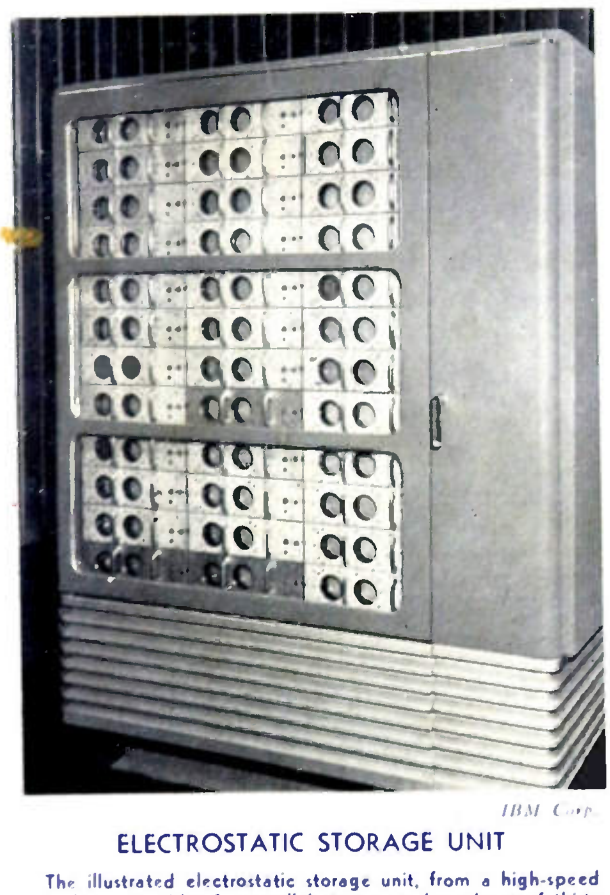
Proceedings of the I·R·E

KREVSKY

A Journal of Communications and Electronic Engineering

October, 1952

Volume 40 Number 10



IBM Corp.

ELECTROSTATIC STORAGE UNIT

The illustrated electrostatic storage unit, from a high-speed calculator, provides for parallel storage and read-out of thirty six-bit binary words. In it are three-inch cathode ray tubes of special design.

PROCEEDINGS OF THE I.R.E.

Table of Contents, Indicated by Black-and-White Margin, Follows Page 64A

The Institute of Radio Engineers

Meetings with Exhibits

● **As a service both to Members and the industry, we will endeavor to record in this column each month those meetings of IRE, its sections and professional groups which include exhibits.**

Δ

October 29-November 1
Audio Fair Hotel New Yorker, New York, N.Y.
Exhibits: Harry N. Reizes, 67 West 44th Street, New York, N.Y.

Δ

December 10, 11 & 12, 1952
Joint IRE-AIEE Computers Conference Park Sheraton Hotel
Exhibits: Perry Crawford, 373 Fourth Avenue, New York City.

Δ

January 26, 27, 1953
1953 7th Regional IRE Conference, University of New Mexico, Albuquerque, N.M.
Exhibits: Hoyt Westcott, 107 So. Washington St., Albuquerque, N.M.
Chairman: C. W. Carnahan, 3169 41st Place, Sandia Base, Albuquerque.

Δ

February 5, 6 & 7, 1953
Southwestern IRE Conference Plaza Hotel, San Antonio, Tex.
Accept Exhibits

Δ

March 23, 24, 25 & 26, 1953
Radio Engineering Show Grand Central Palace, New York City
Exhibits Manager: Wm. C. Copp, 303 W. 42nd St., New York 36, N.Y.

Δ

April 11, 1953
NEREM—New England Radio Engineering Meeting, University of Connecticut, Storrs, Conn.
Exhibits: H. W. Sundius, The Southern New England Tel. Co., 227 Church St., New Haven, Conn.

Δ

April 18, 1953
Spring Technical Conference of the Cincinnati Section, Cincinnati, Ohio
Chairman: J. P. Quitter, B. Baldwin Company, 1801 Gilbert Ave., Cincinnati 2, Ohio

Δ

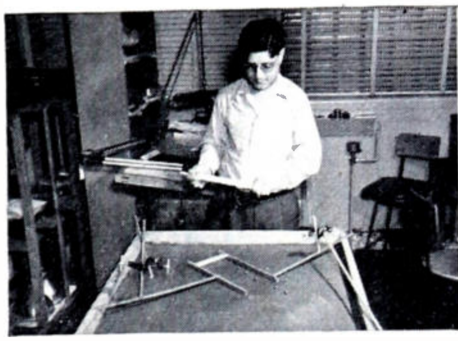
May 11, 12 & 13, 1953
National Conference on Airborne Electronics Hotel Biltmore, Dayton, Ohio.
Exhibits: Paul D. Hauser, 1430 Gascho Drive, Dayton 3.

What Rauland means by "Perfection Through Research"

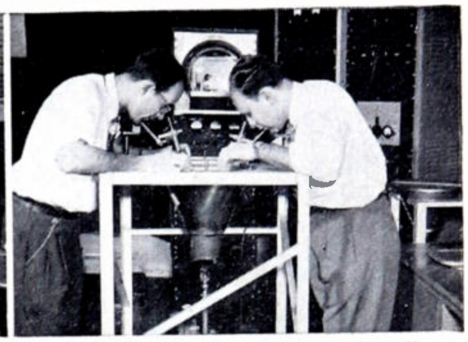
Rauland is one of the few companies devoting so much top engineering talent full time to picture tube improvement and perfection.

The result has been to give you more picture tube advancements since the war than any other manufacturer . . . first chance at the latest developments for companies using Rauland tubes as original equipment . . . and a real selling edge at the retail level because of the extra satisfaction which Rauland advantages offer.

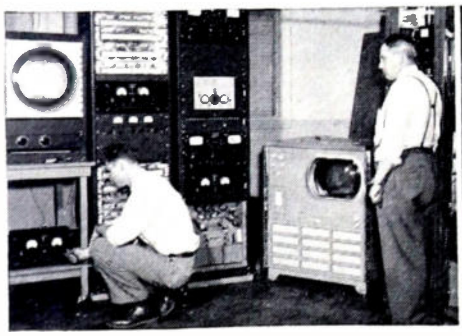
That's why so many alert manufacturers look to Rauland for the best in picture tubes.



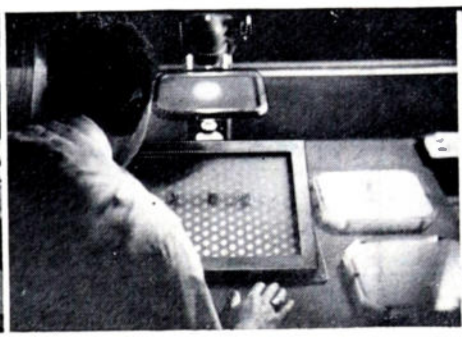
Rubber model for studying electron optical designing—basis for Rauland's exclusive Indicator Ion Trap.



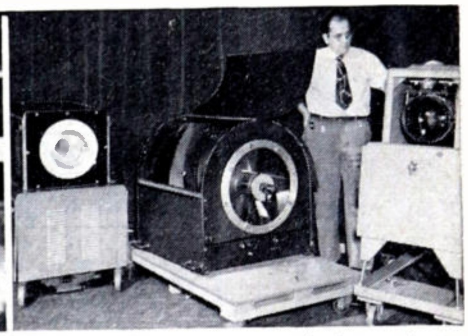
Alignment of the screen and parallax mask of tri-color tube containing approximately a million fluorescent dots.



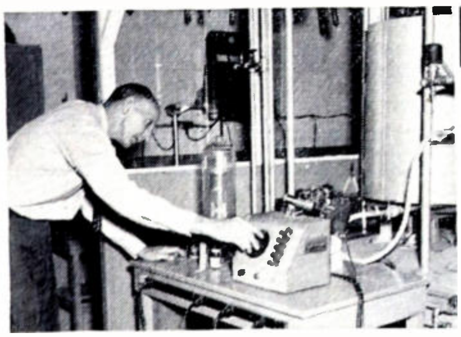
All-electronic tri-color tube in electronic receiver system (left) in comparison with mechanical system (right).



Inspection and checking of perforations .0075" in diameter in masks of tri-color picture tubes.

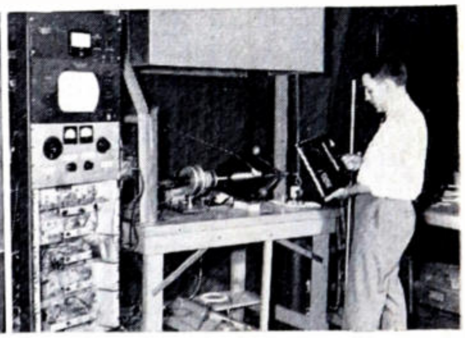


Rauland large-screen projectors using three different optical systems, all of which give theater-size pictures.



Careful study of the formation of thin metallic films in a vacuum . . . basis for the aluminizing of tubes.

Examination with polarimeter permits careful control of strains for superior glass-to-metal sealing.



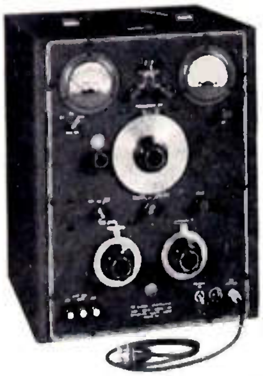
A physicist using a Rauland-developed radiation meter in checking X-ray radiations from cathode ray apparatus.

THE RAULAND CORPORATION

Perfection Through Research

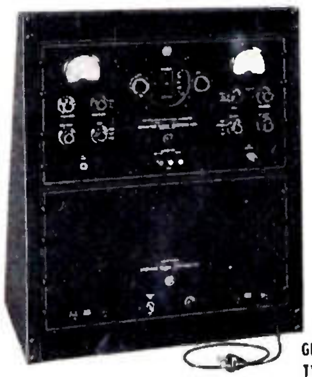
4245 N. KNOX AVENUE • CHICAGO 41, ILLINOIS

PRECISION ELECTRONIC MEASURING INSTRUMENTS

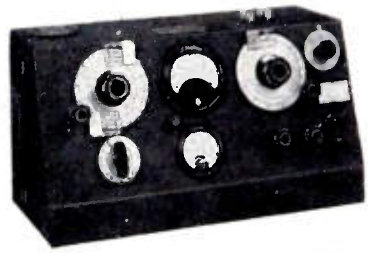


FM-AM SIGNAL GENERATOR TYPE 202-B

Q-METER TYPE 190-A



SIGNAL GENERATOR TYPE 211-A



Q-METER TYPE 160-A

Signal Generators

Frequency and Amplitude Modulated
Crystal Monitored
For Mobile Communications Receivers

Q Meters

Low, Medium and High Frequencies
Very High Frequencies

Univerters

Low, Medium and High Frequencies
VHF for Glide Slope Receivers

Write for complete information

Q METERS

Type	Freq. Range	Q Range	Tuning Capacity Range	Q Accuracy	Price
160-A	50 kc to 75 mc	20 to 625	30–450 mmf	5% to 30 mc	$625.
190-A	20 mc to 260 mc	5 to 1200	7.5 to 100 mmf	5% to 100 mc	$625.

FM-AM SIGNAL GENERATORS

Type	Freq. Range	Output Range	Modulation FM	Modulation AM	Application	Price
202-B	54–216 mc	0.1 to 200,000 µv	0–240 kc	0–50%	General	$975.
202-C	54–216 mc	0.1 to 200,000 µv	0–240 kc	0–50%	Mobile	$1090.
202-D	175–250 mc	0.1 to 200,000 µv	0–240 kc	0–100%	Telemetering	$980.

FM SIGNAL GENERATOR (For Mobile Communications Receivers)

Type	Freq. Range	Output Range	Modulation	Application	Price
206-A	146–176 mc	0.1 to 200,000 µv	0–250 kc	Mobile	$910.

OMNI-RANGE SIGNAL GENERATOR (Crystal Monitored)

Type	Freq. Range	Output Range	Modulation	Application	Price
211-A	88–140	0.1 to 200,000 µv	0–100% am	Omni-Range Rcvrs.	$1800.

GLIDE SLOPE TEST SET

Type	Freq. Range	Output Range	Modulation	Accessory to	Price
212-A	329–335 mc	0.1 to 200,000 µv	0–100% am	211-A	$875.

UNIVERTERS

Type	Freq. Range	Output Range	Modulation FM	Modulation AM	Accessory to	Price
207-A	0.1 to 55 mc	0.1 to 100,000 µv	0–240 kc	0–50%	202-B 202-C	$345.
207-B	0.1 to 55 mc	0.1 to 100,000 µv	0–240 kc	0–50%	202-D	345.
207-C	0.1 to 50 mc	0.1 to 100,000 µv	0–240 kc		206-A	345.

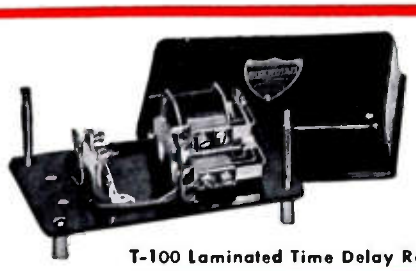
BOONTON RADIO *Corporation*

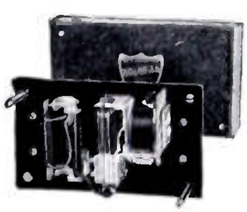
BOONTON·N·J·U·S·A·

Relays BY GUARDIAN

COVER A WIDE RANGE OF APPLICATIONS . . .

In equipment that guides warplanes in flight or conditions personnel with radio diathermy—"Relays by Guardian" are most popular control units. For example, radio diathermy machines usually employ a time delay relay such as the Guardian T-100 to "warm up" the filaments of oscillator type tubes. The time delay is adjustable between 10 and 60 seconds; contact capacity is 1500 watts on 110 v., 60 cycles non-inductive A.C. Power Consumption of coil and time delay during closing of thermostat blade is approximately 10 VA and after closing, 5.5 VA. A similar relay, the Guardian T-110 may be equipped with an extra set of open or closed contacts. Both relays may be used in applications requiring the changing of circuits after a predetermined interval. Consult Guardian wherever automatic control is desired for making, breaking, or changing the characteristics of electrical circuits.

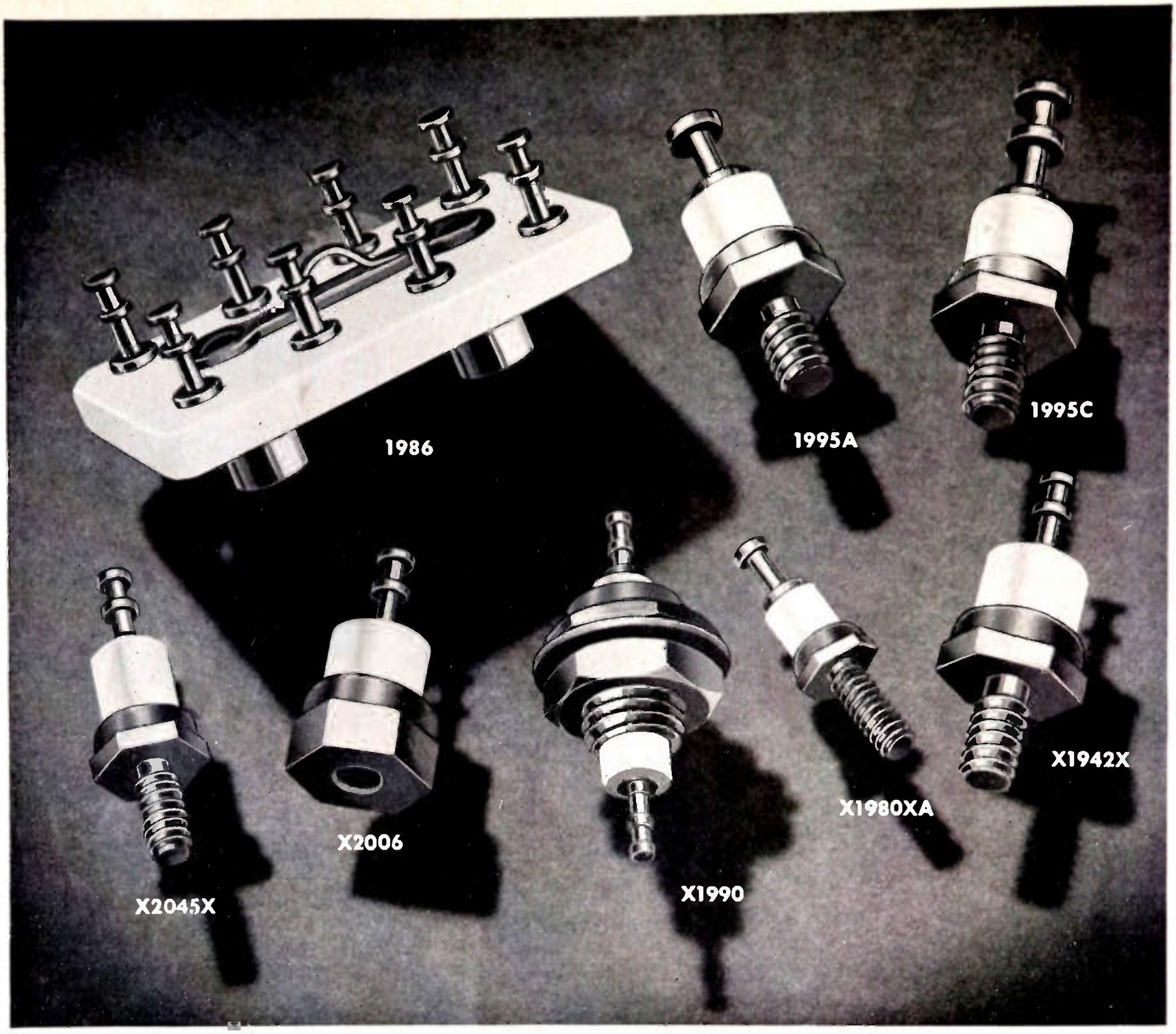

T-100 Laminated Time Delay Relay


T-110 Time Delay Relay (not laminated)

GUARDIAN ELECTRIC

1628-L W. WALNUT STREET CHICAGO 12, ILLINOIS

A COMPLETE LINE OF RELAYS SERVING AMERICAN INDUSTRY



Units shown magnified approximately 2½ times

Make sure of meeting government "specs" . . . see C.T.C. for ceramic insulated components

You have to be 100% on-the-beam if your equipment is to withstand the conditions it must undergo in military service.

That's why manufacturers using electrical and electronic components turn to C.T.C. for their ceramic insulated units. Our long experience and constant dealing with government requirements have gained us a wide acceptance as an outstanding supplier to those working on U. S. contracts . . . especially for the armed forces.

Whatever your needs in ceramic insulated terminals, feed-throughs or terminal boards you can depend on C.T.C. We meet the most exacting government standards for materials, tolerances, finishes, moisture prevention and anti-fungus treatment. Finishes on metal surfaces for instance, can be hot tinned, electro-tinned, cadmium plated, silver plated or gold plated to your requirements. All ceramic units in our standard line are grade L-5, silicone impregnated.

C.T.C. offers a consulting service at no extra charge to help solve your special problems. For all specifications and prices, write to Cambridge Thermionic Corporation, 456 Concord Avenue, Cambridge 38, Mass. West Coast Manufacturers contact: E. V. Roberts, 5068 West Washington Boulevard, Los Angeles 16 and 988 Market Street, San Francisco, California.

CAMBRIDGE THERMIONIC CORPORATION

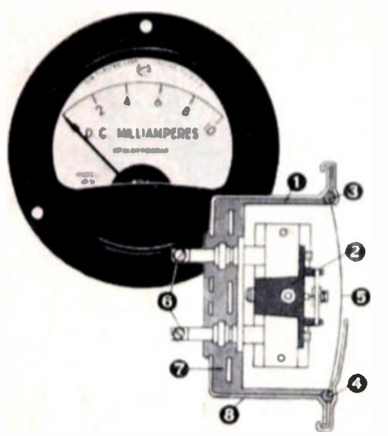
custom or standard . . . the guaranteed components

NEWS and NEW PRODUCTS

October 1952

Meters

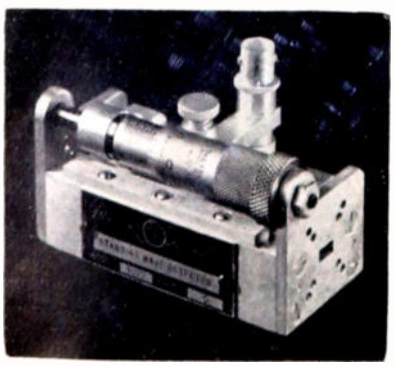
Sun Electric Corp., 6323 Avondale Ave., Chicago 31, Ill., announced today that they are now producing meters to meet Military Specification MIL-M-10304.



Some of the features of the new "Ruggedized" Meters are: Meter movement shock mounted and housed in rubber-lined case, observation window rubber grommeted and sealed to rubber lining of case providing hermetical seal of high dielectric materials; terminals side-tapped and provided with tinned binding screws to facilitate wiring with or without wire lugs or by pressure, soldering or both. Meters are available as dc voltmeters, ammeters, milliammeters, and microammeters and also may be ordered as rectifier type ac instruments.

Standing Wave Detector

A new standing-wave detector (Type 1022) has been developed by **Microwave Associates, Inc.,** 22 Cummington St., Boston 15, Mass. It is designed for precision low-level impedance measurements in the millimeter region when used with a suitable source and amplier. VSWR's as low as 1.01 can be read accurately in the region from 34 to 36 kmc.



The unit consists of a slotted section of RG 96/U waveguide milled from a solid piece of brass and silver plated. A movable carriage is provided carrying a spring loaded adjustable coupling probe, a silicon diode detector socket, and coaxial output fitting which will mate with a UG-88/U or equivalent BNC cable connector. This carriage rests on a semicylindrical polished bearing surface and is positioned by a micrometer drive. A total longitudinal probe displacement of 0.750 inch is available. The probe position can be read accurately to 0.001 inch. The total insertion length of the unit is 3.25 inches.

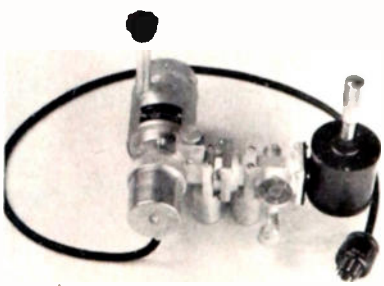
For flexibility two sets of tapped holes are available to mate with UG/U and 600/U waveguide connectors or Microwave Associates' special waveguide choke flanges P 1001.

A high level version of this standing-wave detector, as well as a roving stub tuner of similar design, are also available.

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

Spectrum Analyzer Adapter

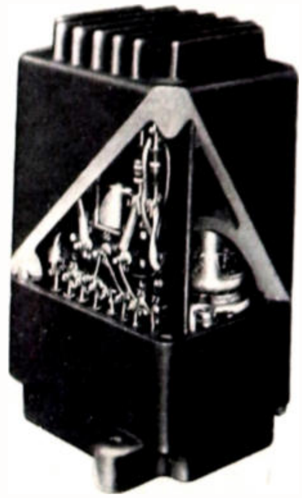
Microwave Associates, Inc., 22 Cummington St., Boston 15, Mass., has made available its Spectrum Analyzer Adapter, Type P-530. This adapter consists of a complete set of rf plumbing to convert any existing S or X band spectrum analyzer to the frequency range centered at 35,000 mc. Although the unit is designed for primary use from 34 to 36 kmc., it will operate satisfactorily over a substantially larger range. The input sensitivity will vary with the noise figure of the spectrum analyzer receiver with which it is used. However, threshold sensitivities of 70 db below 1 milliwatt are to be expected.



The unit is composed of a 2K25 klystron operating at X-band and powered from the spectrum analyzer power supply. The oscillator output is fed through a variable attenuator to a mixer where its fourth harmonic is mixed with the incoming millimeter signal. The signal input enters through a UG-381/U connector and RG-96/U waveguide. The I F output of the mixer is a type N connector. A calibrated X-band reaction type wavemeter is provided which covers the X-band fundamental of the 2K25. A calibration reading of the wavemeter is also supplied.

Plug In Units

The development of Unistage is announced by **Technical Development Corp.,** 4058 Ince Blvd., Culver City, Calif. Unistage is a self contained unit assembly with all components necessary to a functional circuit. Sizes are available to accommodate any combination from 1 tube to 4 tube circuits.



Unistage is offered to the design engineer in its basic form, so that any desired circuitry may be installed. This basic unit comprises essentially: the die cast aluminum housing; a terminal board having a large number of single and through terminals which are coded for easy assembly of components; the tube plate which allows the use of standard miniature and noval sockets; and the tube well or wells. Production and process engineering information is included. Technical Development Corp., also custom-manufactures completed Unistage units with the specified circuitry.

Half-Wave Vacuum Rectifier

Tube Department, Radio Corp., of America, 415 S. Fifth St., Harrison, N. J., recently announced a half-wave vacuum rectifier tube (6AX4-GT) of the heater cathode type. It is intended particularly for use as a damper tube in horizontal deflection circuits of television receivers.

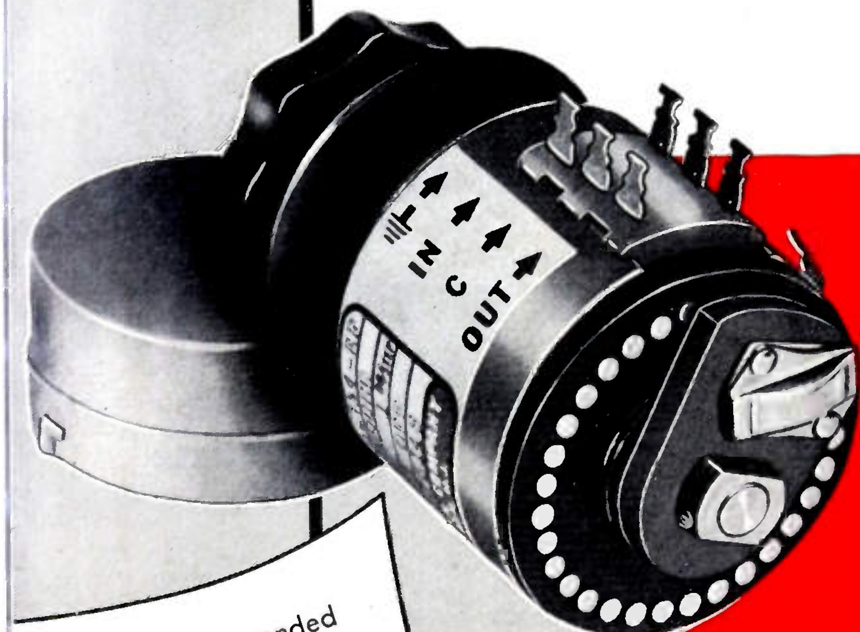
Designed to withstand negative peak pulses between heater and cathode of as much as 4000 volts with a dc component up to 900 volts, the 6AX4-GT provides flexibility in choice of deflection circuits.

(Continued on page 44A)

THESE ARE THE

REASONS

WHY DAVEN IS THE LEADER IN THE AUDIO ATTENUATOR FIELD



LA-130 Series,
Ladder Network
30 Steps, 1¾ Diameter

1. "KNEE-ACTION" ROTOR . . . gives positive contact and low contact resistance under all conditions.
2. BRASS CASE OF 2-PIECE CONSTRUCTION . . . gives excellent shielding and allows more compact equipment design.
3. "LOCK-TITE" DUST COVER . . . designed to withstand severe vibration and at the same time allow easy accessibility.
4. ENCLOSED ROLLER-TYPE DETENT MECHANISM . . . gives positive indexing, but does not increase depth of unit.
5. LOW-LOSS MOLDED TERMINAL BOARD . . . has high resistance to leakage.

Greatly expanded production facilities enable Daven to make delivery from stock on a large number of standard attenuator types.

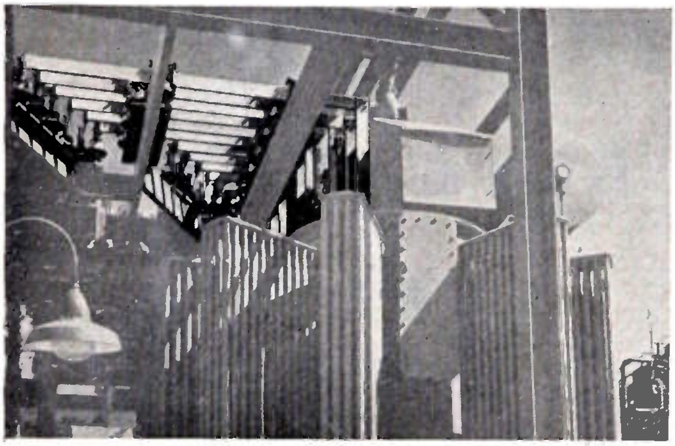
Write for complete catalog data.

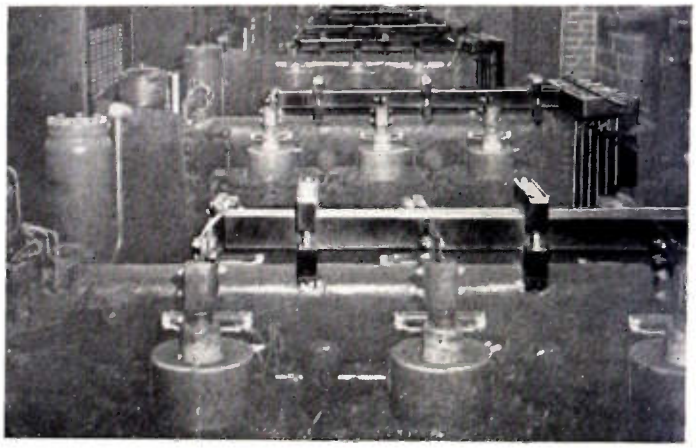
THE DAVEN CO.

WORLD'S LARGEST MANUFACTURER OF ATTENUATORS

195 Central Ave., Newark 4, N. J.

It takes a Lot of Copper to make a Lot of Chemicals!





These photographs show a lot of copper bus bar in a new plant of a great chemical company, whose name and location cannot be disclosed. The copper carries heavy currents to electro-chemical equipment for the production of valuable products used in national defense and in industry. Revere furnished 325,000 pounds of bus bar for this service, the bar going into substations, rectifier stations, and cell houses. In addition, at the time of installation the Revere Technical Advisory Service collaborated with the customer in working out some difficult details in the design of switches. If you need electrical conductors, remember that copper has the highest electrical conductivity of all the commercial metals, that Revere makes bus bar, and that the Revere Technical Advisory Service is always ready to work with you on any problem concerning copper and its alloys or aluminum alloys. Call the nearest Revere Sales Office.

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COPPER AND BRASS INCORPORATED

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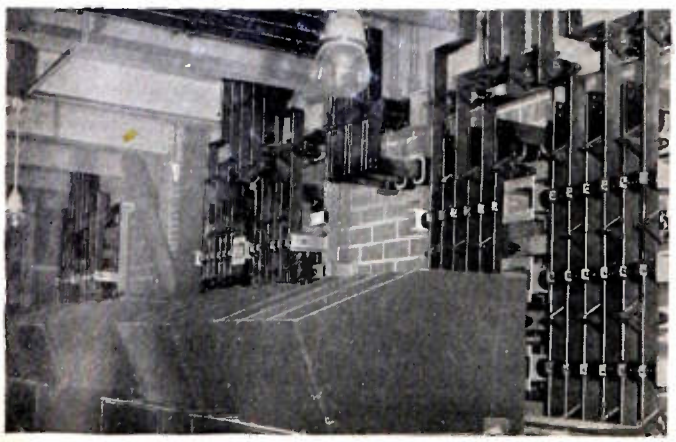

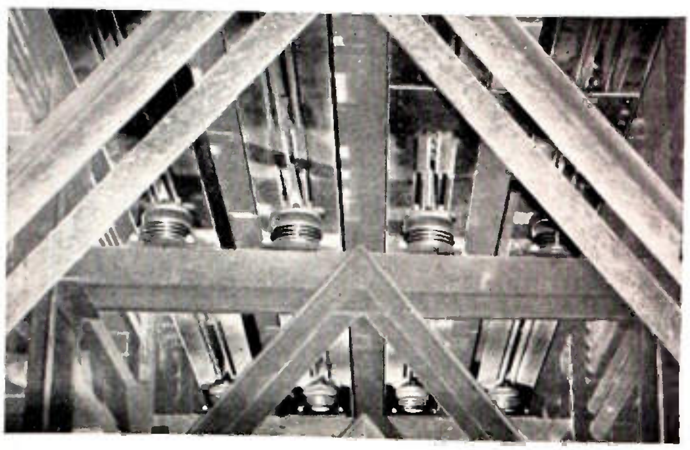


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STACKPOLE CARBON COMPANY, St. Marys, Pa.
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RC42
Insulated types. Write for Bulletin for complete details.
A DEPENDABLE SOURCE OF RESISTOR SUPPLY for over 20 YEARS

HAVE YOU AN

"impossible" requirement

FOR TECHNICAL CERAMICS?

These parts are enlarged approximately one and one half times. ▶

Many people are kind enough to say we're the first to try for any "impossible" technical ceramic. It's probably true that we've made more different sizes, types and shapes than anybody.

Through cooperation, and a little give and take on both sides, we've been able to make a lot of "impossible" ceramics. If you have an "impossible" requirement, let us know. We might be able to work it out with you. Anyway, we'd be caught trying.

We don't make a thing but technical ceramics. We've been doing it for over fifty years. It might pay you to give us first crack at anything in technical ceramics.

P. S. A couple of new plants in production now! On most things we can give you pretty fast deliveries.

51ST YEAR OF CERAMIC LEADERSHIP

AMERICAN LAVA CORPORATION

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E+TH *and*

ZENITH RADIO CORPORATION

ZENITH RADIO

PHONE BERKSHIRE 7-7100

6001 DICKENS AVENUE CHICAGO 39, ILLINOIS

CABLE ADDRESS "ZENITHRAD" ALL CODES

August 14, 1952

Antara Chemicals Division
General Dyestuff Corporation
435 Hudson Street
New York 14, N. Y.

Gentlemen:

We consider our new "Super-X" radio chassis to be the finest, most dependable in radionic history — second to none for power and sensitivity. For all our IF circuits we make our own IF Units — using threaded cores molded of G A & F Carbonyl Iron Powders. We use your E and TH types — as we have for some 10 years.

In our Armstrong FM Core Tuning Units we have recently switched from your SF type to your newly developed "J" Powder. We find that we get the same stability with a higher Q value.

We are always on the search for finer materials. We are happy to acknowledge and give credit, when we find them.

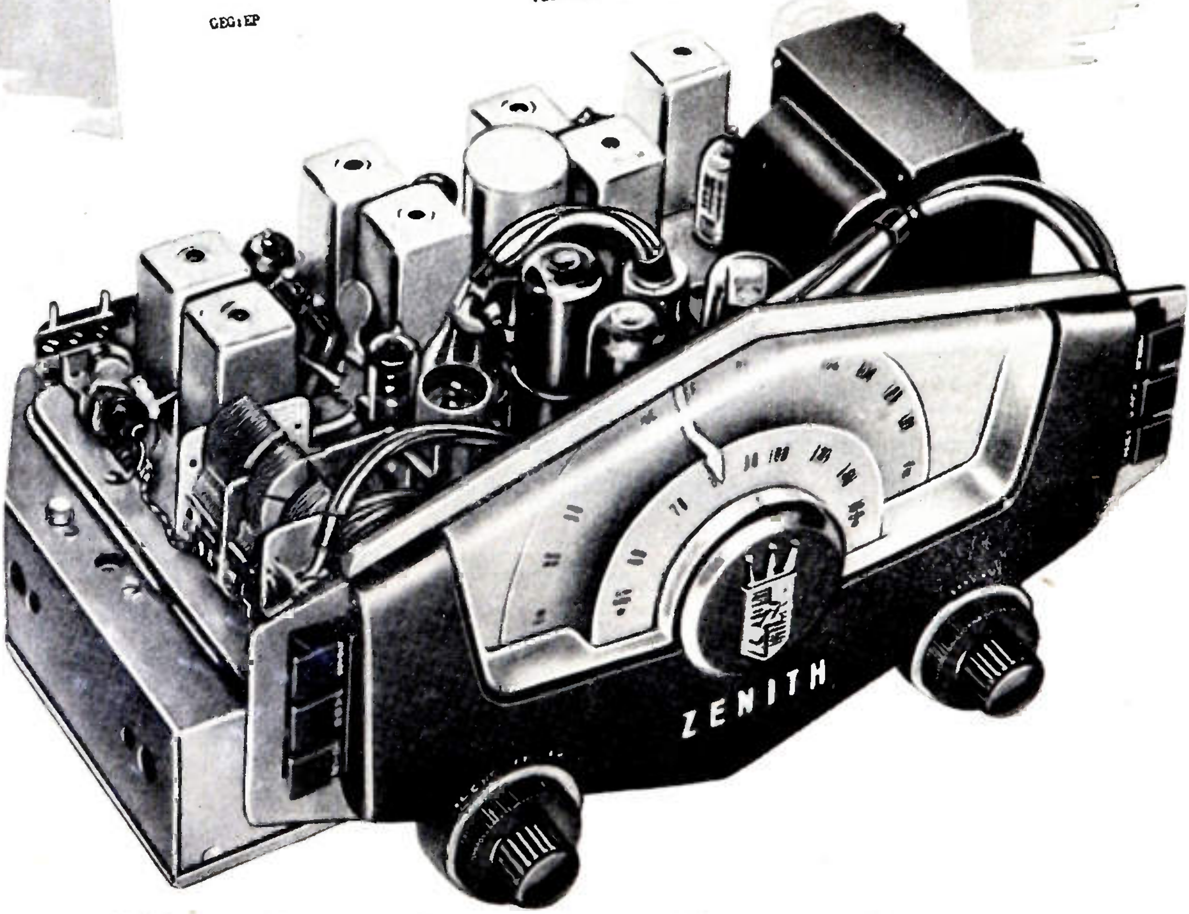
Very truly yours,

ZENITH RADIO CORPORATION

G. E. Gustafson

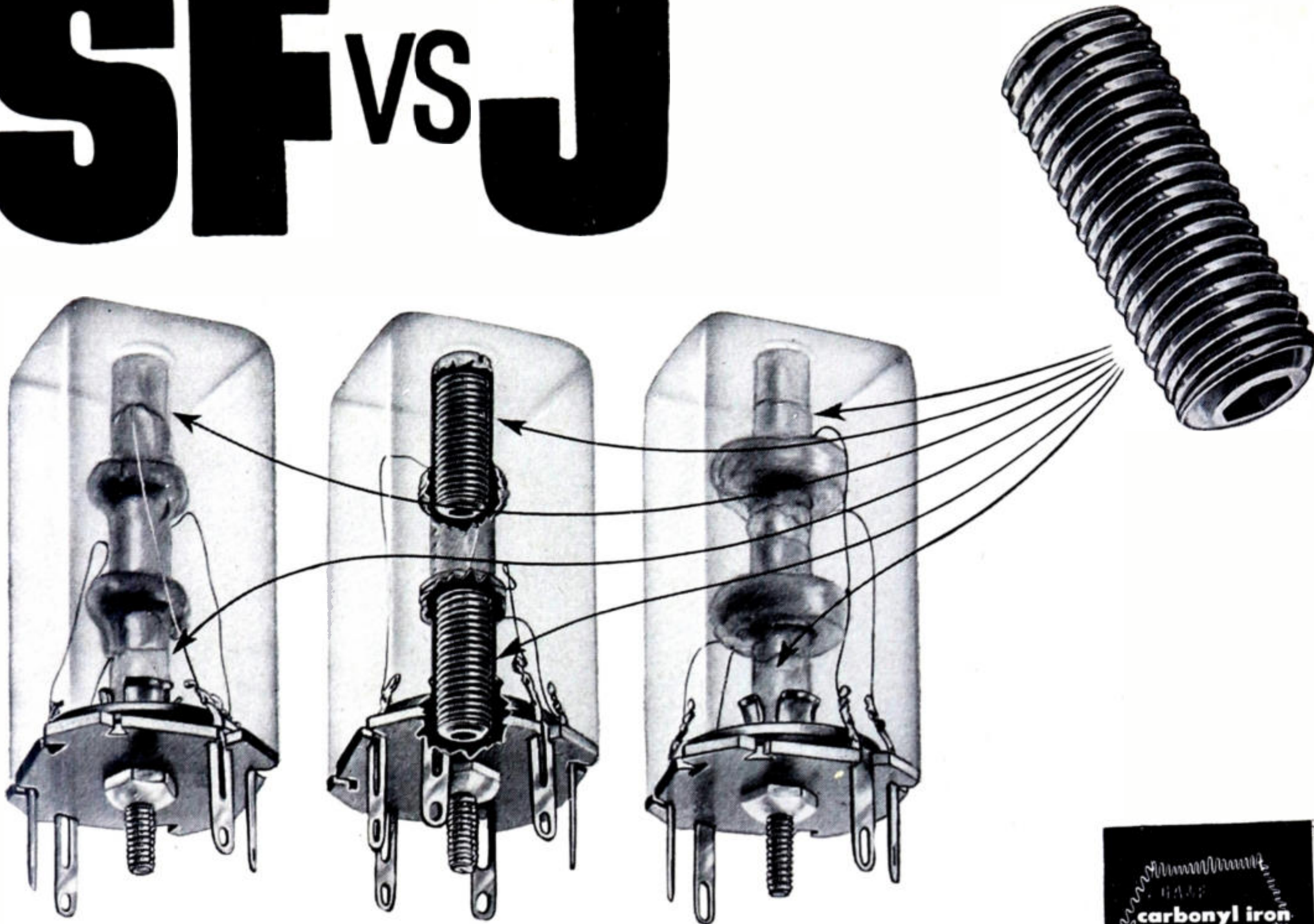
G. E. Gustafson
Vice President in Charge of Engineering

GEG:EP



G A & F® Carbonyl

SF VS J



G A & F Carbonyl Iron Powders are used to produce cores for transformer and inductor coils of every form—to increase Q values, to vary coil inductances, to reduce the size of coils, to confine stray fields and to increase transformer coupling factors.

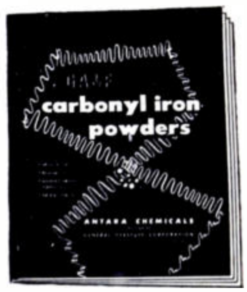
For use in TV and in Radio, including FM, the extremely small size of the particles is of enormous value, since eddy currents develop only within each particle—proportional to the square of the particle diameter. In core-making, the particles are insulated from each other by coating them, before compounding, with an efficient insulating agent.

These powders are microscopic, almost perfect spheres of extremely pure iron. They are produced in seven carefully controlled types, ranging in average particle-size from three to twenty microns in diameter.

Similarly, their properties vary, making them useful in many different applications. Engineers have commented on the fact that cores made from these powders lend themselves to smoothness of adjustment and to ease of grinding. The new Ferromagnetic Powder "J" was designed for high Q cored coils at VHF. At high frequencies, it has the lowest losses for its relatively high permeability.

* * *

We are proud to serve the Zenith Radio Corporation . . . We urge you to ask your core maker, your coil winder, your industrial designer, how G A & F Carbonyl Iron Powders can increase the efficiency and performance of the equipment or product you make, while reducing both the cost and the weight.



THIS WHOLLY NEW 32-PAGE BOOK offers you the most comprehensive treatment yet given to the characteristics and applications of G A & F Carbonyl Iron Powders. 80% of the story is told with photomicrographs, diagrams, performance charts and tables. For your copy — without obligation—kindly address Department 32.

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Iron Powders . . .

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it's news!

EAI's Dataplotter . . .

An Electronic System That Converts Digital Data To An Analog Plot . . .

Here is a system that will save countless man-hours and costs, and will insure accurate and clear presentation of data.

This new Dataplotter, designed and developed by Electronic Associates Inc., will automatically plot a cartesian curve composed of incremental points or symbols from IBM card data at maximum machine reading speed.

It will accept data from other inputs — Magnetic tape, keyboards, digital computers, etc.

It will retain at all times the basic accuracy of the digital system.

Here's what the Dataplotter system consists of:

- Variplotter Model 205G
- Digital-to-analog converter, Model 417
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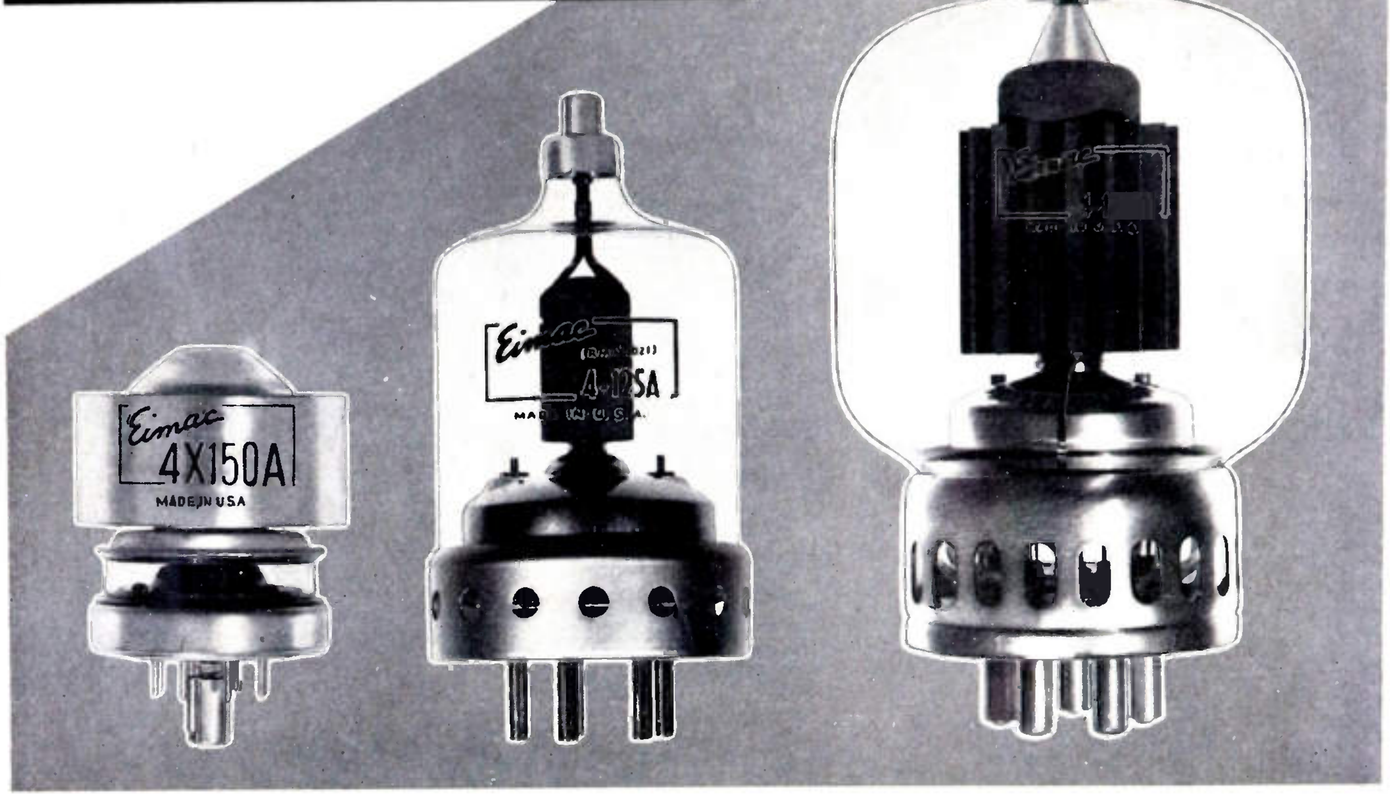
Eimac Power Tetrodes

Dependable

Economical



Years of dependable operation have established Eimac tetrodes as economical, incomparable performers. Economical because of low driving power, long life and simple circuit requirements. Incomparable because of the many Eimac features, including high power gain, ability to withstand great amounts of mechanical and thermal shock and stability of operation. Eimac tetrodes range in plate dissipation ratings from 65 to 20,000 watts and operate over the spectrum from audio frequencies to the ultra high frequencies of television. Eimac tetrodes are used as oscillators, modulators or amplifiers by those who demand the ultimate in transmitter performance.

We invite consultation concerning your electronic problems and needs. For free information about any of Eimac's complete line of power tetrodes-write our application engineering department.

Now available for 25 cents is the Eimac application bulletin number eight, "The Care and Feeding of Power Tetrodes". This 28-page booklet was written by vacuum tube engineers to help you get the most out of your tetrodes.

EITEL-McCULLOUGH, INC.

SAN BRUNO, CALIFORNIA

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

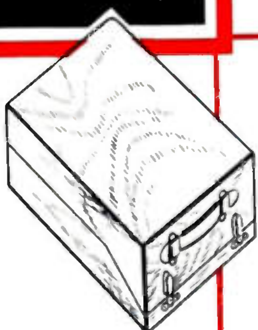
Eimac TUBES

marion's NEW metertester

MODEL M-2

MEASURES SENSITIVITY AND RESISTANCE

for testing and calibration of D. C. instruments in the laboratory and on production lines



Marion's New Metertester (Model M-2) retains proven Marion features but increases application flexibility. In addition to improved circuitry for sensitivity measurement it also measures internal resistance of sensitive instruments without exceeding full scale rating of the instrument under test.

FEATURES

- Regulated Power Supply
- Stepless Vacuum Tube Voltage Control
- Illuminated 8½" Mirror-Scale Standard Instrument, Hand Calibrated
- Marion Ruggedized Null Indicator movement for bridge balance indication
- Decade of .1% accurate Manganin Wire Wound Resistors
- Direct Reading Bridge Circuit using Helipot
- Complete. No accessories required

SPECIFICATIONS

ACCURACY: Overall better than ¼ of 1%
RESISTANCE RANGE: 0-5000 ohms
POWER SOURCE: 115V A C 60 cycles
CASE SIZE: 15⅛" x 10⅛" x 5⅜"
WEIGHT: 15 lbs.

SENSITIVITY RANGES

0-25UA	0-200UA	0-800UA	0-10 MA
0-50UA	0-400UA	0-1 MA	0-100 Volts
0-100UA	0-500UA	0-5 MA	

The New M-2 Model can also be used for additional purposes, such as a precise source of DC current and voltage and as a precision Wheatstone bridge in the 0-5000 ohm range.

For further information write Marion Electrical Instrument Co., 407 Canal Street, Manchester, N. H., U. S. A.

marion meters

Reg. U. S. Pat. Off.

MANUFACTURERS OF RUGGEDIZED, HERMETICALLY SEALED AND STANDARD PANEL INSTRUMENTS

TYPE 252, JAN-R-19, Type RA20

2 watt, 1 17/64" diameter variable wirewound resistor. Also available with other special military features not covered by JAN-R-19. Attached Switch can be supplied.

	RA20, JAN Shaft Type SD		RA20 High Torque, JAN Shaft Type SD	
Resistance	CTS Part	JAN-R-19 TYPE	CTS Part	JAN-R-19 TYPE
50±10%	B8079	RA20A1SD500AK	X3496	RA20A2SD500AK
100±10%	W6929	RA20A1SD101AK	L9388	RA20A2SD101AK
250±10%	X3497	RA20A1SD251AK	M9879	RA20A2SD251AK
500±10%	W6931	RA20A1SD501AK	X3498	RA20A2SD501AK
1000±10%	W6932	RA20A1SD102AK	X3499	RA20A2SD102AK
1500±10%	W6933	RA20A1SD152AK	M9809	RA20A2SD152AK
2500±10%	W6934	RA20A1SD252AK	L9103	RA20A2SD252AK
5000±10%	W6935	RA20A1SD502AK	L9104	RA20A2SD502AK
10,000±10%	W6936	RA20A1SD103AK	H8979	RA20A2SD103AK

TYPE 25, JAN-R-19, Type RA30 (May also be used as Type RA25)

4 watt, 1 17/32" diameter variable wirewound resistor. Also available with other special military features not covered by JAN-R-19. Attached Switch can be supplied.

	RA30, JAN Shaft Type SD		RA30 High Torque, JAN Shaft Type SD	
Resistance	CTS Part	JAN-R-19 TYPE	CTS Part	JAN-R-19 TYPE
50±10%	X3502	RA30A1SD500AK	W2837	RA30A2SD500AK
100±10%	X3503	RA30A1SD101AK	X3504	RA30A2SD101AK
250±10%	X3505	RA30A1SD251AK	X3506	RA30A2SD251AK
500±10%	X3507	RA30A1SD501AK	M7566	RA30A2SD501AK
1000±10%	X3508	RA30A1SD102AK	S2444	RA30A2SD102AK
1500±10%	X3509	RA30A1SD152AK	X3510	RA30A2SD152AK
2500±10%	X3511	RA30A1SD252AK	S2736	RA30A2SD252AK
5000±10%	Q1409	RA30A1SD502AK	X3512	RA30A2SD502AK
10,000±10%	X3513	RA30A1SD103AK	R1561	RA30A2SD103AK
15,000±10%	X3514	RA30A1SD153AK	L9107	RA30A2SD153AK

Immediate delivery from stock

JAN-R-94 AND JAN-R-19 TYPE MILITARY VARIABLE RESISTORS

167 types

Preference given to orders carrying military contract number and DO rating. Other JAN items or special items with or without associated switches can be fabricated to your specifications. Please give complete details on your requirements including electrical and mechanical specifications.

UNPRECEDENTED PERFORMANCE CHARACTERISTICS

Designed for use in military equipment subject to extreme temperature and humidity ranges including jet and other planes, guided missiles, tanks, ships and submarines, telemetering, microwave, portable or mobile equipment and all other military communications.

For further information, write for Stock Sheet No. 162

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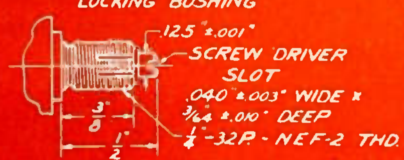
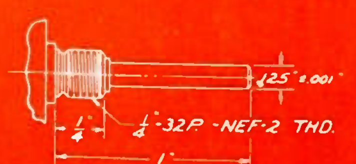
CTS SHAFT TYPE LT-2
LOCKING BUSHING



MOUNTING HARDWARE ASSEMBLED
MOUNTING NUT 3/8" HEX. x 3/32"
LOCK NUT 3/8" HEX. x 1/8"
LOCK WASHER #1914A

CTS SHAFT TYPE RE



MOUNTING HARDWARE ASSEMBLED
MOUNTING NUT 3/8" HEX. x 3/32"
LOCK WASHER #1914A

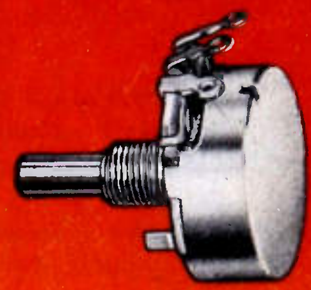
TYPE 65

½ watt 70° C, ¾" diameter miniaturized variable composition resistor.

Resistance	CTS Part CTS Shaft Type RE	CTS Part Locking Bushing CTS Shaft Type LT-2
250±10%	X3516	X3530
500±10%	X3517	X3531
1000±10%	X3518	X3532
2500±10%	X3519	X3533
5000±10%	X3520	X3534
10,000±10%	X3521	X3535
25,000±10%	X3522	X3536
50,000±10%	X3523	X3537
100,000±10%	X3524	X3538
250,000±10%	X3525	X3539
500,000±10%	X3526	X3540
1 Meg±20%	X3527	X3541
2.5 Meg±25%	X3528	X3542

TYPE 95, JAN-R-94, Type RV4

2 watt 70°C, 1⅛" diameter variable composition resistor. Also available with other special military features not covered by JAN-R-94. Attached Switch can be supplied.



Resistance	JAN-R-94 TYPE RV4 JAN Shaft Type SD	JAN-R-94 TYPE RV4 JAN Shaft Type RJ	CTS Part Non-JAN Locking Bushing CTS Shaft Type LT-1
100±10%	RV4ATSD101A	RV4ATRJ101A	W3160
250±10%	RV4ATSD251A	RV4ATRJ251A	W3161
500±10%	RV4ATSD501A	RV4ATRJ501A	W3162
1000±10%	RV4ATSD102A	RV4ATRJ102A	W3166
2500±10%	RV4ATSD252A	RV4ATRJ252A	W3163
5000±10%	RV4ATSD502A	RV4ATRJ502A	W3164
10,000±10%	RV4ATSD103A	RV4ATRJ103A	W3167
25,000±10%	RV4ATSD253A	RV4ATRJ253A	W3168
50,000±10%	RV4ATSD503A	RV4ATRJ503A	W3169
100,000±10%	RV4ATSD104A	RV4ATRJ104A	W3170
250,000±10%	RV4ATSD254A	RV4ATRJ254A	W3171
500,000±10%	RV4ATSD504A	RV4ATRJ504A	W3172
1 Meg±20%	RV4ATSD105B	RV4ATRJ105B	W3173
2.5 Meg±20%	RV4ATSD255B	RV4ATRJ255B	W3165
5 Meg±20%	RV4ATSD505B	RV4ATRJ505B	W3159

TYPE 45, JAN-R-94, Type RV2

¼ watt, 15/16" diameter variable composition resistor. Also available with other special military features not covered by JAN-R-94. Attached Switch can be supplied.

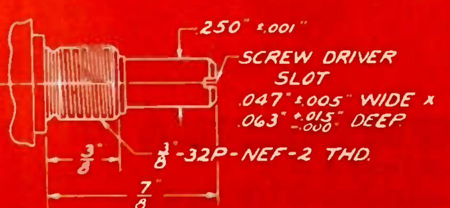
Resistance	RV2, JAN Shaft Type SD — CTS Part	RV2, JAN Shaft Type SD — JAN-R-94 TYPE	CTS Part Non-JAN Locking Bushing CTS Shaft Type LT-1
100±10%	A5876	RV2ATSD101A	A5922
250±10%	A5877	RV2ATSD251A	A5923
500±10%	A5878	RV2ATSD501A	A5924
1000±10%	A5879	RV2ATSD102A	A5925
2500±10%	A5880	RV2ATSD252A	A5926
5000±10%	A5881	RV2ATSD502A	A5927
10,000±10%	A5882	RV2ATSD103A	A5928
25,000±10%	A5883	RV2ATSD253A	A5929
50,000±10%	A5884	RV2ATSD503A	A5930
100,000±10%	A5885	RV2ATSD104A	A5931
250,000±10%	A5886	RV2ATSD254A	A5932
500,000±10%	A5887	RV2ATSD504A	A5933
1 Meg±20%	A5888	RV2ATSD105B	A5934
2.5 Meg±20%	A5889	RV2ATSD255B	A5935

TYPE 35, JAN-R-94, Type RV3

½ watt, 1⅛" diameter variable composition resistor. Also available with other special military features not covered by JAN-R-94. Attached Switch can be supplied.

Resistance	RV3, JAN Shaft Type SD — CTS Part	RV3, JAN Shaft Type SD — JAN-R-94 TYPE	CTS Part Non-JAN Locking Bushing CTS Shaft Type LT-1
100±10%	A5861	RV3ATSD101A	A5907
250±10%	A5862	RV3ATSD251A	A5908
500±10%	A5863	RV3ATSD501A	A5909
1000±10%	A5864	RV3ATSD102A	A5910
2500±10%	A5865	RV3ATSD252A	A5911
5000±10%	A5866	RV3ATSD502A	A5912
10,000±10%	A5867	RV3ATSD103A	A5913
25,000±10%	A5868	RV3ATSD253A	A5914
50,000±10%	A5869	RV3ATSD503A	A5915
100,000±10%	A5870	RV3ATSD104A	A5916
250,000±10%	A5871	RV3ATSD254A	A5917
500,000±10%	A5872	RV3ATSD504A	A5918
1 Meg±20%	A5873	RV3ATSD105B	A5919
2.5 Meg±20%	A5874	RV3ATSD255B	A5920
5 Meg±20%	A5875	RV3ATSD505B	A5921

JAN SHAFT TYPE SD



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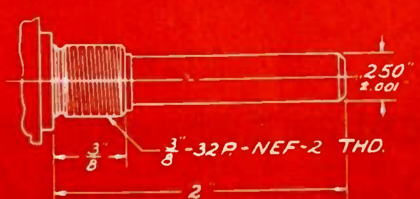
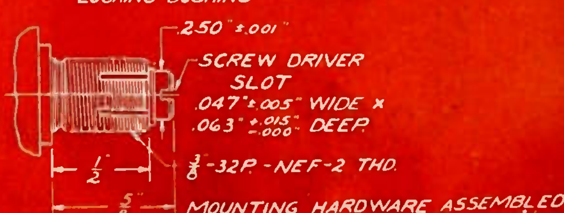
MOUNTING NUT 9/16" HEX. × 3/32"
LOCK WASHER #1920A

JAN SHAFT TYPE RJ



MOUNTING HARDWARE ASSEMBLED
MOUNTING NUT 9/16" HEX. × 3/32"
LOCK WASHER #1920A

CTS SHAFT TYPE LT-1
LOCKING BUSHING



MOUNTING HARDWARE ASSEMBLED
MOUNTING NUT 9/16" HEX. × 3/32"
LOCK NUT 1/2" HEX. × 5/32"
LOCK WASHER #1920A

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MYCALEX is a highly developed glass-bonded mica insulation backed by a quarter-century of continued research and successful performance. Both pioneer and leader in low-loss, high frequency insulation, MYCALEX offers designers and manufacturers an economical means of attaining new efficiencies, improved performance. The unique combination of characteristics that have made MYCALEX the choice of leading electronic manufacturers are typified in the table for MYCALEX grade 410 shown below. Complete data on all grades will be sent promptly on request.

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- **ELECTRODES, METAL INSERTS CAN BE MOLDED-IN**
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Dielectric constant, 1 megacycle	9.2
Loss factor, 1 megacycle	0.014
Dielectric strength, volts/mil	400
Volume resistivity, ohm-cm	1×10^{15}
Arc resistance, seconds	250
Impact strength, Izod, ft.-lb/in. of notch	0.7
Maximum safe operating temperature, °C	350
Maximum safe operating temperature, °F	650
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Coefficient of linear expansion, °C	11×10^{-6}
Tensile strength, psi	6000

MYCALEX is specified by the leading manufacturers in almost every electronic category


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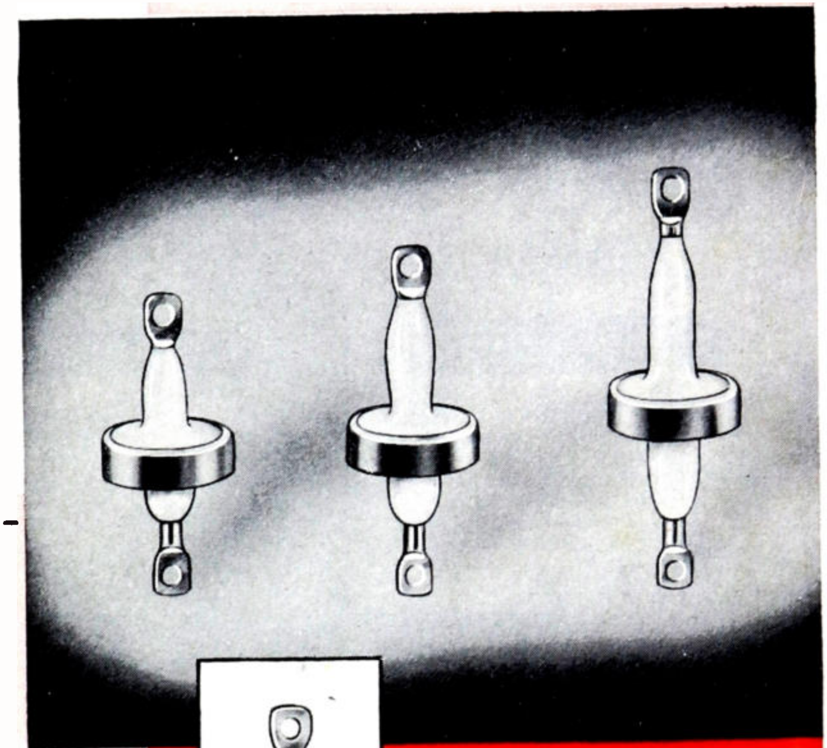
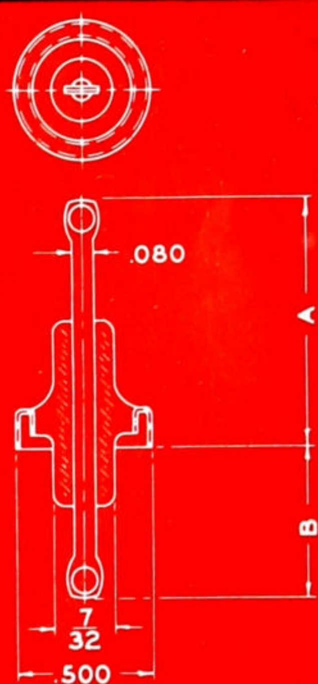


RUGGED FOOLPROOF



HERMETIC's new high voltage terminal is a glass-metal header that can take up to 10,000 volts and yet occupies only ½" diameter on the mounting surface. It is amazingly rugged and will insure foolproof operation because of the diaphragm construction of the mounting. It is available with solid or tubular center conductors. Applications are many, among which are its use in transformers, choke coils, capacitors and pulsed circuits.

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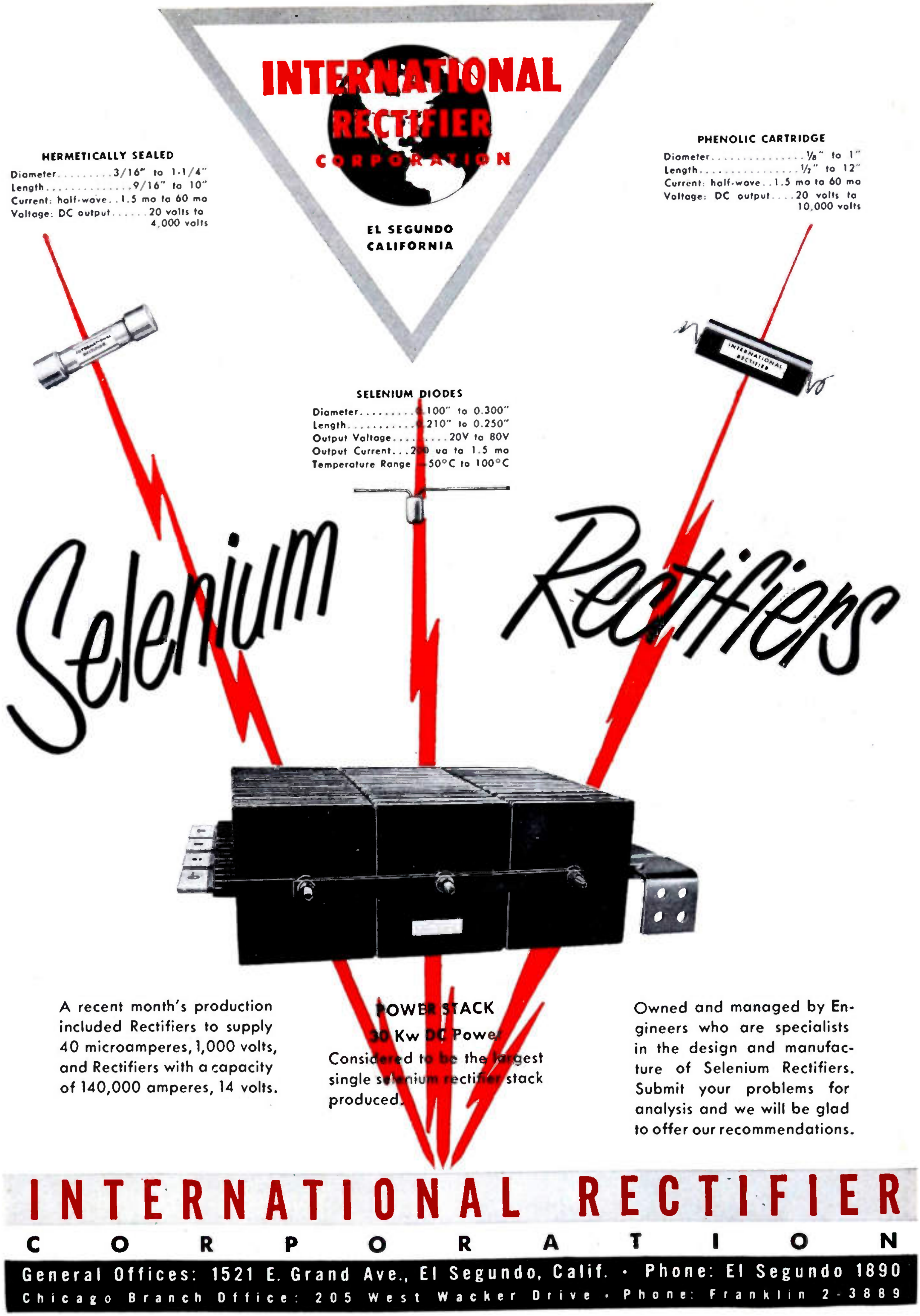

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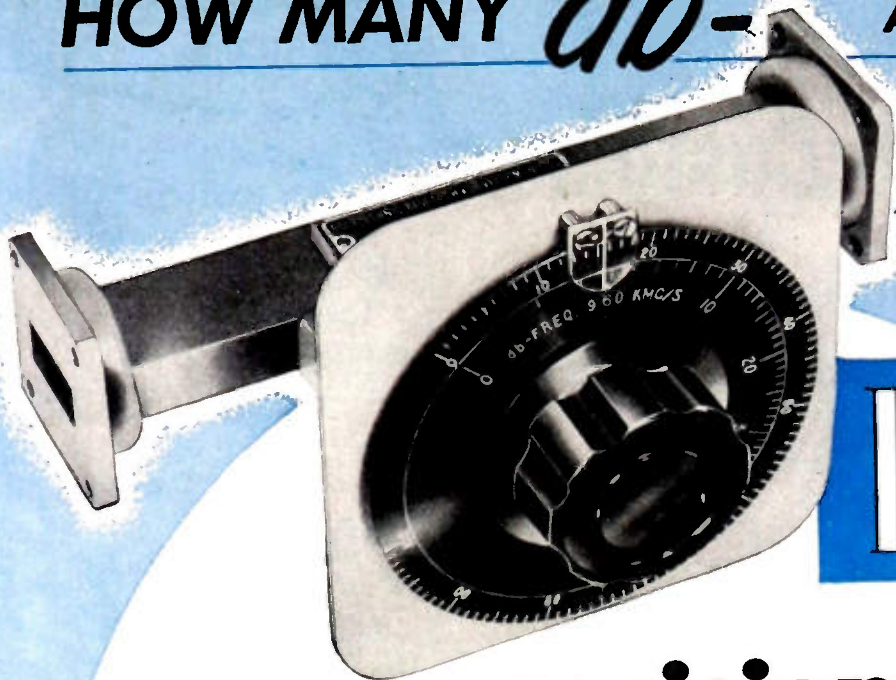

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precision variable attenuators

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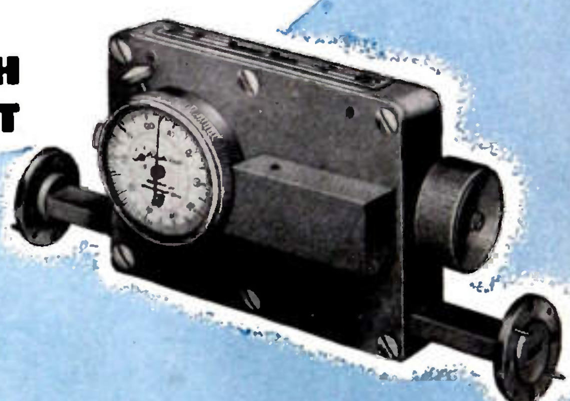

Polytechnic RESEARCH & DEVELOPMENT COMPANY, Inc.

PRD

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1941 WL-530 in Radar at Pearl Harbor

USERS OF WESTINGHOUSE TUBES GET FIRST BENEFITS FROM MANY NEW TUBE DEVELOPMENTS

These are only a few of the "firsts" that Westinghouse created in the electronic tube industry. In each case, designers using Westinghouse Tubes gained advantages by having first chance to use these innovations.

Today, Westinghouse still pioneers in electronic tubes and tube making. For instance, Westinghouse 40 KV and 20 KV rectifying tubes are under 9 ounces, only 2¾" high. Designers seeking the ultimate in space and weight savings will find them in these new WL-6102 and WL-6103 tubes.

Radical new developments in other power tubes and receiving and television picture tubes are now being engineered at the *NEW* Westinghouse Electronic Tube Division at Elmira and Bath, New York.

NEW SERVICE, NEW DISTRIBUTION

Westinghouse plans for Electronic Tube Division expansion are in operation. New service facilities, new warehousing policies, and new distributors are opening rapidly.

New merchandising methods will aid distributors in serving industrial users—many of these business-building programs are totally new in the tube industry. Here, as elsewhere, Westinghouse plans to provide industry leadership in service.

It pays in profits to deal with Westinghouse and with Westinghouse distributors. For full information on how Westinghouse can help you with problems of design, service, or supply, call your nearest Westinghouse representative, or write to Department C-110

RELIATRON™ TUBES

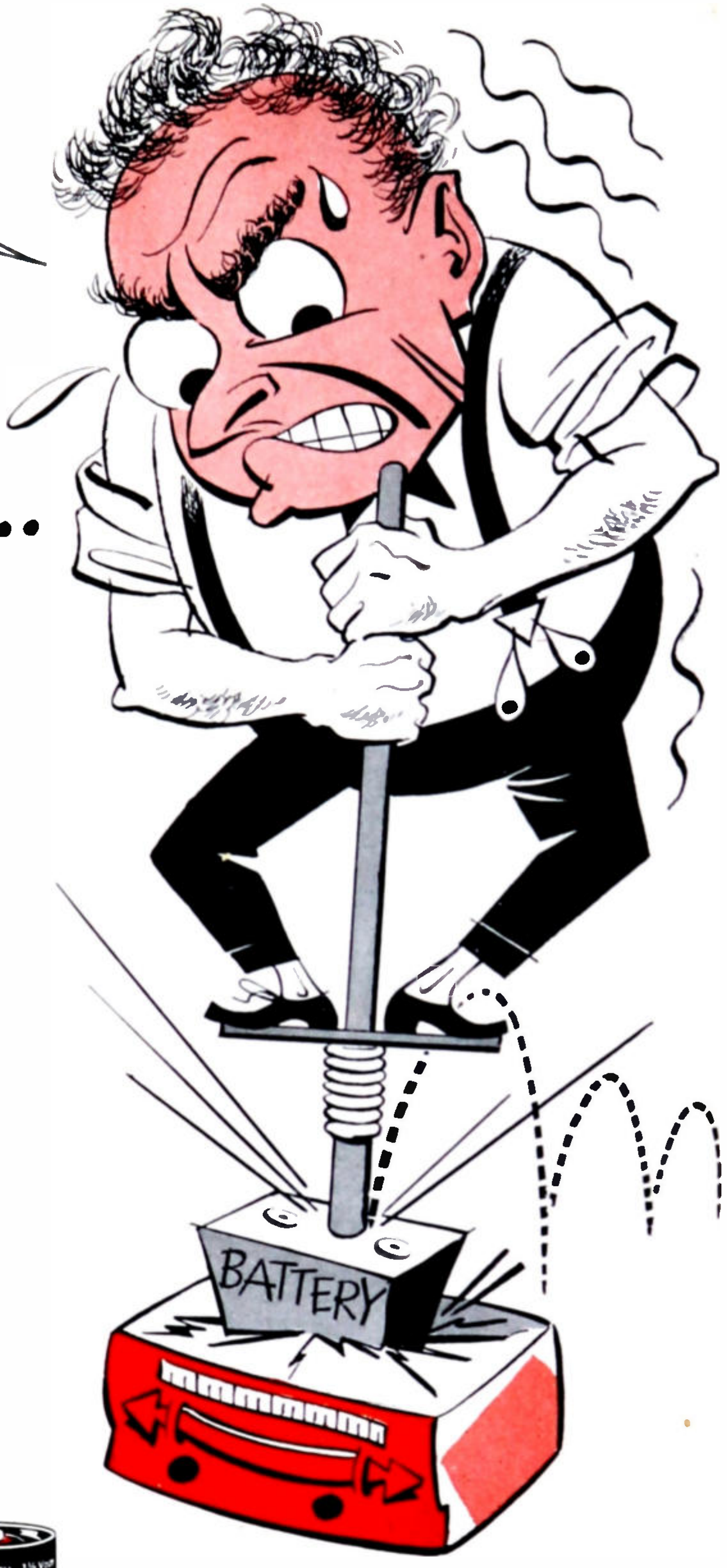
YOU CAN BE SURE...IF IT'S Westinghouse

ELECTRONIC TUBE DIVISION

Westinghouse Electric Corporation
Box 284, Elmira, N. Y.

You can use a pogo stick to make it fit...

BUT it's simpler to design the radio around the battery!



National Carbon offers a complete range of "Eveready" "Nine Lives" radio batteries. Just design your new model receiver—any type or size—around *standard*, compact, long-lasting "Eveready" batteries and forget you ever had a battery problem.

Users prefer them, too. They enjoy *better* listening *longer* . . . and when replacements *are* necessary, "Eveready" brand radio batteries are sure to be available wherever radio batteries are sold.

Write to our Battery Engineering Department for full details and specifications of "EVEREADY" radio batteries.

"Eveready" No. 964 "A" battery and "Eveready" No. 477 "B" battery for "personal" receivers—lowest priced complement of its size on the market—feature lowest cost per hour of listening plus a new high in balanced life of the two components.

EVEREADY
TRADE-MARK
RADIO BATTERIES

The terms "Eveready", "Nine Lives" and the Cat Symbol are registered trade-marks of Union Carbide and Carbon Corporation

NATIONAL CARBON COMPANY
A Division of Union Carbide and Carbon Corporation
30 East 42nd Street, New York 17, N. Y.

District Sales Offices: Atlanta, Chicago, Dallas, Kansas City, New York, Pittsburgh, San Francisco

IN CANADA:
National Carbon Limited, Montreal, Toronto, Winnipeg

electronic wire and cables for standard and special applications

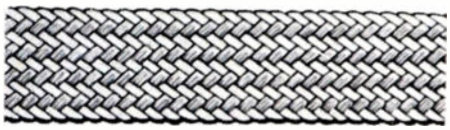
Whether your particular requirements are for standard or special application, choose *LENZ* for the *finest* in precision-manufactured electronic wire and cable.

GOVERNMENT PURPOSE RADIO AND INSTRUMENT HOOK-UP WIRE,
plastic or braided type, conforming to Government Specification JAN-C-76, etc., for radio and instruments. Solid or flexible conductors, in a variety of sizes and colors.

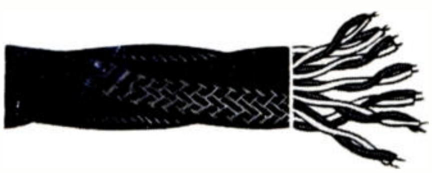
RADIO AND INSTRUMENT HOOK-UP WIRE,
Underwriters Approved, for 80° C., 90° C. and 105° C. temperature requirements. Plastic insulated, with or without braids.

RF CIRCUIT HOOK-UP AND LEAD WIRE
for VHF and UHF, AM, FM and TV high frequency circuits. LENZ Low-Loss RF wire, solid or stranded tinned copper conductors, braided, with color-coded insulation, waxed impregnation.

SHIELDED MULTIPLE CONDUCTOR CABLES
Conductors: Multiple—2 to 7 or more of flexible tinned copper. Insulation: extruded color-coded plastic. Closely braided tinned copper shield. For: Auto radio, indoor PA systems and sound recording equipment.



SHIELDED COTTON BRAIDED CABLES
Conductors: Multiple—2 to 7 or more of flexible tinned copper. Insulation: extruded color-coded plastic. Cable concentrically formed. Closely braided tinned copper shield plus brown overall cotton braid.



SPECIAL HARNESSES,
cords and cables, conforming to Government and civilian requirements.

SHIELDED JACKETED MICROPHONE CABLE
Conductors: Multiple—2 to 7 or more conductors of stranded tinned copper. Insulation: extruded color-coded plastic. Closely braided tinned copper shield. Tough, durable jacket overall.

JACKETED MICROPHONE CABLE
Conductors: Extra-flexible tinned copper. Polythene insulation. Shield: #36 tinned copper, closely braided, with tough durable jacket overall. Capacity per foot: 29MMF.

TINNED COPPER SHIELDING AND BONDING BRAIDS
Construction: #34 tinned copper braid, flattened to various widths. Bonding Braids conforming to Federal Spec. QQ-B-575 or Air Force Spec. 94-40229.

