that

$$\left|\frac{\Delta Y_{in}}{Y_{in}}\right| = \left|\frac{\Delta Y_{out}}{Y_{out}}\right|\frac{1}{(\text{Power attenuation})}.$$

Even though the admittances Y_{in} and Y_{out} must be real, the variations do not have to be real and, in general, they will not be real. If it were possible to find the values of a, and b and Y_{out} for the particular network, it would be possible to calculate the relationship between the real and imaginary components of the admittance variation. However, if only the power attenuation is known, only the relation between the absolute magnitudes of $\Delta Y_{in}/Y_{in}$ and $\Delta Y_{out}/Y_{out}$ can be found.

New Design for a Secondary-Emission Trigger Tube*

C. F. MILLER†, SENIOR MEMBER, IRE, AND W. S. McLEAN‡

Summary—**Aspects of the design of a nongaseous miniature trigger tube are given. A triode input section produces a primary electron beam which impinges on a dynode to produce secondary electrons. These are collected by two different output elements which may be used separately or as a unit. A surface having long life and stability is described. Suggested applications include its use as a relaxation oscillator, multivibrator, pulse inverter, triangular-wave generator, and dynatron.**

Introduction

An earlier paper[1] described the application of secondary-emission phenomena to the problem of obtaining trigger action. A design of tube was shown which served well for the basic principles involved, but which left something to be desired in terms of high-frequency operation and miniaturization. The tube described in this present paper satisfies these last two requirements. It carries the designation NU-1032-J.

Significance of the "First-Zero" Voltage

It has become common practice, in working with secondary-emission surfaces or dynodes, to designate the quality by the value of the "first-zero" voltage taken from the dynatron characteristic. Such a characteristic is shown as Fig. 1, curve A. The "first-zero" voltage is the lower of two voltages at which the current to the dynode is zero. For the construction used, at this voltage there will be a large difference of potential between the collector grid and the dynode, so that all the secondary electrons will be drawn away from it. Thus, this "first-zero" voltage is that voltage at which the ratio of secondary to primary electrons (δ) is equal to unity. It is also the voltage for the primary electrons at which their velocity is just great enough to produce one secondary electron for each primary electron.

Tube Design

The main problem which confronts the development of any type of secondary-emission tube is one of stability during life. Maintaining a low and consistent value of "first-zero" voltage is the prime requisite for a good trigger tube. In 1938, it was found[2] that contamination given off by the cathode caused poisoning of the dynode surface and consequent instability in secondary-emission tubes. Theory and experiments showed that this contamination traveled in straight lines. With this in mind, the first stable secondary-emission amplifier was developed by sheilding the dynode from the cathode and causing the primary electrons to travel in curved paths.

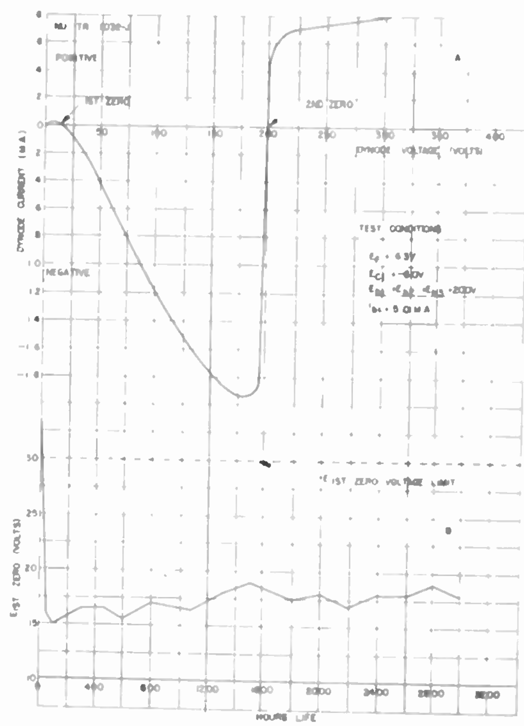

Fig. 1—Dynode characteristic curve A and first-zero value during life, curve B.

This same principle has been applied in the design of the present tube. The final tube geometry was decided on after an extensive rubber-model investigation. The principal elements and their relationship to each other are shown in Fig. 2. These include the cathode, control grid, and slotted anode which make up the triode section; also the shield, anode No. 2 or collector grid, dynode, and anode No. 3. The slot in the plate forms a beam of primary electrons. The influence of the shield, which is at cathode potential, causes the primary electrons to travel in curved paths, passing through anode No. 2 and finally striking the dynode. Particular care was paid to obtaining an even density and spread of electrons over the surface of the dynode in order to prevent overloading any one portion of the surface. Electrons passing through the middle of the slot strike

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† National Union Radio Corp., Orange, N. J.
‡ Western Electric Co., New York, N. Y.

[1] A. M. Skellett, "Use of secondary electron emission to obtain trigger or relay action," *Jour. Appl. Phys.*, vol. 13, pp. 519–524; August, 1942.

[2] J. L. H. Jonker and A. J. W. M. v. Overbeck, "The applications of secondary emission in amplifying valves," *Wireless Eng.*, vol. 15, pp. 150–157; March, 1938.

the middle of the dynode, and those through the outer limits spread evenly to each side of the center.

The rubber-model trajectories were checked by means of a tube containing a fluorescent dynode which showed the same spread of electrons on the dynode as did the rubber model. One of the innovations on this type of tube is the use of anode No. 3. Previous designs depended entirely on the collector grid for collecting the secondaries, and, consequently, many secondaries in passing through the collector grid struck the first anode. Anode No. 3 prevents this loss in that it shields the first anode almost completely from secondaries passing through the collector grid and eliminates the conditions which were conducive to oscillation by the secondaries about the collector grid.

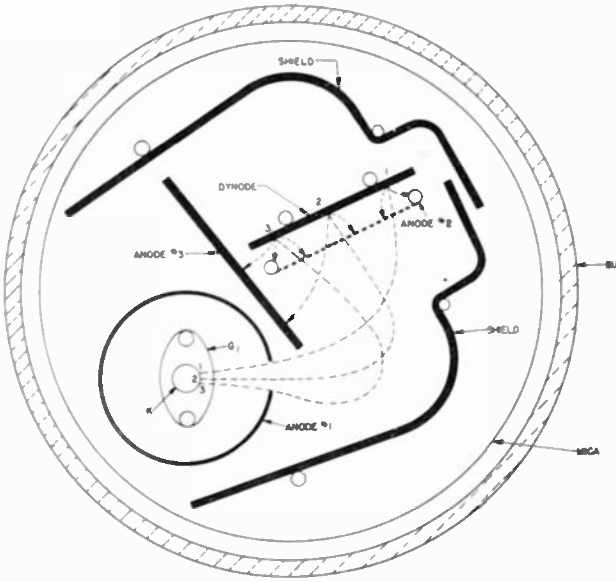

Fig. 2—Cross section showing the elements of the trigger tube.

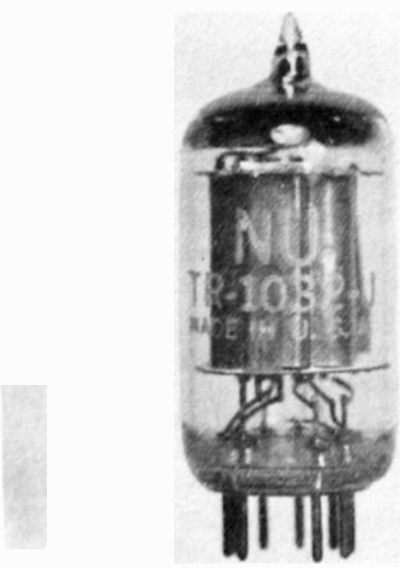

Fig. 3—Photograph of the miniature trigger tube.

The dynode surface combines a base metal of oxygen-free copper or silver with a thin composite layer of barium and magnesium oxides. The barium and magnesium are evaporated onto the copper under vacuum conditions. The base metal is highly polished on the active side before the layers of the other metals are applied. Special care is taken not to contaminate the copper in any way once it is polished. The thin layers of barium and magnesium are oxidized by exposing the structure to oxygen. It is believed that some of the pure metal diffuses through this oxide layer to the surface during the aging process.

A photograph of the tube is shown in Fig. 3.

Life Tests

The arbitrary limit for an acceptable value of secondary-emission ratio for a trigger tube is a "first zero" of +30 volts or less. Curves A and B in Fig. 1 show what may be expected of the tube with respect to life and consistency; curve A is the dynode characteristic taken after 3,000 hours life. Curve B shows that the "first-zero" value has been good throughout life. The initial drop (from approximately 35 to 16 volts) indicates that the surface was "aged-in" during this time, and after a few hours the "first zero" reached an average value varying between 15 and 19 volts. The tests indicate that this particular surface will maintain consistent values of δ under 0.5 watt per square centimeter dissipation.

These life tests have shown that it is possible to make dynodes of the kind used herein that last for 5,000 hours or better, maintaining constant characteristics.

Applications

The basic trigger circuits for the use of this tube are described in the earlier paper[1] referred to above.

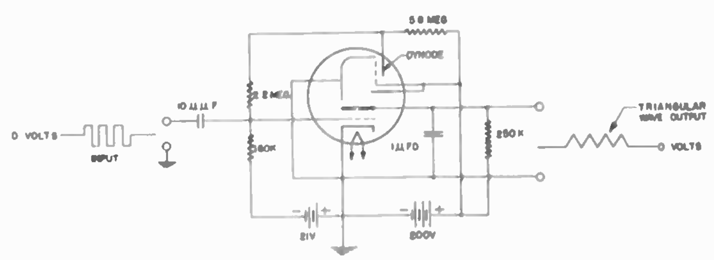

Fig. 4—Triangular-wave generator, $f = 5{,}000$ cps.

Fig. 4 shows the tube used as a triangular-wave generator. The action of the circuit is as follows: The square-wave input is differentiated into triggering

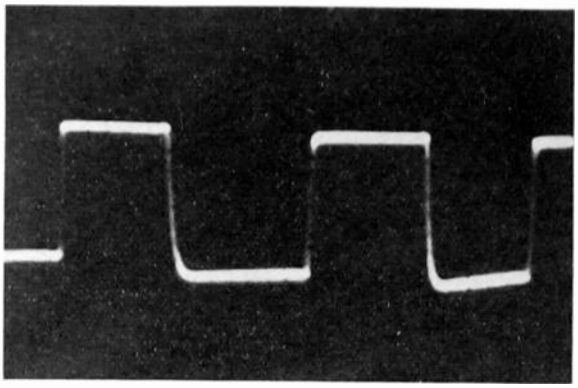

Fig. 5—Square-wave output for 100 kc.

pulses by the input RC circuit. These pulses trigger the tube "on" and "off," as in the circuit described earlier. When the tube is triggered "off," the 0.1-μf capacitor is charged exponentially through the plate resistor (250,000 ohms). When the tube is triggered "on," the output capacitor discharges exponentially through the tube.

Fig. 5 shows the output wave form when the tube is triggered at 100 kc. The leading edge has a rise time of approximately 0.2 microsecond. The full capabilities of this tube have not yet been determined; however, it is believed that it can be made to trigger at frequencies as high as 1 Mc since no gas-deionization time is involved.

Electronic Techniques Applied to Analogue Methods of Computation*

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Summary—**This paper describes in detail the electronic devices and principles that have been developed for the California Institute of Technology (CIT) electric analogue computer. This is a general-purpose, large-scale computer applicable to a wide range of linear and nonlinear ordinary algebraic or differential equations and linear and nonlinear partial differential equations.**

In addition to the basic principles of the computer, a detailed discussion is given of those elements considered to be of particular interest. These include the devices for generating the arbitrary functions of the independent variable (the excitation functions), the amplifiers for producing active linear elements such as negative impedances and for representing the nonsymmetrical terms of the matrix specifying the differential equations, the multipliers for producing arbitrary functions of the dependent variables (nonlinear elements).

Performance data on these devices are presented, together with analogies and solutions of representative types of problems.

Introduction

THE ELECTRIC analogue computer, which has been developed by the California Institute of Technology provides a device having as wide a field of application as is considered practical with the electric analogue principle. It is thought that the greatest field of usefulness of such a computer lies in the accuracy range that is limited to the order of 1 per cent, which covers the vast majority of problems in the general field of engineering analysis. With this accuracy-limitation, great simplification in the computer and its operation results from the use wherever possible of R, L, C, and transformer circuits in simulating the linear terms of both algebraic and differential equations.[1,2] This is in contrast to computers which employ electronic amplifiers for all integration or differentiation. They provide higher accuracy, but are restricted in their field of application as compared to that practicable with a computer of the type discussed here.

Electric analogue computers are limited to one independent variable that can be continuously represented. However, by the use of finite-difference methods, such as must be employed for digital computers, more independent variables can be handled. The CIT computer is designed to handle partial differential equations with up to three independent variables. Two basic circuit techniques are employed. For the first of these, one independent variable is continuously represented as time on the computer.[1,2] In the other, all independent variables are represented by finite-difference methods, in which case the analogous circuit becomes a steady-state ac or dc mesh.[3] Both techniques are applicable to both ordinary and partial differential equations. It is the application of the first technique that requires most of the electronic equipment, such as devices to generate suitable functions of the independent and dependent variables, multipliers for multiplying any two variables together, and amplifiers to provide the proper power-output levels from these devices and to produce negative impedance terms and the unilateral or unsymmetrical terms of a matrix or servosystem.

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† California Institute of Technology, Pasadena, Calif.

[1] G. D. McCann, "The mechanical transients analyzer," *Proc. NEC*, vol. 2, p. 372, 1946.

[2] E. L. Harder and G. D. McCann, "A large scale general-purpose electric analog computer," AIEE Technical Paper 48-112.

[3] G. D. McCann and C. H. Wilts, "Application of electric analog computers," *Jour. Appl. Mech.*, 1949.

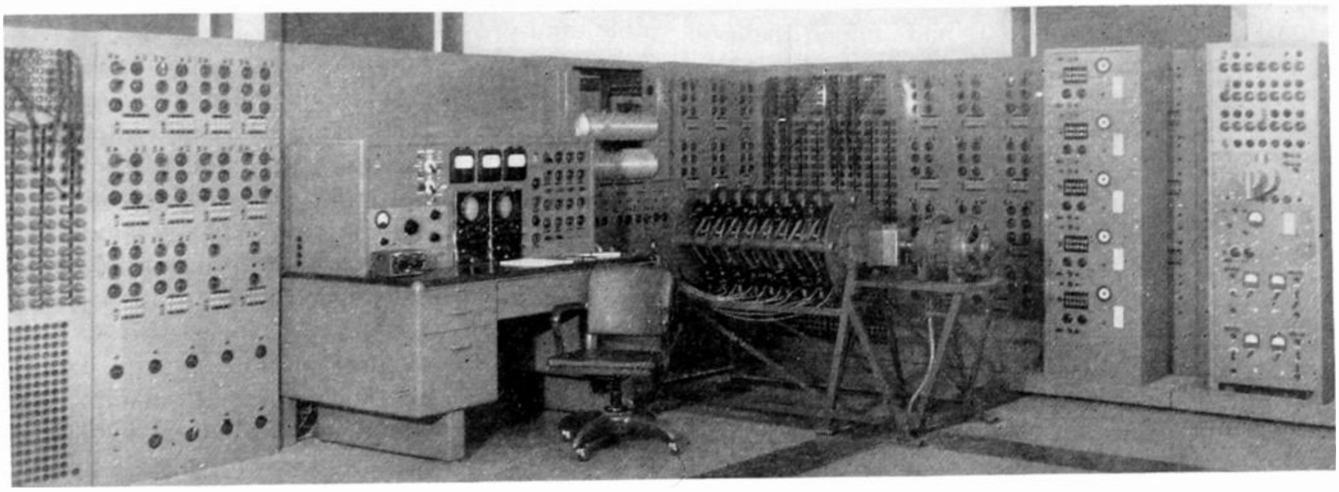


Fig. 1—A general view of the California Institute of Technology computer.

It is these recently developed basic electronic elements of the computer that are to be described in this paper. The general arrangement of the computer, a photograph of which is shown in Fig. 1, has been described in a previous paper.[2]

Examples of Computer Techniques

The required specifications for the analogue elements is best considered by first illustrating their application to specific problems and describing the analogue techniques employed. The basic analogue elements employed in the computer are listed in Fig. 2 in terms of their symbolization as employed for circuit layouts. Typical analogous circuits for ordinary differential equations are given in Figs. 3 and 4.

As illustrated in the examples, R, L, and C circuit elements are usually employed for all linear "self-impedance" terms. They are also employed for bilateral mutual terms that can be formed with junction points between the meshes simulating the simultaneous equations and for transfer functions. Special transformers[2] are used wherever possible for bilateral mutuals requiring circuit isolation or, in conjunction with amplifiers, to provide a four-terminal unilateral element.

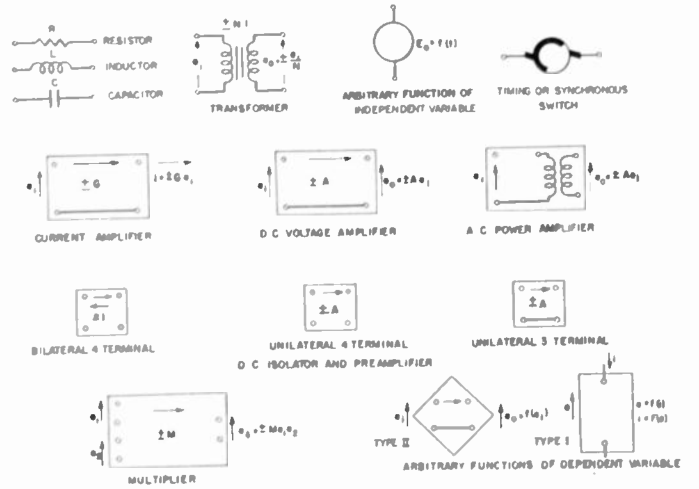


Fig. 2—Symbolization of the basic computer elements.

The use of such circuit elements fixes the impedance level and the rate of computation. To minimize stray coupling and stray pickup problems and enable the quick setup of a wide range of circuits, the circuit impedance base must be kept low and the power level reasonably high. Unintentional impedances, such as the output impedance of a voltage amplifier, must be kept below a few ohms if they are series elements, and above several hundred thousand ohms (and in extreme cases, 10 or 20 megohms) if they are shunt impedances. In order that the elements are sufficiently close to being ideal inductors, etc., the natural frequencies of the analogue circuits must lie within certain limits. A normal operating frequency range of from 10 cps to 1 or 2 kc was chosen.[2] However, because of this wide frequency range, transformers have certain limitations. It is difficult to keep the ratio of the series to magnetizing impedance of a transformer sufficiently low over such a wide operating frequency range. This, together with

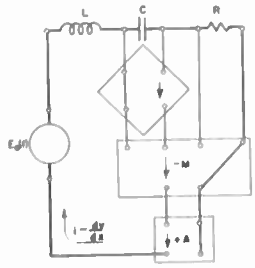


Fig. 3—Illustrative analogy for an ordinary differential equation:

$$F_0(x) = k_1 \frac{d^2y}{dx^2} + f_1(y) \frac{dy}{dx} + k_2 y.$$

Analogous circuit equation:

$$E_0(t) = L \frac{d^2q}{dt'^2} + \frac{AMR}{C} F_1(q) \cdot \frac{dq}{dt'} + \frac{q}{C} \left\{ R \frac{dq}{dt'} \text{ negligible} \right\}.$$

Numerical conversion formula:

$$y = a \frac{F_0}{E_0} q; \quad L = \frac{a}{n^2} k_1; \quad \frac{I}{C} = ak_2; \quad \frac{AMR}{C} \cdot F_1(q) = \frac{a}{n} f_1(q)$$

where
a = impedance-base change factor
n = independent-variable conversion factor.
(One unit of x in its system of units equals $1/n$ units of t' seconds.

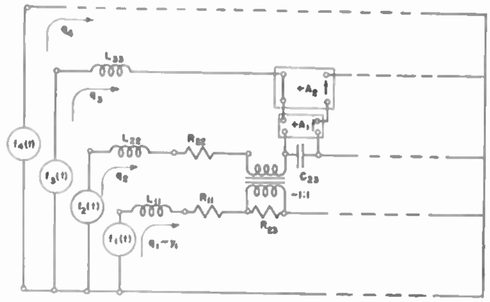


Fig. 4—Illustrative analogy for simultaneous equations:

$$f_1(t) = K_{11}\frac{d^2y_1}{dt^2} + k_{11}\frac{dy_1}{dt} + k_{12}\frac{dy_2}{dt} + \cdots$$

$$f_2(t) = K_{22}\frac{d^2y_2}{dt^2} + k_{22}\frac{dy_2}{dt} + k_{12}\frac{dy_1}{dt} + \cdots$$

$$f_3(t) = K_{33}\frac{d^2y_3}{dt^2} + k_{23}y_2 + \cdots$$

$$f_4(t) = \cdots.$$

the tendency of transformer-coupled feedback circuits to "motorboat," constitute the major difficulties sometimes encountered in the use of transformers. It is, therefore, occasionally necessary to use four-terminal "isolators" which operate properly from zero frequency (dc) to the upper limit of 1 or 2 kc (see Fig. 4). The general philosophy regarding accuracy has been to achieve 1 per cent accuracy in all elements. Thus, an inductor should exhibit a variation in effective inductance which is less than 1 per cent of its value throughout the range in operating level and operating frequency (in this case, 100 to 1,000 cps). In addition, its effective resistance should be less than 1 per cent of its reactance at all frequencies and levels (i.e., its Q should be greater than 100). Similarly, all of the electronic equipment must have less than 1 per cent amplitude distortion throughout the operating range. Furthermore, the frequency response of the equipment must not vary more than 0.1 db throughout the frequency range of the computer. It was considered satisfactory if the phase shift were less than 1° throughout the frequency range. However, since these devices are frequently used in cascade and in feedback circuits, the phase and amplitude characteristics must be controlled to eliminate oscillations at high and low frequencies. To prevent low-frequency "motorboating," all electronic equipment except the arbitrary function of the independent-variable (forcing-function) generators was designed for operation from zero frequency (dc) to the upper frequency limit of the computer. To simplify elimination of high-frequency oscillations, it was found desirable to extend the high-frequency limit of all equipment as far as practicable.

Power Amplifiers

Three types of amplifiers are shown in Fig. 2—ac power amplifiers, dc current generators, and dc voltage amplifiers. Although they are frequently used for other purposes, their most common uses are: (1) Ac power amplifier: to amplify the output of the forcing-function generators (arbitrary function of the independent variable) and provide a low source impedance. (2) Dc current generator: to produce a current as a "forcing function." (3) Dc voltage amplifier—to produce negative or unilateral impedances, and to amplify the output of generators of an arbitrary function of the dependent variable. Typical examples of the use of the dc amplifiers are given in Figs. 3 and 4. The dc amplifier input impedances must be kept high so that they will not distort the circuit across which they are to be connected. However, since they usually can be used with a small input signal, the circuits can be so set up that 500,000 ohms is a sufficiently high input impedance. For most problems, power levels corresponding to a maximum current of 30 to 50 ma and a voltage of 50 to 100 volts are adequate.

Two types of dc voltage amplifiers were found desirable. One of these has a negative or 180° phase-reversal characteristic, and is to be used particularly for servomechanism problems. It has a maximum gain of 50 and is usually used in groups of three or more. The other is a positive-gain amplifier. For this a gain of 100 is usually adequate when it is used to produce a negative or unilateral impedance, since the accuracy limitation of the computer is 1 per cent. However, in limited cases, larger gains are required. In order to provide for this additional gain, the amplifier can be used with a preamplifier, which has a gain variable from 0 to 10. It is considered adequate if the current amplifiers can deliver about 50 ma and 100 volts. Their effective source impedance should be a megohm or more.

As can be seen from the illustrative circuits (Figs. 3 and 4), it is necessary that all analogue elements be capable of operation with no connection between normal ground terminal and true system ground, or they must be provided with isolating elements at their terminals. As will be discussed later, it was found practicable to design the electronic equipment for operation with no connection to system ground, thereby affording considerable simplification.

Multipliers

The multipliers must be capable of multiplying together any two variable voltages having frequencies up ti 1 or 2 kc. Atypical examples of their use is given in Fig. 3. Recalling the over-all 1 per cent accuracy requirements of the computer, it is desirable, as a *minimum* requirement, that the multipliers be capable of forming products of functions such that the instantaneous product obtained for any input voltages deviates from the correct value by an amount less than 1 per cent of the maximum product obtainable.

Arbitrary Function Devices

Because of the wide range of possible functions, it is desirable to have several methods of generating functions of both time, the independent variable, and the dependent variables. For the "independent variable" or time functions, six square-wave voltage generators, consisting of storage-battery circuits, and six variable-frequency sine-wave generators are provided. These are described in footnote reference 2. In addition, however, it is necessary to be able to generate any time function that can be plotted or expressed mathematically. Many important nonlinear functions of the dependent variable, such as saturation curves, square-law characteristics, sudden discontinuities, etc., can be generated with relatively simple diode circuits. In addition, however, provision must be made for perfectly arbitrary functions.

Description of Electronic Components

Generator of Arbitrary Functions of the Independent Variable

Since the computer operates in the audio-frequency range, it is natural to consider the possible application of the various principles employed for sound recording, including the light-beam–photocell principle, magnetic tape, or wire recording and conventional disk recording and playback techniques. All of these methods were tried experimentally and found practical. As will be discussed later, the light-beam–photocell principle could be most easily and practically developed into a form suitable to the needs of the computer, and was adopted for one form of "forcing-function" generator. However, the use of any of the above principles leads to one practical difficulty. If the function of the independent variable is unknown and the only known data are some actual solutions obtained from experimental tests, it is desirable to have a generator in which the arbitrary function can be varied until, by trial and error, a correct solution is obtained.

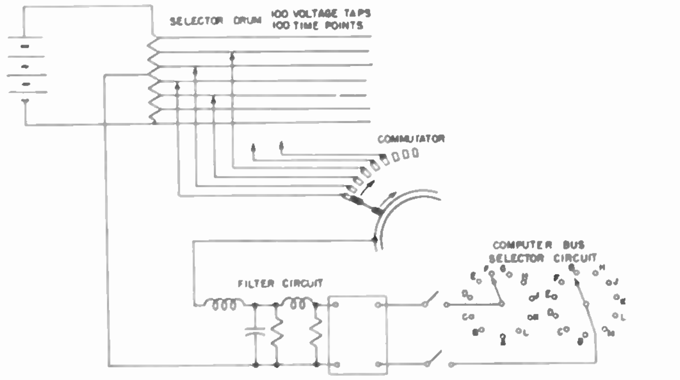


Fig. 5—Schematic diagram of the arbitrary forcing-function generator Type I; variable gain amplifier, 1-ohm output impedance, 50 volts at 1-ampere output.

Type I

The device illustrated in Fig. 5, although limited in the maximum frequency which it can represent, does provide for rapid changes in the forcing function. As shown in the schematic diagram (Fig. 5), 100 voltage taps are provided which may be selected by the adjustment of 100 switches, each of which connect to a segment of a commutator, thus providing 100 points in time. A filter circuit provides for a smooth transition between points. The 100 voltage taps are rods placed around the periphery of a circular drum, over which are 100 circular rings each connecting to a segment of a mechanical commutator. Adjustment of the function generator is accomplished by rotating the rings around the drum. Three of these devices are provided in the computer. They may be used separately, or two may be ganged in series to provide up to 200 points in time.

Type II

The generator employing the photoelectric principle is illustrated in Fig. 6. A photographic film is produced

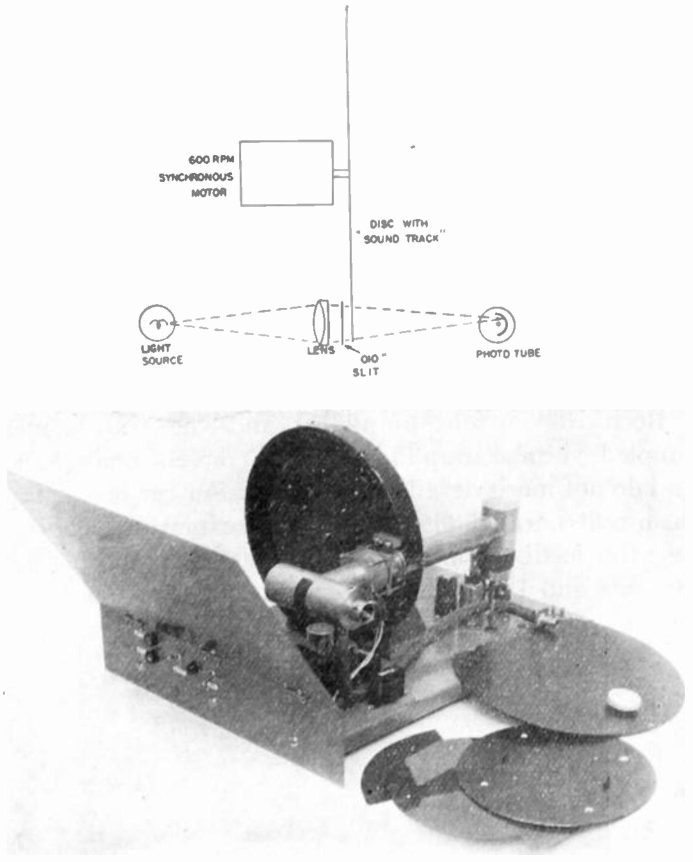


Fig. 6—(a) Schematic diagram of the generator of an arbitrary function of independent variable. (b) Photograph of a generator of an arbitrary function of independent variable.

with a circular "sound track" having a variable width proportional to the desired function. This film and a narrow slit are interposed between a light source and a phototube. When the film is rotated at synchronous speed (10 rps), a voltage is produced which is a periodic function of time, and which is, during each cycle, proportional to the desired function of the independent variable. The film diameter is 11 inches, with provision for a maximum track width of $\frac{3}{8}$ inch. A simple ac amplifier is used to increase the signal level to about 1 volt.

There are six principal sources of distortion in such an optical-photoelectric system. Two of these result from geometrical properties of the system. One arises if the width of the "sound track" is made exactly proportional to the desired function. In this case, the angle subtended by the track is not exactly proportional to the desired function, but is related by a simple trigonometric expression which departs appreciably from linearity if the angle becomes too large. Since the light transmitted is proportional to the angle, a lack of linearity results. The other distortion is a frequency-sensitive amplitude distortion. It arises because of the finite slit width and limits the detail possible because of the averaging effect over the width of the slit. These distortions were made less than 0.1 per cent by maintaining suitable ratios of critical dimensions.