PA AND INTERCOMMUNICATION CABLE
Conductors: #22 stranded tinned copper. Insulation: textile or plastic insulated conductors. Cable formed of Twisted Pairs, color-coded. Cotton braid or plastic jacket overall. Furnished in 2, 5, 7, 13 and 25 paired, or to specific requirements.

CHECK LENZ FIRST!

Lenz Electric Manufacturing Co.
1751 N. Western Ave., Chicago 47, Illinois
Our 48th Year in Business

cords, cable and wire for radio ◆ p. a. ◆ test instruments ◆ component parts

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 10A)

Attentuation Networks

The Daven Company., Dept. AN, 191 Central Ave., Newark 4, N. J., announces the availability of its new "T" or Balanced "H" Attenuation Networks, Series 690. These networks are designed for use in general laboratory and production testing where ruggedness, flexibility and reliability are of prime importance.

Series 690 has a frequency range from zero to 50,000 cps. These networks consist of "plug in" impedance-adjusting networks in compact assemblies with either two or three attenuation controls, depending upon the loss-per-step desired. The "plug in" impedance-matching networks may be obtained in a wide range of impedance and loss, with special impedances and losses available to meet requirements.

The level of operation is 20 db (0.6 watts) maximum input, with accuracy of resistor units calibrated at 1 per cent. The type of winding for these attenuation networks is the noninductive, card type.

A feature of the Series 690 is the use of silver alloy for the contacts, slip-rings and the Daven patented "knee-action" switch rotor. The rotor is tamper-proof and provides low and uniform contact resistance over the life of the unit.

The networks are available in either portable or rack models.

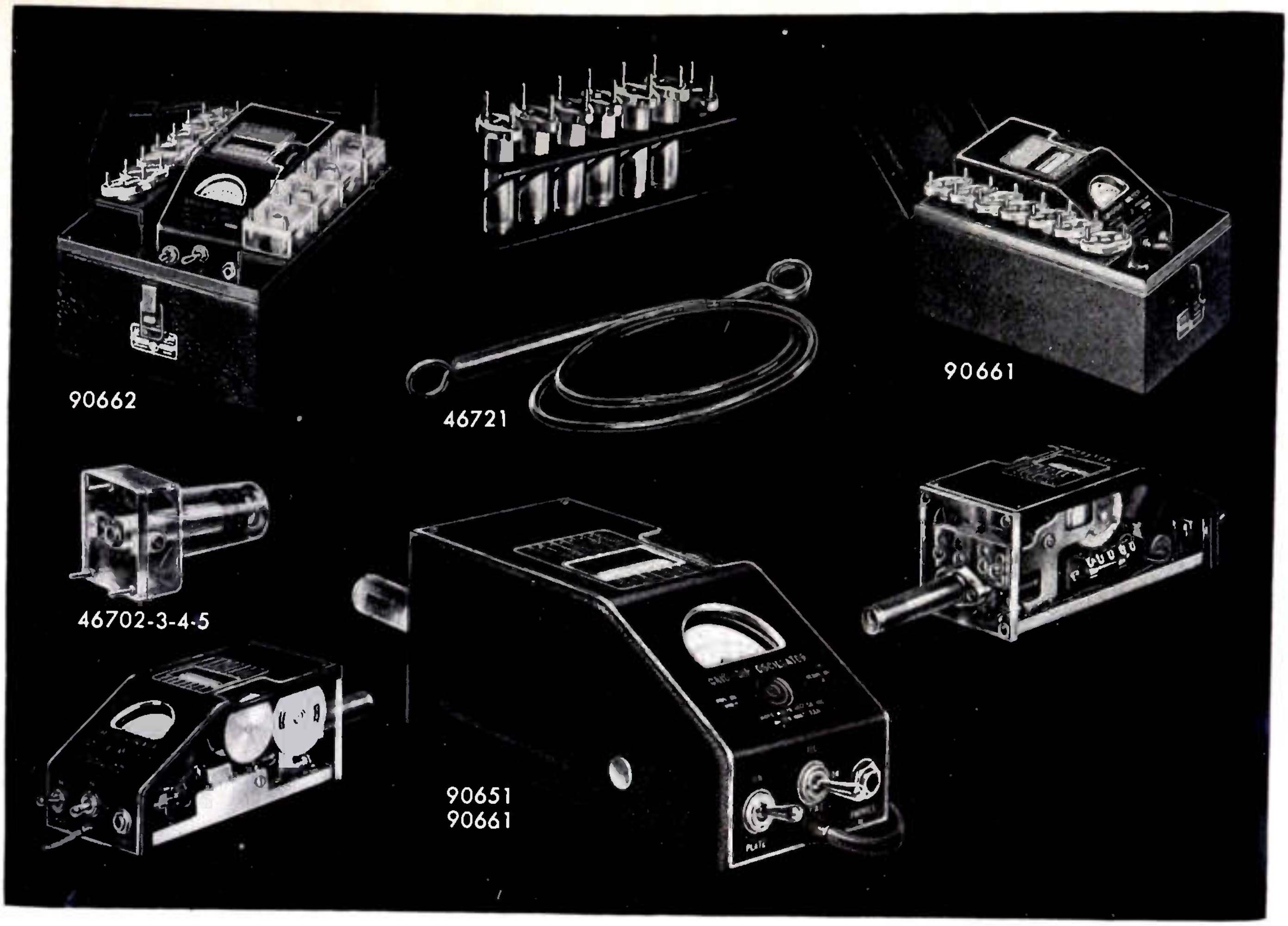
Single Mike Switch for Dual Communication

The **American Radiotelephone Company Inc.,** 3505 Fourth St., N. St. Petersburg, Fla., has designed a switch that provides a means of switching an operator's headset microphone from one communication system, such as a land line telephone, to the transmitter of a second. It holds open the connection simultaneously with the first system. Thus, a single headset microphone can be used with any two communication systems.

Called the American Radio-Telephone Microphone Control, this new unit works by a foot operated micro-switch so the operator's hands are left free.

The manufacturer states this new unit is adaptable for use in dispatching stations where directions are received by land line telephone and communicated by an operator to mobile units by radio. Claim is also made that it can be used between any two electronic communications systems to increase operator efficiency through faster message transmission.

(Continued on page 54A)



Designed for Application

Grid Dip Meters

Millen Grid Dip Meters are available to meet all various laboratory and servicing requirements.

The 90662 Industrial Grid Dip Meter completely calibrated for laboratory use with a range from 225 kc. to 300 mc. incorporates features desired for both industrial and laboratory application, including three wire grounding type power cord and suitable carrying case.

The 90661 Industrial Grid Dip Meter is similar to the 90662 except for a reduced range of 1.7 to 300 mc. It likewise incorporates the three wire grounding type cord and metal carrying case.

The 90651 Standard Grid Dip Meter is a somewhat less expensive version of the grid dip meter. The calibration while adequate for general usage is not as complete as in the case of the industrial model. It is supplied without grounding lead and without carrying case. The range is 1.7 to 300 mc. Extra inductors available extends range to 220 kc.

The Millen Grid Dip Meter is a calibrated stable RF oscillator unit with a meter to read grid current. The frequency determining coil is plugged into the unit so that it may be used as a probe.

These instruments are complete with a built-in transformer type A.C. power supply and interminal terminal board to provide connections for battery operation where it is desirable to use the unit on antenna measurements and other usages where A.C. power is not available. Compactness has been achieved without loss of performance or convenience of usage. The incorporation of the power supply, oscillator and probe into a single unit provides a convenient device for checking all types of circuits. The indicating instrument is a standard 2 inch General Electric instrument with an easy to read scale. The calibrated dial is a large 270° drum dial which provides seven direct reading scales, plus an additional universal scale, all with the same length and readability. Each range has its individual plug-in probe completely enclosed in a contour fitting polystyrene case for assurance of permanence of calibration as well as to prevent any possibility of mechanical damage or of unintentional contact with the components of the circuit being tested.

The Grid Dip Meters may be used as:

1. A Grid Dip Oscillator
2. An Oscillating Detector
3. A Signal Generator
4. An Indicating Absorption Wavemeter

The most common usage of the Grid Dip Meter is as an oscillating frequency meter to determine the resonant frequencies of de-energized tuned circuits.

Size of Grid Dip Meter only (less probe): 7 in. x $3\frac{3}{16}$ in. x $3\frac{3}{8}$ in.

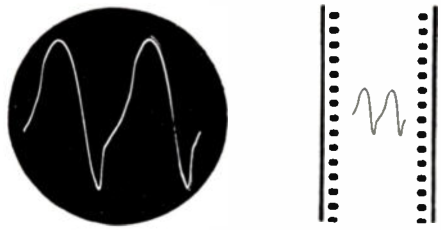
JAMES MILLEN M MFG. CO., INC.

MAIN OFFICE AND FACTORY

MALDEN, MASSACHUSETTS, U.S.A.

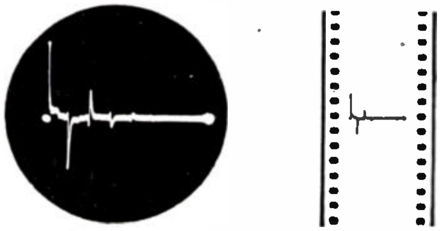
Want an oscilloscope camera NOW?

Fairchild Oscillo-Record Cameras are now available *from stock* for immediate shipment. With these units you can make *permanent* photographic records of oscilloscope traces, thereby eliminating possible errors in making hand sketches from memory. In time-saving and convenience alone, these cameras will pay for themselves many times over.



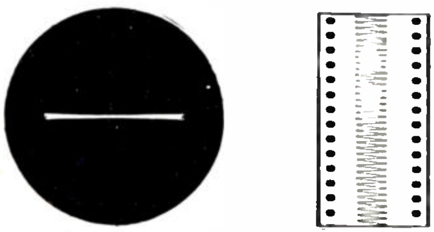
Scope Image — Film Recording

1. Single-frame photography of stationary patterns using a continuously running sweep.



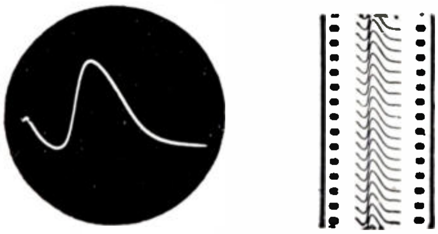
Scope Image — Film Recording

2. Single-frame photography of single transients using a single sweep.



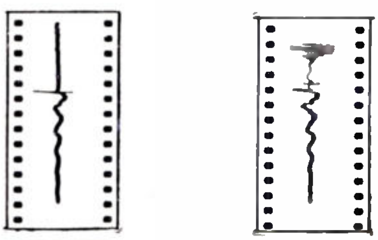
Scope Image — Film Recording

3. Continuous-motion photography employing film motion as a time base.



Scope Image — Film Recording

4. Continuous-motion photography employing oscilloscope sweep as a time base.



FILM MOTION TIME BASE — FILM MOTION AND SCOPE SWEEP

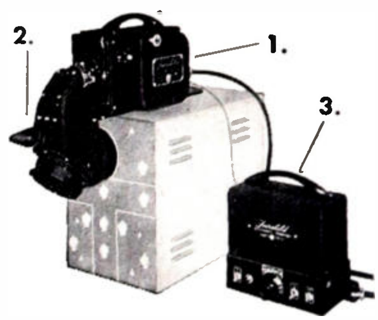
5. Continuous-motion photography employing combination of film motion and oscilloscope sweep as a time base.

Complete information about applications and operation of both the Fairchild Oscillo-Record Camera and the Fairchild-Polaroid Oscilloscope Camera is available. Write today to *Fairchild Camera and Instrument Corporation, 88-06 Van Wyck Boulevard, Jamaica 1, New York, Department 120-18F2.*

FAIRCHILD OSCILLO-RECORD CAMERA IS UNUSUALLY VERSATILE

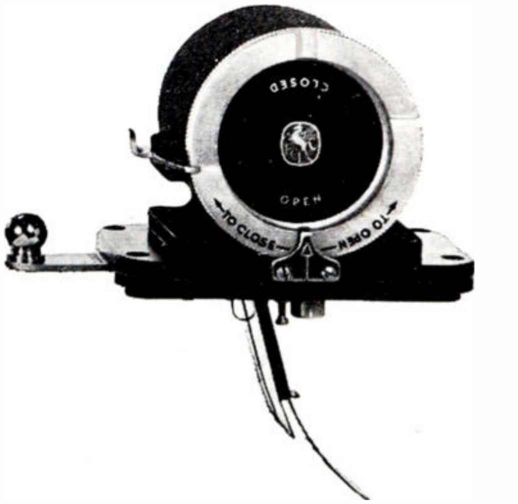
Users of the Fairchild Oscillo-Record Camera like its versatility. Designed for both still and continuous-motion photography on 35-mm film, it records non-recurring phenomena that are too rapid for visual study, others that are so slow that continuity is lost, and the occasions where very high-speed transients are combined with very slow-speed phenomena. For some idea of the types of jobs this instrument can do, study the examples at the left. Each solves a particular problem. Oscillo-Record camera users especially like its:

- **CONTINUOUSLY VARIABLE SPEED CONTROL** — 1 in/min. to 3600 in/min.
- **TOP OF SCOPE MOUNTING** that leaves controls easily accessible.
- **PROVISION FOR 3 FILM LENGTHS**—100, 400 or 1,000 feet.



1. Camera, 2. periscope, 3. electronic speed control. Accessories include 400- and 1,000-ft. film magazines, magazine adaptor and motor, universal mount for camera and periscope, binocular split-beam viewer.

FAIRCHILD TAKE-UP CASSETTE FOR SHORT RUNS



Where only a few pictures are required for quick development and study, a small Take-up Cassette is available as an accessory. The convenience afforded by this unit results in the saving of considerable time in handling short runs and reduces film wastage to a minimum. It is easily attached to the top of the camera by means of an adapter. A built-in knife permits short lengths of exposed film (up to 10 feet) to be cut off and removed with the cassette for developing.

FAIRCHILD

OSCILLOSCOPE RECORDING CAMERAS

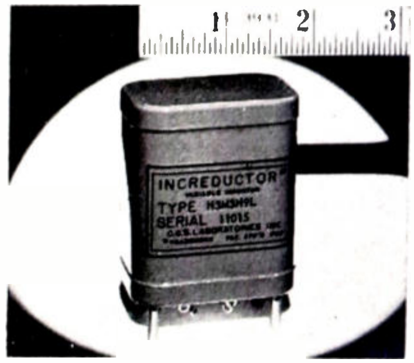
News—New Products

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

(Continued from page 44A)

Remote Control RF Inductor

The H3M 3H9L INCREDUCTOR* controllable inductor has been added to the line of available Increductor units manufactured by **C. G. C. Laboratories, Inc.,** 391 Ludlow St., Stamford, Conn.

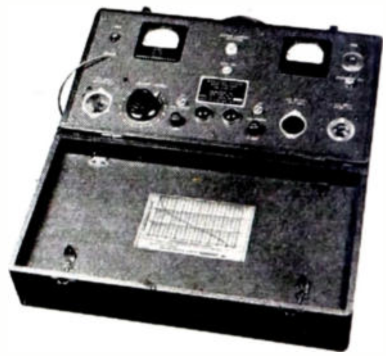


The maximum inductance of this unit, at zero control current, is 30 mh which can be reduced to 1/400 of this value with 100 ma peak control current. The corresponding frequency variation is 20 times the starting frequency. The approximate range of starting frequencies for zero current is between 10 and 100 kc. The H3M 3H9L type features a linear relationship between frequency shift and control curret extending over at least a 5:1 change of frequency.

Data sheet upon request.

* Trademark

Testor



Transitron, Inc., 154 Spring St., New York 12, N. Y., is now in production on an insulation tester with continuously variable output from 0 to 15 kv dc. The instrument will perform accurate resistance measurements and nondestructive breakdown tests on all types of materials. Tests may be made on motor windings, ignition systems, wiring harnesses, insulators, connectors, etc. Point of breakdown is clearly visible. In addition, the unit can be used as a laboratory high-voltage power supply.

Panel meters provide direct indication of the applied voltage and leakage current through the test specimen with an accuracy of 2 per cent. The tester can be used with complete safety. It will operate on 115 v 60 cps and/or 6, 12, 24 and 32 volts dc.

(Continued on page 75A)

FERROXCUBE ·3C cores are nickel-free

APPLICATIONS:

- I-F TRANSFORMERS
- PERMEABILITY TUNING DEVICES
- LOW-LOSS INDUCTORS
- SATURABLE CORE REACTORS
- HORIZONTAL OUTPUT TRANSFORMERS
- DEFLECTION YOKES
- TELEPHONE LOADING COILS

When your drawings call for Ferroxcube 3C cores for your TV deflection yokes and horizontal output transformers, you can forget about procurement problems. These ferrite cores are nickel-free . . . . and delivery will be made exactly as scheduled by you!

Improved temperature stability, high saturation flux density, and high permeability are among the other advantages of Ferroxcube 3C.

Complete technical data is yours for the asking in Engineering Bulletin FC-5101A, available on letterhead requests. ★ ★ ★ ★ ★ ★ ★ ★ ★

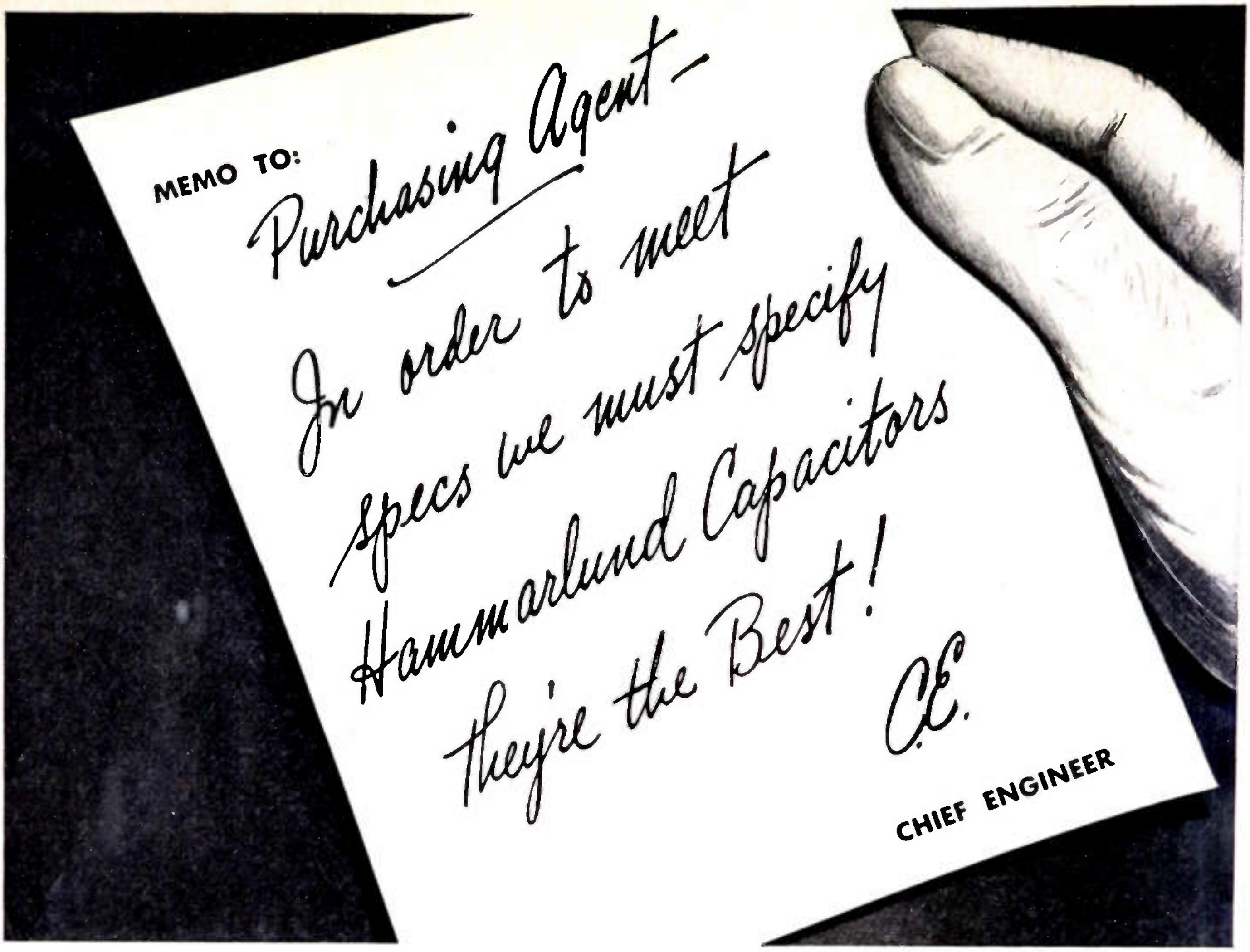
FERROXCUBE

THE MODERN CORE MATERIAL

FERROXCUBE CORPORATION OF AMERICA

A Joint Affiliate of Philips Industries and Sprague Electric Co., Managed by Sprague

SAUGERTIES, NEW YORK

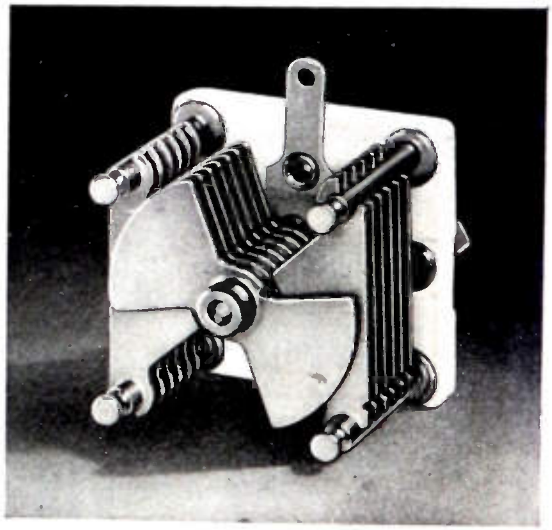


They've always been the best!

VARIABLE CAPACITORS

Hammarlund Capacitors, backed by 42 years of design, engineering and production experience, are today recognized by the military services, electronic manufacturers and research engineers, as the finest quality capacitors available. Millions of them are in use by almost every important manufacturer of electronic equipment.

See for yourself. Write today for the 1952 CAPACITOR CATALOG. It has the complete story.



HAMMARLUND

MORE THAN 40 YEARS EXPERIENCE COUNTS!

THE HAMMARLUND MANUFACTURING CO., INC.
460 WEST 34th STREET • NEW YORK 1, N. Y.

NEW EXPANDED LINE –

OF E-I STANDARD terminals

48 STANDARD TYPES NOW AVAILABLE TO SIMPLIFY YOUR DESIGN PROBLEMS, SPEED DELIVERIES, REDUCE COSTS!

16 ADDITIONAL TYPES!

The increasingly popular E-I STANDARD LINE of SEALED TERMINALS now includes 16 additional types making a total of 48 items that can be ordered direct from stock with prompt delivery preassured. Our application engineers believe that this new expanded group of standard items could readily solve the majority of sealed terminal problems thereby eliminating much of the time and expense involved in custom design and production.

All 48 types are currently being specified in great numbers for an extremely wide range of applications, thus users of these types benefit by the additional economy of large scale production. For complete information covering all 48 types write today for Bulletin 949-A.

TYPE: **AAA-30W-HS**

TYPE: **AAA-30W-HP**

TYPE: **RR-60W-PP**

TYPE: **AAA-30W-SX**

TYPE: **CCS-80W-XP**

TYPE: **CCS-80W-HP**

TYPE: **AA-40W-SP**

TYPE: **AB-60T-SX**

TYPE: **C-80W-2/56P**

TYPE: **AB-60W-SS**

TYPE: **C-75T-SS**

TYPE: **AB-60T-LX**

TYPE: **B-60W-SS**

TYPE: **ABS-40W-HH**

TYPE: **ABS-40W-XP**

TYPE: **ABS-40W-HP**

E-I . . . YOUR HEADQUARTERS FOR HERMETICALLY-SEALED MULTIPLE HEADERS, OCTAL PLUG-INS, TERMINALS, COLOR CODED TERMINALS, END SEALS, etc. WRITE FOR CATALOGS.

E I

ELECTRICAL INDUSTRIES

MANUFACTURERS OF SPECIALIZED ELECTRONIC EQUIPMENT

44 SUMMER AVENUE, NEWARK 4, NEW JERSEY

DIVISION OF AMPEREX ELECTRONIC CORPORATION

RCA
VIDICON

TECHNICAL FEATURES

- ✓ Only $6\frac{1}{4}''$ long and $1''$ in diameter
- ✓ Provides 400-line resolution
- ✓ Spectral response approaching that of the eye
- ✓ Sensitivity permits televising with 100-200 foot-candles of illumination
- ✓ Designed for use with commercially available camera lenses
- ✓ Operates with low dc voltages

ACTUAL SIZE →

NEW, small-size camera tube for low-cost industrial television

Now, the RCA-developed 6198 Vidicon extends the advantages of television coverage to countless industrial users . . . opens the door to simplified television camera designs.

The small size and simplicity of operation of this television camera tube facilitates the design of compact and low-cost television camera equipment—including equipment for closed-circuit, portable, and remote-control applications.

Produced in the same plant by the same skilled hands that make the RCA Image Orthicon camera tube, the RCA-6198 offers the detail of 400-line picture quality at low unit cost. It employs magnetic focus and deflection, and operates with relatively low dc voltages.

Utilizing a photoconductive layer as its light-sensitive element, the RCA-6198 has a sensitivity which permits televising scenes with 100 to 200 foot-candles of incident illumination. The photoconductive layer has a spectral response characteristic approaching that of the eye. The dimensions of the useful area of this layer are such that stock camera lenses can be employed. The size and location of the layer permit a wide choice of commercially available lenses.

The following components, designed for use with the RCA-6198 Vidicon, are also available:

RCA-216D1 Deflecting Yoke
RCA-217D1 Focusing Coil
RCA-218D1 Alignment Coil
RCA-233T1 Horizontal Deflection Transformer
RCA-234T1 Vertical Deflection Transformer

For complete data on the RCA-6198 Vidicon and associated components, write RCA, Commercial Engineering, Section JR47, Harrison, N. J., or contact your nearest RCA Field Office.

FIELD OFFICES: (East) Humboldt 5-3900, 415 S. 5th St., Harrison, N. J. **(Midwest)** Whitehall 4-2900, 589 E. Illinois St., Chicago, Ill. **(West)** Madison 9-3671, 420 S. San Pedro St., Los Angeles, California.

THE FOUNTAINHEAD OF MODERN TUBE DEVELOPMENT IS RCA

RCA

RADIO CORPORATION of AMERICA
ELECTRON TUBES **HARRISON, N. J.**

TMK. ®

Changes of address (with advance notice of fifteen days) and communications regarding subscriptions and payments should be mailed to the Secretary of the Institute, at 450 Ahnaip St., Menasha, Wisconsin, or 1 East 79 Street, New York 21, N. Y.

* Numerals in parentheses following Directors' names designate Region number.

PROCEEDINGS OF THE I.R.E.®

Published Monthly by

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VOLUME 40 — *October, 1952* — NUMBER 10

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

Responsibility for the contents of papers published in the PROCEEDINGS OF THE I.R.E. rests upon the authors. Statements made in papers are not binding on the Institute or its members.

Ernst Weber

DIRECTOR 1952-1954

Ernst Weber was born in Vienna, Austria, on September 6, 1901. He received the diploma of electrical engineering in 1924, from the Technical University in Vienna, and the Ph.D. degree in 1926, from the University of Vienna. On the basis of his studies as a research engineer at the Austrian Siemens-Schuckert Company, Dr. Weber received the Sc.D. degree in 1927 from the Technical University in Vienna.

Upon request, Dr. Weber transferred to the Siemens-Schuckert Company in Berlin-Charlottenburg in 1929, where he also was appointed lecturer at the Technical University. In 1930, he was invited, as visiting professor, to the Polytechnic Institute of Brooklyn, and in 1931, was appointed to the permanent position of research professor of electrical engineering in charge of graduate study. From 1942 through 1945, he was head of graduate study and research in the electrical engineering department of the Institute.

During World War II, Dr. Weber was named official investigator by the Office of Scientific Research and Development. In recognition for his contributions to its research group, he was awarded the Presidential Certificate of Merit. The Microwave Research Institute of the Polytechnic Institute of Brooklyn, and the Polytechnic Research and Development Company grew out of this wartime research. Dr. Weber has been head of the Microwave Institute since 1945, and was recently elected president of the Polytechnic Company.

Dr. Weber has contributed scientific papers to technical publications on electromagnetics, linear and nonlinear circuits, and magnetic amplifiers.

A Member of the IRE in 1941, Dr. Weber became a Senior Member in 1946, and was elected Fellow in 1951. He has been actively associated with numerous IRE Committees. He is also a fellow of the American Institute of Electrical Engineers and the American Physical Society.

Radio Spectrum Conservation

Like most arts, or even sciences, radio communication has grown without full pre-testing of its various aspects, coordination between them, and logical evolution of its methods and equipment. Progress has been only partly systematically and pragmatically guided; often enough advances have been of empirical origin. As a result, the utilization of the radio-frequency bands and channels has been partly a matter of afterthoughts, compromises, and reluctant adaptations to the existing limitations. Even though much time and capable thought have wisely been devoted to channel and band allocation, and with resulting benefit to the effectiveness and utility of radio communication, present day demands threaten to exceed the supply of frequencies presently available.

The time had clearly come for an over-all survey and appraisal of the channel resources of radio, and of the best ways in which to use them. Although it would presently be impracticable to make certain basic changes to the frequency-allocation structure, a recapitulation of the progress and errors of the past would be of considerable value in avoiding mistakes in the future and insuring the continued expansion of our radio communication facilities.

Such a comprehensive survey has recently been completed under the auspices of the Joint Technical Advisory Committee (formed and encouraged by The Institute of Radio Engineers and the Radio-Television Manufacturers Association). In its report, the Committee has studied the radio spectrum in detail, developed the preferred methods of its use, pointed out certain past limitations and present problems, and offered valuable information for the guidance of those entrusted with future frequency allocations. The task of the Committee was a monumental one, and the results of the Committee's efforts are correspondingly basic and constructive.

This report has now been published in a volume entitled "Radio Spectrum Conservation" which was prepared by a selected group of contributing experts under the coordinating guidance of the Joint Technical Advisory Committee and which contains the analyses and conclusions described above. Further information as to the availability and contents of this volume appears on page 1256 of this issue.

The work of the Joint Technical Advisory Committee should prove to be an important contribution to the conservation of one of our most precious natural resources, the radio spectrum.

—*The Editor*

Radio: A Coalescence of Science and the Arts*

I. S. COGGESHALL†, FELLOW, IRE

ONE OF THE MORE enduring rewards realized by an engineer active in the field of radioelectronics is his simultaneous immersion in the atmospheres of science and the arts.

Although it is not a new thing under the sun to find a personal blend of inventive and artistic attributes embodied in a Leonardo or a Morse, it has remained for recent generations to produce engineers who have made all art their debtor by lifting the dynamic range of sound by the decibel, and of light by the lumen, and putting both under manipulative control.

During the long history of music and the graphic arts, men were denied the aid of mathematics and applied physics in the solution of their problems. Stradivari had to grope empirically for the golden tonality of his justly famous violins. The pipe organ of antiquity, the classical reed, string, and pipe instruments, the harpsichord and piano, and finally Edison's phonograph similarly were products of cut-and-try, long before designs could be based on the laws of nature. All of them together left unplumbed the thresholds of hearing and feeling.

The vacuum-tube amplifier and its associated electrical circuits and acoustical equipment—brain-children of members of this Institute—changed all that. Today the frequency gamut of sound has been dissected and related to the physiology of hearing and the psychology of listening. Every portion of the spectrum has been segregated, scrutinized, analyzed, and synthesized, using the tools of Fourier, Helmholtz, Pupin, de Forest, and a host of others. Pitch, volume, and color are brought under complete control. With amplifiers, resonance and delay circuits, and filters, we forge our waveforms and hammer them into shape, add dimension through reverberation and binaural handling. Like building blocks we expand and compress frequency envelopes and volume ranges, translate them in position, modulate them to carriers, record them on discs, film, wires, and tapes, and, finally, tailor-make their reproduction to the dictates of time, space, occasion, and audience preference.

The nearer to perfection our electro-acoustic instrumentalities have approached, the closer has been the effect of the music reproduced or projected to satisfying the soul. In a happy blending of the practical with the beautiful, the realm of music has been vastly expanded and enriched by the influence of radioelectronic techniques.

Portrait of Lee de Forest presented to the IRE by the artist, Harriet de Forest, daughter of the famous inventor.

Great musicians have lent their enthusiasm to the mastery of the new medium. Engineers and artists and their organizations have collaborated, in their quest for fidelity of transmission and reproduction, to place the microphone symbolically at the pinnacle of science and the arts.

As a result, throughout an expanded breadth of the audible spectrum, throughout an augmented depth of dynamic range of volume, the nuances of musical inflection now play directly upon the human emotions. Engineering has attained, through art, a poignant contact with man's aspiring. The achievement of simultaneously bringing millions of people into the orbit of the new relationships is a cultural revolution of the first magnitude.

Yet I would be traitor to my trust if I did not express on your behalf a longing for something even better than we have now. The equivalent of Beethoven's Ninth or the Ride of the Valkyries has not yet been written for electronics presentation in the future! For the hundred-piece orchestra and the Metropolitan stage as now constituted, without the benefit of electronic devices, are already dated hold-overs of a by-gone day.

Modern ingenuity must be capable, not necessarily of replacing, but certainly of augmenting the oboe and the piccolo. Musical "prime-movers," containing electro-acoustical circuits in place of strings, bladders, and columns of air must be somewhere around the corner, offering agreeable sounds never before heard on earth.

Again, when the second violin section of an orchestra consists of eight identical instruments playing in identical time, pitch, and volume, the eight-fold amplification and slight phase displacements of the individual strings are dearly bought in the form of salaries instead of a box-full of electrical components. How usefully, at the same expense, the other seven could be employed at tasks which the composer of the future will assign to them, when he is relieved of the present payroll restraint on his composition!

Then think of the audience of the future as having dropped its accustomed shackles to a point-source of sound (even though it be a 100-point concentration on a stage). Through control from an electronic console, manipulated under the baton of the conductor, the music could leave the stage, be made to surge overhead, march up and down the aisles, saturate the people in the audience with its omnipresence, envelope them in its ethereal volume, move them with its power, and play upon the spinal cords of their emotions!

Enough of what can be done with a few acoustical watts. Time will permit of my referring only briefly to the fact that the kinescope and the photoelectric cell are now performing a parallel miracle in the graphic arts, doing for sight what the microphone has done for sound.

From Giovanni Bellini through Rembrandt to Constable, a long succession of painters experimented in emotional contrasts of light with shade without finding anything much blacker than black or whiter than white. Hogarth created new effects by dragging multicolor pigments across the canvas instead of mixing them on the palette; Monet adopted a brush technique to present an admixture of color on canvas directly to the eye, and Van Gogh let pure sunshine bathe the mosaic of his work in a

* Decimal classification R060. Original manuscript received by the Institute, May 14, 1952. Speech given at the ceremony of presentation to the IRE of a portrait of Lee de Forest by the artist Harriet de Forest, May 7, 1952.

† Western Union Telegraph Company, 60 Hudson St., New York, N. Y.

sort of secondary emission. Yet none of them increased by one iota the stimulus to the rods and cones of the retina which do our seeing for us.

Then radio and photography joined in meeting the problems of developing and printing sound track on motion-picture film, and the decibel found itself struggling in solutions of silver and hypo. First, in applications to light and shade, then to the color problems of the movies and television, radio moved boldly into a new career in colorimetry and photoelectric science. Here its mission is to do for the eye what it has already done for the ear. Again it finds itself dealing in physiology, this time the physiology of seeing; and in psychology, this time the psychology of perceiving. And here its work is unfinished, but it holds promise of not only developing instrumentalities beyond the comprehension of the masters, but of extending their benefits to the masses.

Progress in electronics, then, goes beyond sight and sound into the realm of feeling in its most hallowed sense. It is a great thing to be privileged to work, thus, with the ultimate fiber of civilization. The radio technician is permitted to share with the artist, and with the stone mason who builds a cathedral, a sense of participation in works which transcend the limitations of time and space and become, in fact, supernatural.

Since it was my privilege, during my term of office as President of The Institute of Radio Engineers, to deal with the artist, Harriet de Forest, in the initial stages of the execution of the portrait of her father, Lee de Forest, which she is presenting to the Institute, I have been pleased to speak a few minutes by way of assuring her that, as an artist among engineers, she is among friends of her art.

Added to that is the respect in which we all hold Dr. de Forest, recipient of the Institute's Medal of Honor in 1922 for major contributions to the communications arts and sciences; President of the Institute in 1930; President, in 1909–1910, of its predecessor, the Society of Wireless Telegraph Engineers; and, as reflected in Dr. de Forest's autobiography, one who himself revels in the art forms which his vacuum tube and associated circuits have brought to the service and satisfaction of mankind.

Adding further to the pleasure of having the artist and her daughter present is the drama inherent in the circumstance that the portrait of a man whose name a President of the United States has said will outlive his own has been executed by his talented and beloved first-born, and is about to be presented to an organization which, for so many years, has been close to his heart.

Technical Writing Grows into New Profession: Publications Engineering*

ROBERT T. HAMLETT†, SENIOR MEMBER, IRE

Summary—**Engineering-level technical writing is described as requiring, foremost, the skills and knowledge of an engineer and, secondly, the ability to write well. For this combination of work the term "Publications Engineer" is proposed. The writer's participation in an engineering project is outlined on a time basis, starting with the sources of information and completed with delivery of the printed work. Satisfying aspects of the field are discussed and the future is predicted as of growing value to the engineering profession as a whole.**

Introduction

THE TREMENDOUS EXPANSION in the size and productiveness of the engineering profession has been due, in a large measure, to the ability of research and development engineers to enlist other engineers for special tasks or services related to their basic problems. It was not so many years ago that an *engineer* was *the engineer*—he was charged with responsibility for all *engineering* work on a project. This was possible because the end result of his engineering work was usually a single unit or instrument which operated without "tie-in" or reference to other equipment. He found time somehow to solve all of the engineering problems that arose in connection with his "brain child."

But the modern era of "systems" rather than "instruments" has changed the engineering approach to a very marked degree. One hears now about systems engineers, product engineers, project engineers, standards engineers, administrative engineers, test engineers, field engineers, production engineers, packaging engineers, industrial engineers, and so on. What has happened? Simply that the individual engineer cannot any longer carry all the burdens of the job of "engineering" of a system or even of a single instrument which ties into a system. While a very gifted engineer, possessing high skill in many branches of engineering, may still be able to visualize and guide the work on his project, he is no longer able to carry on the many individual investigations, attend the frequent engineering conferences, plan the fiscal and field-testing programs, solve the production and packaging problems, or create the publications which are necessary.

This ability of the engineer to pass on responsibility to other engineers has given rise to still another field of specialization within the engineering profession—that of TECHNICAL WRITING. (See Fig. 1.) The products of this new field are instruction books, training manuals, engineering reports, technical data sheets, and many other types of technical information, a sampling of which appears in Fig. 2. The workers in this field are referred to as "Technical Writers," "Engineering Writers," "Specification Writers," "Technical Report Writers," and the like. This author prefers to call the workers in this field "Publications Engineers," in keeping with other well-established titles such as "Standards Engineer," "Test Engineer," and "Field Service Engineer." This new title will be used throughout the article.

What Is a Publications Engineer?

The principal reason why this author prefers the new title "Publications Engineer" to that of "Technical Writer" is that it more clearly designates the duties of such a worker, and also places him in a proper professional status with fellow engineers, where he rightly belongs. For he is an engineer first, and secondly a writer. The term "Technical Writer," as commonly accepted, refers to a writer who writes material on technical subjects to various levels of intelligence but who is not usually concerned with the actual publication processes and problems.

The Publications Engineer is an engineering specialist who relieves other engineers of the major portion of the responsibility for production of all publications required as a result of the engineers' work. The Publications Engineer writes technical material, plans and directs preparation of copy, and carries through on all details concerned with actual production of the publication. It is necessary to repeat that he is first an engineer, then a writer, and finally, a publication man.

Engineers have always labored under the stigma that they cannot write well. It is a common attitude, even in precollege education, to assume that because the student is superior in mathematics he must be inferior in English. This affects the student's attitude and he very naturally uses it as an excuse for not seriously studying the subject in which he is prejudged to be inferior. When the "superior" math student goes to engineering school, it is a foregone conclusion that there is very little that can be done to help him there. However, he is given one or possibly two courses in English (especially "arranged" for engineers) early in his college work. No further attempts are made to help him overcome a deficiency which will handicap him throughout his entire career.

There is no doubt that some engineers cannot write—but some lawyers, some accountants, and some doctors cannot write well! Some doctors do not develop a pleasing "bedside" manner, so they become fine surgeons or specialists. So some engineers do not take time to write well, and because of this other engineers now find an interesting and well-paid profession.

* Decimal classification: R071. Original manuscript received by the Institute, April 21, 1952.
† Engineering Department Head for Publications, Sperry Gyroscope Co., Great Neck, L. I., N. Y.

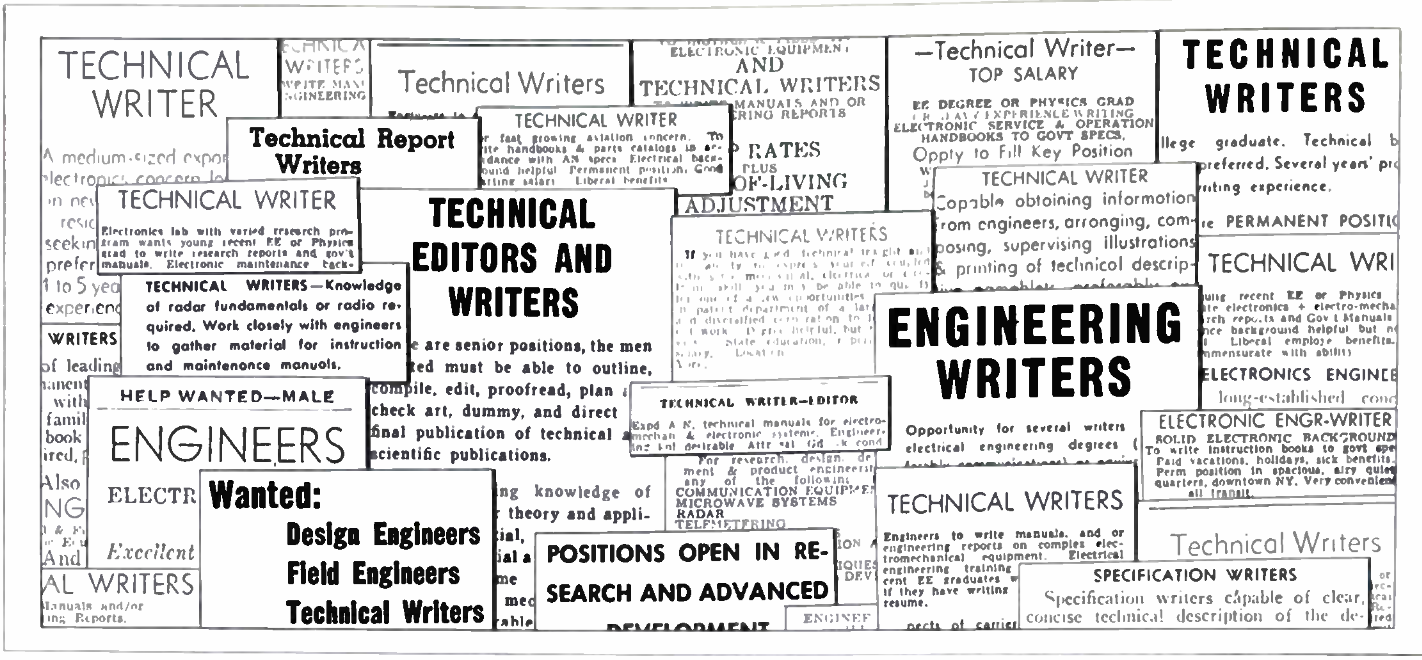

Fig. 1—Demand for engineer-writers, identified herein as "Publications Engineers," is evidenced by these classified advertisements selected at random from newspaper and magazine employment sections.

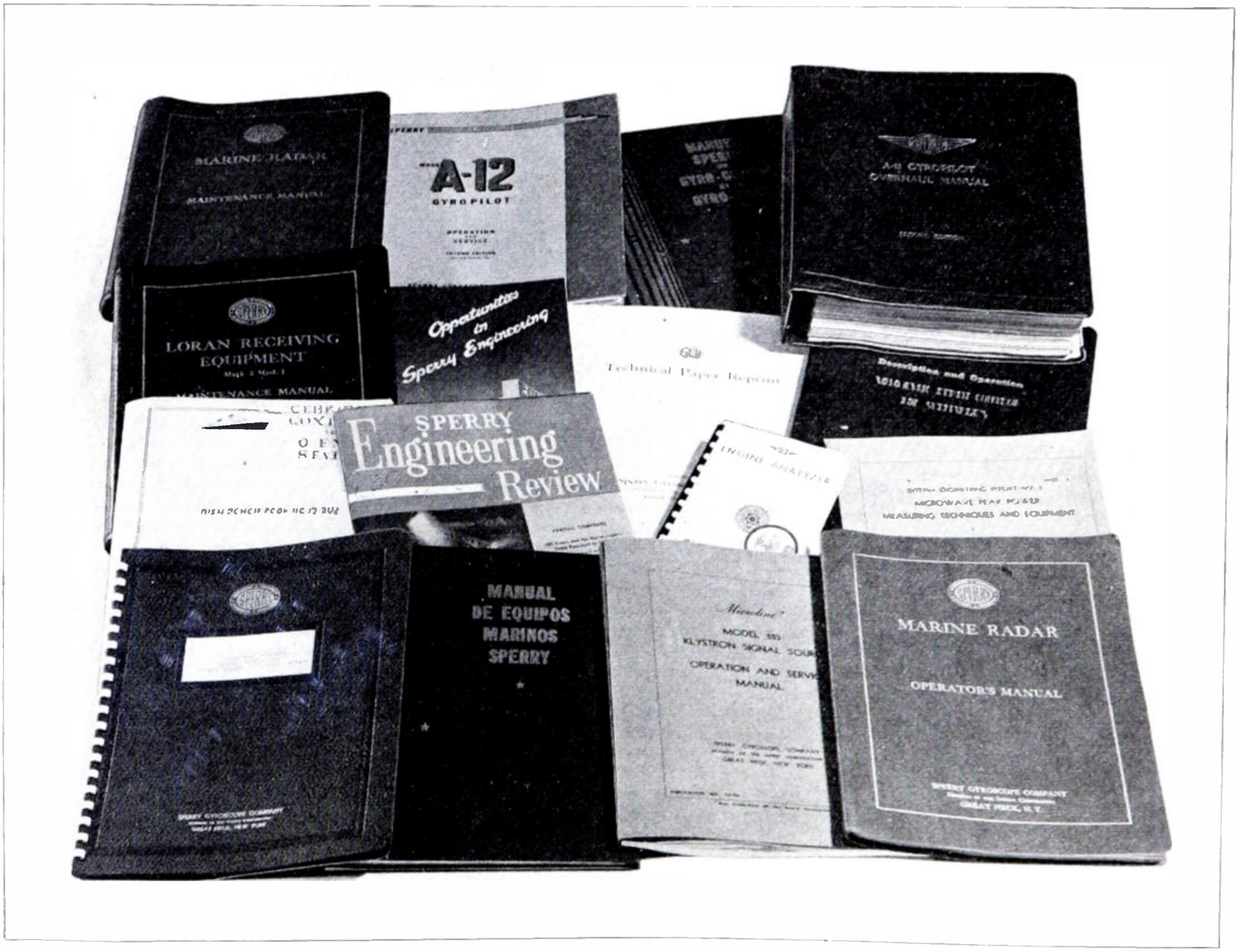

Fig. 2—Publications Engineers produce a variety of printed matter requiring the combined skills of the engineer and the writer.

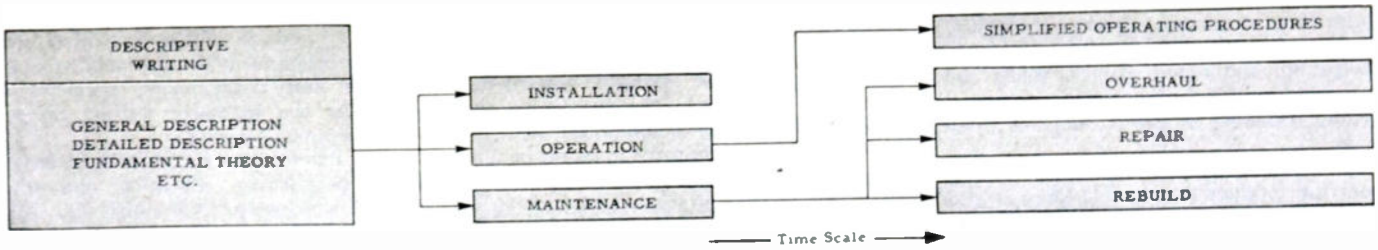

Fig. 3—Typical writing assignments on a system.

The Publications Engineer must be an engineer who has writing aptitude. This aptitude may have never become very obvious because of the misguidance and lack of encouragement received during his education. The author has seen many engineers, who felt certain that they were below average in writing aptitude, develop into excellent writers of technical material. No one can doubt that the engineering profession would be in a much better position if there were more effective writers among us. (The same might be said for speakers.)

The Publications Engineer must be an engineer with unquenchable thirst for learning. If he is a mechanical engineer, he must be learning more about electronics; if he is an electrical engineer, he must be learning about aerodynamics, hydraulics, and the like. He is constantly challenged to describe something about which he knows practically nothing. But with his basic engineering education under his hat, he tackles each unknown with some confidence that he can understand and interpret it for others who may know more or less about it than he does. Many fine technical descriptions result when engineers who are educated in one field begin to write on subjects in other engineering fields; they use analogies which help the reader in applying the description to his own experience.

The Publications Engineer must have a working knowledge of the advantages and disadvantages of many types of reproduction processes, such as spirit duplication, mimeograph, Photostat, blueline, and blueprint, Ozalid, and offset printing and letterpress printing. He is familiar with type faces, paper stock, cover materials, binding methods, and the like. He understands the problems involved in production of copy by typewriters, Varitypers, typesetting, and phototype. He has a practical knowledge of the arts of photography and retouching, and he guides technical illustrators in visualizing and rendering special illustrations for use with his written words.

All of his talents and acquired knowledge are combined in the process of preparing a publication that must meet government or commercial specifications covering content, format, practicability, and literary standards. He is at the same time an engineer, a writing specialist, a publications expert, and a student of psychology!

Variety of Work

When the young Publications Engineer has overcome his inferiority complex in tackling new writing projects, he finds the variety of writing assignments to be one of the most attractive features of his job. It is a familiar complaint among engineers that they become too specialized and know too little of what is taking place in the scientific world around them. While no scientist can hope to keep abreast of the tremendous evolution of technical achievements now taking place, the Publications Engineer finds real satisfaction in testing and adding to his knowledge in many different fields. As an example, in the author's company the skilled Publications Engineer develops a descriptive knowledge in such varied fields as radar,

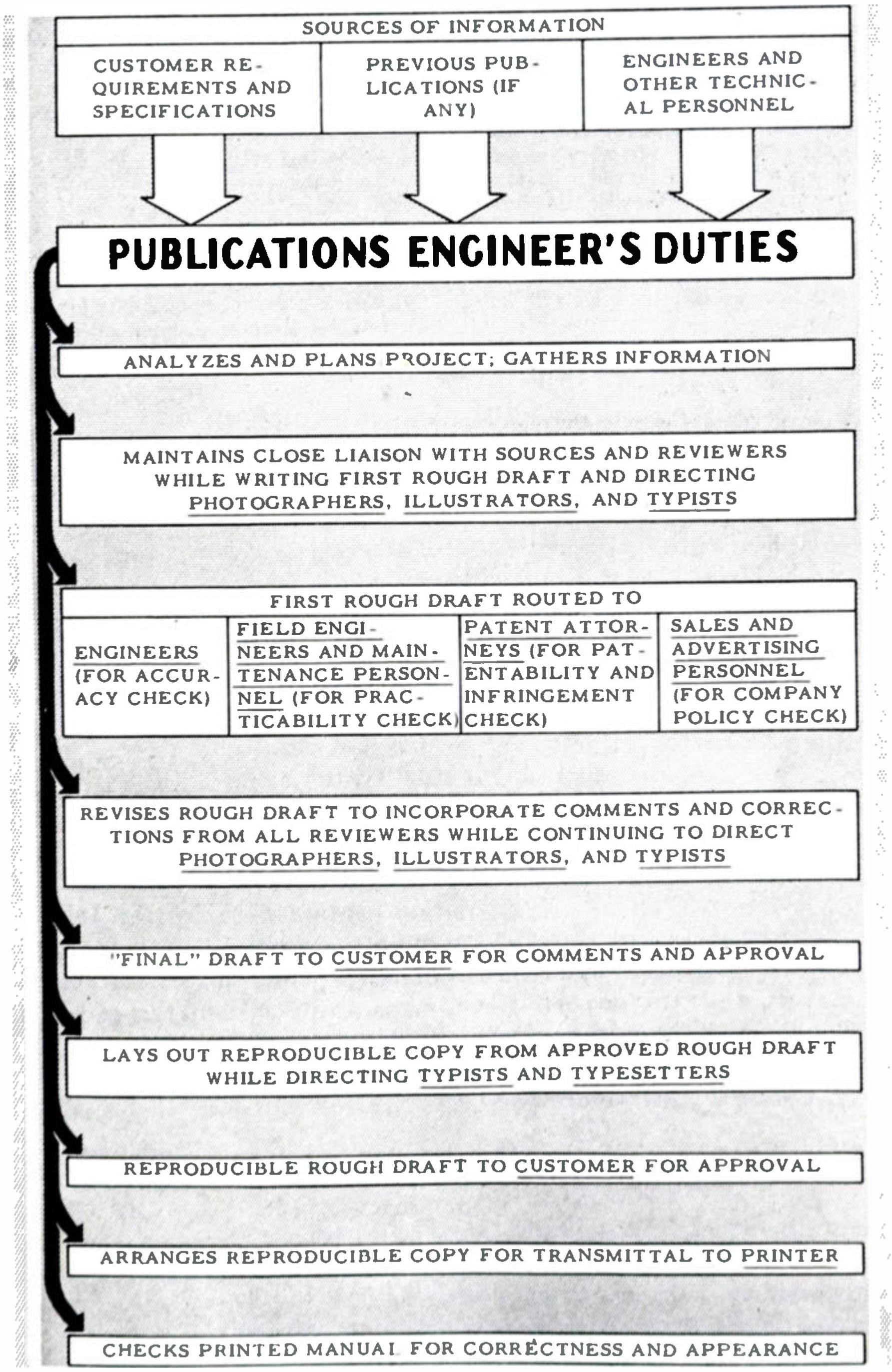

Fig. 4—Publications Engineer participates in every phase of manual's "life cycle"; he must gain broad knowledge of product's engineering, manufacture, and customer's application. Persons with whom he deals directly are underscored.

hydraulics, servomechanisms, gyroscopics, computing mechanisms, ballistics, optics, navigation, and aerodynamics. When the occasion demands, he becomes, for a time, a writing specialist in one or more of these fields.

In addition to the variety of writing from the product standpoint, there is also much variation in the material to be gathered on any one product or system. Fig. 3 illustrates some of the writing assignments on a single system. Some of the assignments require the Publications Engineer to work intimately with the equipment; in some cases he completely disassembles and reassembles the units. In other cases, he accompanies the equipment on trial runs or field tests. These experiences give a "practical" satisfaction to those who like to feel that they are not just "theoretical" writers.

Another attractive feature of the Publications Engineer's work lies in the variety of contacts which he makes in the course of the development and approval of a publication. Fig. 4 shows a typical "life story" of an instruction book prepared for the Armed Services. The underlining in the diagram gives an indication of the many individuals concerned in the preparation or approval of the publication prior to its final printing; the Publications Engineer works constantly with all of those shown.

The Future for Publications Engineers

Young engineers often raise the question as to the future of Technical Writing or Publications Engineering. There are several factors which appear to be of importance in attempting to predict the future—but to the author they all look favorable toward increasing opportunity for this new profession. First, the complexity of equipment and systems certainly will continue to increase; automatic control is the ultimate goal of nearly all future instrumentation, and with such control always comes increased technical complexity. With increasing complexity there is greater need for more complete instructional material. As one associate put it, "the equipment becomes more complex but the intelligence of the average user remains the same." Second, granted that complexity will increase, there is the immediate following condition that the equipment will be much more costly and must be repaired rather than replaced. This adds again to the need for publications which will be adequate for the purpose. The funds allocated for publications will necessarily increase, but will still be a very small portion of the total cost of the equipment. Third, if the caliber of engineering graduates coming into Publications Engineering is maintained or raised, there will be a broadening in the scope of their work since they themselves will develop opportunities for using their special skill to supplement the work of other engineers. This is a very important responsibility in any new profession—to develop and broaden the particular skills and to offer them to others.

Conclusion

Publications Engineering is a new profession which has grown rapidly in the past few years because of the increasing complexity of equipment and the inability of the research and development engineers to undertake the extensive writing projects which became necessary.

The Publications Engineer must have a sound engineering education and must possess writing aptitude—although it is pointed out that the possession of this aptitude may not be realized by many young engineers.

The Publications Engineer develops a knowledge of the reproduction and printing processes, and can guide the publication through all of its stages from rough draft to its printed form.

The variety of work assignments and personal contacts appeal greatly to certain engineering graduates. Some of the writing arrangements cover theoretical aspects, others are along practical lines where the writer works closely with the equipment in the factory or in the field.

The "personal-satisfaction" factor is quite high for the Publications Engineer since his assignments are usually of short duration, compared to those of the engineer, and he "sees" the final results of his labors at more frequent intervals.

Finally, the future of this new profession looks promising because of the trend towards more complex equipment and the accompanying requirements for more complete handbook and engineering report coverage. The future also depends upon the efforts which Publications Engineers make to find new areas of service to the engineering profession.

Correction

F. R. Abbott, author of the paper, "Design of Optimum Buried-Conductor RF Ground System," which appeared on pages 846–852 of the July, 1952 issue of the Proceedings of the I.R.E., has requested that the editors publish the following corrections:

Delete superscript 2 from the end of the sentence above equation (14) on page 848.

Delete subscripts r and u from J in Fig. 1 on page 848.

Bridges Across the Infrared-Radio Gap*

MARCEL J. E. GOLAY†

The following paper is presented as a part of a series of valuable tutorial papers. It is published with the approval of the Tutorial Papers Subcommittee of the IRE Committee on Education.—*The Editor.*

Summary—The wide gap which exists between the generation and detection of coherent radiation on one hand, and the generation and detection of submillimeter incoherent radiation on the other hand, is discussed from the standpoint of spectroscopy.

While the resolving powers possible with microwave spectroscopy can be indefinitely extended with refinements in technique, the resolving powers possible with infrared spectroscopy are subject to an upper limit determined by the size of the external optics utilized. In the far infrared at wavelengths greater than 10 microns the resolving powers obtained in the past have been subject to further instrumental limitations, but the attainment of the optically allowable resolving powers appears possible now with recent developments in infrared spectroscopic instrumentation.

Introduction

In the past ten years the radio engineer has succeeded in decreasing the wavelength of coherently generated radiation from around 10 cm down to a little over 1 mm. This hundred-fold extension of his mastery over the radio spectrum has been accompanied by the spectacular technical advances in communication and radar with which we are familiar. However, as we shall proceed to still shorter wavelengths, it is safe to predict that because of the increasing opacity of the atmosphere to such radiations, our interest in these will become more purely scientific.

We designate by the name of "microwave spectroscopy" the vigorous new science which has been made possible by the recent extension of radio techniques to the cm and mm spectral regions. Molecular-absorption spectroscopy forms one important branch of this new science, and represents the one scientific endeavor which is shared by the microwave spectroscopist and the infrared spectroscopist. Both want to know something about the frequencies at which molecules resonate, and, in order to find out, both must generate radiation of a known frequency and measure it after passage through an absorption cell filled with these molecules.

In the 6-mm wavelength region, where we know both how to generate and heterodyne coherent or cw radiation, the powers and sensitivities available for microwave absorption studies are overwhelmingly adequate.

In the region around 1 mm we know only how to generate pulsed, narrow-banded radiation; the microwave spectroscopist working in this region has borrowed the detector of the infrared spectroscopist until recently, and is only now learning how to detect this radiation electrically.

At wavelengths less than 1 mm the molecular spectroscopist relies entirely on far infrared sources, on optical elements such as gratings, prisms, or interferometers, and on far infrared detectors. It may be illustrative to make an estimate of the number of orders of magnitude which separate radio generators from infrared generators and radio detectors from infrared detectors.

The power generated within a frequency band Δf by an infrared source at the absolute temperature T and having an area approximately $\frac{1}{2}$ wavelength square, which can be called a point source, is given by the expression

$$W = kT\Delta f. \tag{1}$$

For the purpose of comparison, assume a klystron generating 1 watt of coherent 3-cm radiation. Associate with this radiation a bandwidth determined by the uncertainty in our relative knowledge of this frequency with respect to a former measurement. One cps can be taken as a reachable figure for Δf if an excellent "flywheel," such as a quartz frequency standard, is utilized with loving care, as it should be in certain phases of microwave spectroscopy. Substituting these values for W and Δf in (1) and solving for T, we obtain the temperature needed for the same emission by an infrared point source. This temperature is $T = 7 \times 10^{22}$°C. Such a temperature does not exist any place in the world and is some 19 orders of magnitude greater than any practical laboratory temperature.

So much for generation. The infrared detection picture is equally rough. In the case of coherent radio detection, the least detectable power ΔW is also given by the second member of (1), except for the noise factor of the detector which should multiply this second member. Assuming 1 second for our measurement, or $\Delta f = 1$, and a detector at room tempertaure with a noise factor of 10, we obtain

$$\Delta W = 4 \times 10^{-20} \text{ watts.}$$

On the other hand, far infrared detectors are the so-called thermal detectors, and their greatest possible sensitivity can be calculated as though they were broadbanded radio receivers sensitive in the entire frequency spectrum. In this case we cannot use the formula above, which would give us infinity when Δf is infinite. The quantum conditions must be observed for the short

* Decimal classification: R111.2×535. Original manuscript received by the Institute, December 5, 1951; revised manuscript received, June 11, 1952.
† Signal Corps Engineering Laboratories, Fort Monmouth, N. J.

wavelengths of the spectrum, and when this is done we obtain for the square of the radiation fluctuations to and from the detector,

$$\overline{\Delta W^2} = 16kk_1T^5S\Delta f = 4k_1T^4S\cdot 4kT\Delta f, \quad (2)$$

where k_1 is the Stefan-Boltzmann radiation constant and S is the sensitive area of the detector. An analogy can be noted between this expression and the expression for the shot noise, $\overline{\Delta i^2} = 2\ i\ e\Delta f$. The radiation flow to and from the detector, k_1T^4S, is analogous to the current i, and the analogue to the charge of the electron is $4kT$, which represents, therefore, a measure of the average grain size in the quantized stream of radiation to *and* from the detector. At room temperature, for 1-mm wavelength for which we can assume a half-mm square for S, and for an observation time of 1 second or $\Delta f = 1$, the square root of this expression gives the smallest power detectable by an ideal thermal detector

$$\Delta W = 2.5 \times 10^{-12} \text{ watts.}$$

When this value is compared with the value obtained above for the case of radio detection, thermal detection is seen to be eight orders of magnitude less sensitive than radio detection. The presence of ΔW^2 as a square term implies ignorance of the nature of the radiation detected, and the radio spectroscopist who must resort to thermal detection of his high-quality coherent or nearly coherent radiation is rather in the position of a violinist who takes his Stradivarius to the pawnshop, to be given an estimate of its value based on the BTU content of the violin's wood. Yet, at submillimeter wavelengths, it may be the best he can do.[1]

With these premises, it will be assumed that the radio spectroscopist has a two-fold interest in infrared techniques. When he can generate, but not detect coherent radiation, he has a direct interest in knowing what sensitivity he can expect from practical thermal detectors. In the shorter wavelength region, in which he can neither generate nor detect coherent radiation, he has a physicist's interest in learning what the infrared spectroscopist can accomplish in the way of advance scouting of the spectrum in the region below 1-mm wavelength. Accordingly, some developments in the thermal detection of radio waves, and some other developments in purely infrared spectroscopic instrumentation will be reviewed in what follows.

Discussion

The two best known infrared detectors, the thermopile and the bolometer, come within one and a half orders of magnitude of reaching the sensitivity limit given by (2). However, the full realization of such a practical sensitivity is predicated upon their being effectively blackened for complete absorption of incoming radiation. What looks like a good black in the visible region could be a very poor black in the 10-micron region. Actually, the effective blackening of the detectors in the 10-micron region is not difficult to accomplish. On the other hand, when these blacks must be deposited on a metallic sheet, as is the case for the thermopile and the bolometer, the condition that the electrical vector of the radiation be nearly zero at the surface of this sheet entails a corresponding decrease in the effective absorptivity of the black deposits for increasing wavelengths.

This difficulty does not exist in the pneumatic radiation detector, which operates on the old principle of the gas thermometer and comes within a half order of magnitude of the ultimate sensitivity permitted by (2).

The essential element of this detector is the radiation absorber which is placed in the center of a gas chamber. This radiation absorber consists of a broadbanded radio antenna in the form of a metallic sheet. The resistance of any square of this sheet matches approximately the parallel connected resistance of free space on both sides of it. This resistance of space, a full-fledged constant of nature, just like the speed of light, the charge of the electron, or Planck's constant, has the value $4\pi.30 = 377$ ohms, and its value is implied in the second Maxwell equation. Therefore, a metallic sheet of 188.5 ohms plays a role quite similar to that of a lumped resistance $R_0/2$ placed across a transmission line with a surge impedance R_0; one half of the energy reaching it will be absorbed, while a quarter each will be transmitted and reflected. (When this half resistance of free space is multiplied by the square of the electronic change, the quantity obtained is an action numerically equal to Planck's constant divided by 137, the famous number which, to this day, has been a major challenge to the theoretical physicists. It is noteworthy that the speed of light is *not* involved in this important relationship.)

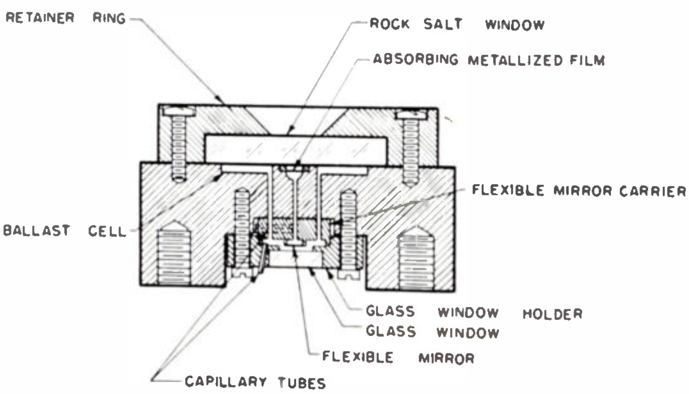

Fig. 1—Receiving head of pneumatic infrared detector.

Fig. 1 illustrates the principal elements of the pneumatic detector. The broadbanded radio antenna described above is the metallized film in the center of the small gas cell. It is heated by incoming radiation; the

[1] The concept of coherence or incoherence of electromagnetic radiation is utilized qualitatively only in this article, but mention should be made of the quantitative treatment of coherence which Gabor gives in his discussion, "Communication theory and physics," *Phil. Mag.*, vol. 41, pp. 1161–87; November, 1950, in which he utilizes the number of quanta in a unit cell of the frequency-time space as a measure of coherence.

gas of the cell becomes heated, expands, and this expansion of the gas deflects the small flexible mirror, which has a surface tension of the order of that of a water bubble. For the detection of mm radio waves the large rocksalt window is replaced by a 0.1-mm fused quartz window, which permits a waveguide to approach the detecting antenna, encountering only a minimum of radiation leakage.

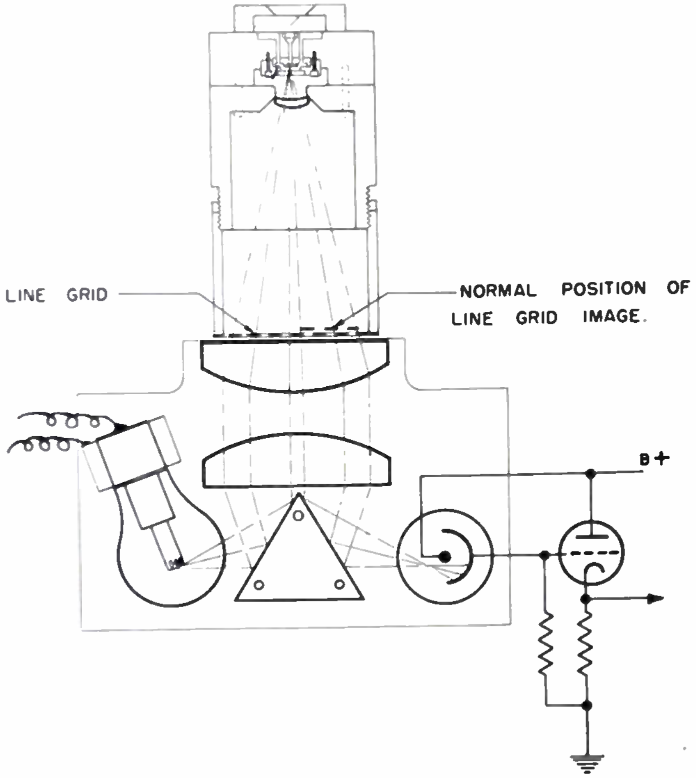

Fig. 2—General assembly of pneumatic infrared detector.

Fig. 2 illustrates the Schlieren-like optical system by means of which the deflections of the flexible mirror are converted into variations of light on a photocell. This optical detecting system permits the delivery of considerably more energy at the photocell output than is needed to cause the deflection of the flexible mirror, and therefore constitutes a preamplifier free of the flicker of thermionic tubes. Thus, the noise present in the photocell output represents the brownian motion of the flexible mirror, and can be verified to nearly disappear when the grid is displaced to a position of insensitivity, the residual noise being the shot noise of the photocell.

Because of the tight thermopneumatic coupling of this flexible mirror with the broadbanded antenna in the gas cell, a good part of the brownian noise of the flexible mirror reflects the fluctuations of the radiative interchange between this antenna and the background to which it is exposed. This is what permits this detector to approach within one-half order of magnitude the ultimate limit of broadbanded detection. In practice, it is used in connection with the so-called radiation chopping method, and the ac signals obtained are amplified, rectified, filtered, and recorded.[2]

While this detector was developed for infrared applications, the use of a broadbanded antenna suggested that it would be equally sensitive to mm radio waves. This possibility was verified for wavelengths up to 3 mm by Townes,[3] who has realized with it a sensitivity of 10^{-11} watts for his molecular absorption studies with the $1\frac{1}{2}$-mm wavelength harmonics of a pulsed magnetron. Such weak powers cause deflections of the flexible mirror which are of the order of 20 trillionths of an inch.

Since Townes' original experiments at 1.5 mm, it has been found that these harmonics of a pulsed magnetron,[4] and even higher harmonics down to 1.1-mm wavelength,[5] can be detected with more sensitivity with a silicon detector, because advantage can be taken of the small duty factor of the magnetron in order to gate the detector during the emitting period only, and thus obtain a 30- to 33-db lower noise than for full-time sensitivity. For the detection of yet shorter radio waves, or of considerably longer trains of pulsed mm waves, thermal detection will probably remain a useful method while awaiting improvements in electrical detection.

The detector just described constitutes, therefore, a kind of infrared-radio link which the microwave spectroscopist can utilize when he works at the short wavelength frontier of the radio spectrum, where he can generate coherent or narrowbanded (pulsed) radiation, but cannot detect it electrically.

Were it possible to effect the complete thermal insulation of a tuned dipole, or of an array of tuned dipoles, so as to restrict the energy interchange between these dipoles and the radiative background to the spectral region to which they are tuned, and were some means available to measure sensitively the temperature of these dipoles, the arrangement so postulated would constitute a thermal detector *not* limited by the fluctuations of (2), but limited by the smaller fluctuations of energy interchange within the tuned spectral range. Unfortunately, considerable physical difficulties attend the realization of such tuned circuits, because of the high-frequency surface resistivity of available conductors, which does not vanish even at superconductive temperatures. Fine-wire bolometers with a linear resistance of the order of the impedance of space per wavelength might permit escape from the limitations of broadbanded detectors given by (2), but when it is considered that these wires are poorly coupled with the impedance of space on account of their inductance, that even if

[2] M. J. E. Golay, "Theoretical considerations in heat and infrared detection," *Rev. Sci. Inst.*, vol. 18, p. 347; 1947; "A pneumatic infrared detector," *ibid.*, p. 357; "The theoretical and practical sensitivity of the pneumatic infrared detector," *ibid.*, vol. 20, p. 816; 1949.

[3] J. H. N. Loubser and C. H. Townes, "Spectroscopy between 1.5 and 2 mm wavelength using magnetron harmonics," *Phys. Rev.*, vol. 76, p. 178; 1949.

[4] J. H. N. Loubser and J. A. Klein, "Absorption of mm waves in nd.," *Phys. Rev.*, vol. 78, p. 348; 1950.

[5] J. A. Klein, J. H. N. Loubser, A. H. Nethercot, and C. H. Townes, "Magnetron harmonics at millimeter wavelengths," *Rev. Sci. Instr.*, vol. 23, p. 78; 1952.

placed in a good vacuum they have a thermal conductive path to ground through the terminating leads, and that bolometers have an inherently poor detecting sensitivity,[6] little appears to be gained in this direction.

Let us now examine what the infrared spectroscopist can accomplish with his over-all instrumentations, either with his classical instrumentation, or by means of recent developments and possible future developments. It will be clear, of course, that there is no instrumental long wavelength limit to the infrared technique, in the sense that there is an instrumental short wavelength limit to the radio technique. There is no point in working only with the infrared technique where the radio technique can be made to work.

It should be said at the outset that the comparison made before between the orders of magnitude involved in the generation and detection of radiation by the radio and infrared methods was slightly unfair to the infrared methods. The infrared spectroscopist has, in the past, done better than might be inferred from the foregoing, and there is room for further improvements. This is because the strength of the radio technique implies a weakness which the infrared technique does not have. Coherent generation requires that the essentially active elements of radio sources and detectors be not larger than, for example, a half wavelength, and also that they be monochromatic. That is, from an optical viewpoint, they must be point sources of a single-line spectrum and point detectors sensitive within a spectral band made as narrow as desired. As this limitation does not apply to infrared sources and detectors, the infrared spectroscopist can improve his lot by a few precious orders of magnitude, and he can do this in three steps, the benefits of which are, fortunately, cumulative.

The first step, which was taken almost intuitively at the birth of spectroscopy, consists of building spectrographs with elongated entrance slits instead of pinhole entrances, and the use of elongated entrance and exit slits in infrared monochromators followed as a matter of course. Thus, the radiative outputs of infrared monochromators are some two precious orders of magnitude higher than if pinholes were utilized, and this has permitted realizing up to the 10-micron wavelength the resolving powers inherent in the optical elements of these instruments, which is determined by the difference of path of the extreme rays in the collimated bundle.

The second step permitted the infrared spectroscopist consists in replacing both the single entrance and exit slits of his monochromator by two arrays of n slits each. With proper slit spacing, radiation of a specified wavelength—controlled by the position of the optical elements—will go in and out pairs of corresponding entrance and exit slits which have the same ordinal number in their respective arrays. Without any further precautions, the n-fold radiative output thus obtained for the specified wavelength range would be contaminated by the radiation of other wavelengths passing in and out noncorresponding slits; but this difficulty can be resolved by modulating the apertures of all the slits in accordance with functions of time which have orthogonal properties. This modulation can be effected by means of rotating discs with concentric slots of varying amplitude, and, in fact, a suitably selected portion of these slots can constitute the slits themselves.

With the arrangement just described, the radiation within a specified narrow spectral range passed by the monochromator, and no other, will be characterized by a specific time modulating, so that a measure of this radiation can be obtained. It must be noted that the old concept of a monochromator and a detector as two separate entities is being replaced here by that of a monochromator-detector combination, in which the "monochromator" does not yield a narrowbanded output which is then measured, but by means of which the measure of a narrowbanded component is obtained.

The choice of orthogonal functions deserves some discussion. An obvious choice would be sinusoidal functions. Thus if the xth entrance and exit slits are sinusoidally modulated at the frequencies $f_1+x\Delta f$ and $f_2+x\Delta f$, with $f_2-f_1>n\Delta f$, the radiation of specified wavelength passed by all corresponding pairs of slits with the same ordinal number, and no other radiation, will be characterized by a sinusoidal modulation of frequency f_2-f_1, and the problem of measuring the corresponding signal components in the detector output presents no difficulty.

Furthermore, every other spectral range will be likewise uniquely characterized by a sinusoidal modulation at the frequency $f_2-f_1+k\Delta f$, and individual selection and simultaneous measurement of these various radiations could be made by means of as many selective circuits. This simultaneous measure of many spectral elements constitutes the third step which is permitted an infrared spectroscopist, and just as the first two steps were permitted by the spatial extension of infrared sources, this step is permitted by spectral extension of these sources.

This third step still awaits realization, but by foregoing it, a wider choice of orthogonal functions becomes available for the second step. Thus, convenient two-value (1 and 0, open- and closed-slit) functions can be selected, and the experimental work done with slits modulated in accordance with such functions has demonstrated the general practicality of the system.[7] Mention can also be made of spectrometric systems with fixed and extended entrance- and exit-slit patterns, in which the measure of a specified narrowbanded range is obtained as the differential measure of two relatively large and broadbanded bundles of radiation passed by the exit-slit pattern of the monochromator.[8,9]

[6] Ref. 2, *loc. cit.*, pp. 351–352.

[7] M. J. E. Golay, "Multislit spectrometry," *Jour. Opt. Soc. Amer.*, vol. 39, p. 437; 1949.

[8] J. Strong, "Experimental infrared spectroscopy," *Physics Today*, vol. 4, p. 14; 1951.

[9] M. J. E. Golay, "Static multislit spectrometry," *Jour. Opt. Soc. Amer.*, vol. 41, p. 468; 1951.

Fig. 3 has been drawn to indicate the order of magnitude of increase in resolving power which the application of the second and third steps can be expected to yield. The conditions postulated for this were relatively conservative. First-order spectra from 2,000-line gratings were postulated for the whole range, which is equivalent to the postulation of a maximum resolving power of 2,000, obtained with gratings ranging in length from 2 cm at 10-micron wavelength to 1 m at ½-mm wavelength (resolving powers as high as 10,000 at 10 microns, and obtained with larger gratings, were reported at the last Ohio State Symposium on Molecular Structure by Peters, of Michigan University). If source temperatures higher than the postulated 300°C temperature are utilized, as well as larger time constants, and if the requirement of a S/N ratio of one thousand is relaxed, the use of the fairly convenient second step alone may go far in pushing to the mm region the wavelength at which the resolving power of the external optics is nearly realized, thus bridging what can be bridged of the spectroscopic gap between the infrared and the radio waves.

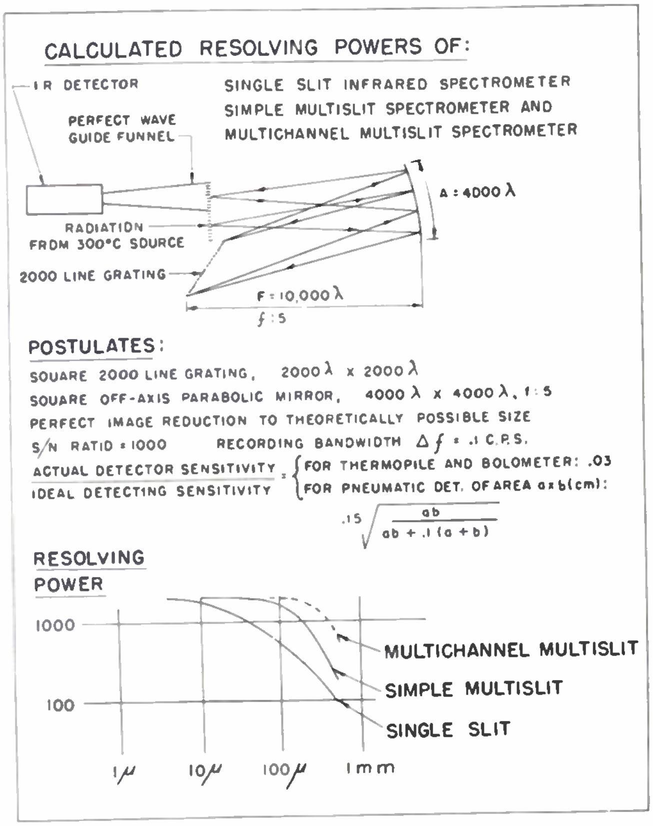

Fig. 3—Resolving power of infrared gratings.

The handicaps under which the infrared spectroscopist must labor are illustrated by Fig. 4, in which the curves of Fig. 3 have been redrawn in the context of the resolving powers already obtained, or foreseeably obtainable in the two neighboring spectral regions.[10]

Conclusion

When compared to radio sources, infrared sources have an extraordinarily small power per unit frequency. Likewise, infrared detectors have an extraordinarily small sensitivity when compared to radio detectors.

Since coherent detection is impossible in the infrared region, little basic progress is likely towards more sensitive detectors. The resolving powers to be expected will remain a function of the path difference of extreme collimated rays in the external optics used.[11] Also, spatial and spectral extension of infrared sources can be utilized more fully to develop spectrometric instrumentation permitting a near realization of the theoretical resolving power of the optical elements used, up to the shortest radio waves.

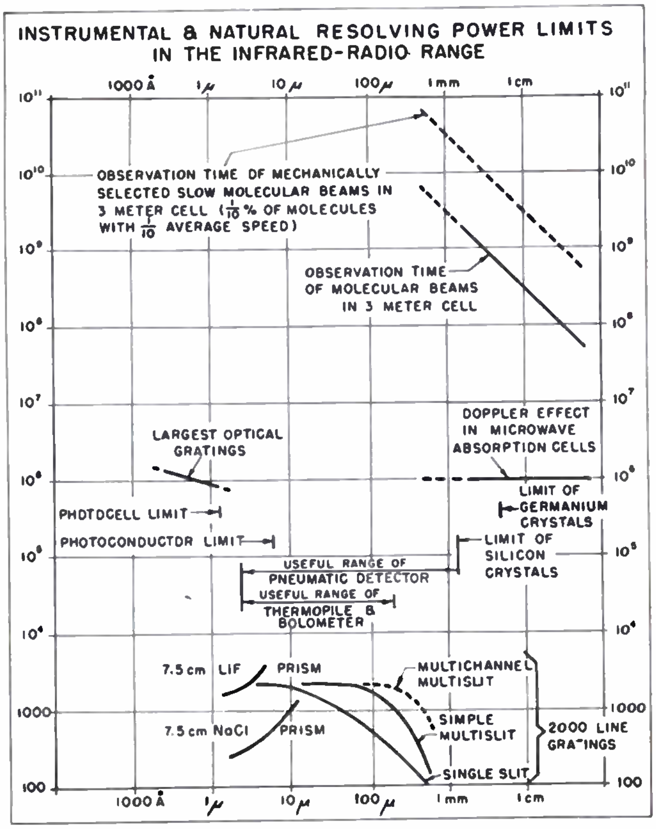

Fig. 4—Instrumental and natural resolving power limits in the infrared-radio range.

[10] Since the drawing of the dotted curve in the upper right corner, which was meant as an eventual target for the microwave spectroscopy instrumentalist, Dicke and Newell have successfully demonstrated a new method for sharpening the resonance curve of a microwave absorption cell without resorting to the molecular-beam technique. With their method, the neighborhood of the lower left end of this dotted line appears reachable in the near future. (G. Newell, Jr. and R. H. Dicke, "A method for reducing the doppler breadth of microwave absorption lines," *Phys. Rev.*, vol. 83, p. 1064; August 15, 1951.)

[11] A few-fold increase in the resolving power of given dispersing elements can be obtained by an application of the multipass method, in which partially dispersed spectra are returned to these dispersing elements for further passes. Walsh has reported recently (*Jour. Opt. Soc. Am.*, vol. 42, p. 496; 1952) his successful application of this method. Multislit multipass spectrometers are discussed by the present author in a letter to the editor of the *Jour. Opt. Soc. Am.*, to be published shortly.

Technique of Trustworthy Valves*

E. G. ROWE†

Summary—**The introduction discusses the reasons for requiring trustworthy valves, and a definition of reliability is given. Following a survey of world progress in this subject, a section is devoted to the design problems involved, with illustrations of the equipment used. Next follows a description of manufacturing procedures and the different techniques necessary, together with information on testing methods. In conclusion, the author impresses the need for greater co-operation from the user.**

Introduction

TRUSTWORTHY VALVES are frequently mechanical redesigns of commercial types intended to provide improved performance under vibration and shock. They are manufactured and tested with such precision as to assure reliable operation for longer lives than would be obtained with their commercial counterparts.

Causes of failures of commercial types are examined and redesigns are made to avoid existing weaknesses. Manufacturing processes are refined to produce more uniform quality and acceptance tolerances are tightened as a continuing check on fabricating accuracy and on the purity of raw materials. The need for long, continuous manufacturing runs of given types of trustworthy valves to achieve maximum uniformity of product is stressed.

The manufacture of valves for commercial radio sets is controlled primarily by the need to produce them at a minimum cost, and despite this limitation it has been possible also to maintain a low failure rate during the normal life of equipment.

However, the rapid increase in the number of television sets with their average complement of 20 valves against radio-set complement of 5 valves has revealed that the failure risk can be embarrassing and, consequently, manufacturers have been steadily improving their quality within the close confines set by economics.

In addition, the fighting services, after the natural lapse of time consequent upon cessation of hostilities, have become increasingly aware of the limitations set on their fighting strategy due to the unreliability of their electronic equipment, and up to the present the valve has come in for most attention on this account.

I. Definition of Reliability

An engineer interested in the subject of "reliable" valves very quickly realizes that any attempt to give an accurate definition of reliability depends very much on the application for which the valve is intended. Is one to refer to a 1,000-hour vibrational condition, to a non-vibrational life of 50,000 hours or more, to a high shock reliability for a very short period, to a long shelf life? One of the best general definitions is that a reliable valve is characterized by a very high probability that it will operate normally when taken from stock and installed in equipment for which it was intended, and by an extremely low probability that it will fail during subsequent operation in that equipment for some definite period of time.

In many cases, the work involved can progress along similar lines to achieve any or all of the above requirements, but for the purposes of this paper, the discussions will relate only to valves to be manufactured in large quantities and required to operate reliably for not more than one or two thousand hours under conditions where extremes of ambient temperature may be encountered either in storage or in operation, and where the valve and equipment may be subject to considerable mechanical shock or vibration. To be still more specific, the immediate objective is to achieve valve types having characteristics corresponding to existing designs but with a failure risk of the order of 1 per cent in 1,000 hours. This achievement is planned to be obtained by refinements of design, material, and manufacturing methods.

II. History of World Progress

Aeronautical Radio, Inc. were the first to sponsor a project to improve the life of the valves that they used. By keeping track of the valves in aircraft equipment, they were able to analyze the failures and the information was used to improve the design of particular valve types.

In parallel with this, the United States Armed Forces have been working along similar lines but more slowly due to the fact that the range and complexity of their electronic equipment is so much greater.

The major programs have been as follows: (a) Bureau of Ships W series, which are strengthened counterparts of existing types and are primarily intended to withstand gun-fire shock; (b) Aeronautical Radio, Inc., series of miniature types; (c) Radio Corporation of America special red tubes; (d) Sylvania and Raytheon subminiature tubes.

Work in Great Britain has been centered on types in the Services Preferred List and mainly on miniatures and subminiatures. Standard Telephones and Cables, Ltd. has concentrated on the miniatures and novals, as will be seen in Table I, but it is evident that the production of the whole range of trustworthy types must still take some substantial time, and careful consideration has therefore been given to interim measures. This has

* Decimal classification: R330. Original manuscript received by the Institute, May 15, 1952. This is one of a class of papers published through arrangements with certain other journals. It appeared in expanded form in the November, 1951 issue of *Journal of the British Institution of Radio Engineers*.

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consisted of selective treatment of standard types and, contrary to the experience in the United States, it has been possible to devise vibrational and aging methods that will select a markedly improved product from ordinary commercially manufactured stock. This has been proven on a number of miniatures and novals and on three GT types, and is based on the selection of many tens of thousands of valves. Therefore, until the full trustworthy program is completed, it is considered that customers will get a worthwhile improvement by special selection methods.

TABLE I

Trustworthy Type Valves Available from Standard Telephones and Cables, Ltd., March, 1951

Trustworthy type	*Commercial equivalent*	*Description*
6180	6SN7GT	Double triode, $\mu=20$
5749	6BA6	Vari-mu radio-frequency pentode
5750	6BE6	Heptode frequency changer
6042	25SN7GT	Double triode
6057	12AX7	Double triode
6058	6AL5	Double diode, separate cathodes
6059	6BR7	Low-noise amplifier, pentode
6060	12AT7	High slope double triode
6061	6BW6	Output pentode, 6V6GT characteristics
6062	5763	Radio-frequency amplifier, pentode
6063	6X4	Full-wave rectifier
6064	8D3	High-slope radio-frequency pentode gm = 7.5mA/V at Ia = 10mA
6065	9D6	Vari-mu radio-frequency pentode
6066	6AT6	Double-diode triode, $\mu=70$
6067	12AU7	Double triode, $\mu=20$

III. Design Considerations

A valve is basically a mechanical structure that has to possess electrical properties that must be maintained throughout its life. Because it is built up from metal, mica, and glass, it is subject to failures common to all structures made of these materials, such as breakage, distortion, loosening of component parts, and the like. In addition, the mechanical weaknesses directly affect electrical and chemical properties.

Complete analysis of many large groups of standard commercial valves returned as failures from operational equipment has shown that the faults occurring are as follows: (a) electrical failures such as noise, instability; (b) mechanical failures of the assembly giving short-circuits and open-circuited elements; (c) mechanical failures of the heater structure; (d) glass faults.

With the exception of group (a), these failures agree quite well with the anticipated failures that can be concluded from a careful survey of our own static-life-test results. Group (a) failures are caused by vibration and are a preliminary to Group (b), being due to a loosening of the structure permitting mechanical movement of the component parts. We can, therefore, subdivide into three main groups: (1) mechanical faults, often aggravated by vibration; (2) heater faults; (3) glass faults.

Improvements to Mechanical Design

It was fairly obvious that redesign efforts had to be directed towards shorter and more rigid structures that would be more stable under vibration.

The equipment shown in Fig. 1, which was developed by Moss of Electronic Tubes Ltd., has been used for studying the mechanical properties of various valve designs. In this, the valve is mounted on a moving-coil vibrator and the alternating-current output of the valve when operated in class-A conditions is examined by means of an oscilloscope connected across an

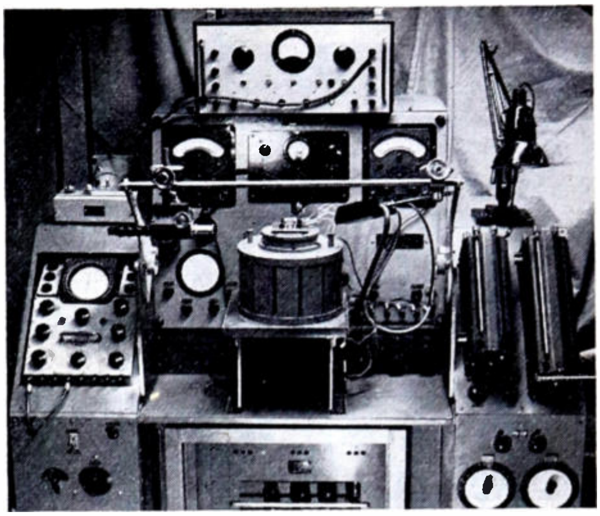

Fig. 1—Apparatus with which vibration tests are made.

anode load resistor. Means are provided for applying a small calibrating signal to the grid of the valve under test so that the noise can be equated to an equivalent voltage on the signal grid, this interpretation of the output overcoming variations in valve gain.

The frequency of vibration of the valve is raised slowly from 15 to 3,000 cps with the amplitude of vibration adjusted to give an equivalent acceleration of $2\frac{1}{2}$ grams.

Photographic records of the noise produced over a frequency range from 200 to 3,000 cps by a valve at three stages in its development are shown in Fig. 2. These diagrams have two main features: (1) sharp resonance peaks, usually in the high-frequency region, which represent true resonance and are usually associated with grid and cathode vibrations; (2) a general noise level in the low-frequency region, which is associated with looseness of the structure and is the more serious fault as this causes noise and instability leading ultimately to short circuits.

A further feature of this testing machine is the provision of a telescope and stroboscopic lamp enabling a vibrating component to be studied when it is in resonance.

Modifications to the structural design of valves in the general direction of improved rigidity include the following main features: (a) tight mica holes for the grids and the use of double micas of optimum thickness; (b) locking straps in the micas to hold the grids; (c) locking the bottom insulator to the stem in as many positions as possible; (d) locking the anodes into the micas by welding straps across the anode lugs rather than bending them down or twisting them; (e) general changes giving greater strength to the anodes and improved location and fixing methods; (f) shortening of the valve mount by using increased diameter cathodes; (g) minimizing the number of welds.

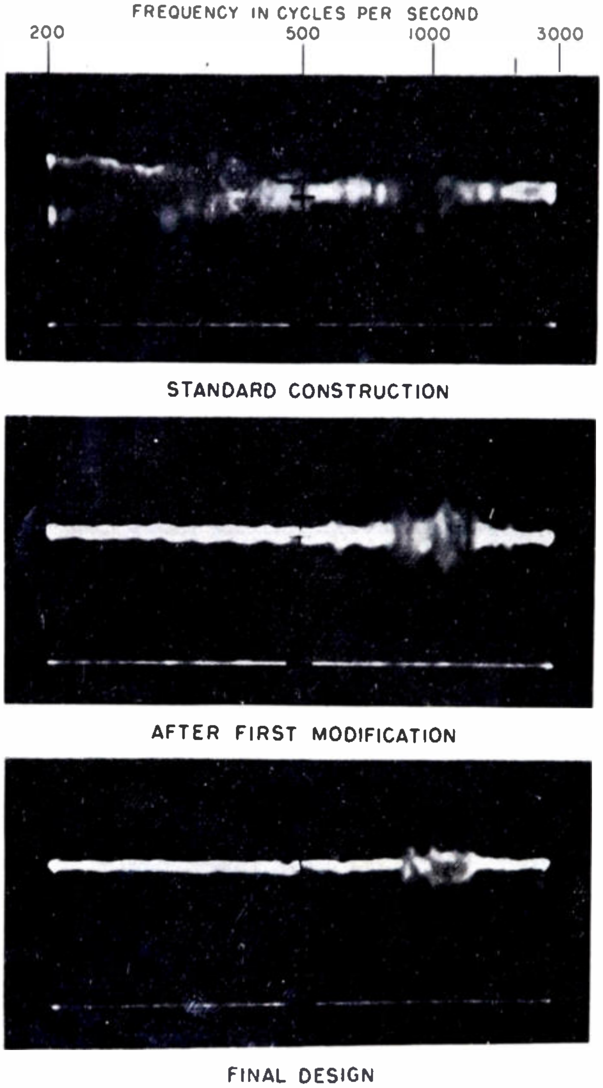

Fig. 2—Photographic record of noise output over frequency range from 200 to 3,000 cps. (a) Standard construction; (b) After first modification; (c) Final design.

By applying many of the modifications described, it is possible to produce a resonance diagram as shown in Fig. 2 (c) having a very low general level of noise but still showing a few sharp peaks in the upper-frequency region (above 1,000 cycles). These latter resonances are, in general, much lower in height than those of the original valve but usually occur at the same frequencies, indicating that they are fundamental resonances of components and can only be removed by a complete change of design technique. As the immediate aim is to produce valves of similar electrical properties (including capacitances) to an existing range, complete redesign is generally out of the question.

Other machines that are of value in this design work are described later in Section VI.

There is much additional work to be carried out. Not only is there scope for improvements on grid making and for closer controls on internal bulb diameters, but there are major projects involved in the studies of cathode base metals and oxide coatings.

Improvement to Heaters

Heater failures resulting from open circuits and short circuits have been due to three causes: (a) excessive core temperature; (b) movement of the heater inside the cathode under vibration conditions which damages the heater coating at the cathode ends; (c) incipient fracturing of the core wire at any sharp bends.

It was established that, while satisfactory for normal receiving valves, it would be advisable to lower the operating temperature by the use of thicker core wires for the trustworthy valves. These thicker core wires necessitated the use of a longer length of heater wire, which meant extra loops giving a tighter fit into the cathode. The improvement from this change was not as much as was expected, as the heaters still had sharp bends that led to breakage, and therefore the reverse helical heater has now been adopted permitting the use of a large wire size having no sharp bends to cause incipient failure.

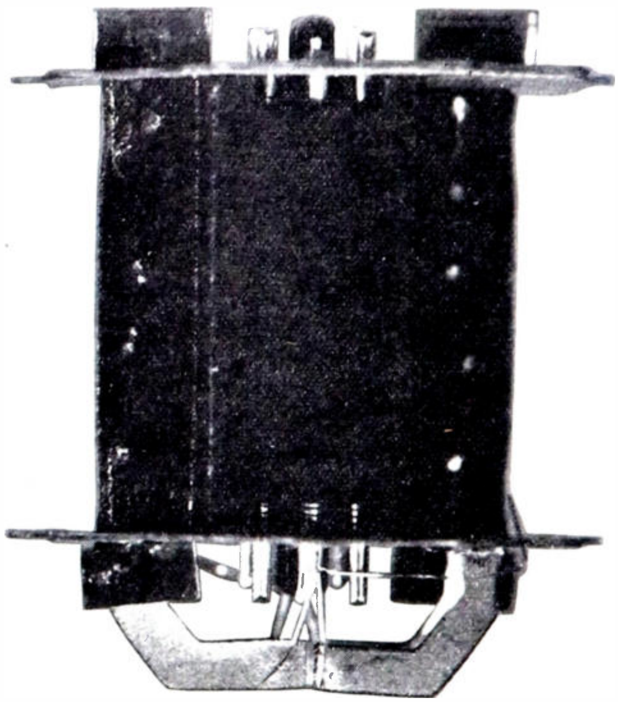

Fig. 3—Desirable heater mounting.

This heater design is now being used for most of the trustworthy valves. It has been found that the core temperature can be kept down to a safe figure and that the heater is flexible and has no sharp bends. In addition, it fits the round cathode sleeves exceedingly well and is easy to insert into the cathode during assembly without damage to itself.

A further design feature was the use of heater bars that lock into the bottom mica and project below the cathode. These can be seen in Fig. 3. The heater can then be inserted into the subassembly, welded in place, and inspected before the subassembly is mounted on the glass base. Furthermore, this arrangement allows the stem wires to be welded to the heater bars near the periphery of the mica, and thus gives valuable support to the valve assembly.

Examples of normal and trustworthy valve designs are shown in Figs. 4 and 5.

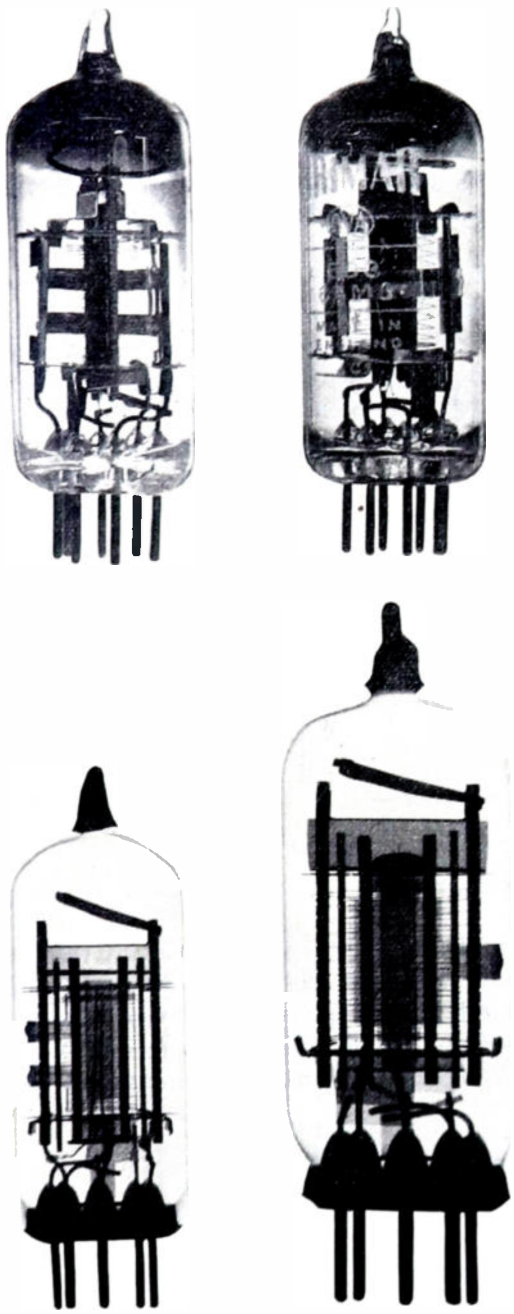

Fig. 4—(a) and (b) External and X-ray views of the 8D3; (c) and (d) external and X-ray views of its counterpart, the trustworthy 6064, illustrate use of locked grids, strengthened anodes, improved heaters, and more rigid mountings.

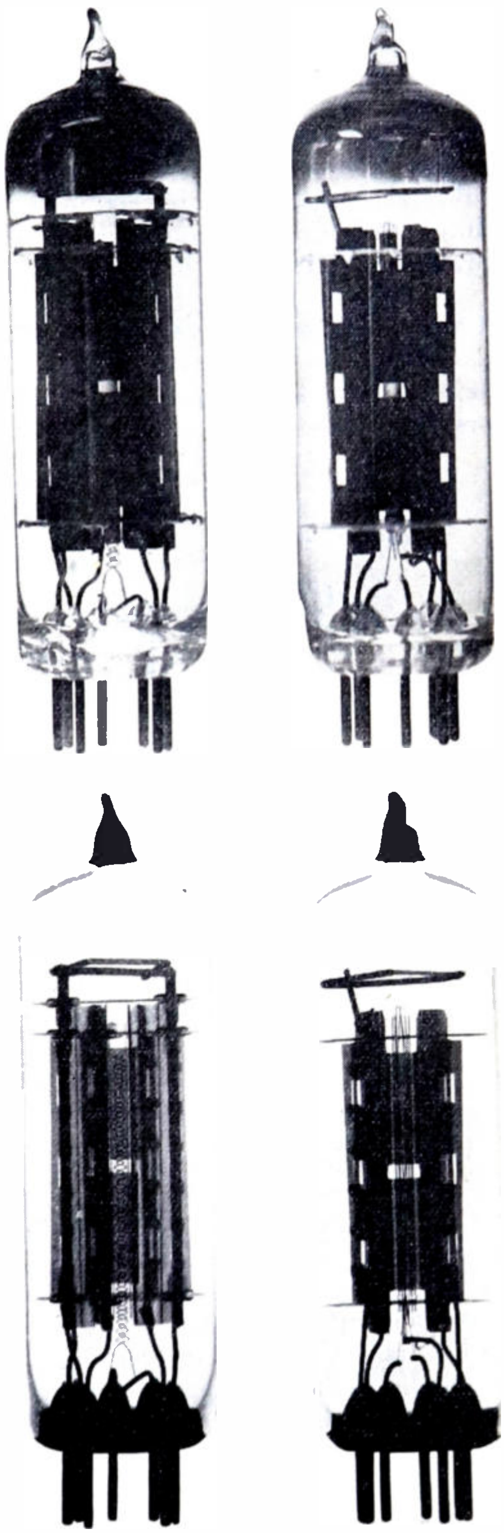

Fig. 5—(a) and (b) External and X-ray views of the 6X4; (c) and (d) external and X-ray views of the trustworthy 6063 show the inclusion in the latter of getter safety mica additional supports, improved heaters, and more rigid mountings.

IV. Glass Problems

As an envelope material, glass has the advantages of transparency, chemical inertness, electrical insulation, hardness, and the ability to be welded by flame heating; but it is a brittle rather than a tough material, and therefore requires a considerable number of refinements in technique to ensure freedom from cracking in service.

Glass is strong in compression and relatively weak in tension, and therefore small surface scratches and any crevices are likely to become cracks under quite small stresses. In addition, residual stresses can lead to ultimate failure of glass structures. The two best known are "mismatch stress" arising from the differential contraction of two glasses welded together, and "thermal strain" due to the differential contraction of glass near a weld, in relation to the glass farther away from the weld.

It will also be realized that in addition to glass-to-glass seals there are the problems of glass-to-metal seals which are an inevitable consequence of leading metal conductors through the glass.

Field failures are best studied by considering glass structures in three groups: the pinch types, the loctal types, and the miniature types. Their principal weaknesses are shown in Fig. 6. The relative occurrences in the field are best shown in Table II.

During the fabrication of a valve, a continuous watch has to be maintained in order to limit the shrinkage both internally in the factory and subsequently in the field. The tests employed are a combination of routine laboratory and practical inspection tests on the floor.

For example, in the laboratory, photo-elastic techniques are used to detect differences in thermal expansions of glasses and also to measure stresses in the bulb walls, while the factory uses as its main control the well-known B test, in which the pins are distorted by the insertion of a conical steel plug, after which the whole is plunged in boiling water. The glasswork must not crack.

TABLE II

Approximate Relative Occurrences of Glass Faults in Service

Fault	*Possible reason*	*Percentage of group faults*		
		Pinch	*Loc-tal*	*Mini-ature*
Cracked base	Thermal strain or expansion mismatch	—	—	65
Cracked bulb	Rings or patches of strain	40	95	20
Crack from tip off	Crevice	40	—	
Cracked dome	Strain ring	10	—	—
Bad tip off	Crevice	—	—	10
Miscellaneous	Leaks, etc.	10	5	5

Expansion Mismatch

There is not usually much trouble in the field from expansion mismatch because the glasses are suitably chosen to avoid it. It is very important that the glass stem shall always be slightly compressed by the bulb and Table III and Fig. 7 show the results from combining different commercial glasses to make miniature valves. α is the coefficient of thermal expansion.

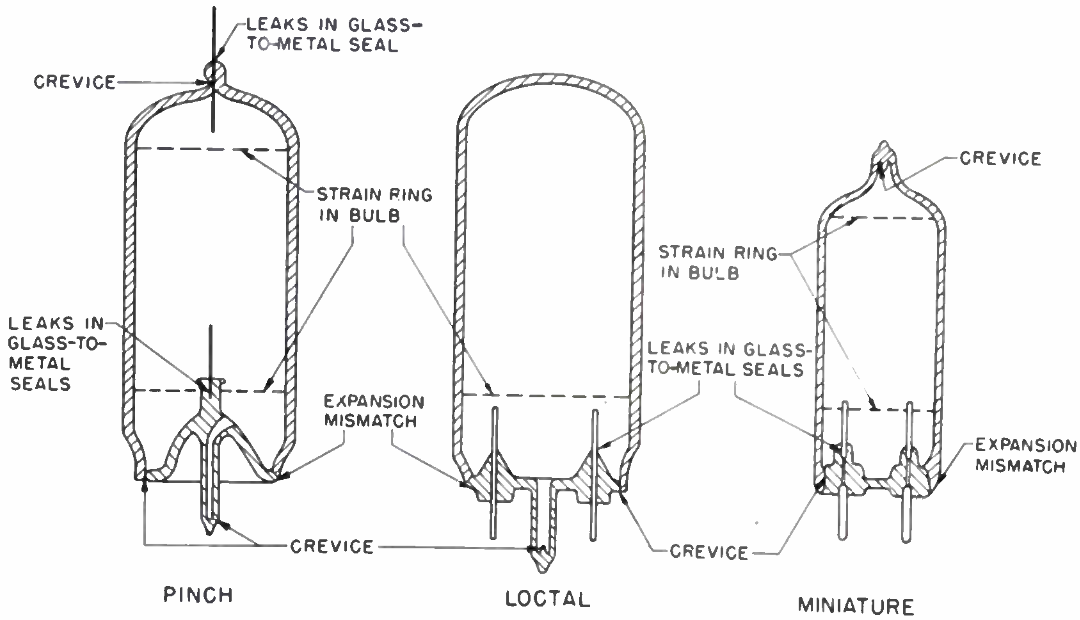

Fig. 6—Principal points of weakness in glasswork.

TABLE III

Effect of Expansion Differences between Bulb and Stem Glasses of Miniature Valves

Bulb, lime-soda glass	*Stem, lead glass*	
	Lead glass A $\alpha = 92$	*Lead glass B* $\alpha = 90$
Glass C $\alpha = 98$	Satisfactory	Cracked bulb
Glass D $\alpha = 96$	Satisfactory	Satisfactory
Glass E $\alpha = 92$	Cracked stem	Satisfactory

To ensure satisfactory matching, it is important to maintain routine laboratory tests on sealing stress and thermal expansion on all incoming supplies of glass.

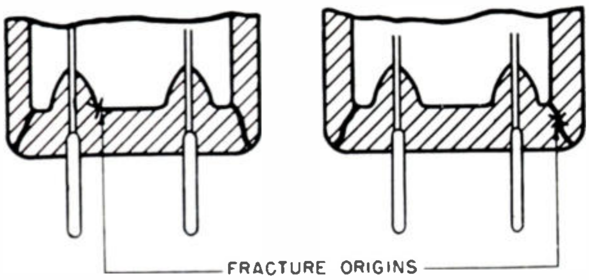

Fig. 7—Fracture origins in miniature valves resulting from differences in bulb and stem glasses. Cracked stem at left occurs when both glasses have same thermal expansion; cracked bulb at right is for bulb glass of $\alpha = 98$ and stem of $\alpha = 90$ where the differences are too great.

Thermal Strains

To prevent failure due to thermal strain, it is essential to minimize the "strain rings" caused by the sharp thermal gradient due to hot joining of the glasses.

The distribution of stress in a strain ring is complex but, in general, can be regarded as a zone of high tension with the body of the glass protected by a thin layer of glass in compression at the surface. The problem is complicated because where the glass is likely to be subjected to tension forces during life, as is likely in valve sockets, strains are deliberately set up in the base of the miniature valve to increase its strength.

These stresses are controlled by strain-viewer observation against established standards.

Crevices

The term crevices is used to describe areas where the glasses join and can form acute angles between their surfaces. An expert operator is able to detect such internal crevices in most glasswork by careful observation of the external contour.

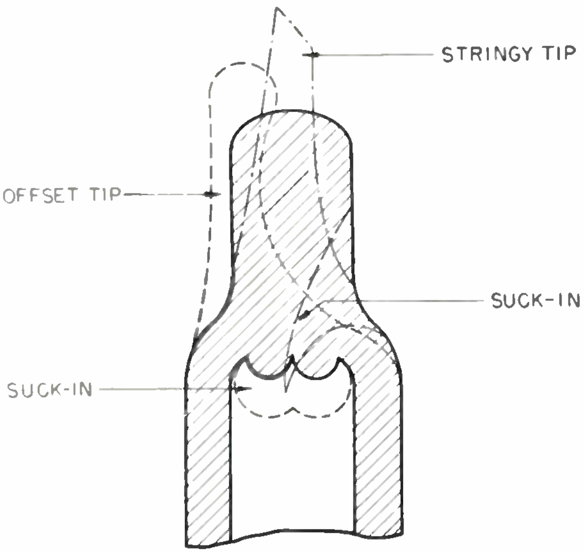

Fig. 8—Normal tip-off shown solid. Some common defects shown in broken lines.

Crevices in the stem-to-bulb seal can be detected by plunging the cold valve into boiling water, thereby starting cracks from any flaws on the inside surface.

In the case of "tip-off," it will be seen from Figs. 8 and 9 that inevitably there is a potential source of weakness at this point and additional assistance is given to the factory by large chart diagrams and definite external dimensional requirements.

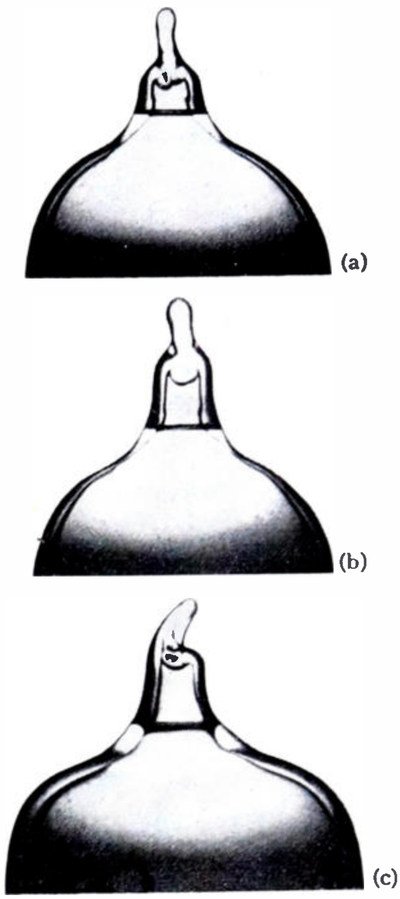

Fig. 9—(a) Normal tip. (b) Air sucked in. (c) Offset tip.

Leaks

Leaks usually occur because of faulty glass-to-metal seals. Vacuum tightness on these seals depends on a chemical reaction between a tightly adherent oxide layer on the metal and the glass melted on during the sealing operation. The stages of the reaction are usually accompanied by color changes in the seal, and thus one test of vacuum tightness is a visual observation of the seal color. As a further aid, the valve may be immersed in a mobile fluorescent liquid which is sucked into the leaky hole and made visible under ultraviolet light.

Fine leaks can only be detected by storage and characteristic deterioration, but it may be possible to accelerate this by storage in hydrogen at 100 pounds per square inch (7 kg per cm²).

Successful results depend on very careful control of the seal wire used, by tests such as microscopic examination for fissures, and glass-sealing tests for coefficient of expansion.

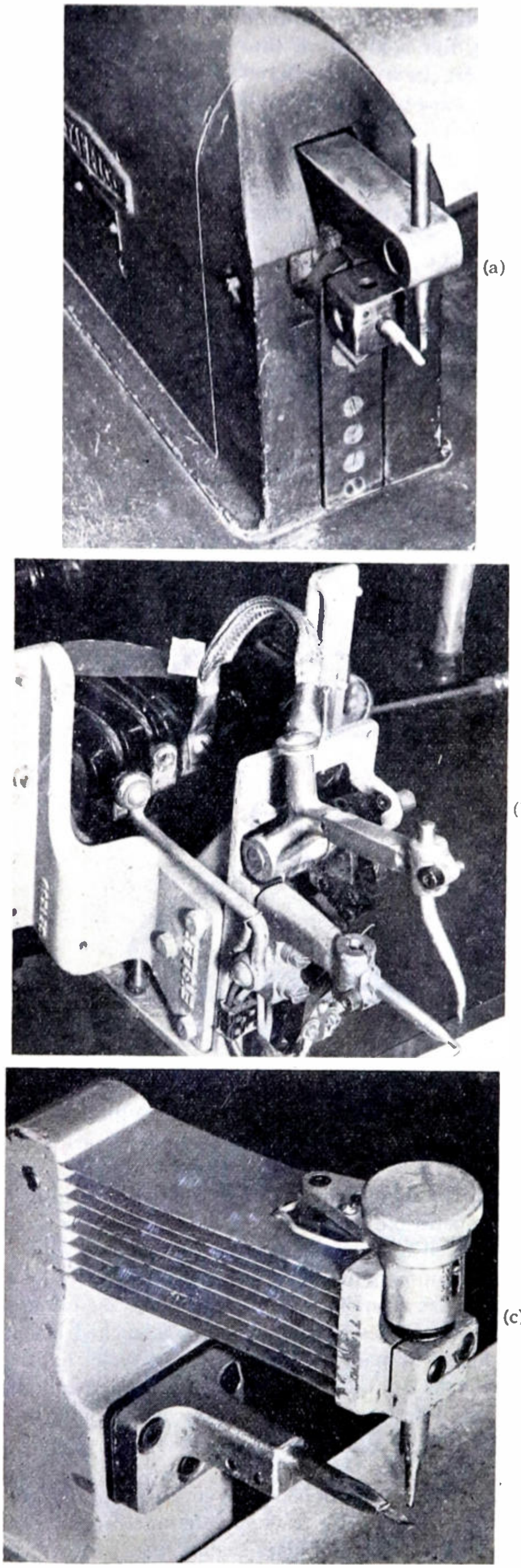

Fig. 10—Welder heads of three designs. (a) the Stanelco welder; (b) the Eisler welder: (c) the Slee welder.

V. Manufacture of Trustworthy Valves

The present state of the art is that the engineer can design reliability into a valve exclusively by calculation and special testing carried out on laboratory equipment. The major problems of reliability then resolve into the adequate control of materials and processes and of the inherent variability of the individual operator.

To meet the first requirements, the raw material standards are made stricter than is usual and all processing, particularly of coated cathodes and heaters, is more careful and thorough. All machinery used is subjected to routine maintenance controls of a higher requirement than normal.

On assembly operations, the most important variable is resistance welding which often relies on the skill of the operator to control the weld. A valve usually has about twenty welds and therefore a high proportion of potentially faulty valves can result from quite a low proportion of faulty welds. We have now designed a welding head and associated timing equipment to give a reliable and repeatable weld.

Fig. 10 shows the comparisons between various welders, Slee being our latest development. In addition, benefit has resulted from a close study of the operations involved whereby the job has been graded to a number of accurately preset welding machines, restricting the work variation demanded from any one unit.

The second variable is a more elusive one, described as "lint." This is a fibrous contaminant which enters the valve structure during the assembly stage and becomes carbonized during subsequent processing, causing leakage and noise in the finished valve.

This trouble can be mitigated only by observance of the utmost cleanliness. All components are kept covered and special trays are used between the cleaning operations and the assembly stage. Valves are assembled under a glass cover (see Fig. 11) which is slightly pressurized by a dry air stream so that the normal air current is away from the valve, and the assembly operators wear special overalls made of lint-free cloth. Finally, completed assemblies are placed in their bulbs immediately after inspection and the bulbed assemblies are kept in closed boxes. The valves are sealed in and exhausted with the minimum of delay.

It is of considerable importance that the inspection should be capable and thorough. As may be seen in Fig. 12, this work is done using binocular microscopes and is controlled by chart methods.

It will be appreciated that the uniformity of product that is desirable in a trustworthy valve can be achieved only by continuous and long production runs. It is only by being ruthless that the standard can be maintained, and therefore both the beginning and the end of such a run may be diverted for use as normal radio valves or else scrapped rather than run the risk of premature failures in the field.

VI. Testing of Trustworthy Valves

Adequate control has to be maintained by batch

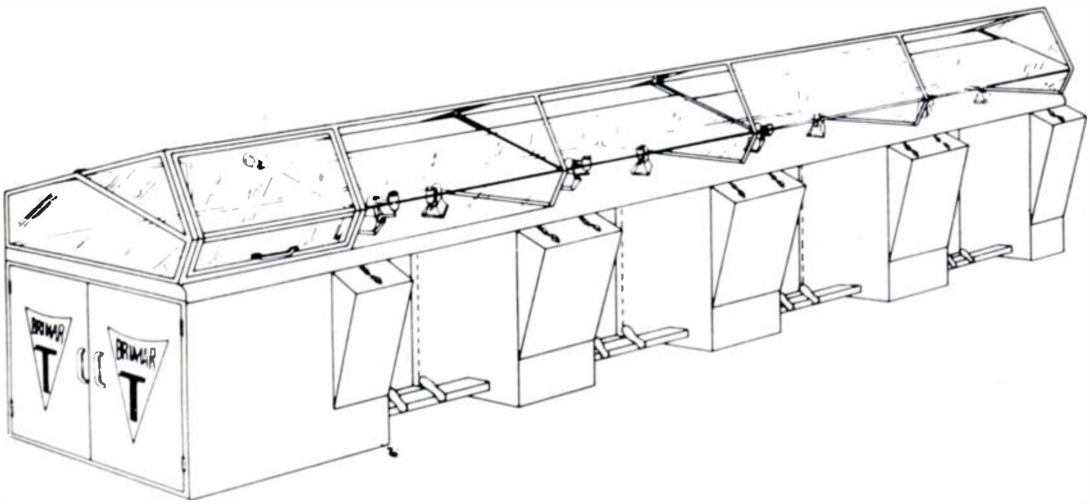

Fig. 11—Bench for assembly of trustworthy valves.

Fig. 12—Inspection position for trustworthy valves. Operator looks into microscope through a glass panel.

manufacture and by progressing valves accordingly through the test procedure. These tests consist of group-A tests, which are factory tests to ensure that the valves are uniform and line up with the design specification, and group-B tests, which are sampling tests to check on the level of quality.

Group-A Tests

These tests are mostly 100-per cent tests on which the basis of batch rejection will be affixed, say 5-per cent rejection from the batch.

Such testing starts at the completion of the assembly stage with a visual inspection of the mounts using a binocular microscope.

At the sealex section, special checks are carried out to ensure satisfactory glass quality of the completed valves. These consist of a visual inspection of the tip-off for quality and a thermal shock test to the completed valve which will show up strained bulbs and bases.

The valves then pass to the activation stage and tests are applied here to detect short circuits and any lint in the valve by applying 250 volts to each electrode in turn with all other electrodes grounded.

A heater-flash is carried out where a high voltage is applied instantaneously to each valve for a period of ten seconds. This serves to sort out possible heater failures.

Each valve, after activation, is subjected to a test for electrical characteristics, a short-circuit test, and a vibration test conducted at 50 cps at an amplitude of ± 0.020 inch. The vibration test is continued for a period of at least one minute during which the valve is operated under class-A conditions and the noise output across an anode load resistor is monitored. All valves are then given a 10-hour life run under class-A conditions, followed by a underheating test of cathode activity and a repeat electrical test.

Figures are recorded on a sample of twelve valves measured before and after the 10-hour run. These figures are compared to assess what changes in characteristic have taken place during this period.

Group-B Tests

These are quality-control tests made on samples taken from the batch at various stages in manufacture to determine that the original quality is being maintained, and they are, in most cases, destructive tests.

At the sealex operation, envelopes are tested for the presence of strain rings by a diamond scratch procedure which will cause spontaneous cracks if excessive strain is present.

After the valves have been given their 10-hour run and retested for electrical characteristics, sample batches of valves are taken and subjected to the following range of mechanical tests:

Resonance Test. The sample is checked for resonance on the design apparatus and points of resonance are recorded and compared in position and height with the standard laid down by the design engineer.

Shock Test. For the shock test, a sample valve is mounted rigidly on a moving platform that is subjected to impact shock from a falling hammer. The apparatus

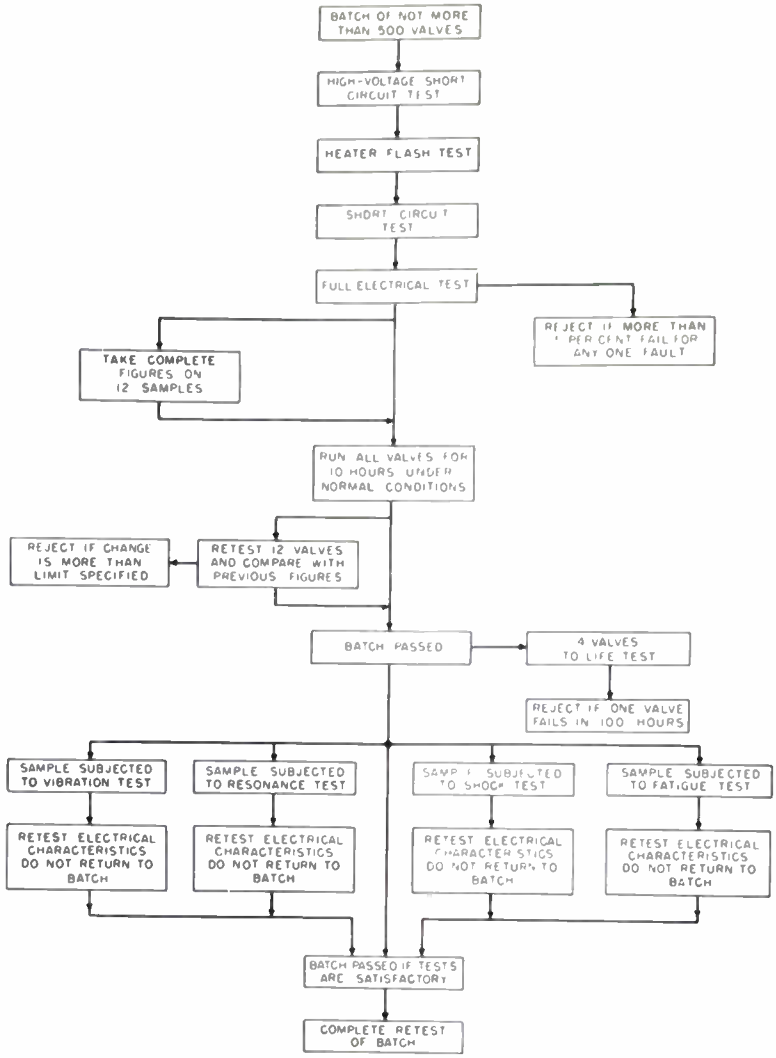

Fig. 13—Testing procedure for trustworthy valves.

is shown in Fig. 14. The valve is operated cold and is given five blows in each of the three planes of the valves, i.e., (a) horizontally with the major axis of the

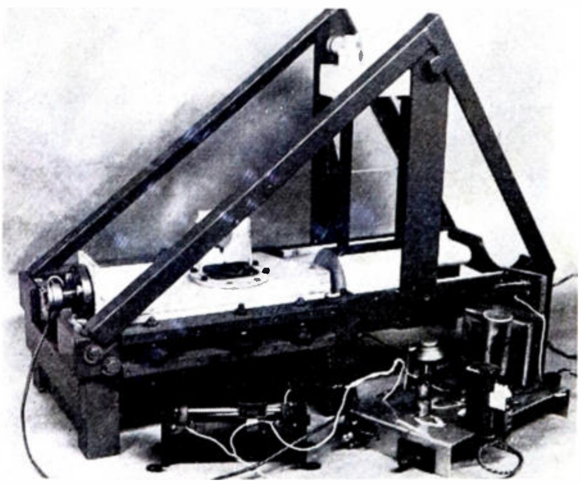

Fig. 14—Apparatus for impact shock testing.

valve perpendicular to the plane of vibration; (b) horizontally with the major axis of the valve parallel to the plane of vibration; (c) vertically.

In the case of the vertical position, the shock is applied from both ends of the valve. The magnitude of shock imparted to the valve during this test is of the

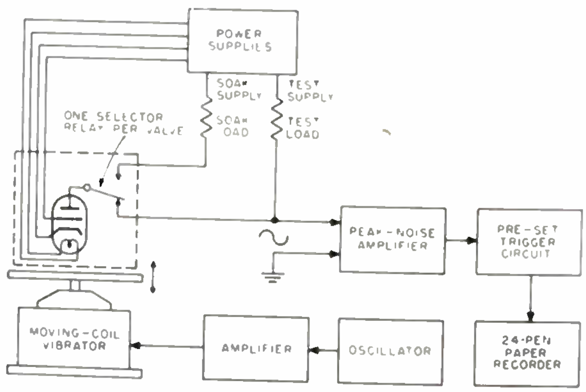

Fig. 15—Schematic of fatigue-test machine.

order of 1,000 grams and the valve is monitored for short circuits occurring during the shock. After shock tests, all valves are electrically retested.

Fatigue Test. The valves are vibrated in the three standard positions for prolonged times of the order of 90 hours each at various spot frequencies and ampli-

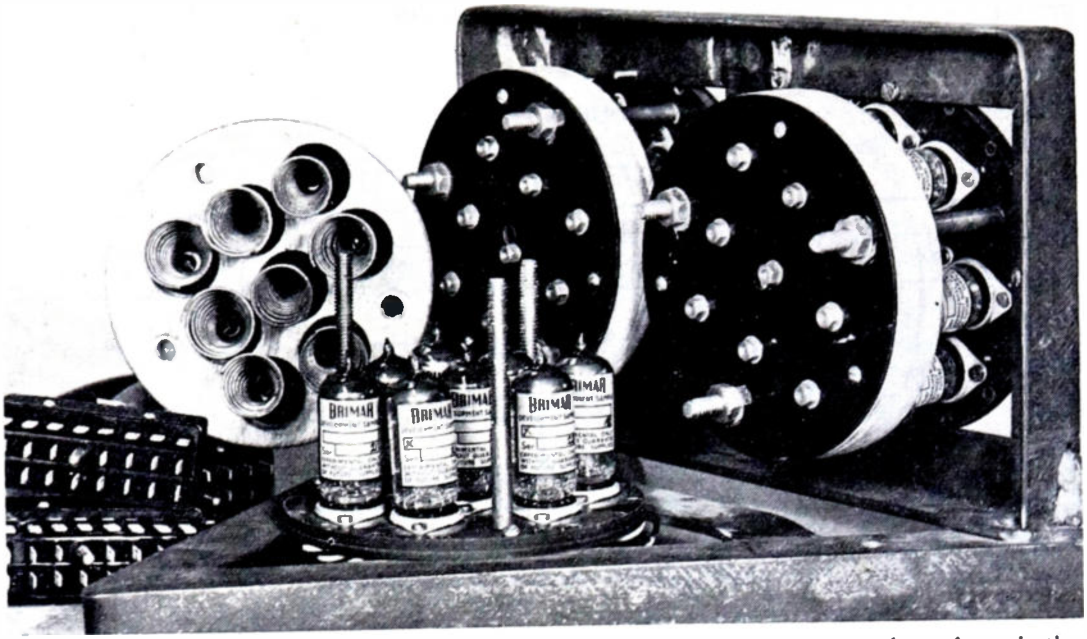

Fig. 16—A group of valves is mounted on each of the three disks that are fastened to a frame in the standard vibration positions. Three of these frames are shown mounted on a vibration table in Fig. 17.

tudes giving accelerations of the order of 2 to 3 grams. During the vibration period, the heaters are switched on and off at a 5-minute cycle and the valves are monitored in class-A operating conditions for excessive noise in the anode resistor. It is considered that the occurrence of noise in an anode resistor under vibration is indicative of a loosening of an electrode that will ultimately cause a short-circuit or open-circuit failure. In this manner, the noise monitoring gives an early indication of a probable failure. Figs. 15, 16, and 17 refer to this equipment.

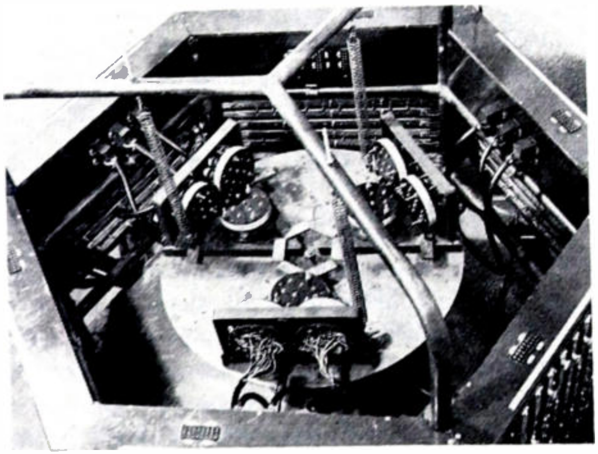

Fig. 17—Vibration table on which three frames, each carrying three groups of valves, are mounted. Wiring from the sockets goes to the connecter strips on the outer periphery of the machine.

Life Test. A batch of the valves is run on normal static life test for a period of 2,000 hours starting after the initial electrical test. The results of the life test are studied at the conclusion of the vibration tests (usually two weeks after the valves have been placed on life) to give an indication of the early life behavior of the batch.

At the conclusion of the above range of tests, the remainder of the batch, which has not been subjected to the vibration test, is retested once more to determine any failures occurring on storage.

Throughout the above tests, a double sampling system is employed and the fate of the batch as a whole depends on the results of the samples subjected to each of the individual mechanical tests. (See Fig. 18.)

By this means, it is possible to ensure that the quality rating of the valve in manufacture remains up to the standard required by the design engineer. Full records are kept on each batch so that at any time later reference may be made to the intial test history of a valve.

VII. Assistance from the Customer

Conservative Operation

It must be appreciated by the equipment designer that a valve cannot have the safety factor of other components, and in case of misuse will often act as the circuit fuse. Knowles of Westinghouse in a recent article stated this truth in the following words: "If valves could form a union, the first thing they would do would be to strike on the grounds of discrimination, speedup, and hazardous working conditions."

It is essential that designers should not use valves when another device would be more suitable, should be especially careful to select the correct valves for the job, should operate all valves at conservative ratings, and should always design for a "safe" failure.

Reliability can only be achieved by the closest cooperation between equipment designer and valve manufacturer. The valve makers' advice not only on the

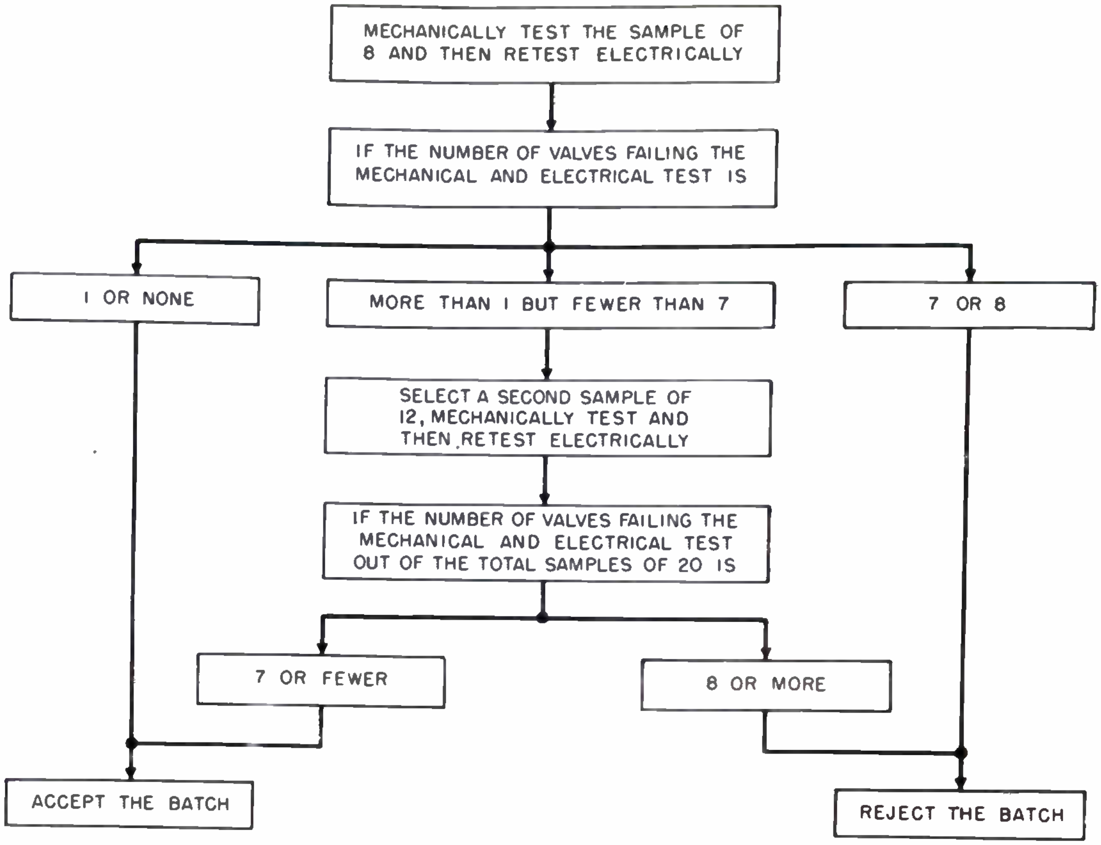

Fig. 18—Double sampling procedure.

conditions of use, but on the methods of connection, valve-holder tolerances, soldering requirements, and the like, must be scrupulously observed if the desired results are to be obtained.

Another aspect of this co-operation is to get assurances that circuits will accept valves made in the widest possible electrical limits. The proportion of reliable valves required will steadily increase and, in the event of an emergency, will immediately become 100 per cent of the valve manufacturers' production. Thus, any necessity for the selection of a limited range of characteristics would be little short of calamitous.

Type Diversity and Continuous Production

It is a well-known axiom that high reliability can only be obtained from valves having a high yield in manufacture, i.e., a low production shrinkage, and that such a state can be achieved only by uninterrupted production over a considerable period of time. It is therefore imperative that the valve maker shall have adequate orders to enable him to achieve this condition and that the number of types involved shall be as restricted as possible. The success of the project depends on circuit designers confining themselves to a short list and being prepared to use more valves of these types rather than employing still another type that may be more elegant technically.

Field Reports

No matter how difficult it may be to organize, the valve maker must be given adequate information from the field regarding the performance of his valves, because it is these data that enable him to maintain an accurate correlation between field conditions and the many test machines designed for factory usage. It is appreciated that with equipment distributed all over the world, often in the hands of semi-trained personnel, this requirement is not easy to meet; but it has been solved by both the American and Canadian air lines and determined attempts are being made in this country to see that the valve manufacturer is not hamstrung from lack of information.

VIII. Shape of Things to Come

While the main effort at present is directed towards making reliable replacements of existing types, it is very important to consider the way to go in the future.

The biggest stumbling block to ultimate reliability of glass-based valves is the valve holder. The valve manufacturer has recognized the problems of incompatability between valves and valve holder and has compromised by specifying the use of a wiring jig to centralize the socket contacts during circuit assembly and of a pin-straightening jig for the valve pins before insertion into the valve holder.

Despite all this, considerable evidence has been secured that semi-skilled personnel can cause a "mechanical insertion loss" of 3 per cent or greater, and while this can be reduced by careful education, the requirement of 1 per cent in 1,000 hours is easily swamped by this single possibility.

Because of the inevitability of this loss, it is probable that the reliable conventional type of valve of the future will have flying leads and will be soldered into the circuit. The size of the envelope will be dependent on the dissipation requirement, and the subminiature will be employed for low-dissipation needs with the miniature (18.5-mm, 0.73-inch) and noval (20-mm, 0.79-inch) types being used for better characteristics and higher dissipation.

Photographs of such types, which are now available, are shown in Fig. 19.

IX. Conclusion

In conclusion, evidence so far obtained shows that early life failures, which are the cause of most of the heartburnings, are due almost entirely to mechanical and glass troubles.

If these are eliminated by attention to design and manufacturing methods, together with a short life run, there is a great hope that for at least 1,000 hours the failures will be negligible. However, correlation between factory tests and field experience can only be achieved with large-scale usage of these better valves, after which it is quite possible that, with user co-operation, the valve maker will be able to issue guarantees showing actual failure rates.

It is a hazardous thing to prophesy, but within five years the reliability of the valve can be such that the present criticisms will then be directed towards other components. As reliability work must inevitably take considerable time, adequate pressure should be maintained on manufacturers of other components to ensure that all the constituent parts of an equipment keep in step on this matter.

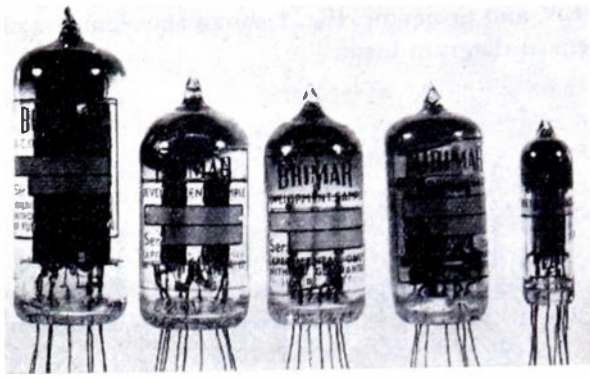

Fig. 19—Flying lead valves having, from left to right, the electrical characteristics of the 6CH6, 12AU7, 6AL5, 8D3, and a subminiature pentode.

X. Acknowledgments

Thanks are due to many members of the staff of Brimar Valve Division, who have done the work described and have assisted in the preparation of this paper, and also to the Admiralty for permission to publish parts of the work done on their behalf.

An Experimental System for Slightly Delayed Projection of Television Pictures*

PAUL MANDEL†, ASSOCIATE, IRE

Summary—**An experimental system is presented, comprising a high-definition flying-spot scanner, microwave relay, transcriber, fast film processor, and projector. A description is given for each part of the system and the performances obtained.**

I. Introduction

THE PRINCIPAL DIFFICULTIES in presenting television pictures to large audiences in the past has been insufficient illumination on the large screen, and also lack of high contrast and fine details of the picture. Illumination of the screen must be in the order of several thousand lumens. The contrast required is at least 50 to 1, while fine details should be comparable with those of standard film projection.

The various systems that have been described in the literature may be divided into three groups: direct projection, eidophor, and intermediate film. The direct projection of television pictures has been reported by several research laboratories.[1] The eidophor system[2] is favored by the Swiss school, but the intermediate film system, first proposed by Fernseh,[3] is now regaining favor and is the type used in this paper.

* Decimal classification: R583. Original manuscript received by the Institute January 3, 1952; revised manuscript received March 5, 1952.
† La Radio-Industrie, Paris, France.

[1] V. K. Zworykin and W. H. Painter, "Development of the projection kinescope," Proc. I.R.E., vol. 25, pp. 6-8; August, 1937.
Knoll, "Kathodenstrahl Bildübertragungsröhren Telefunken-Hausmitteilungen.—vol. 81, pp. 65-79; 1939.
E. Schwartz "Entwicklung der Braunschen Fernsehröhre.—Fernseh Hausmitteilungen Band 1, Heft 4, pp. 123-129; 1939.
P. Mandel, "An experimental television projector," Proc. I.R.E., vol. 37, pp. 1462-1467; December, 1949.
Communication of T. W. Lance at the International Television Congress, Milan, 1949.

[2] F. Fischer-Thieman, "Theoretische Betrachtungen.—Schweizer Archiv für Angewandte Wissenschaft. Heft 1 und 2; 1941; Heft 11 und 12; 1941; Heft 1; 1942; Heft 5-7, 10, 1942.

[3] G. Schubert, "Das Zwischenfilmverfahren."—Fernseh Hausmitteilungen, Band 1, Heft 3, pp. 65-72; 1939.

The project was started in January, 1949, based on the French high-definition standard of 819 lines. A description of the complete apparatus and of the results obtained follows.

II. General Description

The complete system consists of a high-definition flying-spot scanner, microwave relay, registering apparatus, and projector. Fig. 1 shows the elements of the system in diagram form.

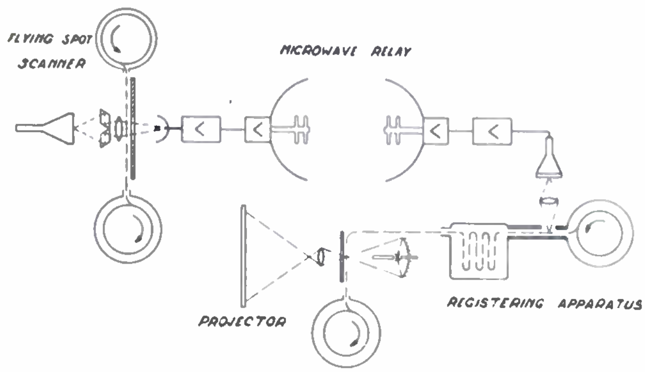

Fig. 1—Schematic disposition of the elements of the system.

The circuit begins with a 35-mm film in the flying-spot scanner which generates the television signal. The microwave relay carries this signal to the desired location.

In the registering apparatus the picture is reproduced on a cathode-ray tube where it is photographed by a synchronously driven camera. The exposed film is driven continuously through the developing, fixing, and drying device and emerges completely processed in 60 seconds. The film then passes through a standard high-powered movie projector for display on the large screen. Special 16-mm film is used in this part of the equipment. Since the sound is simultaneously registered, the film can be kept in stock as long as desired for future projection.

A. Choice of Analyzer

It was felt at the beginning that it would be necessary to have a standard television picture analyzer of very high performance, one capable of maintaining this performance during the time required for the development of the other parts of the system. As the primary source of the picture, the 35-mm standard film was chosen. The reasons for this choice are its high resolution, extended contrast range, invariability with time, ease of optical control, and facility of comparison of the final picture with the initial picture under identical conditions.

The analysis of the film is performed by a "flying-spot" scanner. The well-known advantages of this type of analyzer are the absence of dark spots, perfect linearity of the electro-optical response, and well-defined black level in the video signal.

On the other hand, the disadvantages of the flying-spot scanner are that it requires uniform speed of the film across the window, special high-precision optical system, and compensation for the inertia of the fluorescent screen on the cathode-ray tube. However, these disadvantages were not considered insurmountable. The excellent resolution and the negligibly low noise level which were finally obtained justified the choice of this scanning method.

III. High-Definition Flying-Spot Scanner

A photograph of the flying-spot apparatus is shown in Fig. 2. Fundamentally, the flying-spot scanner is an apparatus for generating a television signal from a film.

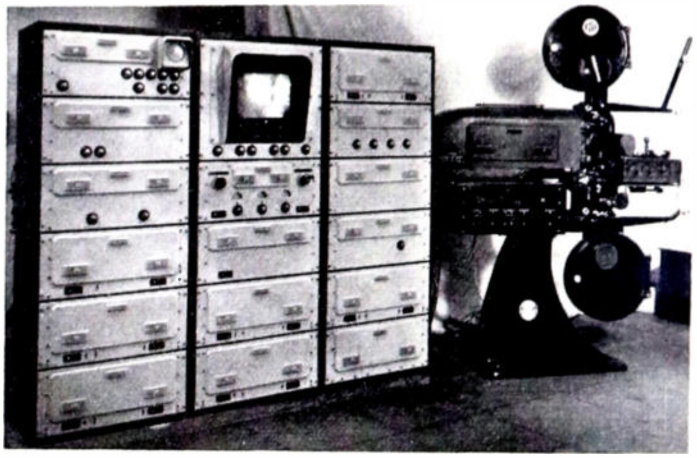

Fig. 2—View of the flying-spot scanner.

It consists of a cathode-tube light source, lens system, shutter, film drive, photoelectric cell, and amplifier. A diagram of the optical system is shown in Fig. 3.

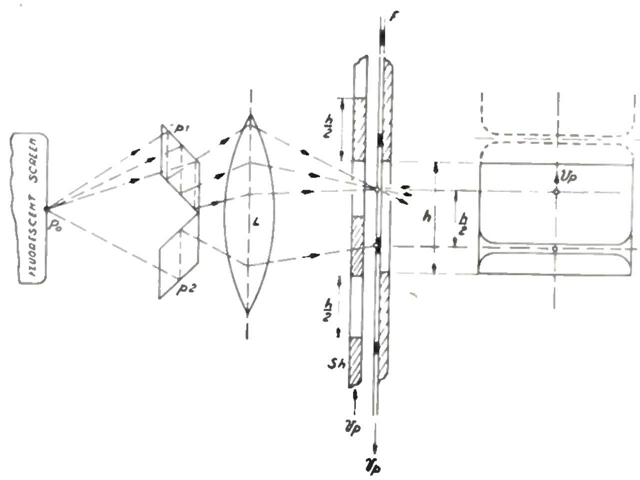

Fig. 3—Split optical system of the flying spot scanner.

A. Optical System

The primary light source is formed by the spot P_0 as it traces two interlaced frames of constant brilliance on the fluorescent screen of the high-tension cathode-ray tube. The time for each frame is 1/50 second, corresponding to the French standard.

A part of the luminous flux issued from the spot is split by the two prisms p_1p_2, and with the aid of the lens system, produces two identical images on the film, the separation between the images being exactly equal to one-half the picture height. The synchronous shutter S_h has equidistant apertures equal to one-half the picture height and is disposed in the immediate vicinity of the window. The shutter is driven in phase with the picture and permits interlaced scanning of the film by alternatingly interrupting one of the two images.[4]

The film is driven at the rate of 25 frames per second in order to allow the scan by two interlaced frames of 1/50 second for each picture on the film. The vertical speed V_p of the spot picture on the film is equal to the film speed, but in the opposite direction. The luminous flux passing through the film is proportional to the transparency of the film at the scanned point. The resulting luminous flux is collected by a condensing lens and projected to a photoelectric cell.

The optical quality of the prisms must be very high so as to avoid limitation of the resolution by lack of exact superposition of two successive frames. The lens system has been especially designed to keep the aberrations to a minimum in spite of the split system, and to assure uniform brilliance of both scanning spots across the window. Any difference in the luminosity of the spots would result in a disturbing 25-cps flicker in the televised picture.

The luminous efficiency of the split optical system cannot exceed 0.50. The measured value was 0.40. The difference is explained by the absorption in the optical system, the reflection from the surfaces, and by the aperture limitation of the lens system caused by the prisms. In order to keep the absorption losses to a minimum, the optical system was made of special glass with very slight absorption from 3,000 angstroms up. The lens system is coated to diminish reflections in the blue-green region of the spectrum.

B. Residual Flicker

It was found that the residual 25-cps flicker, when using the full aperture of a carefully designed optical system, was prohibitive. It was possible to reduce the flicker to a satisfactory level by the periodic modulation of the intensity of the scanning spot,[5] or by periodic modulation of the video gain by means of a 25-cps signal of suitable wave form.[6] The first method has its limitation in the saturation effect of the fluorescent screen, the second in the noise level of the multiplier. A third method of reducing the flicker is by limiting the aperture of the system on the front side. But the amplitude of the signal is correspondingly reduced. All three methods give satisfactory results in connection with a well-designed optical system. The influence of film shrinkage can be quickly compensated for by a slight axial displacement of the prism system or of the cathode-ray tube. It was not considered necessary to make this adjustment automatic.

C. Film Speed

One of the requirements for high resolution is the best possible uniformity of film speed. Two methods are used to achieve this. First, the rotational speed of the projector is stabilized by an elastically-coupled high-speed fly-wheel system. Second, the teeth and cylindrical surface of the main sprocket drum are carefully ground to obtain the highest possible uniformity of the film movement. As a result of these measures, there was no noticeable loss of detail due to the lack of superposition of successive frames. Stability observations using standard 35-mm film prints indicated a picture stability at least as good as that obtained by standard intermittent projectors using the "Maltese Cross" driving system.

D. Cathode-Ray Scanning Tube

The performance of the entire system depends largely on the quality of the scanning tube. Extremely small spot size (about 0.05 mm = 0.002 inch) and high current density are desired from an electro-optical point of view. High luminous efficiency and a small time constant of the fluorescent screen are of considerable value in facilitating electrical compensation for the luminous lag of the screen.[7] Uniformity of focus on the screen is obtained by limiting the sweep deflection angle. This uniformity is necessary to obtain constant luminosity (particularly violet and ultraviolet) for each point of the scanned picture because the efficiency of most screen materials depends on the current density in the spot. The ideal screen material would have an optical response presenting no lag behind the electronic excitation.

A simple 150-mm (6-inch) cathode-ray tube was developed to meet the above requirements. The tube is magnetically focused and has a triode gun. An especially treated, very pure Z_nO_2 powder was selected for the screen material after experimenting with many different types.

The active diameter of the spot was measured with the aid of a resolution pattern and was found to be 0.07 mm. The beam current was 50 μa and the accelerating potential was 30 kv. Commercially available tubes gave similar results except for the lower focus uniformity at the corners of the picture.

E. Compensation

There are two methods of compensation for the time lag of the screen material. One method uses only the ultraviolet light separated by a Wratten-type 18 filter. The resulting time lag is of the simple exponential type

[4] R. Möller, "Mehrfach-Filmablastung."—Fernseh Hausmitteilungen 1942, Band 2, Heft 5, pp. 129–133.
[5] French Patent P.V. 598,127; October 14, 1950.
[6] French Patent PV 48,755; November 18, 1950.
[7] R. D. Kell, "An experimental simultaneous color TV system," Proc. I.R.E., vol. 35, pp. 861–875; September, 1947.

which may be easily compensated by a suitable RC network. But the disadvantages exceed the advantages. The loss of luminosity through the filter, and the heavy absorption of this part of the spectrum in the lens system, increase the noise-to-signal ratio considerably.

Moreover, the crystal structure of the screen is more apparent, in the form of a sandy background on the picture. Therefore, it was not found desirable to use only ultraviolet radiation.

The second method is to use the total light output and to provide the much more complicated compensation required. Compensation is accomplished in this case by three suitable RC networks incorporated into the first stages of the video amplifier.

This method has given excellent results. The compensation for the screen material has proved very stable. Slight adjustments are required only after replacement of a scanner tube or after considerable modification of the beam current or accelerating potential.

F. Video Amplifier

The bandwidth of the video amplifier is 15 mc, 3 db down. No phase correction is used. The nominal bandwidth required is 10.5 mc.

The input to the amplifier comes from the multiplier type of photoelectric cell. The over-all noise level depends upon the sensitivity of the photocathode and upon the quality of the photo multiplier. The actual noise level is negligibly low. When standard black and white motion-picture prints are used, the noise is imperceptible from a viewing distance greater than three times the height of the picture.

The output of the amplifier is proportionate to the transparency of the film, owing to the linear response of the photocathode, multiplier, and the video amplifier. On the other hand, the luminous output of a cathode-ray tube is not proportional to the modulating voltage.[8] Whether the cathode-ray tube is used as the usual monitor or is photographed makes no difference. Therefore, a precorrection called "gamma correction" is incorporated in the amplifier.

The gamma correction is particularly important when a negative print is scanned. The correction is made by the use of the nonlinear characteristics of a suitably chosen amplifier tube.

The black reference level of the television signal is formed by the suppression of the scanning beam at the end of each line and each frame. In the formation of the gamma correction it is necessary to re-establish the black level. This is done by means of a synchronously driven clamping system[9] on the grid of the corrector stage. The degree of correction can be changed by the displacement of the reference level on the nonlinear portion of the tube characteristics.

Great care was taken to keep permanently the geometrical distortion of the picture at a low level by the use of passive correction networks and by avoiding the nonlinear regions of the sweeping tube characteristics. The passive networks include inverse feedback in the scanning circuits.[10] The geometrical distortions are less than ±4 per cent in each field element, representing 1/64th of the picture area.

G. High-Voltage Power Supply

The 30-kv dc power supply for the scanner tube is operated as a quadrupler, using cascaded filament transformers and rectifier tubes. A photograph of the power supply is shown in Fig. 4.

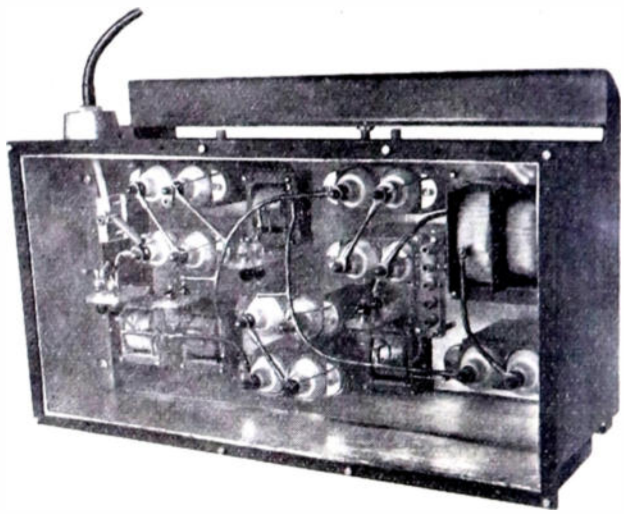

Fig. 4—Inside view of the 30-kv supply.

Two rectifier tubes are used in parallel in each stage in order to minimize the probability of an interruption due to heater failure. The entire compact unit is immersed in oil to reduce the size. The breakdown voltage of the polyethylene high-tension cable is sufficiently high to permit the use of an oil light metallic cone in place of the usual large ceramic insulator.

IV. Microwave Relay

The broadband microwave relay was developed to provide point-to-point transmission for distances in the order of twenty-five miles. The choice of its characteristics was guided by the desire to obtain a portable, simple and stable apparatus, capable of long time operation with a minimum of maintenance personnel. The carrier frequency of 940 mc allows reasonably high gain from the parabolic mirrors and at the same time facilitates the attainment of 5 watts of peak power. The above considerations and the large bandwidth of 10.5 mc led to the use of amplitude modulation. A photograph of one element of the relay is shown in Fig. 5.

Both the transmitter and the receiver have the same

[8] P. Mandel, "Appareillage de Télévision à 1015 lignes," *Bull. S.F.E.*, 6e série, Tome V, n° 47; May, 1945.

[9] L. W. Morrison, "The radar receiver."—*Bell Sys. Tech. Jour.*, vol. 26, p. 754.

[10] P. Mandel, "Tecnica a Sviluppo Dei ricevitori di Televisione—Televisione Italiana, vol. 1, pp. 5–12; March, 1950. Also, French Patent P.V. 631 598; July 5, 1952.

external appearance for the dipole, parabolic mirror, pedestal, and the box on the back of the mirror. Their uhf circuits are located at the back of the respective mirror in the weatherproof box.

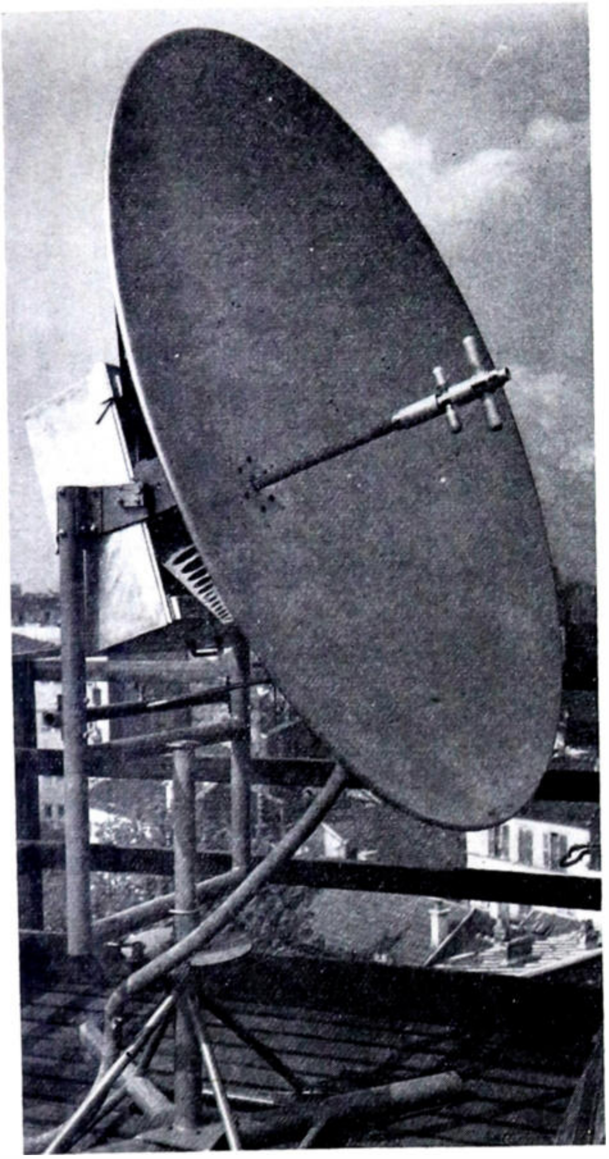

Fig. 5—View of the relay transmitter.

A. Transmitter

The basic outline of the transmitter is shown in Fig. 6.

The uhf circuit consists of the master oscillator, buffer, and final amplifier. The final amplifier is modulated on the cathode at a fixed black level. The modulator and black level clamp are pictured in Fig. 5.

All the uhf circuits are of the coaxial-cavity type. The thermal frequency drift of the master oscillator is compensated by a suitable choice of methods for the construction of the cavity. The remaining drift does not produce a noticeable deterioration of the received picture in spite of the partial suppression of one sideband in the receiver. Thus no warm-up time is required before transmission. All the power amplifiers are of the grounded-grid type using lighthouse tubes.

A monitor on the transmitter output is provided by a small, loosely coupled, wide-band cavity with a silicon crystal detector and a cathode follower.

The output of the final amplifier is connected directly to the quarter-wave dipole located in the focus of the mirror. A parasitic dipole reflector reinforces the radiation from the main dipole to the mirror. Quarter-wave chokes are used to prevent currents on the outside of the dipole supports.

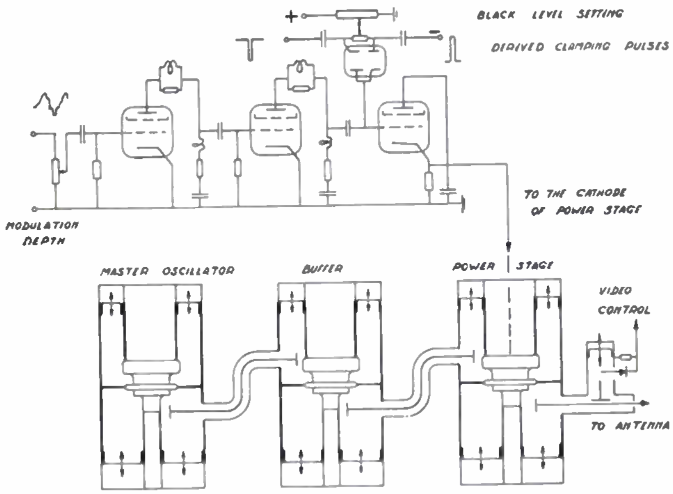

Fig. 6—Simplified schematic of the relay transmitter.