Four other sources of distortion are due to imperfections in the apparatus: variable sensitivity of the phototube surface, imperfect slit, variation of source light intensity with direction, and variation in transmissibility of film. Distortion arising from variable sensitivity of the phototube surface is minimized by focusing the light from the source to one spot on the phototube by means of a simple achromatic lens. Distortion due to the other three causes is believed to be the major source of error. For a perfect film, the generator shown in Fig. 6 with a $\frac{5}{8}$-inch sound track has a maximum over-all distortion of 0.3 per cent.

Power Amplifiers

Both the current amplifiers and the transformer-coupled voltage amplifiers are of conventional design and do not merit detailed discussion. For the latter, four push-pull–parallel 6L6's with approximately 60 db of negative feedback met the special output requirement of 50 volts and 1 ampere with an internal impedance of 1 ohm. The amplification varies less than 0.1 db, and the phase shift is considerably less than 1° from 10 to 2,000 cps.

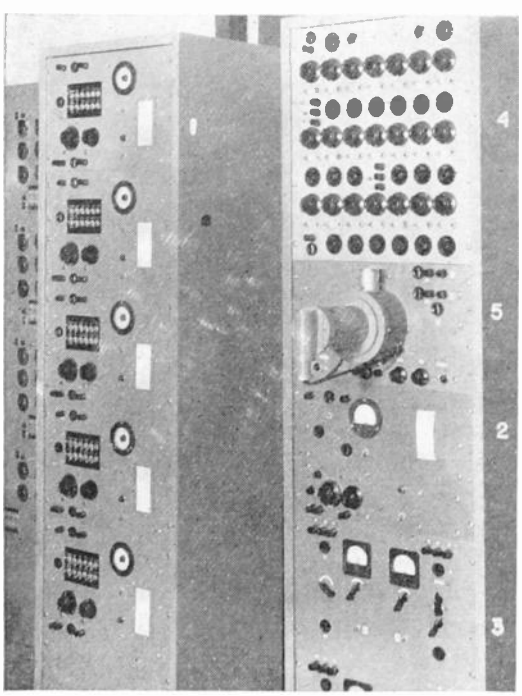
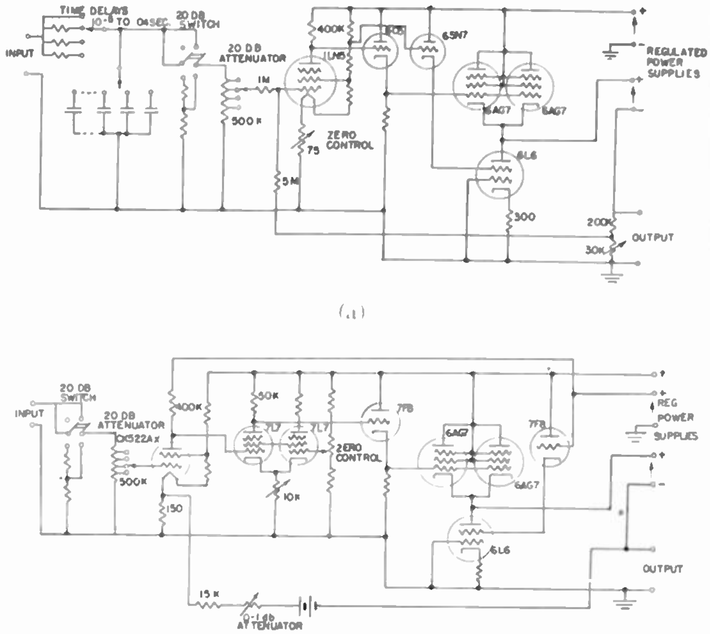


Fig. 7—Photograph of amplifiers and arbitrary-function generators. (1) Negative gain servo amplifiers. (2) Positive-gain dc amplifier. (3) Voltage and current limiters. (4) Dependent-variable function generator—Type I. (5) Dependent-variable function generator—Type II.



(a)

(b)

Fig. 8—Schematic diagrams of dc amplifiers. (a) Servo or negative-gain amplifier. (b) Positive-gain amplifier.

The dc voltage amplifiers presented the most difficult problems. It is of great importance to minimize the drift in output voltage. In typical servo problems, amplifiers may be cascaded with an over-all gain of many thousand, so that drifts of even a few millivolts at the input may be objectionable. After considerable experimental and theoretical work on the matter, satisfactory amplifiers were developed. Since the main source of drift lies in the power supplies and the first stage of an amplifier, it is necessary to use well-regulated power supplies and to choose carefully the tube for the first stage. Filament-type tubes with well-regulated filament current were found superior to the well-known Miller circuit. However, since the simplest source of regulated voltage is the high-voltage power supply, a low-current tube is desirable to avoid excessive power requirements. Some of the negative-gain servo amplifiers are shown in Fig. 7, and the circuit diagrams for both types of amplifiers are shown in Fig. 8. In practice, the 1LN5 pentode with a filament current of 0.040 ampere was found satisfactory for the negative-gain amplifier. A subminiature hearing-aid tube, the CK522AX, is used for the positive-gain amplifier, since its filament current is only 0.020 ampere, while its drift and microphonism characteristics are very satisfactory. This tube cannot be used for the negative-gain amplifier because of its inability to provide a sufficiently large output voltage in a one-stage amplifier.

After about a two-hour warm-up period, the drift as referred to the grid of the first tube settles down to a value that does not exceed 1 millivolt per hour. A typical drift record is shown in Fig. 9. The sharp pips

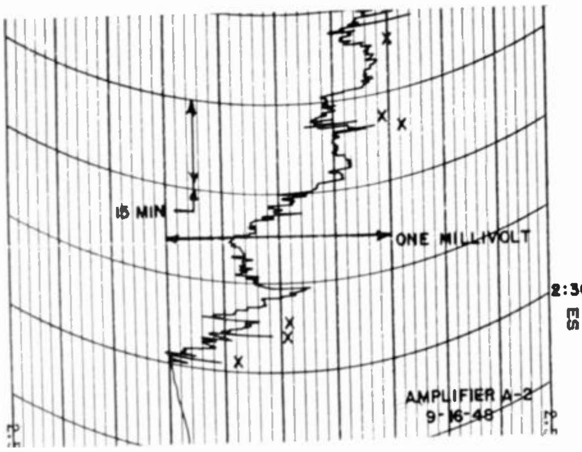
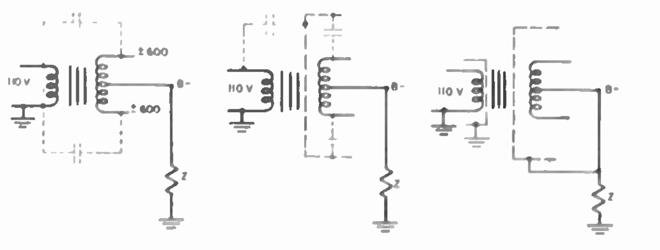


Fig. 9—Typical drift characteristic of a negative-gain amplifier.

marked by X were due to supply-line switching surges. Since this drift stability exceeds the computer requirements, no attempt was made to obtain further improvement.

The other major problems in the design of the dc amplifiers arise because the normal ground terminal ($B-$) of the amplifier cannot be connected to system ground unless isolators are used. It is conceivable that the analogue circuit impedance to ground may be as high as several thousand ohms, and that the voltage to ground may be as large as 30 or 50 volts. Two undesirable effects can arise. Unless the grid of the input tube of the amplifier is very well shielded from true ground by the $B-$ bus, the capacitance between grid and ground, in series with the normal high impedance from grid to $B-$, results in a grid signal if a voltage exists between $B-$ and true ground. It was considered desirable to have all amplifier chassis connected to true ground, so that shielding was required within the chassis. Without resorting to complete electrostatic shields, the resulting transfer characteristics for the positive-gain amplifier gave rise to an output voltage equal to approximately 0.1 per cent of the signal between $B-$ and true ground. This was considered satisfactory.

The other effect, a 60-cps ripple appearing on the $B-$ terminal if an impedance is connected between $B-$ and true ground, is more difficult to eliminate. This ripple normally arises from unbalanced capacitance between the ends of the high-voltage winding and the primary winding, as shown in Fig. 10(a). With normal "balanced-winding" procedures, a ripple of about 0.1 volt appears if the impedance Z (Fig. 10) is made equal to 1,000



(a) (b) (c)

Fig. 10—Shielding requirements for power-supply transformers.

ohms. This corresponds to an unbalanced capacitance of a few hundred $\mu\mu$f. If a single shield, as shown in Fig. 10(b), is used, conditions are not appreciably improved. If, for example, the shield is connected to true ground, the unbalanced capacitances to the shield still exist. If the shield is connected to $B-$, then the capacitance between primary winding and shield still gives rise to an objectionable signal in the output. By careful choice of connections the ripple voltage may be reduced to 40 or 50 millivolts, but this is much larger than is permissible.

A very satisfactory solution results if two complete electrostatic shields are provided, as shown in Fig. (10(c). Because of construction difficulties, perfect shielding is not economically practicable. However, commercial manufacturers have provided transformers in which the unbalanced leakage capacitance is less than 0.05 $\mu\mu$f with an increase in cost ranging only from 50 to 100 per cent. This unbalanced capacitance gives rise to a ripple signal of about 0.1 millivolt per thousand ohms impedance.

The other characteristics of the amplifier are achieved by conventional means. The required low output impedance is achieved by use of a cathode-follower output and considerable negative feedback. The available output current is increased substantially above the value for a conventional cathode follower through use of a pentode as cathode impedance. The important characteristics of the amplifiers are listed in Table I.

TABLE I
PERFORMANCE DATA OF DC VOLTAGE AMPLIFIERS

	Negative-Gain Amplifier	*Positive-Gain Amplifier*
Gain	0.5 to 50 in. 1-db steps	1 to 100 continuously variable
Maximum voltage output	±125 volts peak	±125 volts peak
Maximum current output	±65 ma peak	±65 ma peak
Input impedance	400,000 ohms	500,000 ohms
Output impedance	25 ohms	0.9 ohms
Ripple level	1 mv rms	1 mv rms
Stability to line voltage fluctuations*	7 mv/10 volts	0.04 mv/10 volts
Drift characteristics*	1 mv/hour**	1 mv/hour**
60 cps ripple across 1,000 ohms between B- and true ground	1 mv	0.1 mv

* Voltage referred to grid of input tube.
** Typical values after 1-or 2-hour warm-up period.

DC Isolator and Preamplifier

Since isolators must, in general, operate with no connection to system ground, power transformers for such devices require the elaborate shielding discussed previously. Thus it is desirable for such isolators to operate without 60-cps ac power supplies, if convenient. On the other hand, the signal voltage available is frequently small, so that a preamplifier may be desirable to minimize the noise-to-signal ratio of the isolator.

Through the use of an amplitude-modulated carrier device, a very simple isolator results, with negligibly

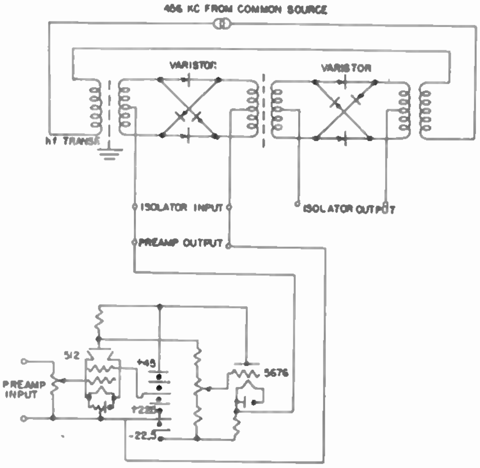


Fig. 11—Schematic diagram of a preamplifier and dc isolator. (NOTE:—Without the preamplifier, the isolator is bilateral in action.)

small capacitance coupling from input to output. Varistor bridges, used as balanced modulators as shown in Fig. 11, have given compact isolators with no 60-cps power supplies. The design of such a unit is not as straightforward as might be supposed, since the phase-shift requirement for the detected output carries over into a similar requirement regarding relative phase shifts of carrier and sidebands. However, through the use of nonresonant circuits, the phase-shift performance of the isolators can be made satisfactory.

A simple battery-operated amplifier is also shown in the circuit of Fig. 11. Both of these units can be mounted on a very small chassis. The isolator produces a useful signal of about 0.7 volt peak-to-peak and has satisfactory linearity and negligible phase shift in the range from zero frequency (dc) to 1,000 cps.

Generator of an Arbitrary Function of the Dependent Variable

A somewhat more difficult problem is to produce an arbitrary function of the dependent variable. As has been pointed out earlier, the dependent variable can be obtained as a voltage existing at some point in the analogue circuit. The problem is, then, to produce a voltage which is an arbitrary function of another voltage. Two devices have been developed for this purpose.

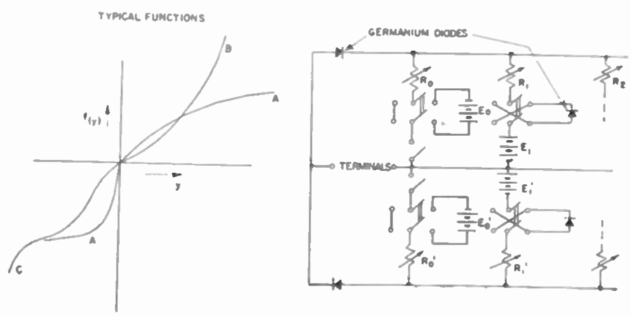


Fig. 12—Schematic diagram of a generator for the arbitrary function of a dependent variable—Type I.

One of the generators (Type I) utilizes crystal diodes and batteries in a manner shown in Fig. 12. The adjustable resistors are made relatively large compared to the forward impedance of the crystals, so that the variations of the latter with current do not affect the over-all impedance appreciably. If the switches are all thrown to the left in Fig. 12, each parallel branch will not conduct current until the terminal voltage rises above the battery voltage of that branch. The resulting current versus voltage characteristic is composed of a series of straight lines, with the slope of successive segments decreasing as the voltage rises. Such a characteristic is shown in curve A of Fig. 12.

If the switches are thrown to the right, a different characteristic results. Each branch produces a circulating current which passes through R_0. The battery voltage E_0 is provided to give zero terminal voltage when no external current enters the terminals. As voltage is applied, the current which flows is determined by the effective resistance of all branches in parallel. As soon as the terminal voltage rises above the lowest branch battery voltage, the current in that branch ceases to flow, and the effective impedance of the device is raised. As the voltage exceeds the highest battery voltage, the impedance becomes equal to R_0 alone, the maximum value. The resulting current versus voltage characteristic is one of constantly increasing slope, as shown in curve B. By combining branches of each type, a characteristic may be achieved whose slope is alternately increasing and decreasing, provided the slope always remains positive. Curve C in Fig. 12 is typical.

The generator of this type which is used with the computer has 22 parallel branches, 11 for positive and 11 for negative voltages. A current-voltage characteristic composed of 22 straight-line segments can be obtained. Such a characteristic can be made to approximate many typical functions with surprisingly good accuracy. Simpler and more special-purpose versions of this device have also been developed, in the form of abrupt current and voltage limiters. These are shown in Fig. 7 together with the Type I generator.

The other arbitrary-function generator (Type II) is based upon a principle developed elsewhere. It utilizes a cathode-ray tube and photocell. The quiescent position of the electron beam is set to one side of the tube and the photocell output is amplified and applied to the plates with a polarity which will drive the beam to the other side. If an opaque template is placed on the surface of the tube, as shown in Fig. 13, the beam spot will be driven below the edge of the template until the light reaching the phototube is just sufficient to produce a signal which will deflect the beam to the edge of the template.

The generator has several limitations which merit discussion. The finite spot size results in some distortion for which compensation is difficult, since distortion depends on the gain of the amplifier, intensity of the spot, the height of the pattern, location of the quiescent position, etc., all of which may vary from time to time. In addition, the halo on the screen contributes an appre-

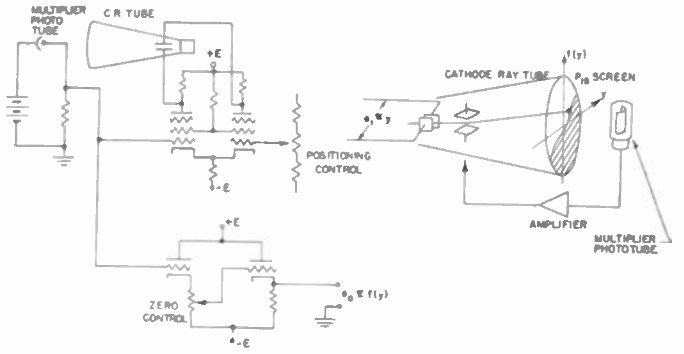


Fig. 13—Schematic diagram of a generator for the arbitrary forcing function of a dependent variable—Type II.

ciable signal which may cause clipping of sharp high peaks of the template, since the halo signal can be picked up on both sides of the peak with the spot completely hidden. A third source of error lies in the parallax resulting from a spot on one side of the cathode-ray-tube face and a pattern on the other side. Careful proportioning of over-all height and width of the pattern, location of the quiescent position, and adjustment of gain minimize these difficulties, but a definite limitation in accuracy exists. Another limitation lies in the inability of the beam to rise or fall abruptly. Thus, if the dependent variable passes rapidly through a region where the arbitrary function changes abruptly, the output may lag behind the ideal curve. This effect can be minimized by a reduction in vertical scale of the template. Since the time lag is small compared to the period of highest frequency used on this computer, the effect on solutions is usually negligible.

The Type II function generator as developed for the computer is shown in Fig. 7. Its frequency response is best defined in terms of the maximum speed of rise of the cathode-ray beam. Under typical operating conditions, this corresponds to 40 microseconds per inch of deflection along the $f(y)$ axis. This figure was obtained with a P5 phosphor screen having a decay time of approximately 20 microseconds. Screens of this type provide sufficiently fast response characteristics for general use on the computer.

Multiplier

The multiplication of two arbitrary voltages can be accomplished in many ways. The method which has been adopted for the CIT computer utilizes a double-modulation and subsequent detection scheme (see the circuit of Fig. 14). A balanced modulator is used to produce sidebands through modulation of a carrier by one of the product terms. The resulting voltage is of the form $E_1 \cos(\omega_c t)$. This voltage is used as the carrier for a second modulator, which gives a sideband output $E_1E_2 \cos(\omega_c t)$. The carrier voltage is, of course, suppressed in both cases through the use of a balanced modulator. For convenience, a varistor bridge is used for the first, and electron tubes for the second modulator. Detection is accomplished with another variator bridge, minimizing the power-supply requirements of the unit. Isolation between inputs and output is automatically supplied by the high-frequency transformers. It will be noted, however, that the second input voltage is not isolated from system ground. If such isolation is required for both inputs, an additional iso'ator must be used. Since phase shift of the sidebands results in phase shift of the detected signal, care must be taken to minimize this effect. The first modulator, with untuned transformers, performs very well in this respect. Since tuned circuits are required in the second modulator, for several reasons, the phase shift here is minimized by overcoupling of the transformer in the plate circuit of the modulator.

In the computer, groups of the dc isolators and the multipliers are supplied by a common high-frequency generator, thereby permitting small, compact units.

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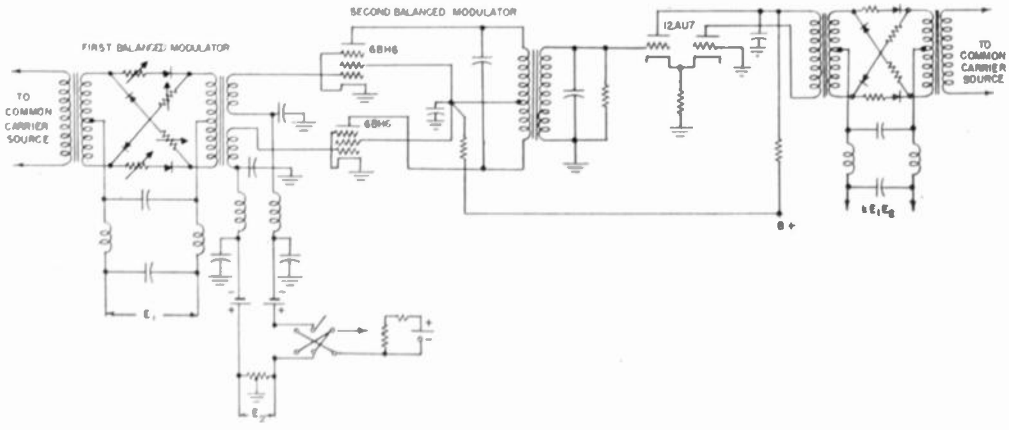

Fig. 14—Schematic diagram of a multiplier.

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Mr. Campbell was in the armed forces from 1942 to 1946, serving during the latter part of that period as a first lieutenant with the Army Security Agency. Since completing his course at McGill, he has been employed as development engineer by Minshall-Estey Organ, Inc., located in Brattleboro, Vt.

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From 1933 to 1935, he was demonstrator in physics at Cambridge University, and until 1937 he was demonstrator in physics at McGill University. From 1937 to 1938, he was a research associate at Cruft Laboratory, Harvard University. From 1938 to 1940, he was an assistant professor of physics at Acadia University, Wolfville, N. S. From 1940 to 1944, he was a research assistant and lecturer in physics at Queen's University, Kingston, Ont. He was an assistant professor of physics from 1944 to 1946 at Queen's University. During the period 1946 to 1947, he was an assistant professor of electrical engineering at McGill University, becoming an associate professor in 1947.

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Fred D. Clapp (S'43–A'44–M'45) was born on July 2, 1914, in Hanford, Calif. He became interested in radio at the age of 14 and obtained an amateur license shortly thereafter. In 1931, he decided to pursue radio and electronics as a life work.

Fred D. Clapp

There followed a succession of positions as police radio technician, broadcast station studio engineer, chief engineer and others. Upon the advent of war, Mr. Clapp entered the ESMWT training courses given at San Mateo Junior College and Stanford University in Calif., becoming an instructor during the first part of 1943. Upon termination of the program, he became associated with the University of California Radiation Laboratory, where he spent some time on electronic projects relating to the electromagnetic separation of uranium isotopes for the atomic bomb. A brief period near the end of the war was taken up as a radar field engineer for the Western Electric Company, New York, N. Y.

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❖

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For a biography and photograph of Eugene F. Grant see page 877, of the July, 1948, issue of the Proceedings of the I.R.E.

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R. C. Mathes

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D. A. McLean was born in Golden, Colo., on July 15, 1905. Since graduation from the University of Colorado in 1929 with the degree of B.S. in chemical engineering, he has been a member of the technical staff of the Bell Telephone Laboratories. He has been primarily concerned with the chemical aspects of electrical insulation with particular reference to capacitor dielectrics, and has published a number of papers in this field. Among the patents he has been granted is one covering the now widely used method of stabilizing chlorinated capacitor impregnating compounds by the addition of small quantities of anthraquinone or similar materials.

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William S. McLean

During the period 1941–1942, he was employed by the General Electric Co., Schenectady, N. Y. on the Student Test Course. During this time, he worked in the Pittsfield, Schenectady, and Philadelphia plants of General Electric, testing various types of electric power equipment. In 1942 he went on active duty in the United States Navy as a radar maintenance officer specializing in fire-control radar, attended several of the Navy radar and fire-control schools, and later served two years in the Southwest Pacific doing maintenance work.

Upon his discharge from the service in 1945, Mr. McLean was employed by National Union Radio Corporation in Orange, N. J., where he worked on the design and development of various types of secondary-emission tubes. In 1948 he was employed by the radio division of the Western Electric Co. in their field engineering force, and is now engaged in radar installation and servicing work.

❖

Carl F. Miller (A'28–M'47–SM'48) was born in Germany on January 17, 1901. He was graduated from State College, Stuttgart, Germany, in 1926, with a degree in mechanical engineering. In 1927 he emigrated to the United States, and he became a citizen in 1933.

Carl F. Miller

From 1927 to 1929, he was employed by the Federal Telephone and Telegraph Co., Buffalo, N. Y., as an engineer for the development of superheterodyne receivers. He joined the Sylvania Electric Products Co., Emporium, Pa., in 1929, and developed the first 6.3-volt line of tubes for automobile receivers; later he invented the lock-in tube and supervised the early production stage. In 1940 he became associated with the Corning Glass Works, Corning, N. Y., as design engineer for their multiform division.

National Union Radio Corp., Orange, N. J. appointed Mr. Miller as chief engineer of their Power Tube Plant in 1943, and since then he has been engaged in the development of electron tubes of various types.

❖

H. A. Sauer was born on Staten Island, N. Y., on May 3, 1908. He attended Cooper Union and New York University, from which institutions he received the B.S. degree in chemical engineering in 1930, and the M.S. degree in physics in 1939, respectively.

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❖

René Snepvangers was born in Antwerp, Belgium, in 1900. He attended various Belgian Universities until 1922. From 1928 to 1934 he was employed as an engineer with the Bell Telephone Manufacturing Company in Antwerp; then he became a consulting engineer for the Belgian Branch of the United Electrical Works, "Tungsram," of Budapest. In 1939, he came to the United States, joining the RCA Research Laboratories as a specialist in phonograph development. In November, 1944, Mr. Snepvangers became associated with CBS Recording Laboratories, and is now the senior recording research engineer.

René Snepvangers

❖

Charles H. Wilts was born in Los Angeles, Calif., on January 30, 1920. He received the B.S. degree in 1940 and the Ph.D. degree in 1948 in electrical engineering from the California Institute of Technology. From 1945 to 1947, he was a National Research Council Fellow. He is the author of three papers on application of analogue computers to engineering problems. Presently, he is an assistant professor of applied mechanics at the California Institute of Technology.

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ACOUSTICS AND AUDIO FREQUENCIES

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621.395.61/.62 1852
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621.395.61 1853
An Omnidirectional Microphone—J. K. Hilliard. (*Audio Eng.*, vol. 33, pp. 20–21, April, 1949.) The microphone is 0.6 inch in diameter and 0.4 inch thick; it weighs less than $\frac{1}{4}$ oz. It is based on electrostatic rather than electromotive principles, and is little affected by blast, iron filing damage, or magnetic induction. In sound reinforcement applications, 4 db more gain is obtainable than with other nondirectional microphones.

621.395.61 1854
Single-Element Unidirectional Microphone—H. F. Olson and J. Preston. (*Jour. Soc. Mot. Pic. Eng.*, vol. 52, pp. 293–302; March, 1949.) Developed for use in sound-film recording, with the following characteristics: single-ribbon type; the back of the ribbon is coupled to a damped folded pipe and an acoustical imped-

ance in the form of an aperture; improved cardioid directional pattern; increased output; reduced weight; reduced wind-noise response.

621.395.623.7 1855
Modern Loudspeaker Technique—M. Alixant. (*Radio Tech. Dig.* (Franç), vol. 3, pp. 83–99; April, 1949. Bibliography, pp. 99–101.) A short discussion of the fundamental theory of electrodynamic loudspeakers, with tabulated data for the principal permanent-magnet alloys and lists giving some details of permanent-magnet and electromagnetic loudspeakers manufactured in France. Modern types of American and British loudspeakers are reviewed briefly.

621.395.625.2:621.396.712 1856
Disc Recording for Broadcast Stations—W. J. Mahoney. (*Audio Eng.*, vol. 33, pp. 9–13, 46; April, 1949.) Discussion of the system designed for new studios at WSAI, Cincinnati, and of the practical problems involved in designing a complex system of recording equalizers.

ANTENNAS AND TRANSMISSION LINES

621.315.212:621.397.5 1857
The London-Birmingham Television Cable: Part 2—Cable Design, Construction and Test Results—H. Stanesby and W. K. Weston. (*P.O. Elec. Eng. Jour.*, vol. 42, part 1, pp. 33–38; April, 1949.) Details of the experimental work on the design and manufacture of the cable, the production of factory lengths, and the jointing technique used for completing repeater sections. Test results on typical sections are also discussed. Part 1: 1279 of June.

621.392.26† 1858
Reflection of H_1 Waves at a Sudden Change of the Cross-Section of a Rectangular Waveguide—K. Fränz. (*Frequenz*, vol. 2, pp. 227–231; September, 1948.) A formula for the reflection coefficient is derived which depends on the waveguide dimensions and is analogous to that for reflection at the junction of two parallel Lecher-wire systems of the same breadth but with their ends at different levels. At the reflection point, a local capacitive field exists.

621.396.67 1859
The General Problem of Antenna Radiation and the Fundamental Integral Equation, with Application to an Antenna of Revolution: Parts 1 & 2—G. E. Albert and J. L. Synge; J. L. Synge. (*Quart. Appl. Math.*, vol. 6, pp. 117–156; July, 1948.) Part 1: Radiation in a finite cavity is discussed, and an integral equation is obtained which has the same form as the basic integral equation for an antenna. The latter equation reduces to a simple explicit form for an antenna with axial symmetry. The application of the results to actual radiating systems is discussed. Part 2: The importance of the gap is emphasized; for a fuller investigation see 24 of 1948 (Infeld). Thin antennas are considered; the current is shown to be approximately sinusoidal outside the gap. A shape term is included for the case when the gap is not at the center, and is calculated explicitly for certain simple shapes. A method of successive approximations is also described for dealing with any antenna, thin or thick.

621.396.67 1860
The Helical Antenna—J. D. Kraus. (PROC. I.R.E., vol. 37, pp. 263–272; March, 1949.) The helix is considered as a fundamental form of antenna and the variation of radiation pattern and type of polarization with change in helix dimensions are discussed. For the axial mode, the radiation pattern and terminal impedance are maintained over a wide frequency range.

621.396.67 1861
A V.H.F. Helical Beam Antenna—H. E. Taylor and D. Fowler. (*CQ*, vol. 5, pp. 13–16; April, 1949.) Design details for an antenna for the 144- to 148-Mc range, with circular polarization and gain of 11.5 to 13 db.

621.396.67 1862
Optimum Design of a Cylindroparabolic Reflector—G. Klages. (*Frequenz*, vol. 2, pp. 151–154; June, 1948.) The Kirchhoff-Huyghens principle is used to calculate the zones on the surface of the reflector which reduce, by interference, the radiation along the axis. The formulas obtained enable reflectors to be designed with optimum gain.