The power gain of the mirror over the isotropic radiator is 20 db and the beam angle for half power is 9 degrees. The voltage standing-wave ratio of the antenna feed is less than 1.15 over the radiated bandwidth of 30 mc.

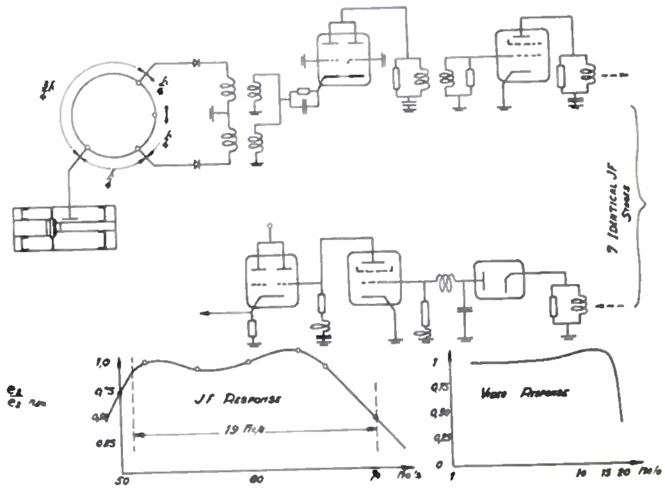

Fig. 7—Simplified schematic and frequency characteristics of the relay receiver.

B. Receiver

The receiver uhf circuit consists of a parabolic mirror, dipole, magic T mixer, and local oscillator. The mirror and dipole are exactly the same as for the transmitter. Since the total gain of the two antennas over dipoles is 36 db and since the transmitter power output is 5 watts, the signal available at the receiver is equivalent to a transmitted power of 20 kw with plain dipoles. A diagram of the receiver is shown in Fig. 7.

The magic T mixer is in the form of a coaxial ring. The push-pull outputs are connected by suitable quarter-wave transformer to the mixer crystals. The resulting intermediate-frequency outputs are combined by bifilar transformers with 1-to-1-turns ratio.

The output is fed to the cathode of the first IF stage (triode). The other six IF stages are pentodes. All seven stages have double-tuned coupling transformers. The video detector is followed by a suitable amplifier and a cathode follower.

The receiver sensitivity for an output of 1-volt peak-to-peak is about 250 μv with less than 1 per cent noise power. Bandwidth response curves are shown in Fig. 7 for the IF and video amplifiers.

C. Remote Control

The transmitter and receiver are each connected by a special cable to the corresponding central rack. Fig. 8 is a photograph of the two control racks.

Fig. 8—Control racks of the relay.

Each of the control racks contains a small oscillograph for monitoring the wave form and a large cathode-ray tube for monitoring the television picture. Also in each control rack are all the power supplies for each unit.

Operationally, only one control is required at each end of the relay. At the transmitter there is a control for adjusting the depth of modulation, depending on the signal lever available. At the receiver is the receiver gain control for setting the output signal at the proper level.

(a)

(b)

Figs. 9(a) and 9(b)—Reproductions of registered films.

V. Registering Apparatus

The registering apparatus is shown in Fig. 10 (a) and (b). On the left may be seen the twin transcriber units. In the center is the synchronous camera, and on the right is the continuous developing apparatus.

A. Transcriber

Fig. 10 (c) shows a close up rear view of the transcriber units. In the center are the twin 150-mm (6-inch) transcriber kinescopes. On the right is a 36-cm (14-inch) cathode-ray tube for monitoring the picture.

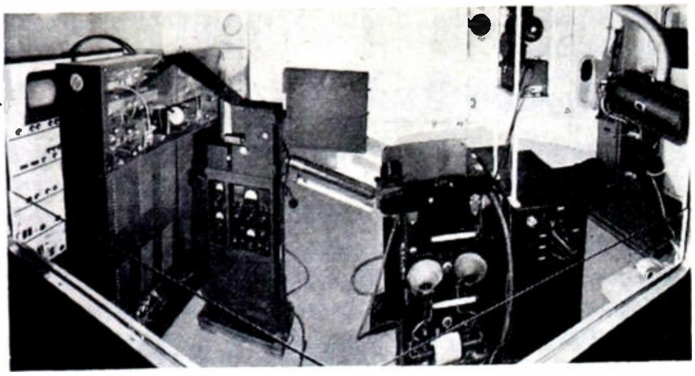

Fig. 10(a)—Registering apparatus and projectors.

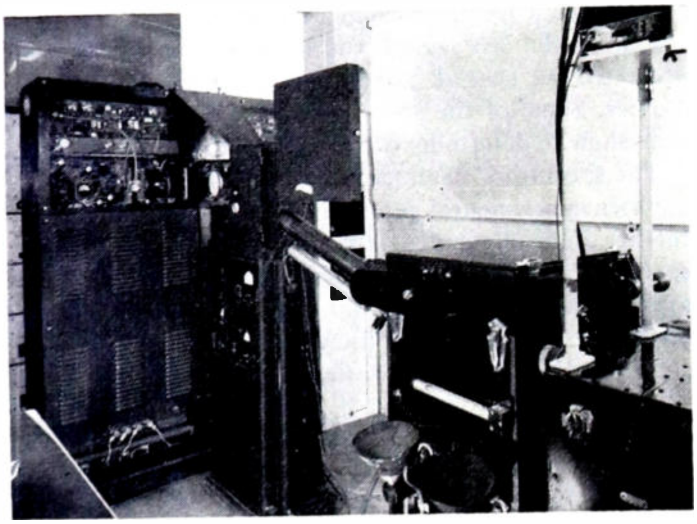

Fig. 10(b)—Registering apparatus kinescopes, camera, and continuous developer.

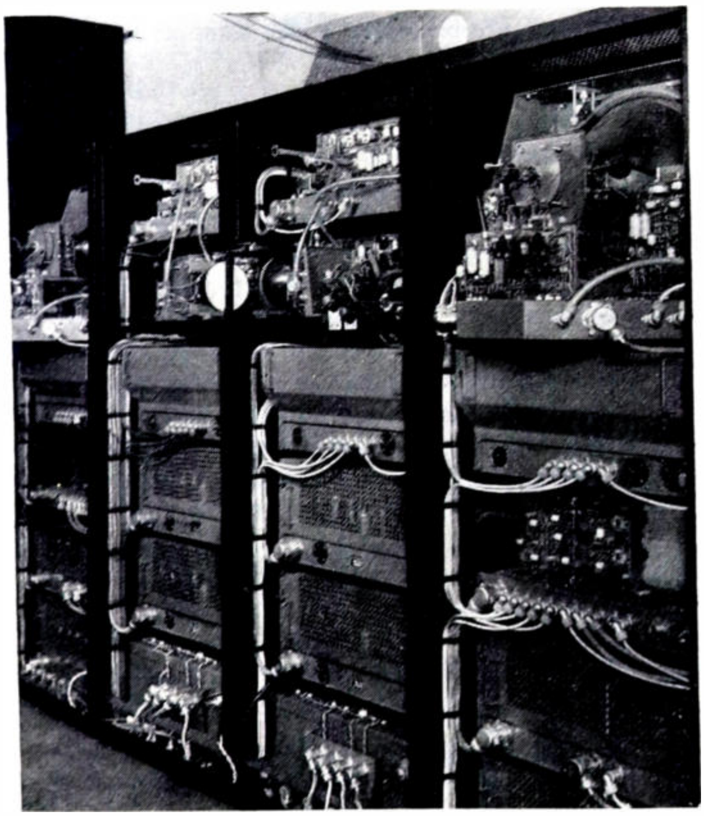

Fig. 10(c)—Rear view of transcriber units.

In the transcriber unit the video signal is amplified and the black level is restored by auxiliary pulses derived from the television signal. The resulting signal, which may be of either positive or negative polarity, as desired, is then applied to the control electrode of the transcriber kinescope.

The kinescope units are twin in order to insure continuity of service in case of failure of a unit. A revolvable mirror is positioned between the kinescopes to direct one of the kinescope pictures to the registering camera.

B. Registering Camera

The synchronously driven camera is equipped with a high-speed transport mechanism (36-degree shutter angle) running in phase with the television picture at the speed of 25 frames per second. Consequently, no single television field is lost in the photographic registering process. The type of 16-mm film used has a highly sensitive low-grain emulsion, especially treated to withstand the relatively high temperatures in the developing apparatus.

The accompanying sound is simultaneously registered on the film in the usual manner. The negative television picture is generally used to obtain directly a positive print ready for projection.

C. Developing

From the camera the exposed film passes continuously through the rapid developing apparatus and emerges completely processed after a time delay of 60 seconds. The apparatus is housed in two cases, as shown in Figs. 10(a) and 10(b). In one case are contained the various liquid solutions through which the film passes. The other case contains the associated circulating pumps, fans, electric heaters, and controls. The film is propelled in a helicoidal path through the various solutions by means of live rolls and guide tubes. The mechanism of this part of the system is particularly simple. The time delay is split up in the following manner:

Developing	6.5 seconds
Washing	5.0
Fixing	19.5
Washing	10.0
Drying	19.0
Total	60.0 seconds

VI. Projectors

The processed film passes through low-power control projectors and then through the high-power principal projector. The projectors are visible in Fig. 10(a) on the right side. The 80-amperes arc lamp in the final projector produces a mean luminous flux of about 2,000 lumens. The resulting illumination on the large screen of 4-meter base (13 feet) is 170 lux, and therefore in the same order as standard movie projection. Figs. 9(a) and (b) are televised pictures as reproduced on the film.

VII. Tests

Preliminary to the field testing of the complete apparatus, each element was tested separately. The resolution of the flying spot pictures was measured with the aid of a test pattern on a high-contrast 35-mm film. The horizontal resolution was found to be 900 picture elements and the vertical 650 elements.

The maximum distance which could be covered by the microwave relay before additional receiver noise could be detected on the picture was 25 miles. There was no measurable loss of resolution, owing to the high-frequency transmission. The stability of transmission was excellent. There was no observable change in picture quality or intensity over a period of several hours.

Tests of the registering apparatus showed a loss of about 10 per cent in resolution. A film of 800 picture elements was produced from an input signal of 900 picture elements. The input signal was generated by the flying-spot scanner using various types of pictures, such as the resolution pattern, gradation scale, and selected subjects on positive or negative prints. Since the total resolution loss was small, no attempt was made at that time to determine the distribution of loss among the various parts, such as the amplifier, kinescope, camera, film, and developing apparatus.

During this part of the work the optimum values were determined and fixed for the various factors, such as the light output of the transcriber tube, *f* number of the camera lens, temperature, and concentration of the film baths.

It is interesting to note that the registered detail was not fixed by the grain of the emulsion or by the spot size of the transcriber kinescope, but by the mechanical vibration of the camera. The upper limit was fixed by the aberration of the lens only when a less sensitive emulsion was used together with a high lens aperture.

The contrast range of the registered film was slightly inferior to that obtainable on a good movie print. Research is being continued to determine the cause.

A. First Field Test

The first field test of the apparatus was at the second "Salon du Cinéma" in Paris, from October 5 to 20, 1950. The test period was two hours each day during the peak visiting hours. An outdoor television camera supplied the signal by cable to the registering apparatus. The registered film was projected to a large screen for the audience in the Salon theater. Very satisfactory results were obtained.[16]

B. Second Field Test

For the second field test the registering apparatus was located at "Cinéma Madeleine" and the microwave relay operated between the theater "Gaumont Palace." At Cinéma Madeleine, television signals from two different sources were registered and projected.

Signals came by microwave relay from TV cameras installed in a studio of Gaumont Palace, and also from a TV receiver tuned to the TV broadcast from the Eiffel-Tower transmitter. The program originated in the studios of La Radiodiffusion Française.

During the last period, the television programs were projected as special attractions between the regular movies. Some of the spectators were interviewed after each show to determine their reaction to the experiment.

The spectators' comments were considered along with our own experimental results in determining the lines of further development of the apparatus. The average spectator's comment was, "slight loss of contrast." The average picture brilliancy was the same as for the regular movie. On the whole, the experimental results were considered as better than satisfactory by the spectators.

Conclusion

The tests indicate that this method of presenting television pictures to large audiences is entirely practical. The loss in fine details is only 10 per cent. The slight loss of contrast should be correctable by further experiment. The slight delay of 60 seconds is certainly not objectionable. The picture brilliancy is the same as for regular movie projection.

Acknowledgment

The accomplishment of this project involved three companies. The electronic apparatus consisting of the flying spot scanner, microwave relay, and the transcriber were developed by the Television Laboratories of La Radio-Industrie S. A. Paris, France, the camera, developing apparatus, and projectors by Andre Debrie, Paris, and the special 16-mm film was produced by the Sté. Pathe-Kodak, Vincennes.

The author is particularly indebted to H. N. Steen and A. Burgert of La Radio-Industrie for their valuable help during the development of the electronic apparatus.

[16] Le Film Français nº 317; November 17, 1950.

Gamma Correction in Constant Luminance Color Television Systems*

SIDNEY APPLEBAUM,† ASSOCIATE, IRE

Summary—**A theoretical study is made of the effects of precorrecting the red, green, and blue image co-ordinates to provide unity over-all gamma. The effect of gamma precorrection is presented with regard to (a) tone rendition on the compatible monochrome receiver, (b) maximum demand of subcarrier and composite signal, and (c) noise interference sensitivity.**

Introduction

THE EFFECTS OF NONLINEARITY in television upon picture quality has received much attention in the literature. Because of the many physical and subjective parameters involved, analysis is necessarily limited to first-order effects. The rest must be determined by experience.

In monochrome television, nonlinear transducers affect both the tone rendition and the signal-to-noise ratio. The brightness-transfer characteristic of the average picture tube compresses the shadow tones and expands highlight tones. However, it has been shown that the nonlinear characteristic of the average picture tube is almost ideal for minimizing noise sensitivity. This indicates that precorrections for the nonlinear characteristics of the receiver should be applied at the transmitter where the noise level is low.

The effects of nonlinear transducers in color television are not only more complex—they are also more perceptible. As a result, the need for gamma or nonlinearity correction in color television, in specific circuits designed for that purpose, has been recognized from the start. Nonlinearity in color television causes chromatic distortion as well as tonal distortion and correcting for it involves three signals as compared with only one. Additional complications arise in a color system designed for compatibility since any method for precorrecting the signal must simultaneously provide correction for both the nonlinear color receiver and the nonlinear monochrome receiver.

This paper attempts to determine the effects of precorrecting the signals in a color television system of the constant luminance type, i.e., a compatible transmission in which three signals are simultaneously transmitted, one of which contains the complete luminance information.

Because of the logarithmic nature of the response of the eye, the subjective effect of a nonlinear picture tube is most conveniently analyzed when the relation between the original image luminance Y and the displayed pictured luminance P is a power law of the form $P=Y^n$. The exponent n, is called the gamma of such a tube. Picture tubes in present use often have nonlinear characteristics which are not of the simple power-law type. Nevertheless, the usual approach to the understanding of their behavior is to approximate the transfer characteristic by a power-law relation over the range of operation. This approximation is justified not only by the simplification in analysis which results, but also by the insensitivity of the eye to the exact nature of the nonlinearity.

The effect of the gamma of a color receiver upon chromatic rendition is shown in the Maxwell triangle (Fig. 1). This figure displays chromaticity in terms of

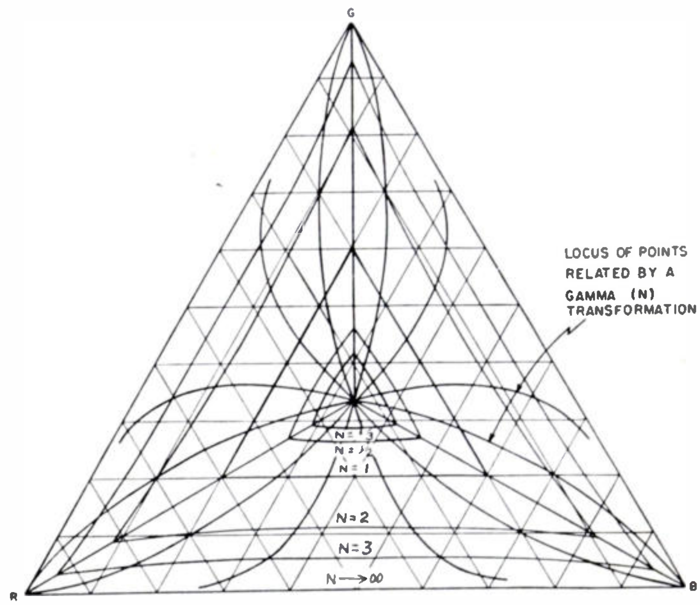

Fig. 1—Effect of gamma on chromaticity.

red, green, and blue co-ordinates, and shows the result obtained when a color point of co-ordinates, r, g, b, is transformed to the point r^n, g^n, b^n. As 'n' is varied, the original point moves along a particular path. For example, a gamma of 2 would transform a point from the $N=1/2$ contour to the $N=1$ contour, or from the $N=1$ to the $N=2$ contour. A number of typical loci are given in the figure. A gamma greater than unity increases the saturation and changes the hue of a color point towards the hue of the dominant primary. Colors which are originally of low saturation are changed mainly in saturation, whereas colors which are originally of high saturation are changed mostly in hue by gamma transformation.

For some scenes, the tonal and chromatic distortion which results from a gamma greater than unity may produce a more pleasing picture subjectively. However,

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since this will vary considerably from scene to scene, it is reasonable to require that the over-all gamma of a color TV system be controllable, with an over-all gamma of unity within the range of variation.

In order to obtain an over-all gamma of unity when nonlinear picture tubes are used, it is apparent that some form of compensation or precorrection is required. This correction may take the form of linearizing the picture tube through feedback, or a "rooting" circuit in the receiver. The alternative to these methods for making the receiver linear, is to make the correction at the transmitter. In the present state of the art, this appears to be the more economical choice. If the over-all system gamma is to be controlled at the studio, a gamma-correcting circuit is necessary at the transmitter. Furthermore, the use of nonlinear picture tubes, with gamma correction at the transmitter, minimizes the noise sensitivity of the system. A disadvantage of precorrection at the transmitter is that it requires standardizing display co-ordinates for the system.

In order to present the results of this paper, it is necessary to assume the reader is familiar with the principles of colorimetry.[1] In the interest of economy of expression matrix algebra is used, but a reading knowledge of the notation is all that is needed. The following discussion is intended only to establish the notation and language to be used in this paper.

Color television systems make use of well-established color-mixture laws. Essentially, these laws state that over a wide range of light intensities any color stimulus can be matched by the additive mixture of proper amounts of a set of 3 other independent color sources. The standard specification of color stimuli is made in terms of three nonphysical primaries (X), (Y), (Z), so chosen that the amount of (Y) light necessary to match a color stimulus is directly proportional to the luminance of the stimulus and that all physical color stimuli are specified by positive amounts of the three standard primaries. The units of measure for the three primaries are chosen so that equal amounts of (X), (Y), and (Z) light would produce the same sensation as an equal-energy white (light with a flat-power spectrum distribution).

This system, therefore, provides us with a three-dimensional co-ordinate system for the specification of color stimuli. In terms of these co-ordinates, a color stimulus, represented by a color point, would be specified as consisting of X amount of (X) light, Y amount of (Y) light and Z amount of (Z) light, i.e.,

$$C = X(X) + Y(Y) + Z(Z). \qquad (1)$$

If a set of primaries (R), (G), (B) is defined in terms of (X), (Y), and (Z), by linear transformation the co-ordinates R, G, B may be found such that

$$C = R(R) + G(G) + B(B). \qquad (2)$$

In color television, color stimuli are conveniently considered to have two intrinsic properties, luminance and chromaticity. Luminance is one-dimensional whereas chromaticity requires two dimensions. Color points of the same luminance lie in a plane parallel to the plane $Y=0$. Color points of the same chromaticity lie on a line through the origin. The chromaticity of a color point is commonly specified by the intersection of the radius vector to the color point with the plane $X+Y+Z=1$. The chromaticity of a point X, Y, Z may be specified in this manner by giving any two of the three co-ordinates of intersection, x, y, z, where

$$x = \frac{X}{X+Y+Z}, \qquad y = \frac{Y}{X+Y+Z},$$
$$z = \frac{Z}{X+Y+Z}. \qquad (3)$$

Chromaticity is commonly specified by giving x and y.

Although the color mixture laws enable the specification of color stimuli in a linear co-ordinate system, they do not specify the sensitivity of the eye to a small displacement (noise) in this co-ordinate system. A great deal of work has already been done to determine the sensitivity of the eye to displacements in the three co-ordinates. Most of this work, however, was done for relatively large fields which were stationary in both time and space. In television noise, we are concerned with fields which subtend small angles, fluctuate in intensity, and appear to have a random motion. Fortunately, experience seems to indicate that results obtained from the stationary measurements still apply to the perceptibility of random noise in television.

Perhaps the best known relation concerning the perceptibility of small color co-ordinate displacements is the Weber-Fechner Law. The law states that over a large intermediate range of luminance, the perceptibility of an achromatic displacement where

$$\frac{\Delta X}{X} = \frac{\Delta Y}{Y} = \frac{\Delta Z}{Z} \qquad (4)$$

is proportional to the percentage change or the Fechner fraction, $\Delta Y/Y$. At very low levels of luminance, the Fechner fraction necessary for perceptibility increases.[2]

More general investigations have been made by MacAdam[3,4] and others on the sensitivity of the eye to general color co-ordinate displacements. MacAdam has also made visual acuity measurements[5] which supply much of the type of information needed for color television. It has been shown that the eye is most sensitive to noise in the Y co-ordinate and least sensitive to noise in the Z co-ordinate. The sensitivity of the eye to X, Y, and

[1] W. T. Wintringham, "Color television and colorimetry," Proc. I.R.E., vol. 39, pp. 1135–1172; October, 1951.

[2] P. Mertz, "Perception of television random noise," *Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 54, pp. 8–34; January, 1950.

[3] D. L. MacAdam, "Specification of small chromaticity differences," *Jour. Opt. Soc. Amer.*, pp. 18–26; January, 1943.

[4] W. R. J. Brown and D. L. MacAdam, "Visual sensitivities to combined chromaticity and luminance differences," *ibid.*, pp. 808–834; October, 1949.

[5] D. L. MacAdam, "Color Discrimination and the influence of Color Contrast on Visual Acuity," Kodak Research Lab., Communication No. 1105.

Z noise on a given background varies with the chromaticity of the background and is roughly proportional to 5:13:2.

In this report, the effect of nonlinear display characteristics upon luminance or 'Y' noise only is evaluated. A complete evaluation of the effect of nonlinear display characteristics upon both luminance and chromaticity noise would involve considerable calculation. The uncertainty in the applicability of the available data, however, does not justify the work necessary.

Co-ordinate Transformations In A Color TV System

In a general color TV system, there are three separate sets of co-ordinates to consider: camera co-ordinates, transmission co-ordinates, and display co-ordinates. This section presents briefly the transformations between these sets of co-ordinates.

The camera is designed to obtain the standard tristimulus co-ordinates (X, Y, Z) of each point of the image. The 'Y' signal represents the luminance variations independent of chromaticity, and is used directly by the compatible monochrome receiver. A color receiver, however, must transform the luminance and chromaticity information it receives into the proper signals to apply to the grids of red, green, and blue tubes. The chromaticities of the receiver primaries may be specified in trichromatic units as

$$\begin{bmatrix} (\text{red}) \\ (\text{green}) \\ (\text{blue}) \end{bmatrix} = \begin{bmatrix} x_r & y_r & z_r \\ x_g & y_g & z_g \\ y_b & y_b & z_b \end{bmatrix} \begin{bmatrix} (X) \\ (Y) \\ (Z) \end{bmatrix}, \tag{5}$$

where (X), (Y), and (Z) are the units or primaries of the tristimulus co-ordinate system. It is most convenient to use a set of red, green, and blue primaries (R), (G), (B) having the chromaticities specified by (5), which have unit lengths such that equal amounts of the primaries specify some reference color. (The reference color used is illuminant C with unit amount of luminance. This will be referred to as unit reference white.) The transformations between *co-ordinates* in the (X), (Y), (Z) and (R), (G), (B) space will be written as

$$\begin{bmatrix} R \\ G \\ B \end{bmatrix} = \begin{bmatrix} c_{11} & c_{12} & c_{13} \\ c_{21} & c_{22} & c_{23} \\ c_{31} & c_{32} & c_{33} \end{bmatrix} \begin{bmatrix} X \\ Y \\ Z \end{bmatrix} = [C] \begin{bmatrix} X \\ Y \\ Z \end{bmatrix}, \tag{6}$$

and

$$\begin{bmatrix} X \\ Y \\ Z \end{bmatrix} = \begin{bmatrix} d_{11} & d_{12} & d_{13} \\ d_{21} & d_{22} & d_{23} \\ d_{31} & d_{32} & d_{33} \end{bmatrix} \begin{bmatrix} R \\ G \\ B \end{bmatrix} = [D] \begin{bmatrix} R \\ G \\ B \end{bmatrix}, \tag{7}$$

where $[C] = [D]^{-1}$.

In a 'constant luminance' system, the transmission primaries are chosen so that one of the co-ordinates is directly proportional to luminance. Since color points of equal luminance lie in a plane parallel to the XZ plane, two of the transmission primaries must lie in the XZ plane. The third primary is chosen to be coincident with the reference-white vector, so that two of the co-ordinates will be zero when colors on the reference-white vector are transmitted.

A set of transmission primaries (A_1), (A_2), (A_3) which satisfies the above conditions is

$$\begin{bmatrix} (A_1) \\ (A_2) \\ (A_3) \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 \\ X_0 & 1 & Z_0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} (X) \\ (Y) \\ (Z) \end{bmatrix}. \tag{8}$$

Note that (A_1) is the same as (X), (A_2) is the unit reference white $(X_0, 1, Z_0)$, and (A_3) is the same as (Z).

The transformations between co-ordinates in the (X), (Y), (Z) and (A_1), (A_2), (A_3) space are

$$\begin{bmatrix} X \\ Y \\ Z \end{bmatrix} = \begin{bmatrix} 1 & X_0 & 0 \\ 0 & 1 & 0 \\ 0 & Z_0 & 1 \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \\ A_3 \end{bmatrix}, \tag{9}$$

and

$$\begin{bmatrix} A_1 \\ A_2 \\ A_3 \end{bmatrix} = \begin{bmatrix} 1 & -X_0 & 0 \\ 0 & 1 & 0 \\ 0 & -Z_0 & 1 \end{bmatrix} \begin{bmatrix} X \\ Y \\ Z \end{bmatrix}. \tag{10}$$

The A_1, A_2, A_3 co-ordinates can also be related to R, G, B, co-ordinates,

$$\begin{bmatrix} R \\ G \\ B \end{bmatrix} = \begin{bmatrix} c_{11} & 1 & c_{13} \\ c_{21} & 1 & c_{23} \\ c_{31} & 1 & c_{33} \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \\ A_3 \end{bmatrix} = [F] \begin{bmatrix} A_1 \\ A_2 \\ A_3 \end{bmatrix}. \tag{11}$$

The A_1 and A_3 co-ordinates contain the chromaticity information only, which in a constant luminance system is transmitted in a channel separate from the luminance channel. The composite video signal using A_1, A_2, A_3 co-ordinates would be proportional to the sum of the luminance signal and the chromaticity subcarrier or

$$A_2 + A_1 \cos \omega_s t + A_3 \sin \omega_s t, \tag{12}$$

where ω_s is the chromaticity subcarrier frequency. The amplitude and phase of the subcarrier corresponding to a given color point may be obtained in (X), (Y), (Z) space by projecting the given color point in the direction of reference white upon the XZ plane. The vector in the XZ plane from the origin to the projection specifies the subcarrier phasor (Fig. 2).

The A_1, A_2, A_3 co-ordinates are only one set of possible transmission co-ordinates. Any other set, however, may be defined in terms of the A_1, A_2, A_3 co-ordinates as follows:

$$\begin{bmatrix} U_1 \\ U_2 \\ U_3 \end{bmatrix} = \begin{bmatrix} \alpha_{11} & 0 & \alpha_{13} \\ 0 & 1 & 0 \\ \alpha_{31} & 0 & \alpha_{33} \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \\ A_3 \end{bmatrix},$$

and (13)

$$\begin{bmatrix} A_1 \\ A_2 \\ A_3 \end{bmatrix} = \begin{bmatrix} \beta_{11} & 0 & \beta_{13} \\ 0 & 1 & 0 \\ \beta_{31} & 0 & \beta_{33} \end{bmatrix} \begin{bmatrix} U_1 \\ U_2 \\ U_3 \end{bmatrix}.$$

The composite signal in U_1, U_2, U_3 co-ordinates is proportional to

$$U_2 + U_1 \cos \omega_s t + U_3 \sin \omega_s t. \qquad (14)$$

The U_1U_3 plane is referred to here as the subcarrier plane. The U_2 co-ordinate is equal to luminance. Although in performing calculations on a given set of transmission co-ordinates it is unnecessary to introduce the A_1, A_2, A_3 co-ordinates, it is convenient to work in 'A' co-ordinates for comparing one set of transmission co-ordinates with another and for obtaining a simple geometric interpretation of the subcarrier.

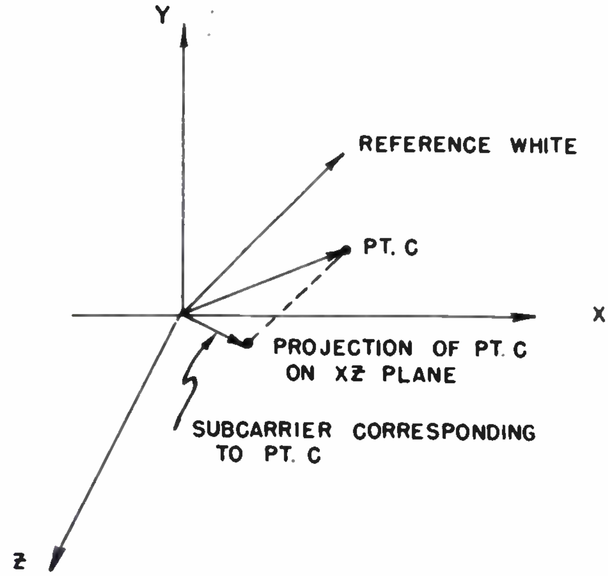

Fig. 2—Subcarrier in $A_1A_2A_3$ transmission co-ordinates corresponding to a color C.

Precorrection

The nonlinear characteristics of kinescopes may be approximated by a power-law of the form,[6]

$$P = P_{\min}\{1 + (K^{1/n} - 1)S\}^n, \qquad (15)$$

where

P = amount of light displayed on tube face,
$K = P_{\max}/P_{\min}$ = over-all contrast ratio,
$P_{\max}$ = maximum amount of light displayed,
$P_{\min}$ = minimum amount of light displayed,
S = normalized signal, varying from 0 to 1 as P varies from $P_{\min}$ to $P_{\max}$.

In a color receiver, there are three display channels. Assuming each has the characteristic of (15) and that equal signals produce the reference white, we may write

$$\begin{aligned} P_r &= P_{\min}\{1 + (K^{1/n} - 1)S_r\}^n \\ P_g &= P_{\min}\{1 + (K^{1/n} - 1)S_g\}^n. \\ P_b &= P_{\min}\{1 + (K^{1/n} - 1)S_b\}^n \end{aligned} \qquad (16)$$

In order to precorrect for these characteristics, it is apparent that the signals S_r, S_g, S_b must be obtained as follows:

$$S_r = \frac{\left(\frac{R}{Y_{\min}}\right)^{1/m} - 1}{k^{1/m} - 1}$$

$$S_g = \frac{\left(\frac{G}{Y_{\min}}\right)^{1/m} - 1}{k^{1/m} - 1}, \qquad (17)$$

$$S_b = \frac{\left(\frac{B}{Y_{\min}}\right)^{1/m} - 1}{k^{1/m} - 1}$$

where

R, G, B = co-ordinates of original image,
$Y_{\min}$ = minimum luminance within range of camera sensitivity,
n/m = over-all system gamma,
$k = K^{m/n}$ = over-all contrast range of camera sensitivity.

For an over-all system gamma of unity, the precorrection must be exactly inverse to the display characteristic, which requires that $m = n$ and $k = K$. Since it is very unlikely that the available contrast range for display (K) will be equal to the contrast range in the camera signal output (k), setting $k = K$ for precorrection will result in unity over-all gamma only in the range for which (16) are valid. Signals outside this range ($S < 0$, or $S > 1$) will be distorted.

A simplified schematic of the operations at the transmitter is shown in Fig. 3.

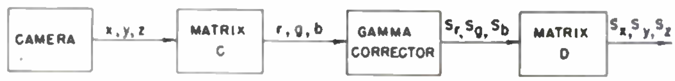

Fig. 3—Operations at a transmitter.

The X, Y, Z co-ordinates of the scanned image are obtained by the camera unit. These are transformed to normalized R, G, B co-ordinates by the 'C' matrix operation, which are then gamma corrected. Following gamma correction is the 'D' matrix operation which converts the gamma corrected R, G, B co-ordinates to corresponding X, Y, Z co-ordinates. With the gamma corrector out, the quantities S_x, S_y, and S_z would be equal to the original X, Y, Z co-ordinates, since the 'C' and 'D' matrix operations are inverse. S_x, S_y, S_z (or S_r, S_g, S_b) may be thought of as the co-ordinates of a virtual image which is a gamma precorrected version of the original image. The co-ordinates of this virtual gamma precorrected image are handled exactly as the co-ordinates of the original picture would be handled without gamma correction. Transmission co-ordinates would be

$$\begin{bmatrix} U_1 \\ U_2 \\ U_3 \end{bmatrix} = \begin{bmatrix} \alpha_{11} & 0 & \alpha_{13} \\ 0 & 1 & 0 \\ \alpha_{31} & & \alpha_{33} \end{bmatrix} \cdot \begin{bmatrix} 1 & -X_0 & 0 \\ 0 & 1 & 0 \\ 0 & -Z_0 & 1 \end{bmatrix} \cdot \begin{bmatrix} S_x \\ S_y \\ S_z \end{bmatrix}, \qquad (18)$$

U_1, U_2, U_3 could also be gotten directly from S_r, S_g, S_b).

[6] B. M. Oliver, "Tone rendition in television," Proc. I.R.E., vol. 38, pp. 1288–1300; November, 1950.

Effect of Gamma Correction on Compatible Black and White Receiver

The compatible monochrome receiver displays the luminance signal S_y as

$$P_y = P_{\min}\{1 + (k^{1/n} - 1)S_y\}^n. \tag{19}$$

If the original image point has the co-ordinates μR, μG, μB where μ is a proportionality factor, then

$$S_r = \frac{\left(\frac{\mu R}{Y_{\min}}\right)^{1/n} - 1}{k^{1/n} - 1}$$

$$S_g = \frac{\left(\frac{\mu G}{Y_{\min}}\right)^{1/n} - 1}{k^{1/n} - 1}, \tag{20}$$

$$S_b = \frac{\left(\frac{\mu B}{Y_{\min}}\right)^{1/n} - 1}{k^{1/n} - 1}$$

and

$$S_y = d_{21}S_r + d_{22}S_g + d_{23}S_b. \tag{21}$$

Since $d_{21}+d_{22}+d_{23}=1$, the displayed luminance is

$$P_y = \frac{P_{\min}}{Y_{\min}}\{d_{21}(\mu R)^{1/n} + d_{22}(\mu G)^{1/n} + d_{23}(\mu B)^{1/n}\}^n. \tag{22}$$

The luminance of the original point is

$$Y = d_{21}(\mu R) + d_{22}(\mu G) + d_{23}(\mu B),$$

which may be combined with (22) to give

$$\frac{P_y}{P_{\min}} = K_c\left(\frac{Y}{Y_{\min}}\right) \tag{23}$$

where

$$K_c = \frac{\{d_{21}R^{1/n} + d_{22}G^{1/n} + d_{23}B^{1/n}\}^n}{d_{21}R + d_{22}G + d_{23}B}.$$

K_c is the factor relating the image luminance to the displayed luminance. It is independent of the intensity factor μ and depends only upon the chromaticity of the image point. In other words, it is independent of luminance. Tone rendition on the monochrome receiver will, therefore, be linear for a constant chromaticity, but a chromaticity dependent factor (K_c) is introduced. For $n=1$, K_c is equal to 1 for all chromaticities. The values of K_c for $n=2$ and $n=3$ are shown in Figs. 4 and 5. For both cases, K_c is unity in the white region and greater than 0.8 over the major portion of the reproducible color triangle. Chromaticities which are close to the display primaries suffer the most attenuation, especially chromaticities in the vicinity of the blue primary.

Effect of Gamma upon Maximum Demand

The chromaticities which can be reproduced at the receiver are restricted to those which lie within the R, G, B triangle. The maximum luminance at which any chromaticity may be reproduced at the receiver depends upon the chromaticity of the illumination of the original scene and equipment limitations at both transmitter and receiver. For the purposes of forming a maximum demand criterion, it is reasonable to assume that the maximum values of the R, G, and B co-ordinates in the

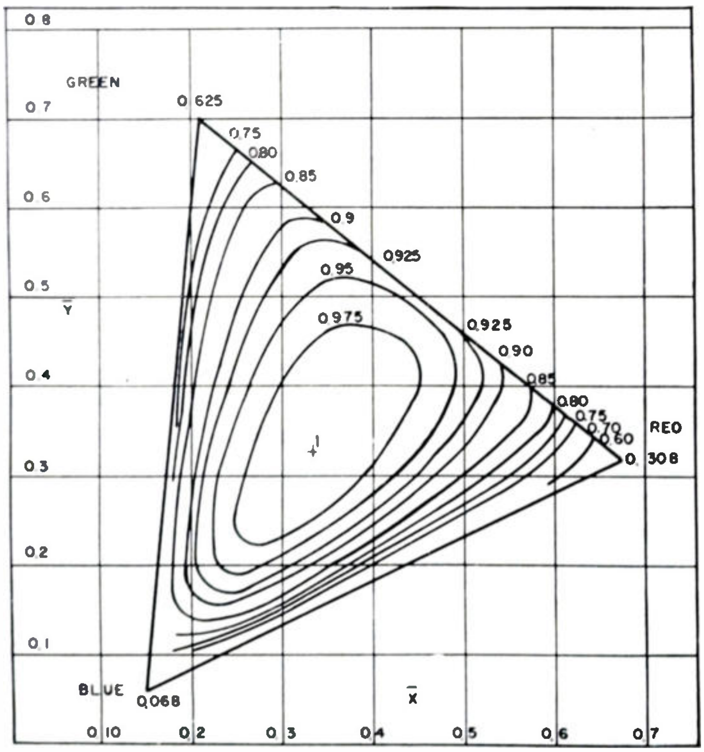

Fig. 4—Chromaticity factor K_c for compatible monochrome receiver with gamma = 2 (see Appendix C).

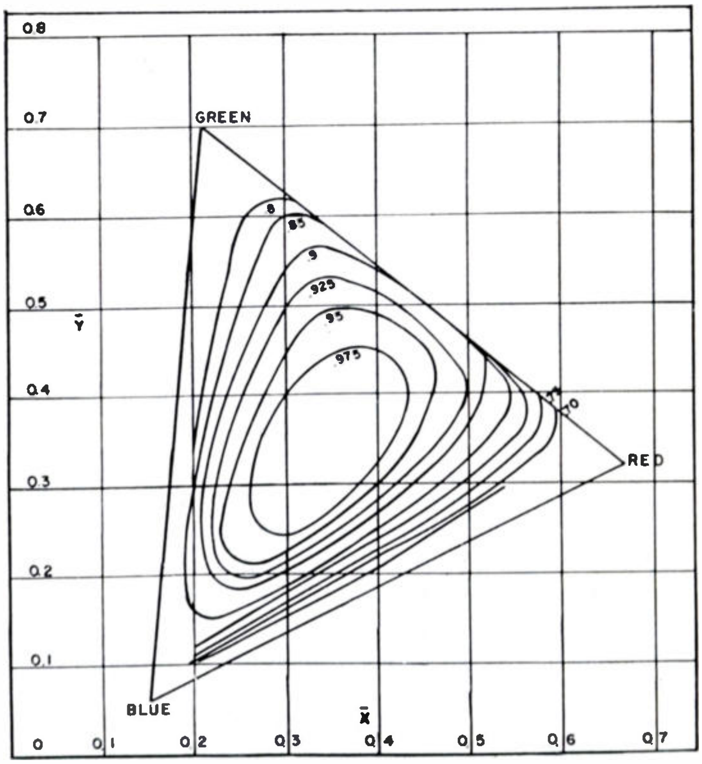

Fig. 5—Chromaticity factor K_c for compatible monochrome receiver with gamma = 3.

original image are equal to the value they take on at the highlight reference white. This is equivalent to assuming that all color points in the original image lie within a cube in the R, G, B space. For this assumption, Fig. 6

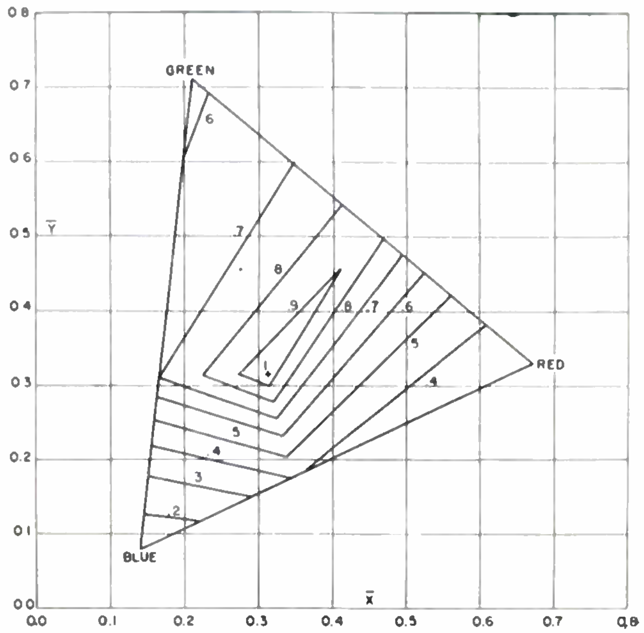

Fig. 6—Maximum luminance obtainable with primaries recommended by NTSC when maximum intensity of each primary occurs on illuminant C.

shows the maximum luminance relative to unity at illuminant C, at which any chromaticity will be reproduced by the primaries recommended for field test by the NTSC. Similar curves were given by MacAdam[7] for other receiver primaries. A maximum demand criterion might also be based upon the maximum visual efficiency of colored passive reflecting materials, first derived by MacAdam.[8] The maximum luminance contours of Fig. 6 based upon the cube in R, G, B space impose a somewhat more severe criterion on a color TV system for all chromaticities, except for those in the yellow-white region, than would maximum luminance contours based upon MacAdam's maximum visual efficiency curves. Since the cube criterion is also easier to visualize and analyze, it is used in this report.

The maximum demand criterion on the R, G, B co-ordinates may be used to find the maximum demand on the chromaticity subcarrier. If the subcarrier consisted of the signals A_1 and A_3 in quadrature, i.e., $U_1 = A_1$, $U_3 = A_3$, then, as remarked before, the subcarrier for any color point is obtained by projecting the point in the reference-white direction on the XZ plane. If the color points of the original image lie within a unit cube in R, G, B space, the co-ordinates S_r, S_g, S_b of the virtual gamma precorrected image also lie within this unit cube [this follows from (20)]. The S_x, S_y, S_z co-ordinates of the gamma corrected image lie within a parallelepiped in X, Y, Z space. The outline of the projection of this parallelepiped in the reference-white direction on the XZ plane gives the maximum demand on the subcarrier. The subcarrier maximum demand contour in A_1, A_3 co-ordinates is shown in Fig. 7 and is obtained by plotting the six points corresponding to the peak obtainable values of the receiver primaries and their complements. Fig. 7 shows that when $U_1 = A_1$ and $U_3 = A_3$, the subcarrier demand is largest on blue and yellow. Any other choice of U_1 and U_3 represents a linear transformation of A_1, A_3 co-ordinate space in which a point-to-point cor-

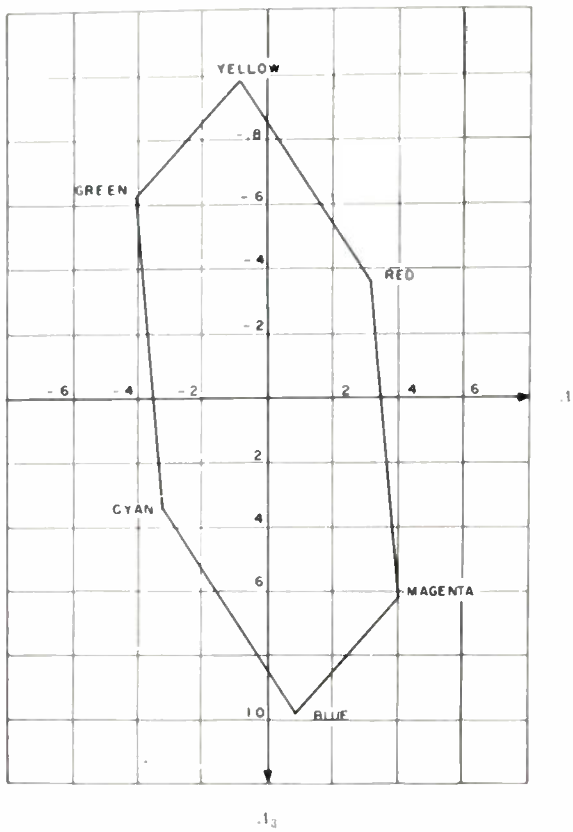

Fig. 7—Maximum demand of subcarrier in A_1A_3 co-ordinates when all colors lie within the unit RGB cube, for primaries recommended by NTSC.

respondence is maintained and in which straight lines are preserved. Because of this, the color points which lie on the maximum subcarrier demand contour in A_1, A_3 co-ordinates remain on the maximum subcarrier demand contour in U_1, U_3 co-ordinates. Therefore, to obtain the maximum subcarrier demand contour for any choice of U_1, U_3, it is only necessary to plot six points in the U_1, U_3 subcarrier plane. Furthermore, since the position of each primary in the subcarrier plane is symmetric about the origin with respect to the position of its complement, only three calculations are necessary. (The complement of a given color is defined here as that color which must be added to the given color to produce unit reference white.) Fig. 8 shows the maximum sub-

[7] D. L. MacAdam, "Quality of color reproduction," Proc. I.R.E., vol. 39, pp. 468–484; May, 1951.
[8] D. L. MacAdam, "Maximum visual efficiency of colored materials," *Jour. Opt. Soc. Amer.*, pp. 361–367; September, 1935.

carrier demand for the U_1, U_3 co-ordinates recommended by the NTSC for field testing.

It is important to investigate also the peak-to-peak swing on the combined video signal, which is proportional to

$$U_2 \pm \sqrt{U_1^2 + U_3^2}. \qquad (24)$$

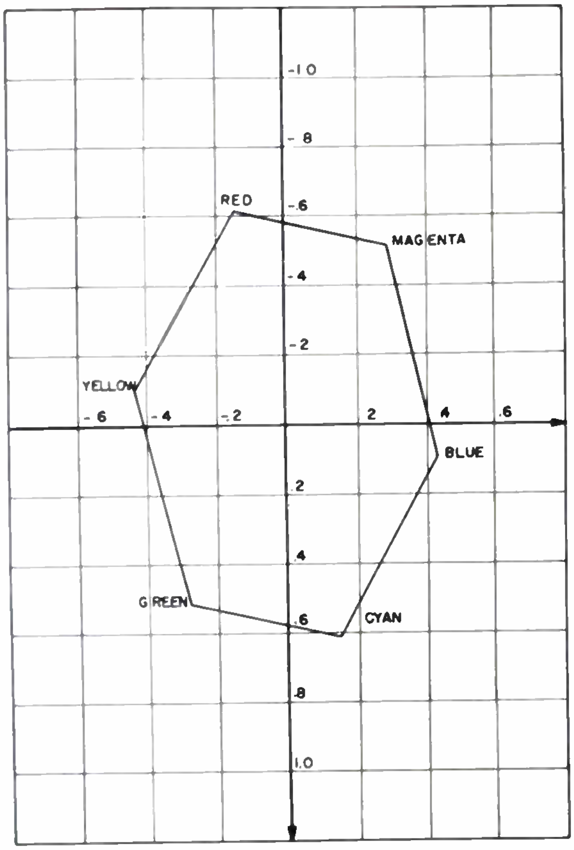

Fig. 8—Maximum demand of subcarrier in U_1U_3 co-ordinates recommended by NTSC.

On a monochrome picture (in reference white), only the luminance component U_2 is present, with a range of 0 to 1. The addition of the subcarrier for a color picture may make the combined signal greater than 1 or less than 0. This corresponds to the 'ultra-white' and infra-black regions, respectively, and would necessitate allowing more set-up in the transmitted signal, as compared with present practice, to prevent interference with the sync, transmitter overload, and intercarrier sound buzz.

The overswing and underswing of the combined signal corresponding to a given color point of the gamma distorted image are given by

$$M_0 = U_2 + \sqrt{U_1^2 + U_3^2} - 1, \qquad (25)$$

and

$$M_u = -U_2 + \sqrt{U_1^2 + U_3^2} \qquad (26)$$

respectively, when these quantities are positive. If, for any color point, M_0 and M_u are both negative, then the combined signal for that color point lies within the range 0–1. If the overswing M_0 is positive, then, the combined signal runs into the 'ultra-white' region. If the underswing is positive, then the combined signal runs into the infra-black region.

Consider the color point C, of the virtual gamma precorrected image, defined by the co-ordinates $[U_1, U_2, U_3]$ whose complement, C_c, with respect to unit reference white $[0, 1, 0$ in the U_1, U_2, U_3 space$]$ is $-U_1$, $1-U_2$, $-U_3$. The overswings and underswings of the complement C_c are

$$M_{0c} = 1 - U_2 + \sqrt{U_1^2 + U_3^2} - 1 = M_u, \qquad (27)$$

$$M_{uc} = (U_2 - 1) + \sqrt{U_1^2 + U_3^2} = M_0. \qquad (28)$$

These relations show that the overswings and underswings of the complement of a given color C are equal, respectively, to the underswings and overswings for the given color. This holds for any constant luminance set of transmission co-ordinates.

If the color point C (U_1, U_2, U_3) has an overswing $(M_0 > 0)$, it is apparent that the color point, C', defined by lU_1, lU_2, lU_3 (where l is a constant greater than 1), has the same chromaticity as C but a greater overswing. Increasing l until the second color point, C', lies in the maximum-demand surface (the three surfaces of the parallelepiped on which $R=1$, $G=1$, and $B=1$), gives the greatest possible overswing for all color points with the same chromaticity as C. The same reasoning applies to the underswings. The maximum possible overswings and underswings occur, therefore, for color points on the maximum-demand surface. If the complement of a

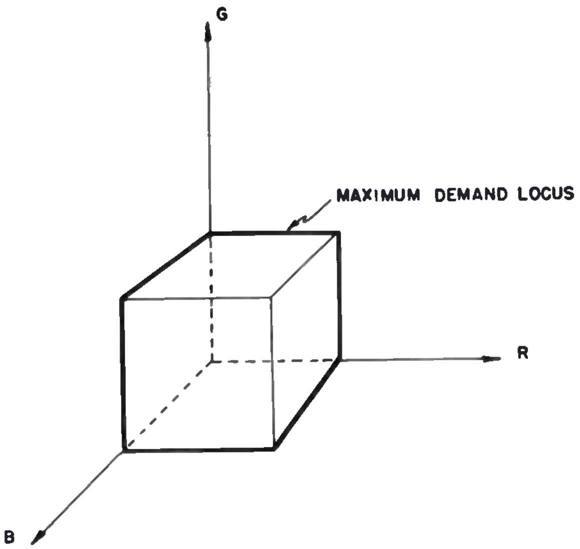

Fig. 9—Maximum demand cube showing locus of points on which maximum subcarrier and composite signal occur.

point in the maximum-demand surface doesn't also lie on the maximum-demand surface, the overswing or underswing of the original point is not the maximum possible. The peak overswings and underswings must occur on the locus of colors which lie on the maximum-demand surface and whose complements also lie on the maximum-demand surface. The desired locus, therefore, is defined by the lines joining the following points of R, G, B space in succession $[1, 0, 0]$, $[1, 0, 1]$ $[0, 0, 1]$ $[0, 1, 1]$, $[0, 1, 0]$, $[1, 1, 0]$. This is shown in Fig. 9. The same locus defines the color points which give maximum subcarrier. Fig. 8 may, therefore, be used to construct geometrically the overswings and underswings which occur on the maximum locus (see Fig. 10).

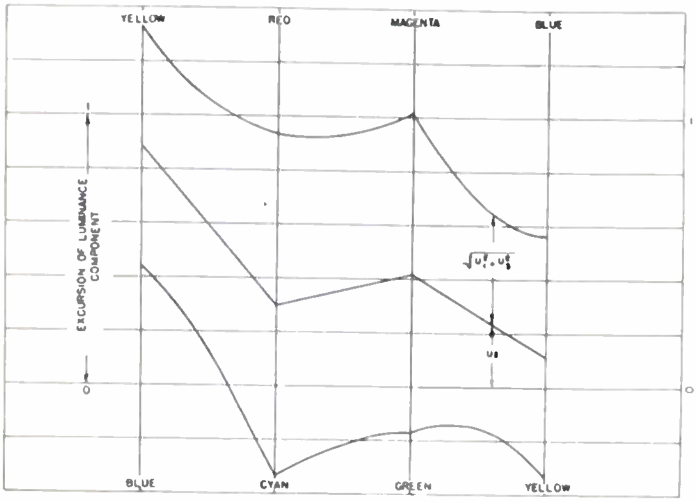

Fig. 10—Excursion of composite signal for color points on maximum demand locus.

The reasoning used above to locate the locus of peak subcarrier and combined signal demand applies to all gamma-distorted images including the linear case $(n=1)$. The amount of gamma correction, however, will affect the frequency of occurrence of peak subcarrier and combined signal. To illustrate this, assume, for example, that the average relative frequency of occurrence of color points of the original scene in the volume $dR\,dG\,dB$ is equal to a constant ν for all points R, G, B in the unit cube. After gamma correction, the point R, G, B is transformed to the point S_r, S_g, S_b. The frequency of occurrence of points in the volume dS_r, dS_g, dS_b at the point (S_r, S_g, S_b), ν_n, is given by

$$\nu_n(S_r, S_g, S_b) = \frac{\partial(r, g, b)}{\partial(S_r, S_g, S_b)}\nu. \tag{29}$$

Using (20), the average density of points along the edges of the cube is found to be

$$\bar{\nu}_n = V_{\min}{}^3 n^2(k-1)k^{(n-1)/n}(k^{1/n}-1)^2\nu. \tag{30}$$

The ratio of this average density to the average density when $n=1$ is

$$\frac{\bar{\nu}_n}{\bar{\nu}_1} = n^2\left(\frac{k^{1/n}-1}{k-1}\right)^2 k^{(n-1)/n}. \tag{31}$$

For a contrast ratio of 30, the frequency of occurrence of peak overswings or underswings is reduced 50 per cent with a gamma precorrection of $n=2$.

Noise Sensitivity

An analysis of the noise sensitivity of monochrome TV systems with display characteristics of the form of (15) has been presented by Oliver[6] and Mertz.[2] Their results are applicable to the compatible monochrome receiver in a color TV system. Since the sensitivity of the eye to small changes in luminance is proportional to the Fechner fraction over a large portion of the range of luminance involved in TV, a measure of the noise or interference sensitivity of a receiver is obtained by finding the logarithmic derivative of its display characteristic. The Fechner fraction produced by a noise voltage ΔS is

$$\frac{\Delta P_y}{P_y} = \frac{dP_y}{P_y dS}\Delta S = \sigma_y \Delta S, \tag{32}$$

where

$$\sigma_y = \frac{d}{dS}(\log P_y) = \text{noise sensitivity at luminance level } P_y.$$

If the minimum perceptible Fechner fraction at a luminance level P_y is denoted by F_y, then the peak signal-$(S=1)$-to-noise ratio, $1/\Delta S$, necessary for the noise to be imperceptible is

$$\frac{1}{\Delta S} > \frac{\sigma_y}{F_y}. \tag{33}$$

If the Fechner fraction F_y were constant over the complete range of luminance involved, σ_y would be a direct measure of signal-to-noise ratio required to make the noise imperceptible at the luminance level for which σ_y is computed. F_y, however, increases at low levels of luminance. Despite this variation in F_y, σ_y is still a useful measure of noise sensitivity in comparing one system with another. Furthermore, the variation in F_y is still sufficiently small so that for ordinary display characteristics $(n<3)$, noise will be most perceptible in the dark regions of the displayed picture (at the minimum displayed luminance). The value of σ_y at the lowest level of presented luminance is, therefore, a measure of the required signal-to-noise ratio for threshold perception of noise.

For an n^{th} power-law receiver, we obtain from (15),

$$\sigma_n = \frac{dP_y}{P_y dS_y} = \frac{n(k^{1/n}-1)}{\left(\dfrac{P_y}{P_{\min}}\right)^{1/n}}. \tag{34}$$

The sensitivity of a linear receiver would be

$$\sigma_1 = \frac{k-1}{\left(\dfrac{P_y}{P_{\min}}\right)}. \tag{35}$$

The ratio of σ_1 to σ_n at the minimum luminance level $(P_y = P_{\min})$ is a measure of the improvement of the n^{th} law receiver over the linear receiver in threshold noise sensitivity. From (34) and (35), this ratio is found to be

$$\left.\frac{\sigma_1}{\sigma_n}\right]_{P=P_{\min}} = \frac{k-1}{n(k^{1/n}-1)}. \tag{36}$$

The improvement is 14 db for a square-law receiver and 18 db for a cubic receiver when the contrast ratio, k, is 80. For more complete results and discussion, the reader is referred to Oliver's article.[6]

In the color receiver, an increment in the co-ordinate S_r will produce an increment in the amount of red light, P_r, which is displayed. From (16), this is found to be

$$\Delta P_r = nP_{\min}(k^{1/n} - 1)\left(\frac{P_r}{P_{\min}}\right)^{(n-1)/n} \Delta S_r. \quad (37)$$

Similar expressions apply to the increments in green and blue light displayed corresponding to increments in the co-ordinates S_g and S_b, respectively. The increments in the R, G, B co-ordinates of the displayed picture P_r, P_g, P_b may be translated into corresponding increments in the X, Y, Z co-ordinates of the displayed picture. By using the relations between P_r and S_r, etc., and dividing through by the luminance, the following relation is arrived at:

$$\begin{bmatrix} \dfrac{\Delta P_x}{P_y} \\ \dfrac{\Delta P_y}{P_y} \\ \dfrac{\Delta P_z}{P_y} \end{bmatrix} = \sigma_n[D] \begin{bmatrix} \left(\dfrac{P_r}{P_y}\right)^{(n-1)/n} & 0 & 0 \\ 0 & \left(\dfrac{P_g}{P_y}\right)^{(n-1)/n} & 0 \\ 0 & 0 & \left(\dfrac{P_b}{P_y}\right)^{(n-1)/n} \end{bmatrix} \begin{bmatrix} \Delta S_r \\ \Delta S_g \\ \Delta S_b \end{bmatrix}, \quad (38)$$

where σ_n is the same quantity defined by (34) and discussed for the compatible receiver.

Since we restrict ourselves here to the noise which enters essentially in the transmission co-ordinates, it is desirable to relate the noise in the displayed image to noise in the U_1, U_2, U_3 co-ordinates. From (11) and (13), the following relation is obtained

$$\begin{bmatrix} \Delta S_r \\ \Delta S_g \\ \Delta S_b \end{bmatrix} = [F] \begin{bmatrix} \Delta A_1 \\ \Delta A_2 \\ \Delta A_3 \end{bmatrix} = [F] \begin{bmatrix} \beta_{11} & 0 & \beta_{13} \\ 0 & 1 & 0 \\ \beta_{31} & 0 & \beta_{33} \end{bmatrix} \begin{bmatrix} \Delta U_1 \\ \Delta U_2 \\ \Delta U_3 \end{bmatrix}, \quad (39)$$

which, when inserted in (38), gives

$$\begin{bmatrix} \dfrac{\Delta P_x}{P_y} \\ \dfrac{\Delta P_y}{P_y} \\ \dfrac{\Delta P_z}{P_y} \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} & h_{13} \\ h_{21} & h_{22} & h_{23} \\ h_{31} & h_{32} & h_{33} \end{bmatrix} \begin{bmatrix} \beta_{11} & 0 & \beta_{13} \\ 0 & 1 & 0 \\ \beta_{31} & 0 & \beta_{33} \end{bmatrix} \begin{bmatrix} \Delta U_1 \\ \Delta U_2 \\ \Delta U_3 \end{bmatrix}, \quad (40)$$

where the h values result from the matrix multiplication. "h_{11}," for example, is given by

$$h_{11} = d_{11}C_{11}\left(\frac{P_r}{P_y}\right)^{(n-1)/n} + d_{12}C_{21}\left(\frac{P_g}{P_y}\right)^{(n-1)/n} + d_{13}C_{31}\left(\frac{P_b}{P_y}\right)^{(n-1)/n}. \quad (41)$$

Since light intensities enter only as ratios in (41), h_{11} is a function of chromaticity only. Further, since there is no necessity for indicating that the color point is a displayed color point, we may drop the 'P' notation and obtain

$$h_{11} = \frac{d_{11}C_{11}R^{(n-1)/n} + d_{12}C_{21}G^{(n-1)/n} + d_{13}C_{31}B^{(n-1)/n}}{Y^{(n-1)/n}}. \quad (42)$$

The expressions for the other h values involve R, G, B, and Y similarly, but with different constants given in the Appendix.

The luminance interference sensitivity to noise in the three transmission channels is

$$\sigma_{yu_1} = \frac{\Delta P_y}{P_y \Delta u_1} = (h_{21}\beta_{11} + h_{23}\beta_{31})\sigma_n, \quad (43)$$

$$\sigma_{yu_2} = \frac{\Delta P_y}{P_y \Delta u_2} = h_{22}\sigma_n, \quad (44)$$

$$\sigma_{yu_3} = \frac{\Delta P_y}{P_y \Delta u_3} = (h_{21}\beta_{13} + h_{23}\beta_{33})\sigma_n. \quad (45)$$

When $U_1 = A_1$, $U_2 = A_2$, $U_3 = A_3$, these expressions reduce to

$$\sigma_{ya_1} = h_{21}\sigma_n, \quad (46)$$

$$\sigma_{ya_2} = h_{22}\sigma_n, \quad (47)$$

$$\sigma_{ya_3} = h_{23}\sigma_n. \quad (48)$$

h_{21} is a measure of the noise coupling from the A_1 chromaticity channel to the luminance channel. A value of 1 for h_{21} would indicate that noise in the A_1 channel is as perceptible (on the chromaticity where h_{21} is 1) as the same noise would be if applied to the luminance channel of the monochrome compatible receiver. Contours of constant h_{21}, for $n = 2$ are shown on the trichromatic diagram, Fig. 11.

Similarly, h_{23} is a measure of the noise coupling from the A_3 chromaticity channel to the luminance channel (Fig. 12).

h_{22} gives the effective luminance interference sensitivity to noise in the luminance channel. Contours of constant h_{22} are shown in Fig. 13 for $n = 2$. For about 95 per cent of the color triangle, h_{22} is between 0.9 and 1. It drops rapidly, however, in the immediate vicinity of the display primaries,— especially the blue primary.

The value of $n = 2$ was selected for the computations of the 'h' factors only to facilitate the calculations. The algebraic definitions of the h factors, however, show that h varies slowly with respect to n except in the immediate vicinity of the primaries. Both h_{21} and h_{23} are zero at the reference white but they vary differently as the saturation is increased in different directions from white. Although this makes a comparison difficult, it may be seen that h_{21} is of the order of 5 to 10 times greater than h_{23}.

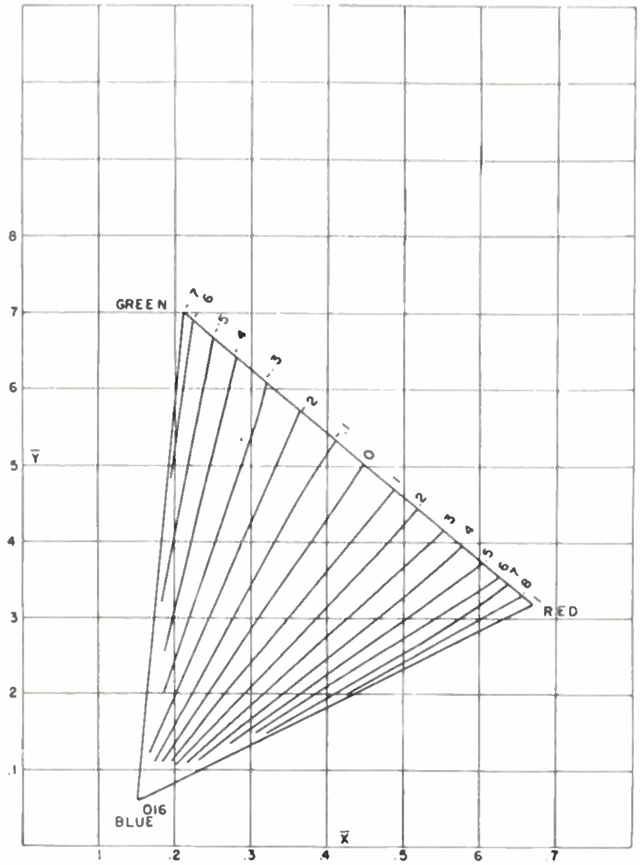

Fig. 11—Chromaticity factor h_{21} for luminance sensitivity to noise in the A_1 channel with gamma = 2.

Noise in the A_1 channel will, therefore, produce 5 to 10 times more luminance noise in the displayed picture than noise in the A_3 channel. Noise in the A_2 channel is more than 1.5 times as effective as noise in the A_1 channel.

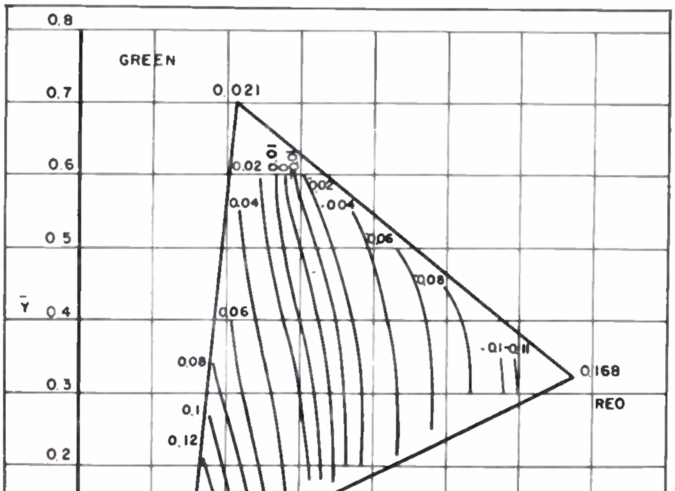

Fig. 12—Chromaticity factor h_{23} for luminance sensitivity to noise in the A_3 channel with gamma = 2.

Conclusions

Gamma correction at the transmitter in red, green and blue co-ordinates results in linear tone rendition on the compatible monochrome receiver. The luminance of saturated colors is attenuated somewhat on the monochrome receiver.

Although the peaks of demand upon the subcarrier and composite video signal are unaffected by gamma correction, the frequency of occurrence of these peak demands is reduced considerably by gamma correction.

The use of nonlinear display characteristics results in noise coupling from the chromaticity channel to the luminance channel. This coupling is zero on a white background and increases roughly with the color saturation.

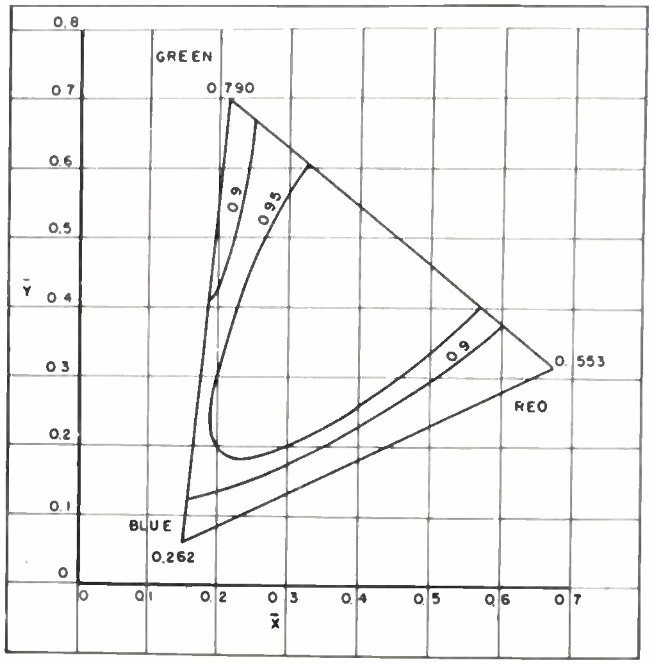

Fig. 13—Chromaticity factor h_{22} for luminance sensitivity to noise in the A_2 channel with gamma = 2 (see Appendix C).

Acknowledgment

The author is especially grateful to H. A. Samulon for his suggestions concerning the scope and presentation and for his encouragement throughout this investigation.

Thanks are also extended to the editorial readers of the IRE for their helpful comments on the presentation.

Appendix

A. For reference purposes, a list of the transformations involved in the co-ordinates recommended for field testing by the NTSC are listed below.

1. Chromaticities of receiver primaries.

$$\begin{bmatrix} (\text{red}) \\ (\text{green}) \\ (\text{blue}) \end{bmatrix} = \begin{bmatrix} 0.670 & 0.330 & 0.000 \\ 0.210 & 0.710 & 0.080 \\ 0.140 & 0.080 & 0.780 \end{bmatrix} \begin{bmatrix} (X) \\ (Y) \\ (Z) \end{bmatrix}$$

2. Transformations to normalized red, green, blue co-ordinates.

Unit reference white $=0.981(X)+(Y)+1.182(Z)$

$$\begin{bmatrix} R \\ G \\ B \end{bmatrix} = \begin{bmatrix} 1.910 & -0.532 & -0.288 \\ -0.985 & 1.999 & -0.028 \\ 0.058 & -0.118 & 0.898 \end{bmatrix} \begin{bmatrix} X \\ Y \\ Z \end{bmatrix}$$

$$\begin{bmatrix} X \\ Y \\ Z \end{bmatrix} = \begin{bmatrix} 0.607 & 0.174 & 0.200 \\ 0.299 & 0.587 & 0.114 \\ 0.000 & 0.066 & 1.116 \end{bmatrix} \begin{bmatrix} R \\ G \\ B \end{bmatrix}$$

3. Transmission co-ordinates.

$$\begin{bmatrix} A_1 \\ A_2 \\ A_3 \end{bmatrix} = \begin{bmatrix} 1 & -0.981 & 0 \\ 0 & 1 & 0 \\ 0 & -1.182 & 1 \end{bmatrix} \begin{bmatrix} X \\ Y \\ Z \end{bmatrix}$$

$$\begin{bmatrix} U_1 \\ U_2 \\ U_3 \end{bmatrix} = \begin{bmatrix} 0.028 & 0 & 0.442 \\ 0 & 1 & 0 \\ -1.675 & 0 & 0.253 \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \\ A_3 \end{bmatrix}$$

B. "h" coefficients of (40).

$$\begin{aligned}
h_{11} &= d_{11}C_{11}\lambda_1 + d_{12}C_{21}\lambda_2 + d_{13}C_{31}\lambda_3 \\
h_{12} &= d_{11}\lambda_1 + d_{12}\lambda_2 + d_{13}\lambda_3 \\
h_{13} &= d_{11}C_{13}\lambda_1 + d_{12}C_{23}\lambda_2 + d_{13}C_{33}\lambda_3 \\
h_{21} &= d_{21}C_{11}\lambda_1 + d_{22}C_{21}\lambda_2 + d_{23}C_{31}\lambda_3 \\
h_{22} &= d_{21}\lambda_1 + d_{22}\lambda_2 + d_{23}\lambda_3 \\
h_{23} &= d_{21}C_{13}\lambda_1 + d_{22}C_{23}\lambda_2 + d_{23}C_{33}\lambda_3 \\
h_{31} &= d_{31}C_{11}\lambda_1 + d_{32}C_{21}\lambda_2 + d_{33}C_{31}\lambda_3 \\
h_{32} &= d_{31}\lambda_1 + d_{32}\lambda_2 + d_{33}\lambda_3 \\
h_{33} &= d_{31}C_{13}\lambda_1 + d_{32}C_{23}\lambda_2 + d_{33}C_{33}\lambda_3
\end{aligned}$$

where

$$\lambda_1 = \left(\frac{R}{Y}\right)^{(n-1)/n} \quad \lambda_2 = \left(\frac{G}{Y}\right)^{(n-1)/n} \quad \lambda_3 = \left(\frac{B}{Y}\right)^{(n-1)/n}.$$

C. The calculations for Figs. 4, 5, 11, 12, and 13 are based upon the primaries shown in the figures with normalization on an equal energy white rather than illuminant C. For the NTSC recommended co-ordinates, the results would differ slightly.

Principles and Applications of Converters for High-Frequency Measurements*

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Summary—**The heterodyne method permits measurements over wide frequency bands with the standards operating at a fixed frequency. The accuracy of such measurements depends upon the performance of heterodyne conversion transducers or converters. Design principles are derived to maximize linearity and dynamic range and minimize zero corrections. These principles have been applied successfully to point-by-point and sweep measurements of delay, phase, transmission, and impedance.**

Introduction

In the measurement of complex electrical quantities at low frequencies, sufficiently pure resistances and reactances can be realized for use in fixed and variable standards in bridges and other measurement equipment. As measurements must be performed at higher and higher frequencies, it becomes increasingly difficult to realize standards which have constant amplitude and constant phase properties over sufficiently broad bands of frequency. Elaborate calibrations become necessary and often, even then, calibration corrections can only be determined with insufficient accuracy.

A successful solution to these problems is the heterodyne method. In this the electrical signal under study is modulated in a frequency converter with a heterodyning frequency to obtain as useful output a signal whose frequency is the difference between the original signal frequency and the heterodyne frequency, and which may be a suitable fixed intermediate frequency. If this converter can be made to transduce linearly the amplitude and phase relationship from the original signal frequency to the fixed intermediate frequency, variable standards may be employed in the fixed intermediate frequency to perform the measurement, thus simplifying the problem of standards design. However, the ability of the converter to act as a linear transducer places the dominant constraint on the final accuracy obtainable.

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We have now a measurement system in which we have introduced an active element into the measurement, and by doing so we have concentrated the accuracy problem in the realization of a linear conversion transducer. Realization of such a transducer enables us to measure with extreme precision any electrical quantity which can be described by an amplitude and phase angle, such as gain, loss, phase shift, delay, reflection, and vector impedance.

The advent of transcontinental coaxial-cable and microwave radio-relay television transmission systems demanded unprecedented accuracy of measurement at high frequencies over many octaves and wide variations in signal voltage. This stimulated investigation into methods to obtain converters which would operate over a wide range of input voltages and wide ranges of frequency with extreme accuracy. As a typical example, a basic phase and transmission measuring set[1,2,3] is shown in Fig. 1. In the "S" converter, the reference signal is

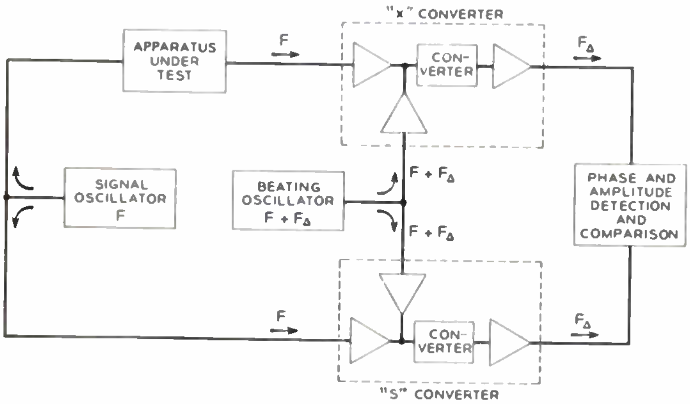

Fig. 1—Basic phase and transmission-measuring circuit.

mixed with the heterodyne frequency to obtain at the IF point a signal of reference phase and amplitude. The "X" converter receives its input signal after it has passed the apparatus under test, and its phase and amplitude is translated to the fixed intermediate frequency again by beating it with the heterodyning frequency. The reference and test signals at the fixed intermediate frequency are then compared by suitable detection devices operating at the fixed intermediate frequency.

The converters must meet two requirements:

(1) A linear relationship must exist between input and output over the range of levels encountered.

(2) Sufficient isolation must exist between the "X" and "S" converters to prevent errors due to cross talk and pickup.

To facilitate the measurement as the measuring frequency is changed, another design aim is desirable:

(3) The phase shift and transmission characteristics of the "X" and "S" converters should be identical as far as possible. Thus when the reference zero conditions at the converter inputs are identical, no corrections have to be made at the converter outputs for differentials between "X" and "S" converter characteristics as the measuring frequency is changed.

The three requirements mentioned will be discussed separately and then in their relationship.

I. Converter Linearity and Dynamic Range

Dynamic range of a converter may be defined as the range of input levels over which a given linearity limit may be met. The lower limit is caused by the presence of noise. The upper limit is determined by the distortion due to overload. Electron-tube converters of the curvature type are considered primarily and are discussed specifically here.

The linear type of converter (linear rectifier) is not discussed here. To achieve sufficient linearity for precise measurement purposes, expedients such as feedback have to be used in practice. Adequate feedback was unrealizable over the wide frequency ranges primarily considered here.

Noise

The smallest input signal which can be utilized by the converter is reached when the effective converter noise is equal to the maximum tolerable departure from linearity. Thus, for example, for a linearity tolerance of 0.01 db, an effective signal-to-noise ratio of 60 db is required for the smallest possible input signal.

With the low impedances encountered in high-frequency wide-band circuits, Johnson noise is usually negligible compared to shot noise.

The shot noise of an electron tube resides in the space current. With sufficient approximation it may be considered as a continuous noise spectrum of equal energy content in any equal-width frequency band. For the purpose of this discussion, we assume that a definite bandwidth has been assigned to the IF detection circuit the choice of which depends upon such factors as speed of response of the detection circuits, low-frequency noise limiting the benefits of further narrowing of bandwidth, and phase and amplitude stability of the band limiting means.

It has been found convenient for computation purposes to refer the noise current to the input grid expressed as the noise of an equivalent resistance noise generator.[4,5] For our purposes we use the equivalent noise resistance computed for amplifier operation (R_n).

[1] D. A. Alsberg and D. Leed, "A precise direct reading phase and transmission measuring system for video frequencies," *Bell Sys. Tech. Jour.*, vol. 28, pp. 221–238; April, 1949.

[2] J. G. Kreer, L. A. Ware, and E. Peterson, "Regeneration theory and experiment," *Bell Sys. Tech. Jour.*, vol. 13, pp. 680–700; October, 1934; and Proc. I.R.E., vol. 22, pp. 1191–1211; October, 1934.

[3] M. Levy, "Methods and apparatus for measuring phase distortion," *Elec. Commun.* (London), vol. 18, pp. 206–228; January, 1940.

[4] W. A. Harris, "Fluctuations in vacuum tube amplifiers and input systems," *RCA Rev.*, vol. 5, pp. 505–524; April, 1941.

[5] W. J. Stolze, "Input circuit noise calculations for FM and television receivers," *Communications*, vol. 27, p. 12 ff.; February, 1947.

The equivalent noise resistance of a triode is[4]

$$R_n = \frac{2.5}{g_m}, \tag{1}$$

where g_m is the actually used mutual transconductance.

The equivalent grid input noise voltage due to shot noise is

$$e_n = \sqrt{1.6 \times 10^{-20} \Delta f\, R_n}, \tag{2}$$

where Δf is the noise bandwidth.