621.396.67:538.56:535.13 1863
The Radiation and Transmission Properties of a Pair of Semi-Infinite Parallel Plates: Part 1—Heins. (*See* 1920.)

621.396.67:621.396.11 1864
Ground Absorption with Elevated Vertical and Horizontal Dipoles—R. E. Burgess. (*Wireless Eng.*, vol. 26, pp. 133–139; April, 1949.) The fraction of the power absorbed in the ground is calculated for dipole heights which are large compared with the wavelength. The dipole is assumed to have a sine-law polar diagram and the energy flow into the ground is evaluated using the Fresnel reflection coefficients which depend upon the wave polarization. The absorption factors are determined for limiting values of ground permittivity. The ground absorption for a vertical dipole is a very slowly varying function of the electrical properties of the ground and is very nearly 0.4 for most practical conditions. It is appreciably greater than that of a horizontal dipole on account of the Brewster phenomenon. The results obtained are compared with those derived earlier by Strutt, Niessen, and Sommerfeld and Renner.

The noise in an antenna system due to thermal radiation from the ground is evaluated directly using thermodynamical principles and is shown to be consistent with the value derived by application of the reciprocity principle.

Brief consideration is given to the ground absorption for antenna arrays.

621.396.671 1865
Driving Point Impedance of a Vertical Cylindrical Radiator and Concentric Ring of Subsidiary Radiators over Perfectly Conducting Earth: Parts 1 & 2—H. Cafferata. (*Marconi Rev.*, vol. 12, pp. 12–20 and 57–67; January to March and April to June, 1949.) Hallén's integral equation method is applied, with particular reference to an approximate solution for the case of 3 small subsidiary radiators symmetrically arranged round the main antenna and with base connections to it, the radius of the ring being small compared to λ. See also 854 of 1946 (Harrison).

621.396.671 1866
Measurement of Gain of Electromagnetic Horns—A. S. Dunbar and M. D. Adcock. (*Jour. Appl. Phys.*, vol. 20, pp. 226–227; February, 1949. Criticism of 1600 of July (Watson and McKinney), and discussion of an alternative method involving measurement only of the ratio of received to transmitted power.

621.396.671 1867
Mutual Impedance of Parallel Aerials—G. Barzilai. (*Wireless Eng.*, vol. 26, p. 73; February, 1949.) Corrections to 648 of April.

621.396.677 1868
Metallic Delay Lenses—S. S. D. Jones and J. Brown. (*Nature* (London), vol. 163, pp. 324–325; February 26, 1949.) Kock's formula (2176 of 1948) for the refractive index n is found experimentally to give a value 18 per cent too low. The new formula

$$n^2 = 1 + (2d/\pi a)\log_e \operatorname{cosec}(\pi b/2d),$$

where a, b and d are dimensions of the lens structure, is based on the assumption that the refracting medium behaves like a transmission line. The formula is valid for the wavelength range within which n is constant. Observed values of n agree with those predicted by this formula within the limits of experimental accuracy.

621.396.677 1869
Directional Antenna for the 152–162-Mc Communications Band—J. S. Brown and V. J. Moffatt. (*Communications*, vol. 29, pp. 14–16, 35; March, 1949.) A corner-reflector type with vertical polarization, in which two sets of parallel rods, spaced 0.1 λ apart in two vertical planes, are used instead of metal sheets. It has a unidirectional radiation pattern, a gain of at least 6 db over a λ/2 dipole and a voltage SWR < 1.5 over the whole frequency band. The feeder cable can have any impedance between 50 and 75Ω.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.3:621.314.6 1870
Voltage Surges Produced in Rectifiers by Starting without Load—T. Douma. (*Communication News*, vol. 9, pp. 121–125; August, 1948.) If a high-voltage rectifier with low damping is started without load, voltage surges up to four times the peak transformer voltage can occur across the tubes, and the voltage on the smoothing capacitors can be twice the peak alternating voltage of the transformer. These surge voltages are at least double those encountered under normal operating conditions. They can be avoided by applying damping to the input choke and providing the smoothing capacitors with a suitable series resistance which can be short-circuited after the rectifier has been switched on.

621.3.09:621.397.5 1871
Attentuation and Phase Distortion and Their Effect on Television Signals—G. Fuchs and V. Baranov. (*Câbles and Trans.* (Paris), vol. 3, pp. 194–207; April, 1949.) An examination of the relations between small attenuation and phase distortions and the deformation of the response curve of a television receiver. In one method, the distortion versus frequency characteristic is treated as a whole, while in a second method this characteristic is split up into a certain number of elementary sections. Practical limits for these distortions are calculated and the theoretical results are applied to 450-line and 1,100-line television systems.

621.314.2.012.8 1872
The Transformer and its Equivalent Representation—P. G. Violet. (*Frequenz*, vol. 3, pp. 1–12; January, 1949.) A general treatment based on quadripole theory. Matrices for the ordinary and the ideal transformer are given and the various equivalent schemes are explained and compared. The application of circuit equivalents for measurement purposes and for filter calculations is considered, in particular for filters whose branches consist of oscillatory circuits. The equivalents for transformers with two secondary windings are given and comparison is made with the normal single-secondary transformer.

621.314.3† 1873
The Basic Considerations in the Design of Push-Pull Magnetic Amplifiers—M. A. Rozenblat. (*Avtomatika i Telemekhanika*, vol. 10, pp. 32–50; January and February, 1949. In Russian.) Discussion of amplifiers of the differential bridge, and transformer types (Figs. 1, 2, and 3 respectively) with loads taking ac. Their operation is discussed and the conditions for obtaining maximum values of both power output and

amplification factor are established. Methods are indicated for determining the optimum operating conditions for a given core material. An example of the design procedure is given, together with experimental results for the amplifier so designed.

621.316.86 1874

Cracked-Carbon Resistors—R. W. Wilton. (*FM-TV*, vol. 9, p. 29; February, 1949.) Brief discussion of the characteristics of resistors made by the Welwyn Electrical Laboratories, by the pyrolitic process of depositing carbon on porcelain. Nonlinearity hardly exists. The resistance value is controlled by the cracking temperature, the hydrocarbon content of the gas mixture, and the time of exposure of the rod in the cracking zone. Such resistors are cheaper than wire-wound resistors for ratings up to 2 to 4 watts, and inductive effects are almost absent from high-value resistances, for which helical grooves may be used. Resistance values are accurate within ±1 per cent, voltage and temperature coefficients are low, and long-term stability is high. High-gain amplifiers using these resistors have little high-level hiss.

621.316.86 1875

Results of an Investigation of the Volt-Ampere Characteristics of Some Nonlinear Resistors—M. A. Topchibashev. (*Avtomatika i Telemekhanika*, vol. 10, pp. 13–24; January and February, 1949. In Russian.) The division of nonlinear resistors into two main groups (varistors and thermistors) is considered; their properties can be best represented by their volt-ampere characteristics. The equation describing these is of the form $I = AV^{\alpha}$ where A is called the coefficient of nonlinearity and α the index of nonlinearity. Analytical and graphical methods for determining A and α are discussed. Results of an experimental investigation of a number of these resistors are summarized in two comprehensive tables showing A, α, and the maximum errors for given voltage ranges.

621.318.42 1876

Axial Field and Supplementary Losses of Inductance Coils without Iron—J. Rezelman. (*Rev. Gén. Élec.*, vol. 58, pp. 154–162; April, 1949.) Formulas and curves are given for circular coils of rectangular winding cross section with experimental results.

621.318.572 1877

A New Pulse-Amplitude Discriminator Circuit—E. H. Cooke-Yarborough. (*Jour. Sci. Instr.*, vol. 26, pp. 96–97; March, 1949.) The circuit described discriminates against pulses smaller than those to be counted. The dead time is a known constant, so that counting losses can be accurately determined without complete knowledge of the resolving time of other equipment in the counting chain. These counting losses are calculated and the condition for the counting rate to be independent of the input pulse width is deduced.

621.318.572:621.396.1 1878

An Electronic Switch for Diversity Reception—H. V. Griffiths and R. W. Bayliff. (*BBC Quart.*, vol. 4, pp. 57–64; April, 1949.) The dual-diversity switch has a discriminator which monitors the signals received on two receivers and controls a gating circuit which blocks the output from the receiver carrying the weaker signal and passes that from the receiver carrying the stronger signal. Circuit details are given of this switch and of a similar device for operation with three receivers (triple-diversity).

621.319:679.5 1879

Design and Performance Characteristics of "Electrets"—E. D. Padgett. (*Tele-Tech*, vol. 8, pp. 36–39; 63; March, 1949.) See also 1307 of June.

621.319.4 1880

Capacitors with Cylinders of Different Expansion Coefficients for Compensation of the Dependence of Capacitance on Temperature—A. Rohmann. (*Funk und Ton*, vol. 3, pp. 230–233; April, 1949.) Design formulas are given, with numerical examples and tabulated temperature coefficients of capacitance for capacitors with outer cylinder of brass (37 per cent Zn) and inner cylinder either of copper or of brass with Zn content ranging from 28.1 to 37 per cent. The coefficient is zero for a capacitor with Cu inner cylinder and radii ratio of 1.164, and also for one with brass inner cylinder (31.1 per cent Zn) and radii ratio of 1.02.

621.319.4:621.315.59 1881

On the Use of Semiconducting Liquids for Impregnating Paper Capacitors—Renne. (*See* 1969.)

621.392 1882

General Impedance-Function Theory—P. I. Richards. (*Quart. Appl. Math.*, vol. 6, pp. 21–29; April, 1948.) Discussion of methods of extending lumped-constant impedance-function theory to distributed-constant circuits.

621.392:621.396.619.13 1883

A Simple Method of Calculation for Electrical Circuits Carrying Frequency-Modulated Voltages—J. W. Alexander. (*Communication News*, vol. 9, pp. 33–38; December, 1947.) An approximate simplification of Vellat's method (79 of 1942), accurate in general within 10 per cent for the final result. The important parameters governing the accuracy are the "swing-factor," or ratio of the maximum frequency deviation to the maximum modulation frequency, and the bandwidth factor, or ratio of the bandwidth to the double swing. Calculations are made as for a circuit in which the voltage varies sinusoidally at frequency f, and f is replaced by the instantaneous frequency in the final result. The distortion for a tuned circuit with and without limiter, and for a bandpass filter, is calculated by this and by other methods and the results are compared.

621.392.43 1884

A Balanced-to-Unbalanced Matching Unit for High Frequencies—J. W. Whitehead. (*Jour. Sci. Instr.*, vol. 26, pp. 71–73; March, 1949.) A circuit for matching an unbalanced circuit to a balanced resistive circuit is investigated theoretically and practically. Component values for a particular case are derived. Experimental units on the same principles have been built with certain inductances and capacitances variable; they have been used successfully with a 3.5-kw transmitter throughout the frequency range 2.5 to 20 Mc. These units can easily be constructed, and adjustments for large changes of frequency can quickly be made since coil changing is unnecessary.

621.392.43 1885

Impedance Transformation by Means of Cables—W. Burkhardtsmaier. (*Funk und Ton*, vol. 3, pp. 151–167 and 202–213; March and April, 1949.) Discussion of the use of sections of cable, with characteristic impedance variable along their length, as impedance transformers. Two types are considered; in the first, the impedance varies stepwise from one end to the other, while in the second, the impedance variation is uniform. Such transformers can be used with advantage not only for long and medium waves, but also for short waves if suitably designed. Design formulas are derived and practical types of equipment for concentric cables and twin conductors are described. In many cases, better results can be obtained by the addition of compensation units at each end of the transformer section. The design of such units is also considered.

621.392.5 1886

On the Synthesis of the Most General Passive Quadripoles—R. Leroy. (*Câbles and Trans.* (Paris), vol. 3, pp. 141–158; April, 1949.) The matrices of the impedances or admittances of passive quadripoles are "positive real" matrices. The construction of quadripoles from such matrices is straightforward when their elements comprise only two types of impedance, but this is not the case when the matrices include resistances, inductances, and also capacitances. The possibility of obtaining passive quadripoles corresponding to such matrices was shown by Gewertz (*Jour. Math. Phys.*, vol. 12, pp. 1–254; 1932–1933) whose paper appears to-day to be largely ignored by technicians. Gewertz studies a particular class of matrices in which the determinant of the even parts of the elements is zero and for which he gives a method of reduction. He also shows that the representation of every "positive real" matrix can be referred to a matrix of this class.

The present paper, by more systematic application of the theory of functions and of "positive real" matrices, simplifies the results of Gewertz and gives physical reasons for certain properties, sometimes established by formal calculus, which may not appear altogether satisfactory. Gewertz' method is also applied to the realization of effective impedances. See also 2539 of 1940, 1627 and 3017 of 1941 (Cauer) and 1887 below.

621.392.5 1887

Reactive Ladder Networks—J. Ville. (*Câbles and Trans.* (Paris), vol. 3, pp. 159–176; April, 1949.) A theoretical analysis based on the properties of the input impedance of ladder networks terminated by an ohmic resistance. This impedance only defines a network with a limited number of cells, which are all-pass. All-pass cells which cannot be realized by ladder construction are essentially of the lattice type.

Systematic use is made of shunt frequencies, for which the shunt arms of the ladder network are short-circuited, cutting out the terminal ohmic resistance; the input impedance then resembles a reactance and the shunt frequencies can, therefore, be determined *a priori* by finding the frequencies for which the input impedance has the properties of a reactance. The shunt frequencies are roots of the equation whose first member is obtained by equating to zero the even part of the input impedance. Comment on certain parts of the paper is made by R. Leroy. See also 2190 of 1948 and 1886 above (Leroy).

621.392.52 1888

Introduction to the Operating-Parameter Theory of Filter Calculation—C. Wisspeintner. (*Frequenz*, vol. 2, pp. 131–140, 154–163, 190–199, and 210–214; May to August, 1948.) The principal results of the mathematical theory of filters are presented in as simple a form as possible to make them readily available to the practical engineer. Quadripole theory is first considered and in this the usual complex quantities are denoted by German characters, while Roman type is used in the treatment of the operating-parameter theory. The wave-parameter theory of quadripoles includes discussion of the properties of various types of filter. In practice, it is seldom possible to terminate a filter with its characteristic impedance and, in consequence, reflections occur which affect the attenuation both within and outside the pass band. The operating-parameter theory takes account of such reflections. The relation between over-all, current, and voltage attenuation is discussed and also echo attenuation. The determination of the no-load resistance for a given attenuation is explained in detail with the aid of numerous diagrams. Application of the theory is illustrated by numerical calculations for many different types of filter. Results for a particular band-pass filter are in fairly good agreement with measurements.

621.392.52 1889
Variable Bandwidth Crystal Filters—B. Sandel. (*Radiotronics*, pp. 78–87; September and October, 1948.) The behavior of the equivalent circuit of a quartz crystal is discussed. For best over-all selectivity, the shunt capacitance in this circuit should be neutralized and practical arrangements for achieving this are described. To vary the bandwidth, a tuned circuit should be used as the load for the filter; the dynamic impedance of this circuit can be altered by the insertion of series or parallel resistors. The stage gain is calculated, with due allowance for bandwidth change. A design example is discussed in detail.

621.392.52 1890
The Input Impedance of Some Low Pass Filters with Resistance Terminations, with Reference to Class "B" Modulator Applications—H. R. Cantelo. (*Marconi Rev.*, vol. 12, pp. 41–56; April to June, 1949.) The properties of certain *m*-derived filters terminated by a load of constant resistance are examined at frequencies up to 1.5 times the cutoff frequency. With a preferred configuration of filter, the secondary leakage inductance of the modulation transformer can be included as part of the filter system. Formulas and curves are given for 3 types of filter. The attenuation and phase shift between input and output of the filter system are also calculated.

621.392.52:621.395.44 1891
Telephony Filters and the Effect of Termination and Losses on Their Characteristics: Parts 1 & 2—J. F. Schouten. (*Communication News*, vol. 9, pp. 61–69 and 97–106; April and August, 1948.) If we assume that all the elements of a reactive network have the same loss-factor, the problem of calculating the effect of these losses on the network can be reduced to that of determining a function of a complex argument when the function is known for an imaginary argument. Three methods of solution are discussed: (a) the method of the perpendicular derivative, based on the differential expressions of Cauchy and Riemann, (b) representation, based on Laplace's differential equations, of the attenuation and phase of a filter by means of soap-film models, and (c) an interpretation, based on Cauchy's integral, analogous to a blurred image in optics.

Within certain limits, a relation exists between the attenuation and phase in a network, as a function of frequency; this relation is ultimately based on the causal relation between the input and output voltage. The soap-film method is applied to several common types of simple filter which can be considered as prototypes of more complex filters.

621.392.52:621.396.621:621.396.5 1892
Application of Crystal Filters in [Telegraphy] Receivers—F. Maarleveld. (*Tijdschr. ned. Radiogenoot.*, vol. 14, pp. 41–55; March, 1949. Discussion, p. 56. In Dutch with English Summary.) The filters are designed for an if of 100 kc and bandwidths of 50, 250, 500, and 1,500 cps respectively. Design formulas are tabulated. See also 2041 below.

621.395.667:621.397.6 1893
Phase and Amplitude Equalizer for Television Use—E. D. Goodale and R. C. Kennedy. (*RCA Rev.*, vol. 10, pp. 35–42; March, 1949.) Describes an equalizer of variable frequency response. Two amplifiers having resonant-circuit loads of variable damping and resonance frequency are used for amplitude equalization. To compensate for the phase distortion in the applied signal and for that produced in the amplitude equalizer, amplifiers with variable phase shift and constant gain are used.

621.396.611.3 1894
Oscillator Power Variation and Frequency Pull-In—L. S. Schwartz. (*Tele-Tech*, vol. 8, pp. 30–32, 57; January, 1949.) An analysis of the effect of coupling variation for a mismatched feeder and an oscillator. Curves show how the oscillator power varies with coupling coefficient and SWR. Results of measurements using a lighthouse triode and cavity resonator are also included.

621.396.615 1895
Frequency Variation of a Crystal-Bridge Oscillator—W. Herzog. (*Arch. Elek.* (Ubertragüng), vol. 2, pp. 357–361; December, 1948.) A simple bridge arrangement in the grid circuit of a tube oscillator enables a considerable variation of the crystal resonance frequency to be achieved. An auxiliary parallel circuit may be used to suppress one of the two possible oscillation frequencies.

621.396.615 1896
Variable Frequency R-C Oscillators—F. Butler. (*Electronic Eng.*, (London), vol. 21, pp. 140–142; April, 1949.) Discussion of phase-shift types having a frequency-selective network associated with a suitable amplifier. A practical oscillator circuit for the frequency range 250 to 5,000 cps is described in detail.

621.396.615:621.384.612.1† 1897
Pulsed Oscillator for F.M. Cyclotron—J. W. Burkig, E. L. Hubbard, and K. R. MacKenzie. (*Rev. Sci. Instr.*, vol. 20, p. 135; February, 1949.)

621.396.615:621.396.619.13 1898
Single-Valve Frequency-Modulated Oscillators: Parts 1 & 2—K. C. Johnson. (*Wireless World*, vol. 55, pp. 122–123 and 168–170; April and May, 1949.) Part 1: A coil in the anode circuit of a pentode, coupled inductively to the cathode circuit, is used to modulate the effective value of the main tuning inductance without disturbing the oscillatory circuit seriously. The amplitude of oscillation can be very nearly constant over frequency ranges at least as great as ±15 per cent and at central frequencies up to 10 Mc, since all the effects of circuit capacitance can be tuned out and the tube input capacitance is unimportant.

Part 2: Details of design and operation.

621.396.615.17 1899
Theoretical and Experimental Study of a Generator of Periodic Slave-Frequency Electrical Pulses—R. Legros. (*Rev. Gén. Elec.*, vol. 58, pp. 143–154; April, 1949.) Two circuits are discussed, of which one is an amplifier-limiter transforming an input voltage of variable wave form, amplitude, and frequency into a square-wave voltage of the same frequency, but almost constant wave form and maximum amplitude. The second circuit is a pulse generator operated by this square-wave output and providing pulses of duration short compared with their recurrence period. Amplifier-limiter circuits using (a) diodes, (b) pentodes as limiters are investigated; neither diodes nor pentodes are entirely satisfactory by themselves, but a combination of the two gives excellent results. Pulses obtained by use of a *RC* differentiator circuit are shown to be practically independent, as regards peak voltage and pulse shape, of the initial voltage wave form applied to the amplifier limiter.

621.396.615.17:621.317.755 1900
A Linear Resistance-Charged Gas Relay Time Base—E. J. B. Willey. (*Electronic Eng.* (London), vol. 21, p. 101; March, 1949.) A circuit designed to avoid the type of transformer ordinarily used, which has one winding very well insulated and also screened to minimize the risk of pickup in the sensitive grid and cathode circuits of the relay. Full component details are given.

621.396.615.17:621.317.755 1901
A Wide-Range Saw-Tooth Generator—P. G. Sulzer. (*Rev. Sci. Instr.*, vol. 20, pp. 78–80; January, 1949.) Description of a circuit which can be used as a timebase at frequencies between 15 cps and 500 kc, with good linearity and rapid flyback. A cathode-coupled multivibrator is modified by including a parallel *RC* combination in the cathode lead of the first tube.

621.396.615.17:621.385.032.24 1902
Grid Current with RC Coupling—H. T. Ramsay. (*Wireless Eng.*, vol. 26, pp. 113–118; April, 1949.) A mathematical analysis of the effect of grid current on the performance of a cathode-coupled multivibrator (Schmitt) circuit, to find the optimum values for the components for operation at a given minimum frequency. Experimental results support the theory.

621.396.645 1903
High-Frequency Amplifier Stages, with Particularly High Sensitivity and Cut-Off Sharpness, for Short-Wave and Ultra-Short-Wave Apparatus—H. Rückert. (*Arch. Elek.* (Übertragung), vol. 3, pp. 24–31; January, 1949.) Discussion with particular reference to input-circuit design, noise factor, and the effect of feedback. Results with a circuit of high sensitivity are given and also a few data for some suitable pentodes.

621.396.645 1904
10-kw. F.M. Broadcast Amplifier—J. R. Boykin. (*FM-TV*, vol. 8, pp. 18–20; December, 1948.) A grounded-grid amplifier, whose anode voltage is only 3,700 volts. Two triodes, Type WL3X2500A-3, are used. An apparent efficiency of 100 per cent is achieved in the frequency band 88 to 108 Mc.

621.396.645:537.311.33:621.315.59 1905
Some Novel Circuits for the Three-Terminal Semiconductor Amplifier—W. M. Webster, E. Eberhard, and L. E. Barton. (*RCA Rev.*, vol. 10, pp. 5–16; March, 1949.) Equivalent circuits for dc and ac are proposed for a transistor. The ac circuit is a T-network of resistances between base, emitter (positive electrode), and collector (negative electrode) with a generator in series with the collector. The generator has an emf μ times the voltage developed across the emitter resistance by the emitter current. Three amplifier connections are analyzed. The first, already described by Bardeen and Brattain (264 of February) has the base as a common electrode, with input to the emitter and output from the collector. The two new connections, which have input and output impedances more suitable for most purposes, have the input to the base, and output respectively from the emitter and the collector. These connections are applied to a direct-coupled amplifier, a 2-terminal oscillator, a relaxation oscillator, and a flip-flop circuit.

621.396.645:539.16.08 1906
A Simple Counter-Tube Amplifier—H. Goldstein. (*Z. Angew. Phys.*, vol. 1, pp. 329–330; March, 1949.) A 2-tube amplifier suitable for operating a counting mechanism is described; its usefulness is limited by relatively low gain. A similar 3-tube circuit has adequate gain for most purposes. Both amplifiers are designed for 200-volt anode supply.

621.396.645:621.317.733 1907
An Amplifier for Use with Conductance Bridges—D. J. G. Ives and R. W. Pittman (*Trans. Faraday Soc.*, vol. 44, pp. 644–646; September, 1948.) Two amplifying channels are used, one being fed with the bridge signal plus the inseparable interference and the other with interference alone, collected by an antenna near the bridge. The outputs are combined to cancel the interference, leaving a clear signal

in the final output. Full circuit details are given of an amplifier sensitive to a few microvolts at 1,200 to 5,000 cps.

621.396.645:621.383 1908

The Development of a Photoelectric A.C. Amplifier with A.C. Galvanometer—J. M. W. Milatz and N. Bloembergen. (*Physica, 's Grav.*, vol. 11, pp. 449–464; March, 1946. In English.) An ac galvanometer is fed by the same ac source as that for a synchronous motor driving a sector disk which interrupts the light falling on a photo cell. This device is sensitive only to 50-cps ac of the right phase. Fluctuations due to thermal effects were practically eliminated by cooling the photo cell with liquid air.

621.396.645.029.3 1909

Compact 6AS7G Amplifier for Residence, Audio Systems—C. G. McProud. (*Audio Eng.* vol. 33, pp. 17–19, 40 and 16–19; March and April, 1949.) Circuit and construction details of an amplifier having output power about 6 watts, switching for selecting radio or phonograph input, sufficient gain and low-frequency equalization for low-level magnetic pickups, separate high- and low-frequency tone controls, and remote control facilities.

621.396.645.371 1910

Negative-Feedback Amplifiers—C. F. Brockelsby. (*Wireless Eng.*, vol. 26, pp. 43–49; February, 1949.) When negatve feedback is applied to an amplifier with two or more stages the frequency characteristic develops peaks at the edges of the band if a certain small amount of feedback is exceeded. These peaks get higher and sharper with increasing feedback until self-oscillation begins. The flatness of the frequency characteristic can be improved, without using filters, by properly distributing the gain between the stages of the amplifier. The frequency characteristics of 2- and 3-stage amplifiers are discussed theoretically; results are confirmed experimentally. It is suggested that a 3-stage amplifier should have two identical wide-band stages and one narrow-band stage; this is contrary to the usual practice.

621.396.662 1911

Fixed-Frequency F.M. Tuners—F. A. Spindell. (*FM-TV*, vol. 9, pp. 16–17, 32; April, 1949.) A circuit diagram of the basic unit, model RP-23, is given and discussed. Models RP-24 and RP-25 have additional control units for selecting in sequence two preset audio volume levels and for switching on and off. The first half of a 12AX7 is associated with a bridged-T network and acts as a narrow-band amplifier at the control frequency. The second half of the 12AX7 amplifies the signal, which is applied to a 12AU7 operating a sensitive relay. This relay energizes a sequence relay which selects the required audio output level.

621.396.69:061.3 1912

Developments in Components—(*Wireless World*, vol. 55, pp. 133–138; April, 1949.) Brief details of exhibits at the Radio Component Manufacturers' Federation 1949 exhibition. See also *Overseas Eng.*, vol. 22, p. 310; April, 1949; and *Electronic Eng.*, vol. 21, pp. 146–149; April, 1949.

621.396.615.17/.18+621.317.35 1913

Waveform [Book Review]—Chance, Hughes, MacNichol, Sayre, and Williams. (*See* 2006.)

GENERAL PHYSICS

535+621.396 1914

Interference, Diffraction and Spectral Resolution in Optics and Radio—G. S. Gorelik. (*Uspekhi Fiz. Nauk*, pp. 407–415; November, 1948. In Russian.) A unified conception of the phenomena common to both sciences.

535.215:621.383 1915

On Photoelectric Voltages in Light-Absorbing Materials—A. P. Snoek and C. J. Gorter. (*Physica, 's Grav.*, vol. 11, pp. 426–432, February, 1946. In English.)

535.215:621.383 1916

On the Interpretation of Observations on the Photoelectric Voltages with Intermittent Light—C. J. Gorter, L. J. F. Broer, and A. P. Snoek. (*Physica, 's Grav.*, vol. 11, pp. 401–411, February, 1946. In English.)

537.521 1917

On the Dielectric Properties of a Gas Discharge—E. E. Salpeter and R. E. B. Makinson. (*Proc. Phys. Soc.*, vol. 62, pp. 180–188; March 1, 1949.) Formal expressions are derived for the space currents and electrode currents obtained by applying a sinusoidal voltage to a pair of electrodes projecting into a gas discharge.

537.521.7 1918

Experimental Investigations of the Electrical and Optical Phenomena in Spark Breakdown in Gases—E. Fünfer. (*Z. Angew. Phys.*, vol. 1, pp. 295–304; March, 1949.)

538.122 1919

Lines of Magnetic Force Made Visible—(*Radio Tech.* (Vienna), vol. 25, pp. 259–260; April, 1949.) A special gas-filled diode, when brought near the poles of a magnet, shows the lines of force as luminous streaks. The electrons inside the tube travel along a narrow path coincident with a tube of force and ionize the gas in the tube. The field can be plotted by moving the tube to different positions relative to the magnet poles. A tube with 8 anodes, giving 8 narrow electron beams, is described.

538.56:535.13:621.396.67 1920

The Radiation and Transmission Properties of a Pair of Semi-Infinite Parallel Plates: Part 1—A. E. Heins. (*Quart. Appl. Math.*, vol. 6, pp. 157–166; July, 1948.) A plane monochromatic electromagnetic wave is incident upon the plates, which are assumed thin and perfectly conducting. The amplitude and phase of the wave traveling between the plates are calculated. The structure is regarded as a two-dimensional receiving antenna and its radiation pattern is obtained. Its properties as a transmitting antenna will be considered later. See also 2756 and 3504 of 1947 (Carlson and Heins).

538.569.4:546.331.31–145.1 1921

Absorption of Ultra-High-Frequency Waves in Salt Solutions—S. K. Chatterjee and B. V. Sreekantan. (*Indian Jour. Phys.*, vol. 22, pp. 547–552; December, 1948.) For aqueous solutions of $MgCl_2$, $CuSO_4$ and KCl, the absorption maxima in the range 300 to 500 Mc shift toward higher concentration for higher frequencies. See also 3099 of 1948.

53/54 1922

Table of Physical and Chemical Constants [Book Review]—G. W. C. Kaye and T. H. Laby. Longmans, Green and Co., London, 10th edition 1948, 194 pp., 21 s. (*Proc. Phys. Soc.*, vol. 62, p. 209; March 1, 1949.) Considerable revision has been undertaken; the fundamental constants have been corrected in the light of recent determinations, and derived constants re-calculated.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.745:550.385 1923

Magnetic Storms and Solar Activity, 1948—H. W. Newton. (*Observatory*, vol. 69, pp. 38–40; February, 1949.) Provisional sunspot numbers, magnetic storms, and sudden commencements are tabulated and discussed.

523.746:550.38 1924

Observational Aspects of the Sunspot-Geomagnetic Storm Relationships—H. W. Newton. (*Mon. Not. R. Astr. Soc., Geophys. Supplement*, vol. 5, pp. 321–335; March, 1949.) Full paper, summary abstracted in 1370 of June.

538.12:521.15 1925

The Magnetic Field of Massive Rotating Bodies—P. M. S. Blackett. (*Phil. Mag.*, vol. 40, pp. 125–150; February, 1949.) Continuation and elaboration of 3112 of 1947.

550.38 "1947.10/.12" 1926

Cheltenham [Maryland] K Indices for October to December, 1947—In 379 of March the author's initials should be R. E.

550.384.4:551.510.535 1927

Daily Magnetic Variations Near the Equators—D. F. Martyn. (*Nature* (London), vol. 163, pp. 685–686; April 30, 1949.) The observed anomalies appear to require the existence in the ionosphere of a region of high electric conductivity, about 15° of latitude wide, encircling the world approximately midway between the magnetic equator (line of zero inclination) and the geomagnetic equator (equidistant from the magnetic poles). See also 1936 and 2790 of 1948 (Egedal) and 710 of April (Chapman).

551.510.52 1928

The Structure of the Temperature Field in a Turbulent Stream—A. M. Obukhov. (*Bull. Acad. Sci. U.R.S.S., Sér. Géogr. Géophys.*, vol. 13, no. 1, pp. 58–69; 1949. In Russian.) Results obtained are mentioned in 2024 below.

551.510.535 1929

On the Connections between an Oxygen-Atom Zone, Ionized Layers and Radiation due to the Mögel-Dellinger Effect—E. Schröer. (*Z. Met.*, vol. 1, pp. 110–113; January and February, 1947.) Calculations based on the most probable values of dissociation and recombination coefficients indicate that the oxygen-atom zone extends upward from a height of about 100 km. Recombination takes place so slowly that the zone persists through the night. The *F*-layer may be due to photoionization of the N_2 molecule. Radiation in the band 661 to 765Å could produce the *E*-layer by photoionization of O_2; radiation in the band 910 to 1010 Å could, similarly, produce a layer at a height of 60 to 100 km with ion concentration normally insufficient for the reflection of electromagnetic waves.

551.510.535 1930

Ionospheric Disturbances and Their Terrestrial effects—G. Leithäuser. (*Funk und Ton*, vol. 3, pp. 127–143; March, 1949.) A general discussion, with particular reference to observations in Germany of ionosphere layer heights, aurora and sporadic-*E*, and correlation with propagation phenomena and geomagnetic effects.

551.510.535:523.3:621.396.11 1931

Moon Echoes and Penetration of the Ionosphere—Kerr, Shain, and Higgins. (*See* 2030.)

551.510.535:551.557 1932

Atmospheric Currents at a Height of 120 km—C. Hoffmeister. (*Z. Met.*, vol. 1, pp. 33–41; November and December, 1946.) Observations extending from 1922 to 1945 are tabulated and discussed. At heights above 100 km in latitude 50° N, currents are observed in summer from SSW to SW with stable velocities of about 50 miles per second and relatively little dispersion in either direction or velocity. In winter two current systems are found, one identical with that of summer but deviated slightly westward, the other consisting of polar-air currents from NW to NE. Velocities are higher in winter than in summer, with mean values of 65 miles per second for the SW currents and 89 miles per second for the northerly. The dispersion in the observed values of direction and velocity is much greater in winter than in summer. 21 per cent of the observed velocities exceeded 100 miles per second; velocities above

200 miles per second were seldom noted; the highest value was 267 miles per second. Possible explanations of the effects are suggested.

551.510.535:621.317.79 1933
Sweep Frequency Ionosphere Equipment—Sulzer. (*See* 2004.)

551.510.535:621.396.11 1934
The Ionosphere over Mid-Germany in February 1949—Dieminger. (*Fernmeldetech. Z.*, vol. 2, p. 94; March, 1949.) Mean values of the limiting frequencies and heights of the *E* and *F* layers, deduced from measurements at Lindau über Northeim, are plotted and discussed. From February 12 to 21, the limiting frequency of the F_2 layer was unexpectedly high and on three occasions about midday exceeded 14 Mc.

551.510.535:[621.396.61+621.396.621 1935
Ionosphere Sounding [equipment]—Maguer (*See* 2069.)

551.593.9 1936
The Luminescence of the Atmosphere—I. A. Khvostikov. (*Uspekhi Fiz. Nauk.*, no. 3, pp. 372-386; November, 1948. In Russian.)

LOCATION AND AIDS TO NAVIGATION

621.396.9:523.3 1937
Detection of Radio Signals Reflected from the Moon—J. H. DeWitt, Jr., and E. K. Stodola. (Proc. I.R.E., vol. 37, pp. 229–242; March, 1949.) The moon is considered as a radar target and formulas are obtained for the attenuation between transmitter and receiver antennas. An experimental equipment and measuring technique used for obtaining reflections from the moon are described and results obtained are given. The possibility of using signals reflected from the moon for communication purposes is considered. See also 2915 of 1948 (Grieg, Metzger, and Waer).

621.396.93 1938
V.H.F. Automatic Direction Finder—(*Engineer* (London), vol. 187, p. 197; February 18, 1949.) A short account of the remotely operated installation at Brussels airport. Probable error should be 1 degree or less, with a range of 100 nautical miles to aircraft at 10,000 feet. The development of such equipment was fully discussed by Cleaver (1078 of May).

621.396.93:519.271 1939
Theory of Error Distribution: Application to Radio Navigational Aids—P. F. Duncan (*Wireless Eng.*, vol. 26, pp. 49–52; February, 1949.)

621.396.93.029.62 1940
Development of Single-Receiver Automatic Adcock Direction-Finders for Use in the Frequency Band 100–150 Megacycles per Second—R. F. Cleaver. (*Elec. Commun.*, vol. 25, pp. 337-362; December, 1948.) Reprint, with minor additions, of 1078 of May.

621.396.93.088 1941
H.F. Transmitter for D.F. Measurements—B. G. Pressey. (*Wireless Eng.*, vol. 26, pp. 124-128; April, 1949.) Discussion of the principles of design and the details of construction and operation of an elevated screened-loop transmitter. The loop can be rotated about a horizontal axis to produce a wave of known polarization for the measurement of the polarization error of direction-finding systems. The parallax error that occurs if a dipole is used as the radiator is thus eliminated.

621.396.932:621.396.9 1942
The Development of Shipborne Navigational Radar—R. F. Hansford. (*Jour. Inst. Nav.*, vol. 1, pp. 118–141; April, 1948. Discussion, pp. 141-147.) Survey based on two papers presented at the International Meeting whose report was noted in 3135 of 1947. Some of the navigational data has been elaborated and the account has been brought up to date; possible future developments are discussed.

621.396.933 1943
Note on a Short-Range Radio Position-Finding System Using Modulated Continuous Waves—R. F. Cleaver. (*Elec. Commun.*, vol. 25, pp. 363–372; December, 1948.) Reprint, with minor additions, of 96 of February.

621.396.933.2 1944
Radio Distance-Measuring Equipment for Aerial Navigation—(*Engineering* (London), vol. 167, p. 187; February 25, 1949.) Long summary of paper abstracted in 1402 of June (Busignies).

621.396.933.2:621.396.826:621.396.619.16 1945
Elimination of Reflected Signal Effects in Pulsed Systems—D. O. Collup. (*Tele-Tech*, vol. 8, pp. 38–40, 64; January, 1949.) Multipath reflections can increase the apparent width of the shorter radar pulses in systems which depend on pulse width or pulse spacing to convey information. The incoming direct pulse can, however, be used to make the receiver insensitive to pulses whose amplitude is not constant over the whole pulse width. An echo-suppression circuit of this type for a radar transponder beacon is described.

621.396.933.24 1946
Rotating Field Beacons—F. de Fremery. (*Communication News*, vol. 9, pp. 2–20; September, 1947.) Beacons rotated mechanically cannot provide information quickly enough for antenna navigation. This difficulty can be overcome by rotating the field electrically, and comparing its phase with that of a pulsating field, but the two fields must not have the same frequency as they could not then be separated in the receiver. The properties of simple and modulated rotating fields are discussed. If two fields have harmonically related frequencies, the phase angle between them can be measured accurately with simple apparatus at the receiver, and this phase angle is a multiple of the azimuth. A rough bearing is obtained from modulation with fields of frequencies f_0 and $2f_0$ rotating in the same sense; the accurate bearing, which by itself would be ambiguous, is obtained from modulation of fields of frequencies f_0 and nf_0 rotating in opposite senses. Apparatus for transmitting and receiving such fields is described.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788 1947
Construction and Theoretical Analysis of a Direct-Reading Hot-Wire Vacuum Gauge with Zero Point Control—J. A. H. Kersten and H. Brinkman. (*Appl. Sci. Res.*, vol. A1, no. 4, pp. 289-305; 1949.) Description of a gauge connected to a special bridge, unaffected by voltage fluctuations or changes in ambient temperature. Pressures between 10^{-6} and 10^{-1} mm Hg can be measured.

531.788.7 1948
Design of an Ionization Manometer Tube—D. L. Hollway. (*Elec. Commun.*, vol. 25, pp. 373–385; December, 1948.) An abridged version was abstracted in 3916 of 1947.

535.37 1949
Luminescent Solids [Phosphors]—H. W. Leverenz. (*Science*, vol. 109, pp. 183–195; February 25, 1949.) A general discussion of the properties of typical phosphors, luminescence mechanism, and luminescence emission spectra.

535.37:535.61-15 1950
Inertia Effects in Infra-Red Sensitive Phosphors—F. R. Scott, R. H. Thompson, and R. T. Ellickson. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 64 67; January, 1949.) Discussion of results of measurements of the time required for certain phosphors to acquire maximum brightness when exposed to infrared radiation.

535.37:621.315.61.011.5 1951
The Optical and Electrical Properties of Zinc Silicate Phosphors—E. Nagy. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 42–49; January, 1949.) A close connection was found between dielectric loss and luminescence. See also 1051 of 1948 (Szigeti and Nagy).

537.311.33 1952
Electron Diffraction and Rectification from Silicon and Pyrite Surfaces—J. M. Cowley and J. L. Symonds. (*Trans. Faraday Soc.*, vol. 44, pp. 53–60; January and February, 1948.) For the best rectification, the crystal lattice should be almost perfect and free from fracture or mosaic structure.

538.213 1953
Permeability Decrease with Increasing Frequency—H. J. van Leeuwen. (*Physica*, 's *Grav.*, vol. 11, pp. 35–42; January, 1944. In Dutch, with French summary.) Theoretical discussion.

538.221 1954
The Magnetic Structure of High-Conductivity Alloys: Part 1—On Certain Peculiarities of the Magnetization Curves and Hysteresis Loops of Alnico and Vicalloy—L. A. Shubina and Ya. S. Shur. (*Zh. Tech. Fiz.*, vol. 19, pp. 88–94; January, 1949. In Russian.) Part 2, 1955 below.

538.221 1955
The Magnetic Structure of High-Conductivity Alloys: Part 2—The Effect of Thermal Treatment on the Electrical Resistance of Alnico—V. I. Drozhzhina, M. G. Luzhinskaya, and Ya. S. Shur. (*Zh. Tekh. Fiz.*, vol. 19, pp. 95–99; January, 1949. In Russian.) Part 1, 1954 above.

538.221 1956
The Structure and Properties of the Alloy Cu_2MnIn—B. R. Coles, W. Hume-Rothery, and H. P. Myers. (*Proc. Roy. Soc.* A, vol. 196, pp. 125–133; February 22, 1949.)

538.221:621.317.4.029.62 1957
Ferromagnetism at Very High Frequencies: Part 2—Method of Measurement and Processes of Magnetization—Johnson and Rado. (*See* 1994.)

538.221:621.318.22:538.652 1958
The Cause of Anisotropy in Permanent Magnet Alloys—K. Hoselitz and M. McCaig. (*Proc. Phys. Soc.*, vol. 62, pp. 163–170; March 1, 1949.) Magnetostriction measurements on samples of Alcomax II indicate that, in the absence of a field, the domain magnetization is along the easy crystallographic direction which makes the smallest angle with the axis of anisotropy.

539.23:537.311 1959
Measurements of the Electrical Resistance of Thin Films of Copper, Silver and Lead—A. van Itterbeek and L. de Greve. (*Physica*, 's *Grav.*, vol. 11, pp. 78–90; February, 1944. In French.) Results are presented for films of thickness up to 100 mμ from room temperature down to the boiling point of liquid hydrogen. The temperature coefficient of resistance for Cu and Ag films is much less than that of the ordinary metal and becomes negative for a thickness of 3 mμ Cu or 4.5 mμ Ag.

539.23:537.311 1960
Measurements of the Electrical Resistance of Superposed Metallic Films—A. van Itterbeek and L. de Greve. (*Physica*, 's *Grav.*, vol. 11, pp. 465–469; March, 1946. In French, with English summary.) The resistance of Ag/Cu or Cu/Ag films agrees fairly well with that calculated for resistances in parallel, except when the film first deposited is extremely thin.

539.23:537.311:546.74 1961

Some Measurements on Thin Films of Nickel—A. van Itterbeek and L. de Greve. (*Physica, 's Grav.*, vol. 11, pp. 470–474; March 1946. In French, with English summary.) For the principal results see 343 and 3305 of 1946.

546.281.26 1962

On the Causes of the Nonlinearity of the Volt-Ampere Characteristic of Carborundum—V. I. Pruzhinina-Granovskaya. (*Zh. Tekh. Fiz.*, vol. 19, pp. 100–110; January, 1949. In Russian.)

546.431.82 1963

The Dielectric Behavior of $BaTiO_3$ Single-Domain Crystals—W. J. Merz. (*Phys. Rev.*, vol. 75, p. 687; February 15, 1949.)

549.514.51 1964

Production of Large Artificial Quartz Crystals—I. Franke and M. H. de Longchamp. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 1136–1137; March 28, 1949.) An autoclave process is described briefly. A seed crystal is arranged in a cool part and quartz or fused silica in the hottest part of the alkaline solution in the vessel, the pressure in which is maintained at 60 atmospheres and the temperature at 200 to 300°C. Homogeneous crystals 3×2 ×0.5 cm, with good electrical properties, are produced in 3 to 4 weeks.

549.514.51:534.133 1965

Notes on the Frequency-Temperature Relationship of Some Low Frequency Quartz Plates—D. Fairweather and N. J. Beane. (*Marconi Rev.*, vol. 12, pp. 68–80; April to June, 1949.) Longitudinal oscillations are mainly considered; various types of cut are discussed. Flexural types of vibration are likely to increase in importance but more information is required. The Bell Telephone Laboratory terminology for angles of cut (see 1995 of 1944) is used. The constants controlling the form of the frequency versus temperature curves are derived by a method closely related to that of Mason (3518 of 1940). Design data are deduced from these and similar curves, and the relationships encountered in practice are discussed.

620.197 1966

The Climatization of Radio Equipment—M. A. (*Radio Tech. Dig.* (Franç.), vol. 3, pp. 115–120; April, 1949.) A note reviewing modern methods of protection against extremes of temperature, humidity, fungi and insects, together with a list of 70 recent references.

620.197:621.319.45 1967

On the Climatization of the Electrolytic Capacitor—H. E. Miquelis. (*Radio Tech. Dig.* (Franç.), vol. 3, pp. 77–82; April, 1949.) A few details are given concerning capacitors of several types; one type can withstand temperatures from −60° to +90°C in store or during transit and functions satisfactorily in the range from −40° to +70°C.

621.315.59:537.311.33:621.396.645 1968

Germanium—Important New Semiconductor—W. C. Dunlap, Jr. (*Gen. Elec. Rev.*, vol. 52, pp. 9–17; February, 1949.) An account of the chemical, physical, and electrical properties of Ge, and its application in the transistor.

621.315.59:621.319.4 1969

On the Use of Semiconducting Liquids for Impregnating Paper Capacitors—V. T. Renne. (*Zh. Tekh. Fiz.*, vol. 19, pp. 218–224; February, 1949. In Russian.)

621.315.61.011.5:546.431.82 1970

Anomalous Dielectric Properties of Polycrystalline Titanates of the Perovskite Type—J. R. Partington, G. V. Planer, and I. I. Boswell. (*Phil. Mag.*, vol. 40, pp. 157–175; February, 1949.)

621.315.612.011.5 1971

A New Type of Dielectric Polarization, and Losses in Polycrystalline Dielectrics—G. I. Skavani and A. I. Demeshina. (*Zh. Eksp. Teor. Fiz.*, vol. 19, pp. 3–17; January, 1949. In Russian.) Experiments were conducted with materials having a distorted crystal lattice. The materials were obtained by sintering TiO_2 with oxides of the second group of metals. Materials with small additions of SrO, CaO, BaO, and ZnO have a high dielectric constant (of the order of 1,000) at frequencies from 10 to 20 kc. The loss angle is greater than for pure TiO_2, and has pronounced frequency and temperature maxima. Variations of ϵ and tan δ with increase of concentration of the alkaline earths are shown graphically. The experimental results are in agreement with a theory of relaxation polarization developed in earlier papers. The activation energy of loosely coupled ions and the frequency of their oscillations, as calculated from the experimental data, have abnormally low values.

621.315.618.015.5:537.52 1972

Breakdown Voltage of Rare-Gas/Nitrogen Mixtures between Hot Electrodes—J. A. M. van Liempt and W. D. van Wijk. (*Physica, 's Grav.*, vol. 11, pp. 167–178; March, 1944. In German.) The dependence of the breakdown voltage on the nitrogen content, the gas pressure, and the temperature and separation of the electrodes is shown graphically. The results are discussed with reference to the design of gas-filled glow lamps.

621.775.7 1973

Powder Metallurgy—G. Fitzgerald-Lee. (*Electronic Eng.*, vol. 21, pp. 87–90; March, 1949.) The development of the art is outlined and basic principles are briefly discussed. Methods of controlling grain growth are considered with particular reference to the production of tungsten wire with properties suitable for lamp filaments. Various applications are mentioned.

666.1.037.5 1974

The Technique of Glass-to-Metal Sealing with Special Reference to Vacuum-Tight Seals—A. G. Long. (*Jour. Soc. Glass Tech.*, vol. 30, no. 137, pp. 67–89; 1946.) A general survey ranging from wire seals used in tube pinches to cylindrical seals some 4 inches in diameter, from the standpoint of the author's personal experience. Practical manufacturing details are discussed and illustrated.

666.1.037.5:621.385.032.5 1975

The Electrode Leads of Transmitting Valves—E. G. Dorgelo. (*Communication News*, vol. 9, pp. 38–44; December, 1947.) Survey and discussion of various methods of sealing.

MATHEMATICS

517.53:621.392.52 1976

Splitting of an Analytical Function into an Even and an Odd Component, and also a Method of Determining, for a Known Even Component, the Odd Component and the Function as a Whole—Correction to 1422 of June, the title of which should read as above.

518.5 1977

Multiplication and Division by Electronic-Analogue Methods—E. M. Deeley and D. M. MacKay. (*Nature* (London), vol. 163, p. 650; April 23, 1949.) A multiplier should have (a) symmetry of response to positive and negative inputs, (b) an absolute indication of zero input, (c) independence of normal changes in electronic characteristics, and (d) rapid and accurate response. A multiplier having a cro with an axial magnetic field is here described; the transverse velocity of the electron stream is proportional to the voltage V_x applied to the X-plates of the cro. The magnetic field H produces a deflecting force initially in the Y-direction, proportional to HV_x. This is counteracted by a photoelectric feedback system; the spot can thus be held close to the X axis, and the voltage V_y is then proportional to the product of V_x and the current i_H producing the field H. Conversely, the photo cell can be used to control H so that i_H is proportional to V_y/V_x. Preliminary experimental results with this technique are satisfactory.

518.5:512.25 1978

A Twelve-Equation Computing Instrument—C. F. Berry and J. C. Pemberton. (*Instruments*, vol. 19, pp. 396–398; July, 1946.) An instrument of the decade-potentiometer type for solving linear simultaneous equations by successive approximations.

518.5:517.944 1979

A Note on Analog Computer Design—J. V. Bronzo and H. G. Cohen. (*Rev. Sci. Instr.*, vol. 20, pp. 101–102; February, 1949.) In certain types of partial differential equations, the spatial derivatives may be replaced by finite difference expressions. By using Laplace transforms and matrices, the original equations can be reformulated as a set of linear simultaneous equations. A simpler procedure sometimes results from the application of a "similarity transformation" to the matrix of these equations before a computer is designed to solve them.

517.564.3(083.5) 1980

Tables of Bessel Functions of Fractional Order. Vol. 1 [Book Review]—National Bureau of Standards. Columbia University Press, 1st edition 1948, 418 pp. (*Phil. Mag.*, vol. 40, p. 124; January, 1949.) $J_\nu(x)$ is tabulated to 10 places of decimals for $\pm\nu = \frac{1}{4}, \frac{1}{3}, \frac{2}{3}$, and $\frac{3}{4}$, and $x = 0(0.001)0.9(0.01)25.0$. Auxiliary tables are provided to extend the range to values of x up to 30,000. See also 1708 of July.

518.2 1981

Practical Five-Figure Mathematical Tables [Book Review]—C. Attwood. Macmillan, London, 1948. 74 pp., 3s. (*Wireless Eng.*, vol. 25, p. 267; August, 1948; *Nature* (London), vol. 163, p. 306; February 26, 1949.) The problem of the regions where mean proportional parts are untrustworthy has been resolutely tackled by adjusting the interval of tabulation to suit the rate of change of the function tabulated. "The book . . . can be thoroughly recommended."

MEASUREMENTS AND TEST GEAR

531.761+621.3.018.4(083.74) 1982

Standard Frequency Broadcasts from Hawaii—(*Tech. Bull. Nat. Bur. Stand.*, vol. 33, pp. 39–40; March, 1949.) An experimental station on the island of Maui, Territory of Hawaii, now broadcasts standard frequencies (5, 10, and 15 Mc), time announcements, standard time intervals and standard musical pitch (440 cps) with call sign WWVH. Simultaneous reception of WWV and WWVH in certain areas should not cause interference.

531.761+621.3.018.4(083.74) 1983

The Atomic Clock—(*Tech. Bull. Nat. Bur. Stand.*, vol. 33, pp. 17–24; February, 1949.) The operation depends upon the constant natural frequency associated with the vibrations of atoms in the NH_3 molecule. Accuracy is within 1 part in 10^7; theoretical considerations indicate a potential accuracy within 1 part in 10^9 or 10^{10}, according to the type of atomic system and spectrum line used. The clock consists essentially of a 100 kc crystal oscillator, a frequency multiplier, a frequency discriminator, a frequency divider, a special 50-cps clock, and a waveguide absorption cell containing NH_3 at a pressure of 10 to 15 μ Hg. The fundamental frequency of the oscillator is first multiplied up to 270 Mc by means of standard low-frequency tubes. It is then multiplied up to 2,970 Mc by means of a frequency-multiplying klystron, which is also modulated by a FM oscillator generating a signal at 13.8±0.12 Mc. The FM output is multiplied in a Si crystal rectifier to 23,870.4±0.96 Mc and fed to the ammonia absorption cell. As the frequency of this modulated control signal sweeps across the absorption-line frequency (23,870.1 Mc) of the NH_3 vapor, the signal reaching the Si rectifier dips because of the absorption, giving a negative output pulse. The output of the FM oscillator at 13.8±0.12 Mc is also fed to a receiver together with a 12.5-Mc signal from the quartz-crystal multiplying chain. When the signal sweeps across the proper frequency (12.5 Mc+the 1.39 Mc if of the receiver) a second output pulse is generated. If the time interval between these two pulses is incorrect, a control signal is generated in a discriminator circuit, and fed to a reactance tube which forces the quartz-crystal circuit to oscillate at the correct frequency. See also *Radio-Electronics*, vol. 20, pp. 74–76; March, 1949; and 1712 of July (Huston and Lyons).

621.317.2:621.397.6 1984

TV Distribution System for Laboratory Use—J. Fisher. (*Communications*, vol. 29, pp. 8–9, 43; February, 1949.) Description of a centralized system for producing the standard RMA composite video signal and distributing it to a number of laboratories with minimum distortion. Sources of signal include (a) local television broadcasting stations, (b) monoscope signal, and (c) picture signal from a crt flying-spot scanner.

621.317.3+621.317.7]:061.3 1985

Papers Digested for Conference on High-Frequency Measurements—(*Elec. Eng.*, vol. 68, pp. 251–257; March, 1949.) Authors' summaries of most of the papers read at the conference.

621.317.3:621.395.44 1986

Maintenance Measurements on Carrier Telephony Equipment—J. de Jong. (*Philips Tech. Rev.*, vol. 8, pp. 249–256; August, 1946.) Discussion of apparatus used and operational requirements.

621.317.324†:621.318.4 1987

Coils as H.F. Measurement Probes for Absolute Field-Strength Determinations—E. Roeschen. (*Funk und Ton*, vol. 3, pp. 167–172; March, 1949.) Measurements with 5 different types of coil indicate that a small single-layer cylindrical coil is particularly suitable for such absolute measurements, since its effective area can be determined accurately from its mean geometrical diameter.

621.317.329:538.122 1988

[Electrolyte] Tank Model for Magnetic Problems of Axial Symmetry—R. E. Peierls and T. H. R. Skyrme. (*Phil. Mag.*, vol. 40, pp. 269–273; March, 1949.) Magnetic problems often involve vortices and cannot, therefore, be directly represented in the electrolyte tank. For 2-dimensional problems, the conjugate field can be studied; this technique is here applied to systems which are very nearly plane and have axial symmetry.

621.317.335.3†+621.317.374 1989

An Optical Method for Measuring the Dielectric Constant and Dielectric Losses of Solid Dielectrics in the Centimetre Wavelength Range—L. L. Odynets. (*Zh. Tekh. Fiz.*, vol. 19, pp. 120–125; January, 1949. In Russian.) Theoretical discussion of a method based on measurements of the transparency of a slab of the dielectric, with a description of the apparatus (Fig. 1) used for measuring the transparency of glass and ebonite for $\lambda=5.5$ cm. The method would be even more suitable for shorter wavelengths since the required over-all dimensions of the measuring apparatus would be smaller; liquid dielectrics could also be used.

621.317.34:621.315.1/.2 1990

No-Load and Short-Circuit Measurements for Determining the Transmission Characteristics of Open Lines and Cables—O. Naumann. (*Arch. Tech.* (Messen), pp. T22–T23; March, 1949.) The approximate methods used for calculating the low-frequency transmission characteristics of cables from no-load and short-circuit measurements of voltage, current, and power are not applicable at high frequency. More accurate methods given by Kaden (1063 of 1937), by Sommer for very low frequencies (4070 of 1939), and by Goldschmidt for high-frequency (1724 of 1943) are discussed.

621.317.353:621.396.619.13 1991

The Determination of the Distortion in a Frequency-Modulator—F. L. H. M. Stumpers and W. W. Boelens. (*Communication News*, vol. 9, pp. 107–109; August, 1948.) As the modulation voltage is increased, the amplitudes of the components of the output frequency spectrum vary and pass through the value zero. The zero points can be used for measuring frequency deviation. The displacement of the zeros, caused by nonlinearity of the modulation characteristic, can be used for the measurement of odd harmonics. A shift in frequency of the output spectrum as a whole can be used to measure even harmonics.

621.317.372 1992

Improved Accuracy with a "Q"-Meter by the Use of Auxiliary Components—A. C. Lynch. (*Electronic Eng.* (London), vol. 21, pp. 91–93; March, 1949.) Accuracy can be considerably increased by using a mirror galvanometer in parallel with the meter giving Q values, a second galvanometer, suitably shunted, in series with the meter measuring the resonant-circuit response and a variable capacitor with a fine scale of high calibration accuracy in parallel with the fitted tuning capacitor. The principle used is that of the reactance-variation method described by Hartshorn and Ward (351 of 1937), but here the input voltage is varied in known ratios, and the voltage in the circuit is brought to a fixed value by detuning. A circuit diagram is given. A Q-meter using this principle is described; accuracy is discussed, and possible methods of further improvement are suggested.

621.317.39.029.6:536.33 1993

Conditions for Maximum Sensitivity of U.H.F. Radiometers—J. L. Steinberg. (*Onde Élec.*, vol. 29, pp. 160–166; April, 1949.) Discussion of apparatus of the type described by Dicke (475 of 1947). With optimum noise factor and input circuit, and a certain amount of negative feedback, it is possible to measure the apparent temperature of a receiving antenna to within about 0.3°C. Apparatus designed for investigation of 1,200-Mc radiation, constructed at the physical laboratory of the École Normale Supérieure, uses a dipole-fed parabolic reflector 3 meters in diameter mounted on an equatorial support. The superheterodyne receiver has a bandwidth of 20 Mc and a noise factor of 3.5. Further details of this equipment will be published later.

621.317.4.029.62:538.221 1994

Ferromagnetism at Very High Frequencies; Part 2—Method of Measurement and Processes of Magnetization—M. H. Johnson and G. T. Rado. (*Phys. Rev.*, vol. 75, pp. 841–864; March 1, 1949.) Part 1: 3182 of 1947 (Johnson, Rado, and Maloof.)

621.317.7.029.63/.64(083.74) 1995

Microwave Measurement Standards—(*Jour. Frank. Inst.*, vol. 247, pp. 156–161; February, 1949.) Discussion of standards and calibration services available or methods being developed at the National Bureau of Standards for frequencies from 300 Mc to over 100,000 Mc. These include: (a) a frequency standard for 300 to 40,000 Mc accurate within 1 part in 10^8, (b) methods and equipment for dielectric measurements at frequencies near 1,000, 3,000, 9,000, and 24,000 Mc, (c) a primary standard of attenuation consisting of a waveguide operating at a frequency below cutoff, which can be used for any microwave frequency by means of a development of the if substitution method, (d) a microwave power-measuring console nearly completed, (e) investigation of the accuracy of bolometers and thermal noise sources as power standards, (f) development of primary and secondary frequency standards using spectrum lines of gases, (g) a prototype atomic clock (see also 1983 above), and (h) measurement of spectrum lines as secondary frequency standards.