A signal output current derived from the modulation process is attenuated with respect to a signal output current arising from simple amplification by the ratio of the mutual transconductance g_m to the effective conversion transconductance $g_{c_{eff}}$ of the tube, which is a function not only of the transconductance slope, but also a function of the driving voltages.[6]

Hence, we may write the equivalent converter noise resistance

$$R_c = R_n\left(\frac{g_m}{g_{c_{eff}}}\right)^2. \tag{3}$$

We define as modulation loss in db

$$\eta_m = 20 \log \frac{g_m}{g_{c_{eff}}}. \tag{4}$$

Consequently, to obtain the lowest noise for a given tube, we must strive to operate it with the minimum modulation loss compatible with overload restrictions (to be discussed later). It has been shown[6] that in the ideal case of the parabolic converter, the minimum modulation loss which can be obtained for maximum overload limited output current is

$$\eta_m = 12 \text{ db}. \tag{5}$$

Overload Effects

By beating two frequencies F and $F+F_\Delta$ together in a nonlinear device, we obtain as useful output the difference frequency F_Δ. F_Δ is produced by all even-order modulation products.[6] Expansion of the coefficients of the modulation products further shows that only F_Δ due to the second-order product (square law) is linearly related to the change in amplitude of one of the inputs when the other remains fixed, provided that all effects on the plate current, other than variations in the external voltage applied to the control grid, are negligible. As the coefficients of the fourth and higher even-order products are not linear functions of the amplitudes of F and $F+F_\Delta$, their presence causes a nonlinearity of the frequency converter and is a primary source of distortion. It should be noted that all odd-order products do not contain any terms in F_Δ, and thus do not contribute any distortion in absence of other effects.

[6] T. Slonczewski, "Transconductance as a criterion of electron tube performance," *Bell Sys. Tech. Jour.*, vol. 28, pp. 315–328; April, 1949.

The range of input voltages Δe, over which an electron-tube mutual transconductance characteristic is linear, determines the maximum signal which may be applied without causing overload.[6] However, no vacuum tubes are parabolic down to the virtual cutoff E_0 (Figs. 2 and 3). This prevents attainment of the minimum modulation loss $\eta_m = 12$ db, which requires swinging the grid to the virtual cutoff point E_0.

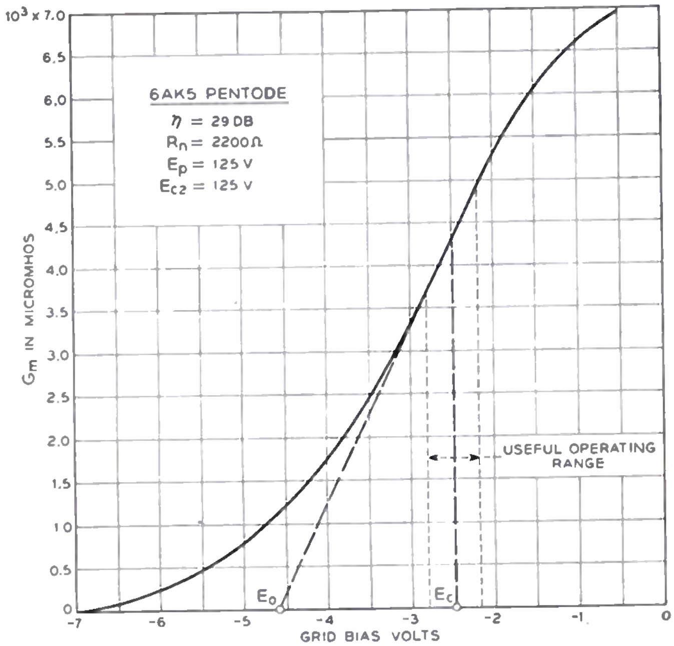

Fig. 2—6AK5 characteristics.

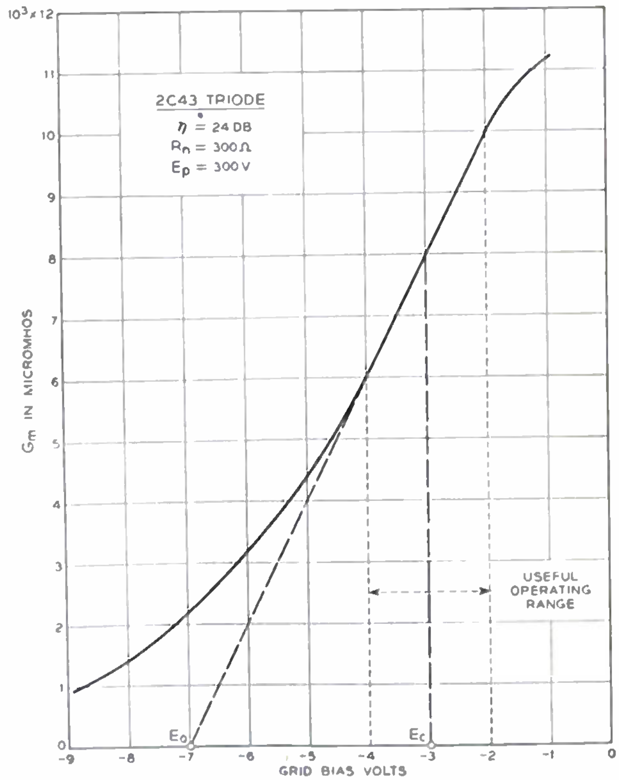

Fig. 3—2C43 characteristics.

Unfortunately, tubes are not specifically designed to achieve parabolic characteristics. As a result, the designer must choose among available tubes those which come closest to his requirements. While he may select likely candidates from inspection of transconductance characteristics, the final determination must be made on the experimental basis to locate fine structure varitions which can only be found by actual measurement.

In practice, distortion increases rapidly outside the well-defined linear transconductance range. Thus, with the maximum usable input voltage range Δe established, maximum undistorted output current is obtained when both input signals, F and $F+F_\Delta$, are of equal magnitude, having a peak value of $\Delta e/4$ each; hence,

$$i_{F\Delta\max} = \frac{\Delta e}{4} g_{c\,\mathrm{eff}}. \tag{6}$$

Dynamic Range

From (2), (3), and (6) we can write for the signal-to-noise ratio of the overload limited input signal

$$\frac{S}{N} = 5 \times 10^9 \frac{(g_{c\,\mathrm{eff}}/g_m)\Delta e}{\sqrt{\Delta f R_n}}. \tag{7}$$

In (7) only $g_{c\,\mathrm{eff}}/g_m$, Δe, and R_n are controlled by the electron-tube characteristics. Thus, a figure of merit or dynamic range factor may be defined from (7) as

$$D_r = \frac{(g_{c\,\mathrm{eff}}/g_m)\Delta e}{\sqrt{R_n}}. \tag{8}$$

For the specific case of the triode, substituting (1) into (8),

$$D_r = 0.63(g_{c\,\mathrm{eff}}/g_m)\sqrt{g_m}\,\Delta e. \tag{9}$$

Equations (8) and (9) show clearly the importance of minimum modulation loss, and a large Δe. Paralleling of tubes increases D_r only proportionally to the square root of increase in g_m.

From the foregoing, the following factors must be balanced against each other to obtain maximum dynamic range:

(a) Overload capacity (Δe).
(b) Inherent internal noise (R_n).
(c) Modulation loss (η_m).

In combination, these factors point in the direction of tubes capable of drawing large plate currents.

Preamplifier Noise Contribution

When buffer amplifiers, to be discussed later, precede the converters, they often are broad-band devices which are capable of transmitting the difference F_Δ as well as the desired signal. The noise spectrum of same bandwidth centered on the difference frequency F_Δ contains the same amount of energy as the one centered on F or $F+F_\Delta$. The noise band centered on F_Δ passes through the converter by the process of amplification; the noise band centered on F or $F+F_\Delta$ passes through the converter only by the process of frequency conversion. Thus, resulting signal-to-noise ratio is again degraded by ratio η_m of amplification to conversion gain.

This effect is minimized in the case of a converter balanced for both F and $F+F_\Delta$. The noise band F_Δ is suppressed the same way the signal frequencies F or $F+F_\Delta$ are suppressed. A band elimination filter, eliminating the noise-band centered on F_Δ ahead of the converter, also eliminates this effect. Bandwidth considerations to be discussed later may make it inadvisable to use a balanced converter or to construct a noise-band elimination filter. Then a considerable premium exists in obtaining a tube with minimum ratio of amplification to conversion gain, to minimize the effect of the buffer noise contribution. Usually, tubes suitable for service as buffer or preamplifiers have no lower internal noise current than the converter tube. When it becomes necessary that low levels be measured (and thus that gain ahead of the converter is desired to minimize the effects of the converter noise contribution), no benefit is derived unless F_Δ noise-band contribution of preamplifier is eliminated by filtering or balanced modulation.

Loading Effects

In a triode converter the output voltages appearing in the plate circuit can cause a loading effect by remodulation.[7] This secondary modulation is equivalent to an error signal, hence constituting a source of nonlinearity. The output voltages can be held below their critical value if the load impedance of the triode is made small. Thus independence from plate remodulation is achieved at the cost of low voltage output. The plate-load impedance must be designed to:

(a) prevent the desired signal F_Δ due to primary modulation from remodulating in the tube.
(b) prevent F and $F+F_\Delta$ in the plate-output circuit to produce by modulation F_Δ following a modulation law other than square, and differing in phase from F_Δ due to primary modulation.

Occasionally an additional loading effect is of practical importance. One of the products of square-law modulation is a dc component which is the useful output in case of a square-law rectifier. If cathode bias is used in the converter tube, this dc current produces changes in bias dependent upon the signal, hence introducing nonlinearity. When this cathode reaction must be reduced, a number of techniques are available, such as bleeding the cathode resistor, electronic decoupling by use of a cathode coupled amplifier, and so on. Inherently, any of these methods simultaneously reduce the dc feedback which compensates for plate-current drifts, thus eliminating the main adantages of cathode bias. Fixed bias operation then often becomes more attractive. This sometimes results in more economical circuitry and eliminates the difficult problem of satisfactory signal-frequency by-passing in case of converters which operate over a large number of octaves.

[7] The term "remodulation" refers to unwanted modulation caused by the nonlinearity of the plate current—plate-voltage characteristic.

The discussion has emphasized triode rather than pentode or tetrode converters, even though the latter have found applications as parabolic converters because they permit development of sizable output voltages without those output voltages causing nonlinearity by remodulation in the plate output circuit.

However, pentodes and also tetrodes have two fundamental weaknesses. Caused by the screen current, their internal noise current is substantially higher than the noise current in comparable triodes of the same transconductance.[4,5] Furthermore, pentodes and tetrodes exhibit "anomalous modulation"[8] for low signal voltages, which causes nonlinearity for signals approaching the noise limit of the converter. This effect is absent in triodes.

In addition, in some triodes the usable parabolic range is larger and extends closer to the grid cutoff E_0, thus permitting higher inputs and a lower modulation loss η_m.

The isolation advantages of pentodes can be combined with the noise and anomalous modulation advantages of triodes by the use of the "cascode"[9] at a small sacrifice in the noise figure. As the cascode circuit requires two tubes, the second tube can as well be applied to regain the output level sacrificed in a single-triode converter.

The superiority of the triode converter was experimentally verified during the development of measuring apparatus for coaxial carrier and radio-relay television transmission systems, where numerous tubes were investigated for suitability as converters. In the range from 50 kc to 80 mc for linearities of 0.01 db, dynamic ranges exceeding 30 to 35 db were found unattainable with pentodes. With the 2C43 lighthouse triode, dynamic ranges from 50 to 60 db for linearities of 0.01 db have been attained.

II. Isolation of "X" and "S" Converters

Assume that $F+F_\Delta$ is the heterodyning frequency and F the signal frequency. An error in measurement results if F can reach either converter by any stray path. One primary stray path by which F can reach the "X" converter from the "S" converter, and vice versa, exists through the common connection furnishing $F+F_\Delta$ to both "X" and "S" converters. The transmission path for $F+F_\Delta$ voltage is given unidirectional properties by buffer amplifiers, which have the necessary attenuation in the reverse transmission direction. To achieve this, careful attention to shielding and grounding problems is required, particularly at higher frequencies. Pentodes or tetrodes rather than triodes are usually indicated for buffer service, the shielding action of the screen grid reducing the coupling between the plate and the control grid.

[8] Anomalous modulation exists when the amplitudes of the modulation products fail to decrease as predicted by the Taylor Series expansion[6] when the input amplitudes are decreased. This may be caused by fine-grain structure variations of the transconductance characteristic due to screen structure effects and secondary emission phenomena.

[9] H. Wallman, A. B. Macnee, and C. P. Gadsden, "Low noise amplifier," Proc. I.R.E., vol. 36, pp. 700–708; June, 1948.

III. Bandwidth and Tracking Considerations

A complete converter may consist of high-frequency buffer amplifiers, a converter, and fixed intermediate frequency amplifier stages (Fig. 1). As measurement is made, it is inconvenient to establish a new reference zero for the measuring set every time the measuring frefiuency is changed. A considerable premium exists to minimize this time-consuming adjustment. In sweep-type measurements the degree of suppression of the zero characteristic is a controlling factor. If we make both "X" and "S" converters identical as to phase and transmission characteristic, no differential phase and transmission change would take place as the measuring frequency is changed and no change in system zero would occur. It is not necessary that both converters be flat in phase and transmission; it is adequate if they are identical.

In a specific case, the realization of the isolation requirements required a single-stage buffer facing the signal frequency F and a two-stage buffer facing heterodyning frequency $F+F_\Delta$. $F+F_\Delta$ and F were combined in a plate-load resistor common to both the F buffer and the second stage of the $F+F_\Delta$ buffer.

In a wide-band amplifier the high-frequency cutoff is determined by the stray shunt capacities present and the low-frequency cutoff by the plate, screen, and cathode by-passing impedances. Stray-capacitance compensation is usually done by shunt peaking, or series peaking interstage design. In order to obtain a reasonably simple and uncritical adjustment, an interstage as simple as possible is preferable. To achieve duplication of phase and transmission characteristics between the two converters to the order of 0.01 db and 0.1°, a practical compromise between phase and transmission requirements is to drop the plate-load resistance until the top frequency of operation of the buffer interstages is no higher than 0.6 f_c, where f_c is the uncompensated cutoff of the interstage (shunt reactance equal to plate-load resistance). This represents only a relatively small sacrifice of the bandwidth which might be obtained considering either phase or transmission requirements alone.

With a given tube, the minimum realizable shunt capacitance of the interstage is fairly fixed, and with the top frequency of operation, the maximum plate load impedance which may be used is then fixed. Considering bandwidth, noise, and overload, a buffer-stage gain in the vicinity of unity usually will give optimum dynamic range; hence high figure of merit tubes are indicated for wide-band high-frequency operation. The limit of benefit from a high-transconductance tube is reached when the plate-load resistance drops so low that the residual stray wiring inductances become larger than the shunt capacitance compensating inductances required.

Wide-band by-passing of plate, screen, and cathode circuits requires special attention. To avoid uncon-

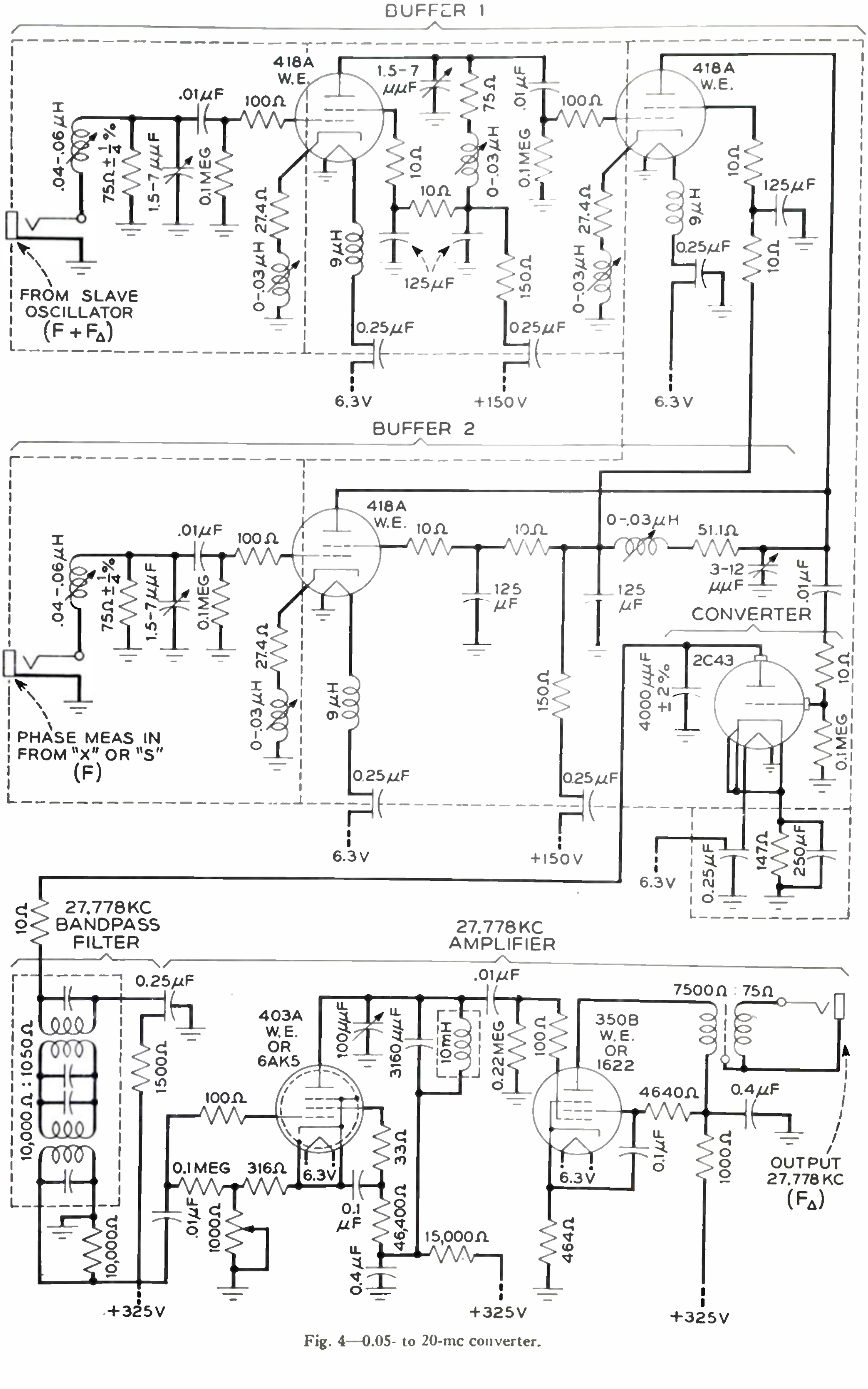

Fig. 4—0.05- to 20-mc converter.

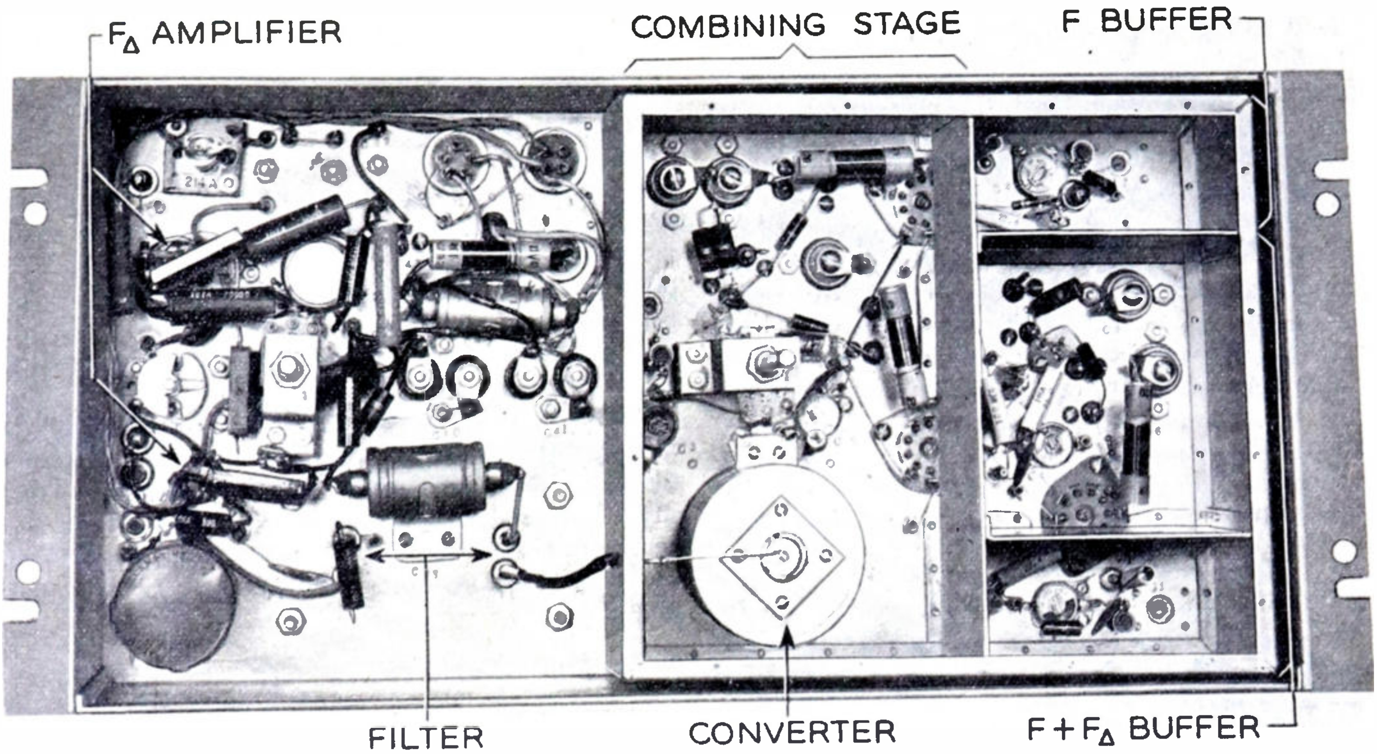

Fig. 5—Converter, front view.

trolled transmission irregularities, in-band parasitic resonances must be avoided. This is most easily done by choosing a by-pass condenser whose capacitance and internal inductance resonate below the transmission band. Such a condenser will then have inductive reactance for in-band frequencies and can be absorbed as part of a shunt-peaking inductance. Low-inductance electrolytic condensers have proven very satisfactory for this purpose.

The adjustment of the interstage networks to align the "X" and "S" converters with respect to each other is relatively simple. By choosing a top frequency of less than $0.6 f_c$, enough margin is provided so that the phase slope of the interstage may be varied by adjusting the interstage reactances, and a phase characteristic linear with frequency may still be maintained. The interstage between the two $F+F_\Delta$ buffers mentioned previously contains $F+F_\Delta$ only; thus linear phase differentials between the two modulators can be mopped up at this point. The combining stage of F and $F+F_\Delta$ buffers contains both F and $F+F_\Delta$. Thus if the phase slope of this interstage varies, only the phase change over the interval F_Δ effects a net change in phase relation between the two converters. As F_Δ is normally small, the net phase effect of this interstage is usually negligible, and this stage can then be used to mop up transmission differentials without affecting the phase differential.

IV. The Actual Converter Design

To illustrate the application of the principles outlined, some of the pertinent design details of converters operating in the range of 0.05 to 20 mc are given. The schematic is shown in Fig. 4, photographs in Figs. 5 and 6. They are part of a measuring set which is a further development of one described previously.[1,10] The set is

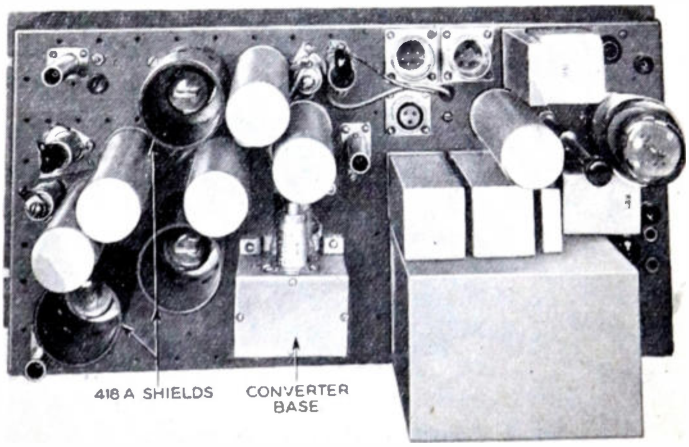

Fig. 6—Converter, rear view.

used for gain, loss, phase, envelope delay, reflection coefficient, and vector impedance[11] measurements. In the most recent development using a similar approach to the converter problem, the top frequency of operation of this measuring set was extended to 80 mc.

[10] "1950 engineering developments," *Elec. Eng.*, vol. 70, p. 22; January, 1951.

[11] D. A. Alsberg, "A precise sweep frequency method of vector impedance measurement," Proc. I.R.E., vol. 39, pp. 1393–1400; November, 1951.

1. The Converter

The 2C43 lighthouse triode was found to be the best tube available for yielding maximum dynamic range and bandwidth. To prevent secondary modulation in the plate circuit, the external plate impedance for the difference frequency F_Δ of 27.778 kc was chosen as 1,050 ohms. The impedance at 50 kc, at which the output filter has 55-db discrimination, already is dropped to 125 ohms, with the impedance dropping further until parasitic inductances raise the impedance again. This rise in plate impedance at increasing frequency is reflected into the converter input circuit by the Miller effect and must be kept small to prevent detuning of the input circuit. To accomplish this, the tube plate conductor and a 4,000-$\mu\mu$f by-pass condenser were formed into a coaxial structure providing a low-inductance path for the high-frequency (Fig. 7). The 4,000-$\mu\mu$f by-pass was

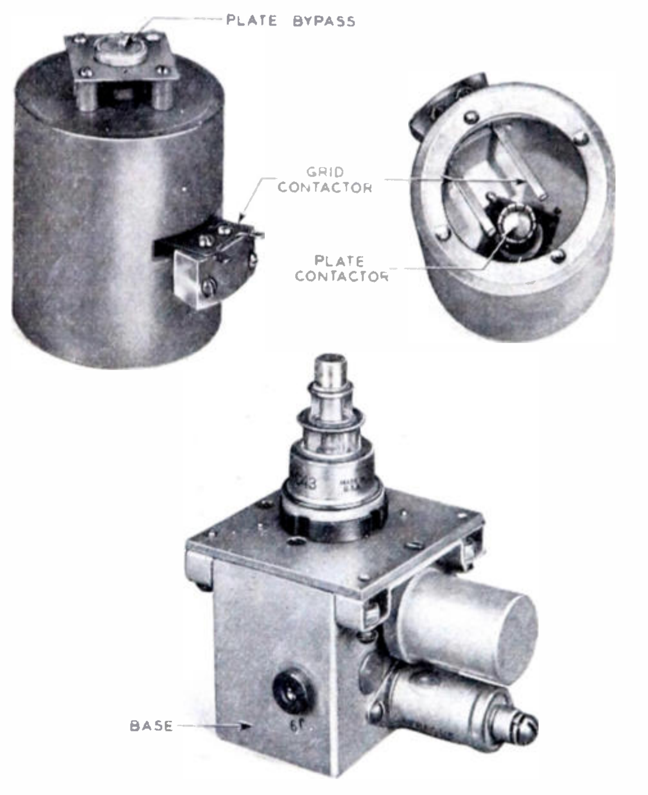

Fig. 7—Converter tube-mounting details.

made part of the input tuning capacity of the 27.778-kc band-pass filter. However, a parasitic resonance then occurred at about 13 mc where the 4,000-$\mu\mu$f condenser went into a parasitic resonance with the inductance of the connecting lead to the filter and the residual inductance of the filter. By introducing 10ω dissipation in the connecting lead, the effect of this resonance was minimized. A similar resonance could be shown to exist at about 500 mc where the tube capacitance resonated with the by-pass inductance.

The cathode biasing was accomplished by a cathode-biasing resistor, by-passed by a 250-μf electrolytic low-inductance condenser. As the frequency of operation is increased beyond 20 mc, the unavoidable cathode inductance to ground introduces enough cathode feedback of the same order as encountered in cathode input grounded-grid operation, opposing the modulation process, to make cathode input grounded-grid operation more attractive, as at the same time the reaction of the plate on the input circuit would be reduced. From experimentation conducted up to 80 mc, it is estimated that grounded-grid converter operation may be superior at frequencies above 100 mc.

Balanced converter operation was not considered because of the difficulty of obtaining balanced inputs through either transformers or phase inverters which could be reproduced to sufficient accuracy in both converters to permit tracking of the frequency characteristics of the two converters to 0.01 db and 0.1° from 0.05 to 20 mc.

2. The Buffer Amplifiers

In order to meet the tracking requirement, the buffer top frequency should be 0.6 f_c. For 20-mc top frequency, the minimum cutoff then is 33.3 mc. Several tubes were tried which met this requirement. They failed, however, to meet 0.01-db linearity and stability requirements. It was noted that tubes such as the 404A would show changes in transmission of the order of 0.01 db after removing and reapplying a 0.25-volt signal. The 418A tetrode (G_m = 25,000 μmhos) exhibited none of these defects, and was used even though its cutoff frequency was far in excess of the requirements. A small amount of cathode feedback (4.5 db) was introduced by leaving the cathode biasing resistors unby-passed. This also avoided the difficulty of satisfactory cathode by-passing. For plate and screen by-passes 125-μf low-inductance electrolytic condensers were used. Paper condensers exhibited spurious resonances within the transmitted band, and by-pass type condensers of adequately large capacitance to satisfy phase requirements at the low end of the band were not available in time to be used in the design.

The wiring of the interstages could not be sufficiently controlled to duplicate wiring inductances in a pair of converters, in particular as the wiring inductances were slightly in excess of the required interstage capacitance compensation. To control the phase characteristic adequately, it was found necessary to adjust both cathode and plate circuit inductances. A special hairpin type variable inductor was designed which had a minimum inductance of less than 0.01 μh and a maximum inductance of about 0.03 μh (Fig. 8).

A 120-db isolation requirement was met by choosing the circuit constants such as to reduce the capacitance coupling through the vacuum tube to produce more than 60-db isolation per stage. Common-ground impedances

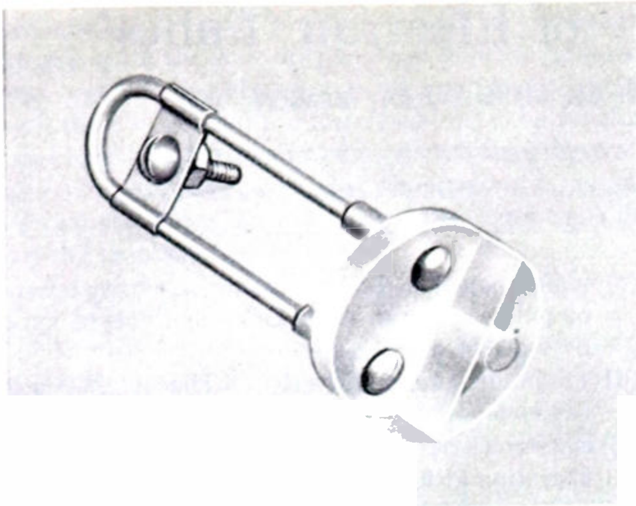

Fig. 8—Hairpin variable inductor.

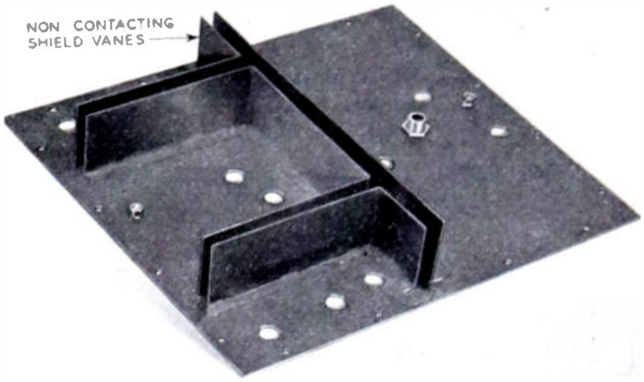

Fig. 9—Shield cover.

were minimized by choosing a solid-copper ground plane.

Electromagnetic coupling was controlled by controlling all shield junctions and cracks. A crack in the shielding can be considered as a waveguide with the lowest mode of transmission controlled by the largest width of the crack. All cracks encountered are substantially shorter than the shortest wavelength used (15 meters); it is thus a waveguide operating below cutoff, which has a constant attenuation of 56.8 db for a length equal to its largest width. Lips were added to all shield junctions to assure adequate attenuation. One single lid covers all shield compartments to facilitate access for maintenance. Copper vanes are added to the cover to form between compartments a labyrinth which acts as an attenuator, preventing coupling between compartments (Fig. 9).

It was found that the 418A electron tubes which were used as buffer amplifiers radiated an electromagnetic field which existed in a volume bounded by the ground plane and a plane ¾ inch distant from the ground plane. A blackened solid copper shield surrounds the buffer tube with none of the ventilation holes penetrating into the radiated field, thus eliminating coupling of the buffers through this field. This shield meets both the electrical shielding and thermal dissipation requirements of the buffer.

The net noise contribution of the 3 buffer stages effective at the converter input was equivalent to a resistance noise of 750 ohms. This compares with a resistance noise R_n of only 300 ohms residing in the 2C43 converter. The noise contribution of the buffer could be eliminated by a noise-rejection filter, with a resulting increase of dynamic range of about 5 db. However, it was impractical to design a F_Δ (27.778-kc) noise-rejection filter with no spurious resonances from 50 to 20,000 kc (9 octaves) and the noise limit set by the 418A tubes was accepted.

Conclusion

The introduction of frequency converters as a critical transmission element into the active part of a measuring device for complex electrical quantities provides high accuracies operable over wide bands of frequency and aids substantially in sweep-type measurements. To obtain optimum results, careful attention to noise, overload, and bandwidth effects is required. It has been shown that the dynamic range is increased by obtaining low modulation loss, high input capability, and high transconductance in the converter tube. Application of the design principles discussed has resulted in converters operating in the frequency range of 0.05 to 80 mc, linear to 0.01 db and accurate to 0.1° over dynamic ranges of as much as 60 db.

Concerning the Reliability of Electron Tubes*

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This paper appears with the approval of the IRE Professional Group on Quality Control. —The Editor.

Introduction

THE MAJOR PORTION of electron-tube research and development supported by tube manufacturers has long been directed toward mass producibility of low-cost tube types for which a large visible market exists. Since the broadcast-receiver industry has provided the most extensive market, quality of normal tube production has been maintained at a level commensurate with economic factors prevailing in this industry. It has frequently been found impractical to adopt suggested modifications of design or materials which might increase tube life or reduce failures in service because such changes would result in considerably higher costs or lower production output.

In recent years, however, there has been a growing demand for tubes with improved reliability for widely diverse usages in both military and industrial equipment. Unfortunately, two serious obstacles have impeded the development of such tubes. In the first place, the market has been so small as to furnish little incentive to manufacturers for investing the additional millions of dollars required for engineering, equipment design, and production machinery. This problem has been mitigated in some cases by the sponsoring of tube-development programs by the government or, less frequently, by the equipment designer. It may soon become much less severe if the market for reliable tubes continues to exhibit the rapid growth evidenced within recent months.

Defining "Reliability"

Another deterrent to the development of such tubes consisted in the difficulty of defining the objective of "reliability." Although customer specifications for reliable tubes may be identical, interpretations of reliability vary as widely as the applications for which the tubes are intended. The general term "reliability" has been used at various times to denote such qualities as

(a) an unusually low rate of sudden inoperable failures (as distinguished from gradual or marginal failures which may be detected and controlled) in normal operation for a comparatively short period;
(b) continuous stability of performance, with a minimum rate of deterioration of electrical characteristics over a specified period;
(c) maintenance of balance between characteristics of pairs of tubes, or sections of duotype tubes, over a specified period;
(d) minimum amount of drift of characteristics between operating cycles in on-off applications;
(e) characteristics stability and/or low failure rate in operation at high ambient temperature;
(f) resistance to high levels of impact shock;
(g) resistance to mechanical vibration for sustained periods.

The (a) type of reliability is generally desired for usage in expendable equipment where the expected unit-life is comparatively brief. An extreme case of this emphasis on lack of inoperable failures for a short time is encountered in guided-missile applications, where the demand is for zero failures over a total operating time which may often be less than one hour, and is rarely more than ten hours. Airborne electronic equipment offers an example of a case where tubes are required to give unusually dependable service for an intermediate length of time, and with the additional factors of adverse conditions of shock, vibration fatigue, and/or comparatively high ambient temperature (thus adding items (b), (e), (f), and (g) from the list above). Uniform effectiveness of control circuits often depends upon the maintenance of balance between the characteristics of pairs of tubes, or of sections of duotype tubes, discussed in (c), which is a special condition of the performance stability of (b). The flip-flop circuits of electronic computers, with their intermittent cycling of voltages applied to tubes, illustrate the necessity for the (d) type of reliability.

Reliability and "Long-Life"

A previous paper published on the subject of evaluating the life expectancy of premium subminiature electron tubes[1] embodied the familiar concept that "long life" (in the order of 5,000 hours) would attain the multiplicity of objectives implied by the term "reliability." This theory was based primarily upon a consideration of the occurrence of failures in service. It is apparent that when tubes are designed to give satisfactory service for several thousand hours, some precautions must also be taken to insure that inoperable failures will be held to a minimum. Obviously, the resultant high mechanical quality of tube structures will contribute in some degree to tube reliability in most applications. Conversely, any

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[1] E. M. McElwee, "Statistical evaluation of life expectancy of vacuum tubes designed for long-life operation," *Sylv. Tech.*, vol III, no. 2; April, 1950.

nprovements effected to satisfy specific requirements such as impact shock, vibration, high temperature) will lmost certainly benefit normal life operation. To this xtent, there is a definite interrelation between reliabil- :y and long life. It remains to be shown that there are ther factors to be considered in the attempt to satisfy hese two objectives, and that the interdependence be- ween the two is not all-inclusive.

Statistical Evaluation of "Life"

Early evaluations of long-life quality of electron tubes vere made in accordance with the JAN-1A rating of ercentage of total possible tube hours.[2] This method of :valuation was generally acceptable to the industry for several reasons.

(1) It had been standardized by the JAN committee, was widely known and understood, and was in general use.

(2) It served adequately as a basis of comparison for relative life quality of different tube types or tube lots.

(3) It included all types of failures ordinarily considered to terminate tube life—shorts, open elements, air tubes, or any variation of characteristics beyond life-test end-point limits.

However, more recent investigations have revealed that this type of life evaluation is not sufficient for the majority of applications for reliable tubes since it provides no information concerning early life failures. The assurance that a group of tubes will amass a certain minimum number of life hours within a specified period is of little value to the designer of expendable equipment, who wants only to limit the number of failures which will occur within the few hours of operation required. It is also insufficient for the man who is concerned primarily with the problem of how many failures of a specific nature will occur in a specialized circuit application every day, or every week. The type of data most often requested by customers is that which indicates cumulative percentage failures for a specified life period, or rates of failure for defects due to various causes. In order to supply this information to the industry, more extensive analysis of life-test data is necessary along lines different from those previously followed.

Analysis of Failure Data

Cumulative percentage-failure curves are plotted in Fig. 1 for two heterogeneous groups of indirectly heated cathode-type subminiature vacuum tubes which were life-tested at rated operating voltages at normal room temperature (approximately 30°C). The larger group of 1,290 tubes of early design is the same group for which a curve of JAN percentages was shown in the paper previously published.[2] The main advantage of the type of curve of Fig. 1, which shows actual percentage of tube failure throughout life, is the facility with which the desired information may be extracted from the curve and applied to problems of specific usages. This type of curve also presents a clearer picture for comparing the quality of two lots of tubes, as demonstrated by the plotting of the two curves of Fig. 1.

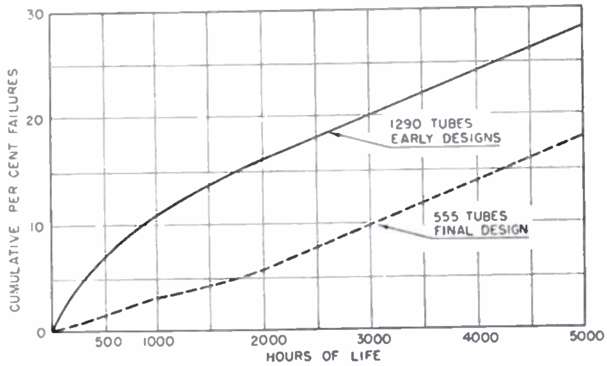

Fig. 1—Per cent failures on life. Subminiature tube types. 30°C ambient temperature. The use of this type of curve facilitates comparison of the long-life quality of the two groups of tubes. Both groups consist of tubes made in pilotline production and include such types as diode detectors, half-wave rectifiers, triodes, RF pentodes, power-output pentodes, and pentode mixers. All tubes were life tested at rated operating conditions at normal room temperature. Failure percentages include all types of defects, both inoperable and "out-of-limits."

More pertinent information as to individual types of failures to be expected within specific portions of the life curve may be indicated by showing the relative percentage of failures due to various causes on a rate-of-failure curve, as shown in Fig. 2. A curve such as this

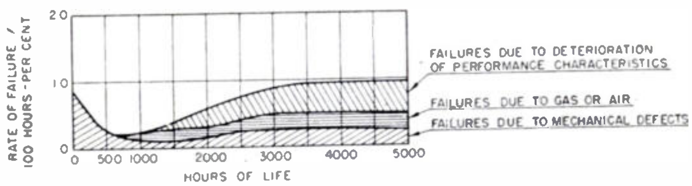

Fig. 2—Breakdown of failures on life. 555 premium subminiature tubes. 30°C ambient temperature. Rate of failures due to various causes. This type of curve provides more specific information concerning numbers and types of failures encountered within definite portions of the life curve. Categories of failures may be further defined as (a) inoperable tubes in which obvious structural defects have been observed, such as open welds, interelectrode shorts, and the like; (b) tubes which are inoperable due to gas arcs or air leaks, or tubes with sufficient grid current to affect performance characteristics; (c) tubes whose performance characteristics fail to meet life-test end-point limits. The latter category may include inoperable tubes when tests fail to reveal any mechanical defect or the presence of gas or air within the tube.

permits the equipment designer to consider only the types of defect which will cause failure in a specific application. In many cases, tubes which fall below life-test end-point limits may not seriously impair performance of the unit, and only inoperable failures need be considered. When data are available on thousands of tubes of a single type, it becomes possible to restrict the categories of failures to much more accurate description; to show, for example, which elements of a tube are open or shorted, to what extent electrical characteristics have

[2] Average life percentage at X hours =

$$\frac{\Sigma\ \text{(life hours for each tube*)}}{X\ \text{hours (number of tubes started)}} \times 100$$

* The life hours for any individual tube shall be the total number of hours that tube has completed without failure, and shall be a maximum of X hours for each tube.

deteriorated, and so on. The accumulation of this type of information offers additional advantage in guiding the efforts of development work toward the reduction of the specific types of failures which are most objectionable in a particular application.

Because of the scarcity of available information concerning performance of tubes in field usage, it is often necessary for a manufacturer to conduct laboratory tests at conditions other than regular life-test specifications in order to investigate failures which may occur in specialized applications. When tubes are to be subjected to intermittent operation of heaters, for example, a manufacturer will test tubes under severe cycling conditions to assure reliable operation of heaters. An important instance of a special requirement to be met was encountered with subminiature tubes developed for use in airborne equipment,[3] where the ambient-temperature range rises as high as 175°C. Special heat chambers were constructed to test subminiatures at a 175°C ambient temperature to obtain the data used to plot the failure curve shown in Fig. 3. Various other applications may necessitate tests under such conditions as impact shock, sustained vibration, high altitude, pulsed voltages, or voltages other than normal operating conditions.

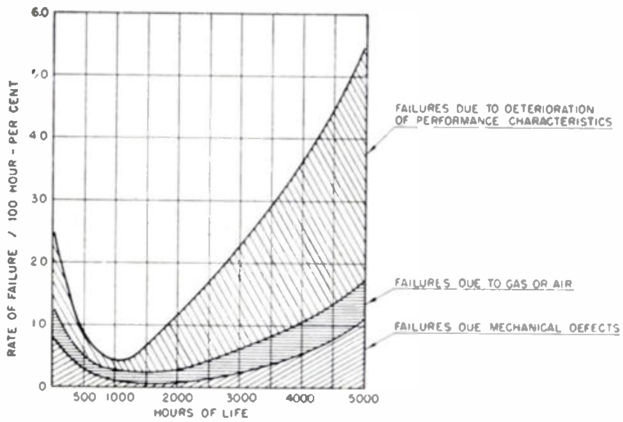

Fig. 3—Breakdown of failures on life. 141 premium subminiature tubes. 175°C ambient temperature. Rate of various failures on high-temperature life. A comparison of this curve with the curve of Fig. 2 illustrates the effects of high ambient temperature upon long-life performance.

New Concepts

Such an investigation of individual tube failures introduces several new concepts of tube operation. It becomes apparent that the dual objectives of reliability and long life are not necessarily attained simultaneously. Analysis of individual failures indicates that there are actually several component failure curves which influence different periods of life operation. The curves of Fig. 2, which are based on the results of life-tests at rated voltages on premium subminiature tubes made in pilotline production, indicate that early failures are due almost entirely to mechanical defects. The curve of these inoperable failures apparently is highest during the first hundred hours of life, decreases for about 1,500 hours, rises slightly to 3,000 hours, and is fairly level thereafter. Failures resulting from the evolution of gas and/or the deterioration of electrical characteristics are negligible until after 1,000 hours of life, and then show a gradual increase as life continues. There appears to be an interdependence between the gas evolution and deterioration of characteristics, with both resulting from some cumulative effect within the tubes which build up to the failure-causing point only after a certain period of tube operation. Operation at high ambient temperature decreases reliability to some extent in early hours, and it appears to accelerate tube failures to a marked degree after approximately 1,500 hours of life.

Application to Future Program

An immediate result of this type of analysis is to guide the direction of contemporary research and development programs along the particular lines dictated by specific applications. When certain usages require a line of tubes which will be highly dependable for 1,000 hours of life, development efforts may be concentrated on elimination of the mechanical failures normally encountered within that period. Considerable progress has already been made in this field by such improvements as simplified mount designs and tightened quality-control systems. When other applications require tube operation for a 5,000- or 10,000-hour period, research may be concentrated on the investigation of designs and materials which will achieve stability of electrical characteristics for at least that period of time. This type of investigation includes work on such basic factors as cathode materials, coating compositions, degassing processes, and the like. Development programs have been initiated on Sylvania premium subminiatures to investigate methods of attaining both early dependability and long-life performance.

The type of tube-failure analysis discussed here should also result in more specific definition of the type of reliability desired for particular applications, and thus make possible a comprehensive examination of suitable methods of evaluating "reliability." Such artificial ratings as the JAN average life percentages are admittedly adequate for certain uses, and undoubtedly are the best method available to the industry at the present time. But it is evident that some new type of evaluation will eventually have to be devised to take into account both the rate of occurrence of inoperable failures and the rate of deterioration of performance characteristics.

A significant advance toward the attainment of tube "reliability" can be accomplished by thus identifying the difficulties which may be encountered along the way. Considerable progress has already been made in producing reliable tubes, and in some cases it has outdistanced the work on reliability of other electronic components. Further progress is possible, however, only when a realistic attitude is maintained toward the problems which must be solved.

[3] This work was done under contract with the Air Matériel Command.

General Considerations in Regard to Specifications for Reliable Tubes*

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This paper is published with the approval of the IRE Professional Group on Quality Control.—*The Editor.*

Summary—**The urgent need to design and manufacture electron tubes which will give more reliable performance under the stresses imposed by complex military and industrial electronic gear demands a new approach to the objective of a more nearly perfect tube. The specification for such a tube should so accurately describe its characteristics that the finished product, manufactured in accordance with the description, will fully satisfy the needs of the equipment designer. It should also enable him to incorporate the tube into an assembly of a high order of reliability. Inspection and application manuals formulated in accordance with the same objectives, and definitely correlated with each other, should be considered as a part of the specification. The lot acceptance system, coupled with adequate sample size, is recommended as an integral feature of the proposed type of specification. Properly employed, it is more effective than 100-per cent screening, and also provides the manufacturer with incentives to *build* quality into his product. Life testing based on adequate sample size is also strongly recommended.**

Statement of Problem

RELIABILITY of electron tubes has been for some time a matter of concern both to the manufacturers and the users of tubes. In the past two years, however, the problem has been receiving major attention on an increasing scale. The reasons are readily apparent.

In the military services and industry alike, there has been a rapid trend toward vastly more complicated electronic equipment performing hitherto undreamed-of functions. As a result, the number of electron tubes used in individual equipments has mushroomed to a point where as many as 18,000 tubes are required for a single installation. What is more important, tubes in these new applications are subjected to severe stresses, often under very adverse environmental conditions. Yet, the only tubes available for the new and complex devices were those designed and manufactured to meet the modest requirements of 5-tube and 6-tube receivers for home entertainment purposes. This was an understandable situation since the new electronic developments have taken place over a surprisingly brief span and, in most cases, before either the tube manufacturers or the equipment manufacturers fully realized that they were witnessing a major trend. But the unhappy result of the situation has been, in effect, that a child has been employed to do a man's job.

To build equipment employing a large number of vacuum tubes is a relatively simple task insofar as the time element is concerned. It is another matter, however, to improve the techniques of designing and manufacturing vacuum tubes from a point where the tubes are satisfactory for mere home-entertainment requirements in 5- and 6-tube combinations to a level where they will meet the rigorous demands of complex electronic equipment increasingly utilized and heavily depended on for military and industrial purposes. Such improvement obviously will require an entirely new approach to the objective of a more nearly perfect electron tube. That approach, in turn, will entail for the manufacturers of tubes substantial economic problems which must be weighed against the vital advantages to be gained. In this discussion, however, the writers are concerned not with economics or manufacturing techniques, but with the general nature of the type of specification that is considered prerequisite to greater tube reliability.

Concept of Approach to Perfection

At the outset, we wish to emphasize one major factor which is of paramount importance in devising a format for a specification. This factor is the desire to approach perfection. Perfection, or in this case complete reliability with zero probability of failure, is an ideal which can only be approached. The degree to which it can be approached depends in part on the investment which industry can make in the interest of manufacturing a more reliable product, and, in part, on the efficiency of techniques for the measurement of tube performance and characteristics. Both of these factors must be weighed against the strategic importance of the electronic device in which the tube is utilized. Let us keep this concept of *approach to perfection* in mind as we discuss and consider specifications.

Purpose of a Specification

To begin, it is first logical to ask, "What is the purpose of a specification?" It would appear that a reasonable answer to this question is: "To describe so precisely the product being purchased that the finished product, manufactured in accordance with the description, will fully satisfy the needs of the purchaser." In the case of an electron tube, the initial purchaser is the equipment-design engineer. He, however, designs the equipment for the ultimate user or purchaser and, conse-

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† Aeronautical Radio, Inc., Washington, D. C.
‡ Bureau of Ships, Dept. of the Navy, Washington, D. C.

quently, the electron-tube specification becomes a subdivision of the equipment specification. This is as it should be, since it is obvious that over-all reliability cannot be obtained without complete harmony between the tube design and its use in the equipment design.

The above discussion leads to only one conclusion: That the electron-tube specification should describe the product in such a way that the equipment designer can incorporate the tubes along with other components into a completed assembly of a high degree of reliability—and, ultimately, *predictable* reliability. (It is implicit in this conclusion, of course, that the equipment designer, in using the tube toward the objective of the ultimate customer, must use it within the limitations indicated in the description.) The foregoing observation may seem to be a statement of the obvious. *It is all too true, however, that tube specifications which have been written in the past have only limited usefulness to the equipment designer.* The reasons for this inadequacy are discussed later in this paper.

Integration of Inspection and Application Manuals with Specification

The writers are fully aware that it may not be possible to formulate a single document which not only describes the product in accordance with the above objectives, but also can serve as a manual for inspection and acceptance of the product. The latter has often been the concept of a specification in the past. The hope is, however, that, with these objectives in mind, it will be possible to devise an inspection manual definitely correlated with an application manual which can then be considered an integral part of the specification.

Obviously, to write a specification to fill all the needs mentioned above is difficult. The authors cannot suggest easy ways in which this can be accomplished. However, the difficulty of the task does not detract in the least from the necessity of thinking of a specification in these terms. Certainly the objective can never be accomplished if this is not done.

The following discussion may at first seem over-simplified and somewhat academic. It is introduced here for the purpose of clarifying objectives, which, rather than the proposal of solutions for detailed problems, is our aim in this paper.

The Ideal Product and the Range of Values

One purpose of describing a product which will satisfy the needs of the purchaser is to achieve, as far as possible, initial interchangeability as between tubes produced by different manufacturers—in other words, to endeavor to achieve standardization. This is an important objective of a specification because it enables the equipment designer to know, irrespective of the make of tube, what deviations from the ideal can be expected. In the past, this appears to have been a major objective of specifications. However, earlier specifications have fallen short of providing this information because they have indicated only the *range* of values which might be expected. They have generally failed to present the other and more important part of the picture,—i.e., to single out the point in the range of values which characterizes the ideal product and to indicate the percentage distribution of the product within the range.

The consequence of this situation has been that the equipment designer, for the most part, has mistakenly assumed that an individual tube is equally likely to have characteristics at any given point in the range of values as at any other point. To reduce this to concrete terms, let us take as a simple example, an electron tube for which the specification permits a tolerance range for transconductance of 2,500 to 4,500 micromhos. Too often the equipment designer has made the mistake of assuming that an individual tube is just as likely to fall in the 2,500 category as in the 2,800 or the 3,600 category, for example. In other words, he has assumed that a curve plotting the characteristics of the tubes in any given lot would be horizontal. This is an obvious fallacy. It has been demonstrated that deviations of individual items from the ideal follow well-established statistical laws. In actual fact, a curve plotting the characteristics of the tubes which we have taken as an example would be more likely to resemble a normal distribution curve. The disparity between the assumed and the actual situations is illustrated in Fig. 1.

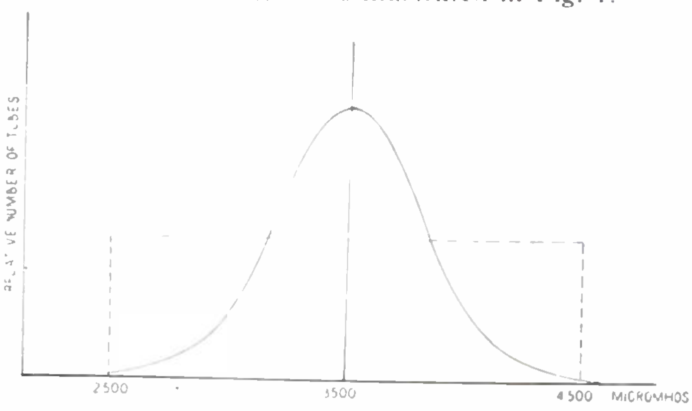

Fig. 1

Role of the Specification in Approach to Perfection

It is an elementary and fundamental fact that no two things in the physical world are exactly alike. But if real progress in tube reliability is to be achieved, there must be a persistent and unrelenting effort to make two things as nearly alike as is economically practicable—that is, to make electron tubes which resemble the *ideal* tube to the highest degree permitted by existing economic factors.

Our ability to duplicate an object as exactly as practicable depends (1) on the amount of effort and expense we are willing to devote to this duplication, and (2) on our ability to discern precisely the differences between two objects. The latter ability depends in turn upon our measurement technique and precision. How well we succeed in our effort to achieve exact duplication is logically measured by the deviation of the completed bject from the ideal which is our objective. But even iven the necessary effort and expense, plus accurate easurement techniques, it seems inconceivable that we an hope for success in making two or more objects as like as practicable unless we consciously aim at a ingle, clearly defined objective. It is equally inconeivable that we can accurately aim at any goal unless ve have before us a specification completely and preisely describing the object which we seek to duplicate. Older specifications presenting only a range of values lo not supply such a description. They are inadequate, herefore, in two ways: First, they do not show the percentage distribution of the product within the allowable range of values, and thus fail to describe the product to the equipment designer as it really is. Second, they fail to give the tube manufacturer a precise objective toward which to direct his efforts.

A specification which provides precise objectives, sometimes referred to as "bogies," and which specifies allowable deviations from these bogies in terms of the statistical laws which the product obeys, should overcome both of the above-mentioned objections. The equipment designer should be materially aided in having before him an accurate indication as to what he may expect from the product he is using, since the deviation determines the percentage of the product which might be expected to fall at any particular part in the range of variation. This is not a new concept, but has been proposed for vacuum tubes both by Davies[1] and Steen.[2] It is hoped that by reiteration this concept will become appreciated by all—not merely tube manufacturers, but equipment manufacturers as well, since such a description of the product, when properly interpreted from a statistical standpoint, can give more realistic design information than has been generally available in the past. This should make the job of designing reliable electronic equipment an easier one.

Systems for Acceptance

Most of the foregoing discussion has dealt with the question of how electron-tube characteristics should be specified. Once the method for such specification is established, the next question which arises is the system to be used for acceptance of the product.

The Fallacy of Screening

Under older types of specifications, individual tubes falling outside of the specified range of values were rejected, while those falling within the range were accepted. This type of test was performed as to many characteristics on 100 per cent of the tubes. The purpose, according to popular fallacy, was to insure that no tubes would be outside the specified range. This general acceptance procedure shall be referred to as "screening."

The fallacy of the screening procedure lies in the fact that many characteristics of vital importance to reliability (such as microphonics, RF noise, shorts, and continuity) were of a type where the accuracy of the test used for discerning the difference between a good tube and a bad one was of a very low order. Consequently, screening tests were exceedingly ineffectual in eliminating bad tubes from the product. It is quite safe to say that if 15 per cent of the tubes were rejected and removed from a lot in the first test; and if the remaining 85 per cent were tested a second time, the per cent rejections would differ only slightly from the first test. Obviously, the first screening process did *not* insure that the tubes were within the specified range. In fact, it had only a very small effect on the quality of the product. It has often been stated that quality cannot be tested into a product. The above example illustrates this quite clearly.

It should be clear at this point that 100-per cent testing on a screening basis does *not* insure a good product. How, then, might a specification be written that would insure a good product and, at the same time, provide incentive for the tube manufacturer to build quality into it? The answer appears to lie in the system of lot-acceptance testing. This, also, is not a new concept, but it is one which needs emphasis.

Lot-Acceptance System

Under the lot-acceptance system, a specification would state the maximum percentage of tubes having a certain defect which a lot may contain to allow it to be accepted. Let us take an example. First, assume that a specification is written which will not allow acceptance of lots having more than 5-per cent defects due to shorts. Now, let us assume that a lot submitted for inspection has 15-per cent defects; the specification would require that this lot be rejected. It may be argued, and rightly so, that such a criterion would result in rejection of a great many tubes which may be perfectly satisfactory. This objection can only mean, of course, that the test method is not adequate to discern between the good tubes and the bad ones. If this be true, we most certainly cannot obtain reliability by accepting the lot. The manufacturer should be allowed to take this lot, reprocess them, or 100 per cent inspect them as many additional times as he wishes. If, after such reprocessing or retesting, the lot passes the acceptance criterion (which, after one rejection, should be tightened considerably to reduce the random chance of a marginal lot being accepted), that particular lot may then be accepted.

Sampling

It will be noticed that in the preceding paragraph no mention has been made of sampling. This has been an intentional omission, since in many instances where the lot-acceptance testing method has been used, the tests have been performed on a sampling basis. There is a tendency, therefore, to think of the lot-acceptance method and sampling as inseparable. In practice, however, this is by no means true. Sampling, properly employed, can be a potent incentive to the manufacturer

[1] J. A. Davies, "Quality control in radio-tube manufacture," Proc. I.R.E., vol. 37, pp. 548–556; May, 1949.

[2] J. R. Steen, "The JETEC approach to the tube-reliability problem," Proc. I.R.E., vol. 39, pp. 998–1000; September, 1951.

to *build* quality into the product. As such, it should play an important role in the type of tube specification proposed in this paper.

Sample Size

It must be emphasized that proper use of the sampling technique is essential to full realization of its benefits. In the past there has been considerable misunderstanding on the part of many regarding the accuracy of the results to be obtained through the lot-acceptance method. These misconceptions have stemmed largely from the fact that all too often sample sizes used in lot-acceptance testing have been too small to insure accuracy. In other words, because the sample has been too small to be representative, the probability of accepting a lot containing defectives several times the acceptable quality level has been considerably too high. This, however, is a criticism of sampling techniques and sample size. It is not a criticism of sampling itself or of the lot-acceptance method. The solution is not to abandon sampling, but to improve the sampling technique—specifically, by using samples large enough to insure an adequate degree of accuracy.

It is obvious, of course, that in deciding upon the sample size, other factors which affect the accuracy of testing must be given proper weight. Such factors include fatigue and adverse psychological reaction experienced by persons engaged in testing large quantities of a product. To the extent that these factors reduce testing efficiency, they must be considered in determining the sample size which will produce satisfactory results.

Sampling as an Incentive to Manufacturers

As previously indicated, an important argument in favor of sampling, coupled with the lot-acceptance method, is that it can provide additional incentives to the manufacturer for building a good product. If he can build tubes of such consistently high quality that they can be accepted on the basis of a single sampling test, he will effect substantial savings in testing time, reprocessing time, and shrinkage. Once he can look forward to such economies, the incentive is provided to place tight manufacturing controls all along his production line and to build quality into the tubes. He will realize in actual fact that it costs less to build good tubes than bad ones.

Life Tests—Importance of Sample Size

One further element of a reliable tube specification which merits special comment is that relating to life-test information. Such data would seem to be of vital importance, particularly if life tests can be designed to simulate as much as possible field operating conditions.

Here again, sample size is the key to obtaining useful data. In the past, life-test samples have been so small that data obtained from them has practically no statistical significance. Undoubtedly this condition is attributable primarily to the cost of conducting life tests. The writers have no easy solution to this problem in manufacturing economics. Certainly the cost of life testing on a 100-per cent basis, or a basis even approximating 100 per cent, would be high. On the other hand, it seems clear that the size of samples heretofore used for life tests has been so inadequate as to be equally unacceptable. Somewhere between these extremes a satisfactory compromise should be found.

Another factor which has contributed to the inadequacy of life-test samples has been the belief that the life test is a destructive test. Such a view is justified, of course, if life tests are run for a sufficiently long period. There is, however, a large and rapidly accumulating body of information with respect to standard-line receiving tubes which indicates that most failures in well-designed tubes and applications are of a random nature, rather than of a deterioration type, during a period of many thousands of hours. Consequently, a 100- or 500-hour life test is of relatively little significance from the standpoint of deterioration, but can give a wealth of information on random failures. It would seem, therefore, that larger samples can be justified and that shipment of samples having completed life test does not reduce the reliability of the lot. On the contrary, in most cases reliability is actually improved, since many of the random failures, which occur early in life, have been eliminated. Therefore, life tests which provide a tight control on random failures are to be highly recommended. Such tight control can only be obtained by sample size consistent with the quality level to be achieved.

Also to be considered in determining sample size for life tests is the fact that when attributes alone are being evaluated, a larger sample is necessary than when variables are being considered. The reason for this is apparent from the definitions of these two terms, which may be stated as follows: Attributes are those types of characteristics, such as defective filament in an electron tube, which are usually described by saying that the condition either exists or it does not exist; i.e., there are no varying *degrees* of the condition. Variables are those types of characteristics, such as transconductance, which are generally expressed in varying degrees; i.e., the object being tested may possess the characteristic only to a small extent, to a moderate extent, or to a very large extent.

Conclusion

Plainly, the type of electron-tube specification which will accomplish all the objectives described in this paper cannot be achieved easily and quickly. Its development must necessarily be a gradual, step-by-step process. The test of the specification in each step of its development must obviously be in terms of the field performance of the electronic equipment in which the tubes are employed. In other words, the value of the specification should be measured against its purpose as previously defined: "To describe so precisely the product being purchased that the finished product, manufactured in accordance with the description, will fully satisfy the needs of the purchaser." In this case, the purchaser is the ultimate user of the electronic equipment.

Resonance Characteristics by Conformal Mapping*

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Summary—The analytical expression $f(\lambda)=a\lambda+b+c/\lambda$, where $=\sigma+j\omega$, may be called the "resonance function." It represents, for example, the impedance of the series-connected LRS circuit and the admittance of the parallel-connected CGΓ circuit.

The conformal mapping of the λ-plane onto the $f(\lambda)$-plane yields figure of considerable usefulness in clarifying the meaning of the resonance function for complex frequencies. Typical examples illustrate the application of the figure to the parallel circuit and one of its intrinsic generalizations.

A brief discussion of the mapping of the reciprocal of the resonance function is included for completeness.

I. THE RESONANCE FUNCTION

THE FOLLOWING expression with real coefficients,

$$f(\lambda) = a\lambda + b + c/\lambda, \tag{1}$$

may be called the "resonance function." Now λ may in general be considered a complex frequency variable; it is thus instructive to map the λ-plane onto the $f(\lambda)$-plane. In order to obtain a single "universal plot," (1) may be transformed in the following way. First define

$$\omega_0 = \sqrt{c/a}, \tag{2}$$

and then divide (1) by the constant $\sqrt{ac}=a\cdot\omega_0$, giving

$$\frac{f(\lambda)}{a\omega_0} = \frac{a\lambda}{a\omega_0} + \frac{b}{a\omega_0} + \frac{c}{a\omega_0\lambda}. \tag{3}$$

If the frequency λ is considered as complex, then $\lambda=\sigma+j\omega$, and (3) may be written as

$$\frac{f(\lambda)}{a\omega_0} = \frac{\sigma+j\omega}{\omega_0} + \frac{b}{\sqrt{ac}} + \frac{\omega_0}{\sigma+j\omega}, \tag{4}$$

or finally as

$$F(\Lambda) = F(\Sigma + j\Omega) = \Lambda + \frac{b}{\sqrt{ac}} + 1/\Lambda, \tag{5}$$

where $\Sigma=\sigma/\omega_0$ and $\Omega=\omega/\omega_0$, the real and imaginary parts of the normalized complex frequency variable.

A plot of the $\Lambda=(\Sigma+j\Omega)$-plane in the form of a rectangular Cartesian grid onto the $F(\Lambda)$-plane then yields an interesting and useful figure. Since only real coefficients of (1) are considered, it is clear that the real term $b/\sqrt{ac}$ in (5) merely amounts to a shift of the plot along the real axis of the $F(\Lambda)$-plane. The conformal map in Fig. 1, which is drawn for the case $b=0$, thus represents (1), and may be written as

$$F_0(\Lambda) = \Lambda + 1/\Lambda \tag{6}$$

without loss of generality.

* Decimal classification: R141.2. Original manuscript received by the Institute, February 15, 1952; revised manuscript received May 21, 1952.

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Not only is Fig. 1 a plot of intrinsic beauty, but it places in a very clear light

(a) the real or complex conjugate roots of the resonance function (5) and (1),

(b) the variation of the function (1) for purely imaginary ranges of the variable $\lambda=\pm j\omega$ (This may be interpreted as the impedance $Z(j\omega)$ of a series-connected LRS circuit, or the admittance $Y(j\omega)$ of a parallel-connected CGΓ network for purely imaginary or $j\omega$-frequencies.),

(c) the variation of the function (1) for complex ranges of the variable $\lambda=\sigma+j\omega$ (This may be interpreted as the impedance $Z(\sigma+j\omega)$ of the series-connected LRS circuit or admittance $Y(\sigma+j\omega)$ of the parallel-connected CGΓ network for complex frequencies $\lambda=\sigma+j\omega$.),

(d) the impedance or admittance of two distinct groups of networks which may be derived from the original resonance function (1).

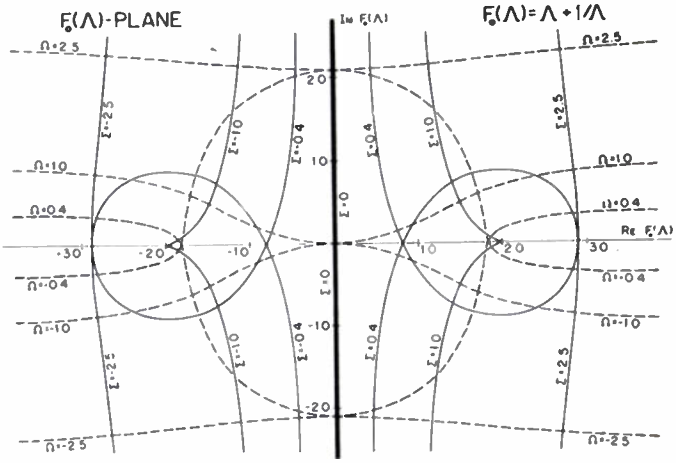

Fig. 1—Conformal map of the resonance function, $F_0(\Lambda)=\Lambda+1/\Lambda$.

Fig. 2, a more detailed version of Fig. 1, is necessary for applications involving numerical magnitudes. The examples in the following sections illustrate the applications of the conformal map to electric networks, utilizing Fig. 2.

II. PARALLEL-CONNECTED CGΓ NETWORK APPLICATIONS

The following example illustrates the generalized admittance of the parallel-connected CGΓ network for complex frequencies.

Example 1—Resonance Function with Complex Conjugate Roots, Poles at Origin, and Infinity

The admittance of the parallel network shown in Fig. 3, given by the following equation

$$Y(\lambda) = C\lambda + G + \Gamma/\lambda, \tag{7}$$

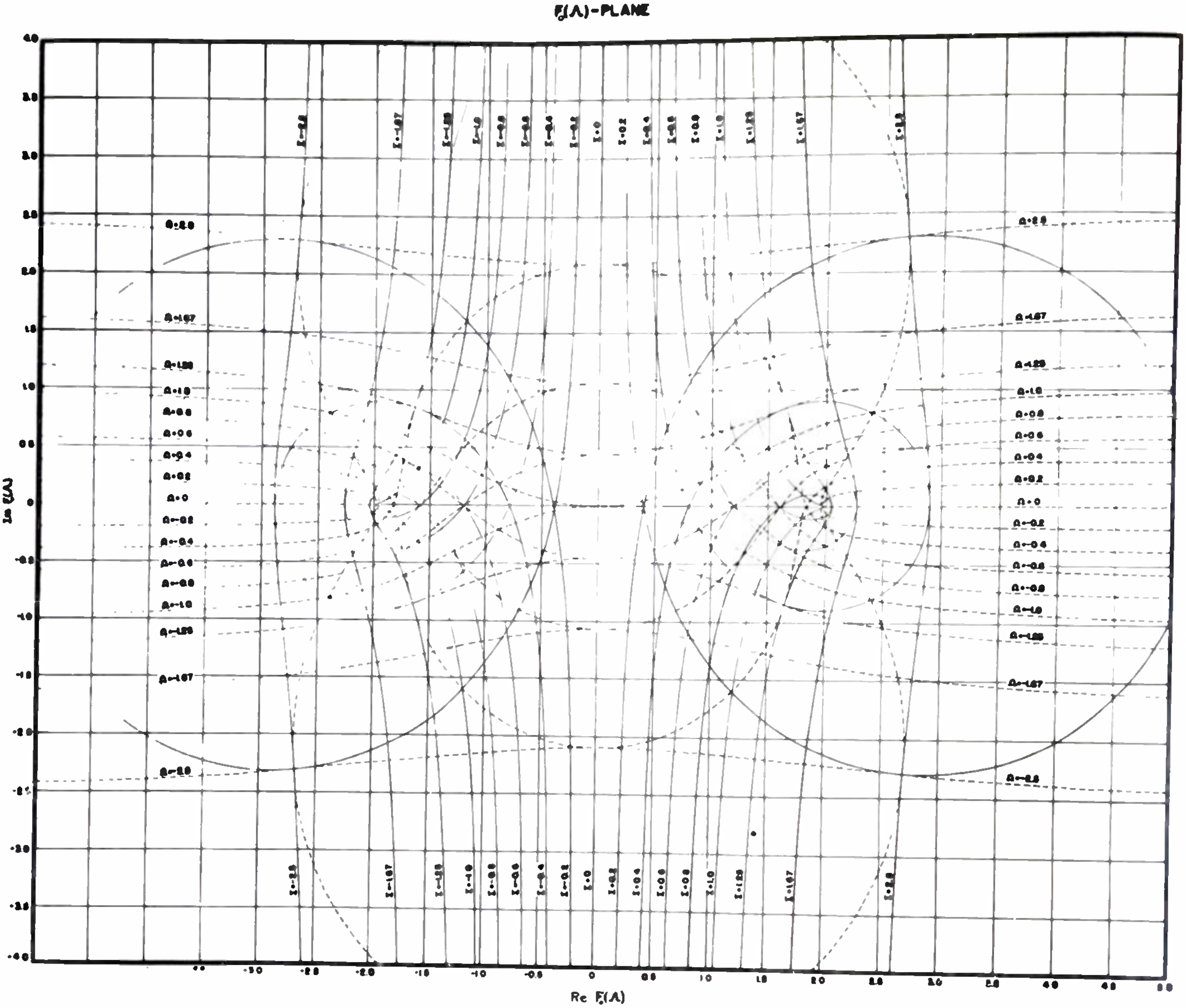

Fig. 2—Detailed conformal map of the resonance function, $F_0(\Lambda) = \Lambda + 1/\Lambda$.

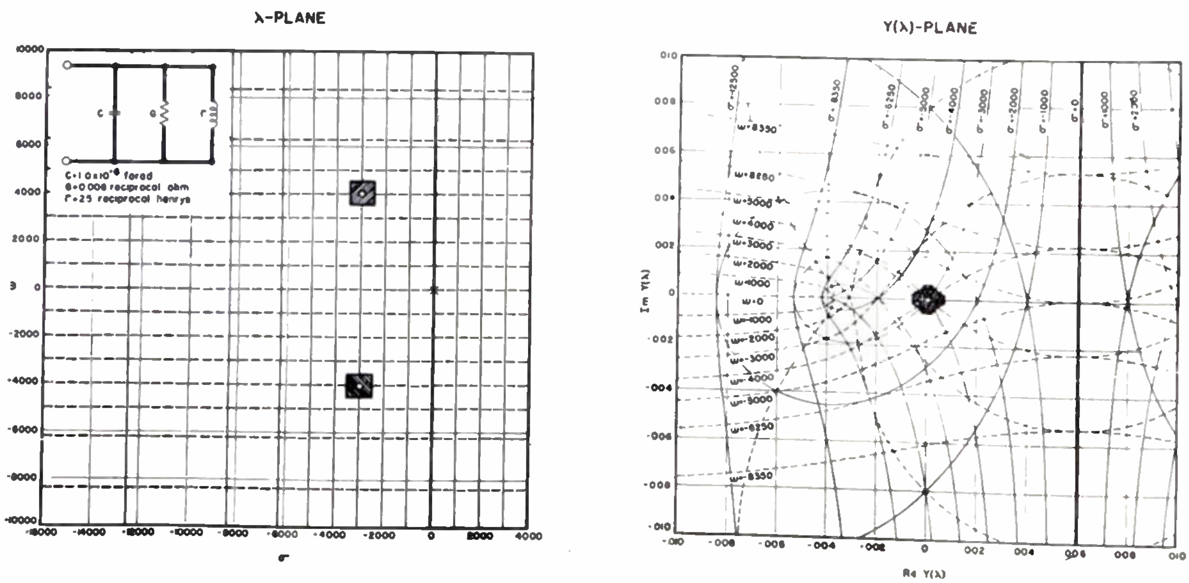

Fig. 3—Resonance function with complex conjugate roots and poles at origin and infinity.

ıas complex conjugate roots at $\lambda_{1,2} = -3{,}000 \pm j4{,}000$ and poles at the origin and infinity. These critical frequencies of the network are conveniently represented by their positions in the λ-plane, the roots by circles, and the poles by crosses. The roots of the equation represent the complex frequencies which yield zero admittance for the network or $Y(\lambda) = 0$, and can also be interpreted as being the free modes of oscillation of the open-circuited network. This means that the separate neighborhoods of the λ_1 and λ_2 roots in the λ-plane map as overlapping areas in the neighborhood of the *origin* of the $Y(\lambda)$-plane. This is shown in Fig. 3 by the cross-hatched areas in the λ-plane and $Y(\lambda)$-plane. For the particular magnitude of shunt conductance chosen in this example, the free modes of oscillation correspond to an exponentially damped sinusoid.

The nature of the variation of the admittance function $Y(\lambda)$ for both complex and purely imaginary ranges of the λ-variable are most easily visualized from the plot of the $Y(\lambda)$-plane, Fig. 3. The admittance for generalized or complex frequencies is given by the curved lines in the figure.

The heavy straight line at $Re\ Y(\lambda) = 0.006$ reciprocal ohm represents the admittance $Y(j\omega)$ of the parallel circuit for $j\omega$ frequencies, corresponding to the heavy line in the λ-plane at $\sigma = 0$. This locus is the well-known "vector" diagram of the parallel circuit. Similar diagrams for the impedance of the series circuit may be found in several standard textbooks.[1,2,3]

III. Intrinsic Network Applications

It might be supposed from the example given above, for which the admittance for $j\omega$ frequencies is a straight line, (and similarly for the corresponding series-connected LRS circuit), that the conformal plot of Fig. 1 is of but limited "practical" import. Such, however, is not the case. The curvilinear graphs also represent the admittance or impedance of many other networks for purely $j\omega$ frequencies.

Thus, referring for the moment to Fig. 4(b), the two networks shown therein also fall within the purview of (1) and Fig. 1. For the series-connected network, the impedance is

$$Z(\lambda) = \lambda L + r + R + 1/(\lambda C + g), \qquad (8)$$

while for the parallel-connected network the admittance is

$$Y(\lambda) = \lambda C + g + G + 1/(\lambda L + r). \qquad (9)$$

When the condition $r/L = g/C$ is specified, (8) and (9) are broadly of the same analytical form as (1), and thus are also represented by the plot in Fig. 1. However, the arbitrary magnitude of the network element R in the impedance expression (8) for the series-connected network indicates that the condition $r/L = g/C$ can always be satisfied. The same is true for the admittance of the parallel-connected network (9), due to the presence of the arbitrary element G. The following example illustrates the application of Fig. 1 to such more complicated networks.

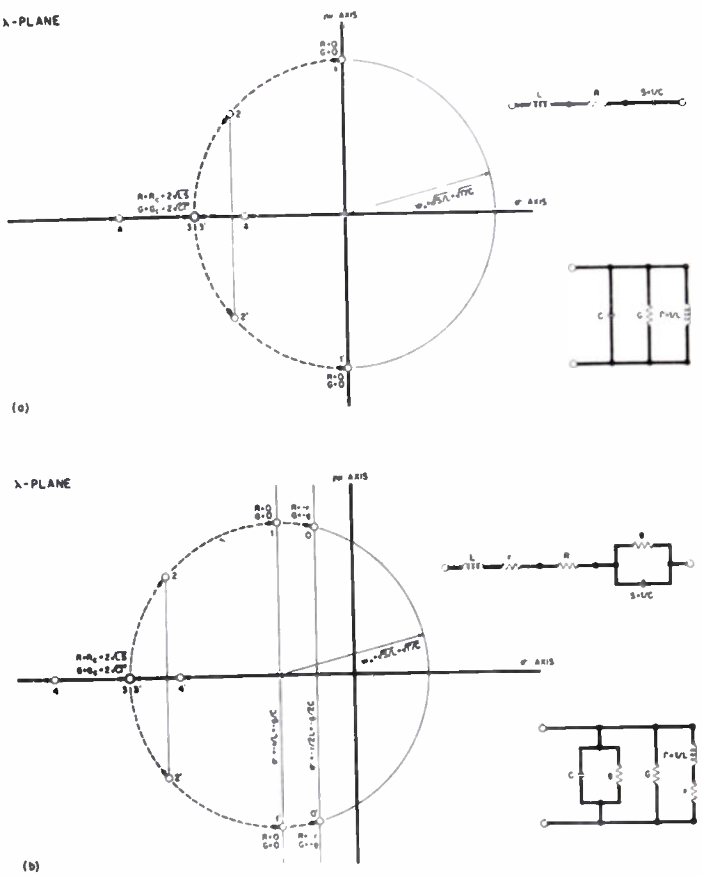

Fig. 4—Loci of roots and poles of the resonance function. (a) The simple series and parallel networks. (b) The more general series and parallel networks.

Example 2—Resonance Function with Complex Conjugate Roots and Poles on Negative Real Axis and at Infinity

A particularly interesting version of the parallel-connected group of circuits shown in Fig. 4(b) results when $G = -g$. This is illustrated by the circuit in Fig. 5, for which $g = -G = 0.003$ reciprocal ohm. This choice of parameter magnitudes results in the physical (although not analytical) absence of the shunting conductances g and $-G$. The roots of the network are at $\lambda_{1,2} = -1{,}500 \pm j4{,}770$, with poles at $\sigma = -3{,}000$ and infinity. The heavy curvilinear graph now represents the admittance $Y(j\omega)$ of the network for $j\omega$ frequencies.