621.317.71/.72 1996

An Instrument for the Measurement and Time Integration of Small Voltages and Currents—I. A. D. Lewis and A. C. Clark. (*Jour. Sci. Instr.*, vol. 26, pp. 80–84; March, 1949.) Description with full component details. Potentials varying between 0 and 0.5 volt and their time integrals can be measured within 3 per cent. Currents can be measured with similar accuracy; full-scale deflection on the lowest range is 0.01 μamp.

621.317.71:621.385 1997

Reduction of Noise in Thermionic Electrometers with Mechanical Conversion—H. den Hartog and F. A. Muller. (*Physica,'s Grav.*, vol. 11, pp. 161–166; March, 1944. In Dutch, with English summary.) Continuation of 858 of 1945.

621.317.715.004.64:538.22 1998

Non-Ferrous Copper Wire for Moving-Coil Meters—P. G. Moerel and A. Rademakers. (*Philips Tech. Rev.*, vol. 8, pp. 315–319; October, 1946.) Effects due to Fe impurity in the Cu wire of the coils is discussed and methods of producing wire with extremely low Fe content are described.

621.317.738 1999

A Note on the Measurement of Four-Terminal Inductances by Astbury's Method—L. H. Ford. (*Jour. Sci. Instr.*, vol. 26, pp. 108–109; March, 1949.) Astbury's method (noted in 1932 Abstracts, p. 48) is unsuitable in its original form for frequencies above af, but if the fixed resistance ratio arms of the bridge are replaced by a Kelvin-Varley potential divider of known phase defect, the method can be used for frequencies up to 100 kc. Experimental results on a 10-μh coil show good agreement at all frequencies from 1 to 100 kc with values obtained by other methods.

621.317.755 2000

A 3-Beam Micro-Oscillograph for Display of Oscillations up to 10,000 Mc/s—v. Fe. (*Frequenz*, vol. 3, pp. 19–22; January, 1949.) A short description of Lee's instrument (1692 of 1946) with a tabular comparison with the Rogowski, Siemens, and A.E.G. oscillographs.

621.317.761 2001

Heterodyne Frequency Meter for High Frequencies—L. Liot. (*Radio Franç.*, pp. 14–17; April, 1949.) Description, with circuit details, of an instrument for the rapid measurement of frequencies from 5 to 1,000 Mc. The local oscillator, which is of the Lecher-line type and uses a Type 955 triode, has a frequency range of 69 to 220 Mc; within this range unknown frequencies are measured by a direct zero-beat method, while harmonic methods are used outside the range.

621.317.772.029.54/.58:621.396.645.37 2002

A Phase Meter for the Frequency Band 100 kc/s–20 Mc/s—W. T. Duerdoth. (*P.O. Elec. Eng. Jour.*, vol. 42, part 1, pp. 43–46; April, 1949.) Intended primarily for measuring the phase change round the feedback loops of amplifiers. An accuracy within ±3° is possible provided that the loop gain or loss does not exceed 40 db. The magnitudes of the voltages whose phases are to be compared are made equal by means of two special variable-gain amplifiers which cause the same phase change in received signals whatever their respective gains. The design of these amplifiers and of the necessary frequency changer is discussed and circuit diagrams are given.

621.317.784 2003

Pulse Power Measurement by a Heterodyne Method—L. S. Schwartz. (*Communications*, vol. 29, pp. 26–27; February, 1949.) Pulse width and repetition rate need not be known. The rf pulse and cw oscillations of nearly the same frequency are applied to the square-law detector of a synchroscope receiver. The cw is adjusted so that the peak value of the variable component of the detection voltage equals the amplitude of the envelope of the rf pulse. The rf pulse power is then 6 db above the cw power at the point of entrance into the detector. Sources of error are discussed; accuracy is within a few per cent.

621.317.79:551.510.535 2004

Sweep Frequency Ionosphere Equipment—P. G. Sulzer. (*Jour. Appl. Phys.*, vol. 20, pp. 187–196; February, 1949.) The equipment records ionosphere virtual height as a function of frequency over the range 1 to 25 Mc. Special features of the device are high power output, good receiver sensitivity, and anti-jamming circuits. Detailed circuit diagrams are included. See also 2240 of 1948 (Thomas and Chalmers).

621.317.791 2005

A Self-Checking Wobbulator—J. H. Vogelman. (*Communications*, vol. 29, pp. 28–31; February, 1949.) A portable test set for frequencies between 5 and 100 Mc, comprising a FM signal generator, frequency meter, and cro which can be used to measure the gain, frequency, bandwidth, and tuning characteristics of wide-band if and rf amplifiers and receivers.

621.317.35+621.396.615.17/.18 2006

Waveforms [Book Review]—B. Chance, V. Hughes, E. F. MacNichol, D. Sayre, and F. C. Williams (Eds). McGraw-Hill, London, 785 pp., £3. (*Wireless Eng.*, vol. 26, p. 139; April, 1949.) Vol. 19 of the MIT Radiation Laboratory series. The book is more descriptive than analytic and deals mainly with nonlinear circuits. The action of electronic switches, frequency multipliers and dividers, and counting circuits is discussed.

621.317.35+621.396.619+621.396.822 2007

Frequency Analysis, Modulation and Noise [Book Review]—S. Goldman. McGraw-Hill, London, 1948, 434 pp., 36s. (*Electronic Eng.* (London), vol. 21, p. 152; April, 1949; PROC. I.R.E., vol. 37, p. 541; May, 1949.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

535.61-15:621.383 2008

Direct Recording of Spectra in the Region 1.2 μ to 3 μ using the Lead Sulfide Photo-Conductive Cell—R. C. Nelson. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 68–71; January, 1949.) See also 3330 of 1947 (Cashman).

621.317.39 2009

Electronic Gauges—J. Schwartz. (*Microtecnic* (Lausanne), vol. 3, pp. 10–18; January and February, 1949. In English.) Conclusion of 1736 of July.

621.365.5+621.365.92 2010

High-Frequency Heating—S. W. Scherer. (*Communication News*, vol. 9, pp. 45–55, December, 1947.) Discussion of both induction and dielectric heating, and of their applications. Several examples are given of apparatus designed for particular applications.

621.38.001.8 2011

The Electronic Brain—W. R. Ashby. (*Radio-Electronics*, vol. 20, pp. 77–80; March, 1949.) Reprint of 1144 of May.

621.38.001.8 2012

Electronics in the Service of Industry—(*Radio Tech. Dig.* (Franç.), vol. 3, pp. 105–111; April, 1949. Bibliography, pp. 111–113.) Brief general discussion, with some details of photo electric counters and opacity meters.

621.38.001.8:061.3 2013

Papers Digested for Conference on Electronic Instrumentation—(*Elec. Eng.*, vol. 68, pp. 246–251; March, 1949.) Authors' summaries of most of the papers read at the conference.

621.384.611.1† 2014

A 20-MeV Betatron—W. Bosley, J. D. Craggs, D. H. McEwan, and J. F. Smee. (*Proc. IEE*, part I, vol. 96, pp. 85–86; March, 1949.) Discussion on 179 of February.

621.384.611.1† 2015

A New Type of [9-MeV] Betatron without an Iron Yoke—A. Bierman. (*Nature* (London), vol. 163, pp. 649–650; April 23, 1949.) Two coils are used in series, with a sealed-off glass acceleration tube between them. The dimensions are chosen so that the magnetic field at a point distant r from the center is proportional to r^{-n}, where $0<n<1$. The flux required within the electron orbit is obtained by means of a small iron core along the axis. The current through the coils is obtained by periodical discharges of a 6.5-$\mu\mu$f capacitor across a spark gap; these discharges occur every few seconds and initiate damped 2.5 kc oscillations, with a peak current of about 5,000 amperes. The whole betatron only weighs about 50 kg and the simple construction makes the cost low.

621.384.612.1† 2016

Design of the Radiofrequency System for the 184-Inch Cyclotron—K. R. MacKenzie, F. H. Schmidt, J. R. Woodyard, and L. F. Wouters. (*Rev. Sci. Instr.*, vol. 20, pp. 126–133; February, 1949.) See also 1712 of 1948.

621.384.612.1†:621.396.615 2017

Pulsed Oscillator for F.M. Cyclotron—J. W. Burkig, E. L. Hubbard and K. R. MacKenzie. (*Rev. Sci. Instr.*, vol. 20, p. 135; February, 1949.)

621.384.621.1† 2018

The High-Voltage Electrostatic Generator at the Atomic Energy Research Establishment—R. L. Fortescue and P. D. Hall. (*Proc. IEE*, part I, vol. 96, pp. 77–85; March, 1949.) A 5-Mv generator of the pressurized Van de Graaf type.

621.385.833 2019

Aberration Correction with Electron Mirrors—E. G. Ramberg. (*Jour. Appl. Phys.*, vol. 20, pp. 183–186; February, 1949.) Formulas for spherical and chromatic aberration are applied to a concave electron mirror with concentrated field distribution. The aberration coefficients of such mirrors are so large that this method of correction has serious practical difficulties.

621.396.615.17:615.849 2020

Electromedical Stimulators—O. B. Sneath and E. G. Mayer. (*Wireless World*, vol. 55, pp. 129–132; April, 1949.) Pulses are required of lengths between 1 second and 10 μseconds, with repetition rates between 1 and 50 per second and maximum output voltage of the order of 100 volts at 100 mamp. Circuits for producing such pulses are discussed.

621.398 2021

Radio Control of Mobile Miniatures—E. L. Safford, Jr. (*CQ*, vol. 5, pp. 18–21, 71; April, 1949.) A system for starting, stopping, and steering model boats or cars.

PROPAGATION OF WAVES

538.566 2022

Unification of the Formulae Representing the Principle of Huyghens for Electromagnetic Waves—F. Croze and G. Darmois. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 824–826; March 7, 1949.) Three conditions must be satisfied if a system of formulas is to be a physically and mathematically correct expression of Huyghens' principle. The formulas proposed successively by Love, Macdonald, Larmor, Bromwich, Schelkunoff, L. de Broglie, Novobatzky (for a finite surface), and Franz all satisfy these three conditions; in consequence, they can all be referred to a common form, which is here given. The condition that the secondary waves from the various elements of the surface considered must be pure electromagnetic waves is not satisfied by the formulas of Kirchhoff, nor by those proposed successively by Ignatowsky, Tonolo, Tedone, Kottler, Stratton, and Chu and (for a closed surface) Novobatzky. This is shown by considering the case where the surface in question is a wave surface, either plane or of large radius. See also 1462 and 1463 of June.

538.566.2 2023

On the Propagation of Waves in an Inhomogeneous Medium—O. E. H. Rydbeck. (*Chalmers Tekn. Hogsk. Handl.*, no. 74, 35 pp. 1948. In English.) A mathematical theory of propagation in an inhomogeneous or stratified medium is developed to determine the limiting conditions in which more approximate theories are applicable. For a slightly inhomogeneous medium, first- and higher order approximations to the wave equations are obtained and their usefulness is estimated by application to special cases for which exact independent solutions exist. Large variations in refractive index are then considered. The results are compared with those of other methods. The theory is applied to the propagation of magneto-hydrodynamic waves in the sun and to the duct propagation of radio waves in the lower troposphere.

621.396.11 2024

The Effect of Pulsations of the Refractive Index of the Atmosphere on the Propagation of Ultra-Short Waves—V. A. Krasil'nikov. (*Bull. Acad. Sci. U.R.S.S., sér. Géogr. Géophys.*, vol. 13, no. 1, pp. 33–57; 1949. In Russian.) The usual interpretation of fading as a result of interference between the direct and indirect rays cannot be strictly correct, especially when transmission extends beyond the horizon. It is suggested that pulsations of the refractive index of the atmosphere due to temperature fluctuations may have an important bearing on fading. It is shown mathematically that when the scale of pulsations is smaller than the wavelength, the waves are dispersed and when it is greater than the wavelength, path variations occur which lead to variations in intensity at the receiving point. Using the methods of geometrical optics and taking into account the results obtained by A. M. Obukhov (1928 above) formulas are derived for determining fluctuations of the amplitude and phase at the point of reception. Experimental results available in the literature are discussed in detail and are in general agreement with the proposed theory. It is concluded that: (a) when transmission is within the horizon, usw fading can be explained by the temperature fluctuations of the atmosphere; (b) fading when transmission is somewhat beyond the horizon is mainly caused by the interference between the two components but it is also necessary to take into account the effect of the temperature fluctuations on both components; (c) at very great distances, when reception is carried out only on the indirect ray, the temperature-fluctuation effect again becomes predominant.

621.396.11:551.5 2025
An Extension of Macfarlane's Method of Deducing Refractive Index from Radio Observations—A. W. Straiton. (*Jour. Appl. Phys.*, vol. 20, p. 228; February, 1949.) Comment on 2894 of 1947. The second set of height-gain measurements there suggested can be replaced by a set of height versus phase measurements; a suitable method of measuring phase was discussed in 3225 of 1948 (Straiton and Gerhardt).

621.396.11:551.510.535 2026
Absorption of Radio Waves Reflected at Vertical Incidence as a Function of the Sun's Zenith Angle—E. W. Taylor. (*Jour. Res. Nat. Bur. Stand.*, vol. 41, pp. 575–579; December, 1948.) Analysis of Central Radio Propagation Laboratory records, extending over 3 years, indicates that absorption depends approximately linearly on the cosine of the sun's zenith angle.

621.396.11:551.510.535 2027
A Note on the Ionospheric Absorption Problem—L. G. McCracken. (*Jour. Appl. Phys.*, vol. 20, pp. 229–230; February, 1949.) The formula for the total nondeviating E-layer absorption derived by Best and Ratcliffe (1748 of 1938) is here obtained without using certain of their approximations. Appleton (395 of 1938) obtained a similar formula.

621.396.11:551.510.535 2028
The Ionosphere over Mid-Germany in February 1949—Dieminger. (*See* 1934.)

621.396.11:551.510.535 2029
Changes in Radio Reception during Sunspot Period '45–47—H. T. Stetson. (*Tele-Tech*, vol. 7, pp. 29, 73; December, 1948.) Summary of Amer. Astr. Soc. paper. Variations of F_2 critical frequencies and of field intensities at 5 and 10 Mc with variations of sunspot activity are recorded for the recent rise period. Diurnal field-intensity changes to be expected during the decline period are discussed.

621.396.11:551.510.535:523.3 2030
Moon Echoes and Penetration of the Ionosphere—F. J. Kerr, C. A. Shain, and C. S. Higgins. (*Nature* (London), vol. 163, pp. 310–313; February 26, 1949.) The possibility of using reflections from the moon to extend the study of the ionosphere is investigated. Stations VLC9 and VLB5 operating at frequencies near 20 Mc were used. The equipment and the experimental procedure are described. Echoes were obtained in 13 out of 15 trials. Wide variations in the amplitude of echoes were recorded; on no occasion did the amplitude exceed the theoretical value. Fading periods of 1 second or less were observed, similar to those experienced with ionospheric reflections. Echo frequencies were about 50 cps higher than those of the transmitted signal, mainly because of the Doppler effect of the earth's rotation. As the moon rose, echoes could not be obtained as soon as could be expected from calculations based on vertical-incidence ionospheric soundings.

Analysis of 20-Mc solar noise also shows ionospheric effects, with the significant difference that solar noise energy is received at unexpectedly low elevations of the sun.

621.396.812.029.62 2031
On Normal and Abnormal Refraction of Ultra-Short Waves in the Atmosphere—H. H. Klinger. (*Z. Met.*, vol. 2, p. 86; March, 1948.) Field-strength records of usw transmissions ($\lambda = 1.3$ meters) over a 63-km path, about 40 per cent beyond the optical range, are discussed briefly. The considerable increases of field strength occasionally observed are accompanied by relatively little fluctuation and can be attributed to certain local characteristics of the lower atmosphere.

621.396.812.029.64 2032
Effect of the Atmosphere on Microwaves—H. H. Klinger. (*Z. Met.*, vol. 2, pp. 314–316; October, 1948.) Measurements of received signal strength for $\lambda = 1.36$ cm showed strong absorption due to rain or even a slightly damp atmosphere, whereas 3.6-cm waves under the same conditions were not appreciably affected. Further investigations with wavelengths of 1 to 10 mm are proposed.

621.396.812.029.64 2033
Low-Level Atmospheric Ducts—R. F. Jones, J. S. McPetrie and B. Starnecki. (*Nature* (London), vol. 163, p. 639; April 23, 1949.) Comment on 1167 of May. Jones explains the presence of the ducts observed both in cold and warm weather in terms of the previous history of the air at heights between 200 and 2,000 feet. McPetrie and Starnecki consider that although Jones' hypothesis is satisfactory for periods of high wind-velocity, low air temperature rather than high wind-velocity is the main cause of the cold-weather ducts.

621.396.812.029.64 2034
Oversea Propagation on Wavelengths of 3 and 9 centimeters—J. S. McPetrie, B. Starnecki, H. Jarkowski, and L. Sicinski. (PROC. I.R.E., vol. 37, pp. 243–257; March, 1949.) For other accounts of results obtained at the same sites see 518 of 1947 (Megaw) and 2329 of 1948 (McPetrie and Starnecki).

621.396.812.3 2035
A Peculiar Type of Rapid Fading in Radio Reception—N. S. Subba Rao and Y. V. Somayajulu. (*Nature* (London), vol. 163, p. 442; March 19, 1949.) Discussion of the "flutter phenomenon," a variation in intensity observed at Waltair, India, when receiving broadcast transmissions on wavelengths of 41 and 60 meters. The effect is observed occasionally on a wavelength of 19 meters but is completely absent at medium frequency. The flutter occurs only during the hot season (February to June) and begins about sunset; the fluctuation frequency is very low (0.3 to 0.5 cps) before sunset, and increases to 2 to 2.4 cps early in the night, thereafter remaining constant for long periods. The ionosphere must be the cause of this phenomenon, which may be due to the rapid movement of ionic clouds across the F layer. See 3279 of 1946 (Wells, Watts, and George).

621.396.812.3:551.510.535 2036
Short-Range Fading of Broadcasting Transmissions—W. Gerber and A. Werthmüller. (*Tech. Mitt. Schweiz. Telegr. Teleph. Verw.*, vol. 25, pp. 1–12; February 1, 1947. In German.) The reflecting properties of the ionosphere in the medium-wave band are discussed. The fine structure of the fading diagram is attributed to interference effects, while the main features are principally determined by absorption and are related both to the sunspot period and to the season, with maximum fading effects in spring and autumn and minimum occurrence in summer and winter. The solar effects are superposed on the seasonal effects and are greatest at sunspot minimum and least at sunspot maximum. Curves are given showing the average fading effects for the Beromünster transmitter as observed at St. Gallen, at different times during the evening, from April, 1936, to November, 1946. The curves all show maxima about 1944, with subsidiary peaks about 1941 and minima near 1938.

621.396.812.3.029.54 2037
Space-Wave Absorption and Large-Scale Weather Conditions—G. Falckenberg and E. Lauter. (*Z. Met.*, vol. 2, pp. 259–265; September, 1948. Transmissions from Kalundborg on $\lambda = 1{,}250$ meters were observed at Warnemünde 180 km distant. In the daytime, the transmissions were almost completely absorbed by the ions in the lower part of the ionosphere, but at night the absorption showed wide variations related to projections into the ionosphere of equatorial or polar air masses. A connection between the absorption and air-pressure variations at a height of 9 km was also established.

RECEPTION

621.396.619.13:621.392 2038
Distortion of Frequency-Modulated Signals in Electrical Networks—F. L. H. M. Stumpers. (*Communication News*, vol. 9, pp. 82–92; April, 1948.) Long summary of part of the thesis noted in 2221 of 1947. See also 1886 of 1948.

621.396.621+621.396.61]:551.510.535 2039
Ionosphere Sounding [equipment]—Maguer. (*See* 2069.)

621.396.621:621.392.52:621.396.5 2040
Application of Crystal Filters in [telegraphy] Receivers—Maarleveld. (*See* 1892.)

621.396.621:621.396.5 2041
A Modern Receiver for Radiotelegraphy—C. T. F. van der Wyck. (*Tijdschr. ned. Radiogenoot.*, vol. 14, pp. 27–39; March, 1949. Discussion, p. 40. In Dutch, with English summary.) Description, with block diagram, of the receiver, and discussion of (a) considerations leading to its design, (b) automatic tuning control, and (c) conditions for a stable circuit. See also 3510 of 1948 and 1892 above.

621.396.81:621.396.9 2042
Signal/Noise Ratio in Radar—S. de Walden. (*Wireless Eng.*, vol. 26, pp. 140–141; April, 1949. Comment on 2899 of 1948 (Levy).

621.396.826:621.396.933.2:621.396.619.16 2043
Elimination of Reflected Signal Effects in Pulsed Systems—Collup. (*See* 1945.)

621.396.822+621.317.35+621.396.619 2044
Frequency Analysis, Modulation and Noise [Book Review]—Goldman. (*See* 2007.)

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 2045
Evaluation of Transmission Efficiency According to Hartley's Expression of Information Content—A. G. Clavier. (*Elec. Commun.*, vol. 25, pp. 414–420; December, 1948.) Transmission efficiency, defined by extending Hartley's expression for telegraphic signals to telephony in presence of noise, is calculated for the main pulse-transmission systems. The resulting expressions enable the various systems to be compared, although simplicity and cost of equipment will also be important factors for deciding their relative merits. See also 515 of March (Landon) and 1361 of June (Shannon).

621.395.43/.44 2046
Study of the General Characteristics of the L. T. T. 12-Channel Carrier-Current Systems for Overhead Lines—H. Pech. (*Câbles and Trans.* (Paris), vol. 3, pp. 177–193; April, 1949.) General problems in connection with such systems are discussed, including (a) attenuation within the prescribed transmission band, (b) characteristic impedance of actual lines, (c) crosstalk, and (d) noise. A description is given of equipment constructed for the French Post Office. The frequency band covered is 30 to 150 kc. Details are included of the modulation and demodulation processes, terminal equipment, filters, equalizers and carrier-frequency generators. Test results on prototype equipment will be published later.

621.396.1:621.397.5 2047
Allocation of Frequencies for the Television Service F. C. McL. (*BBC Quart*, vol. 4, pp. 54–56; April, 1949.) The frequency band at present allocated is 41 to 66.5 Mc; it is hoped that the upper limit will shortly be raised to 68 Mc. For all new stations, asymmetric-sideband transmission will be used to enable 5 exclusive channels to be obtained in the full band, but transmissions from Alexandra Palace will remain unchanged.

621.396.619.13/.14:534.78 2048
Ratio of Frequency Swing to Phase Swing in Phase- and Frequency-Modulation Systems Transmitting Speech—D. K. Gannett and W. R. Young. (Proc. I.R.E., vol. 37, pp. 258–263; March, 1949.) Theoretical and experimental results are discussed. The ratio was found to vary with different voices, with the microphone and circuit characteristics, and with the kind of volume regulation used.

621.396.65.029.63 2049
Choice of Suitable Heights [of stations], Distance and Wavelengths in Planning Decimetre-Wave Links—A. Grün. (*Fernmeldetech. Z.*, vol. 2, pp. 69–72; March, 1949.) In the case of decimeter wave systems, fading is due chiefly to interference between the direct ray and that reflected from an intermediate point of the earth's surface. Calculations show that in order to obtain as great a received field strength as possible, with little fading, the wavelength for a given distance between stations should not be too small and optimum heights should be chosen for the antennas. Formulas are given from which the optimum heights can be calculated, with curves for ranges of 50 and 60 km respectively and wavelengths from 10 to 100 cm.

621.396.712.2 2050
WMGM Master Control Equipment Design—M. E. Gunn. (*Audio Eng.*, vol. 33, pp. 24–28, 40; March, 1949.) Details, including block diagram, of a high-power broadcast station installation.

621.396.9:523.3 2051
Detection of Radio Signals Reflected from the Moon—DeWitt and Stodola. (*See* 1937.)

621.396.619+621.317.35+621.396.822 2052
Frequency Analysis, Modulation and Noise [Book Review]—Goldman. (*See* 2007.)

SUBSIDIARY APPARATUS

621.316.722.077.65:621.396.68 2053
Commutation in Rectifiers Using Relay Tubes: Parts 1 and 2—T. Douma. (*Communication News*, vol. 9, pp. 70–81 and 110–120; April and August, 1948.) A detailed discussion of the behavior of the well-known 3-phase Graetz circuit, taking account of the effect of leakage inductance and self-capacitance in the power transformer. Means of protecting this transformer from steep-fronted voltage surges are also indicated. The control characteristic of grid-controlled rectifiers is discussed in an appendix.

621.316.726 2054
Frequency Correction Equipment for Railway Signalling Supplies—(*Engineer* (London), vol. 187, pp. 184–186; February 18, 1949.) A 10-kva induction motor has a 440-volt three-phase stator and a single-phase rotor, the speed of which is controlled by a frequency-selective relay so that the rotor output frequency is between 49.7 and 50.3 cps even though the frequency of the stator input varies from 47.5 to 51.4 cps. An electronic frequency corrector is also described which rectifies the variable-frequency supply and reconverts the dc into ac at the required frequency by means of inverting tubes and a timing unit. Both kinds of equipment are operated by remote automatic control. See also *Engineering* (London), vol. 167, pp. 331–332; April 8, 1949.

TELEVISION AND PHOTOTELEGRAPHY

621.397.26 2055
Ultrafax—D. S. Bond and V. J. Duke. (*RCA Rev.*, vol. 10, pp. 99–115; March, 1949.) See also 1203 of May.

621.397.5 2056
Polycast System for TV on U.H.F. R. M. Wilmotte and P. A. Demars (*FM-TV*, vol. 8, pp. 26–28, 46; December, 1948.) For the United States uhf television band, the power required to obtain satisfactory coverage at all points within say 30 miles of a single transmitter is prohibitive. "Polycasting" is a suggested alternative, using ten to fifteen 200 watt transmitters within the area, each covering an area of 10 miles radius.

621.397.5:535.88:791 2057
Theater Television—(*Jour. Soc. Mot. Pic. Eng.*, vol. 52, pp. 243–267; March, 1949. Bibliography, pp. 268–272.) Report, in language as little technical as possible, of the Theater Television Committee of the Society of Motion Picture Engineers on the present state of the art.

621.397.5:621.3.09 2058
Attenuation and Phase Distortion and Their Effect on Television Signals Fuchs and Baranov. (*See* 1871.)

621.397.5:621.315.212 2059
The London-Birmingham Television Cable: Part 2—Cable Design, Construction and Test Results—Stanesby and Weston. (*See* 1857.)

621.397.5:621.396.1 2060
Allocation of Frequencies for the Television Service—McL. (*See* 2047.)

621.397.6:621.385.832 2061
The Graphechon—A Picture Storage Tube—L. Pensak. (*RCA Rev.*, vol. 10, pp. 59–73; March, 1949.) 1949 IRE National Convention paper. Describes the combination, within one envelope, of a cathode-ray tube and an iconoscope, enabling a picture written once on the common screen to be scanned continuously for 1 to 2 minutes. The picture is recorded on the screen by use of the fact that the thin insulating layer on the iconoscope target becomes conductive at the point of impact of a high-velocity electron beam. The picture is read by the conventional iconoscope method. The graphechon was designed originally in connection with the teleran navigation system.

621.397.6:621.395.667 2062
Phase and Amplitude Equalizer for Television Use—Goodale and Kennedy (*See* 1893.)

621.397.61:621.3.015.3 2063
Standardization of the Transient Response of Television Transmitters—R. D. Kell and G. L. Fredendall. (*RCA Rev.*, vol. 10, pp. 17–34; March, 1949.) It is suggested that picture-monitoring receivers for vestigial-sideband television transmitters should now be standardized on the basis of response to a double-sideband signal modulated by a square wave. Tolerances for vestigial-sideband transmitter performance could then be defined by the shape of the monitor receiver response when the transmitter is modulated by a square wave. Phase-correction networks for reducing the distortion introduced by the vestigial-sideband system and by the restricted bandwidth used are also discussed.

621.397.61-182.3 2064
The WMAL TV Mobil TV Unit—F. W. Harvey and E. D. Hilburn. (*Communications*, vol. 29, pp. 8–11, 31; March, 1949.) An illustrated description.

621.397.62 2065
Experimental Television Receiver (*Telev. (Franç.)*, no. 45, pp. 13–16, 21; March, 1949.) Circuit diagrams and complete details of a receiver using a small electrostatic cathode-ray tube with a screen only 7 cm in diameter, and capable of receiving either the present 450-line or the proposed 819-line transmissions.

621.397.62:621.396.662 2066
Simplified TV Receiver Channel Switching Mechanism J. A. Hansen (*Tele-Tech*, vol. 7, pp. 36–38, 72; December, 1948.) A compact 12-channel superheterodyne tuner which uses a sliding carriage to carry the tuning elements. The sound if is about 21.75 Mc and the vision if 26.25 Mc. Performance is discussed.

621.397.743(73) 2067
Television in 1949—Stations and Networks [in the U.S.A.]—(*Tele-Tech*, vol. 7, December, 1948. Supplement.) Map and brief tabulated data of stations now existing or under construction.

621.397.823:629.135 2068
Television Interference by Aircraft—A. H. Cooper. (*Wireless World*, vol. 55, pp. 142–145; April, 1949.) The use of directive receiving antennas reduces the region within which an aircraft causes interference, but a greater all-round reduction of interference is usually obtained by raising the receiving antenna. Removal of the dc component from the received signal eliminates fluctuations in picture brightness; the changes in contrast due to the interference are not then so noticeable. The design of a filter to attenuate frequencies between 1 cps and 25 cps, to mitigate some disadvantages of removing the dc component, is discussed.

TRANSMISSION

621.396.61+621.396.621]:551.510.535 2069
Ionosphere Sounding [equipment] P. Maguer. (*Radio Franç.*, pp. 7–12; April, 1949.) Description, with complete circuit diagrams, of (a) a transmitter giving 50 μsecond pulses with peak power of 1.6 kw and recurrence frequency of 50 per second, and (b) a wide-band superheterodyne receiver with good sensitivity, in which a cro sweep circuit of simple design gives a height-scale accuracy approximating to that obtained with a crystal oscillator and its somewhat complex frequency-division circuits.

621.396.61:621.316.726 2070
Frequency Control in Transmitters—H. B. R. Boosman and F. H. Hugenholtz. (*Communication News*, vol. 9, pp. 21–32; September, 1947.) Discussion of various methods of obtaining high stability of frequency and accuracy of tuning, including methods involving the principles of frequency-adjustment control and decade tuning. A number of discriminator circuits and circuits suitable for synchronization with a high harmonic of a control frequency are also described.

621.396.61:621.316.726 2071
Telesynchronization with Standard Frequency—L. Rohde and R. Leonhardt. (*Fernmeldetech. Z.*, vol. 2, pp. 85–90; March, 1949.) Detailed discussion of methods of phase and frequency synchronization, with particular reference to the control of common-wave broadcasting transmitters. Methods of deriving a control voltage from the phase or frequency difference between the standard and the local generator are described. The case where the ratio of the standard and local frequencies is that of two reasonably small integers is also considered.

VACUUM TUBES AND THERMIONICS

621.383 2072
Lead Selenide Photoconductive Cells—C. J. Milner and B. N. Watts. (*Nature* (London),

vol. 163, p. 322; February 26, 1949.) Cells prepared by an experimental chemical deposition process are found to have a broad sensitivity maximum at 3 to 4 μ. Appreciable sensitivity has so far only been exhibited at low temperatures.

621.383 2073
Bismuth Sulphide Photocells—B. T. Kolomiets. (*Zh. Tekh. Fiz.*, vol. 19, pp. 126–131; January, 1949. In Russian.) The photo cells are made with synthetic Bi_2S_3. Various experimental characteristics are plotted. This type of cell is particularly suitable for automatic control applications.

621.383 2074
Photoelectric Multipliers—S. Rodda. (*Jour. Sci. Instr.*, vol. 26, pp. 65–70; March, 1949.) Discussion of fundamental principles, emissive materials, electrode shapes, current fluctuations, dark current, fatigue phenomena, practical circuits, applications, etc.

621.383:535.215 2075
On the Interpretation of Observations on the Photoelectric Voltages with Intermittent Light—Gorter, Broer, and Snoek. (*See* 1916.)

621.383:535.215 2076
On Photoelectric Voltages in Light-Absorbing Materials—Snoek and Gorter. (*See* 1915.)

621.383:621.396.645 2077
The Development of a Photoelectric A.C. Amplifier with A.C. Galvanometer—Milatz and Bloembergen. (*See* 1908.)

621.385 2078
Standard Valves [Book Review]—Standard Telephones and Cables, 1947, 328 pp., 15s. (*Nature* (London), vol. 163, p. 387; March 12, 1949.) Details and technical data of ordinary tubes, cathode-ray tubes, cold-cathode relays, v.m. tubes and disk-seal uhf triodes manufactured by the above firm. Brimar tubes are not included.

621.385 2079
New Miniature American-Type Valves Made in France—M. Leroux. (*Radio Prof.* (Paris), vol. 18, pp. 6–9 and 18–20, 23; March and April, 1949.) Full technical details, with operational characteristics and practical circuits. The tubes include pentodes 6AG5, 6AU5, 6BA6, 6AK5, output tubes 6AQ5 and 6AK6, double-diode triode 6AT6, diode 6AL5, double triode 6J6, high-μ triode 6J4, rectifier 6X4, and ac/dc tubes 12BE6, 12BA6, 12AT6, and 12AV6 which are identical with the corresponding 6-volt tubes except that their heater current is 0.15 ampere at 12.6 volts.

621.385 2080
A New Series of Small Radio Valves—G. Alma and F. Prakke. (*Philips Tech. Rev.*, vol. 8, pp. 289–295; October, 1946.) Discussion of the "A" series or "Rimlock" tubes, 22 mm in diameter, for which overheating of the cathode is avoided by joining the glass or metal bulb to the flat glass base with a glaze or cement which becomes plastic at a comparatively low temperature.

621.385 2081
New Post-War German Valves—(*Radio Tech.* (Vienna), vol. 25, p. 249; April, 1949.) Brief particulars of 15 tubes, including type of base and of body and, in some cases, heater voltage and current. Output data are included for 3 new developments: UEL71 and VEL11, 2-watt tetrode/pentodes, and UL2, a 1.5-watt output pentode.

621.385 2082
Series of Modern Valves for F.M. Broadcasting and for Television—J. Becquemont. (*Onde Élec.*, vol. 29, pp. 145–151; April, 1949.) Discussion of methods of achieving small interelectrode distances, high electron densities, high anode dissipation, low reaction capacitance and output impedance, high gain and complete separation of input from output. The use of Dilver-P, an alloy comparable to kovar, together with a glass of the same expansion coefficient has made possible the construction of a series of tubes in which the various electrodes are carried on concentric cylinders, which form the base connections. The operation of welding the Dilver-P cylinders to the intermediate glass rings to form the tube base is carried out in a high-temperature furnace, using a graphite mandrel. This construction gives short connections to the electrodes. Illustrations are given of a 500-watt triode and a 2-kw tetrode with anode cooling fins.

621.385 2083
Theory and Applications of Trochotrons—H. Alfvén; L. Lindberg; K. G. Malmfors; T. Wallmark; and E. Åström. (*Kungl. Tekn. Högsk. Handl.*, (Stockholm), no. 22, 106 pp; 1948. In English.) Includes 5 separate but coordinated papers. The first, "On Trochoidal Electronic Beams and Their Use in Electronic Tubes (Trochotrons)," by Alfvén, is an integrating paper. The equations of motion of charged particles in a nearly homogeneous magnetic field perpendicular to a nearly homogeneous electric field are discussed, and the properties of the resulting trochoidal beams are considered in detail. The arrangement of a simple trochotron is described. The cathode is an electron gun, placed between an L-shaped anode and a straight "rail" electrode at nearly cathode potential. The electron beam travels between the "rail" and a number of "boxes." The anode forms one side of the last box, while the side of the first box remote from the cathode is connected to the rail. Electrodes forming sides of the boxes, at right angles to the rail, are called spades; those forming sides parallel to the rail are called plates. All spades and plates, and the anode, are normally at a potential of +200 volts relative to the cathode, except the spade connected to the rail. The beam may enter the first box and be collected by the first plate. The beam can be made to enter one of the other boxes by altering the voltage of the corresponding spade. For earlier work, see 3800 of 1945 and 1205 of 1948.

The second paper, "Design and Properties of Trochotrons," by Wallmark, discusses the trochotron in greater detail, with illustrations of the way in which the beam behaves when some electrode voltages are changed by steps. Operating conditions, the size of the electrode system, electrode materials, etc., are considered.

The third paper, "Design of Trochotron Circuits," by Lindberg, discusses the way in which external pulses may be used to operate each spade as a self-locking switch. Applications to counters, chronoscopes, and pulse-time modulation are also discussed.

The fourth paper, "Experimental Investigation on an Electron Gas in a Magnetic Field," by Åström, discusses experimental results concerning currents which flow to electrodes whose potential is negative relative to the cathode. These "negative currents" are always accompanied by noise. They occur even if there is no alternating voltage on the electrodes. This phenomenon is related to cutoff effects in magnetrons.

The fifth paper, "On the Instability of an Electron Gas in a Magnetic Field," gives theory primarily proposed as an explanation of Åström's results, but it may also throw light on the origin of solar and cosmic noise.

621.385 2084
Valves with Resistive Loads—S. W. Amos. (*Wireless Eng.*, vol. 26, pp. 119–123; April, 1949.) Expressions are derived for the maximum undistorted output that can be delivered to a purely resistive load by a tube with given high-voltage supply. Although the expressions are derived for the cathode-follower circuit, many of the formulas may be applied to tubes with the load in the anode circuit.

621.385 2085
Some Recent Developments in the Technique of Radio Valve Manufacture—J. W. Davies, H. W. B. Gardiner, and W. H. Gomm. (*Proc. Inst. Mech. Eng.* (London), vol. 158, no. 3, pp. 352–363; 1948. Discussion, pp. 364–368.) A general account of the mechanical aspects of the manufacture of large transmitting and high-frequency tubes, with particular attention to the making of spiral, squirrel-cage, and planar grids, glass-to-metal joints of the annular type, using chucks for holding the component parts, and the assembly and alignment processes.

621.385:061.3 2086
Digests of Papers Presented at Conference on Electron Tubes—(*Elec. Eng.*, vol. 67, pp. 589–600; June, 1948.) Authors' summaries of most of the papers read are given. Full texts are being published in the *Proceedings of the Conference on Electron Tubes* (price $3). The *Report on Electron Tube Survey* (price $2) prepared by an AIEE Committee, contains data on which much information discussed at the conference was based. Both are obtainable from AIEE Order Department, 33 West 39 Street, New York 18, N. Y.

621.385:621.396.822 2087
The Noise Factor of Grounded-Grid Valves—A. van der Ziel and A. Versnel. (*Philips Res. Rep.*, vol. 3, pp. 255–270; August, 1948.) A mathematical treatment of grounded-grid tubes in which part of the output noise current also flows in the input circuit, giving partial noise suppression. Triodes, pentodes, and secondary-emission tubes are considered, and the effects of circuit and dielectric losses, transit time, field inhomogeneities, partition noise, and secondary-emission noise are discussed.

Measurements made on tubes for $\lambda = 7.25$ meters agree with theory for curves of noise factor against antenna resistance, and for noise resonance curves. Loose antenna coupling favors a low noise factor, but complete noise suppression is impossible. Suppression is most effective for grounded-grid triodes over a narrow frequency band, so that this type of tube is less useful for wide-band working. The noise factor for secondary-emission tubes is much greater than that for triodes. See also 249 of February.

621.385.029.63/.64 2088
Some Slow-Wave Structures for Traveling-Wave Tubes—L. M. Field. (PROC. I.R.E., vol. 37, pp. 34–40; January, 1949.) The gain per unit length of a traveling-wave tube is determined by a structure factor. Four types of structure are compared, namely (a) helix, (b) disk-loaded rod, (c) apertured-disk, and (d) helical waveguide. Amplifier performance of helix tubes at 10,000 Mc and wide-tuned oscillator performance (1.5 to 1) with second-harmonic output around 20,000 Mc are considered. Design, construction, and performance of a 10,000-Mc tube of type (b) are also discussed.

621.385.029.63/.64 2089
Effect of the Transverse Electric Vector in the Delay Line of the Traveling-Wave Valve: Part 1—O. Doehler and W. Kleen. (*Ann. Radioélec.*, vol. 4, pp. 76–84; January, 1949.) The hypotheses used in different theories of the linear behavior of the traveling-wave tube are briefly reviewed and the effect of the transverse electric field of the delay line on the interaction between the beam and the wave is examined. This effect appears to have been neglected hitherto. The radial field causes five waves to be excited instead of the three found on the

hypothesis of a purely longitudinal field; the gain of the amplified waves can be appreciably increased by the action of the radial field. Particular types of delay line where such effects predominate are discussed and the change of gain due to the space charge and to the absorption of electrons by the delay line is studied. To be continued. See also 1543 of June (Pierce).

621.385.032.216 2090
Fluctuation Effects of Emission from Oxide-Coated Surfaces—R. P. Bien and Yang Yo-Han. (*Science Rec.* (Nanking), vol. 2, pp. 65–70; October, 1947. In English.) Fluctuations observed bear no relation to heating current but can be explained statistically.

621.385.032.216 2091
Thermionic Emission from Oxide Coated Cathodes—D. A. Wright. (*Proc. Phys. Soc.*, vol. 62, pp. 188–203; March 1, 1949.) Discussion of an experimental investigation of the emission from Ba/Sr oxide cathodes and from thoria cathodes under pulsed and dc conditions. Semiconductor theory can explain the emission and conductivity of cathode coatings. The part played by the interface layers between the coating and the metal to which it is applied is also considered. For earlier work see 606, 1980, and 1981 of 1948.

621.385.032.216 2092
Poisoning in High-Vacuum Oxide-Cathode Valves—G. H. Metson and M. F. Holmes. (*Nature* (London), vol. 163, pp. 61–62; January 8, 1949.) The ionizing action of the current after leaving the cathode surface is suggested as the prime factor leading to cathode deterioration. Results of life tests on two batches of a particular type of pentode are discussed. Those arranged as triodes with anodes strapped to screen and suppressor grids and primed with 200 volts failed after 500 to 1,000 hours, whereas those connected as diodes, with control grids acting as collectors and +2 volts applied between grid and cathode to give a cathode current of about 12 milliamps showed little deterioration after 6,000 hours.

621.385.032.216 2093
Resistance of Oxide Cathode Coatings for High Values of Pulsed Emission—W. E. Danforth and D. L. Goldwater. (*Jour. Appl. Phys.*, vol. 20, pp. 163–173; February, 1949.) The potential variation of fine ribbon probes embedded in standard BaO or SrO coatings was observed for 19 tubes. Potential gradients were found adequate to admit dielectric breakdown as a cause of sparking. The resistance of SrO cathodes is several times that of BaO or mixed-oxide cathodes. Superposition of dc upon pulsed emission causes a marked decrease in resistance.

621.385.032.216 2094
Work Functions and Conductivity of Oxide-Coated Cathodes—G. W. Mahlman. (*Jour. Appl. Phys.*, vol. 20, pp. 197–202; February, 1949.)

621.385.032.42 2095
A New Air-Cooling System for Transmitting Valves—W. L. Vervest. (*Communication News*, vol. 9, pp. 92–96; April, 1948.) See also 2673 of 1948 (de Brey and Rinia).

621.385.032.5:666.1.037.5 2096
The Electrode Leads of Transmitting Valves—E. G. Dorgelo. (*Communication News*, vol. 9, pp. 38–44; December, 1947.) Survey and discussion of various methods of sealing.

621.385.2:621.396.822 2097
Noise Spectrum of Temperature-Limited Diodes—D. B. Fraser. (*Wireless Eng.*, vol. 26, pp. 129–132; April, 1949.) A simple and completely general derivation of the formula for the fluctuation currents in plane and cylindrical diodes, with explicit allowance for the transit time. When transit time is negligible, the formula reduces to that of Schottky.

621.385.2:621.396.822 2098
Noise Spectrum of a Diode with a Retarding Field—J. J. Freeman. (*Jour. Res. Nat. Bur. Stand.*, vol. 42, pp. 75–88; January, 1949.) A general expression is derived for the spectrum generated by the random emission of electrons having arbitrary trajectories within a waveguide. A numerical solution is obtained for the potential distribution within a plane diode; results are shown as a series of curves, and compared with von Laue's results. The equivalent mean-square fluctuation current due to the space charge within a diode is deduced for (a) linear potential distribution, and (b) the distribution which occurs near the beginning of the retarding field. In case (a), the equivalent noise temperature of the diode conductance is equal to the cathode temperature.

621.385.2.032.216 2099
Some Characteristics of Diodes with Oxide-Coated Cathodes—W. R. Ferris. (*RCA Rev.*, vol. 10, pp. 134–149; March, 1949.) The Epstein-Fry-Langmuir equation for the space-charge current in a plane diode and the Boltzmann equation for the retarding-field current are used to obtain a set of universal characteristic curves for plane diodes; $V_r|e/kT|$ is taken as the abscissa. These curves agree with experimental measurements when a series resistance is assigned to the oxide cathode. Universal curves of incremental conductance and tables of functions used are given.

621.385.3.029.64 2100
Electronics of Ultra-High-Frequency Triodes—R. R. Law. (Proc. I.R.E., vol. 37, pp. 273–274; March, 1949.) An empirical relation is deduced for anode efficiency as a function of frequency, voltage, and interelectrode spacing.

621.385.3.032.29:621.317.335.2† 2101
Triode Interelectrode Capacitances—E. E. Zepler and J. Hekner. (*Wireless Eng.*, vol. 26, pp. 53–58; February, 1949.) The variations of grid-to-cathode and grid-to-anode capacitance with working conditions were investigated experimentally. The effects of the mutual conductance, amplification factor, and supply voltages are shown graphically. A theory is given which is in fair agreement with these results.

621.385.032.29:621.317.335.2† 2102
Interelectrode Capacitance of Valves—B. L. Humphreys and E. G. James. (*Wireless Eng.*, vol. 26, pp. 26–30; January, 1949.) Discussion of measurements made under different operating conditions, on two types of tube—DET 22 and E 1714—whose active elements are all cylindrical and coaxial. The measurements were made on a rf bridge at a frequency of 1 Mc. The increase of grid-to-cathode capacitance with increasing anode current was much greater than that expected theoretically and depended greatly on the grid-to-cathode geometry. Anode-to-grid capacitance decreased very slightly with increasing anode current.

621.385.832:621.397.6 2103
The Graphechon—A Picture Storage Tube—Pensak. (*See* 2061.)

621.396.615.142:621.316.726 2104
Frequency Stabilization of V.M. Valves—H. Borg. (*Wireless Eng.*, vol. 26, pp. 59–73; February, 1949.) A simplified discussion of general principles mainly in connection with continuous-wave microwave oscillators. Frequency-control systems are considered in which an error voltage, generated between a standard reference frequency source and the oscillator to be stabilized, is used to correct frequency variations of the oscillator. An application is described in which the frequency of a v.m. tube at 9,360 Mc is stabilized by comparison with a crystal, giving short-term stability, relative to the crystal, of the order of +100 cps.

621.396.615.142.2 2105
Valves for Communication on Frequencies above 1,000 Mc/s: Part 1—H. Schnitger. (*Fernmeldetech. Z.*, vol. 2, pp. 51–56; February, 1949.) A general outline of the principles of klystrons of the normal and the reflex type, with a short account and illustration of the 2-chamber klystron of the German Post Office; this gives an output of 100 watts on a wavelength of 9 cm.

621.396.645:537.311.33:621.315.59 2106
Germanium—Important New Semiconductor—Dunlap, Jr. (*See* 1968.)

621.396.645:537.311.33:621.315.59 2107
Some Novel Circuits for the Three-Terminal Semiconductor Amplifier—Webster, Eberhard, and Barton. (*See* 1905.)

621.396.645:537.311.33:621.315.59 2108
The Double-Surface Transistor—J. N. Shive. (*Phys. Rev.*, vol. 75, pp. 689–690; February 15, 1949.) Emitter and collector point-contacts bear on opposite faces of a thin wedge of Ge and a third contact of larger area is provided on the base of the wedge. Separation between the points should not exceed 0.1 mm. Families of curves are given which facilitate correct choice of the dc operating voltage and give complete information for determining the dynamic input and output impedances and the forward and backward transfer impedances about any selected operating point. An explanation of the action of this type of transistor is given. See also 913 of April (White) and back references and 2109 below.

621.396.645:537.311.33:621.315.59 2109
Investigation of Hole Injection in Transistor Action—J. R. Haynes and W. Shockley. (*Phys. Rev.*, vol. 75, p. 691; February 15, 1949.) The impedance changes at the collector point of a transistor were investigated by applying an intermittent potential to multiple emitter points and observing the resulting probe-current variations. The results are explained in terms of the movement of positive particles with a mobility of about 1.2×10^3 cm/sec per v/cm.

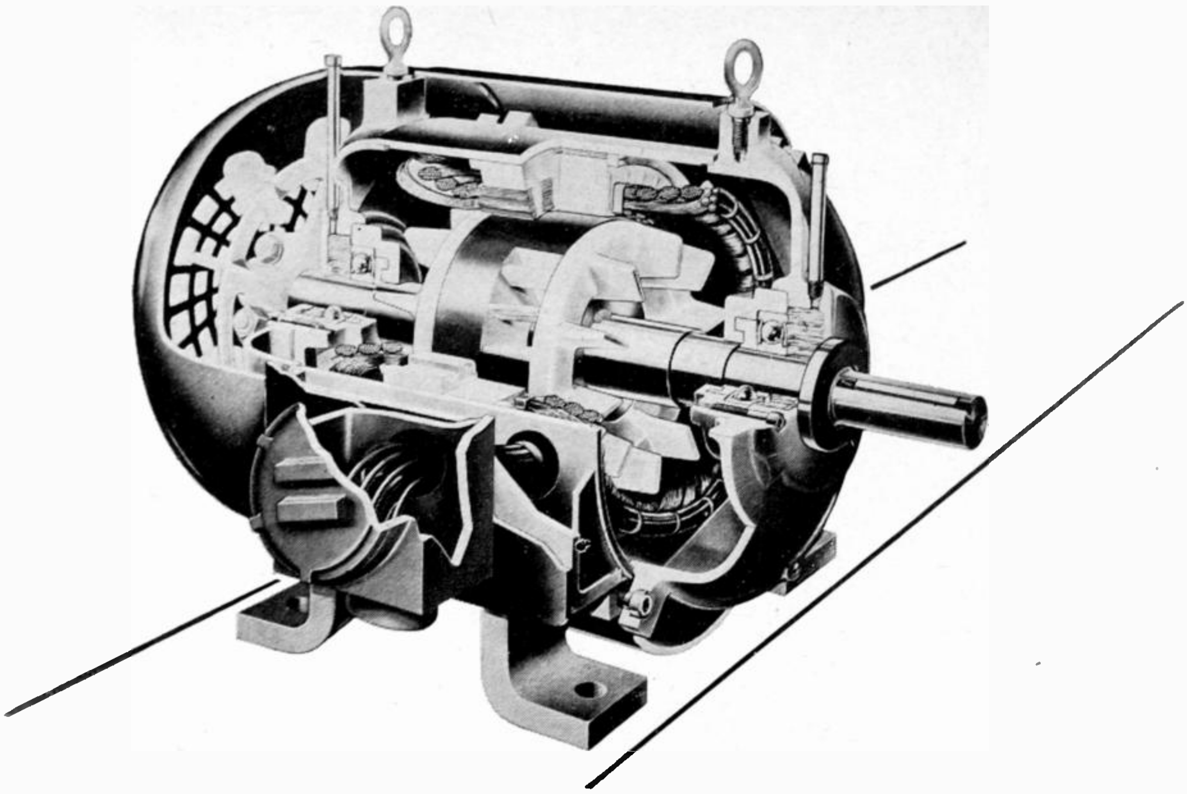
621.383 2110
Photoelectric Cells in Industry [Book Review]—R. C. Walker. Pitman and Sons, London, 501 pp., 40s. (*Electronic Eng.* (London), vol. 21, p. 68; February, 1949.) Typical uses of such cells are discussed from the point of view of the practical man whose purpose is to use electrons rather than to theorize about them.

MISCELLANEOUS

621.39 2111
What the S.C.E.L. [Signal Corps Engineering Laboratories] is Doing—H. A. Zahl. (*FM-TV*, vol. 9, pp. 13–22; February, 1949.) A general survey of work in many fields. Development work is the primary task, but basic and applied research work related to the military effort is also undertaken.

621.396 2112
Radio Progress during 1948—(Proc. I.R.E. vol. 37, pp. 286–322; March, 1949.) A general review with a bibliography of 855 references.

621.396 Popov 2113
Alexander S. Popov—J. B. Thornton. (*Wireless Eng.*, vol. 26, pp. 141–142; April, 1949.) Comment on 1842 of 1948 (G. W. O. H.).



SECTION MEETINGS

Atlanta

"Discussion of and Tour Through WAGA Television Station," by P. Cram, Chief Engineer, Television Station WAGA; May 20, 1949.

Baltimore

"Interference Problems," by E. W. Chapin, Chief Engineer, Federal Communications Commission: "Advantages of FM Transmission of Television Programs in the UHF Band," by W. K. Roberts, Assistant Chief Engineer, Federal Communications Commission; Business Meeting; Nomination of Officers; May 24, 1949.

Beaumont-Port Arthur

"Electron Optics and Its Applications," by R. H. Biser, Faculty of Lamar College; May 24, 1949.

Buenos Aires

Election of Officers; April 8, 1949.

Buffalo-Niagara

Nomination of Officers; May 12, 1949.

"A Flow Meter for Granular Substances," by J. F. McCullough; "A Magnetic Thickness Gauge," by R. Rowe, Carborundum Company; "Comparisons Between American and British Television," by D. Swaine, Colonial Radio; Election of Officers; May 18, 1949.

Boston

"A New Long-Playing Disk Recording System," by P. C. Goldmark, Columbia Broadcasting System, October 26, 1948.

"A New Method for Measuring the Product of Two Voltages Using a Single Vacuum Tube," by D. B. Sinclair, Assistant Chief Engineer, General Radio Company, November 18, 1948.

"The Electronic Theory of the Transistor," by W. Shockley, Bell Telephone Laboratories; December 16, 1948.

"The Physical Concepts in the Statistical Approach to Communication Problems," by R. M. Fano, Massachusetts Institute of Technology; January 20, 1949.

"Traveling-Wave Amplifiers," by H. G. Rudenberg, Raytheon Manufacturing Company; February 17, 1949.

"Radio Telemetering," by C. H. Hoeppner, Raytheon Manufacturing Company; March 24, 1949.

"Psycho-Acoustic Aspects of Speech Compression," by J. C. R. Licklider, Faculty of Harvard University; April 21, 1949.

"Television Station Installation and Operation," by W. H. Hauser and S. V. Stadig, Radio Station WBZ; Election of Officers; May 26, 1949.

"Microwave Spectroscopy," by R. Karplus and E. Fletcher, Faculty of Harvard University; June 17, 1949.

Cedar Rapids

"The Baldwin Electronic Organ," by A. F. Knoblaugh, The Baldwin Piano Company, May 12, 1949.

Cincinnati

Election of Officers; June 15, 1949.

Cleveland

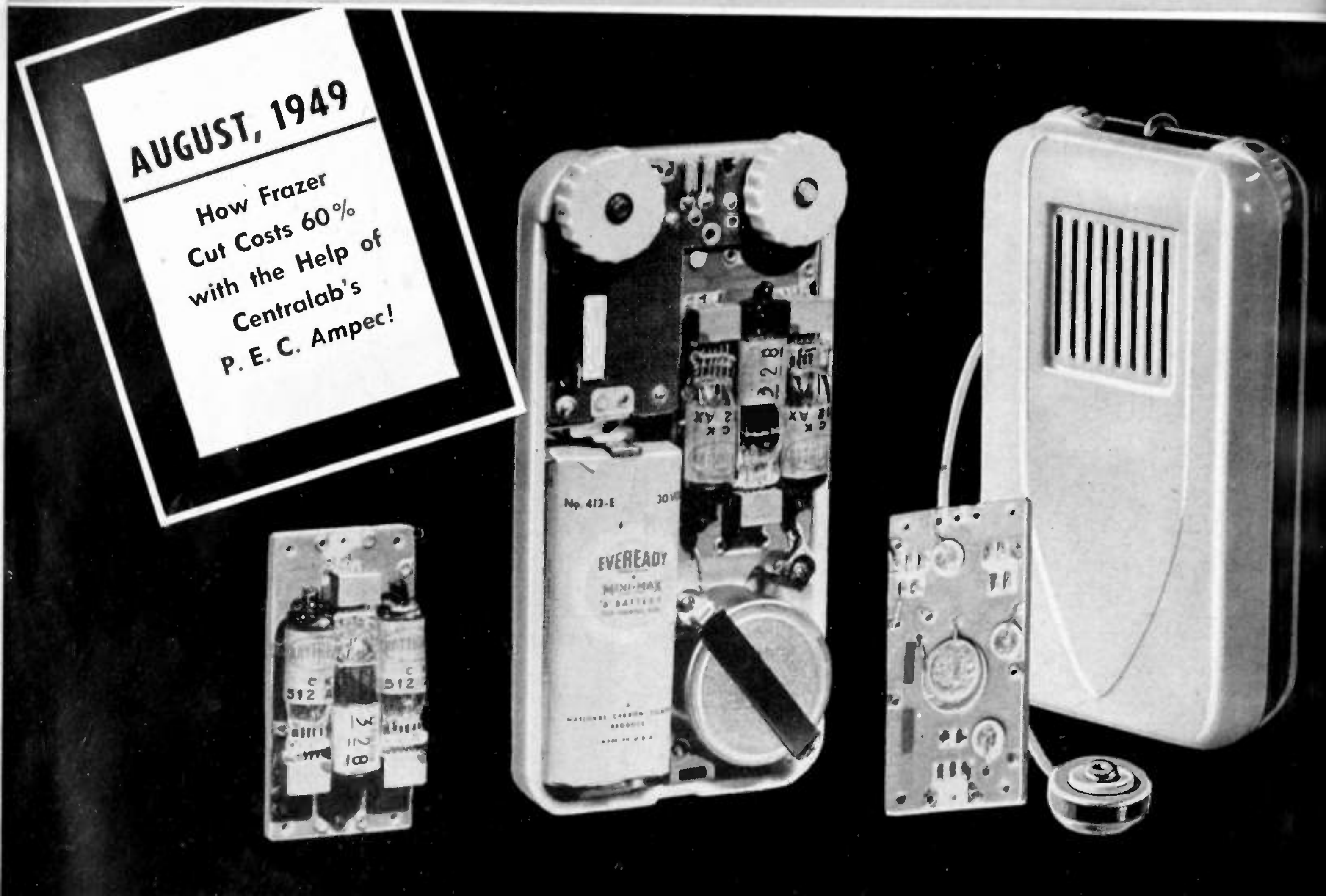
Inspection Tour; "Description of WNBK and WTAM Studios," by E. Leonard, Engineer-in-Charge, National Broadcasting Company Cleveland Facilities; and A. Hammerschmidt, Supervisor, Television Operations; May 26, 1949.

Connecticut Valley

Election of Officers; June 4, 1949.