The admittance diagram $Y(\lambda)$ in Fig. 5 shows graphically the well-known fact that a "tank circuit" exhibits unity power-factor at a $j\omega$-frequency which differs from

[1] W. L. Everitt, "Communication Engineering," McGraw-Hill Book Co., Inc., New York, N. Y., p. 49; 1932.

[2] R. M. Kerchner and G. F. Corcoran, "Alternating Current Circuits," John Wiley and Sons, Inc., New York, N. Y., p. 101; 1943.

[3] Cruft Laboratory Electronics Staff, "Electronic Circuits and Tubes," McGraw-Hill Book Co., Inc., New York, N. Y., p. 40; 1947.

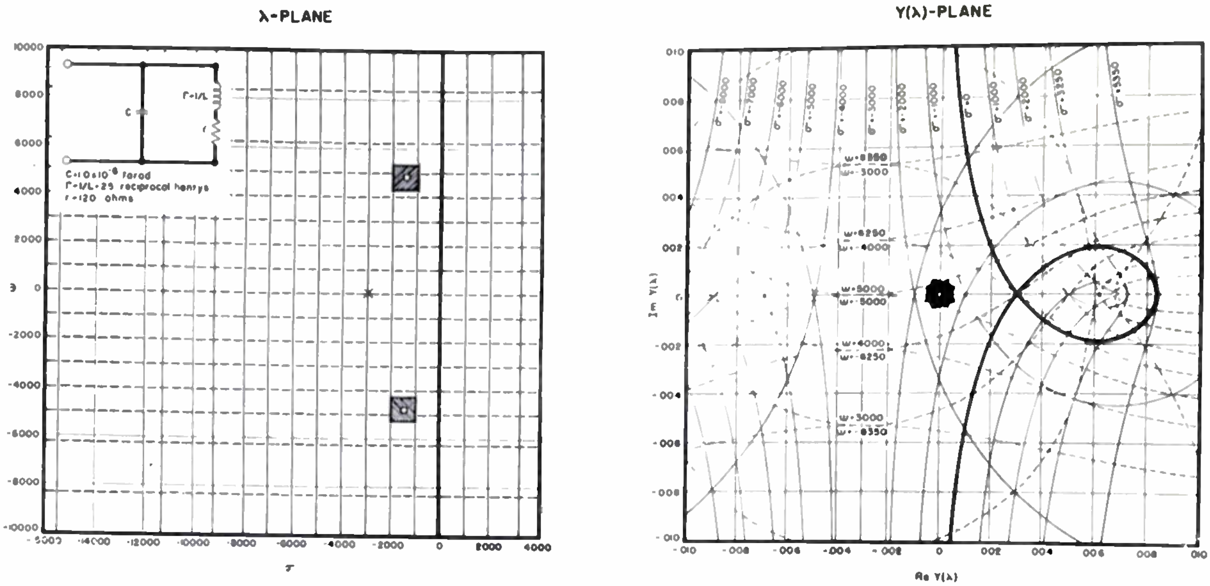

Fig. 5—Resonance function with complex conjugate roots and poles on negative real axis and at infinity ("tank circuit").

the frequency for minimum-admittance (maximum-impedance) magnitude. This is of some concern in the operation of output circuits of radio-frequency amplifiers, particularly for high-power applications.

IV. Relation Between Network Groups

The close relationship between the groups of networks, represented by Example 1 as contrasted with Example 2, can perhaps be seen in the clearest light from a consideration of the loci of the roots and poles for the networks in the λ-plane. The locus of the roots (and the position of the poles) of the simpler parallel CGΓ (or series LRS) networks is shown in Fig. 4(a), and is known to be a circle.[4,5]

More general circuits implicit in (1) are illustrated in Fig. 4(b). In these, the resistances and conductances associated with the reactive elements must be in the relation $r/L = g/C$; however, since the networks include the arbitrary elements G or R, this constraint is not as rigid as might be supposed. The loci of the roots and the position of the poles merely shift by an amount given by $\sigma = -r/L = -g/C$. The radius of the circular portion of the locus remains unchanged, as is clearly evident from a comparison of Fig. 4(a) with Fig. 4(b).

V. Reciprocal Functions

A conformal map of the reciprocal of the resonance function $F(\Lambda)$ would also be useful, particularly, for example, as an aid in visualizing the generalized impedance of the simple CGΓ parallel circuit. Fig. 6 is the conformal map of the reciprocal function $A(\Lambda) = 1/F_0(\Lambda)$, for special case in which $b=0$ [(1) and (5)].

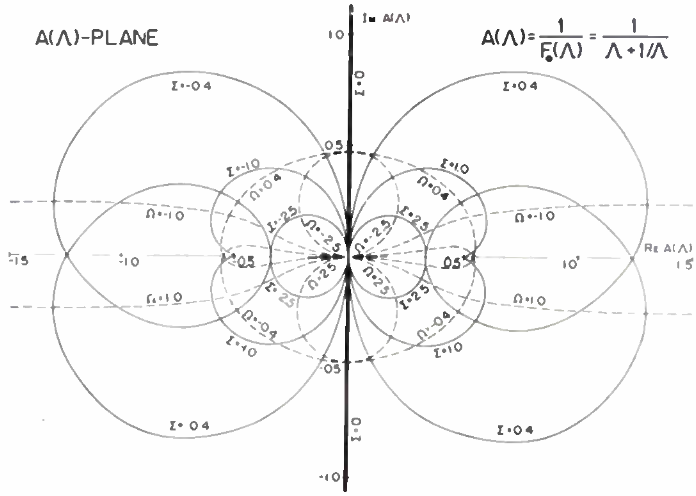

Fig. 6—Conformal map of the reciprocal function $A(\Lambda) = 1/F_0(\Lambda)$.

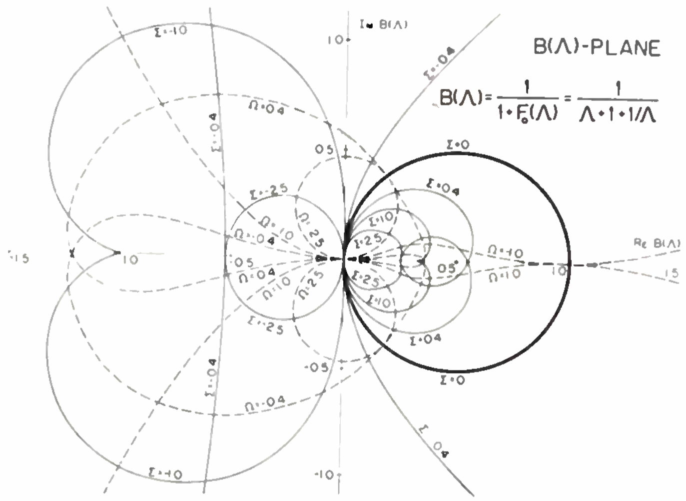

Fig. 7—Conformal map of the reciprocal function $B(\Lambda) = 1/(1+F_0(\Lambda))$.

[4] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., New York, N. Y., p. 27; 1945.

[5] E. A. Guillemin, "Communication Networks," John Wiley and Sons, Inc., New York, N. Y., vol. 1, p. 181; 1931.

A universal plot of the resonance function $F(\Lambda)$, Fig. 1, is feasible since the constant term in this function merely represents a shift of the mapping along the real axis in the function plane. The reciprocal function $1/F(\Lambda)$ apparently cannot be generalized in the same manner. This is clear from a comparison of Fig. 6 with Fig. 7, the latter being a conformal map of the function

$$B(\Lambda) = 1/(1 + F_0(\Lambda)) = 1/(\Lambda + 1 + 1/\Lambda), \quad (10)$$

which is the reciprocal of $F(\Lambda)$ for the special case $b = \sqrt{ac}$ (5).

VI. Conclusions

A measure of the generality of the resonance function is indicated by the specific examples which illustrate some uses of the mapping of the function. However, the examples given by no means exhaust the fields of application of the conformal plot since it is not limited to purely electric networks. For example, the resonance function has its counterpart in purely mechanical systems. Likewise, the resonance equation also arises in the consideration of the stability of certain electric[6] and electro-mechanical[7] networks employing electron-tube amplifiers. In these applications, the entire λ-plane and the $f(\lambda)$-plane are physically significant.

In conclusion, it should be pointed out that from the didactical standpoint the figures are invaluable and they apparently are not readily available elsewhere in the literature.[8] For example, the conformal maps illustrate a number of network theorems on physical realizability.[9]

[6] P. M. Honnell, "The generalized transmission matrix stability criterion," *Trans. AIEE*, vol. 70; 1951.

[7] P. M. Honnell, "The application of feedback to an electromechanical transducer for seismograph testing," *Bull. Seismol. Soc. Amer.*, vol. 40, pp. 217–231; July, 1950.

[8] These figures do not appear in the comprehensive list given by Betz. A. Betz, "Konforme Abbildung," Springer-Verlag, Berlin/Gottingen/Heidelberg; 1948.

[9] H. W. Bode, *op. cit.*, chapt. VII.

Scanning-Current Linearization by Negative Feedback*

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Of the wide variety of known methods of scanning current linearization employed in television systems, a number use feedback, and experience indicates that appropriate application of negative feedback offers the most effective approach to the attainment of sensibly perfect linearity.

The problem of generating a linear current change from a given source of constant direct potential difference is essentially one of "current integration" and may be expressed in the alternative forms

(form I) $i(t) = R^{-1}p^{-1}E$

(form II) $i(t) = (Rp)^{-1}E$,

where

E = available *pd* (usually a high-tension voltage),
$p \equiv d/dt$ is the time-derivative operator,
R^{-1} = voltage-current factor,
$i(t)$ = required scanning current form.

Correspondingly, it is found that practical circuit developments for the realization of $i(t)$ fall into either of two categories, according to whether the current integration is performed indirectly by performing the component processes

(IA) $e(t) = pE$ (i.e., voltage integration)

(IB) $i(t) = R^{-1}e(t)$ (i.e., voltage-current conversion)

in cascade (the "driven"-type scanning system), or is performed directly in a single stage (the "sawtooth-current" oscillator). Each of the basic processes may, with advantage, be performed by a negative feedback system, the appropriate type of feedback for each case being

(IA) derivative voltage feedback,

$$e_f(t) = kpe_0(t),$$

(IB) direct (i.e., without waveshaping) current feedback,

$$e_f(t) = Ri_0(t),$$

(II) derivative current feedback,

$$e_f(t) = Rpi_0(t),$$

subscripts f, $_0$ denoting feedback and output quantities, respectively. In principle, the required degree of linearity may then be obtained merely by use of sufficient loop gain, subject to the existence of a network capable of defining β in the desired manner and to the maintenance of an adequate margin against Nyquist instability.

In certain voltage-feedback amplifier configurations, typified by the Miller-Blumlein integrator, the over-all gain approaches, with increasing μ, the ratio of two branch impedances, thereby allowing exact derivative voltage feedback, process (IA), to be obtained by the use of a pure capacitor C in the appropriate branch with a resistor R in the other, giving $\beta = 1/Cp/R = 1/CRp = 1/\tau p$, as required. A convenient method of deriving a feedback voltage proportional to coil current in process (IB) without inserting a resistor in the path of the scanning current is to connect a series CR branch having the same time-constant (L/r) as the scanning coil in parallel with the latter, and take the feedback voltage from across the capacitor. Then

$$e_f(t) = \frac{1 + \frac{L}{r}p}{1 + CRp} ri_0(t) = ri_0(t), \text{ for } L/r = CR.$$

Transposing C and R provides derivative current feedback, as required in process (II). Thus,

$$e_f(t) = \frac{1 + \frac{L}{r}p}{1 + CRp} CRrpi_0(t) = (CRr)pi_0(t)$$

$$\text{for } L/r = CR.$$

Certain forms of type (II) system may be derived by carrying out the voltage-current conversion process within the feedback loop of type (IA).

In general, the driven-type system is preferred in vertical scanning systems, but the direct method is more appropriate for the high-efficiency resonant-return circuits now used for horizontal scanning. Apart from the general advantages accruing from the application of substantial negative feedback, its most attractive feature from the design point of view is the possibility of restricting major control of linearity to a quite small group of components. In contrast, other methods generally involve mutual cancellation of the various factors contributing to nonlinearity and require consideration of linearity at almost every point in the system.

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Broad-Band Matching with a Directional Coupler*

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Summary—This paper presents the results of a theoretical and experimental study of a waveguide matching technique which allows a directional coupler to be located any distance away from the discontinuity causing the original mismatch and a broad-band match to still be obtained.

Design curves are included which give the required coupling coefficient of the directional coupler and the power loss for a given initial mismatch and desired vswr reduction. Experimental confirmation of the theory is also presented.

I. Introduction

IN WAVEGUIDE work it is often necessary to reduce standing waves resulting from a discontinuity some distance away because it may not be desirable to work directly at the discontinuity. An example is the mismatch due to energy reflected from a parabolic reflector coming directly back into the feed. A fairly broad-band match can be obtained by use of a vertex plate mounted on the parabola (i.e., by working *at* the discontinuity), but this degrades the minor lobe structure considerably. For this reason a different matching method would be very desirable.

The method described here is illustrated in Fig. 1. A portion of the incident energy is taken off by the directional coupler into arm 1. (The coupler will be assumed to have perfect directivity so that there is no coupling of the incident wave into arm 2.) By proper adjustment of the phase and amplitude of the voltage returned to the waveguide by the coupler, it should be possible to reduce the original standing wave over a band of frequencies.

The disadvantage in this method comes from the fact that only a portion of the energy originally tapped off by the coupler is returned to the waveguide. The remaining energy must be absorbed in the termination on arm 2 to prevent resonance effects. However, there may be situations where the advantages arising from a reduction in vswr more than offset the resulting loss.

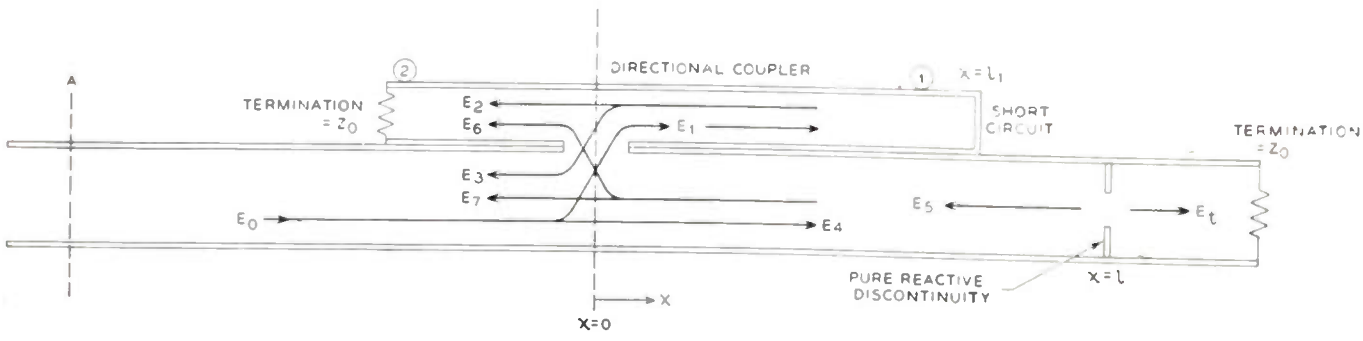
As is well known, another scheme for reducing the vswr over a broad band is to insert an attenuator between the source and the discontinuity. For a given initial vswr and reduction in vswr, however, the loss for this scheme always exceeds that resulting from the directional coupler scheme, and, in fact, for a complete cancellation of the vswr, the loss would be infinite.

II. Theory

It has been shown elsewhere[1] that $(E_1/E_0)_{x=0}=jc$, so that evaluating the voltages in Fig. 1 at $x=0$, and setting $E_0=1$, we get

$$\begin{aligned}
E_1 &= jc\\
E_2 &= -jc\sqrt{1-c^2}\exp(j2\beta l_1)\\
E_3 &= c^2\exp(j2\beta l_1)\\
E_4 &= \sqrt{1-c^2}\\
E_5 &= a\sqrt{1-c^2}\exp(j2\beta l+j\psi)\\
E_6 &= jac\sqrt{1-c^2}\exp(j2\beta l+j\psi)\\
E_7 &= a(1-c^2)\exp(j2\beta l+j\psi)
\end{aligned} \qquad (1)$$

where $\beta=2\pi/\lambda_g$, λ_g = guide wavelength. The amplitude



$q = ae^{j\psi}$ = REFLECTION COEFFICIENT OF THE DISCONTINUITY AND Z_0 IN PARALLEL

$c = \left|\frac{E_1}{E_0}\right|$ = VOLTAGE COUPLING COEFFICIENT OF THE DIRECTIONAL COUPLER

R = VSWR LOOKING TO THE RIGHT AT A

R_0 = VSWR BEFORE MATCHING $= \frac{1+a}{1-a}$

$L = \frac{\text{POWER INPUT}}{\text{POWER OUTPUT}} = \left|\frac{E_0}{E_t}\right|^2$

Fig. 1—Method of matching with a directional coupler.

* Decimal classification: R117.12×R310.4. Original manuscript received by the Institute, January 25, 1952; revised manuscript received May 22, 1952.

† Bell Telephone Laboratories, Inc., Holmdel, N. J.

[1] "Principles of Microwave Circuits," MIT Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., p. 299; 1948.

of the transmitted wave, E_t, is given by $[1-c^2]^{\frac{1}{2}}|1+q| = [(1-c^2)(1-a^2)]^{\frac{1}{2}}$.

The total wave traveling to the left in the main guide past $x=0$ is E_3+E_7. In order for this to be as small as possible, E_3 must be out of phase with E_7, or

$$2\beta l + \psi - 2\beta l_1 = n\pi, \ n = \pm 1, \pm 3, \pm 5, \cdots. \quad (2)$$

If (2) is satisfied, a complete match may be obtained by setting $|E_3| = |E_7|$. This occurs for a critical value of the coupling, c, equal to

$$c_o^2 = \frac{a}{1+a} = \frac{R_0 - 1}{2R_o}. \quad (3)$$

Two cases may now be distinguished, depending on whether $c > c_0$ (over coupling) or $c < c_0$ (under coupling).

Case I: $0 \leqq c \leqq c_0$.

In this case $|E_3| \leqq |E_7|$, and the vswr looking to the right at A is (assuming (2) holds)

$$R = \frac{1 + |E_7| - |E_3|}{1 - |E_7| + |E_3|} = \frac{1 + a(1-c^2) - c^2}{1 - a(1-c^2) + c^2}$$

$$= \frac{R_o(1-c^2)}{1 + c^2 R_o}. \quad (4)$$

The power ratio, L, is

$$L = \frac{R+1}{R}\,\frac{R_o+1}{4} \quad (5)$$

obtained by substituting for c its value as given by (4).

Case II: $1 \geqq c \geqq c_0$.

Now $|E_3| \geqq |E_7|$, thus R becomes

$$R = \frac{1 - |E_7| + |E_3|}{1 + |E_7| - |E_3|} = \frac{1 + c^2 R_o}{R_o(1-c^2)}, \quad (6)$$

and

$$L = (R+1)\frac{R_o+1}{4}. \quad (7)$$

Equations (4) to (7) have been plotted for values of R_0 from 1 to 9 db, shown in Figs. 2 and 3. The bounding curve on the right in Fig. 3 represents the loss resulting from the initial mismatch. The additional loss due to this matching scheme is given by the difference in loss between that for vswr $= R_0$ and for vswr $= R$.

If the actual coupling of the directional coupler built for use in a certain system turns out to be larger than desired, an effective decrease in coupling may be obtained by inserting an attenuator in the coupler arm containing the short circuit. The loss will be that which would be computed for the actual coupling value and initial vswr, R_0.

It is clear that for the directional coupler matching method to be effective over as wide a frequency band as possible, the two electrical lengths involved, βl and βl_1, must have the same frequency characteristics.

The frequency response of the directional coupler also affects the wide-band performance. By using couplers having many coupling elements, satisfactory performance can usually be obtained, however.

Another factor governing the frequency response may be seen by examining (2). Using this expression to de-

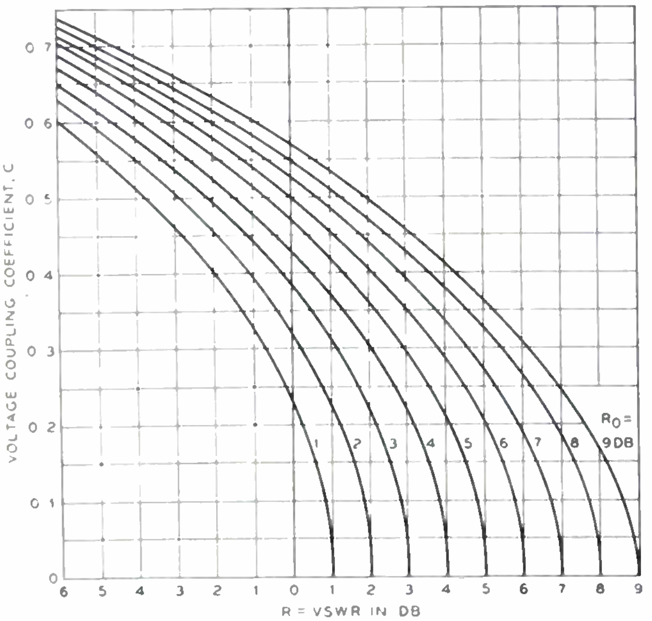


Fig. 2—Directional coupler voltage coupling coefficient, c, required for given initial vswr (R_0) and final vswr (R). The region to the right of the line $R=0$ corresponds to coupling less than critical ($c \leqq c_0$); that to the left corresponds to coupling greater than critical ($c \geqq c_0$). Critical coupling (c_0) is the coupling which reduces the standing wave to 0 db.

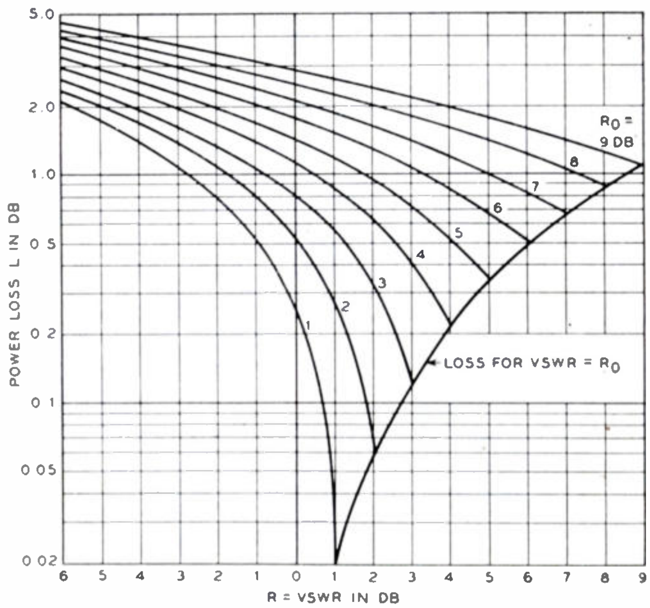


Fig. 3—Power loss L for given initial vswr (R_0) and final vswr (R). The region to right of the line $R=0$ corresponds to coupling less than critical ($c \leqq c_0$); that to the left corresponds to coupling greater than critical ($c \geqq c_0$). Critical coupling (c_0) is the coupling which reduces the standing wave to 0 db.

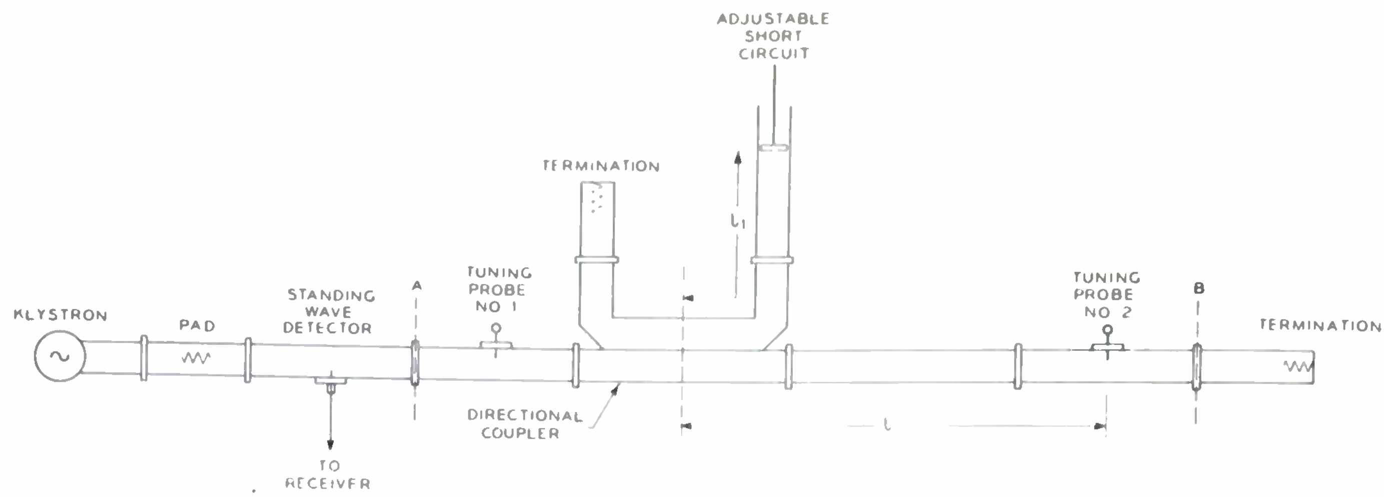


Fig. 4—Physical set-up used in experimental verification of the theory.

termine l_1

$$l_1 - l = \frac{\lambda_g}{4}\left(\frac{\psi}{\pi} - n\right). \tag{8}$$

Obviously the quantity in parenthesis should be as small as possible for broad-band operation, which indicates that $n = \pm 1$. The choice in sign depends on the sign of ψ, and for $\psi \neq 0$, one of the two choices will usually result in better broad-band operation. If $\psi = 0$, the choice is immaterial. The degradation in vswr due to the right-hand side of (8) will, in general, be less than 1.0 db for bandwidths up to 10 per cent, and for initial vswr less than 9 db for any value of ψ, providing the sign of n is chosen properly.

III. Experimental Results

The experimental set-up is shown in Fig. 4. Tuning probe #2 was set to give an initial vswr, R_0, of 5.5 db at 4,200 mc; the measured frequency variation is shown by curve 1, Fig. 5. Tuning probe #1 and the directional coupler were removed for this test. Curve 2 shows the vswr variation measured when probe #1 was used to give an initial match at 4,200 mc, with a separation between probes of $9\lambda_g$ at 4,200 mc.

The coupling of the directional coupler used was 0.542 at 4,000 mc, 0.50 at 4,200 mc, and 0.457 at 4,400 mc. Ignoring the $\lambda_g/4$ effect arising from (8), these values of coupling in connection with the vswr due to probe #2 should give final values of R equal to 1.3 db at 4,000 mc (over coupled), 0.3 db at 4,200 mc (over coupled), and 0.9 db at 4,400 mc (under coupled). The directivity of the coupler was 20 db or better over the band, so that any effect arising from imperfect directivity should be negligible.

Curve 3 of Fig. 5 shows the variation with frequency of the vswr using the directional coupler $9\lambda_g$ from probe #2 (probe #1 removed). Here l_1 was set to give a minimum vswr at 4,200 mc, but obviously this particular l_1 does not correspond to the optimum value of n. By setting l_1 to other values which gave a minimum vswr at midband, and taking frequency runs, an optimum setting was found. Curve 4 shows the vswr variation for this case. The difference in l_1 for curves 3 and 4 was one guide wavelength at 4,200 mc. Note that the vswr of 0.4 db at 4,200 mc is very close to the predicted value of 0.3 db. At the band edges the vswr is 0.7 db higher than predicted; this may be mainly accounted for by the $\lambda_g/4$ term of (8). At 4,200 mc with l_1 set to give 0.4 db vswr, a loss of 1.7 db was measured. From the curves, a loss of 1.6 db was predicted.

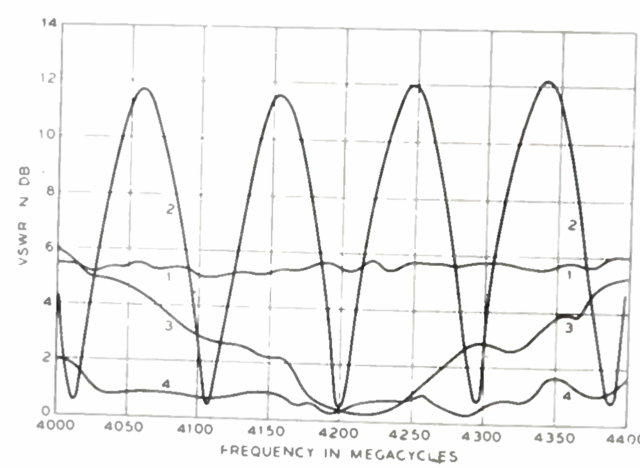


Fig. 5—Experimental results, showing variation of the vswr (R) versus frequency for: (1) discontinuity alone; (2) discontinuity with matching probe $9\lambda_g$ away; (3) discontinuity with directional coupler matching; (l_1 not adjusted properly); (4) same as (3), but l adjusted for best broad-band matching.

IV. Conclusions

A method of using a directional coupler to cancel standing waves has been investigated theoretically and experimentally. The method will give a broad-band match which is independent of the relative location of the discontinuity and directional coupler. Good agreement with the theoretical design curves has been obtained.

Synthesis of Narrow-Band Direct-Coupled Filters*

HENRY J. RIBLET†, ASSOCIATE, IRE

Summary—**This paper describes a general synthesis procedure for the design of narrow-band, direct-coupled filters, based on an approximate first-order equivalence between direct and quarter-wave coupled filters. Thus a quarter-wave coupled filter, whose bandwidth is a few per cent wider than required, serves as a prototype. The approximations underlying the synthesis procedure for quarter-wave coupled filters, given by Lawson and Fano and applied by Mumford to the "maximally flat" case, are re-examined and justified. The transmission characteristics of a six-cavity filter of total Q of 37.6 is computed exactly with excellent agreement with the prototype performance. Experimental confirmation is given.**

Introduction

A NUMBER of articles have been written in the past few years discussing the theory of narrow-band, waveguide filters, assuming constant shunt susceptances. Pritchard,[1] Fano and Lawson,[2] and Hessel *et al.*[3] have treated the problem of cascades of identical elements. The first two authors consider filter arrangements which they describe as quarter-wave coupled filters, while Hessel *et al.* discuss the case of identical direct-coupled cavity filters and treat a modification of this case which, under certain conditions, is equivalent to a quarter-wave coupled filter.

Unfortunately, filters consisting of identical elements have pass-band characteristics which deteriorate rapidly with increasing number of filter elements. Lawson and Fano[4] have shown how the general synthesis procedure described by Darlington[5] may be extended to narrow-band waveguide filters of both the quarter-wave and the direct-coupled types. For the case of quarter-wave coupled filters, Mumford[6] has given explicit formulas for the cavity Q's required for "maximally flat" performance, and has included the effects that the frequency sensitivity of the quarter wavelength connecting lines have on the final design. W. D. Lewis and W. W. Mumford[7] in a recently issued patent have given formulas for the susceptances and their spacings for maximally flat direct-coupled filters. Although they also use an equivalent quarter-wave coupled filter as a prototype, no information is given as to the method of derivation so that it is difficult to determine the range of usefulness of their formulas or the possibility of extending their method to the general synthesis problem.

The Problem

The synthesis procedure introduced by Lawson and Fano proceeds in two steps: In the first, a suitable section of the waveguide filter, consisting of line lengths and reflecting elements, is replaced by a low-frequency analogue whose behavior is linear in some frequency variable. In the second, the total frequency behavior of a cascade of low-frequency analogues is obtained by additions and inversions. This latter operation is readily reversed by the method of continued fractions, so that waveguide-element values may be obtained for a large class of rational polynomials in the frequency variable.

For example, in the case of quarter-wave coupled filters, the resonant cavities are equivalent to shunt resonant elements and give rise to the addition operation, while the quarter wavelengths of line give rise to the inversion operation. It is clear that the first statement requires that the input admittance Y_i of a resonant cavity when terminated in an admittance Y_0 may be written

$$Y_i = Y_0 + j4Q\Omega, \tag{1}$$

where Q is a constant and Ω is the frequency variable. Actually, the approximation in (1) is only approximately true since Q is a function of Y_0. The inversion operation given by the quarter wavelength line is true only to the zeroth order. Mumford has shown how a useful first-order approximation is obtained, however, by associating certain resonant elements with the inversion operation.

When it comes to combining these filter elements in cascade, the problem becomes rather delicate. We are asked to add and invert linear functions of Ω and attach significance to the higher power of Ω encountered when all of the underlying approximations are, at best, of the first order. Nevertheless, it appears that this procedure can be justified in the case of quarter-wave coupled filters of sufficiently high Q.

Background

Any waveguide network consisting of known reflecting elements separated by known line lengths may be completely explained by means of conventional formulas. Unfortunately, the numerical effort involved be-

* Decimal classification: R143×R310. Original manuscript received by the Institute, February 18, 1952; revised manuscript received June 2, 1952.

† Microwave Development Laboratories, Inc., Waltham 54, Mass.

[1] W. L. Pritchard, "Quarter wave coupled waveguide filters," *Jour. Appl. Phys.*, vol. XVIII, pp. 862–872; October, 1947.

[2] R. M. Fano and A. W. Lawson, "Microwave filters using quarter wave couplings," Proc. I.R.E., vol. 35, pp. 1318–1323; November, 1947.

[3] J. Hessel, G. Goubau, and L. R. Battersby, "Microwave filter theory and design," Proc. I.R.E., vol. 37, pp. 990–1002; September, 1949.

[4] A. W. Lawson and R. M. Fano, "The Design of Microwave Filters," Microwave Transmissions Circuits, M.I.T. Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., vol. 9, pp. 696–703; 1948.

[5] S. Darlington, "Synthesis of reactance 4-Poles," *Jour. Math. Phys.*, vol. XVIII, pp. 257–353; September, 1939.

[6] W. W. Mumford, "Maximally flat filters in waveguide," *Bell Sys. Tech. Jour.*, vol. XXVII; October, 1948; also pp. 684–713; October, 1938.

[7] U. S. Patent No. 2,585,563; February 12, 1952. The existence of this patent was pointed out to me by S. B. Cohn.

comes prohibitive for synthesis purposes as soon as the number of sections exceeds three or four. Clearly then, any general synthesis procedure will depend on convenient approximations. Because of the importance of appreciating the limitations of these approximations, they will be derived directly from the exact transmission-line formulas without the intervention of equivalent low-frequency circuits.

Basic to this discussion is the admittance transformation effected by a shunt susceptance jB surrounded by equal line lengths l. Using the notation of Hessel *et al.* and putting $\tan\phi = -2/B$, $\theta = 2\pi l/\lambda g$, and $\psi = \phi + 2\theta$, we obtain

$$Y_i = \frac{-j(\cos\psi + \cos\phi) + \sin\psi Y_0}{\sin\psi + j(\cos\phi - \cos\psi)Y_0}, \tag{2}$$

where Y_0 is the terminating admittance and Y_i is the input admittance.

It will be found that the waveguide filters discussed in the previous references are comprised either of resonant elements which at the selected frequency leave the output impedance essentially unchanged, or of antiresonant elements which give an impedance inversion at the selected frequency. In fact, except for shunt resonant elements, all the resonant elements can be further resolved into antiresonant elements. These, then, are the basic building blocks of microwave filters. Only two of these are required to cover the cases. The quarter wavelength of line has already been referred to and the other consists of the shunt susceptance surrounded by a suitable short length of line.

Our immediate problem is that of obtaining the admittance transformations effected by these antiresonant elements. Since, as has already been observed, the representation of a resonant cavity as a shunt-tuned susceptance is not strictly valid even to the first order, that is, to the first power in the frequency variable Ω, our calculations need not be carried beyond this order. The result for the quarter-wave line is obtained from (2) by putting $\phi = \pi/2$ and $2\theta = \pi/2$. Then

$$Y_i = \frac{j - \pi/2\Omega Y_0}{-\pi/2\Omega + jY_0}. \tag{3}$$

Here Ω is the frequency variable $(\lambda g_0 - \lambda_g)/\lambda g$, where λg and λg_0 denote the variable guide wavelength and the guide wavelength at resonance, respectively. The elements in (3) were obtained by taking the first two terms, in the expansions in increasing powers of Ω of the corresponding elements in (2).

The antiresonant element associated with the shunt susceptance jB is obtained from (2) by selecting $\psi = 0$, for $\lambda g = \lambda g_0$. Then the corresponding admittance transformation is given, to the first order in Ω, by

$$Y_i = \frac{j(1 + \cos\phi) + \phi\Omega Y_0}{\phi\Omega + j(1 - \cos\phi)Y_0}. \tag{4}$$

The form of (3) and (4) is characteristic of antiresonant elements.

Quarter-Wave Coupled Filters

Fig. 1 shows a waveguide cavity consisting of two antiresonant shunt susceptances separated by two anti-

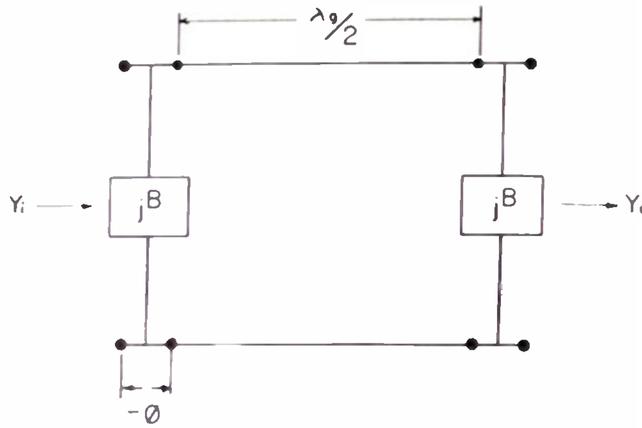


Fig. 1—Schematic resonant cavity.

resonant quarter wavelengths of line. The first-order admittance transformation is readily found to be

$$Y_i = \frac{j\dfrac{\pi(1 + \cos\phi) - 2\phi}{1 - \cos\phi}\Omega + Y_0}{1 + j\dfrac{\pi(1 - \cos\phi) - 2\phi}{1 + \cos\phi}\Omega Y_0}. \tag{5}$$

For high Q cavities, B is large and ϕ and 1-cos ϕ tend to zero. Only then does (5) reduce to the form of (1) with Q, given by

$$Q = \frac{\pi(1 + \cos\phi) - 2\phi}{4(1 - \cos\phi)}. \tag{6}$$

A simpler expression for Q is obtained from (5) by putting $Y_0 = 1$ (it is approximately in the pass band of a filter) and expanding by the binomial theorem. Then

$$Q = \frac{\cos\phi(\pi - \phi)}{\sin^2\phi}. \tag{7}$$

For future reference, a table of values of Q for negative values of B is given in Table I, while Fig. 2 gives ϕ as a function of B.

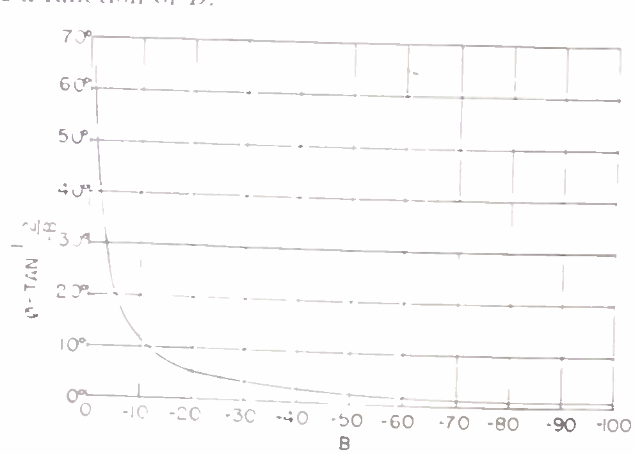


Fig. 2—ϕ versus B.

The quarter wavelength line has the first-order behavior given by (3), which can be rewritten as

$$Y_i = j\frac{\pi}{2}\Omega + \frac{1+\left(\frac{\pi}{2}\Omega\right)^2}{Y_0 + j\frac{\pi}{2}\Omega}. \tag{8}$$

Neglecting the second-order term in Ω, the approximation given by Mumford is obtained in which a variable length quarter-wave (at resonance) line is replaced by a fixed quarter wavelength line terminated in equal resonant elements.

TABLE I

$-B$	Q	$-B$	Q
1	1.14	11	91.35
2	3.33	12	108.74
3	6.91	13	127.89
4	11.97	14	147.73
5	18.59	15	170.42
6	26.76	16	195.70
7	36.45	17	217.64
8	47.88	18	246.93
9	60.71	19	274.54
10	75.18	20	303.27

$$Q = \frac{(\pi-\phi)\cos\phi}{\sin^2\phi}$$

The problem now remains of justifying a continued fraction expansion in increasingly higher powers of Ω when the characteristics of the elements of the filter are obtained only by neglecting all powers of Ω greater than the first.

Consider a section of a quarter-wave coupled filter consisting of a resonant cavity of loaded Q, Q_1 followed by a quarter wavelength of line and another cavity of loaded Q, Q_2. Then the input admittance is given by

$$Y_i = j\left(4Q_2 + \frac{\pi}{2}\right)\Omega + \frac{1+\left[\left(\frac{\pi}{2}\Omega\right)^2\right]}{j\left(4Q_1+\frac{\pi}{2}\right)\Omega + Y_0},$$

where a second-order term to be neglected is placed in the square brackets. In the event of narrow-band filters, where Q would seldom be less than 5, the error in the coefficient of Ω^2 in Y_i due to neglecting $[(\pi/2)\Omega]^2$ is less than 1 per cent in the terms involved, and this approximation is readily justified. Neglecting the terms in $\pi/2$ entirely, however, would give rise to a 15-per cent error in this coefficient. The success of the approximation for the quarter-wave coupled case is seen to depend on the fact that the significant frequency variable terms appear with large coefficients so that higher powers of $Q\Omega$ can not be neglected.

Direct-Coupled Filters

The analysis of direct-coupled filters given by Hessel *et al.* depends on an antiresonant element, readily obtained from (2), which consists of two quarter wavelength lines on each side of an antiresonant shunt element. Its first-order admittance transformation may be written

$$Y_i = \frac{j(1-\cos\phi) + (\phi-\pi)Y_0}{(\phi-\pi)\Omega + j(1+\cos\phi)Y_0}. \tag{9}$$

This may be rewritten

$$Y_i = j\frac{\pi-\phi}{1+\cos\phi}\Omega + \frac{\frac{1-\cos\phi}{1+\cos\phi} + \left[\frac{\Omega(\pi-\phi)}{1+\cos\phi}\right]^2}{Y_0 + j\left(\frac{\pi-\phi}{1+\cos\phi}\right)\Omega}.$$

If we neglect the squared power of Ω, such an antiresonant element can be represented as a quarter wavelength of line of characteristic admittance $\sqrt{1-\cos\phi/1+\cos\phi}$ with equal tuned elements at each end. Accordingly, one might attempt to proceed to a synthesis of direct-coupled filters in this way. Unfortunately, for values of Ω near the half-power points of narrow-band filters, it will be found that $1-\cos\phi$ and the squared term in (15) are of the same order of magnitude so that this procedure does not appear to be justified.

Following a suggestion made by Hessel *et al.*, the synthesis procedure in this article proceeds by constructing a prototype quarter-wave coupled filter, and then constructing an equivalent direct-coupled filter. This depends on the fact that a quarter-wave coupled filter is, in fact, a direct-coupled filter in which the coupling is produced by suitably spaced pairs of susceptances. It turns out that the quarter-wave spacing, together with the excess phase, is just the spacing required to make an antiresonant element of the pair of susceptances. Our problem then is the establishment of the equivalence between a pair of antiresonant susceptive elements, corresponding to ϕ' and ϕ'', sandwiched between three quarter wavelengths of line and a single antiresonant element placed between quarter wavelengths of line. The first-order admittance transformation for the first case is readily found to be

$$Y_i = \frac{j(1-\cos\phi')(1-\cos\phi'') + A\Omega Y_0}{B\Omega + j(1+\cos\phi')(1+\cos\phi'')Y_0}, \tag{10}$$

where

$$A = \phi' + \phi'' + (\pi/2-\phi'')\cos\phi' - (\pi/2-\phi')\cos\phi'' - \frac{3\pi}{2} - \frac{\pi}{2}\cos\phi'\cos\phi''$$

$$B = \phi' + \phi'' - (\pi/2-\phi'')\cos\phi' + (\pi/2-\phi')\cos\phi'' - \frac{3\pi}{2} - \frac{\pi}{2}\cos\phi'\cos\phi''.$$

It is clear from (9) and (10) that we obtain zeroth order equivalence by selecting the single susceptance so that

$$\frac{1-\cos\phi}{1+\cos\phi} = \frac{(1-\cos\phi')(1-\cos\phi'')}{(1+\cos\phi')(1+\cos\phi'')}. \tag{11}$$

An exact first-order equivalence requires that all the corresponding coefficients in (9) and (10), when suitably normalized, be identical. Although this is true in the limit of large susceptances, for the six-cavity filter to be described later, it was found that the coefficients of Ω in (10) were generally 6 or 7 per cent larger than those in (9). This difference is readily taken into account by a change in the frequency variable. For practical design purposes, it is believed that selecting a quarter-wave coupled prototype filter having a few per cent wider bandwidth than ultimately required will serve most engineering purposes.

Illustrative Example

Fig. 3 gives the prototype quarter-wave coupled filter selected to have the insertion gain characteristic, $P_0/P_L = 1 + (65.2\Omega)^{12}$, and the equivalent direct-coupled

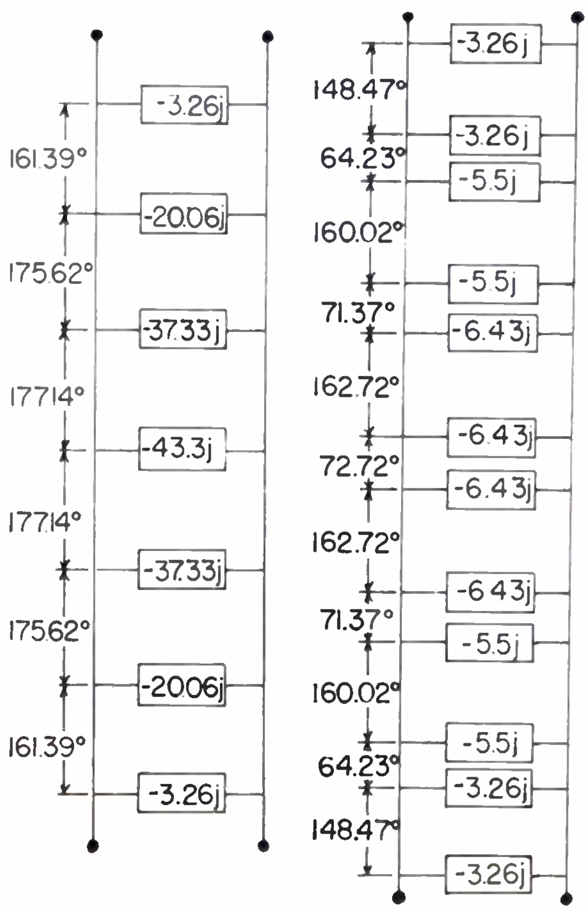


Fig. 3—Prototype and final design.

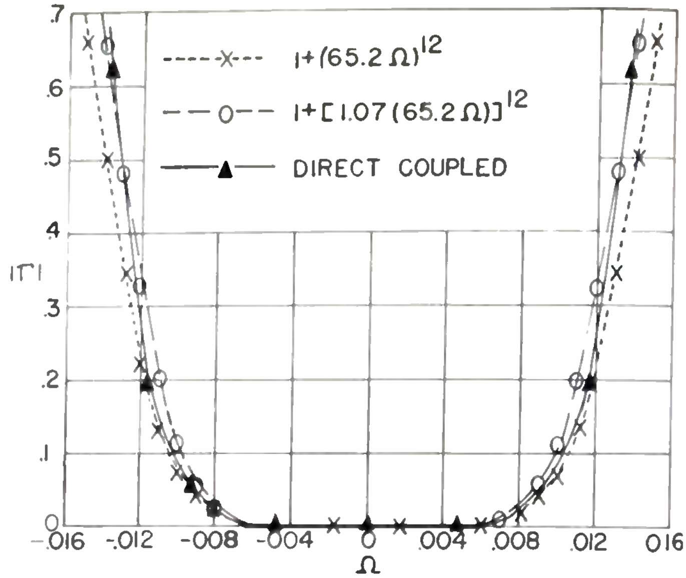


Fig. 4—Computed insertion loss.

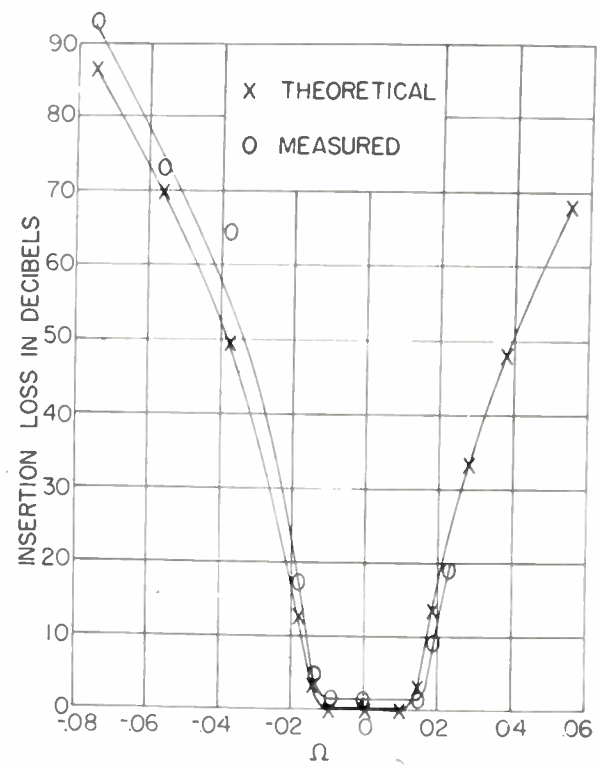


Fig. 5—Measured insertion loss.

filter. Fig. 4 gives a plot of the absolute value of the reflection coefficient of this direct-coupled filter obtained without approximations, using transmission-line formulas. The difference in bandwidth between the direct-coupled filter and its prototype is clearly shown. Figs. 5 and 6 give actual measured behavior of the six-cavity direct-coupled filter compared with a prototype filter having a narrowed bandwidth. It is interesting that this performance was obtained without any elements extraneous to the theory, and without a swept-frequency signal generator.

Concluding Remarks

The general synthesis of direct-coupled narrow-band waveguide filters may be based on the use of a prototype quarter-wave coupled filter of slightly greater bandwidth, without using more restrictive approximations than those underlying the theory of the quarter-wave coupled filter. This depends upon the fact that the two types of filters are very similar in principle. Most differences will resolve themselves down to the relative merits of obtaining a given reflection coefficient with a single susceptance or with plurality suitably spaced susceptances.

Any direct-coupled filter having an odd number of cavities will have a resonant cavity at its center. For a three-cavity direct-coupled maximally flat filter of total Q equal to 50, the Q of the central cavity is 1,852.3. In general, the central cavity of such a filter is of the order of the square of the total Q of the filter. This is an interesting example which illustrates the problems to be encountered in attempting to infer the total Q of a circuit from the Q's of its components.

ACKNOWLEDGMENT

The author wishes to acknowledge the assistance of Mrs. Jean Rockwood who made the calculations and graphs used in this paper and of Mr. T. S. Saad who supervised the construction and testing of the six-cavity filter whose insertion-loss characteristics are shown in Figs. 5 and 6.

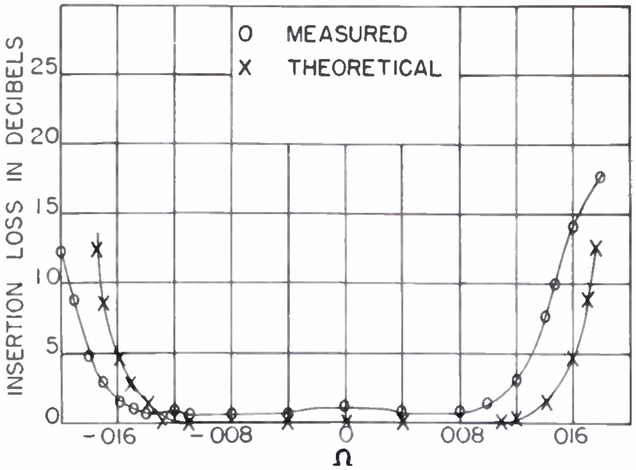


Fig. 6—Measured insertion loss.

Optimum Filters for the Detection of Signals in Noise*

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Summary—**A detection system usually contains a predetection filter whose function is to enhance the strength of the signal relative to that of the noise. An optimum predetection filter is defined in this paper as one which maximizes the "distance" between the signal and noise components of the output (subject to a constraint on the noise component) in terms of a suitable distance function $d(x, y)$. In a special case, this definition leads to the criterion used by North, and yields filters which maximize the signal-to-noise ratio at a specified instant of time. North's theory of such filters is extended to the case of nonwhite noise and finite memory (i.e., finite observation time) filters. Explicit expressions for the impulsive responses of such filters are developed, and two examples of practical interest are considered.**

I. INTRODUCTION

CONSIDERABLE EFFORT has been devoted in recent years to the development of optimum methods of detection of weak signals in noise. Broadly speaking, the purpose of detection is to establish the presence or absence of a signal in noise, or, more generally, to obtain an estimate of a quantity associated with the signal, *e.g.*, the instant of occurrence of a pulse. In general, a detection system comprises a predetection filter whose function is to enhance the strength of the signal relative to that of the noise, and thereby facilitate the detection process.

A special but rather inclusive type of such filters is the main concern of the present paper. As background, a brief review of the published (or available) work on the detection problem is presented in the following.

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In a report published in 1943, North[1] developed a theory of optimum filters—now commonly referred to as "North filters"—based on the maximization of the predetection signal-to-noise ratio. A central result in North's theory is that, in the case of white additive noise, the signal-to-noise ratio is maximized by a filter whose impulsive response has the form of the image of the signal to be detected. A similar result—formulated in terms of so-called "matched" filters—was obtained independently by Van Vleck and Middleton.[2]

More recently, Lee, Cheatham, and Wiesner[3] have described a method of detection of periodic signals based on the use of the correlation analysis. In many respects, the results obtained by this method are essentially equivalent to those obtained by the use of the conventional integration technique, which in turn may be deduced from the theory of North filters. Work along somewhat similar lines has also been reported by Leifer and Marchand.[4]

A more sophisticated approach to detection—closely paralleling the classical Neyman-Pearson theory of testing statistical hypotheses—was initiated by Siegert,[5]

[1] D. O. North, "Analysis of the Factors which Determine Signa /Noise Discrimination in Radar," Report PTR-6C, RCA Laboratories; June, 1943.

[2] J. Van Vleck and D. Middleton, "A theoretical comparison of the visual, aural, and meter reception of pulsed signals in the presence of noise," *Jour. Appl. Phys.*, vol. 17, pp. 940–971; November, 1946.

[3] Y. W. Lee, T. P. Cheatham, Jr., and J. B. Wiesner, "Application of correlation analysis to the detection of periodic signals in noise," PROC. I.R.E., vol. 38, pp. 1165–1172; October, 1950.

[4] M. Leifer and N. Marchand, "The design of periodic radio systems," *Sylvania Technologist*, vol. 3, pp. 18–21; October, 1950. See also, PROC. I.R.E., vol. 39, pp. 1094–1096; September, 1951.

[5] "Threshold Signals," MIT Rad. Lab. Series, McGraw-Hill Book Co., Inc., New York, N. Y., vol. 24, chap. 7; 1950.

and, recently, was further developed by Schwartz.[6] Another statistical theory involving the determination of the *a posteriori* probability distribution of the signal was recently advanced by Woodward and Davies.[7] In many practical situations the usefulness of the latter theory is restricted by the fact that its application requires much more statistical information about the signal and noise than is generally available, while in those cases where the necessary information is available the mathematical computations and the mechanization of the detection process present formidable difficulties. One exception is the case where the noise is additive, white, and Gaussian. In this case the calculations are relatively simple, and the theory leads to the conclusion that the optimum detector consists essentially of a North filter followed by a nonlinear detecting device.

Of the methods mentioned above, the last two are probabilistic in nature, that is, they make use of the probability distributions of the signal and noise. By contrast, detection procedures involving the use of North filters correlation analysis, integration technique, and the like are nonprobabilistic, and hence are inherently much simpler and, in principle, less efficient than the probabilistic procedures.

Despite the theoretical superiority of probabilistic over nonprobabilistic methods, the latter are generally of greater practical utility for two reasons: First, the information about nth-order probability distributions of the signal and noise—which is required by probabilistic methods—is difficult to obtain and to handle for any but the Gaussian type of random function. More important, in many practical cases the pertinent probability distributions are lacking in temporal or spatial stability, or both; in other words, the probability distributions change from day to day or are dependent on the location of the noise source. In such cases, it is clearly unrealistic to base the design of an optimum detector on probability distributions that are assumed to be time and space invariant.

The main advantage of nonprobabilistic methods is that they require relatively little statistical information about the noise—the power spectrum or the correlation function being usually sufficient—and are less critically dependent upon the stability of signal and noise characteristics. Their chief weakness is that they are optimum in only an arbitrary although reasonable sense, and do not make use of such information about the probability distributions as might be available.

The present paper has a twofold purpose: first, to formulate a rather general predetection filtering criterion of a nonprobabilistic type which, through specialization, might be adapted to a wide range of practical cases; and second, using a special form of this criterion, to extend the theory of North filters to the case of nonwhite noise and finite memory (i.e., finite observation time) filters.

In connection with the extension of North's theory, the case of nonwhite noise, it should be noted that such an extension was recently described by Dwork.[8] However, Dwork's results do not resolve the problem of nonwhite noise, since his filters are not, in general, physically realizable. By contrast, the extension described in this paper always leads to physically realizable filters.

II. Criteria of Optimum Filtering

In assessing the performance of filters, predictors, detectors, and many other devices, it is convenient to use a suitable distance function, $d(x, y)$, as a measure of the disparity between two functions $x(t)$ and $y(t)$. For practical purposes, the following three types of distance function are of greatest utility: (The functions $x(t)$ and $y(t)$, appearing below, are assumed to be defined over a long interval of time $(0, T_0)$.)

(a) $d(x, y) = \max \{ | x(t) - y(t) | \}$ = maximum value of the magnitude of the difference between $x(t)$ and $y(t)$.

(b) $$d(x, y) = \frac{1}{T_0} \int_0^{T_0} | x(t) - y(t) | \, dt \tag{1}$$

(c) $$d(x, y) = \left\{ \frac{1}{T_0} \int_0^{T_0} [x(t) = y(t)]^2 dt \right\}^{1/2}. \tag{2}$$

Of these, the distance function of type (c) is of widest applicability, and is also the easiest to handle analytically. It will be recognized as simply the rms value of the difference between $x(t)$ and $y(t)$.

In order to place in evidence the similarities as well as the differences between the criteria of optimum performance for the predetection filter on the one hand, and the conventional[9] filter on the other, it will be helpful to consider first a typical conventional filter F whose input, $u(t)$, consists of the sum of a signal $s_i(t)$ and a noise $n_i(t)$, and whose output, $v(t)$, is required to be as close as possible—in terms of a suitable distance function $d(x, y)$—to the input signal $s_i(t)$. Such a filter may be said to be *optimum* if

$$d[v(t), s_i(t)] = \text{a minimum}; \tag{3}$$

for all $s_i(t)$ in some class $S_i = \{s_i(t)\}$ and all $n_i(t)$ in some class $N_i = \{n_i(t)\}$.

For a distance function of type (c), this formulation of optimum filtering reduces to the familiar minimum mean-square-error criterion. For the linear case, the

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output of F, $v(t)$ consists of the sum of the responses of F to $s_i(t)$ and $n_i(t)$, which are denoted by $s_0(t)$ and $n_0(t)$, respectively. Thus, $v(t)=s_0(t)+n_0(t)$, and, on assuming that $s_i(t)$ and $n_i(t)$ are stationary and independent, (3) reduces to[10]

$$\overline{[s_0(t)+n_0(t)-s_i(t)]^2} = \text{a minimum}, \tag{4}$$

where the bar indicates a long-term time average and the classes S_i and N_i consist of stationary random functions having fixed correlation functions $\Psi_s(\tau)$ and $\Psi_n(\tau)$, respectively.

In the case where F is a predetection filter, the situation is different in that the purpose of F is to facilitate the detection of $s_i(t)$, rather than to reproduce $s_i(t)$. Accordingly, in the case of a predetection filter, it is reasonable to assess the performance in terms of the "distance" between the signal component $s_0(t)$ and the noise component $n_0(t)$ in the output of F.[11] More specifically, by analogy with (3), a predetection filter F will be said to be *optimum* if

$$d[s_0(t), n_0(t)] = \text{a maximum}, \tag{5}$$

for all $s_i(t)$ in some class S_i and all $n_i(t)$ in some class N_i, subject to a constraint[12] on $n_0(t)$ (or $s_0(t)$). The constraint may usually be expressed in terms of the "distance" between $n_0(t)$ and the zero signal. Thus, the quantity to be maximized by F becomes

$$R = d[s_0(t), n_0(t)] - \lambda d[n_0(t), 0] = \text{a maximum}, \tag{6}$$

where λ is a constant (Lagrangian multiplier).

The above criterion is, in principle, sufficiently general to cover a wide variety of practical cases. However, only a few types of distance function can be handled analytically. Of these, the most important is the distance function of type (c), with which the expression for R reduces to

$$R = \overline{[s_0(t)-n_0(t)]^2} - \lambda\overline{n_0^2(t)} = \text{a maximum}. \tag{7}$$

In what follows, attention will be confined to the case where $s_i(t)$ is a signal of known form. It is expedient, then, to replace the time averages in (7) by ensemble averages, with t held constant at a fixed value t_0 (relative to a temporal frame of reference attached to the signal $s_i(t)$). Straightforward calculation yields for this case.

$$R = s_0^2(t_0) - \mu\overline{n_0^2(t)} = \text{a maximum}, \tag{8}$$

where μ is a constant equal to $\lambda-1$, and the bar indicates the time average. (It is tacitly assumed that $n_0(t)$ is ergodic, in which case the ensemble and time averages are identical.)

It is readily seen that when F is a linear filter, (8) is equivalent to maximizing the signal-to-noise ratio

$$\rho = \frac{s_0^2(t_0)}{\overline{n_0^2(t)}} = \text{a maximum}, \tag{9}$$

which is the criterion used in North's theory. Thus, when the distance function is of type (c) and F is linear, the general filtering criterion (6) reduces to the North criterion.

It is also clear that the maximization of R, as expressed by (8), is equivalent to the minimization of

$$Q = \overline{n_0^2(t)} - \lambda s_0(t_0) = \text{a minimum}, \tag{10}$$

where λ is a constant (Lagrangian multiplier). From the mathematical point of view, this is the most convenient form of the predetection filtering criterion, and is the one that will be used in the sequel.

In the following sections, explicit expressions for the impulsive responses of linear, physically realizable, and finite as well as infinite memory predetection filters that are optimum in the sense that they minimize Q (or, equivalently, maximize R and the signal-to-noise ratio ρ) will be developed. The assumptions on the signal and noise are: The signal $s_i(t)$ is a specified but otherwise arbitrary function of time, and the noise $n_i(t)$ is ergodic and has a known correlation function $\Psi_n(\tau)$.

By virtue of the similarity between the predetection filtering criterion (10) and the minimum mean-square-error criterion (4), the mathematics of optimum predetection filters is almost identical with that of optimum filters of the Wiener type. However, instead of following the conventional treatment of Wiener filters, a spectrum-shaping technique which circumvents the use of the calculus of variations and, furthermore, avoids the need for the solution of the Wiener-Hopf equation will be used here. It should be noted that variants of this technique have been employed to considerable advantage in the theory of optimum predictors.[13,14]

III. Determination of the Optimum Filter

The principle of the spectrum-shaping technique is illustrated in Fig. 1. Here $u(t)$ and $v(t)$ represent, respectively, the input and output of a filter F, not necessarily linear, which is optimum in the sense that it maximizes (or minimizes) some quantity Q associated with $v(t)$, on condition that the corresponding input $u(t)$ is a member of a specified class of functions of time, $U=\{u(t)\}$. For convenience in terminology, F will be said to be optimum with respect to the criterion Q and the class of inputs U.

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and, recently, was further developed by Schwartz.[6] Another statistical theory involving the determination of the *a posteriori* probability distribution of the signal was recently advanced by Woodward and Davies.[7] In many practical situations the usefulness of the latter theory is restricted by the fact that its application requires much more statistical information about the signal and noise than is generally available, while in those cases where the necessary information is available the mathematical computations and the mechanization of the detection process present formidable difficulties. One exception is the case where the noise is additive, white, and Gaussian. In this case the calculations are relatively simple, and the theory leads to the conclusion that the optimum detector consists essentially of a North filter followed by a nonlinear detecting device.

Of the methods mentioned above, the last two are probabilistic in nature, that is, they make use of the probability distributions of the signal and noise. By contrast, detection procedures involving the use of North filters correlation analysis, integration technique, and the like are nonprobabilistic, and hence are inherently much simpler and, in principle, less efficient than the probabilistic procedures.

Despite the theoretical superiority of probabilistic over nonprobabilistic methods, the latter are generally of greater practical utility for two reasons: First, the information about nth-order probability distributions of the signal and noise—which is required by probabilistic methods—is difficult to obtain and to handle for any but the Gaussian type of random function. More important, in many practical cases the pertinent probability distributions are lacking in temporal or spatial stability, or both; in other words, the probability distributions change from day to day or are dependent on the location of the noise source. In such cases, it is clearly unrealistic to base the design of an optimum detector on probability distributions that are assumed to be time and space invariant.

The main advantage of nonprobabilistic methods is that they require relatively little statistical information about the noise—the power spectrum or the correlation function being usually sufficient—and are less critically dependent upon the stability of signal and noise characteristics. Their chief weakness is that they are optimum in only an arbitrary although reasonable sense, and do not make use of such information about the probability distributions as might be available.

The present paper has a twofold purpose: first, to formulate a rather general predetection filtering criterion of a nonprobabilistic type which, through specialization, might be adapted to a wide range of practical cases; and second, using a special form of this criterion, to extend the theory of North filters to the case of nonwhite noise and finite memory (i.e., finite observation time) filters.

In connection with the extension of North's theory, the case of nonwhite noise, it should be noted that such an extension was recently described by Dwork.[8] However, Dwork's results do not resolve the problem of nonwhite noise, since his filters are not, in general, physically realizable. By contrast, the extension described in this paper always leads to physically realizable filters.

II. Criteria of Optimum Filtering

In assessing the performance of filters, predictors, detectors, and many other devices, it is convenient to use a suitable distance function, $d(x, y)$, as a measure of the disparity between two functions $x(t)$ and $y(t)$. For practical purposes, the following three types of distance function are of greatest utility: (The functions $x(t)$ and $y(t)$, appearing below, are assumed to be defined over a long interval of time $(0, T_0)$.)

(a) $d(x, y) = \max \{ | x(t) - y(t) | \}$ = maximum value of the magnitude of the difference between $x(t)$ and $y(t)$.

$$\text{(b)}\quad d(x, y) = \frac{1}{T_0}\int_0^{T_0} | x(t) - y(t) | \, dt \tag{1}$$

$$\text{(c)}\quad d(x, y) = \left\{ \frac{1}{T_0}\int_0^{T_0} [x(t) = y(t)]^2 dt \right\}^{1/2}. \tag{2}$$

Of these, the distance function of type (c) is of widest applicability, and is also the easiest to handle analytically. It will be recognized as simply the rms value of the difference between $x(t)$ and $y(t)$.

In order to place in evidence the similarities as well as the differences between the criteria of optimum performance for the predetection filter on the one hand, and the conventional[9] filter on the other, it will be helpful to consider first a typical conventional filter F whose input, $u(t)$, consists of the sum of a signal $s_i(t)$ and a noise $n_i(t)$, and whose output, $v(t)$, is required to be as close as possible—in terms of a suitable distance function $d(x, y)$—to the input signal $s_i(t)$. Such a filter may be said to be *optimum* if

$$d[v(t), s_i(t)] = \text{a minimum}; \tag{3}$$

for all $s_i(t)$ in some class $S_i = \{ s_i(t) \}$ and all $n_i(t)$ in some class $N_i = \{ n_i(t) \}$.

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tion of two linear networks L and L^{-1} which are inverses of one another. (The combination of L and L^{-1} is equivalent to a direct connection.) Now the output of L, which is denoted by $u'(t)$, may be regarded as the input to a composite filter F' which consists of L^{-1} and F. Clearly, if F is optimum with respect to the criterion Q and class of inputs U, then F' is optimum with respect to the criterion Q and class of inputs $U' = \{u'(t)\}$. Consequently, F may be obtained indirectly by first designing a filter F' which is optimum with respect to the criterion Q and class of inputs U' and, then combining F' in tandem with the shaping network L. The advantage of this indirect procedure is that it allows the designer to control the characteristics of the input to F within the limitations imposed by the linearity of the shaping network L. By a proper choice of the shaping network, a complicated optimization problem involving the design of F may, in many cases, be reduced to a simpler problem involving the design of F'.

In applying this approach to optimum filters in the sense of criterion (10), which is equivalent to North's criterion, it is convenient to consider first the relatively simple case in which the filter is not required to have finite memory. This implies that the only condition which the impulsive response must fulfill is that $W(t)$ should vanish for negative t. The case in which the impulsive response is required to vanish outside of a specified interval $0 \leqq t \leqq T$ (finite memory filter) will be considered later in the paper.

Infinite Memory Filters

In this case, it is simplest to use a shaping network L which results in a white noise at the input to F'. The transfer function of such a network may be determined as follows:

If $N(\omega^2)$ is the power spectrum of the input noise $n_i(t)$ and $N'(\omega^2)$ is that of the output of L, then $N'(\omega^2)$ is related to $N(\omega^2)$ by the equation

$$N'(\omega^2) = |H_L(j\omega)|^2 N(\omega^2), \tag{11}$$

where $H_L(j\omega)$ is the transfer function of L. Hence, in order that $N'(\omega^2)$ be a constant, for example unity, it is necessary that

$$|H_L(j\omega)|^2 = \frac{1}{N(\omega^2)}. \tag{12}$$

Expressing $N(\omega^2)$ as the product of two conjugate factors $N_+(j\omega)$ and $N_+^*(j\omega)$ so that $N_+(j\omega)$ and its reciprocal are regular in the right-half of the $j\omega$-plane, it is seen that if $H_L(j\omega)$ is set equal to the reciprocal of $N_+(j\omega)$

$$H_L(j\omega) = \frac{1}{N_+(j\omega)}, \tag{13}$$

then $N'(\omega^2) = 1$. Thus, a shaping network L whose transfer function is given by (13) results in a white noise at the input to F'.

The calculation of $N_+(j\omega)$ is a straightforward algebraic problem, since, in practice, $N(\omega^2)$ is generally of the form

$$N(\omega^2) = \frac{A(\omega^2)}{B(\omega^2)} = \frac{a_0 + a_1\omega^2 + \cdots + a_l\omega^{2l}}{b_0 + b_1\omega^2 + \cdots + b_m\omega^{2m}}, \tag{14}$$

where $A(\omega^2)$ and $B(\omega^2)$ are polynomials in ω^2, and m and l rarely exceed 3. Correspondingly, $N_+(j\omega)$ is of the form

$$N_+(j\omega) = \frac{A_+(j\omega)}{B_+(j\omega)}, \tag{15}$$

where $A_+(j\omega)$ and $B_+(j\omega)$ are real polynomials in $j\omega$ of degrees l and m, respectively. Thus, $A_+(j\omega)A_+^*(j\omega) = A(\omega^2)$ and $B_+(j\omega)B_+^*(j\omega) = B(\omega^2)$, and neither $A_+(j\omega)$ nor $B_+(j\omega)$ has zeros in the right half of the $j\omega$-plane. In terms of these polynomials, $H_L(j\omega)$ reads

$$H_L(j\omega) = \frac{B_+(j\omega)}{A_+(j\omega)}. \tag{16}$$

A simple example will serve to illustrate these relations. Consider a noise whose power spectrum is

$$N(\omega^2) = \frac{\omega^2}{\omega^2 + a^2}. \tag{17}$$

For this case,

$$N_+(j\omega) = \frac{j\omega}{j\omega + a}, \tag{18}$$

and consequently the transfer function of the shaping network is

$$H_L(j\omega) = \frac{j\omega + a}{j\omega}. \tag{19}$$

Now let $W(t)$ and $W'(t)$ be the impulsive responses of F and F', respectively. Referring to Fig. 1, it is seen that $W(t)$ is related to $W'(t)$ by the operational equation

$$W(t) = H_L(p)W'(t). \tag{20}$$

Since $H_L(p)$ is specified by (13), the determination of $W(t)$ is reduced essentially to finding the expression for $W'(t)$. This is a relatively simple problem, as the following analysis indicates:

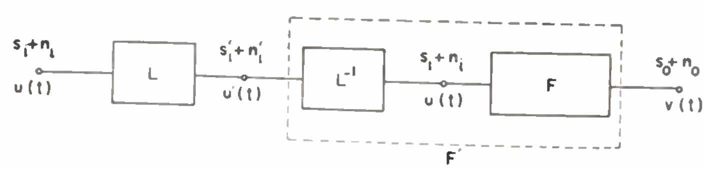

Fig. 1—Principle of spectrum-shaping technique.

Let $s_i'(t)$ and $n_i'(t)$ denote signal and noise components of the input to F'. As a consequence of setting the transfer function of L equal to $1/N_+(j\omega)$, the noise component $n_i'(t)$ is white noise [$N'(\omega^2) = 1$] while the signal component $s_i'(t)$ is related to $s_i(y)$ (signal component of the input to F) by the operational relation

$$s_i'(t) = H_L(p)s_i(t) = \frac{1}{N_+(p)} s_i(t). \tag{21}$$

So far as F' is concerned, the problem is simply that of determining the impulsive response of an optimum filter in the case where the input noise is white. Thus, the quantity to be minimized by F' is

$$Q = \sigma^2 - \lambda s_0(t_0), \tag{22}$$

where σ^2 denotes $\overline{n_0^2(t)}$, i.e., the mean-square value of the output noise; $s_0(t)$ is the response of F' to $s_i'(t)$; t_0 is a specified instant of time; and λ is an arbitrary constant. In terms of $W'(t)$, the expressions for σ^2 and $s_0(t_0)$ are

$$\sigma^2 = \int_0^\infty [W'(t)]^2 dt \tag{23}$$

and

$$s_0(t_0) = \int_0^\infty W'(t)s_i'(t_0 - t)dt. \tag{24}$$

Substituting these in (22) yields

$$Q = \int_0^\infty \{[W'(t)]^2 - \lambda W'(t)s_i'(t_0 - t)\} dt. \tag{25}$$

The minimization of Q can easily be achieved without the use of variational techniques by expressing the integrand in (25) as the difference of a perfect square and a constant. Thus, completing the square and setting $\lambda = 2$ (for convenience) yields

$$Q = \int_0^\infty [W'(t) - s_i'(t_0 - t)]^2 dt - \int_0^\infty [s_i'(t_0 - t)]^2 dt. \tag{26}$$

Since the second term on the right is a constant, Q is a minimum when

$$W'(t) = s_i'(t_0 - t), \quad t \geqq 0. \tag{27}$$

In other words, the impulsive response of F is identical for positive t, with the image of $s_i'(t)$ with respect to $t = t_0/2$. This, as should be expected, is in agreement with the classic result of North's theory in the white noise case.

Once $W'(t)$ is determined, the expression for $W(t)$ can easily be found through the use of (20). Thus, introducing the unit step function $1(t)$ (in order to indicate that $W(t)$ vanishes for $t<0$) and using (20), one obtains the operational relation

$$W(t) = \frac{1}{N_+(p)} 1(t)s_i'(t_0 - t), \tag{28}$$

where $s_i'(t)$ is given by (21) and $N_+(p)$ is given by (15). This relation constitutes a general expression for the impulsive response of a *physically realizable* optimum filter in the nonwhite noise case.

A more explicit expression for the impulsive response of the optimum filter can readily be found by expressing $s_i'(t)$ in terms of its Fourier transform

$$s_i'(t) = \int_{-\infty}^\infty \frac{S(j\omega)}{N_+(j\omega)} e^{j\omega t} df, \tag{29}$$

where $S(j\omega)$ is the Fourier transform of $s_i(t)$. Substituting this in (28) yields the impulsive response

$$W(t) = \frac{1}{N_+(p)} 1(t) \int_{-\infty}^\infty \frac{S^*(j\omega)}{N_+^*(j\omega)} e^{j\omega(t-t_0)} df. \tag{30}$$

From this, the expression for the transfer function of F is found to be

$$H(j\omega) = \frac{1}{N_+(j\omega)} \int_0^\infty dt\, e^{-j\omega t} \int_{-\infty}^\infty df' \frac{S^*(j\omega')e^{j\omega'(t-t_0)}}{N_+^*(j\omega')}, \tag{31}$$

where

ω' = variable of integration, $f' = \omega'/2\pi$;

$(\ \ )^*$ = complex conjugate of $(\ \ )$;

$S(j\omega)$ = Fourier transform of the signal $s_i(t)$ (at the input to F);

$N(\omega^2)$ = power spectrum of the noise $n_i(t)$ (at the input to F);

$N_+(j\omega)$ = factor of $N(\omega^2)$ which, together with its reciprocal, is regular in the right half of the $j\omega$-plane, and is such that $N_+(j\omega)N_+^*(j\omega) = N(\omega^2)$.

Equation (31) is the desired expression for the transfer function of a linear, infinite memory, and physically realizable filter that minimizes Q and also maximizes R and the signal-to-noise ratio ρ at $t = t_0$.

It will be noted that if F is not required to be physically realizable, $W(t)$ need not vanish for $t<0$, and consequently the lower limit in the first integral in (31) should be $-\infty$. Then, (31) reduces to

$$H(j\omega) = \frac{S^*(j\omega)e^{-j\omega t_0}}{N(\omega^2)}, \tag{32}$$

which is identical with the expression given by Dwork.[8]

Finite Memory Filters

In the finite memory case, the situation is complicated somewhat by the requirement that $W(t)$ should vanish not only for $t<0$ but also for $t>T$, where T is a specified constant. Since $W(t)$ is related to $W'(t)$ by the equation

$$W(t) = H_L(p)W'(t), \tag{33}$$

the impulsive response of F' is constrained to be such that $H_L(p)W'(t) = 0$ for $t>T$. This constraint is responsible for the complications arising in the process of optimization of F'.

In the infinite memory case just treated, it was found expedient to set the transfer function of the shaping network L equal to the reciprocal of $N_+(j\omega)$, which results in a white noise at the input to F'. The same choice in the finite memory case, however, would make it rather difficult to take into account the constraint imposed on $W'(t)$. Analysis of possible choices for $H_L(j\omega)$ indicates that, in the finite memory case, it is expedient to set $H_L(j\omega)$ equal to the denominator of $N_+(j\omega)$. In other words,

$$H_L(j\omega) = B_+(j\omega), \tag{34}$$

where $B_+(j\omega)$ is defined as in (15). With this expression for the transfer function, the power spectrum of the noise at the input to F' assumes the following form

$$N'(\omega^2) = A(\omega^2) = a_0 + a_1\omega^2 + \cdots + a_l\omega^{2l}, \tag{35}$$

where $A(\omega^2)$ is the numerator of $N(\omega^2)$ (see (14)). Also, the relation between $W(t)$ and $W'(t)$ becomes

$$W(t) = B_+(p)W'(t). \tag{36}$$

In view of this relation, the requirement that $W(t)=0$ for $t>T$ imposes the following constraint on $W'(t)$:

$$B_+(p)W'(t) = 0 \quad \text{for} \quad t > T, \tag{37}$$

which means that, for $t>T$, $W'(t)$ must be a solution of the differential equation $B_+(p)W'(t)=0$.