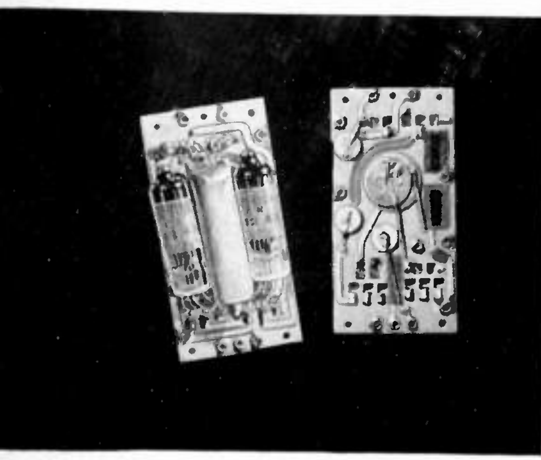
(Continued on page 39A)

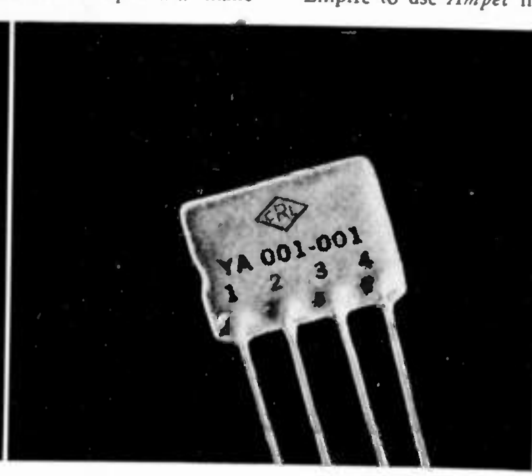
Centralab reports to

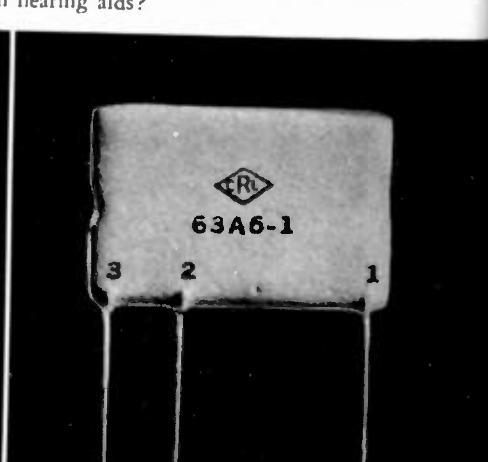


Models courtesy of the Frazer Company

1 Size and weight were important considerations to the Frazer Company, Auckland, New Zealand. That's why this firm switched to Centralab's amazing *Printed Electronic Circuit* for use in the hearing aids it manufactures. Frazer found that *Ampec did* make smaller and lighter units possible. It also found that this tiny 3-stage audio-amplifier helped cut production costs almost 60%. Is it any wonder Frazer is proud of being first in the British Empire to use *Ampec* in hearing aids?





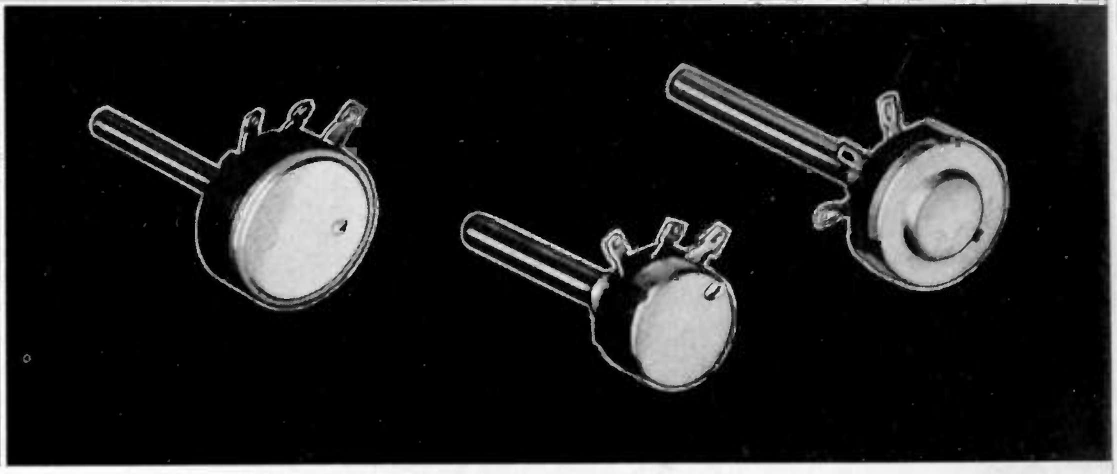


2 Centralab's Ampec, above, is an integral assembly of tube sockets, capacitors, resistors and wiring combined into one miniature amplifier unit.

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

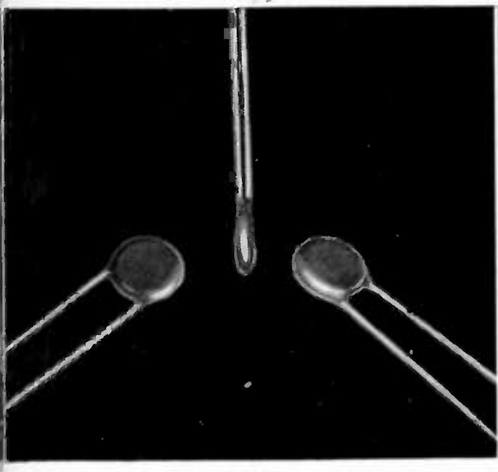
4 This is the new CRL *Vertical Integrator Network* used in TV sets. Variations of this Centralab *Network* are available on special order.

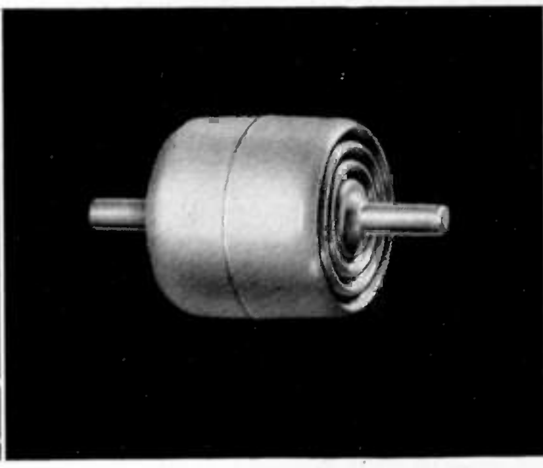
Electronic Industry

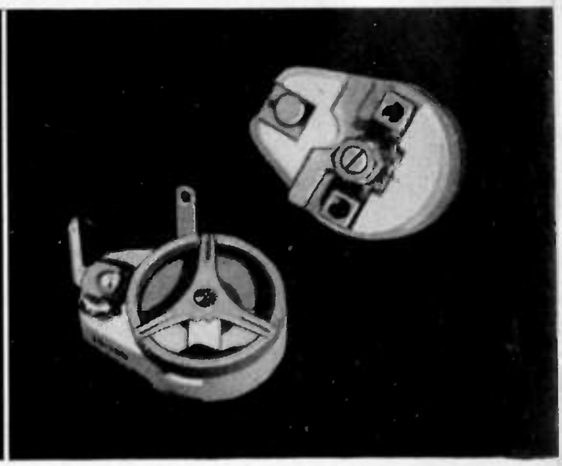


5 Model "1" *Radiohm* control — plain and switch types — is no larger than a dime. Especially designed for miniature applications.

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



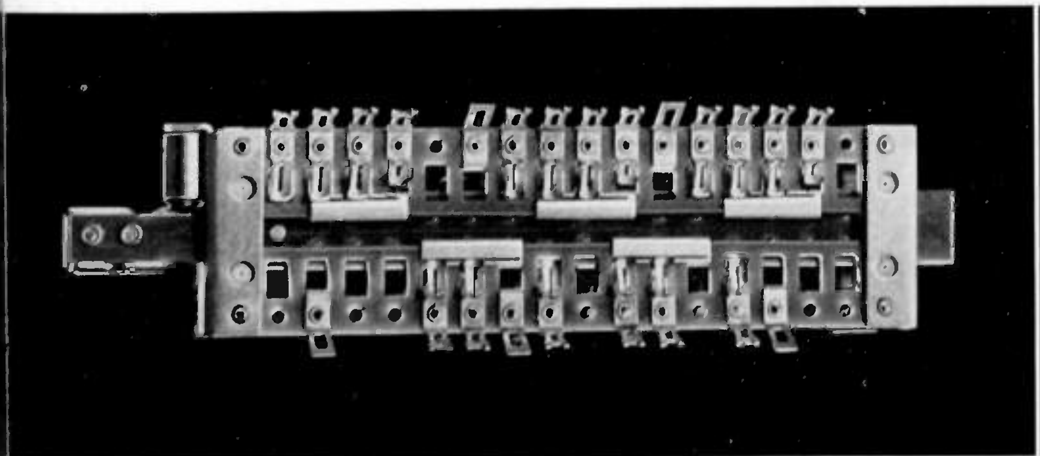


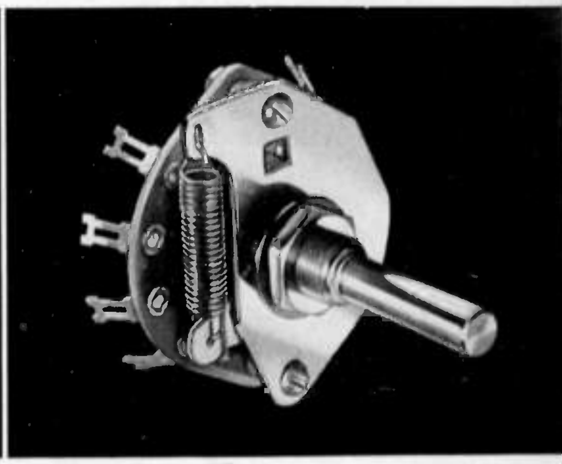


7 For by-pass or coupling applications, check CRL's original line of ceramic disc and tubular *Hi-Kaps*. *Disc Hi-Kaps* are smaller than a dime!

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

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





10 Centralab's development of a revolutionary, new *Slide Switch* gives you improved AM and FM performance! Flat, horizontal design saves valuable space, allows short leads, convenient location to coils, reduced lead inductances for increased efficiency in low and high frequencies. CRL *Slide Switches* are rugged and dependable.

11 Great Step forward in switching is CRL's New *Rotary Coil and Cam Index Switch*. Its coil spring gives you smoother action, longer life.

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Centralab Printed Electronic Circuits

973 — AMPEC — three-tube P. E. C. amplifier.
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999 — PENTODE COUPLATE — specialized P. E. C. coupling plate.
42-9 — FILPEC — Printed Electronic Circuit filter.

Centralab Capacitors

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

Centralab Switches

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

Centralab Controls

42-7 — MODEL "1" RADIOHM — world's smallest commercially produced control.
697 — VARIABLE RESISTORS — full facts on CRL Variable Resistors.

Centralab Ceramics

720 — CERAMIC CATALOG — CRL's steatite and ceramic products.

General

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

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SECTION MEETINGS

(Continued from page 34A)

Dallas-Fort Worth

"Silicones—Electrical Properties and Applications," by Southwest Manager, Dow Corning Corporation; May 17, 1949.

"Pulse Time Modulation," by R. G. Maddox, Federal Telephone and Radio Corporation; June 2, 1949.

Dayton

"IRE Affairs," by S. L. Bailey, President, The Institute of Radio Engineers; Election of Officers; May 12, 1949.

Detroit

"FM—FM Telemetering System," by H. B. Schultheis, Research Engineer, Bendix Aviation Corporation; May 20, 1949.

Fort Wayne

"The Automatic Telephone Exchange," by E. J. Kane, Home Telephone and Telegraph Company, Election of Officers; May 23, 1949.

"Microwaves, Their Development and Use," by M. G. Staton, RCA-Victor Division; "Brain Waves," by C. Mengani, Indiana Technical College; June 13, 1949.

London

"Annual Banquet Meeting," by F. H. R. Pounsett, Stromberg-Carlson Company, Ltd.; February 16, 1949.

"Hysteresis Effects," by J. Young; "Frequency Divider Networks," by D. Aaronson; "Three-Dimensional Cathode-Ray Display," by N. Broten; March 11, 1949.

"Practical Aspects of Modern Directional Broadcast Antenna Arrays," by G. A. Robitaille, Chief Engineer, CFPL; April 29, 1949.

"Practical Discussion of General Antenna Engineering," by J. E. Hayes, Columbia Broadcasting Company; Election of Officers; May 27, 1949.

Los Angeles

Television Symposium and panel discussion; May 21, 1949.

Milwaukee

Election of Officers; June 9, 1949.

New York

"The Stratovision System," by C. E. Nobles, Westinghouse Electric Corporation; December 1, 1948.

"The Electron Wave Tube," by A. V. Haeff, Naval Research Laboratory, and J. R. Pierce, Bell Telephone Laboratories; December 16, 1948.

"A Field Test of UHF Television in the Washington Area," by G. H. Brown, RCA Laboratories; January 5, 1949.

"Instantaneous Audience Measurement System (IAMS)," by P. C. Goldmark, J. W. Christensen, A. Bark, J. T. Wilner, and A. Goldberg; Columbia Broadcasting System; January 19, 1949.

"A New Microwave Triode," by J. A. Morton, A. E. Bowen, W. W. Mumford, and M. E. Hines, Bell Telephone Laboratories; February 2, 1949.

Motion Picture on "Atomic Physics," by J. R. Dunning, Faculty of Columbia University; February 16, 1949.

"Development of a Large Metal Kinescope for Television," by J. Kelar, H. P. Steier, C. T. Lattimore, and R. D. Faulkner, Radio Corporation of America; February 23, 1949.

"Design Problems in Meeting NAB Tentative Magnetic Recording Standards," by W. E. Stewart, Radio Corporation of America; "Noise Factors in Magnetic Tape Recording," by D. G. C. Hare, Consulting Physicist; March 2, 1949.

"A Facsimile Multiplex System for FM Broadcasting Networks," by W. S. Halstead, President, Communications Research Corporation; April 6, 1949.

(Continued on page 40A)

Section Meetings

(Continued from page 39A)

"Acoustic Lenses for Audio-Frequency Applications," by W. E. Kock, Bell Telephone Laboratories; April 18, 1949.

"Image Quality in Photography and Television," by O. H. Schade, RCA Victor Division; May 4, 1949.

"Some Circuit Aspects of the Transistor," by R. M. Ryder, Bell Telephone Laboratories; Election of Officers; June 1, 1949.

North Carolina-Virginia

"Television Station Design," by J. N. Comer, General Electric Company; May 27, 1949.

Pittsburgh

"Control Problems in Nuclear Power Plants," by M. A. Schultz, Westinghouse Electric Corporation; Election of Officers; June 13, 1949.

Portland

"Changing Patterns that Influence Engineering Decisions," by C. W. Leihy, Executive Vice-President, Electrical Publications, Inc; Dedication of Dearborn Hall, May 28, 1949.

Salt Lake

"Directional Antennas," by S. Benson, Radio Institute; June 1, 1949.

San Diego

"Radar System Requirements for Maximum and Minimum Range," by W. S. Ivans, Consolidated Vultee Aircraft Corporation; April 5, 1949.

"Atomic Energy Possibilities," by L. E. Reukema, Faculty of University of California; April 26, 1949.

Television Symposium on "Technical Problems of Studio Production," by R. W. Clark, National Broadcasting Company; "Television Receiver Installation and Service Problems," by L. Borgeson, Radio Corporation of America Service Company, Inc.; Round Table Discussion, led by L. Papernow, Television Broadcasting Company; June 7, 1949.

Seattle

"Distributed Amplifier," by R. A. Wilson, Graduate Student, University of Washington; "Base Reflex Increase," by R. Foss, Graduate Student, University of Washington; "RC Coupled Feedback Amplifiers," by L. D. Barter, Graduate Student, University of Washington; May 26, 1949.

"L-1 Carrier Telephone System," by D. Nutting, Pacific Telephone and Telegraph Company; June 10, 1949.

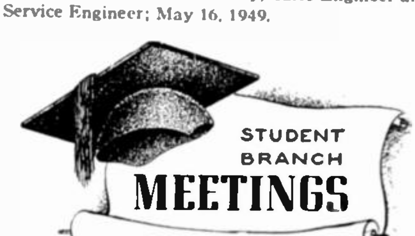
Williamsport

"The Use of Stratovision in the UHF Band," by C. E. Noble, Westinghouse Electric Corporation, May 25, 1949.

SUBSECTIONS

Amarillo-Lubbock

"Westinghouse 50 Kw AM Transmitter," by Messrs. Massey and McHoney, Sales Engineer and Service Engineer; May 16, 1949.



University of California Society of Electrical Engineers—
IRE-AIEE Branch

Field Trip; May 10, 1949.
Field Trip; May 19, 1949.

University of Colorado—IRE Branch

"The Electret," by W. E. Brittin, Faculty of University of Colorado; Nomination of Officers; May 25, 1949.

(Continued on page 42A)

STUDENT BRANCH

MEETINGS

(Continued from page 40A)

GEORGIA SCHOOL OF TECHNOLOGY—IRE BRANCH

"Tour of Transmitter Station," by C. F. Daugherty, Radio Station WSB; April 27, 1949.

"Manufacture of Power Transformers," by B. J. Sturman, Jr., Westinghouse Electric Corporation; Election of Officers; May 3, 1949.

Film, "Stepping Along with T-V"; May 19, 1949.

IOWA STATE COLLEGE—IRE-AIEE BRANCH

Election of Officers; May 24, 1949.

STATE UNIVERSITY OF IOWA—IRE BRANCH

Student Talks; May 11, 1949.
Student Talks; May 18, 1949.
Election of Officers; May 25, 1949.

UNIVERSITY OF MINNESOTA—IRE-AIEE BRANCH

"Color Movies on Hawaiian Islands," by R. T. S. Carter, President, RTS Carter Company; May 10, 1949.

NEWARK COLLEGE OF ENGINEERING—IRE BRANCH

"The Graphic Recorder as a Measuring Tool," by L. P. Reitz, Sound Apparatus Company; May 18, 1949.

NEW YORK UNIVERSITY—IRE BRANCH

Business Meeting and Election of Officers; May 24, 1949.

NORTH CAROLINA STATE COLLEGE—IRE BRANCH

"Opportunities for Engineers in the Federal Service," by E. McCrensky, Director of Personnel, Office of Naval Research; May 25, 1949.

OREGON STATE COLLEGE—IRE BRANCH

"High Voltage Power Arc Test," and movie of test, by E. C. Starr, Faculty of Oregon State College; Election of Officers; May 25, 1949.

"Factors Which Affect Engineering Decisions," by C. W. Leihy, Executive Vice-President, Electrical Publications Inc.; and Banquet; May 28, 1949.

PURDUE UNIVERSITY—IRE BRANCH

"Ceramic Dielectric, Ceramic Capacitors, and Printed Circuits," by B. Marks, Centralab Company; Election of Officers; May 31, 1949.

ST. LOUIS UNIVERSITY—IRE BRANCH

Election of Officers; March 24, 1949.

"Telemetering," by P. T. Ramey, Union Electric Company; April 21, 1949.

"Selenium Rectifiers," by F. A. Waelterman, Vickers, Inc.; April 28, 1949.

"Missouri Society of Professional Engineers," by E. S. Rehagen, Westinghouse Electric Corporation; May 19, 1949.

SAN DIEGO STATE COLLEGE—IRE BRANCH

Election of Officers; May 17, 1949.

"Visual Understanding of Video," by R. T. Silberman, Student; May 23, 1949.

SEATTLE UNIVERSITY—IRE BRANCH

Election of Officers; April 21, 1949.

STANFORD UNIVERSITY—IRE-AIEE BRANCH

Election of Officers; "General Electronics Research Picture at Stanford," by F. E. Terman, Faculty of Stanford University; Student Talks; and Open House; May 25, 1949.

UNIVERSITY OF WASHINGTON—
IRE-AIEE BRANCH

"Northwest Power Problems," by Mr. Lampson, B.P.A.; December 9, 1948.

Business Meeting; February 9, 1949.

Business Meeting and Nomination of Officers; April 8, 1949.

Business Meeting and Election of Officers; May 2, 1949.

WAYNE UNIVERSITY—IRE-AIEE BRANCH

Election of Officers; May 12, 1949.

"RF Induction Heating," by Mr. Cardwell, Westinghouse Electric Corporation; June 2, 1949.

MEMBERSHIP

The following transfers and admissions were approved and will be effective as of August 1, 1949:

Transfer to Senior Member

Baldwin, L. W., Box 118, R.F.D. 1, Oxnard, Calif.
Byers, H. G., 133 Glengarry Ave., Toronto 12, Ont., Canada
Courtney, J. M., 359 N. Datil Dr., Albuquerque, N. Mex.
Davis, A., Jr., 920 East 49, Austin, Tex.
Dennison, B. H., 1314 S. Pollard St., Arlington, Va.
De Shong, J. A., Jr., 1414 N. Austin Bldv., Oak Park, Ill.
Hidy, J. H., 4410 S. Peoria, Box 58, Tulsa, Okla.
Klein, R. M., 2200 Morris Ave., New York 53, N. Y.
Lundahl, T., Columbus, Sherburne, N. Y.
Martin, A. E., 226 West 137 St., New York 30, N. Y.
Masters, R. W., Marlton Pike & Wesley Ave., Erlton, N. J.
McGaughey, J. R., Navy Electronics Laboratory, San Diego 52, Calif.
Perper, L. J., 3255 Ridge Ave., Dayton 5, Ohio
Rochester, N., R.F.D. 2, Wappingers Falls, N. Y.
Rollefson, K. E., 1615 Ridge Ave., Evanston, Ill.
Shepherd, W. G., 2176 Stanford Ave., St. Paul, Minn.
Talmage, F. E., 340 Comly Ave., W. Collingswood, N. J.

Admission to Senior Member

Altovsky, V. A., 88 rue Lecourbe, Paris 15, France
Bartlett, S. C., 74 Etville Ave., Yonkers 2, N. Y.
Bernreuter, H. A., 5208 W. Kinzie St., Chicago 44, Ill.
Bowie, W. G., 302 Houston Ave., Syracuse 10, N. Y.
Dorff, L. A., 92 Hawthorne St., Glen Ridge, N. J.
Fong, L. B. C., National Bureau of Standards, Electronics Division, Washington, D. C.
Given, F. J., Bell Telephone Laboratories, Murray Hill, N. J.
Hoekstra, C. E., 2912 Holton Ave., Fort Wayne 5, Ind.
Hosmer, E. A., 420 Market St., San Francisco 11, Calif.
Kendall, H. C., 143 Edgemont Rd., Rochester 7, N. Y.
Landrey, L. R., 58 Elm St., Lynbrook, L. I., N. Y.
McGraw, J. E., Army Field Forces Board #4, Fort Bliss, Tex.
Merrill, L. L., Clarkson College, Potsdam, N. Y.
Moisson, A. A., 4 Square Charles Laurent, Paris 15, France
Phillipps, R. J., 37 Lexington Dr., Livingston, N. J.
Rawlins, R. E., 11180 Emelita St., North Hollywood, Calif.
Raymond, F. H., 37 Avenue des Courlis, Le Vesinet, (Seine & Oise), France

(Continued on page 44A)

MEMBERSHIP

(Continued from page 43A)

Spandau, E. P., 801 Broad Blvd., Dayton 9, Ohio
Victoreen, J. A., 3800 Perkins Ave., Cleveland 14, Ohio

Transfer to Member

Bedwell, T. H., Physics Department, Florida State University, Tallahassee, Fla.
Bittner, B. J., 3110-42 Pl., Sandia Base Branch, Albuquerque, N. Mex.
Bodle, D. W., Bell Telephone Laboratories, 463 West St., New York 14, N. Y.
Brannen, P. M., 24½ N. Fifth St., Duquesne, Pa.
Florman, L. W., 413 N. Rush St., Itasca, Ill.
Gross, J., Box 973, Nairobi, Kenya Colony, British East Africa
Hancock, N. W., 600-13 St., N. W., Cedar Rapids, Iowa
Hoffman, R. Y., Jr., 1234 Ridgewood Dr., Northbrook, Ill.
Jorgenson, T. O., WEAU Transmitter, R. R. 1, Hy Q, Eau Claire, Wis.
Osbahr, B. F., 206 Eighth Ave., Brooklyn 15, N. Y.
Pfefer, B. L., 428 Woodbine Ave., Syracuse 6, N. Y.
Ratts, B. H., 2506 Terrace Rd., Fort Wayne 3, Ind.
Schull, G. R., 12131 Mayfield Ave., Los Angeles 24, Calif.
Waller, M. J., R.F.D. 1, Foxboro, Ont., Canada
Winkler, E. H., 71 White St., Shrewsbury, Red Bank, N. J.
Young, C. W., 4745 Nogal St., San Diego 2, Calif.
Zupnick, I. N., 768 Linden Blvd., Brooklyn 3, N. Y.

Admission to Member

Allan, D. K., 925 Louisiana Ave., Baton Rouge, La.
Anderson, C. W., 920 Pine St., St. Louis 1, Mo.
Baker, E. E., Jr., Box 379, APO 182, c/o Postmaster, San Francisco, Calif.
Belprez, G. R., 1322 Chalmers Ave., Detroit 15, Mich.
Bettis, E. S., Box 105, Oak Ridge, Tenn.
Bindner, J. T., 1586 Hedding Ct., San Jose 11, Calif.
Chess, R. B., 1830 South 54 Ave., Chicago 50, Ill.
Collier, J. W., 1903 Ridge Pl., Washington, D. C.
Cooke, H. F., Apt. 17, Maxwell Ct., Mains Ave., Syracuse, N. Y.
Dahl, A. H., 114 Disston Rd., Oak Ridge, Tenn.
Danielsen, A. C., 1429 South 51, Milwaukee 14, Wis.
Das, P. N., 1 Bhaduri Lane, P.O. Serampore, Dist. Hooghly, Bengal, India
Diehl, C. E., Box 1535, Salt Lake City 11, Utah
Frenkel, L. J., Jr., 216 South Yale, Albuquerque, N. Mex.
Hassel, E. W., 2911 Erie St., Racine, Wis.
Hershey, J. H., Washington Valley Rd., Morristown, N. J.
Hosker, G. R., c/o Richards-Wilcox Canadian Ltd., London, Ont., Canada
Kilbey, A. R., 20 Dix St., Waltham, Mass.
Layzell, L. M., Radio Planning Department, Flight Ops., Division, K.L.M. Royal Dutch Airlines, Schiphol Airport, Amsterdam, Holland
Lee, R. E., 1923 W. Baltimore St., Baltimore 23, Md.
Mahmoud, A. A., Faculty of Engineering, Fouad I University, Giza, Cairo, Egypt
Mallory, V. L., Continental Electronics Manufacturing Co., 1728 Wood St., Dallas, Tex.
Martens, H. A., 5675 Vreeland Rd., R.F.D. 2, Ann Arbor, Mich.
Michie, J. L., 23 Laburnum Dr., Skelmersdale, Lancs., England
Miller, S. S. 1540-52 St., Brooklyn, N. Y.
Mutter, W. E., Graduate House, Massachusetts Institute of Technology, Cambridge 39, Mass.

(Continued on page 45A)

MEMBERSHIP

(Continued from page 44A)

Platzman, M. M., c/o Video Corporation of America, 229 W. 28 St., New York 1, N. Y.
Rawhouser, R., 65 W. Great Miami Blvd., Dayton 5, Ohio
Ridenour, P. W., OMR #485, Keesler AFB, Miss.
Saifuddin, Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C.
Senecal, T. L., 5 Capitol Pl., Dayton 10, Ohio
Simons, R. F., 368 S. Maple St., W. Hempstead, L. I., N. Y.
Stavely, E. B., 203 Main Engineering Bldg., The Pennsylvania State College, State College, Pa.
Toole, P. C., Union Point, Ga.
Trinkle, W. S., 2324 Ripley St., Philadelphia 15, Pa.
Veneklasen, P. S., 639 West Foothill Blvd., Monrovia, Calif.
Wax, N., Department of Electrical Engineering, University of Illinois, Urbana, Ill.
Winer, J. D., 58 Barker Ave., Eatontown, N. J.
Wright, A., 1123 Maxine Dr., Fort Wayne, Ind.

The following admissions to Associate were approved, to be effective as of July 1, 1949:

Anderson, R. S., 11148½ Truro Ave., Inglewood, Calif.
Beckett, E. J., 4022 Monroe Ave., E. St. Louis, Ill.
Bell, V., 1649 Washington Blvd., Chicago 12, Ill.
Beltz, G. E., 33 Ninth St., McMechen, W. Va.
Bolsey, E. J., 57 Central Ave., Hartsdale, N. Y.
Borowski, E., 477 Second St., Brooklyn 15, N. Y.
Boscoe, A., 2072 W. Sixth St., Brooklyn 23, N. Y.
Cart, R. E., 4870 Sheridan Rd., Chicago 40, Ill.
Chapman, C. I., AFRS-WVTQ & WVTC, APO 25, c/o Postmaster, San Francisco, Calif.
Clause, E. M., 4411 S. E. Windsor Ct., Portland 6, Ore.
Cowgill, L. R., 5050 N. Broadway, Chicago 40, Ill.
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Dean, F. R., 268 Brookline Ave., Boston 15, Mass.
DeHart, W. D., 205½ Market St., Spencer, W. Va.
Dymek, B., 81 Endicott St., Worcester 4, Mass.
Eckert, G. F., 120 W. Schiller St., Chicago 10, Ill.
Faraday, B. J., Sound Division, Naval Research Laboratory, Washington 25, D. C.
Feld, M. M., 209 Ave. P., Brooklyn 4, N. Y.
Filley, F. R., 73 Grand Ave., Akron 2, Ohio
Fink, J. H., 230 W. Fifth St., Emporium, Pa.
Fletcher, C. H., 203 S. Church, Monroe, N. C.
Gillin, J. M., R.D. 1, Phoenix, N. Y.
Hardie, F., Hayestown Rd., Danbury, Conn.
Hoffman, P. L., Jr., 1225½ Cota Ave., Torrance, Calif.
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Kline, S. H., 1710-15 St., San Francisco, Calif.
Krute, E. H., 2349 Glenwood Dr., Port Arthur, Tex.
Kwiatek, W. K., 129 N. Blakely St., Dunmore, Pa.
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Maack, H. E., 3849 S. Albany Ave., Chicago 32, Ill.
McCasland, R. S., 4857 Schubert Ave., Chicago 39, Ill.
McGee, H. A., 3120 Hedgerow Dr., Dallas, Tex.
Monell, L. E., 3433 W. 59 Pl., Los Angeles 43, Calif.
Morrisset, J. B., Box 631, Lubbock, Tex.
Murphy, R. L., 925 Belden Ave., Chicago 14, Ill.
Musicaro, J. S., 1834-71 St., Brooklyn 14, N. Y.
Newhouse, P. D., 2955 N. Lowell Ave., Chicago, Ill.

(Continued on page 47A)

(Continued from page 43A)

Spandau, E. P., 801 Broad Blvd., Dayton 9, Ohio
Victoreen, J. A., 3800 Perkins Ave., Cleveland 14, Ohio

Transfer to Member

Bedwell, T. H., Physics Department, Florida State University, Tallahassee, Fla.
Bittner, B. J., 3110-42 Pl., Sandia Base Branch, Albuquerque, N. Mex.
Bodle, D. W., Bell Telephone Laboratories, 463 West St., New York 14, N. Y.
Brannen, P. M., 24½ N. Fifth St., Duquesne, Pa.
Florman, L. W., 413 N. Rush St., Itasca, Ill.
Gross, J., Box 973, Nairobi, Kenya Colony, British East Africa
Hancock, N. W., 600-13 St., N. W., Cedar Rapids, Iowa
Hoffman, R. Y., Jr., 1234 Ridgewood Dr., Northbrook, Ill.
Jorgenson, T. O., WEAU Transmitter, R. R. 1, Hy Q, Eau Claire, Wis.
Osbahr, B. F., 206 Eighth Ave., Brooklyn 15, N. Y.
Pfefer, B. L., 428 Woodbine Ave., Syracuse 6, N. Y.
Ratts, B. H., 2506 Terrace Rd., Fort Wayne 3, Ind.
Schull, G. R., 12131 Mayfield Ave., Los Angeles 24, Calif.
Waller, M. J., R.F.D. 1, Foxboro, Ont., Canada
Winkler, E. H., 71 White St., Shrewsbury, Red Bank, N. J.
Young, C. W., 4745 Nogal St., San Diego 2, Calif.
Zupnick, I. N., 768 Linden Blvd., Brooklyn 3, N. Y.

Admission to Member

Allan, D. K., 925 Louisiana Ave., Baton Rouge, La.
Anderson, C. W., 920 Pine St., St. Louis 1, Mo.
Baker, E. E., Jr., Box 379, APO 182, c/o Postmaster, San Francisco, Calif.
Belprez, G. R., 1322 Chalmers Ave., Detroit 15, Mich.
Bettis, E. S., Box 105, Oak Ridge, Tenn.
Bindner, J. T., 1586 Hedding Ct., San Jose 11, Calif.
Chess, R. B., 1830 South 54 Ave., Chicago 50, Ill.
Collier, J. W., 1903 Ridge Pl., Washington, D. C.
Cooke, H. F., Apt. 17, Maxwell Ct., Mains Ave., Syracuse, N. Y.
Dahl, A. H., 114 Disston Rd., Oak Ridge, Tenn.
Danielsen, A. C., 1429 South 51, Milwaukee 14, Wis.
Das, P. N., 1 Bhaduri Lane, P.O. Serampore, Dist. Hooghly, Bengal, India
Diehl, C. E., Box 1535, Salt Lake City 11, Utah
Frenkel, L. J., Jr., 216 South Yale, Albuquerque, N. Mex.
Hassel, E. W., 2911 Erie St., Racine, Wis.
Hershey, J. H., Washington Valley Rd., Morristown, N. J.
Hosker, G. R., c/o Richards-Wilcox Canadian Ltd., London, Ont., Canada
Kilbey, A. R., 20 Dix St., Waltham, Mass.
Layzell, L. M., Radio Planning Department, Flight Ops., Division, K.L.M. Royal Dutch Airlines, Schiphol Airport, Amsterdam, Holland
Lee, R. E., 1923 W. Baltimore St., Baltimore 23, Md.
Mahmoud, A. A., Faculty of Engineering, Fouad I University, Giza, Cairo, Egypt
Mallory, V. L., Continental Electronics Manufacturing Co., 1728 Wood St., Dallas, Tex.
Martens, H. A., 5675 Vreeland Rd., R.F.D. 2, Ann Arbor, Mich.
Michie, J. L., 23 Laburnum Dr., Skelmersdale, Lancs., England
Miller, S. S. 1540-52 St., Brooklyn, N. Y.
Mutter, W. E., Graduate House, Massachusetts Institute of Technology, Cambridge 39, Mass.

(Continued on page 45A)

MEMBERSHIP

(Continued from page 44A)

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Rawhouser, R., 65 W. Great Miami Blvd., Dayton 5, Ohio
Ridenour, P. W., OMR #485, Keesler AFB, Miss.
Saifuddin, Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C.
Senecal, T. L., 5 Capitol Pl., Dayton 10, Ohio
Simons, R. F., 368 S. Maple St., W. Hempstead, L. I., N. Y.
Stavely, E. B., 203 Main Engineering Bldg., The Pennsylvania State College, State College, Pa.
Toole, P. C., Union Point, Ga.
Trinkle, W. S., 2324 Ripley St., Philadelphia 15, Pa.
Veneklasen, P. S., 639 West Foothill Blvd., Monrovia, Calif.
Wax, N., Department of Electrical Engineering, University of Illinois, Urbana, Ill.
Winer, J. D., 58 Barker Ave., Eatontown, N. J.
Wright, A., 1123 Maxine Dr., Fort Wayne, Ind.

The following admissions to Associate were approved, to be effective as of July 1, 1949:

Anderson, R. S., 11148½ Truro Ave., Inglewood, Calif.
Beckett, E. J., 4022 Monroe Ave., E. St. Louis, Ill.
Bell, V., 1649 Washington Blvd., Chicago 12, Ill.
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Bolsey, E. J., 57 Central Ave., Hartsdale, N. Y.
Borowski, E., 477 Second St., Brooklyn 15, N. Y.
Boscoe, A., 2072 W. Sixth St., Brooklyn 23, N. Y.
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(Continued on page 45A)

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Newhouse, P. D., 2955 N. Lowell Ave., Chicago, Ill.

(Continued on page 47A)

MEMBERSHIP

(Continued from page 45A)

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Strom, L. D., 57-45 225, Little Neck, L. I., N. Y.
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Stubner, J. W., 2116 Grove St., Glenview, Ill.
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Veronica, D. J., 45 Nevada Ave., Buffalo, N. Y.
Walcher, J. F., 1525 Maple Ave., Wyoming, Ohio
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Werner, W. F., 72 St. James Pl., Brooklyn 5, N. Y.
West, R. E., 1638 Kepzie Ave., Chicago 47, Ill.
White, S. A. A., 74 Huron Ave., Cambridge, Mass.
Wisner, C. V., Jr., 6726 S. Oglesby Ave., Chicago, Ill.
Wodzinsky, W. T., 202 Broadway St., Carnegie, Pa.
Womack, R. H., 124A Grosvenor, Inglewood, Calif.
Zawada, J. D., Chicago Technical College, 2000 S. Michigan Ave., Chicago, Ill.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.
(Continued from page 24A)

Recent Catalogs

• • • A new 49-page catalog in color describing their complete line of relays, classified according to purpose and by contact ratings, by **Leach Relay Co.**, 5915 Avalon Blvd., Los Angeles 3, Calif.

• • • A listing of over 400 transformers and related components has just been released, and is available to interested firms, by **Standard Transformer Corp.**, Elston, Kedzie & Addisons Sts., Chicago 18, Ill.

(Continued on page 48A)

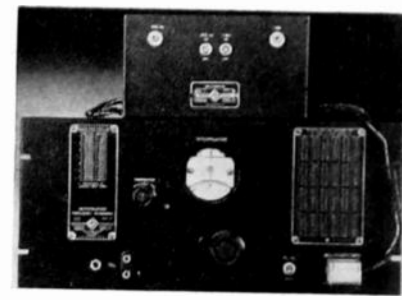
News—New Products

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

(Continued from page 47A)

Interpolator for Uhf Measurements

The Model 1110-A, Interpolating Frequency Standard, designed for use with heterodyne frequency meters in making measurements at uhf up to about 3,000 Mc, is announced by **General Radio Co.**, 275 Massachusetts Ave., Cambridge 39, Mass.



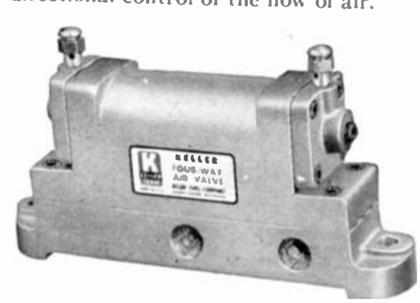
The manufacturing engineers claim that when it is used with a meter, whose accuracy is 0.1 per cent, the accuracy is increased to 0.001 per cent. The Model 1110-A was designed for use with General's Type 720-A heterodyne frequency meter which has a range of 100 to 200 Mc. Frequencies up to 3,000 Mc are measured by using harmonics of the 720-A.

An additional series of harmonics, based on 0.1 Mc fundamental, is also generated, for use with type 620-A meter, in measuring the range between 10 and 300 Mc.

Four-Way Air Valve

A new line of four-way air valves of the balanced piston type is announced by the **Keller Tool Co.**, Grand Haven, Mich.

The valve is actuated by poppet-type control buttons which exhaust air from either end of a balanced piston. The piston is used only as a means of operating a faced slide valve, which is the actual seal for directional control of the flow of air.



A few of the new design features incorporated are: a slide valve of oil resistant rubber (which, the manufacturer claims, have sealing surfaces that improve with use); a stainless steel piston for moving the slide valve; all pipe connections are located in the base, so that working parts are accessible without disconnecting the air lines.

News—New Products

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These air valves are available with ¼", ½", and 1" pipe threads, with capacities of 30, 90, and 350 cubic feet per minute, respectively.

They may be bench mounted, or as an integral part of a fixture with remote controls.

HF Voltmeter for TV, FM, and Radar

A new type, MV-18A, vacuum-tube voltmeter, which measures rf voltage down to 1 millivolt at frequencies between 1 and 200 Mc, has recently been developed by **Millivac Instruments,** P.O. Box 3027, New Haven, Conn.

In the previously mentioned range, it is flat within 10 per cent. When used for higher frequencies, larger response corrections have to be made. Range is 10 microvolts to 2,500 Mc.

For low-voltage measurements, the MV-18A uses germanium "pseudo-thermocuples" as a detector, and a carrier type dc amplifier which converts the dc voltages, into meter readings. Up to 1,000 volts, regular crystal diode rectification is used.

Direct TV and FM field-strength measurements, complete hf signal tracing through TV and FM receivers, at actual operating signal levels, and vhf and uhf laboratory research are applications of this new instrument.

New AF Bridge

The new Model 100 Bridger, an instrument for bridging a vacuum-tube voltmeter, distortion meter, and/or oscilloscope across any part of an af circuit through a shielded cable with none of the load of the meters or the cable on the circuit, is being marketed by **Audio Instrument Co.,** 1947 Broadway, New York 23, N. Y.

The Model 100 has an input impedance of 100 megohms in parallel with 6 $\mu\mu f$ when using 3' shielded input. The output is 200 ohms with one side grounded.

Voltage ratio: output/input is 0.98 (−0.2db) up to 30 volts with a low-capacity circuit, and up to 25 volts with regular shielded cable. This may be ex-

(Continued on page 55A)

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . . .

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
1 East 79th St., New York 21, N.Y.

RADIO ENGINEERS

Outstanding opportunity with progressive, rapidly growing company for radio engineers experienced in high frequency work. An interesting, attractive future assured, but requirements of jobs must be paralleled by necessary skill. If you have a background of high frequency work, tell us about yourself in résumé giving education, experience and salary requirements. Plant located in New Jersey within commuting distance of New York City. Box 569.

ELECTRONICS TEACHER

Electronics teacher to take charge of electronics option at accredited state land grant college in northwest. Salary to $4800.00 for nine months. Write giving picture, education, experience, references and complete personal data to Box 572.

ELECTRICAL ENGINEER - PHYSICIST

Graduate electrical engineer or physicist experienced in microwaves and servomechanism for design and layout of electronic circuits. Write Chance Vought Aircraft, Division United Aircraft Corp., Box 5907, Dallas, Texas.

INSTRUCTOR

There will be an opening in September for an instructor to teach electronics, transmission line and wave guide theory. Salary depends on qualifications and is up to $5600.00 for nine months. Possibility of later appointment at lower rank and salary. Box 573.

ENGINEER

Large active midwestern quartz crystal plant needs first class quartz crystal engineer, well grounded in theory and with complete experience in manufacturing and testing procedures all types quartz units. Send complete detailed information on experience and background in first reply. Salary open. Our employees know of this ad. Box 574.

TECHNICAL WRITER

An opportunity for an experienced technical writer. Preferably an electrical engineering graduate with experience in the field of electronic development. Research laboratory located adjacent to Washington, D.C. Salary commensurate with experience. Box 576.

Positions Wanted By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants, and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The Institute publishes free of charge notices of positions wanted by I.R.E. members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The Institute necessarily reserves the right to decline any announcement without assignment of reason.

ELECTRONIC ENGINEER

B.S.E.E. June 1949, Polytechnic Institute of Brooklyn. Single. 2 years experience as Navy electronic technician. Member Tau Beta Pi, Eta Kappa Nu. Desires position in electronic field; New York metropolitan area preferred. Box 265 W.

FIELD SALES ENGINEER

B.S.E.E. communications major. 5 years experience in microwave, radar, and general communication equipment. Also some aircraft and electronic control experience. Finest references. Box 266 W.

COMMUNICATIONS ENGINEER

B.S.E.E. June 1949, University of Minnesota, communications major. Age 25. Married, no children. 2 years experience GCA and other radar, 2 years airline radio and telephone work. Desires position with future in TV or electronics. Box 267 W.

TELEVISION ENGINEER

Graduated American Television Institute of Technology with B.S.T.E. Age 22. Single. Desires position in design and development work in vicinity of Boston, Mass. Box 269 W.

ENGINEER

June 1949 honor graduate of leading midwest engineering college. B.S.E.E. Harvard-M.I.T. radar course. 3 years experience as research and development officer. MOS 7050. Experience with electronic bombing computers and telephone subs. equipment. Tau Beta Pi, Eta Kappa Nu—President, Student AIEE. Prefer connection with quality conscious firm in southwest U.S.A. Box 270 W.

ELECTRONIC ENGINEER

Ex-flying officer. Industrial engineering and electronic engineering background. Both research and development and manufacturing experience on airborne radar, television and guidance systems. Schools—NCE, University of Dayton, Harvard, Columbia and M.I.T. Box 271 W.

TELEVISION ENGINEER

Graduated American Television Institute of Technology January 1949 with B.S.T.E. Age 23. 1st class FCC license.

(Continued on page 52A)

Positions Wanted

(Continued from page 51A)

2 years Army radio operator mechanic. Desires position as development or TV station engineer. Thorough understanding of RCA and Dumont TV equipment. Box 272 W.

ADMINISTRATIVE OR EXECUTIVE

Retired Naval officer. M.S. Radio Engineering; Harvard 1928. Varied experience in charge reorganizing and operating major units Naval Communication System, radio procurement, electronics planning, radio stations maintenance, etc. Desires position as administrator or executive in communications, radio, or electronics, anywhere in U. S. Available immediately. Box 273 W.

JUNIOR ENGINEER

B.S. Radio Engineer, Chicago Technical College; Member Sigma Phi Delta, March 1949. Age 23; single. Navy experience Electrical Maintenance Technician. Desires position with a good opportunity. Box 274 W.

ENGINEER

M.S.E.E. June 1949. California Institue of Technology; age 23; married. 2 years Navy radio technician. Summer work in small motor design. Member Tau Beta Pi. Desires work in control or telemetering. Box 275 W.

ELECTRONIC ENGINEER

B.E.E. New York University 1948, electronics major. 1 year GE test experience. GE advanced course in engineering. Interested electronic control equipment, servo-mechanisms; computers. New York Metropolitan area. Box 276 W.

ENGINEER

B.E.E. June 1949. Cooper Union. Age 28. Married. 3 years electrical testing and inspection experience including supervision. 3 years Army Signal Corps as radio serviceman. Desires position with future. Box 277 W.

EXECUTIVE ASSISTANT

Business administration plus engineering training. M.B.A. University of Chicago; B.S.E.E., B.S. (math), University of Michigan. Eta Kappa Nu. Single. Age 26. Signal Corps technician; 1 year research, production quality control experience. Desires job in medium size concern. Résumé upon request. Box 278 W.

ENGINEER

B.S.E.E. September 1949, University of Florida. Age 30. Married. 2 years Navy radio-radar technician and school. 1 year Chief Engineer 250 watt radio station. Motor and power transformer experience. Interested sales or construction. Will travel anywhere. Box 279 W.

JUNIOR ENGINEER - PHYSICIST

5½ years of physics, electrical engineering and electronics in Fordham University of Rochester, Harvard, M.I.T. 3 years Naval electronics officer; B.S. in physics, June 1949. Age 25; married; 1 child. Desires employment in electronic engineering, sales or development, anywhere, preferably west coast. Box 280 W.

SALES ENGINEER OR EXECUTIVE ASSISTANT

Former assistant to Vice President of a world wide export organization invites

(Continued on page 53A)

Positions Wanted

(Continued from page 52A)

inquiries from firms having need for addition to sales engineering staff or assistant to top executive. Young, energetic, widely traveled with a solid background of research design and supervision in all phases of electrical and electronic equipment. Capable of handling sales, engineering, production and personnel. Box 281 W.

ELECTRONICS ENGINEER

Guided missile electronics engineer. Currently engaged in production engineering of guided missiles. 3 years experience in development of guidance and control equipment. 4 years radar development experience for Army. Graduate engineer with post-graduate school. Box 282 W.

ENGINEER

Graduate of American Television Institue of Technology, May 1949 with B.S.T.E. Age 24. Single. Presently engaged in post-graduate work in the U.H.F. field. Desires laboratory work in the New York City suburban area. Box 298 W.

ENGINEER

B.E.E., June 1949, Polytechnic Institute of Brooklyn, Tau Beta Pi, Eta Kappa Nu. Age 25. Married. 2½ years in Army Signal Corps. Desires position in development or product engineering. Prefer New York location but will consider work anywhere in United States. Box 299 W.

TECHNICAL WRITER

Development engineer available for writing assignments in radio, television, and allied arts. Technical articles, instruction booklets and similar work treated with strictest confidence on commission basis. Box 300 W.

FIELD ENGINEER

Currently engaged U.S.N. field engineering. For past 9 years have been supervising crews for installation and maintenance of Naval Radar, Sonar, V.H.F., H.F., and communication equipment (receivers and transmitters). Graduate of Naval Matériel School, Washington, D.C., B.S. in Applied Electronics, St. Louis University. Age 30. Married, 3 children. Available June 30, 1949. Box 301 W.

ELECTRONICS ENGINEER

B.S. in Radio Engineering, and graduate work in Electrical Engineering. Former communications and fighter control officer. Desires research in electronics or electrical engineering. Married, 1 child. Location immaterial. Box 302 W.

ENGINEER MANAGER

Harvard M.S. in communication engineering. Directed important and extensive wartime Loran project in Pacific. Recently Development Division Engineering Manager. Desires permanent and responsible position with progressive organization. Box 303 W.

RADIO ENGINEER

Radio Engineer, 24, single, Inter. B.Sc., 3 years experience in electronic research and development, left England July, seeks position in electronic laboratory of American university with view of continuing study. Box 304 W.

RADIO ENGINEER

B.S. in radio engineering, one course to complete for B.S.E.E. Married, no children. Age 26. 2½ years experience telephone carrier and VHF installation

(Continued on page 54A)

Positions Wanted

(Continued from page 53A)

and maintenance. Desires research or design position anywhere in United States. Box 305 W.

TELEVISION ENGINEER

Graduate American Television Institute of Technology May 1949 with B.S.T.E. Age 35, married, 1st class F.C.C. license 5 years electrical maintenance, 3 years radar maintenance, 3 years radio servicing. Desires positions as TV station engineer. Prefer east or mid-west. Box 306 W.

ELECTRONICS ENGINEER

B.S. physics. Age 27. Married, no children. 2 years industrial electronics research and development. Some guided missile development. 3 years Army Radar developing and maintainance. Anywhere in United States, will also consider foreign position. Box 307 W.

ELECTRONICS DESIGN ENGINEER-ELECTRONICS INSTRUCTOR

B.Sc. 1937 University of Chicago. Age 32. Married. 3 years experience in design of radar equipment with eastern manufacturer. 2½ years experience in Navy as radar maintenance officer. Have had teaching experience in physics. Desires position as electronic design engineer in midwest, or electronics instructor in midwestern college. Box 308 W.

ENGINEER

B.E.E. 1948 electronics option Georgia School of Technology. Age 22. 1½ years electronic technician U. S. Navy. 1st class radio telephone F.C.C. license. 1 year industrial experience in design and development of radar components. Member Eta Kappa Nu and Tau Beta Pi. Desires work in development. Box 309 W.

ELECTRONIC ENGINEER

B.S. in E.E., with high distinction, University of Connecticut, June 1949. Experience: 1 year teaching, 4 years electronic technician. Age 28. Research or development position desired. M. Cannizzaro, 22 Division St., Waterbury, Conn.

JUNIOR ENGINEER OR TECHNICAL WRITER

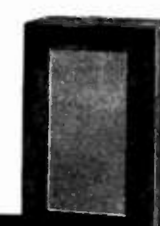
B.S.E.E. June 1949 Illinois Institute of Technology. Single, age 23. Desires position in sales engineering or production in in United States. Willing to travel. 1½ years experience as Navy electronic technician, power plant operator, and electronic inspector. Excellent references. Box 310 W.

News—New Products

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

Recent Catalogs

• • • The May issue of the *Research Worker* describes three popular types of power supplied for cathode-ray tube high voltage. This paper is printed by Aerovox Corp., New Bedford, Mass., from whom a subscription may be obtained by application.



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 49A)

tended to 250 volts by adding Model 104 voltage divider, which has a 10 to 1 ratio.

The cable capacity is almost balanced out by the circuit configuration, and although the capacity is low, the cable retains flexibility.

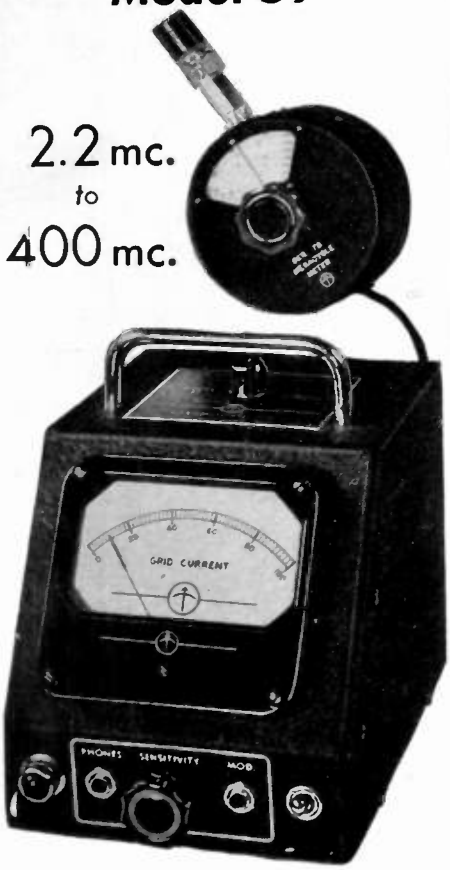
Super Midget Relay

A new low-cost super mdiget relay, Model SM, in which the magnetic circuit elements all perform multiple functions, permitting an appreciable reduction in size without loss of operating efficiency, has been designed and manufactured by Potter & Brumfield Sales Co., 549 W. Washington Blvd., Chicago 6, Ill.

Dimensions of the open SM are 5/8" diameter, by 1 5/16" over-all length. Windings available up to 3,400 ohms, permitting operation up to 75 volts dc with minimum sensitivity of 5 ma at 80 milliwatts. Maximum coil dissipation is 1.75 watts at 83°C rise. Contacts are silver, and rated at 2.5 amperes for 100 operations, 1 ampere for 50,000, or 0.25 amperes for continuous operation on 115 volt 60 cps noninductive load.

In variance to the open type, the SM is available dust sealed in polystyrene 5-pin plug-in enclosure, or hermetically sealed in glass with a miniature 7-pin tube base. The latter can be evacuated and gas filled if desired.

(Continued on page 57A)



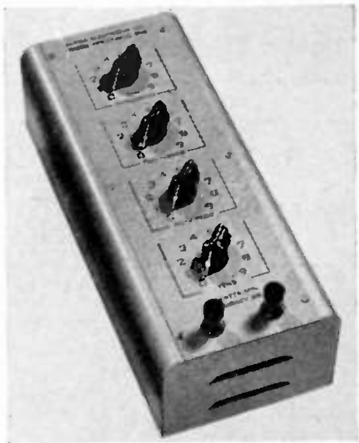
News—New Products

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

(Continued from page 55A)

Decade Resistance Box

A new Model 10 decade power resistance box, suitable for laboratory and test applications, has been developed by **Marma Electronic Co.**, 1632-36 N. Halstead St., Chicago 14, Ill. (a new division of Fidelity Amplifier Co. of the same address).



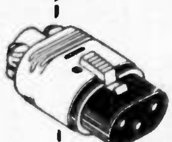
The manufacturer states that the combination of power handling capacity, 2 per cent accuracy, and wide range afforded by a four-decade box makes it practical for replacing the accurate ½ per cent type used as standards, as these cannot be subjugated to any appreciable load without damage.

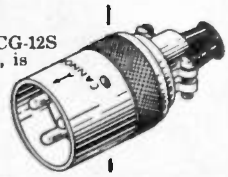
This box will dissipate a minimum of 10 watts, and maximum of 30 watts, depending on setting. Four decades allow for flexibility and provide for a total available resistance of 99,000 ohms. Low inductance lends accuracy of readings on all af and the lower rf. The case is metal and may be grounded, providing additional shielding when necessary.

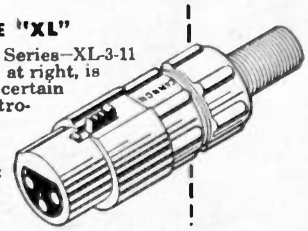
Two New Ballast Tubes

A subminiature for minute applications, and an aircraft ballast tube to withstand the vibrations of airborne radio equipment, have been developed by **Amperite Co., Inc.**, 561 Broadway, New York 12, N. Y.

(Continued on page 59A)







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 57A)

On the right is the subminiature. This tube was designed without prongs, eliminating the necessity of a base. The leads are soldered directly to the leads from the tube.

This tube can be supplied to dissipate any wattage up to 3 watts. Maximum current 0.9 amperes. A 100 per cent increase in voltage will produce a change of less than 5 per cent. An ambient change of $-50°C$ to $+70°C$ will cause a difference in current of less than 2 per cent.

On the left is the aircraft type, which will withstand 40 g, and reacts identically to temperature extremes as the tube described above.

This tube can be supplied to dissipate up to 25 watts, from 60 ma to 3 amperes.

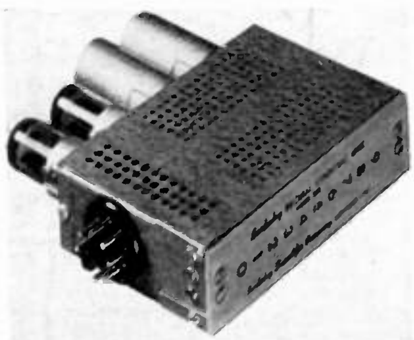
AS EXHIBITED AT THE

SPOTLIGHT THE NEW...

RADIO ENGINEERING SHOW

Electronic Decimal Counting Unit

For use in industrial and commercial applications, the Model 700 Decimal Counting Unit was recently marketed by **Berkeley Scientific Co., 6th St. & Neven Ave., Richmond, Calif.**



To be used primarily in counting, timing, and frequency measurement work, the Model 700 is a four-tube plug-in unit with a scale-of-ten counting circuit capable of rates in excess of 40,000 pulses per second, and of resolving pulse pairs spaced less than 5 microseconds apart.

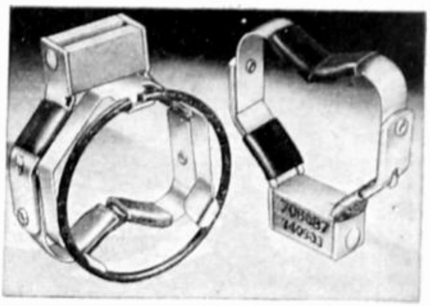
Counts are read from 10 neon lamps on the panel numbered 0 to 9. Each count is indicated directly by illuminating the lamp corresponding to the pulses received. Units may be cascaded to count any number. Thus, by mounting several next to each other, counts may be read directly.

(Continued on page 60A)

(Continued from page 59A)

TV-Tube Beam Benders

Two new types of TV-tube beam benders, which serve to minimize burnt spots on screens, are announced by **Clarostat Mfg. Co., Inc.** Washington St., Dover, N. H.



The series TV-2 has a single permanent bar magnet and is used with 10" tubes with flux densities across the poles of 33 ± 3G and 75 ± 10G.

The series TV-3 is higher in cost, but features two magnets: the bar magnet for the rear and a ring magnet for the front elements. This type is used with 12" and larger tubes, particularly those of bent-gun design.

Both types have rubber covered spring arms for friction fit on 1⅜" to 1½" necks.

(Continued on page 61A)

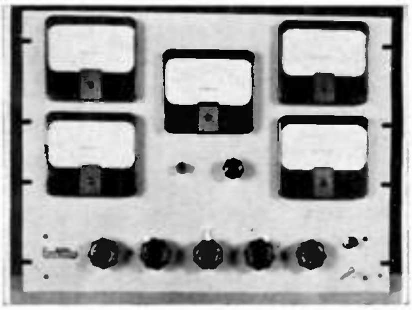
News—New Products

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

(Continued from page 60A)

New Phase Relation Meter

A new Model 108-C Phase Meter, which provides a means for measuring the phase relations existing in directional antenna systems, has been designed and developed by **Clarke Instrument Corp.**, 910 King St., Silver Spring, Md.



With the Model 108-C, provision has also been made for remote monitoring of amplititudes of the currents in the several elements of the array. The phase indication is marked in 2° intervals, however, ½° increments can readily be resolved.

Although intended for operation in the standard band, this meter can be supplied in other frequency ranges.

PROFESSIONAL CARDS

INDEX AND DISPLAY ADVERTISERS

DISPLAY ADVERTISERS

Cornell-Dubilier capacitors might look like others... *but differ where it counts!*

That there's more than meets the eye—when it comes to capacitors — is a fact well known to radio engineers for many years. Anyone who knows his way around in the industry, as you do, is not fooled for a moment by external appearance. It's what's *inside* that counts—which is why you can count on Cornell-Dubilier.

Engineers specify C-D because over a period of 40 years they have learned they can count on C-D capacitors for complete dependability, for long years of trouble-free performance, for really genuine economy. Perhaps that's why an impressive percentage make it a point to specify C-D's. Inquiries cordially invited. Catalog available on request.

Cornell-Dubilier Electric Corporation, South Plainfield, New Jersey, Dept. M. Other plants in New Bedford, Brookline and Worcester, Mass.; Providence, R. I.; Indianapolis, Ind., and subsidiary, The Radiart Corp., Cleveland, Ohio.

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