Turning to the determination of $W'(t)$, one has to express σ^2 and $s_0(t_0)$ in terms of $W'(t)$ and the noise and signal components of the input to F'. If $s_i(t)$ is the signal at the input to F, then the signal at the input to F' is

$$s_i'(t) = B_+(p)s_i(t), \tag{38}$$

and correspondingly at the output of F'

$$s_0(t_0) = \int_0^\infty W'(t)s_i'(t_0 - t)dt. \tag{39}$$

The expression for σ^2 is readily obtained by noting that a term in $N'(\omega^2)$ of the form $a_k\omega^{2k}$ results in a mean-square value component

$$\sigma_k^2 = \int_0^\infty a_k\left[\frac{d^kW'(t)}{dt^k}\right]^2 dt. \tag{40}$$

Hence $N'(\omega^2)$, being the sum of such terms, results in

$$\sigma^2 = \int_0^\infty \left\{a_0[W'(t)]^2 + a_1\left[\frac{dW'(t)}{dt}\right]^2 + \cdots + a_l\left[\frac{d^lW'(t)}{dt^l}\right]^2\right\} dt, \tag{41}$$

which is the desired expression for σ^2.

On substituting (39) and (41) in (22) and restricting the range of integration to $0 \leqq t \leqq T$, the expression for the quantity to be minimized, a, is found to be

$$Q = \int_0^T \left\{a_0[W'(t)]^2 + a_1\left[\frac{dW'(t)}{dt}\right]^2 + \cdots + a_l\left[\frac{d^lW'(t)}{dt^l}\right]^2 - \lambda W'(t)s_i'(t_0 - t)\right\} dt. \tag{42}$$

The determination of a function $W'(t)$ which minimizes this expression is carried out in the Appendix. Once $W'(t)$ has been determined, the impulsive response $W(t)$ of the optimum filter F can be found from the relation

$$W(t) = B_+(p)W'(t), \tag{43}$$

where $B_+(p)$ is defined by (15). The resulting expression for $W(t)$ is given below ($W(t)=0$ outside of the interval $0 \leqq t \leqq T$):

$$W(t) = \frac{1}{N_+(p)} 1(t) \int_{-\infty}^{\infty} \frac{S^*(j\omega)e^{j\omega(t-t_0)}}{N_+^*(j\omega)} df + \sum_{\nu=1}^{2l} A_\nu e^{\alpha_\nu t} + \sum_{\mu=0}^{m-l-1} B_\mu\delta^{(\mu)}(t) + \sum_{\mu=0}^{m-l-1} C_\mu\delta^{(\mu)}(t - T), \tag{44}$$

where, to recapitulate,

$W(t)$ = impulsive response of the optimum filter F;
$N(\omega^2)$ = power spectrum of the noise component of the input to F;
$N_+(j\omega)$ = a factor of $N(\omega^2)$ so that $N_+(j\omega)$ and $1/N_+(j\omega)$ are regular in the right half of the $j\omega$-plane, and $|N_+(j\omega)|^2 = N(\omega^2)$;
$(\quad)^*$ = complex conjugate of $(\quad)$;
$S(j\omega)$ = Fourier transform of the signal component of the input to F;
t_0 = a specified instant of time (relative to the signal);
$2l$ = degree of the numerator of $N(\omega^2)$;
$2m$ = degree of the denominator of $N(\omega^2)$;
A_ν, B_μ, C_μ = undetermined coefficients;
α_ν = roots of the equation $A(-p^2)=0$, where $A(\omega^2)$ is the numerator of $N(\omega^2)$;
T = settling time (length of memory);
$1(t)$ = unit step function;
$\delta(t)$ = unit impulse function;
$\delta^{(\nu)}(t)$ = νth derivative of $\delta(t)$.

In this expression, the terms $\sum_{\nu=1}^{2l} A_\nu \exp(\alpha_\nu t)$ represent the general solution of the differential equation $A(-p^2)W(t)=0$. The terms involving impulse functions of various orders arise from operating with $B_+(p)$ (see (43)) on the discontinuities of $W'(t)$ and its derivatives at $t=0$ and $t=T$. The first term in (44) may be written in a somewhat different but equivalent form which is sometimes advantageous.

$$\text{First term} = \frac{1}{N_+(p)} 1(t)s_i'(t_0 - t), \tag{45}$$

where

$$s_i'(t) = \frac{1}{N_+(p)} s_i(t). \tag{46}$$

There remains the question of the undetermined coefficients A_ν, B_μ, and C_μ. The steps leading to the determination of these coefficients can best be formulated by going back to the minimization of the quantity Q,

$$Q = \sigma^2 - \lambda s_0(t_0), \tag{47}$$

which is achieved with the optimum filter. Previously, this quantity was expressed in terms of $W'(t)$, i.e., the impulsive response of F'. Now it will be necessary to express Q directly in terms of $W(t)$.

The expression for σ^2 in terms of $W(t)$ reads[15]

$$\sigma^2 = \int_0^T \int_0^T W(t)W(\tau)\psi_n(t - \tau)dtd\tau, \tag{48}$$

where $\psi_n(\tau)$ is the correlation function of the noise component of the input to F. Similarly, the expression for $s_0(t_0)$ is

$$s_0(t_0) = \int_0^T W(t)s_i(t_0 - t)dt, \tag{49}$$

where $s_i(t)$ is the signal component of the input to F. Substituting these expressions in (47) gives

$$Q = \int_0^T \int_0^T W(t)W(\tau)\psi_n(t - \tau)dtd\tau - \lambda \int_0^T W(t)s_i(t_0 - t)dt. \tag{50}$$

On applying standard variational formulas, it is found that Q is minimized by a $W(t)$, which satisfies the following integral equation:

$$\int_0^T W(\tau)\psi_n(t - \tau)d\tau = s_i(t_0 - t), \quad 0 \leqq t \leqq T. \tag{51}$$

Thus, the impulsive response of F is the solution of integral (51), whose kernel is the correlation function of $n_i(t)$ (the noise component of the input to F), and whose right-hand member is the image of the signal component $s_i(t)$ with respect to $t = t_0/2$. It is of interest to note that when $T = \infty$ this integral equation reduces to the Wiener-Hopf equation which is encountered in Wiener's theory of prediction.[16] (Equation (51) is similar in form to that encountered in an extension of Wiener's theory described in footnote reference 14.)

Now the expression for $W(t)$ obtained previously (see (44)) is, in effect, the solution of the integral (51). Consequently, the undetermined coefficients in (44) may be determined by substituting $W(t)$, as given by (44), into (51) and treating the resulting equation as an identity. This procedure will be illustrated by an example treated in the next section.

By using the fact that $W(t)$ is the solution of (51), it is possible to obtain a simple expression for the signal-to-noise ratio ρ at the output of the optimum filter. Thus, writing σ^2 in the form

$$\sigma^2 = \int_0^T dtW(t) \int_0^T W(\tau)\psi_n(t - \tau)d\tau \tag{52}$$

and noting that $W(t)$ satisfies the integral equation

$$\int_0^T W(\tau)\psi_n(t - \tau)d\tau = s_i(t_0 - t), \quad 0 \leqq t \leqq T,$$

one obtains

$$\sigma^2 = \int_0^T W(t)s_i(t_0 - t)dt, \tag{53}$$

which, in view of (49), is numerically equal to $s_0(t_0)$. (Note that, when $W(t)$ is the solution of (51), $s_0(t_0)$ is a positive quantity.) This implies that *the mean-square value of the noise output of the optimum filter is numerically equal to the signal output at* $t = t_0$. (The numerical equality does not hold unless $W(t)$ is the solution of (51).)

Now the general expression for ρ is $\rho = |s_0(t_0)|^2/\sigma^2$. Making use of the fact that for the optimum filter $\sigma^2 = s_0(t_0)$, the expression for the signal-to-noise ratio at the output of the optimum filter becomes

$$\rho_{max} = s_0(t_0) = \sigma^2. \tag{54}$$

It is easily verified that, if the filter is not required to be physically realizable, (54) reduces to the expression for ρ_{max} given by Dwork.[8]

IV. Illustrative Examples

Two examples of practical interest will serve to illustrate the theory: First, suppose that $s_i(t)$ is a periodic signal of period T_0, consisting of a train of rectangular pulses of unit height and width d. The settling time is assumed to be equal to an integral multiple of T_0, i.e., $T = kT_0$. The instant t_0 is specified as

$$t_0 = (k - 1)T_0 + d. \tag{55}$$

In other words, the signal-to-noise ratio is to be maximized at the instant immediately following the occurrence of the training edge of the kth pulse. The power spectrum of noise is assumed to be of the form

$$N(\omega^2) = \frac{1}{\omega^2 + a^2}\cdot \tag{56}$$

The first step in determining the optimum filter is to form the expression for $N_+(j\omega)$. For the specified $N(\omega^2)$, $N_+(j\omega)$ reads

$$N_+(j\omega) = \frac{1}{j\omega + a}\cdot \tag{57}$$

Next, it is noted that $2p = 0$ ($2l$ = degree of the numerator of $N(\omega^2)$) and $2m = 2$ ($2m$ = degree of the denominator of $N(\omega^2)$). Thus $m - l - 1 = 0$, and therefore all terms in (44) involving undetermined coefficients are zero. Consequently, $W(t)$ is given by

$$W(t) = (p + a)1(t) \int_{-\infty}^{\infty} (a - j\omega)S^*(j\omega)e^{j\omega(t-t_0)}df, \quad 0 \leqq t \leqq T; \tag{58}$$

or more simply, (using (45)),

[15] H. M. James, N. B. Nichols, and R. S. Phillips, "Theory of Servomechanisms," Rad. Lab. Series, McGraw-Hill Book Co., New York, N. Y., vol. 24, chap. 6; 1947.

[16] N. Wiener, "The Extrapolation, Interpolation, and Smoothing of Stationary Time Series," John Wiley and Sons, Inc., New York, N. Y.; 1949.

$$W(t) = (p + a)1(t)s_i'(t_0 - t), \qquad 0 \leqq t \leqq T, \quad (59)$$

where

$$s_i'(t) = (p + a)s_i(t). \quad (60)$$

From (59) and (60), the impulsive response of the optimum filter is found to be expressed by

$$W(t) = (a^2 - p^2)s_i(t), \quad 0 \leqq t \leqq T, \quad (61)$$

where $s_i(t)$ is the specified pulse train. The form of $W(t)$ is shown in Fig. 2.

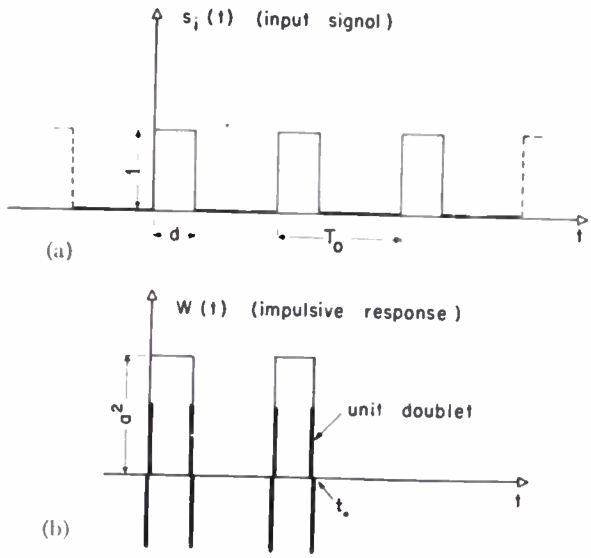

Fig. 2—(a) Form of the signal $s_i(t)$. (b) Form of the impulsive response of the optimum filter.

In the case under consideration, it is worth while to derive the expression for the transfer function $H(j\omega)$ of the optimum filter. From (61), it is readily found that $H(j\omega)$ is given by

$$H(j\omega) = \left(\frac{a^2 + \omega^2}{j\omega}\right)(1 - e^{-j\omega d})(1 + e^{-j\omega T_0} + \cdots + e^{-j\omega(k-1)T_0}); \quad (62)$$

or more simply,

$$H(j\omega) = \left(\frac{a^2}{j\omega} - j\omega\right)(1 - e^{-j\omega d})\left[\frac{1 - e^{-j\omega kT_0}}{1 - e^{-j\omega T_0}}\right]. \quad (63)$$

The bracketed term in this expression represents the transfer function of an ideal "integrator."[17] Thus, the optimum filter consists of a tandem combination of a filter F_1 with transfer function

$$H_1(j\omega) = \left(\frac{a^2}{j\omega} - j\omega\right)(1 - e^{-j\omega d}), \quad (64)$$

and an ideal "integrator" F_2 with transfer function

$$H_2(j\omega) = 1 + e^{-j\omega T_0} + e^{-j2\omega T_0} + \cdots + e^{-j(k-1)\omega T_0}. \quad (65)$$

[17] J. V. Harrington and T. F. Rogers, "Signal-to-Noise improvement through integration in a storage tube," Proc. I.R.E., vol. 38, pp. 1197–1203; October, 1950.

It is of interest to note that the optimum filter has, in general, this structure (that is, a filter followed by an "integrator") whenever the signal $s_i(t)$ is periodic and the power spectrum function $N(\omega^2)$ has a constant for the numerator.

Second Example

In this case the signal $s_i(t)$ is assumed to consist of a single rectangular pulse of unit amplitude and width d. The spectral density function is of the form

$$N(\omega^2) = \frac{\omega^2}{\omega^2 + a^2}. \quad (66)$$

The signal-to-noise ratio is to be maximized at the instant of occurrence of the trailing edge of the pulse. The impulsive response is required to vanish outside of the interval $0 \leqq t \leqq T$.

Following the same procedure as in the preceding example, one finds

$$N_+(j\omega) = \frac{j\omega}{j\omega + a}. \quad (67)$$

Since $2l = 2m = 2$, the unit impulse terms in (44) are zero. The second term, which is the general solution of the differential equation $-p^2W(t) = 0$, is of the form

$$A_0 + A_1 t. \quad (68)$$

The first term in (44) may be written as (see (86))

$$\text{first term} = -\frac{(p + a)}{p^2} 1(t)s_i'(t_0 - t), \quad (69)$$

where

$$s_i'(t) = (p + a)s_i(t) \quad (70)$$

and

$$s_i(t) = 1(t) - 1(t - d). \quad (71)$$

Calculation of this term yields

$$\begin{aligned} \text{first term} &= 1 - \frac{a^2t^2}{2} && \text{for } t \leqq d \\ &= -a^2dt + \frac{a^2d^2}{2} && \text{for } t > d. \end{aligned} \quad (72)$$

Thus the complete expression for $W(t)$ is

$$\begin{aligned} W(t) &= 1 + A_0 + A_1 t - 0.5a^2t^2 && \text{for } 0 \leqq t \leqq d \\ &= A_0 + 0.5a^2d^2 + (A_1 - a^2d)t && \text{for } d < t \leqq T. \end{aligned} \quad (73)$$

It remains to calculate the undetermined coefficients A_0 and A_1. For this purpose, it is necessary to set up the integral (51), of which $W(t)$ is the solution. The kernel of this equation is the correlation function $\Psi_n(\tau)$ of the noise component $n_i(t)$. Using the fact that $\Psi_n(\tau)$ is the inverse Fourier transform of $N(\omega^2)$ (Wiener-Khintchine relation), one readily finds

$$\psi_n = \delta(\tau) = 0.5ae^{-a|\tau|}. \quad (74)$$

Hence, for the case under consideration the integral equation reads

$$\int_0^T W(\tau)[\delta(t-\tau) - 0.5ae^{-a|t-\tau|}]d\tau = 1(t) - 1(t-d), \quad 0 \leqq t \leqq T. \quad (75)$$

Substituting $W(t)$ as expressed by (73) into this equation and requiring that $W(t)$ be the solution of (75), yields two linear equations in A_0 and A_1 which, upon solution, give

$$A_0 = \frac{ad(2aT + 2 - ad)}{2(aT + 2)} \quad (76)$$

and

$$A_1 = \frac{a^2d(2aT + 2 - ad)}{2(aT + 2)}. \quad (77)$$

Substituting these values in (73) yields

$$\begin{aligned} W(t) &= 1 + \frac{ad(2aT + 2 - ad)}{2(aT + 2)} + \frac{a^2d(2aT + 2 - ad)t}{2(aT + 2)} - 0.5a^2t^2 \quad \text{for } 0 \leqq t \leqq d \\ &= \frac{ad(aT + 1)(ad + 2)}{2(aT + 2)} - \frac{a^2d(ad + 2)t}{2(aT + 2)} \quad \text{for } d < t \leqq T. \quad (78) \end{aligned}$$

This is the desired expression for the impulsive response of the optimum filter. A plot of $W(t)$ for $T = 1$ msec, $d = 100$ μsec and $a = 20{,}000$ sec^{-1}, is shown in Fig. 3. It will be noted that $W(t)$, as given by (78), bears

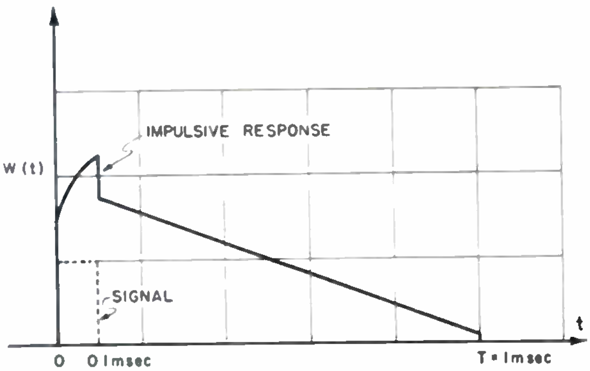

Fig. 3—Impulsive response of the optimun filter for example 2. The signal $s_i(t)$ in this case is a rectangular pulse (indicated by broken lines).

little if any resemblance to the impulsive response suggested by the "matched-filter" formula, namely $W(t) = s_i(t_0 - t)$. This implies that a "matched" filter may be far from optimum in a situation wherein the noise is not white and the filter is required to have a finite memory.

Appendix

The application of a standard variational formula[18] to (42) leads to the differential equation

$$A(-p^2)W'(t) = s_i'(t_0 - t), \quad 0 \leqq t \leqq T, \quad (79)$$

of which $W'(t)$ is a solution. In this equation $A(-p^2)$ represents the numerator of $N(\omega^2)$ with ω^2 replaced by $-p^2$, and

$$s_i'(t) = B_+(p)s_i(t), \quad (80)$$

where $B_+(p)$ is defined in (15).

$W'(t)$ may be written as the sum of two terms, $W_P'(t)$ and $W_G'(t)$, of which $W_P'(t)$ represents a particular solution of (79), while $W_G'(t)$ is the general solution of the homogeneous equation

$$A(-p^2)W'(t) = 0. \quad (81)$$

The general solution $W_G'(t)$ is given by

$$W_G'(t) = \sum_{\nu=1}^{2l} A_\nu' e^{\alpha_\nu t} \quad (82)$$

where the A_ν' are arbitrary constants and the α_ν are the roots of the characteristic equation $A(-p^2) = 0$. (If α_ν is a multiple root of order k, then A_ν' is a polynomial of $(k-1)$st degree in t.)

Since a particular solution is not unique, $W_P'(t)$ may be written in various forms which differ between themselves by terms of the form $A_\nu' e^{\alpha_\nu t}$. Two of the more convenient expressions for $W_P'(t)$ are

$$\begin{aligned} (1) \quad W_P'(t) &= \frac{1}{A_+(p)} 1(t) \frac{1}{A_+(-p)} s_i'(t_0 - t) \\ &= \frac{1}{A_+(p)} 1(t) \int_{-\infty}^{\infty} \frac{S^*(j\omega)}{A_+^*(j\omega)} B_+^*(j\omega) e^{j\omega(t-t_0)} df \end{aligned} \quad (83)$$

and

$$(2) \quad W_P'(t) = \frac{}{A(-p^2)} 1(t)s_i'(t_0 - t). \quad (84)$$

Using the first of these expressions, $W'(t)$ reads

$$W'(t) = \frac{1}{A_+(p)} 1(t) \int_{-\infty}^{\infty} \frac{S^*(j\omega)}{A_+^*(j\omega)} B_+^*(j\omega) e^{j\omega(t-t_0)} df + \sum_{\nu=1}^{2l} A_\nu' e^{\alpha_\nu t}. \quad (85)$$

Substituting this in (43) and replacing $A_+(j\omega)/B_+(j\omega)$ by $N_+(j\omega)$, one obtains the expression for $W(t)$ given by (44). The impulsive terms in (44) arise from operating with $B_+(p)$ on the discontinuities of $W'(t)$ and its derivatives at $t = 0$ and $t = T$.

When (84) rather than (83) is used to represent $W_P'(t)$, the first term in (44) is replaced by

$$\text{first term} = \frac{B(p)}{A(-p^2)} 1(t)s_i'(t_0 - t). \quad (86)$$

(This change affects only the undetermined coefficients A_ν). In the case of the second example in section 4, this form of the first term is more convenient to work with than that appearing in (44).

[18] Courant-Hilbert, "Methoden Der Mathematischen Physik," Interscience Publishers, Inc., New York, N. Y., vol. I, p. 163; 1931.

The Detection of a Sine Wave in the Presence of Noise by the Use of a Nonlinear Filter*

THOMAS G. SLATTERY†

Summary—**This paper reviews the basic theory of the design of nonlinear filters to detect signals in the presence of noise or clutter. It then describes the exact manner of how a nonlinear filter may be instrumented to detect the presence of a sine wave of unknown frequency in random noise.**

I. The Theory of the Design of Nonlinear Filters for the Detection Problem

A. Introduction

Statistical techniques are being applied to an ever growing number of engineering problems. To demonstrate the logical manner in which these techniques combine engineering design specifications and *a priori* information, the general problem of detecting a signal in the presence of noise will be considered. Many papers,[1,2,3,4,5] have recently been written concerning this problem. A résumé of the theory will be presented.

B. Design Equation of Nonlinear Filter for the Detection Problem

In the detection problem the question is asked whether a given time series is part of a series composed of signal plus noise or whether this time series is noise only. Call the time series of signal plus noise $S(t)$ and the time series of noise only $N(t)$. In any physical problem the bandwidth of the given time series is limited, and therefore there is a top frequency present in $S(t)$ and $N(t)$. It then follows that $S(t)$ and $N(t)$ can be completely described in the time interval

$$0 \leqq t \leqq t_1$$

by $2Wt_1$ values, where W is the top frequency present.

We need only deal, therefore, with a finite number of values of the incoming data; since it is not known whether the incoming data arise from the $S(t)$ time series or the $N(t)$ time series, let us represent these values by $X_1, X_2, \cdots, X_m$, where $m = 2Wt_1$. Now in order to determine whether this series of numbers $X_1 \cdots X_m$ arises from $S(t)$ or $N(t)$, we must know something of the statistical nature of these processes. In general, the functions which completely define[6] a random process of the kind we are dealing with are the expressions $P_s(X_1, X_2, \cdots, X_m)$, the joint density function of the signal plus noise, and $P_n(X_1, X_2 \cdots X_m)$, the joint density function of the noise alone.

The problem then becomes one of determining whether the sample of m members arises from a population whose sampling distribution is $P_s(X_1 \cdots X_m)$ or whether it arises from a population whose sampling distribution is $P_n(X_1 \cdots X_m)$.

We shall now define the best way of determining whether the sample $X_1 \cdots X_m$ originated from the time series of signal plus noise or the time series of noise only. First we must define a criterion of best. We can make two kinds of errors, known to statisticians as type I and type II errors.[7] We can make the error of calling noise a signal and the error of calling a signal noise, that is, missing the signal altogether. These errors are associated, respectively, with the terms "false-alarm" probability and probability of nondetection.

We shall therefore call the system of detection "best" which in the long run will hold fixed the false-alarm probability and will minimize the probability of missing the signal.

To fix ideas, we will now introduce a very convenient pictorial representation of the process, first given by Neyman and Pearson.[8,9] For simplicity, we assume that there are only two members X_1 and X_2 in the sample and from these two members X_1 and X_2, which are two values of the incoming signal, we are to determine whether the signal arises from noise alone or contains a target return. Now, if we were to take a long series of such pairs of observations in a real situation, we might get from one process a set of points as plotted in Fig. 1. The crosses denote $S_2(t)$ and the O's denote $N_2(t)$, each point representing, then, a pair of observations. The cluster of O's around the origin will have a density which is exactly equal to $P_n(X_1, X_2)$, and the cluster of crosses will have a density which is equal to $P_s(X_1, X_2)$. Now the

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[1] H. Sullivan, "Proposal for the Study of Non-Linear Prediction, Interpolation and Extrapolation of Time Series," unpublished. (Former Director of Research, Avion Instrument Corp., Paramus, N. J.)

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[5] M. Schwartz, "A Statistical Approach to the Automatic Search Problem," Doctoral Dissertation (Harvard); June, 1951.

[6] H. M. James, N. B. Nichols, and R. S. Phillips, "Theory of Servomechanisms," McGraw-Hill Book Co., Inc., New York, N. Y., chapt. 6, section 6.3; 1947.

[7] A. M. Mood, "Introduction to Mathematical Statistics," McGraw-Hill Book Co., Inc., New York, N. Y., chapt. 12; 1950.

[8] J. Neyman and E. S. Pearson, "On the problems of the most efficient tests of statistical hypotheses," *Phil. Trans.* (London), series A, vol. 231, p. 289; 1931.

[9] J. Neyman, "Basic ideas and theory of testing statistical hypotheses," *Jour. Roy. Stat. Soc.*, vol. 105, p. 292; 1942.

problem is: Supposing that we were given the scatter diagram of Fig. 1 and had one particular point (X_1X_2) which arises from a sample that had been taken from one of the distributions $S(t)$ or $N(t)$. How, then, can we best determine from which distribution it arose?

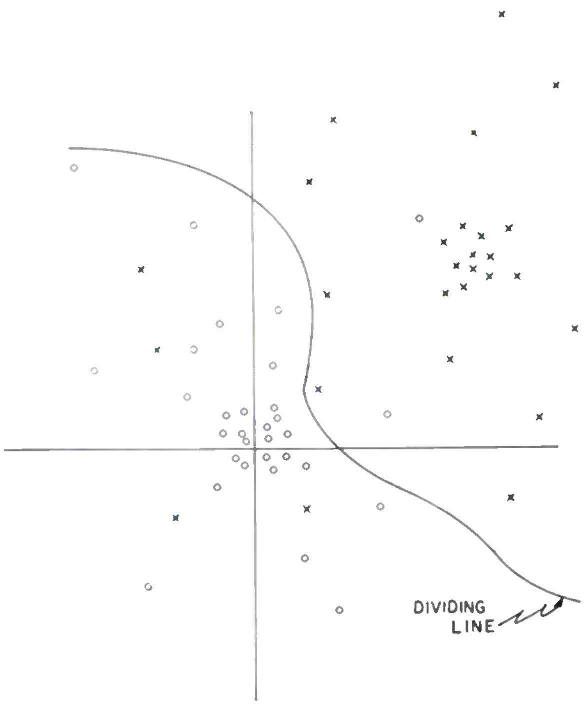

Fig. 1—Scatter diagram.

One possible approach is to draw in advance some curve, and to decide that if the point falls on one side of this curve we shall say that it arises from the distribution $S(t)$ and that if the point falls on the other side, from the distribution $N(t)$. The problem resolves completely, therefore, to selecting the "best" curve between two distributions.

The criterion which we have chosen demands that we fix the probability of saying the distribution is signal plus noise when it is really noise. Call this probability B. This means, in geometrical terms, that the proportion of points in the scatter diagram which lie on the right-hand side of the curve and originate from $N(t)$ is B. Now holding this proportion fixed, we are to draw a curve in such a way as to minimize the number of points due to $S(t)$ which fall on the left-hand side of the curve. This implies minimizing.

$$A = \int P_s(X_1X_2)dX_1dX_2 \tag{1}$$

left of curve
= probability that point from $S(t)$ will be to left of curve,

while holding fixed,

$$B = \int P_n(X_1X_2)dX_1dX_2 \tag{2}$$

right of curve
= probability that point from $N(t)$ will be right of curve.

The problem is, therefore, a calculus of variation problem in which the boundary is to be varied. The solution is found to be

$$P_s(X_1X_2) - \lambda P_n(X_1X_2) = 0, \tag{3}$$

$$\lambda = \frac{A}{1 - B} = \text{likelihood ratio.}$$

This is the equation of the curve separating the two distributions, and defines the best method of estimating whether the distribution from which the sample arose is $S(t)$ or $N(t)$. In particular, the decision is made on the basis of which of the inequality signs applies in the following,

$$P_s(X_1X_2) - \lambda P_n(X_1X_2) \gtrless 0. \tag{4}$$

Now to generalize to the case of m members of the sample, $X_1 \cdots X_m$, the procedure is obvious. In this case a sample space becomes an m dimensional hyperspace and the criterion becomes an $(m-1)$ dimensional hypersurface of this space. The calculus of the variations problem is then exactly analogous to the one defined in (1) and (2), and its solution is given by the equation:

$$P_s(X_1 \cdots X_m) - \lambda P_n(X_1 \cdots X_m) = 0. \tag{5}$$

C. Discussion of the Preceding Results

It has been shown how the best determination can be made of whether a sample of "m" members belongs to one statistical population or another, each statistical population being defined in terms of its m^{th} order probability density function. Criterion for "best" has been chosed to be that which keeps the false-alarm probability constant and at the same time minimizes the error of missing the signal. The results of these computations, moreover, define a filter, for they state that "m" values of a function taken in succession are to be confined in a certain specific way. Now the mechanization of such a filter is always possible, for the questions which come up in linear filters are absent here. In linear filter design it is possible to specify formally a filter in which the energy storage is required to be infinite, or in which the filter is required, because of the mathematical formulisms involved, to predict future quantities which, as yet, do not exist. Such filters are unrealizable, and it is therefore necessary to test for realizability by such criteria as Bode's and Nyquist's. However, the filter defined by (4) merely prescribes a certain function of a finite number of voltages. These voltages must be combined in a specific way and, because of the characteristics of probability density functions, all the quantities involved will be realizable.

The theory developed implied that a finite number of samples of the received signal should be obtained and the operations that were to be performed on these samples prescribed. In many instances[10] it can be seen that the result of the prescribed operations on the samples is equivalent to passing the original signal through a known circuit. When the output exceeds the level prescribed by (4), detection is said to have occurred. An example of this idea is given in the latter half of the paper.

Probably the most important point about this best detection filter is the two design parameters that are part of the basic specifications. In other words, once two specifications are given, namely,

(a) the probability of detection required,

and

(b) the false alarm probability that can be tolerated,

the filter is then completely defined. Notice from the definition of the likelihood ratio, λ, that if it is required that all signals be detected, the solution is to call everything a signal. In that way all signals will be detected, but the false-alarm probability will be unity.

II. The Detection of a Sine Wave in Noise by the Use of a Nonlinear Filter

A. The Importance of the Detection of a Sine Wave in Random Noise

In many applications a signal is transmitted by radar or sonar with a certain carrier frequency. If a target is present within the range of the transmitter, then the signal is reflected to the receiver. The input to the receiver is then to be identified as either a true signal or noise. The frequency of the incoming signal need not be known for the most general solution. If the frequency is known, the probability of detection can be increased, of course, since more *a priori* information is available. Notice that the question that is asked in the detection of a sine wave in noise is a basic detection problem.

There are reasons, other than the detection problem, why the separation of a sinusoid from random noise is important. One of these, the effects of noise on frequency-modulation systems, has been pointed out by Rice[11] and Middleton.[12]

B. The Fundamental Reason Why Nonlinear Filters Are Better than Linear Filters

It is known that linear filters are best only when the input signal has a Gaussian amplitude distribution. Both Shannon[13] and Singleton[2] have given formal proofs. A different argument can be presented as follows: The design of linear filters as described by Wiener is based on the autocorrelation function of the input signal and the autocorrelation function of the noise. The autocorrelation function of a signal can be derived from the second-order joint density function of the signal, i.e.,

$$\phi(\tau) = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} x_1x_2p(x_1x_2)dX_1dX_2 \tag{6}$$

x_1 = amplitude of signal at time t_1,

x_2 = amplitude of signal at time $t_1+\tau$,

$p(x_1x_2)$ = second-order joint density function of amplitude of signal,

$\phi(\tau)$ = autocorrelation function of signal.

Therefore, any signal with different third- or higher-order density functions, but the same second-order density function, will result in the same linear filter. Obviously then, when we *know* the joint density functions of the third or higher order and ignore it, we should not expect to get the best filter. This is exactly what happens when you design only linear filters. Nonlinear filters, on the other hand, are designed on the basis of the n^{th} order joint density function of the signal and noise, and therefore must preserve far more information about the original signal.

C. The Joint Probability Density Function of a Sine Wave Plus White Noise

In order to apply the theory of nonlinear filters, it is necessary to develop the joint probability density function of the signal received by the radar or sonar, i.e., the joint density function of signal plus noise and the joint density function of noise alone. In the detection problem, we are trying to identify a signal of a certain frequency. Furthermore, since we are searching a certain area with the radar or sonar set, we have a certain amount of time "$t_{\max}$," that we can look in a given direction, per scan of the transmitter. In order to determine the best filter, suppose that the amplitude of the incoming wave was sampled. The number of samples that are necessary in order to insure that all the information about the signal is obtained is well known to be $2f_{\max}t_{\max}$, where "$f_{\max}$" is the maximum frequency that will be considered. Therefore, we have a fixed number of samples of the incoming signal to deal with in the time "$t_{\max}$." Let

$$N = 2f_{\max}t_{\max}. \tag{7}$$

Suppose, first, that these N samples described a sine wave plus white Gaussian noise. Let

$$\begin{aligned} S_k &= k^{\text{th}} \text{ value of the received signal,} \\ Y_k &= k^{\text{th}} \text{ value of the sine wave} = E_0 \sin(k\omega\tau+\theta), \\ N_k &= k^{\text{th}} \text{ value of the random noise,} \\ S_k &= Y_k + N_k \\ k &= 1, 2, \cdots, N. \end{aligned} \tag{8}$$

[10] D. L. Durkey, "Optimum Method for Detection of Pulsed Signals in Noise," talks given at 1952 IRE National Convention.

[11] S. O. Rice, "Statistical properties of a sine wave plus random noise," *Bell Labs. Tech. Jour.*, vol. 27; January, 1948.

[12] D. Middleton, "On theoretical signal-to-noise ratios in FM receivers; comparison with amplitude modulation," *Jour. Appl. Phys.*, vol. 20, pp. 334–351; April, 1949. Also, "Spectrum of frequency-modulated waves after reception in random noise," *Quart. Appl. Math.*, vol. 7, pp. 129–173; July, 1949; vol. 8, pp. 59–80; April, 1950.

[13] C. E. Shannon and H. W. Bode, "Linear least square smoothing and prediction theory," Proc. I.R.E., vol. 38, pp. 417–425; April, 1950.

If the amplitude distribution of the noise has a mean of zero and a variance, σ^2, then it is shown in Appendix I that the inequality, $P_s - \lambda P_n > 0$, becomes,

$$\rho = \left\{\left[\sum_{k=1}^{N} S_k \sin(k\omega\tau)\right]^2 + \left[\sum_{k=1}^{N} S_k \cos(k\omega\tau)\right]^2\right\}^{1/2} > K\frac{\sigma^2}{E_0}, \quad (9)$$

where K is a constant that is determined by the following equation,

$$I_0(K) = \lambda \exp\left[\frac{NE_0^2}{4\sigma^2}\right] = \lambda \exp\left[\frac{2f_{\max}t_{\max}E_0^2}{4\sigma^2}\right], \quad (10)$$

and I_0 is the Bessel function of imaginary argument. The latter equation, (10), results only if

$$k\tau = \frac{1}{\omega}\left[\frac{(2k-1)\pi}{4} - \theta\right] \qquad k = 1, 2, \cdots, N.$$

However, if "τ" is selected in this way, it will be possible to instrument the filter without sampling the received signal. Therefore, it is only necessary to insure that it would be *theoretically* possible to select "τ" as prescribed. Such is obviously the case.

D. The Instrumentation of the Nonlinear Filter for the Detection Problem

Examination of (9) indicates that "ρ" is the output of an LC tuned circuit followed by a linear detector "$t_{\max}$" seconds after the received signal has been applied to the input. The complete filter therefore consists of a bank of tuned circuit and linear detectors, each followed by a threshold trigger device such as a blocking oscillator or multivibrator. These LC circuits are tuned to the set of possible frequencies that the signal to be detected could have. They are, of course, linear devices that any engineer would immediately think of to detect a signal of a single frequency. The nonlinear part of the filter is the set of linear detectors and trigger devices. The operating levels of the trigger devices are determined by (9) and (10). The techniques of mathematical statistics have therefore established a logical way of combining the following specifications and pieces of information:

1. Probability of detection required and false-alarm probability that can be tolerated—(λ).
2. *A priori* information, that the signal is a sine wave whose maximum frequency is no higher than $f_{\max}$.
3. *A priori* noise information—(σ^2).
4. Maximum time allowed for detection—($t_{\max}$).

The results given here for the detection of a sinusoid are a special case of a much more general theorum. Professor Fano[14] of M.I.T. has shown that in the general case of additive Gaussian noise, the optimum detector can be obtained by arranging a device to choose the largest output of a bank of tuned circuits. The techniques outlined in this paper determine whether this largest output actually implies detection.

[14] R. M. Fano, "Signal-to-Noise Ratio in Correlation Detectors," Technical Report No. 186, M.I.T. Research Laboratory of Electronics, Cambridge, Mass.; February, 1951.

Appendix I

A. The Joint Density Function of a Sine Wave Plus Random Noise

It is required to find the N^{th} order joint density function of "N" samples of a signal composed of a sine wave plus random noise. The definition of (8) will be used, and the samples are assumed to be "τ" seconds apart. It is further assumed that the phase angle "θ" is uniformly distributed and that the noise has a Gaussian amplitude distribution with mean zero and variance σ^2. Then the joint density function of S_n and θ is

$$\begin{aligned} P(S_1, S_2, \cdots, S_N, \theta) &= P(\theta)P(S_1, S_2, \cdots, S_N/\theta) \\ &= P(\theta)P_n(S_1 - Y_1, S_2 - Y_2 \cdots S_N - Y_N). \quad (11)\end{aligned}$$

If it is assumed that the samples are independent, then

$$P(S_1, \cdots, S_N, \theta) \doteq \frac{1}{2\pi}\left[\frac{1}{[2\pi]^{1/2}\sigma}\right]^N \prod_{k=1}^{N} \cdot \exp\left[-\frac{\{S_k - E_0\sin(k\omega\tau + \theta)\}^2}{2\sigma^2}\right]. \quad (12)$$

If the samples intervals are chosen such that,

$$k\tau = \frac{1}{\omega}\left[\frac{(2k-1)\pi}{4} - \theta\right], \quad (13)$$

and we let

$$\sum_{k=1}^{N} S_k \cos k\omega\tau = \rho\cos\beta, \quad (14)$$

$$\sum_{k=1}^{N} S_k \sin kw\tau = \rho\sin\beta,$$

then

$$P(S_1, S_2, \cdots, S_N, \theta) = \frac{1}{2\pi}\left[\frac{1}{(2\pi)^{1/2}\sigma}\right]^N \exp\left[-\frac{1}{2\sigma^2}\left\{\sum_{k=1}^{N} S_k^2 - 2E_0\sin(\theta+\beta) + \frac{NE_0^2}{2}\right\}\right]. \quad (15)$$

When "θ" is integrated out, the result is,

$$P_S(S_1, S_2, \cdots, S_N) = \left[\frac{1}{\sqrt{2\pi}\,\sigma}\right]^N \exp\left[-\frac{1}{2\sigma^2}\left(\sum_{k=1}^{N} S_k^2 + \frac{NE_0^2}{2}\right)\right] I_0\left(\frac{\rho E_0}{\sigma^2}\right), \quad (16)$$

where I_0 is the Bessel function of imaginary argument.

When (16) is combined with the expression for $P_n(S_1, S_2, \cdots, S_N)$, the relation $P_s - \lambda P_n > 0$ becomes

$$I_0\left(\frac{\rho E_0}{\sigma^2}\right) > \lambda \exp\left[\frac{NE_0^2}{4\sigma^2}\right]. \tag{17}$$

Let a constant "K" be the argument of the Bessel function that results in

$$I_0(K) = \lambda \exp\left[\frac{NE_0^2}{4\sigma^2}\right]; \tag{18}$$

then in order for inequality (18) to be true,

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Simultaneous Radiation of Odd and Even Patterns by a Linear Array*

CHESTER B. WATTS, JR.†, ASSOCIATE, IRE

Summary—A method is described for obtaining, with a broadside linear array, simultaneous radiation of two antenna patterns, one of which is an odd function, and the other an even function of azimuth angle. While the two patterns are not mutually independent, it is possible for them both to be free of minor lobes. The arrangement has an appealing simplicity when used with a slotted waveguide; however, it also has the drawback that the performance is limited to a comparatively narrow band of frequencies. An approximate theory of operation is given, together with some experimental results for an application in connection with runway localizers for instrument landing.

Introduction

It is sometimes desirable to be able to drive a linear antenna array with two signals of different type, simultaneously, and in such a way that the signals are radiated with different field patterns, one an odd function and the other an even function of the azimuth angle. An illustration of two such field patterns is given in Fig. 1. The pattern $f_A(\theta)$ has even symmetry in that

$$f_A(\theta) = f_A(-\theta), \tag{1}$$

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The azimuth angle, θ, is measured from the normal to the array as indicated in Fig. 2. An even pattern has zero rate of change at the center line, while an odd pattern is always characterized by a null at the center line.

Such pairs of odd and even patterns find frequent use in the general field of direction finding, where the odd pattern is used to measure the departure from the center line while the even pattern is used to furnish a phase reference which determines the sensing or polarity of the departure. These null and reference patterns are, of course, not necessarily radiated by the same antenna elements. A good example of this is to be found in the

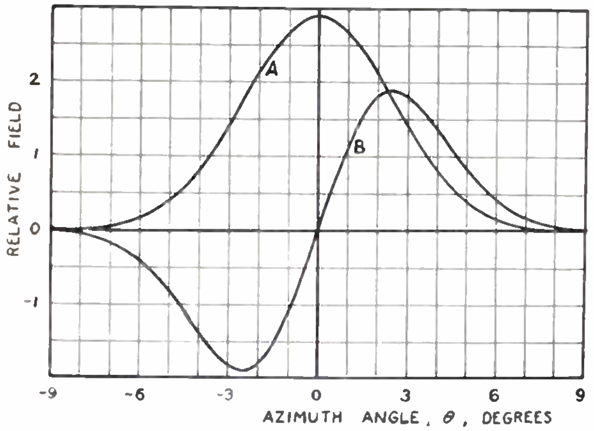

Fig. 1—Curve A: even pattern, $f_A(\theta)$. Curve B: odd pattern, $f_B(\theta)$.

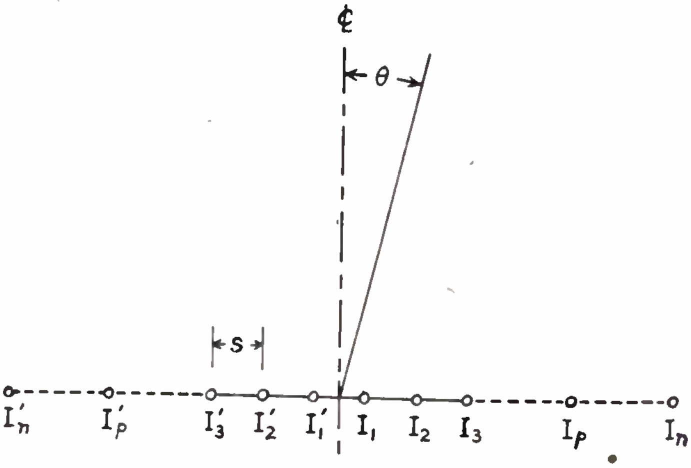

Fig. 2—Symmetrical linear array.

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Description of Method

The symmetrical linear array illustrated in Fig. 2 has a uniform spacing, s, between adjacent elements. The value of s will usually lie between 0.5 and 1.0 wavelength. Symbol I_p represents the current in any element numbered p from the center on the right side, and I_n is the current in the end element. The primed symbols indicate the symmetrical currents on the left side. Now suppose the elements are fed in a manner which may be represented by the equivalent circuit of Fig. 3. The sym-

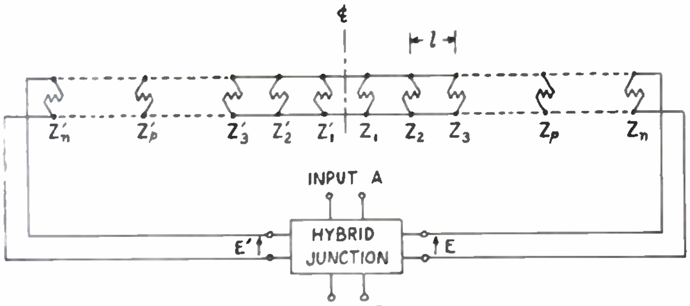

Fig. 3—Equivalent circuit.

bols Z_1, Z_2, $\cdots$, and the like, each represent the input impedance of an element, and may be varied to control the element currents. These impedances are all shunted across the line, the points of connection being uniformly separated by a length of line, l, where

$$l = \frac{\lambda_g}{2} + \delta, \tag{3}$$

in which λ_g is the wavelength in line or waveguide at the operating frequency and δ is a small part of a wavelength.

The hybrid junction is arranged so that a signal applied at input A produces output voltages

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Now if there is complete symmetry about the center line throughout the system, the signal at input A will be radiated as an even pattern while the signal at input B will be radiated as an odd pattern. It should be noted, however, that this statement holds only for an even number of elements, as shown; for an odd number of elements, the reverse is true. The shapes of the odd and even antenna patterns are, of course, determined by the distributions of element currents, and are not mutually independent in this arrangement. Even so, it is possible, as will be shown, to approach patterns of the type illustrated in Fig. 1, which have no minor lobes.

When the method is used with a slotted rectangular waveguide, the physical arrangement of an element may be as sketched in Fig. 4. Numerous other variations are

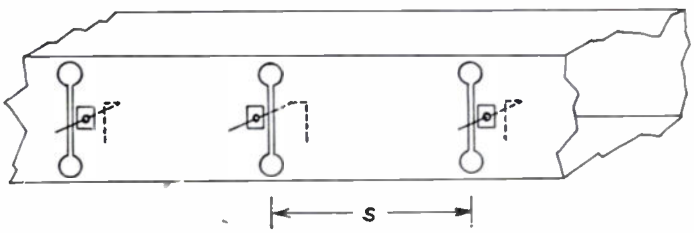

Fig. 4—Physical arrangement of elements.

possible.[2] Here, each element is a transverse slot in the narrow face of the waveguide operating in the dominant mode. The radiated polarization is longitudinal. The slot is end-loaded to bring it near half-wave resonance. The depth of penetration of the hooked probe largely determines the element current for a given field in the guide. Any two adjacent slots have their probes on opposite sides in order to keep the radiated fields in phase.

An Approximate Theory of Operation

To begin with, let us make several assumptions which, while not exactly true in practice, will make this approximate theory much simpler. The assumptions are as follows:

1. The number of elements is large enough to produce patterns of sufficient sharpness that the difference between θ and $\sin \theta$ can be ignored in the region of interest. Also, the number of elements is large enough to allow one to think of a continuous current distribution across the aperture, rather than discrete sources.

2. The coupling of the elements to the waveguide is so light that the distribution of fields within the guide is not appreciably disturbed thereby; that is, in the equivalent circuit, the impedance Z_p is always high compared with the impedance across which it is connected.

[1] P. Caporale, "The CAA instrument landing system," *Electronics*, vol. 18, pp. 116–124; February, 1945; and also pp. 128–135; March, 1945.

[2] R. E. Clapp, "Probe-fed Slots as Radiating Elements in Linear Arrays," M.I.T. Radiation Lab. Report No. 455; January 25, 1944.

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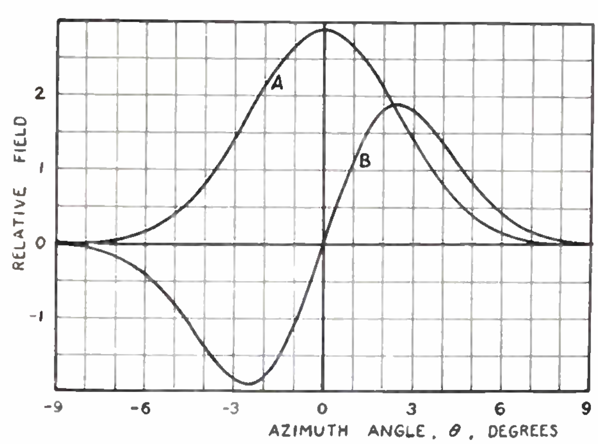

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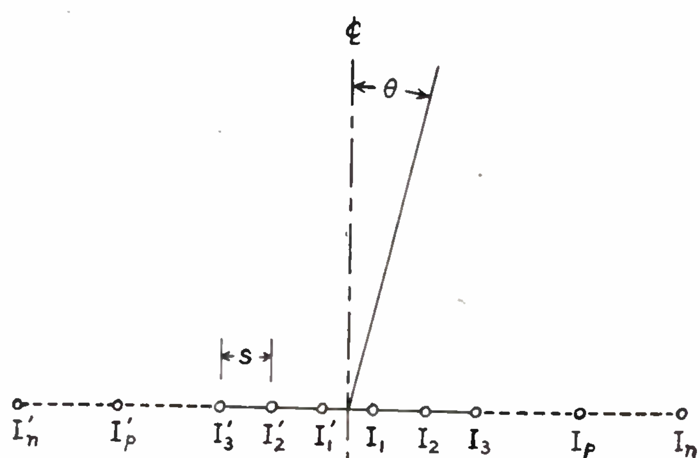

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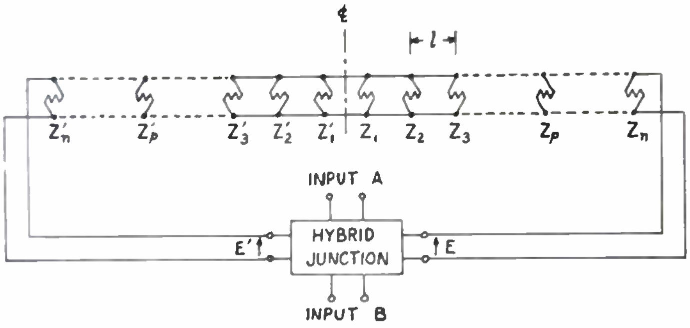

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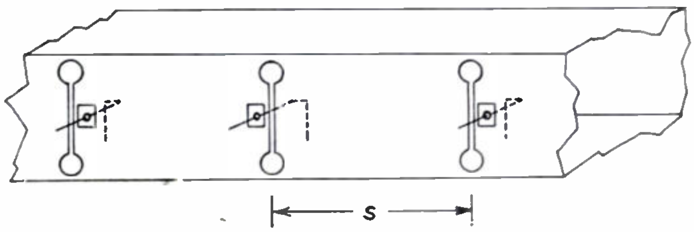

Fig. 4—Physical arrangement of elements.

possible.[2] Here, each element is a transverse slot in the narrow face of the waveguide operating in the dominant mode. The radiated polarization is longitudinal. The slot is end-loaded to bring it near half-wave resonance. The depth of penetration of the hooked probe largely determines the element current for a given field in the guide. Any two adjacent slots have their probes on opposite sides in order to keep the radiated fields in phase.

An Approximate Theory of Operation

To begin with, let us make several assumptions which, while not exactly true in practice, will make this approximate theory much simpler. The assumptions are as follows:

1. The number of elements is large enough to produce patterns of sufficient sharpness that the difference between θ and $\sin \theta$ can be ignored in the region of interest. Also, the number of elements is large enough to allow one to think of a continuous current distribution across the aperture, rather than discrete sources.
2. The coupling of the elements to the waveguide is so light that the distribution of fields within the guide is not appreciably disturbed thereby; that is, in the equivalent circuit, the impedance Z_p is always high compared with the impedance across which it is connected.

[1] P. Caporale, "The CAA instrument landing system," *Electronics*, vol. 18, pp. 116–124; February, 1945; and also pp. 128–135; March, 1945.

[2] R. E. Clapp, "Probe-fed Slots as Radiating Elements in Linear Arrays," M.I.T. Radiation Lab. Report No. 455; January 25, 1944.

3. Distance δ, which is the difference between the element spacing and the half-wavelength in the waveguide, is quite small, so that $\sin 2\pi(n\delta/\lambda_g)$ is not appreciably different from $2\pi(n\delta/\lambda_g)$ and $\cos 2\pi(n\delta/\lambda_g)$ is approximately unity.

Suppose the desired radiation patterns of the array are constant phase and are shaped in amplitude, as shown in Fig. 1, where

$$f_A(\theta) = A\epsilon^{-k_1\theta^2} \tag{6}$$

$$f_B(\theta) = B\theta\epsilon^{-k_1\theta^2}, \tag{7}$$

in which the constant k_1 has the value 0.08 $(\text{degree})^{-2}$. These two shapes happen to belong to a family of functions which have the interesting property of being their own Fourier transform, except for constants.[3] Thus, the current distributions required to produce the patterns have essentially the same shapes as the patterns themselves.

$$I_A(x) = C\epsilon^{-k_2x} \tag{8}$$

$$I_B(x) = Dx\epsilon^{-k_2x^2}, \tag{9}$$

where the variable x represents the distance along the aperture from the center of the array and the constant k_2 is given by the relation

$$k_2 = \frac{\pi^4}{(180)^2\lambda^2 k_1}. \tag{10}$$

Consider now first the application of signal to input A, which results in an even radiation pattern. By symmetry, there must be a null in the waveguide field, $\mathcal{E}_A$, at the center line. Also, by virtue of the simplifying assumptions, there will be standing waves throughout the length of the waveguide, with each element being close to a maximum, as indicated in Fig. 5. Since, with a high

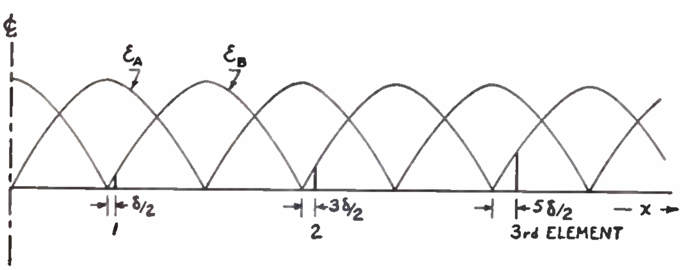

Fig. 5—Relation of element positions and standing waves in the waveguide.

standing-wave ratio, the field $\mathcal{E}_A$ has a phase which is substantially constant with x, except for sharp 180-degree changes across the null points, it is possible to adjust the values of the element impedances, Z_p, to obtain element currents which follow the desired distribution function $I_A(x)$ of (8).

Assuming that the element impedances have been chosen, consider next the application of signal to input B. This time, from considerations of symmetry, there must be a standing-wave maximum at the center line, as indicated by the curve labelled $\mathcal{E}_B$ in Fig. 5. Now, each element is close to a null, the space between element and associated null being proportional to the distance x from the center line. Thus, instead of all the elements being in fields which are approximately the same strength, we have the situation where each element is in a field whose strength is approximately proportional to the distance from the center line. It is apparent, then, that whatever current distribution existed for input A is multiplied by a factor roughly proportional to x for input B. Hence, the desired current distribution $I_B(x)$, (9), is obtained.

Experimental Results

As would be expected, the arrangement which has been described is not suitable for operation over a broad band of frequencies. The simplifying limitation that $n\delta$ be kept small is quickly violated as frequency is varied, particularly for large aperture arrays. However, in actual practice, the limitation is not quite as severe as was indicated in the approximate theory. It has been found, experimentally, that good results may be obtained with values of $n\delta$ up to about $\lambda_g/4$. Also, there is good reason to believe that the bandwidth may be considerably increased by means of periodic loading.[4]

An experimental slotted waveguide array has been constructed for operation in the 110-mc localizer band. The waveguide measures 41 by 78 inches in cross section, and about 105 feet in length. The material is galvanized sheet iron, and the mechanical design and assembly follow techniques which are standard in the fabrication of large air-conditioning ducts. The waveguide is driven at each end by a quarter-wave whip, and an RF bridge constructed of RG-8/U transmission line is used to obtain the hybrid junction. There are 18 slot elements, spaced $70\frac{1}{4}$ inches, or approximately 0.707 wavelength. The slot elements are fed by hooked probes, as indicated in Fig. 4.

Sample records of even and odd field patterns which have been obtained with this array are given in Fig. 6. These patterns were recorded with an airborne linear field meter at a distance of six miles and an altitude of 1,000 feet above ground. The operating frequency was 109.1 mc, with δ equal to 4.8 inches. In each case there is plotted a theoretically obtained pattern (dotted line). It will be noted that the measured patterns are somewhat broader than the calculated, particularly in the case of the odd pattern, Fig. 6(b). This is attributed to a progressive phase error in the element currents brought about by the fact that the slot dimensions were all identical, while the probe penetrations varied from about 3 inches at the ends to about 18 inches toward the middle. The vertical portions of all probes measured 10 inches.

[3] G. A. Campbell and R. M. Foster, "Fourier Integrals for Practical Applications," *Bell Tel. Sys. Tech. Pub.*, Monograph B-584; September, 1931.

[4] A. W. Lines, G. R. Nicoll, and A. M. Woodward, "Some properties of waveguides with periodic structure," *Proc. IEE* (London), vol. 97, pt. III, pp. 263–276; July, 1950.

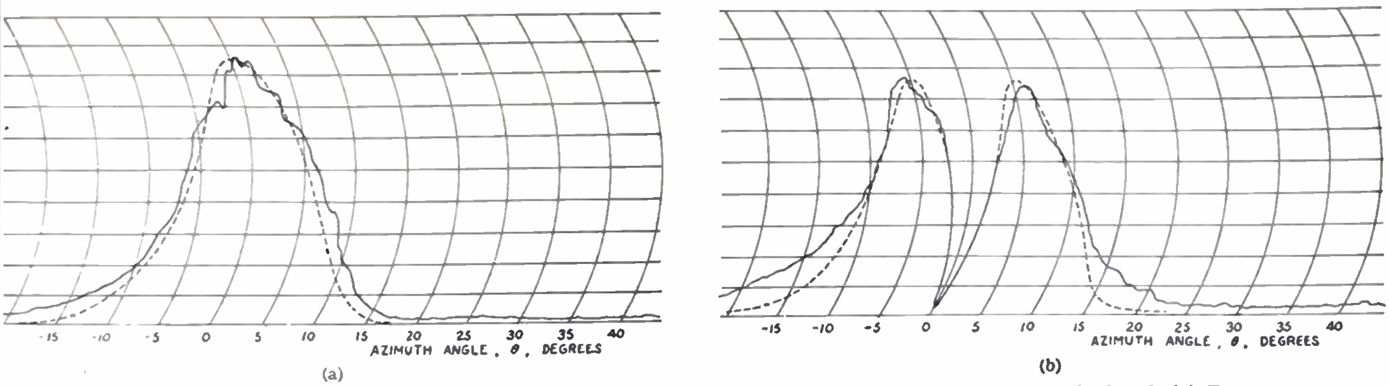

Fig. 6—Sample field recordings for experimental 18-slot array. Solid curve—measured, dotted curve—calculated. (a) Even pattern. (b) Odd pattern.

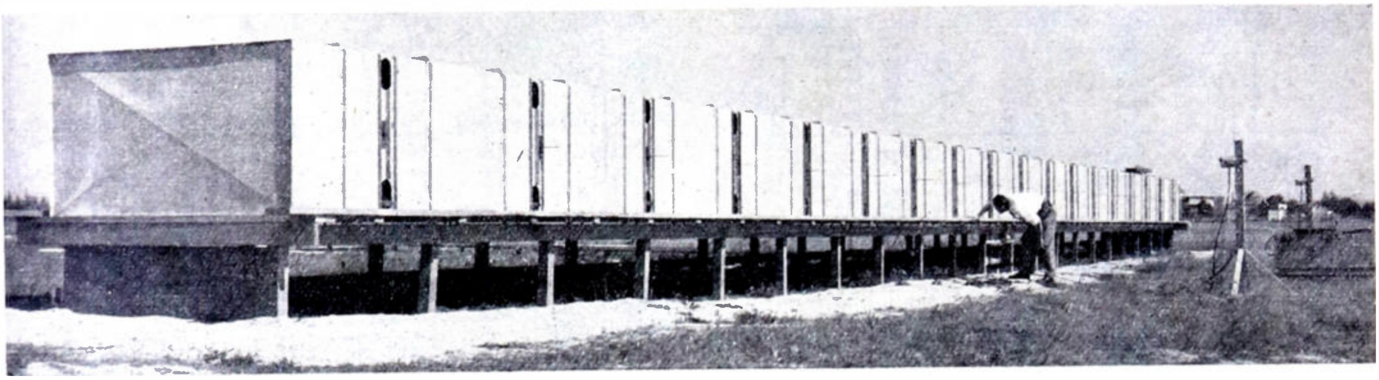

Fig. 7—Slotted waveguide array.

[t has been found that the phase error can be corrected, f desired, by the addition of an adjustable shorting bar on each slot, the slots normally being on the long side of half-wave resonance.

ACKNOWLEDGMENT

The writer is indebted to K. E. Voyles and S. E. Taggart for assistance in the design, construction, and adjustment of the experimental array.

Conductance Measurements on Operating Magnetron Oscillators*

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Summary—**The conductance terms in the equivalent circuit of an operating magnetron oscillator can be experimentally obtained by measuring the magnetron *Q*-factors both in the nonoscillating and oscillatory conditions. A method for determining the "operating" *Q*-factors in a magnetron by measuring the deviations in magnetron output power and oscillating frequency caused by a specified load mismatch is reviewed, and its application to magnetron conductance measurements is discussed. Experimental data are presented to illustrate the method of measurement.**

I. Introduction

The Equivalent Circuit of an Operating Magnetron

THE principal-mode equivalent circuit of a "cold" magnetron, i.e., one in which effects of electrons present in the magnetron interaction space are negligible, is that of any single-loop coupled cavity with sufficient mode separation, and it can be represented as in Fig. 1(a). In this representation, the usual restrictions—that the coupling device has reactance which varies negligibly with frequency and is dissipationless—are implied. Considering the resonator itself, to the left of terminals $A-A_1$, G_c is the load conductance transformed into the cavity by the turns ratio of the coupler, G represents the total cavity losses, and L and C are the lumped parameters of the equivalent tank circuit.

In the case of the oscillating magnetron, Fig. 1(b), G'' is included to account for the change in the total cavity losses under oscillating conditions, so that the cavity losses become G'.[1] Y_e' is the electron-transit admittance, the susceptive portion of which, C_e, can be measured experimentally by observing the frequency

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[1] Primed symbols indicate parameters under conditions of oscillation, as distinct from "cold" values.

shift of the magnetron with varying average plate current (frequency pushing). This capacitive effect of the electron cloud has been analyzed for both the plane and cylindrical structures.[2–5]

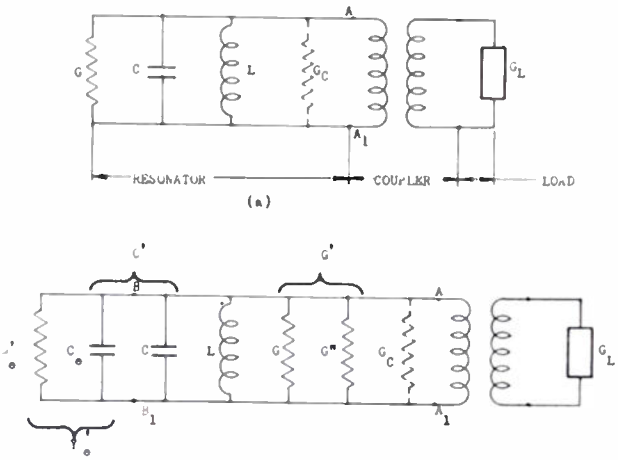

Fig. 1—Equivalent circuits of magnetron oscillator.
(a) Nonoscillating magnetron. (b) Oscillating magnetron.

The electronic conductance G_e', however, is more difficult to determine by the conventional measuring techniques on microwave magnetrons. It is the purpose of this paper to present a method whereby the conductances in the magnetron equivalent circuit can be experimentally obtained.

II. The Measurement of Q-Factors of an Operating Microwave Oscillator

An insight into the properties of an operating microwave oscillator may be gained by analyzing its operating ("hot") Q-factors defined to parallel the usually considered Q-factors for the nonoscillating conditions. From these factors the losses ascribed to various parameters in the equivalent circuit can be calculated.

The method to be presented has been developed elsewhere.[6] Its salient points will be reviewed here so that it may be applied to conductance measurements in the following sections.

Referring to Fig. 1(b), the "hot" Q-factors are defined as follows:

$$Q_L' = \frac{\omega_0'C'}{G' + G_c} = \frac{1}{\omega_0'L(G' + G_c)} \tag{1}$$

$$Q_u' = \frac{\omega_0'C'}{G'} = \frac{1}{\omega_0'LG'} \tag{2}$$

$$\frac{1}{Q_c'} = \frac{1}{Q_L'} - \frac{1}{Q_u'}, \tag{3}$$

where $\omega_0' = 2\pi f_0'$ is the radian frequency of the oscillator.

In terms of these factors, the maximum frequency deviation (pulling) of the oscillator from the matched-load condition when a mismatch of r_L in voltage standing-wave ratio (vswr) is introduced in the load and varied in phase is

$$\Delta f_0' = \frac{f_0'}{4}\left(\frac{1}{Q_L'} - \frac{1}{Q_u'}\right)\left(r_L - \frac{1}{r_L}\right); \tag{4}$$

while the maximum output power variation under similar conditions is

$$\Delta P_0 = P_0\left(\frac{2Q_L'}{Q_u'} - 1\right)(r_L - 1), \tag{5}$$

where P_0 is the oscillator output power for matched-load conditions.

In terms of the coupling parameter β defined by $Q_u' = (1+\beta)Q_L'$, (4) can be written as

$$\Delta f_0' = f_0' \frac{\beta}{4Q_u'}\left(r_L - \frac{1}{r_L}\right) \tag{6}$$

or

$$\Delta f_0' = f_0' \frac{\beta}{1+\beta}\,\frac{1}{4Q_L'}\left(r_L - \frac{1}{r_L}\right); \tag{7}$$

while (5) can be rewritten in the form

$$\Delta P_0 = P_0 \frac{1-\beta}{1+\beta}(r_L - 1). \tag{8}$$

Equations (6), (8) and (7), (8) form two independent sets, each of which can be solved for Q_u' and Q_L', respectively, in terms of $\Delta f_0'/f_0'$, $\Delta P_0/P_0$ and r_L. It will be noted that the quantities f_0', $\Delta f_0'$, P_0, and ΔP_0 are directly measurable on an oscillating cavity, and thus the values of the "operating" Q-factors can be experimentally determined. From (8)

$$\beta = \frac{(r_L - 1) - \dfrac{\Delta P_0}{P_0}}{(r_L - 1) + \dfrac{\Delta P_0}{P_0}}, \tag{9}$$

so that, from (6),

$$Q_u' = \frac{(r_L - 1) - \dfrac{\Delta P_0}{P_0}}{(r_L - 1) + \dfrac{\Delta P_0}{P_0}}\,\frac{f_0'}{4\Delta f_0'}\left(r_L - \frac{1}{r_L}\right). \tag{10}$$

[2] J. C. Slater, "Microwave Electronics," D. Van Nostrand Co., New York, N. Y., chapt. XIII; 1950.
[3] H. W. Welch, Jr., "Effects of space charge on frequency characteristics of magnetrons," Proc. I.R.E., vol. 38, p. 1434; December, 1950.
[4] H. W. Welch, Jr. and W. G. Dow, "Analysis of synchronous conditions in the cylindrical magnetron space charge," *Jour. Appl. Phys.*, vol. 22, p. 433; April, 1951.
[5] G. B. Collins, "Microwave Magnetrons," McGraw-Hill Book Co., Inc., New York, N. Y.; 1948.
[6] M. Nowogrodzki, "The Measurement of Q-Factors of an Operating Microwave Oscillator." Thesis in partial fulfilment of the requirements for the M.E.E. degree at the Polytechnic Institute of Brooklyn; May, 1951.

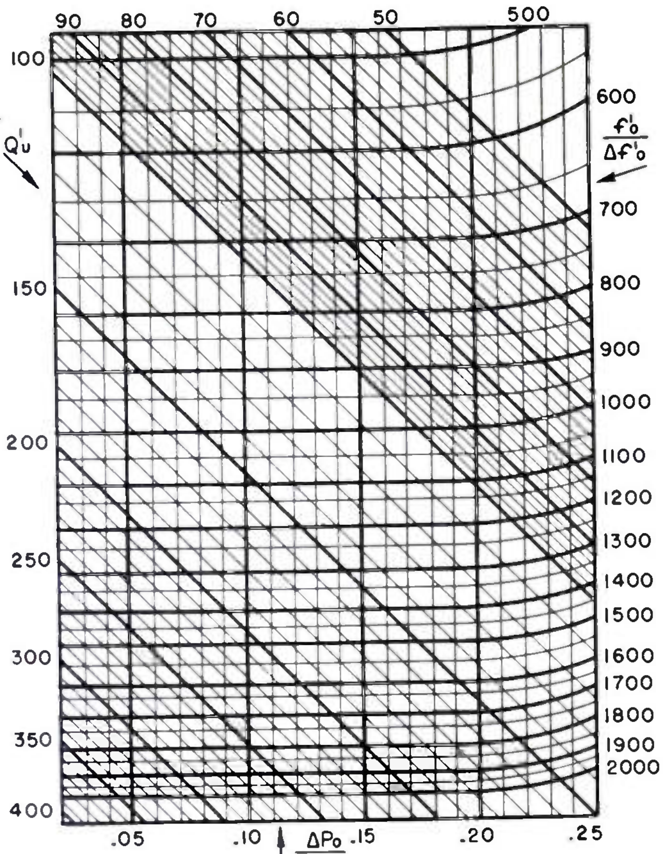

Fig. 2—Chart for determination of Q_u'. Plot of (10) for mismatch vswr of 1.5:1.

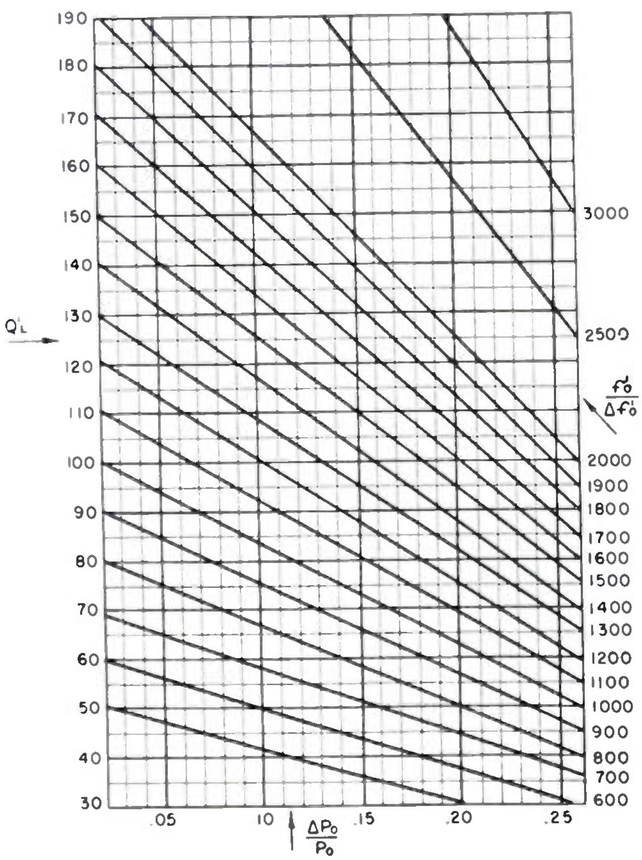

Fig. 3—Chart for determination of Q_L'. Plot of (11) for mismatch vswr of 1.5:1.

Similarly, from (7)

$$Q_L' = \frac{f_0'}{8\Delta f_0'}\left(1 - \frac{\dfrac{\Delta P_0}{P_0}}{r_L - 1}\right)\left(r_L - \frac{1}{r_L}\right). \tag{11}$$

Expressions (10) and (11) become considerably simplified once a value of the load mismatch r_L is decided upon. For magnetrons, the usual case of interest is $r_L = 1.5$, since the magnetron pulling figure is usually defined as the maximum variation in magnetron frequency when a mismatch of $r_L = 1.5$ is introduced in the load and varied in phase. In terms of the symbols used here, the pulling figure thus defined is twice $\Delta f_0'$.

In Figs. 2 and 3, (10) and (11) have been plotted for $r_L = 1.5$. These charts are taken from the cited reference.[7]

III. The Measurement of Magnetron Conductances

The unloaded Q of a "cold" magnetron is defined as

$$Q_u = \frac{\omega_0 C}{G}. \tag{12}$$

This Q-factor can be measured by any of the conventional "cold" measuring techniques.[8] Also, the equivalent lumped capacitance C can be derived with the aid of design formulas and equations available in the literature for the various magnetron geometries.[9] An immediate check on the accuracy with which C has been determined is available by comparing the value of the "cold" resonant wavelength in the desired mode derived by using the calculated value of C, with the resonant wavelength measured on the cold-bench setup. By successive approximations, the calculated wavelength (and therefore the value of C) can be made to agree with the actual value to within, for example, 5 per cent (greater accuracy may not be consistent with the nature of the measurements). Then, using the *measured* values of Q_u and ω_0 and this *calculated* value of C,

$$G = \frac{\omega_0 C}{Q_u}; \tag{13}$$

while L, the lumped equivalent inductance, is

$$L = \frac{1}{C\omega_0^2}. \tag{14}$$

The assumption is now made that this inductance remains essentially constant as the magnetron is caused to oscillate, all frequency changes being ascribed to the susceptive (capacitive) effect of the electron-transit ad-

[7] *Ibid.*, p. 25.
[8] See, for example, C. G. Montgomery, "Technique of Microwave Measurements," McGraw-Hill Book Co., Inc., New York, N. Y., p. 330 *et seq.*; 1947
[9] G. B. Collins, *op. cit.*, chapt. 11.

mittance Y_e'. Under these circumstances, by measuring the oscillating radian frequency ω_0', the lumped capacitance for an *oscillating* magnetron can be easily calculated as

$$C' = \frac{1}{L(\omega_0')^2} = C\frac{f_0^2}{(f_0')^2}. \tag{15}$$

Equation (15) thus describes the susceptive effect of the electron-transit admittance in terms of the measured "cold" and oscillating magnetron resonant frequencies.

In addition, it will be noted that a knowledge of G and measurements of both the "cold" and "operating" Q-factors of the magnetron suffice to determine all conductance terms in the equivalent circuit of Fig. 1(b). Thus, from (1), (2), and (13)

$$G' + G_c = \frac{Q_u}{Q_L'}\frac{f_0}{f_0'}G \tag{16}$$

$$G' = \frac{Q_u}{Q_u'}\frac{f_0}{f_0'}G, \tag{17}$$

from which

$$G_c = Q_u\left(\frac{1}{Q_L'} - \frac{1}{Q_u'}\right)\frac{f_0}{f_0'}G = \frac{Q_u}{Q_c'}\frac{f_0}{f_0'}G. \tag{18}$$

The condition for oscillation requires that the electron-transit conductance G_e' cancel all conductances to the right of terminals B-B_1 Fig. 1(b).[10] Thus

$$G_e' = -(G' + G_c) = -\frac{Q_u}{Q_L'}\frac{f_0}{f_0'}G. \tag{19}$$

The loading effect represented by G'', can be calculated by making use of (12). Thus

$$G'' = G' - G = G\left(\frac{Q_u}{Q_u'}\frac{f_0}{f_0'} - 1\right). \tag{20}$$

Also, the "hot"-"cold" ratio of cavity losses may be of interest. This ratio is given by

$$\frac{G'}{G} = \frac{Q_u}{Q_u'}\frac{f_0}{f_0'}. \tag{21}$$

The "operating" Q-factor measurements can obviously be performed at any desired number of points within the particular magnetron performance chart, and the dependence of the conductances with various imposed conditions on the magnetron can thereby be obtained.

IV. Experimental Data

The empirical data presented in this section are intended merely as an illustration of the measurement method developed above. Average data on a considerable number of magnetrons (and possibly magnetron types) would be required before an attempt at interpretation of the numerical results might be undertaken.

Fig. 4 presents experimental data taken on a 4J58-type magnetron, a fixed-frequency, 12-vane, double-ring-strapped, packaged, pulsed-type tube operating in the X_b-band. The "cold" resonance measurements were made using the Wheeler method,[11] and the following values were obtained:

$$Q_u = 879$$
$$Q_L = 284$$
$$Q_c = 344$$
$$f_0 = 6429 \text{ mc per second}$$
$$\lambda_0 = 4.675 \text{ cm.}$$

The 4J58-type magnetron is one of the few types in which the pulling figure for a load mismatch other than 1.5:1 in vswr is of interest.[12] For this reason $r_L = 1.75$, the JAN-specified mismatch, was used in the measurements.

For $r_L = 1.75$, (10) and (11) become

$$Q_u' = 0.295F\frac{0.75 - P}{0.75 + P} \tag{22}$$

$$Q_L' = 0.1475F(1 - 1.333P), \tag{23}$$

where

$$F = \frac{f_0'}{\Delta f_0'} \text{ and } P = \frac{\Delta P_0}{P_0}.$$

Fig. 5 gives the Q-factors of the oscillating magnetron as a function of its average plate current.

Average plate current (milliamps.)	Average power output (galvanometer divisions)				Frequency (megacycles per second)			
	Matched load	P_{Max}	P_{min}	ΔP_0 (Av.)	Matched load	f'_{Max}	f'_{min}	$\Delta f_0'$ (Av.)
15	39.5	48.0	28.5	9.75	6410	6414	6405	4.5
20	53.5	67.5	38.0	14.75	6408	6413	6403	5.0
25	67.0	86.5	45.0	20.75	6405	6410.5	6399.5	5.5
30	75.0	100.0	53.0	23.5	6401.5	6407	6397	5.0

Fig. 4—Experimental results. Type 4J58 magnetron. (Duty cycle = 0.001; $r_L = 1.75:1$.) One galvanometer division = 3 watts.

[10] See, for example, A. B. Bronwell and R. E. Beam, "Theory and Application of Microwaves," McGraw-Hill Book Co., Inc., New York, N. Y., p. 142; 1947.

[11] H. A. Wheeler, "Measurement of Resonance by the Reflection Loss Method," Hazeltine Corporation Report 9164; 1945.

[12] JAN specification 4J54-59.

Following the Collins curves,[13] estimated values of $C=0.415$ micromicrofarads, $\lambda_0=4.22$ cm were initially obtained. Since the measured cold-resonant wavelength was $\lambda_0=4.675$ cm, this represented a 12-per cent difference in wavelengths. By assuming the error to be attributed mainly to the estimated strap capacitance and modifying this estimated value slightly, new values for C and λ_0 were obtained, namely, $C=0.465$ micromicrofarads, $\lambda_0=4.46$ cm. This represented a difference in calculated and measured resonant wavelengths of the order of 4.7 per cent, which was considered adequate. Thus the values used in subsequent calculations were

$$C=0.465 \text{ micromicrofarads.}$$
$$G=21.8 \text{ micromhos.}$$

I_b (milliamperes)	Q_U'	Q_L'	Q_C'
15	207	139	422
20	172.5	119	384
25	140	100	350
30	157.5	113	400

Fig. 5—Q-factors of oscillating magnetron. (Type 4J58.)

Fig. 6 is a tabulation of conductance values calculated according to the formulas developed in the preceding section. Fig. 7 is a graphical presentation of the variations of G_e' and G'' with magnetron average-plate current.

conditions, its effect could possibly be investigated separately.

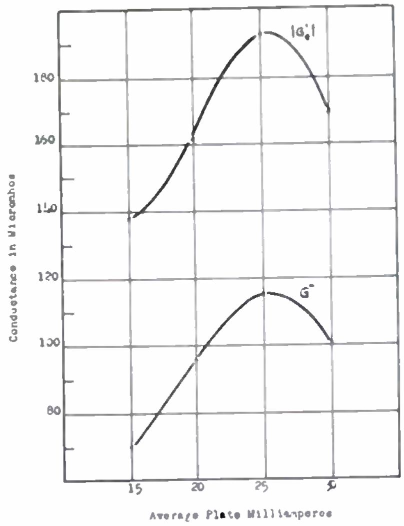

Fig. 7—Electron-cloud effects, oscillating magnetron. (Type 4J58.)

Reference to (21) shows that the ratio Q_u/Q_u' is a good approximation of G'/G since f_0/f_0' is very close to unity. Ratios of 15–25 were measured for Q_u/Q_u' in some S-band magnetrons, notably the 4J47 type, while this ratio in the 725A (X-band) magnetron was found to be approximately 3–5.[14]

An indication of the accuracy of measurements is afforded by the spread in measured values of Q_c' (and G_c).

I_b milliamps	$\dfrac{Q_U}{Q_U'}$	$\dfrac{Q_U}{Q_L'}$	$\dfrac{f_0}{f_0'}$	$\lvert G_e'\rvert$ or $G'+G_c$	G'	$\dfrac{G'}{G}$	G''	G_c
15	4.2	6.32	1.005	138.5	92	4.22	70.2	45.7
20	5.1	7.38	1.004	161.5	118	5.12	96.2	50
25	6.28	8.79	1.003	192	137.8	6.3	116	55
30	5.58	7.78	1.003	170	122.5	5.6	100.7	48.2

Fig. 6—Magnetron conductances. Type 4J58 magnetron. (All conductance values in micromhos.)

V. General Remarks

The magnitude of the operating losses of the magnetron as compared to the "cold" equivalent conductance G is of interest. It will be seen that the ratio G'/G varies between approximately 4 and 6. This comparatively large increase in losses is represented in the equivalent circuit by the term G'', and can perhaps be accounted for by separating the effects of the electrons in the "hub" around the cathode from those in the rotating "spokes." Thus the spokes are thought to contribute the negative conductance G_e', while the hub electrons contribute a loading term G'', which increases as the anode voltage is increased (for constant magnetic field), i.e., with increasing radius of the hub. Since the hub is present both during oscillation and under so-called preoscillation

[13] G. B. Collins, *loc. cit.* It should be noted that according to the computed values tabulated in Table 11.1, p. 465 of this reference, (2d) on p. 464 should read

$$\lambda_r = \lambda_r\sqrt{\frac{C_t}{C_r}}.$$

Since Q_c depends upon the characteristics of the coupler alone (which has been assumed to vary negligibly with frequency in the vicinity of resonance), the "cold" and "operating" values of this Q-factor should remain the same, except, perhaps, for thermal changes. The mean measured value of Q_c' is seen to be 389 (cf. Fig. 5), which is within 13.5 per cent of the "cold" value of 344; while the spread of measured values of Q_c', ascribed to experimental error, is of the order of 10 per cent.

Acknowledgments

The helpful suggestions of Victor J. Stein, of the Amperex Electronic Corporation, are gratefully acknowledged. Thanks are also due to Dr. Simon Sonkin of the City College of New York and the Columbia University Radiation Laboratory, who read this paper in manuscript form. Finally, the author is indebted to the Polytechnic Institute of Brooklyn for permission to reprint the charts shown in Figs. 2 and 3.

[14] M. Nowogrodzki, *op. cit.*, Figs. 9 and 11.

The Magnetic Dipole Antenna Immersed in a Conducting Medium*

JAMES R. WAIT†

Summary—The magnetic dipole antenna or small current-carrying loop immersed in a conducting medium is investigated. Explicit expressions for the fields are derived for the case when there is a spherical insulating cavity enclosing the dipole. The power radiated from the insulated dipole is discussed and an explicit expression for the total power is given for the case when all displacement currents in the media are negligible.

Introduction

THERE HAS BEEN considerable interest in recent years in the study of electromagnetic systems which are immersed in a conducting medium. This condition might arise when the transmitting antenna is underwater and the received signal is to be measured at some other underwater point.

The radiation from antennas immersed in a conducting medium will be radically different from the case of a surrounding nondissipative medium. As in the case of plane-wave propagation,[1] the radiated fields from the antenna will suffer considerable additional attenuation and phase distortion as a result of the finite conductivity of the medium. The power relationships will be vastly modified from the classical free-space case.

The electric or Hertzian uninsulated dipole in a conducting medium of infinite extent has been investigated by Tai.[2] It was here pointed out that difficulty is encountered when the attempt is made to calculate the total power radiated from the dipole due to the very high ohmic losses in its immediate neighborhood. It was concluded that the dipole must be insulated. Tai[3] also considered the biconical antenna at the center of an inhomogeneous region with spherical symmetry.

In this paper the magnetic dipole situated at the center of a spherical insulating cavity will be considered. The power radiated is then equal to power that crosses the wall of the cavity. Expressions for the fields in the external region will be explicitly given.

Derivation of the Fields

A magnetic dipole may be conceived of as an infinitesimal loop of wire of area dA carrying a current I. Such a dipole is situated at the center of a spherical cavity of radius a which has a dielectric constant ϵ_1, magnetic permeability μ_1, and a conductivity σ_1, which can be later allowed to vanish. The external medium is generally dissipative and has electrical properties ϵ_2, μ_2, and σ_2. MKS units are to be used throughout. A time factor $e^{i\omega t}$ is employed.

It is quite evident that the magnetic dipole in this situation will give rise to transverse-electric spherical waves inside and outside the spherical cavity. A conventional spherical polar system of co-ordinates (r, θ, ϕ) is chosen with the magnetic dipole situated at the origin and oriented in the polar direction. The resulting fields are then given by the following equations in terms of a scalar wave function ψ_m:

$$H_{rm} = \frac{1}{i\mu_m\omega}\left(\frac{\partial^2\psi_m}{\partial r^2} - \gamma_m^2\psi_m\right), \tag{1}$$

$$E_{\phi m} = \frac{1}{r}\frac{\partial\psi_m}{\partial\theta}, \tag{2}$$

$$H_{\theta m} = \frac{1}{i\mu_m\omega r}\cdot\frac{\partial^2\psi_m}{\partial r\partial\theta}, \tag{3}$$

$$E_{rm} = E_{\theta m} = H_{\phi m} = 0,$$

where

$$\gamma_m^2 = i\sigma_m\mu_m\omega - \epsilon_m\mu_m\omega^2$$

and the subscript m takes the value 1 or 2 to denote the interior or the exterior region, respectively. The function ψ_m satisfies the equation

$$r^2\frac{\partial^2\psi_m}{\partial r^2} + \frac{1}{\sin\theta}\frac{\partial}{\partial\theta}\left(\sin\theta\frac{\partial\psi_m}{\partial\theta}\right) - \gamma_m^2r^2\psi_m = 0. \tag{4}$$

Now the primary field of the magnetic dipole inside the spherical cavity is well known and the field components are given by

$$H_{r1}^0 = [IdA/2\pi r^3][1+\gamma_1 r]\exp(-\gamma_1 r)\cos\theta \tag{5}$$

$$H_{\theta 1}^0 = [IdA/4\pi r^3][1+\gamma_1 r+\gamma_1^2r^2]\exp(-\gamma_1 r)\sin\theta \tag{6}$$

$$E_{\phi 1}^0 = [-i\mu_1\omega IdA/4\pi r^2][1+\gamma_1 r]\exp(-\gamma_1 r)\sin\theta. \tag{7}$$

The corresponding wave function ψ_1^0 which gives rise to the fields is then

$$\psi_1^0 = [i\mu_1\omega IdA/4\pi r][1+\gamma_1 r]\exp(-\gamma_1 r)\cos\theta. \tag{8}$$

This can be written in terms of the spherical Bessel function[4] of order one of the third type as follows:

* Decimal classification: R120. Original manuscript received by the Institute, February 19, 1952; revised manuscript received May 20, 1952.

† Department of Electrical Engineering, University of Toronto, Toronto, Canada.

[1] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, N. Y., p. 276; 1941.

[2] C. T. Tai, Cruft Laboratory Research Report No. 21, Harvard University, Cambridge, Mass.; 1947.

[3] C. T. Tai, Cruft Laboratory Research Report No. 75, Harvard University, Cambridge, Mass.; 1949.

[4] The spherical Bessel functions of argument z are defined by

$$k_1(z) = (1+z)\exp(-z)/z$$

and

$$i_1(z) = \cosh z - \sinh z/z.$$

$$\psi_1{}^0 = ck_1(\gamma_1 r)\cos\theta, \tag{9}$$

where

$$c = i\mu_1\omega IdA\gamma_1/4\pi.$$

The general solution for the complete wave function ψ_1 inside the cavity and the complete wave function ψ_2 outside the cavity can then be written, respectively, as:

$$\psi_1 = ck_1(\gamma_1 r)\cos\theta + A_1 i_1(\gamma_1 r)\cos\theta \tag{10}$$

and

$$\psi_2 = B_1 k_1(\gamma_2 r)\cos\theta, \tag{11}$$

where i_1 is a spherical Bessel function of order one of the first type. The coefficients A and B can be found by applying the boundary conditions which require the continuity of the tangential electric and magnetic fields across the boundary. That is, the E_ϕ and the H_θ components are continuous at $r=a$. From (2) and (3), this implies that ψ_m and $(1/\mu_m)(\partial\psi_m/\partial r)$ are continuous at $r=a$. This gives rise to two equations for the solution of A_1 and B_1 so that

$$A_1 = [B_1 k_1(\beta) - ck_1(\alpha)]/i_1(\alpha) \tag{12}$$

and

$$B_1 = c\frac{\alpha k_1(\alpha)i_1'(\alpha) - \alpha k_1'(\alpha)i_1(\alpha)}{\alpha i_1'(\alpha)k_1(\beta) - q\beta k_1'(\beta)i_1(\alpha)}, \tag{13}$$

where $\alpha=\gamma_1 a$, $\beta=\gamma_2 a$, and $q=\mu_1/\mu_2$. The dashed quantities indicate a differentiation with respect to the argument of the function.

The wave functions ψ_1 and ψ_2 inside and outside the spherical cavity are then completely specified. By carrying out the operations indicated by (1), (2), and (3), the fields in the region exterior to the cavity may be written

$$H_{2r} = (IdA)_e(1+\gamma_2 r)\exp(-\gamma_2 r)\cos\theta/2\pi r^3 \tag{14}$$

$$H_{2\theta} = (IdA)_e(1+\gamma_2 r+\gamma_2{}^2 r^2)\exp(-\gamma_2 r)\sin\theta/4\pi r^3 \tag{15}$$

and

$$E_{2\phi} = -(IdA)_e i\mu_2\omega(1+\gamma_2 r)\exp(-\gamma_2 r)\sin\theta/4\pi r^2, \tag{16}$$

where $[IdA]_e$ is defined as the equivalent magnetic dipole contained within the spherical cavity. The ratio of $(IdA)_e$ to the moment IdA is given by

$$[IdA]_e/IdA = q\gamma_1 B_1/\gamma_2 C. \tag{17}$$

The above ratio may be considerably simplified in form if the operating frequency and the dimensions of the cavity, and so on, are such that

$$\sigma_1 = 0,\quad |\alpha| = (\epsilon_1\mu_1\omega^2)^{1/2}a \ll 1,\ \text{and}\ \mu_1 = \mu_2 = \mu.$$

The ratio of the magnetic moments is then given by

$$(IdA)_e/IdA = 3\exp(\beta)(3+3\beta+\beta^2)^{-1}. \tag{18}$$

It is quite evident that this ratio approaches the value unity as the magnitude of β approaches zero. In other words, at sufficiently low frequencies the cavity has little or no effect on observed fields in the exterior region.

The Power Dissipated

If the outer medium is highly conducting and the frequency is sufficiently low, the displacement currents play a negligible role. The following approximation is then valid:

$$\gamma_2 = [i\mu\omega(\sigma_2 + i\epsilon_2\omega)]^{1/2} \simeq (i\sigma_2\mu\omega)^{1/2}.$$

The power radiated by the dipoles can then be calculated from the component of the Poynting vector in the radial direction, which is given by

$$P_r = E_\phi H_\theta^*.$$

The real power $p(\theta)$, as a function of θ crossing unit area of the cavity wall, is then given by

$$p(\theta) = \tfrac{1}{2}\text{ real part of }[E_{2\phi}H_{2\theta}^*]_{r=a}.$$

After some algebraic manipulation the expression for $p(\theta)$ is given by

$$p(\theta) = \frac{9}{2}\left(\frac{IdA}{4\pi}\right)^2\frac{\mu\omega}{a^5}\sin^2\theta X(m), \tag{19}$$

where

$$X(m) = \frac{2(1+m)m^2}{(3+3m)^2 + (3m+2m^2)^2} \tag{20}$$

and

$$m = (\sigma_2\mu\omega/2)^{1/2}a.$$

The total power P crossing the wall of the cavity is found by a suitable integration.

$$P = 2\pi a^2\int_0^\pi p(\theta)\sin\theta d\theta = \frac{3}{4\pi}\frac{(IdA)^2}{a^3}\mu\omega X(m). \tag{21}$$

For values of m much less than one, the factor $X(m)$ is approximately given by

$$X(m) \simeq 2m^2/9 = \sigma_2\mu\omega a^2/81. \tag{22}$$

The power is then approximately given by

$$P \simeq (IdA)^2(\mu\omega)^2\sigma_2/12\pi a. \tag{23}$$

It is quite evident that the power dissipated becomes infinite as the radius of the cavity approaches zero. The physical situation of course requires that the cavity radius a is a finite quantity; however, it may be concluded that an insulated cavity enclosing a magnetic dipole-type antenna does specify the amount of power required to maintain a given field in the exterior conducting region.

The author[6] has also considered the transient fields of a magnetic dipole source, immersed in a conducting medium, energized by a positive step-function current. These transient fields might have been derived from the steady-state equations (14), (15), and (16) of this paper.

[6] J. R. Wait, *Geophysics*, vol. 16, p. 213; April, 1951.

The Behavior of Rectangular Hysteresis Loop Magnetic Materials Under Current Pulse Conditions*

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Summary—This paper studies the effect of variation of magnetic and physical parameters on the time needed to change a magnetic toroid from a condition of residual flux density to the opposite condition of saturation flux density. The method of measurement is described, and the experimental results given. These results are compared with results predicted from a solution of the magnetic skin-effect equation. Curves are given which show the variation of switching time with applied magnetic field strength. An equivalent circuit is proposed which dissipates the same amount of energy as the core does during its switching time.

THERE HAS BEEN considerable interest recently in the use of thin-ribbon magnetic toroids in digital computers.[1] Attention has been directed to rectangular hysteresis loop materials because they have two stable, well-defined states of positive residual flux density and negative residual flux density to store binary information. (See Fig. 1.) Suitable materials are cold rolled nickel iron alloys (trade name: orthonik, deltamax, permeron, permenorm 5,000z) and oriented silicon steels (trade name: silectron, hipersil, microsil).

Fig. 1 indicates the pertinent magnetic quantities referred to in this paper. H_M is the amplitude of the applied field strength, B_M is the corresponding flux density, and B_R is the residual flux density corresponding to $H_M = 5$ oersteds. H_C is the coercive force, corresponding to a previously applied $H_M = 5$ oersteds, and $\Delta B/\Delta H$ is the average slope of the sides of the hysteresis loop.

The term "switching time," used in this paper, means the time it takes to change the state of magnetism of a core from negative residual flux density ($-B_R$) to positive saturation flux density ($+B_M$), with the application of a step function of applied field strength (H_M). The factor which determines switching time is the eddy-current losses. A theoretical analysis in which saturation effects were neglected, and the B-H loop linearized, indicated that switching time should vary directly as the square of the thickness of the material, directly as the conductivity, directly as $\Delta B/\Delta H$, and directly as some power between the square root and the first power of H_C, depending on the value of H_M.[2]

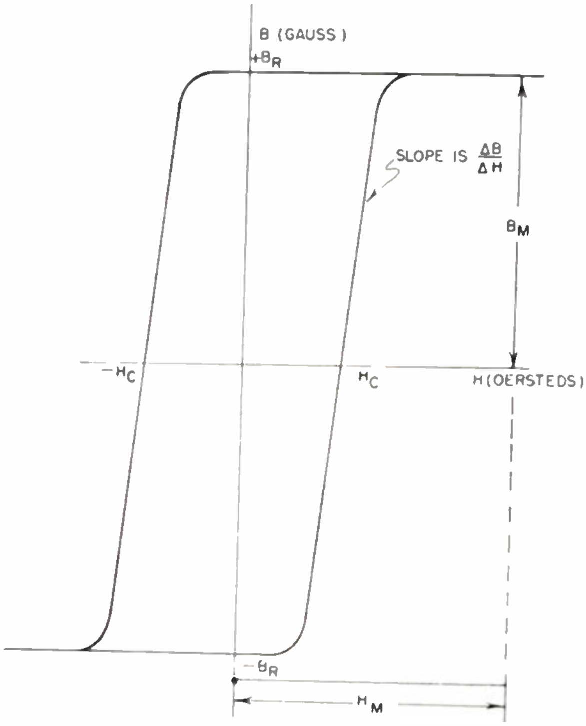

Fig. 1—Definitions of magnetic parameters on the dc hysteresis loop.

The object of this paper is to study experimentally how switching time varies with H_M, how it varies with thickness of material, and how it varies with $\Delta B/\Delta H$.

The cores studied were of orthonik, a rectangular hysteresis loop material, and they were wound-ribbon toroids whose physical dimensions are shown in Table I.

The dc hysteresis loops of cores number 182, 183, 185, and 187 are shown in Figs. 2, 3, 4, and 5, respectively. Notice that cores 182 and 185 have identical magnetic tape thickness, and identical coercivity, but that the slopes of the sides of their hysteresis loops differ by a factor of 3.5 to 1.

TABLE I
Physical dimensions of experimental cores.

Core number	182	184	185	183	186	187
Magnetic tape thickness (inches)	0.0011	0.0011	0.0011	0.0005	0.0005	0.0005
Magnetic tape width (inches)	0.125	0.125	0.125	0.125	0.125	0.125
Number of wraps of magnetic tape	17	14	14	27	20	22
Magnetic path length (cms)	5.47	5.42	5.42	5.40	5.40	5.40
Cross sectional area of iron (cm²)	0.0156	0.0129	0.0128	0.0111	0.0084	0.0091

* Decimal classification: R282.3. Original manuscript received by the Institute, August 31, 1951; revised manuscript received June 11, 1952.

† Magnetics Research Company, Chappaqua, N. Y.

[1] An Wang, "Magnetic delay line storage," Proc. I.R.E., vol. 39, pp. 401–407; April, 1951.
W. Papian, A Coincident Current Magnetic Memory Unit, Project Whirlwind Report R 192, MIT Digital Computer Laboratory, Cambridge, Mass.; Sept. 8, 1950.

[2] Bozorth, "Ferromagnetism," D. Van Nostrand Company, Inc., New York, N. Y., pp. 784–788; 1951.

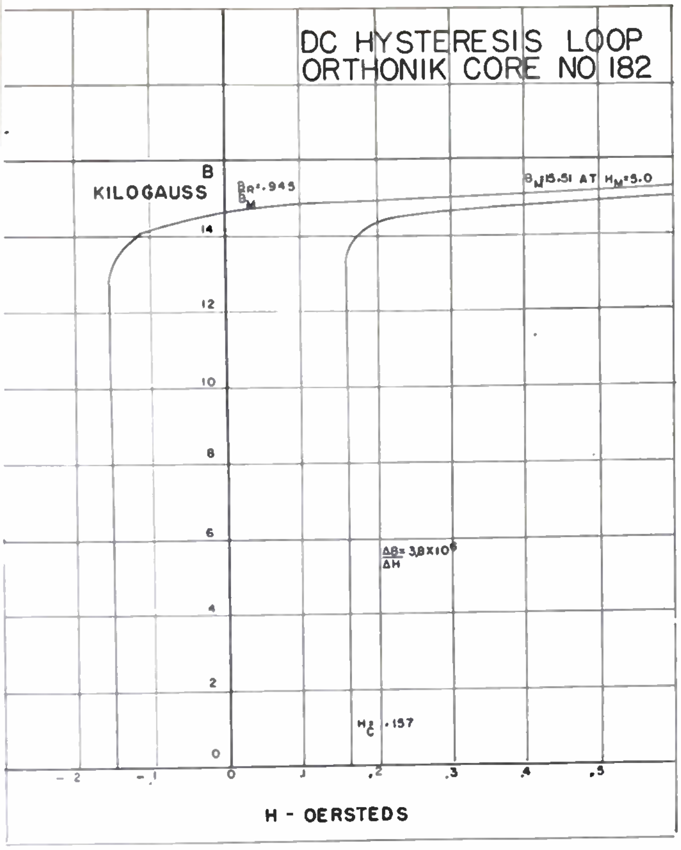

Fig. 2

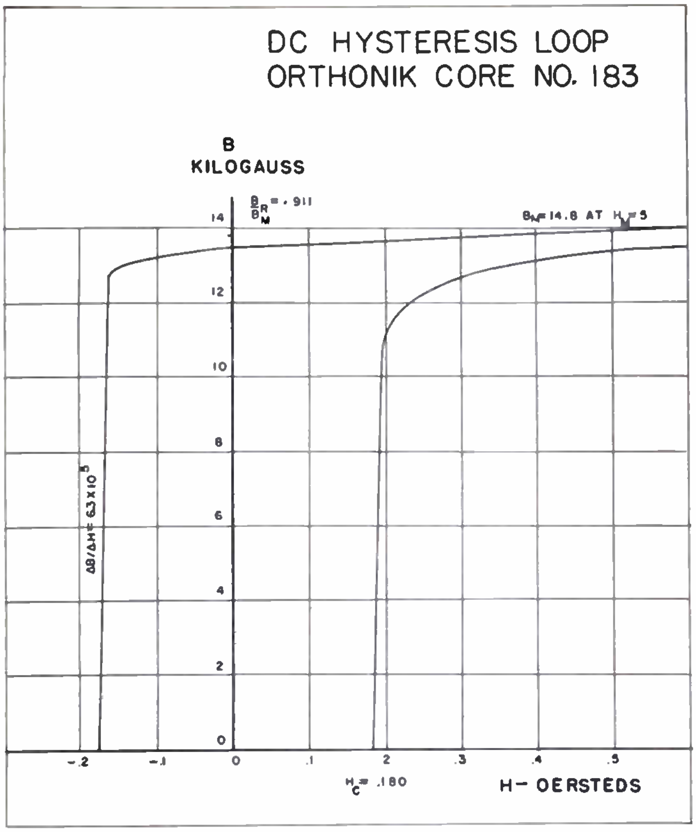

Fig. 3

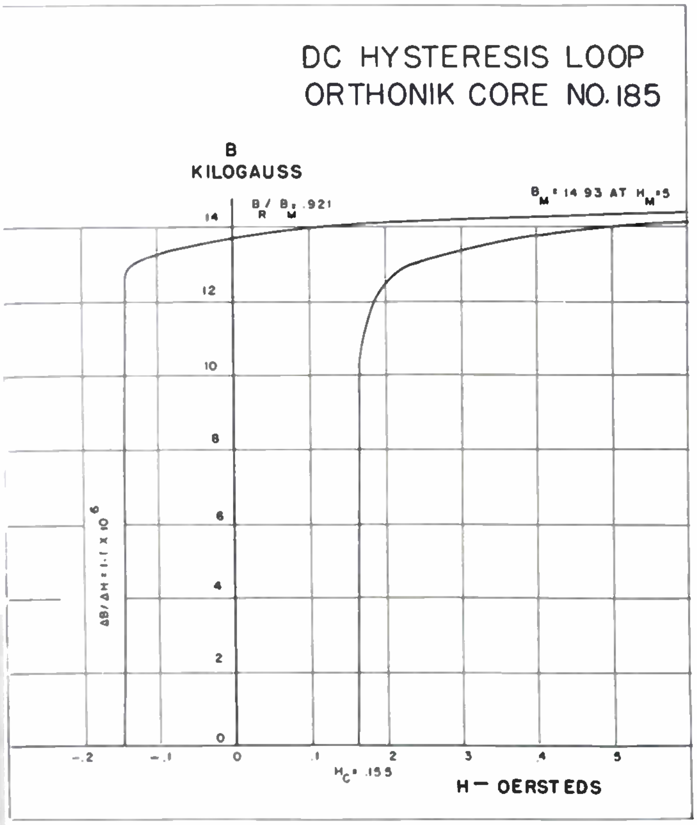

Fig. 4

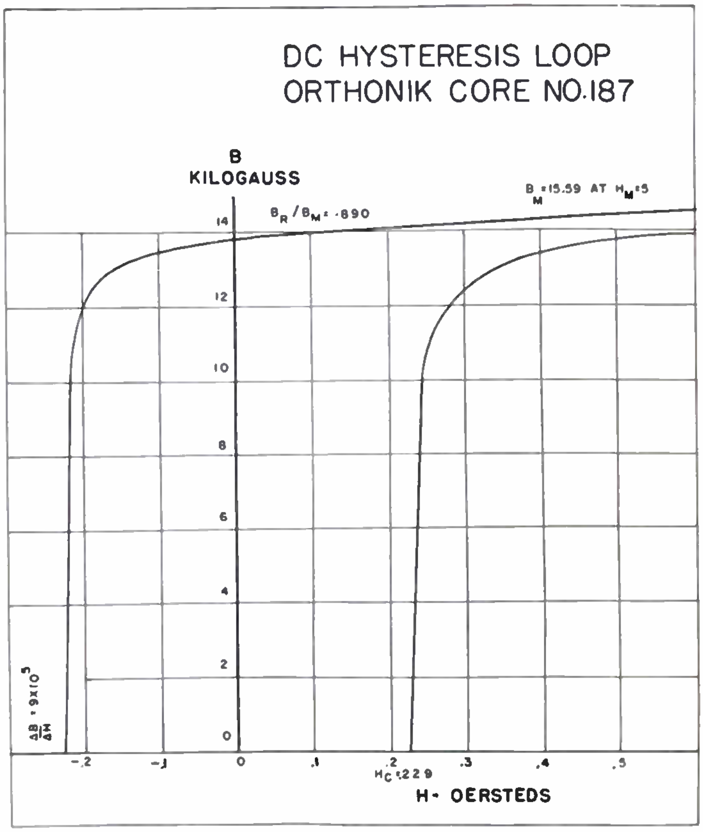

Fig. 5

The winding configuration used in the measurements is shown in Fig. 6. Current pulses were passed through two windings to produce an alternating magnetomotive force of equal positive and negative amplitude. In this way, the cores were cycled between values of negative and positive saturation, and the resulting flux pattern observed through the means of a third winding connected to an integrator circuit. Each current coil was returned to ground through a 10-ohm resistance. The voltage across the 10-ohm resistance was measured, and the driving currents adjusted to give an appropriate value of H_M.

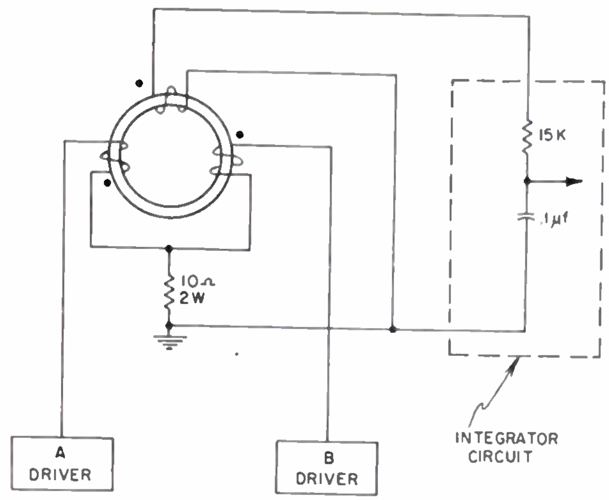

Fig. 6—Experimental arrangement used to obtain applied field strength versus switching-time characteristic.

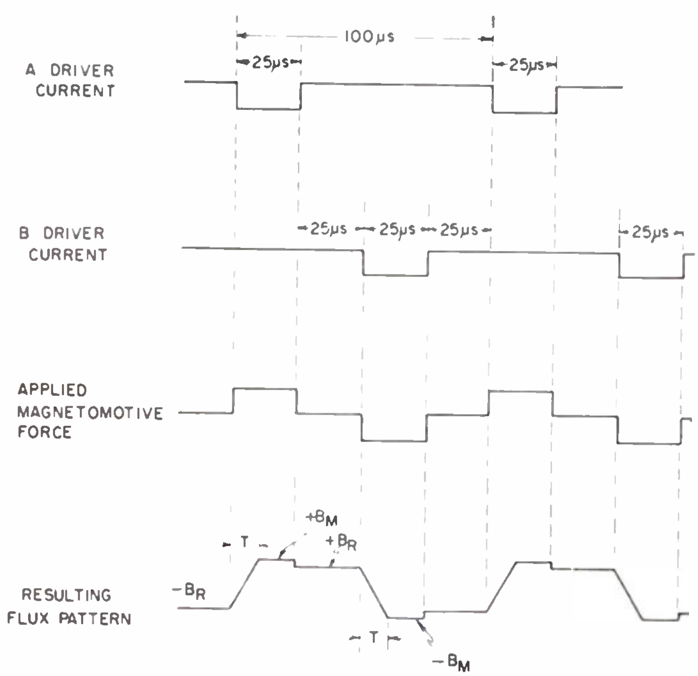

Fig. 7—Time relationship of A driver current, B driver current, applied magnetomotive force, and resulting flux pattern.

Fig. 7 shows the timing diagram of the A and B driver current pulses, the magnetomotive force, and the resulting flux pattern. At the beginning of the A driver current pulse, the core is at negative residual flux density. The flux starts to change to positive saturation flux density $(+B_M)$, reaching this value at switching time, T, after the initiation of the A current pulse. At this time the flux change is complete, and the core remains at $+B_M$ until the end of the A current pulse, when it drops to the value $+B_R$. It remains at the value $+B_R$ until the initiation of the B current pulse, at which time it begins to change to $-B_M$. The flux completes this change at switching time, T, also, for equal values of positive and negative H_M. The flux remains at $-B_M$ until the end of the B current pulse, at which time it assumes the value $-B_R$. The complete cycle is then repeated.

The core under test is connected to the A and B drivers, values of H_M are adjusted, and the resulting switching time measured. The measured range of H_M was between 0.5 and 5 oersteds.

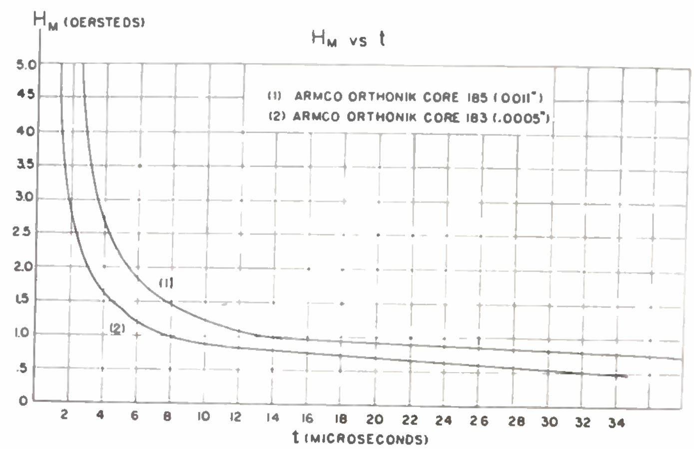

Fig. 8—Applied field strength versus switching-time curves of cores 183 and 185.

Fig. 8 is a plot of H_M versus t for cores number 183 and 185. In the region of H_M from 1.5 to 5 oersteds, the curves are approximately similar in shape, but displaced in time by a factor of about 2. Referring back to the static B-H curves of cores 183 and 185, $\Delta B/\Delta H$ for $183 = 6.3\times 10^5$, and $\Delta B/\Delta H = 1.1\times 10^6$ for 185. According to linear theory, the ratio of the switching time of the 1.1-mil thick sample to the switching time of the $\frac{1}{2}$-mil thick sample should be approximately

$$\frac{\left(\frac{\Delta B}{\Delta H} d^2\right) 185}{\left(\frac{\Delta B}{\Delta H} d^2\right) 183} = 8.45.$$

The experimental result is quite different, the switching time of the 1.1-mil sample being larger than the $\frac{1}{2}$-mil sample by a factor of 1.8 to 2.0, depending on the value of H_M. Fortunately, it is possible to evaluate the effect of $\Delta B/\Delta H$. An examination of the static B-H loop of cores 182 and 185 shows that they have identical values of H_C, but that $\Delta B/\Delta H$ for core $182 = 3.8\times 10^6$ and for core $185 = 1.1\times 10^6$. They differ by a factor of 3.45. Ac-

TABLE II

H_M versus T

	0.0005-inch tape cores				0.0011-inch tape cores				
Core number	183	186	187	Average	182	184	185	Average	
	T	T	T	(T) 0.5	T	T	T	(T) 1.1	$R=\frac{(T)\ 1.1}{(T)\ 0.5}$
Hm									
0.5	35	30	30	31.7	70	66	64	66.7	2.10
1	8.0	6.5	7.0	7.2	15	15	15	15.0	2.08
1.5	4.5	3.5	4.0	4.0	8.5	9.0	8.0	8.5	2.13
2.0	2.8	2.6	2.7	2.7	5.5	6.0	5.3	5.6	2.07
3.0	1.8	1.6	1.7	1.7	4.0	3.8	3.6	3.8	2.24
4.0	1.4	1.3	1.5	1.4	2.8	3.0	2.8	2.9	2.07
5.0	1.3	1.2	1.3	1.3	2.6	2.8	2.6	2.7	2.08

cording to linear theory, this difference should show up n their relative switching time, since both cores are nade of 1.1-mil material. Core 185 should switch approximately 3.45 times faster than 182. Experimental lata does not bear this out (see Table II). The switching ime of core 182 does not differ from 185 by more than 10 per cent over the entire range of H from 0.5 to 5 oersteds. The conclusion is that $\Delta B/\Delta H$ does not play a significant part in determination of switching time.

The best estimate of the effect of change in thickness on switching time will be made by taking the average switching time of the three cores wound with 0.0011-inch thick tape and comparing it with the average of the three cores wound with the 0.0005-inch thick tape over the measured range of H. Table II gives the individual core data and the averages for each set. Alongside the averages is plotted a coefficient R which is the ratio of the average switching time of the 0.0011-inch material to that of the 0.0005-inch material. $R=2.1$, and is independent of H_M over the range 0.5 to 5 oersteds. The ratio of the thickness of the two materials is 2.2. Therefore, to a high degree of approximation, the reduction in switching time is proportional to that in thickness.

An equivalent input impedance of an orthonik core operating under current pulse conditions will now be derived. First, it will be shown that the magnetic core is a dissipative element; second, an average resistance $\overline{R}$ will be found such that during the switching-time interval the same amount of energy is dissipated in $\overline{R}$ by the input current as is dissipated by the magnetic core. The following symbols will be used:

A —Cross sectional area of iron in cm².
L —Mean path length of iron in cm.
N —Number of turns on the input winding.
I —Current flowing in input winding N (amperes).
T_0 —Time duration of current pulse (seconds).
T —Switching time.
R —Equivalent input resistance of core (ohms).
V —Voltage across input winding N.
H_M—Applied field strength (oersteds).
E_{in}—Input energy to core.
E_b —Energy given back by the core at the end of the current pulse.
E_d —Energy dissipated in the core.

1. $$E_{in} = \int_0^{T_0} VI dt = NAI \times 10^{-8} \int_0^{T_0} \frac{dB}{dt} dt$$
$$= NAI(B_R + B_M) \times 10^{-8}.$$
2. In CGS units
$$E_{in} = 0.794 H_M(B_R + B_M)AL \times 10^{-8}.$$
3. $$E_b = 0.794 K H_M(B_M - B_R)AL \times 10^{-8}.$$

K is a number between 0 and 1 and depends on the curve joining B_R to B_M.

4. The energy dissipated in switching the core is the difference between the input energy and the energy returned by the core at the end of the input current pulse.
$$\therefore\ E_d = E_{in} - E_b = E_{in}\left(1 - \frac{E_b}{E_{in}}\right).$$
5. $$\frac{E_b}{E_{in}} = K\left(1 - \frac{B_R}{B_M}\right) \Big/ 1 + \frac{B_R}{B_M}.$$

For typical 1-mil orthonik cores $B_R/B_M=0.92$ so that $E_b/E_{in} \leqq 0.05$.

6. Therefore, E_d is approximately equal to E_{in}. This means that the magnetic core is a dissipative element when used under these current pulse conditions.
7. From the definition of $\overline{R}$, the energy dissipated in it by the current I, during the switching time T, is
$$I^2\overline{R}T = \frac{H_M{}^2L^2\overline{R}T}{(1.26)^2N^2}.$$
8. Equating the energy dissipated in the equivalent resistance to the energy dissipated in the core and solving for $\overline{R}$
$$\overline{R} = \frac{1.26N^2(B_R + B_M)A \times 10^{-8}}{H_M TL}.$$

The value of $\overline{R}$ is inversely proportional to the product of $H_M T$. This product is approximately constant for applied field strengths between 1.5 and 5 oersteds. Therefore, $\overline{R}$ is approximately constant in this range. $\overline{R}$

can be computed from the physical dimensions of the core, the magnetic parameters, and the switching time versus applied field curve.

This equivalent resistance concept has proved very useful in the design of magnetic shift registers and magnetic memory units for computer use.

Conclusions

(1) The ratio of the switching time of 0.0011-inch orthonik to 0.0005-inch orthonik is 2.1, and is independent of H_M in the measured range.

(2) Switching time varies inversely with H_M in the range 1.5 to 5 oersteds.

(3) Switching time is independent of $\Delta B/\Delta H$ for large values of $\Delta B/\Delta H$.

(4) The solution of the skin-effect equation for a linear B-H loop of slope, $\Delta B/\Delta H$, does not give results which agree with experiment.

(5) The core under current pulse conditions is a dissipative element.

(6) An average resistance $\overline{R}$ can be found that dissipates the same amount of energy as the core does during the switching-time interval. This resistance is approximately constant in the range 1.5 to 5 oersteds applied H_M.

Acknowledgment

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Correspondence

A Note on Class C Power-Amplifier Design* (4356)

One of the most straightforward methods of practical class C power-amplifier design is that due to Wagener,[1] particularly when carried out in a series of systematic steps, as indicated by Eastman.[2] This method, though it assumes linear operation during the time-plate current flows, produces results which are within the degree of accuracy usually required in most design problems.

When a preliminary design has been completed, it is often necessary for the designer to modify his design slightly by changing some one or more of the design parameters. The effects of these modifications upon the performance and operation of the amplifier may not be readily discernible without repeating a considerable portion of the analysis.

In Fig. 1 the pertinent design parameters

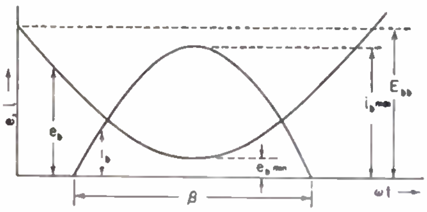

Fig. 1

are indicated. It is assumed that the pulse of plate current i_b is a portion of a sine wave of angular duration β and reaches a maximum value indicated by i_{bmax}. The instantaneous plate voltage e_b reaches a minimum e_{bmin} at the time i_b goes through its maximum. It is further assumed that the voltage across the load $(E_{bb}-e_b)$ is sinusoidal and that the load impedance is resistive at the fundamental frequency.

For a current pulse of angular duration β and maximum amplitude i_{bmax}, a Fourier analysis yields

$$I_b = \frac{i_{b\max}}{\pi}\left(\frac{\sin\beta/2 - (\beta/2)\cos\beta/2}{1-\cos\beta/2}\right) \quad (1)$$

$$I_{pm} = \frac{i_{b\max}}{2\pi}\left(\frac{\beta - \sin\beta}{1-\cos\beta/2}\right) \quad (2)$$

$$\frac{I_{pm}}{I_b} = \frac{\beta - \sin\beta}{2(\sin\beta/2 - (\beta/2)\cos\beta/2)}, \quad (3)$$

where

I_b = dc component of the pulse

I_{pm} = peak value of the fundamental ac component of the pulse.

The plate-circuit efficiency is given by

$$\eta = \frac{P_{\text{out}}}{P_{\text{in}}} = \frac{E_{pm}I_{pm}}{2E_{bb}I_b} = \left(\frac{1-k}{2}\right)\left(\frac{I_{pm}}{I_b}\right). \quad (4)$$

where

100η = plate-circuit efficiency in per cent

E_{bb} = dc plate voltage

$E_{pm} = (E_{bb}-e_{bmin})$ = peak value of fundamental ac component of voltage across the load

$k = e_{bmin}/E_{bb}$.

By making use of the parameters indicated in (1) through (4), the nomogram of Fig. 2 has been constructed. Through the use of this nomogram, one can quickly and easily observe the effect upon the performance of the amplifier brought about by a variation in any one or more of the parameters.

It may also be used to facilitate the original design. For instance, suppose an efficiency of 82 per cent is desired with a value of $k=5$ per cent. Entering the nomogram at (a), the assigned value of η, one proceeds horizontally to (b), the intersection of η with k; thence vertically upward passing

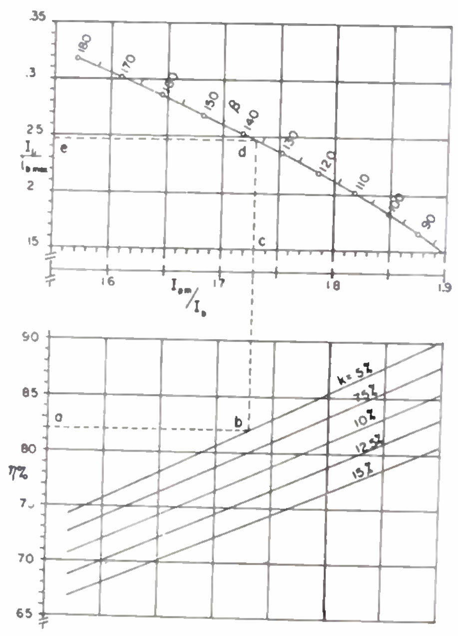

Fig. 2

through (c) to (d), the intersection with the β curve; thence horizontally to (e).

The results obtained from the above operation for this amplifier are then at

(a)	$\eta = 82$ per cent
(b)	$k = 5$ per cent
(c)	$\frac{I_{pm}}{I_b} = 1.73$
(d)	$\beta = 137°$
(e)	$\frac{I_b}{i_{b\max}} = 0.248$.

H. J. Scott
University of California
Berkeley 4, Calif.

* Original manuscript received by the Institute April 7, 1952.

[1] W. G. Wagener, "Simplified methods for computing performance of transmitting tubes," Proc. I.R.E., vol. 25, pp. 47–77; January, 1937.

[2] A. V. Eastman, "Fundamentals of Vacuum Tubes," McGraw-Hill Book Co., Inc., New York, N. Y., 3rd. ed., pp. 407–410; 1949.

Correspondence

On the Isotropic Artificial Dielectric* (4357)

In a recent paper, Corkum[1] has shown curves of the variation of refractive index with the length of two artificial dielectric samples, made up of metal spheres and quartz spheres, respectively. In the same paper, mention is made of the "interface problem" which arises when it is assumed that the physical boundary of the artificial dielectric actually defines the effective interface between the sample and free space.

There is strong evidence that the two effects are interrelated, i.e., that refractive-index measurements made on small samples deviate from the theoretical value precisely because an additional phase shift is introduced at the interface. The discrepancy may be expressed by an adjustment of the sample size, as Corkum suggests. This procedure yields an "equivalent length" of artificial dielectric having the refractive index of a very large sample.

Alternately, the effective sample size may be taken to be equal to the actual size, and an adjustment made in the value of refractive index. A first-order theory has been proposed in the report of an investigation[2] carried out at Yale University under the sponsorship of the Air Force Cambridge Research Center. With this method of attack, an "effective refractive index" is defined by

$$n_{eff} = n + \frac{x}{d}, \qquad (1)$$

where d is the sample length and x is a factor which depends on the geometry of the array and the properties of the obstacle material.

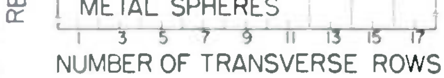

Fig. 1—Measured values of refractive index and theoretical curves for two artificial dielectrics.

The measurements of refractive index n made by Corkum are compared with the theoretical values of n_{eff} obtained from (1) in the accompanying table and also in Fig. 1 above. The measured n evidently approaches values of 1.195 and 1.221 for the metal and quartz spheres, respectively. The corresponding values of x used in the theoretical computations were 0.095 and 0.084; the units were chosen so that the sample length d could be expressed simply in terms of the number of obstacle rows making up the sample. The curves shown in Fig. 1 are actually plots of (1), and are seen to fit the experimental points exceedingly well.

TABLE I

Number of rows	Metal spheres		Diel. spheres	
	n (meas.)	n_{eff}	n (meas.)	n_{eff}
1	1.284	1.290	1.311	1.305
2	1.245	1.243	1.263	1.263
3	1.231	1.227	1.249	1.249
4	1.222	1.219	1.242	1.242
5	1.220	1.214	1.237	1.238
6	1.201	1.211	1.234	1.235
7	1.205	1.209	1.232	1.233
8	1.206	1.207	1.231	1.232
9	1.198	1.206	1.056	1.230
10	1.197	1.205	1.228	1.229
11	1.202	1.204	1.227	1.229
12	1.199	1.203	1.227	1.228
13	1.118	1.202	1.226	1.227
14	1.202	1.202	1.225	1.227
15	1.201	1.201	1.224	1.227
16	1.200	1.201	1.224	1.226
17	1.202	1.201	1.223	1.226

The value of $x = 0.095$ for the array comprising *metal* spheres was obtained as the result of an apparently successful attempt to predict this quantity on the basis of scattering theory. For the array of *quartz* spheres, the same theory yielded a value lower than the $x = 0.084$ used in the present contribution; the latter value must be therefore considered to be of a semi-empirical nature. The theory has not been published in its entirety partly because of this uncertainty, but mainly because of insufficient experimental data available for testing the theory conclusively. A brief mention of it appeared in a paper published in London earlier this year.[3]

CHARLES SÜSSKIND
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* Original manuscript received by the Institute May 28, 1952.

[1] R. W. Corkum, "Isotropic artificial dielectric," Proc. I.R.E., vol. 40, pp. 574–587; May, 1952.

[2] C. Süsskind, "Investigation of obstacle-type artificial dielectrics," final report under Contract No. AF 19(122)-270, Yale University, New Haven, Conn.; June, 1951.

[3] C. Süsskind, "Obstacle-type artificial dielectrics for microwaves," *Jour. Brit. IRE*, vol. 12, pp. 49–62; January, 1952.

Linear Detection of Amplitude-Modulated Signals* (4358)

An interesting mathematical treatment of "linear detection" of amplitude-modulated signals seems to have been overlooked in the literature. This treatment, based on a suggestion by McLean of the Rome Air Development Center, enables one to express the output of an appropriate type of "linear detector" in a form which clearly indicates the theoretical limitations of the process. The type to be considered is an ideal rectifier with a resistance load followed by an isolation device and then by an ideal low-pass filter. The necessity of including an isolation device prevents our conclusions from being directly applied to circuits like the simple diode directly feeding an RC load. Also, it is to be recognized that the term "linear detection" is a misnomer since detection is never a linear process.

The analysis begins by assuming a modulating function given by

$$f(t) = E_m \cos \omega_a t. \qquad (1)$$

The amplitude-modulated wave is

$$e(t) = E_c(1 + m \cos \omega_a t) \cos \omega_c t. \qquad (2)$$

It should be noted that a component of $e(t)$ with a frequency ω_a does not exist, so that a true linear operation on $e(t)$ cannot recover $f(t)$.

After detection, a component of frequency ω_a will exist. For the case of a half-wave rectifier in the system described above, this may be shown as follows: The output of the detector $e_0(t)$ is the positive part of the amplitude-modulated wave, (2), being zero for those values of time when (2) is negative. Since $(1+m \cos \omega_a t)$ is always positive (for $m<1$), $e_0(t)$ is given by the product of $(1+m \cos \omega_a t)$ and the half-wave rectified carrier. Expressing the half-wave rectified carrier in a Fourier series, the output becomes

$$e_0(t) = E_c(1 + m \cos \omega_a t)(1/\pi + 1/2 \cos \omega_c t + 2/3\pi \cos 2\omega_c t + \cdots). \qquad (3)$$

Applying standard trigonometric identities, $e_0(t)$ is seen to contain components with frequency dc, ω_a, $\omega_c - \omega_a$, ω_c, $\omega_c + \omega_a$, $2\omega_c - \omega_a$, $2\omega_c$, $2\omega_c + \omega_a$, and so on. An ideal low-pass filter will separate the desired component of frequency ω_a from the remaining undesired components, provided that

$$\omega_c - \omega_a > \omega_a \text{ or } \omega_a < 1/2\omega_c. \qquad (4)$$

If a full-wave rectifier is used in the system described above, the output will be given by the product of $(1+m \cos \omega_a t)$ and the full-wave rectified carrier. An analysis similar to that used for the half-wave case reveals that the output has components with frequencies dc, ω_a, $2\omega_c - \omega_a$, $2\omega_c + \omega_a$, and so on. An ideal low-pass filter will separate the desired component of frequency from the undesired components, provided that

$$2\omega_c - \omega_a > \omega_a \text{ or } \omega_a > \omega_c. \qquad (5)$$

The preceding analysis shows that an ideal modulating and demodulating system of the type described above is limited in operation by the requirement that the highest frequency component of the modulating function satisfy (4) for the half-wave case and (5) for the full-wave case.

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* Original manuscript received by the Institute April 7, 1952.

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

Sidney Applebaum was born on August 31, 1923, in New York City. He received the degree of B.E.E. from the College of the City of New York in 1944 and immediately joined the Western Electric Company in the radar test planning section. From 1945 to 1947, he served as a radar technician in the Army Air Force.

S. Applebaum

Since 1947 he has been with the General Electric Company. He is a graduate of the three-year General Electric Advanced Engineering Program. After spending an additional year on the program staff, he joined the Electronics Division where he is presently engaged in advanced development in the Electronics Laboratory.

Mr. Applebaum is a member of Tau Beta Pi, Eta Kappa Nu, and R.E.S.A.

❖

Marcus A. Acheson (A'36–M'37–F'41) is a native of Denison, Tex. He received the B.S. deg ee in electrical engineering from Rice Institute in 1924, and the same year was employed by the General Electric Company at Schenectady, N. Y., as engineer in charge of developing high-voltage electronic devices.

M. A. Acheson

Ten years later Mr. Acheson accepted a position as division engineer at Sylvania in Clifton, N. J. The following February he was appointed manager of design and development of proximity fuse tubes for the Navy Bureau of Ordnance, and received an award for exceptional service from the Bureau.

From 1942 until 1948, when he was appointed chief engineer of the Radio Tube Division of Sylvania, Mr. Acheson served as manager of the Advanced Development Laboratories of Central Engineering. In August, 1950 he was transferred to the staff of E. Finley Carter, vice president in charge of engineering.

Mr. Acheson has approximately thirty patents on radio transmitter circuits, water-cooled power tubes, and tubes for television transmission.

❖

D. A. Alsberg (A'46–M'48–SM'52) was born in Kassel, Germany, on June 5, 1917. He obtained his undergraduate training at the Technische Hochschule in Stuttgart, Germany, and completed his work there in 1938.

D. A. Alsberg

After coming to the United States, he was engaged in graduate study at the Case Institute of Applied Science (now Case Institute of Technology) from 1939 to 1940. Following three years as development engineer with several companies in Ohio, he entered the United States Army and served at the Aberdeen Proving Ground in Maryland, and in Europe.

In 1945 Mr. Alsberg joined Bell Telephone Laboratories, where he has been concerned with phase, transmission, and related measurements problems in connection with coaxial-cable carrier and microwave radio-relay development.

❖

Ivan S. Coggeshall (A'26–M'29–F'42) was born in Newport, R. I., on September 30, 1896, and attended Worcester Polytechnic Institute, where he studied electrical engineering. He received the honorary degree of Doctor of engineering from the Institute in 1951.

Ivan Coggeshall

Since 1917 Mr. Coggeshall has been employed by the Western Union Telegraph Co. in New York, specializing in ocean cables. In 1927 he became general traffic manager of the Company's overseas communications; and in 1952, director of planning of international communications.

Mr. Coggeshall is a Lt. Commander in the U. S. Naval Reserve, and served on the Cable Committee of the Board of War Communications during World War II. He is a director of the Mexican Telegraph Co., and the Institute of Radio Engineers, having been president in 1951, and is a member of Tau Beta Pi and the A.I.E.E.

❖

Marcel J. E. Golay was born on May 3, 1902, in Switzerland. He received the Baccalaureat Scientifique from the Gymnase de Neuchatel in 1920, the licenciate in electrical engineering from the Federal Institute of Technology at Zurich in 1924, and the Ph.D. in physics from the University of Chicago in 1930.

From 1924 to 1928 Dr. Golay worked on telephone cables for the Bell Telephone Laboratories. In 1930 and 1931 he worked on automatic telephone systems for the Automatic Electric Company, and since 1931 has been a member of the Signal Corps Laboratories, Fort Monmouth, N. J., working on subaqueous acoustics, terrestrial sound ranging, heat detection, radio relays, and electronic controls. He is now chief scientist of the Squier Signal Laboratory.

M. J. E. Golay

In 1946 Dr. Golay disclosed a new attack on the ideal electrical filter problem and also described a new detector for infrared and short radio waves. In 1948 he announced the principle of multislit spectrometry, which alleviates the intensity problems in far infrared spectrometry. He received the Harry Diamond award in 1951.

❖

Robert T. Hamlett (SM'46) was born in February, 1902 in Fulton, Ky. He received the B.S. degree in electrical engineering in 1928 from the University of Illinois.

R. T. Hamlett

From 1928 to 1941 Mr. Hamlett served in the engineering division of the Western Union Telegraph Company, where some of his research resulted in seven patents covering improvements in electrical networks and communication equipment. Earlier, he had worked in various capacities for power and light and telephone companies.

In 1941 Mr. Hamlett joined the Sperry Gyroscope Company as a technical writer, and in 1942 was made publications editor in charge of aeronautical, aircraft armament, and radio products. In 1945 he was transferred from the Brooklyn to the Garden City Research Laboratories, where he served as engineering publications editor, assuming charge of the publications department in 1947 after Sperry activities were concentrated at Great Neck, L. I., N. Y.

❖

Keith C. Harding was born in Fort Maginus, Mont., on May 31, 1921. He received the B.S. degree in electrical engineering from Indiana Technical College in 1943, and took special courses in electron-tube theory, design, and application at the University of Maryland in 1947 and 1948.

Mr. Harding has been employed since 1943 in the Bureau of Ships, Department of the Navy. From 1943 to 1946, he was asso-

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

ciated with the development of shipboard motor controls and instruments. Since 1947, he has served in the Electronics Division, Electron Tube Section, presently being responsible for Navy tube-development programs and procurement specifications for receiving-type tubes.

K. C. Harding

Mr. Harding is a member of the Senior Engineers Association of the Bureau of Ships.

P. M. Honnell (A'29–M'41–SM'43) was born on January 28, 1908 in Paris, France. He received the B.S. degree in electrical engineering and the E.E. degree from Texas A. and M. College. His graduate studies were continued at the Technischen Hochschule in Vienna; Massachusetts Institute of Technology, where he received the M.S. degree in electrical engineering; California Institute of Technology, where he received the M.S. degree; and St. Louis University, where he received the Ph.D. degree in geophysics.

P. M. Honnell

As a member of the technical staff of Bell Telephone Laboratories for 1930 to 1933, Dr. Honnell developed methods of measuring the harmonic radiation from transatlantic short-wave radio-telephone transmitters. In the Geophysical Division of the Texas Company from 1934 to 1938, he designed apparatus for seismic prospecting.

During World War II, Dr. Honnell was at the Signal Corps School, Fort Monmouth, N. J., and then in the department of chemistry and electricity at the U. S. Military Academy in West Point. Besides teaching and administrative duties, he designed and supervised the installation of the electronics laboratory at the Academy.

Dr. Honnell has also taught at Southern Methodist University and the University of Illinois. He is presently professor of electrical engineering at Washington University, St. Louis, Mo.

A colonel in the Signal Corps, U.S.A.R., Dr. Honnell was awarded the Legion of Merit medal in 1946 for his services at West Point. He is a member of the A.I.E.E., Soc. Exploration Geophysics, Seismological Soc. Amer., Amer. Geophysical Union, Tensor Club of Great Britain, AAAS, Tau Beta Pi, and Sigma Xi.

R. E. Horn (S'52) was born in St. Louis, Mo. on November 30, 1927. He received the B.S. degree in electrical engineering from Washington University in 1950. At that time he was awarded a Fortesque Fellowship from the Westinghouse Electric Corporation, and continued his studies at the University, receiving the M.S. degree in electrical engineering in 1951.

R. E. Horn

Prior to graduation, Mr. Horn spent two years as an electronic technician in the U. S. Navy, and now holds the grade of ensign, U.S.N.R. At present he is an instructor at Washington University and an engineer on a Naval Ordnance research project.

Mr. Horn is a member of Tau Beta Pi and Sigma Xi.

William C. Jakes, Jr. (S'43–A'49) was born in Milwaukee, Wis., on May 15, 1922. He received the B.S.E.E. degree from Northwestern University in 1944, then entered the Navy and served for two years as airborne electronics officer. He returned to Northwestern University in 1946 for graduate study, receiving the M.S. degree in 1947 and the Ph.D. degree in 1949. During part of this time he was employed by the Microwave Laboratory at Northwestern as a research associate.

W. C. Jakes, Jr.

Since July, 1949, Dr. Jakes has been a member of the technical staff of the Bell Telephone Laboratories, Inc., engaged in microwave propagation and antenna studies. He is a member of Sigma Xi, Eta Kappa Nu, and Pi Mu Epsilon.

C. R. Knight (M'45) was born in Salt Lake City, Utah, on September 25, 1918. He received the B.S. degree in electrical engineering from the University of Utah in 1940.

C. R. Knight

From 1940 to 1951 Mr. Knight was employed by the General Electric Company in various capacities in the electron-tube field. His assignments included application engineering and advanced tube-development work. At present he is assistant project director of the Military Tube Project, Aeronautical Radio, Inc.

Mr. Knight is a member of Phi Kappa Phi and Tau Beta Pi.

Paul Mandel (A'45) was born on February 15, 1906, in Szolnok, Hungary. In 1929, he received the diploma of electrical engineering from the Polytechnical High School of Berlin, Germany, and joined the staff of the Dr. G. Seibt Laboratories, engaging in research work on acoustical systems and broadcasting receivers until 1931, when he became associated with the Sachsenwerk A. G. in Dresden as head of the broadcasting receiver development, remaining until 1933. He was engaged from 1936 in television research work for the Compagnie des Compteurs, Montrouge, France, as head of the research group on high-definition television systems, large-screen television projectors, and television receivers. In 1949, he joined the Television Laboratories of the Radio-Industrie in Paris, for further development of high-definition television systems as chief of a development section.

Paul Mandel

Mr. Mandel is a member of the Société des Radioélectriciens, of the Société Française des Électriciens, and a member of the Comité Mixte de Télévision.

Eleanor M. McElwee (M'51) was born in New York, N. Y., on May 15, 1924. She received the A.B. degree in English and mathematics from Ladycliff College in 1944, and has done additional work in science at Cooper Union.

E. M. McElwee

Miss McElwee was employed by the Western Electric Tube Shop, New York City, from 1944 to 1947 as assistant product engineer. She came to Sylvania Electric Products Inc. in 1947, and was in charge of the statistical analysis program and life-testing tubes for the engineering analysis section of the Product Development Laboratories at Kew Gardens, N. Y.

Markus Nowogrodzki (S'47–A'49) was born on September 13, 1920 in Warsaw, Poland. He started his undergraduate training at the Electrotechnical Institute of the University of Grenoble, France. He came to the United States in 1940, and served with the U. S. Army Intelligence Service overseas from 1943 until 1945. He received the B.E.E. degree with honors in

M. Nowogrodzki

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

1948 and the M.E.E. degree, in 1951, both from the Polytechnic Institute of Brooklyn.

From 1948 until 1951 Mr. Nowogrodzki was on the staff of Hazeltine Electronics Corporation, where he was engaged in the development of microwave components and devices for radar equipments. He joined Amperex Electronic Corporation as microwave engineer in 1951, and he is now engaged in the development of magnetron test equipment and measuring techniques in Amperex's Microwave Division.

Mr. Nowogrodzki is a member of Tau Beta Pi and Eta Kappa Nu.

❖

John R. Ragazzini (A'41–M'46–SM'52) was born in New York City on January 3, 1912. He received the B.S. and E.E. degrees at the College of the City of New York in 1932 and 1933, and the A.M. and Ph.D. degrees at Columbia University in 1939 and 1941.

J. R. RAGAZZINI

After associating with one of the New York City Departments, Dr. Ragazzini became an instructor in the School of Technology of the City College. In 1941 he joined the faculty of the School of Engineering at Columbia University, where he is now professor of electrical engineering. He was active in the Division of War Research at Columbia during World War II and later as an engineering consultant in the field of feedback control and noise problems. In addition to his academic duties, he is currently directing a project in air defense at the new Electronics Research Laboratories at Columbia.

Dr. Ragazzini is a member of Phi Beta Kappa, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu, and of the A.I.E.E. and A.S.E.E.

❖

Henry J. Riblet (A'45) was born on July 21, 1913, in Calgary, Canada. He received the B.S. degree in 1935 and the Ph.D. degree in 1939 from Yale University.

HENRY J. RIBLET

From 1939 to 1942, Dr. Riblet taught mathematics first as instructor at Adelphi College and later as assistant professor at Hofstra College. From 1942 to 1945 he was at the Radiation Laboratory, where he was in charge of one of the three design sections of the antenna group. From 1945 to 1949 he was in charge of the antenna and RF groups at the Submarine Signal Company. He is now employed by the Microwave Development Laboratories.

He is a member of the American Mathematical and Physical Societies.

❖

Ernest G. Rowe was born in 1909 in Plymouth, England. He received an honors degree in engineering in 1932 and an M.Sc. degree a year later from London University.

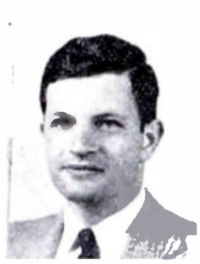
ERNEST G. ROWE

In 1933 Mr. Rowe joined the development staff of the M. O. Valve Company, and later became chief of the valve department. In 1948, he was named chief receiving valve engineer of the Brimar Valve Division of Standard Telephones and Cables Ltd.

Mr. Rowe is a member of the I.E.E. and an associate of the City and Guilds of London Institute, and holds the diploma of the Imperial College of Science and Technology.

❖

Eugene A. Sands (A'48) was born on March 15, 1925 in New York City. He graduated from Harvard College in 1944 with the B.S. degree in physics, and then served three years in the U. S. Navy as a communications officer. He returned to Harvard to earn the M.S. degree in applied physics in 1948.



EUGENE A. SANDS

Afterwards Mr. Sands designed hearing aids for the Sonotone Corporation of Elmsford, N. Y., and then joined the Burroughs Adding Machine Company in Philadelphia. He is now with the Magnetics Research Company of Chappaqua, N. Y., doing work on the design and application of magnetic components to computers and business machines.

❖

Thomas G. Slattery was born in Boston, Mass. in 1918. From 1941 to 1946 he served as a communications officer in the U. S. Air Forces. He received the B.S. degree in electrical engineering in 1948 and the M.S. degree in applied mathematics in 1949, both from the University of Michigan. From 1947 to 1949 he was a research assistant at the Engineering Research Institute, University of Michigan, working on operations research problems for the Office of Naval Research.

T. G. SLATTERY

In 1949 Mr. Slattery became a project engineer at Sperry Gyroscope Company, and was engaged in theoretical design studies for a guided missile project. The following year he joined the staff of the Avion Instrument Corporation of Paramus, N. J., and worked on the design and construction of aircraft fire-control computers.

From 1951 to 1952 Mr. Slattery was a senior engineer with Melpar, Inc., primarily engaged on a countermeasures project and acting as a consultant on radar and telemetering projects. He is presently employed as a project engineer by the Transducer Corporation of Boston, Mass.

❖

James R. Wait was born in 1924 in Ottawa, Canada. During World War II he was a radar technician in the Canadian Army. In 1949 he received the M.A. degree in applied physics, and in 1951 the Ph.D. degree in electrical engineering, both from the University of Toronto. He was employed by the Research Laboratories of the Hydro Electric Power Commission of Ontario in 1948 and by Newmont Exploration, Ltd. of Jerome, Ariz. in 1951 and 1952.

JAMES R. WAIT

Dr. Wait is now in charge of the theoretical section of the Radio Physics Laboratory of the Defence Research Board in Ottawa. He is a member of the Professional Engineers of Ontario, the Canadian Association of Physicists, and the Society of Exploration Geophysicists.

❖

Chester B. Watts, Jr., was born in Washington, D. C., on June 16, 1918. After receiving the B.S. degree in electrical engineering from the Massachusetts Institute of Technology in 1940, he went to work for the Federal Telegraph Company, Newark, N. J., on developing antennas and modulation equipment for instrument landing localizers and radio ranges.

C. B. WATTS, JR.

From 1942 to 1946, Mr. Watts was an engineering project officer in the Aircraft Radio Laboratory, Wright Field. He was awarded the Legion of Merit for work in connection with the development of automatic landing equipment.

Since 1947 Mr. Watts has been employed by the Civil Aeronautics Administration in the development of radio aids to air navigation, principally instrument landing.

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For a photograph and biography of DR. LOTFI ZADEH, see page 1128 of the September issue of the PROCEEDINGS OF THE I.R.E.

Institute News and Radio Notes

Calendar of COMING EVENTS

IRE-RTMA Radio Fall Meeting, Syracuse, N. Y., October 20–22

IRE Annual Broadcast Conference, Franklin Institute, Philadelphia, Pa., October 27

Symposium on Microwave Circuitry, Western Union Telegraph Company Auditorium, New York, N. Y., November 7

Kansas City IRE Technical Conference, President Hotel, Kansas City, Kan., November 21–22

1952 National Conference on Vehicular Communications, Statler Hotel, Washington, D. C., December 3–5

IRE-AIEE-ACM Computers Conference, Park Sheraton Hotel, New York, N. Y., December 10–12

IRE-AIEE Meeting on High Frequency Measurements, Washington, D. C., January 14–16

IAS-IRE-RTCA-ION Symposium on Electronics in Aviation, New York, N. Y., January 26–30

IRE Southwestern Conference and Electronics Show, Plaza Hotel, San Antonio, Tex., February 5–7

1953 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 23–26

9th Joint Conference of RTMA of United States and Canada, Ambassador Hotel, Los Angeles, Calif., April 16–17

1953 National Conference on Airborne Electronics, Dayton, Ohio, May 11–13

Kansas City Section Conference Slated

The fourth annual Regional Papers Technical Conference will be sponsored by the Kansas City IRE Section, November 21–22, at the President Hotel, Kansas City, Kan.

"Electronics in Industry" carries the general theme of the conference. The two-day sessions will cover recent advancements in the fields of industrial communications, audio and video circuitry, instrumentation, antennas and radiation, and radio and television broadcasting.

1952 Student Awards

The IRE Board of Directors has established a plan whereby students in the different IRE Student Branches may be given an award by the local Section. The Annual IRE Student Branch Awards for 1952, as well as the name of the Student Branch in which the student winner was enrolled, are listed as follows:

Student Branch Award Winner	*Student Branch*
M. A. Winkler	Univ. of Akron
R. W. Hellwarth	Princeton Univ.
A. O. Levy, Jr.	Rutgers Univ.
J. L. Preston	Yale Univ.
B. J. Shinn	Univ. of Connecticut
J. E. Dixon	Univ. of Dayton
W. D. Sneddon	Univ. of British Columbia
E. R. Schmidt	Univ. of Illinois
N. A. Bourgeois, Jr.	Tulane Univ.
A. B. Welch	Southern Methodist Univ.
J. S. Lodato	Louisiana State Univ.
J. L. Minck	Univ. of Notre Dame
Jack Schwartz	Illinois Inst. of Technology
E. R. Yoder	Northwestern Univ.
J. W. Long	Univ. of Kansas
J. D. Crane, Jr.	Utah State Agricultural College
William Rouse	Syracuse Univ.
E. W. Messinger	Univ. of Cornell
R. B. Naugle	Univ. of Arkansas
R. B. Kieburtz	Univ. of Washington
P. E. Kammerman	Univ. of Maryland
R. L. Van Allen	George Washington Univ.
Orrin Kaste	Univ. of Wisconsin
R. W. Sackett	Marquette Univ.
R. J. Talbot	Michigan College of Mining and Technology
M. L. Fox	Kansas State College
J. W. Christie	Univ. of Detroit
Joseph Ossanna, Jr.	Wayne Univ.
E. G. Gilbert	Univ. of Michigan
K. L. Johnson	Seattle Univ.
R. M. Wolfe	Univ. of Louisville
G. A. Breakey	Carnegie Inst. of Tech.
F. W. Keay	Univ. of Pittsburgh
John Tomlinson	Pennsylvania State College

Microwave Symposium Slated for November

A symposium on "Microwave Circuits," sponsored by the IRE Professional Group on Microwave Theory and Techniques, is scheduled for November 7, 1952, at the Western Union Telegraph Company Auditorium, New York, N. Y.

Registration fee for the symposium, received before October 24, will be $1.75. The fee will be advanced to $2.00 after that date. Registrations filed after October 12 will be held and can be obtained at the entrance to the auditorium on the date of the meeting. All persons desiring to register for the symposium should send their remittance to: W. M. Goodall, Bell Telephone Laboratories, Inc., Box 107, Red Bank, N. J.

Program

Morning Session: November 7

Andre Clavier, *Chairman*

Introduction and Welcome, Ben Warriner, Chairman of IRE Professional Group on Microwave Theory and Techniques

"A New Chart for the Solution of Transmission Line and Polarization Problems," A. Deschamps, Fed. Tele. Labs.

"Some Coupled-Wave Theory and Application to Waveguides," S. E. Miller, Bell Telephone Labs., Inc.

"Audio Modulation Substitution System for Microwave Attenuation Measurements," J. Korewick, Sperry Gyroscope Co.

"The Microwave Interferometer for Measuring the Time Displacement of a Projectile Within the Barrel of a Gun," H. C. Hanks, Jr., Glen L. Martin Co.

Afternoon Session: November 7

G. C. Southworth, *Chairman*

"Performance of Ferrites in the Microwave Range," A. G. Fox, Bell Telephone Labs., Inc.

"Microstrip—A New Microwave Transmission Technique," H. F. Engelmann, Fed. Tele. Labs.

"Some New X-Band Components for Radar Use," H. J. Riblet, T. S. Saad, and E. Hodge, Microwave Development Labs., Inc.

"Duplexing Filter Design at 2,000 Mc," D. R. Crosby, RCA Victor

"A Standard Waveguide Spark Gap," David Dettinger, Wheeler Labs.

LAST CALL!

AUTHORS FOR IRE NATIONAL CONVENTION!!

Lloyd T. DeVore, Chairman of the Technical Program Committee for the 1953 IRE National Convention, to be held March 23–26, requests that prospective authors submit the following information: (1) Name and address of author, (2) Title of paper, (3) A 100-word abstract and additional information up to 500 words (both in triplicate) to permit an accurate evaluation of the paper for inclusion in the Technical Program.

Please address all material to: Lloyd T. DeVore, c/o IRE Headquarters, 1 East 79 Street, New York 21, N. Y.

The deadline for acceptance is November 17, 1952. Your prompt submission will be appreciated.

Professional Group News

Audio

Recent activities of the Audio Group have included the distribution of PGA-9 Transactions to the members, and petitions for the formation of Group Chapters in Philadelphia and Albuquerque.

At the National Electronics Conference in Chicago, Ill., September 29–October 1, 1952, the Audio Group presented the following papers: "Analog for Loudspeaker Design," by J. J. Baruch and H. C. Lang; "High-Power Audio Amplifiers," by L. F. Deise and H. J. Morrison; and, "Direct Measurement of the Efficiency of Loudspeakers by Use of a Reverberation Room," by H. C. Hardy, H. H. Hall, and L. G. Ramer.

Vehicular Communications

The IRE Professional Group on Vehicular Communications will sponsor the third annual National Conference on Vehicular Communications, December 3–5, 1952, in Washington, D. C.

Inquiries for further information on the Conference may be addressed to the Chairman: W. A. Shipman, Box 215, 109 E. Broad St., Falls Church, Va.

1953 IRE National Convention

Each IRE Professional Group has appointed a representative to serve on the 1953 IRE National Convention Journal Committee.

Radio Telemetry and Remote Control

M. V. Kiebert, Chairman of the IRE Professional Group on Radio Telemetry and Remote Control, has volunteered to speak at IRE Section meetings throughout the country on, "Telemetering Installations and Systems." Mr. Kiebert travels across the nation about once a month and should be contacted directly at Bendix Aviation Corporation, Teterboro, N. J., in regard to speaking engagements.

Instrumentation

The 202-page Proceedings of the Symposium on Progress in Quality Electronic Components, held in Washington, D. C., May 5–7, 1952, containing 48 papers, fully illustrated, are available from the Radio-Television Manufacturers Association, 800 Wyatt Building, 777-14 St., N.W., Washington, D. C. The price is $5.00 per copy.

The Instrumentation Group has appointed R. L. Sink as sponsor representative for the West Coast Symposium on Quality Electronic Components which will be held in May, 1953.

In conjunction with the Cleveland IRE Section, the Group sponsored a technical session at the annual meeting of the Instrument Society of America, in Cleveland, September 8–12, 1952. P. L. Hoover, head of the department of engineering, Case Institute of Technology, was the Group representative for the session.

Information Theory

Nathan Marchand, Chairman of the Information Theory Group has announced that the Group's Administrative Committee is planning a National Conference for this fall. In addition, there will be activities on the West Coast. Time, place, and details will be announced shortly.

Broadcast Transmission Systems

The IRE Professional Group on Broadcast Transmissions Systems will sponsor the second Annual Broadcast Conference, October 27, 1952, at the Franklin Institute, Philadelphia, Pa.

Approximately nine papers of technical interest will be presented, and a luncheon and banquet will be included in the conference program.

Further information may be obtained from the Conference chairman: Lewis Winner, Bryan Davis Publishing Company, 52 Vanderbilt Ave., New York 17, N. Y.

Electronic Computers

The Electronic Computer Group has appointed the following committee for the Joint Computer Conference at the Hotel Park Sheraton, New York, N. Y., December 10–12, 1952: N. H. Taylor (Chairman), J. H. Howard (Secretary), S. N. Alexander (Program Chairman), M. M. Astrahan, P. Crawford, Jr., F. J. Maginnis, J. C. McPherson, A. R. Mohr, W. H. MacWilliams, C. V. L. Smith, V. G. Smith, S. B. Williams.

JTAC Report on Spectrum Conservation Available

A comprehensive survey and appraisal of the utilization of the radio spectrum is now available in a volume entitled "Radio Spectrum Conservation," prepared by a group of well known experts under the auspices of the Joint Technical Advisory Committee.

Subjects covered in the volume are as follows: (chapter 1) history of the allocation of the radio spectrum from 1896 to the present; (chapter 2) propagation characteristics of the radio spectrum from 10 kc to 300,000 mc; (chapter 3) an ideal approach to allocations covering all types of services; (chapter 4) critique of the present allocations from 10 kc to 300,000 mc; (chapter 5) dynamic conservation of spectrum resources covering past lessons and present problems, ideal and actual conditions of spectrum occupancy, technical measures to implement conservation, economic factors, and specific examples.

"Radio Spectrum Conservation" is available from the McGraw-Hill Book Co., 330 W. 42 St., New York 18, N. Y., at $5.00 per copy.

Technical Committee Notes

The **Standards** Committee convened on July 31, under the chairmanship of A. G. Jensen. Chairman Jensen brought to the attention of the Committee the question of appointments of technical committee representatives in connection with the submission of material for the Annual Review. R. R. Batcher, Chairman of the Annual Review Committee, stated that preliminary work with the Annual Review should be commenced immediately. Last year 20 to 25 per cent of the reports arrived after the deadline, causing a delay of about two months. Chairman Jensen advised the members, as chairmen of technical committees, to select a representative, or representatives, and to so advise Mr. Batcher, with a copy to L. G. Cumming, Technical Secretary. The Committee Chairman is also to ascertain that his appointee is willing to do the work and will carry it through to completion, thus eliminating instructions and relevant data to individuals not actually concerned with the work. The Annual Review Committee is presently preparing instructions, etc., which will be distributed as soon as the appointments are made. A. G. Clavier reported on the work being done in his Symbols Committee and the Subcommittee on Graphical Symbols for Semiconductors. Discussion ensued concerning the forthcoming ASA Standard on Graphical Symbols which will be available in final draft form in about three weeks. In order to avoid any conflict in symbols, it was suggested that this document be given a wide circulation. The Symbols Committee will consider this document at its next meeting, after which it should be submitted to the Standards Committee for approval. The Committee then turned its attention to the Definition of Noise Figure (Noise Factor), to be included in the document 51 IRE 17. PS3 and which was given final approval at the last Standards Committee meeting.

On July 11, the **Electron Devices** Committee met under the Chairmanship of G. D. O'Neill. Chairman M. E. Hines of the Ad Hoc Committee on Reorganization of the Committee on Electron Devices, stated that the first meeting of this group would be held early in the fall. During a general discussion of the problems facing this Committee, it was pointed out that the standards work on cathode-ray tubes, storage tubes, and pickup tubes is now the responsibility of a single subcommittee. The Chairman of this subcommittee, R. B. Janes, has begun work on Phototube Definitions and Methods of Test, in collaboration with R. G. Stoudenheimer. Although most of the material inherited from the work of past years requires nearly complete revision, it should be ready for the main Committee by late 1952. Pickup tube standards will be considered in 1953. Approval has been given to the name of the Subcommittee for Test Methods for High-Vacuum Microwave Tubes with the numerical designation 7.11. A list of Microwave Tube Definitions was discussed and a final revision was made. R. M. Ryder reported that the Subcommittee on Semiconductor Devices had organized two task groups, one on Definitions and Letter Symbols, under S. J. Angello, and one on Methods of Test under W. J. Pietenpol.

The **Electronic Computers** Committee

(Technical Committee Notes, cont.)

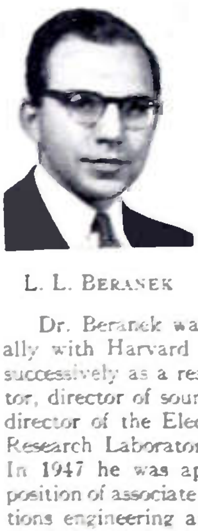
met on June 27, under the Chairmanship of Robert Serrell. Nathaniel Rochester presented a detailed report of the Eastern Definitions Subcommittee's present and projected activities. Liaison will be maintained with an AIEE subcommittee on computer definitions. It is planned to give particular attention to such terms in common use in patent specifications. Attention will also be given to the proper style of all their definitions. During the discussion it was pointed out that all such analog computer terms which do not deal specifically with servomechanisms, should be included within the province of the two Definitions Subcommittees. Specific servomechanisms terms (such as the word "servomechanism" itself) should be handled by the Servo-Systems Committee. The next item discussed concerned the formation of a new Subcommittee on Magnetic Recording. Chairman Serrell announced that he had appointed S. N. Alexander the Subcommittee Chairman. Members of this Subcommittee on Magnetic Recording for Computing Purposes will be appointed by its Chairman. The scope of the new Subcommittee has been tentatively described as follows: "To study the physical characteristics of apparatus and materials used in magnetic recording for computing purposes on continuous media (tapes, drums, etc.) that are significant in determining the performance of these devices. To establish proper technical methods of measuring: (1) relevant physical characteristics of equipment and materials; (2) performance." At the suggestion of Dr. Alexander, a discussion ensued on doing similar work for computer diodes. Chairman Serrell asked Dr. Alexander and Mr. Rochester to consider the situation and make recommendations later.

On June 13, the **Navigation Aids** Committee convened under the Chairmanship of P. C. Sandretto. A list of terms presented by L. M. Sherer was considered by the Committee. Also considered was a list of terms prepared by R. E. Gray, who extracted the definitions from the Radiation Laboratory Wartime Series, volume 19.

The **Sound Recording and Reproducing** Committee, under the Chairmanship of A. W. Friend, met on July 29. Revisions were made in the tentative Standards on Sound Recording and Reproducing: Methods of Measurement of Noise in Sound Recording and Reproducing Systems, 1951 (51 IRE 19. PS1), as the result of comments and suggestions received by the Committee. This material is now ready for Standards Committee action. A. P. G. Peterson, Chairman of Subcommittee 19.1, submitted, by mail, a list of his present personnel and reported that R. E. Zenner had submitted a revision of his material on "Proposed Standards on Frequency Response of Recording-Reproducing Systems with Resistive Terminations" to the Subcommittee. Comments have been made which will delay sending this proposal to the main Committee. The proposal on "Distortion: Definitions and Procedures for Measurement" is still on the agenda; however, no new proposal is yet ready for submission. Lincoln Thompson, Chairman of Subcommittee 19.2, reported that his Subcommittee on Mechanical Recording is still preparing a draft on Record Calibration. It is hoped that this material will be ready for submission to the Committee during 1952. H. E. Roys explained in detail (with the help of Mr. McNaughten, Chairman of the CCIR Committtee on Recording and Reproducing) the general program by which CCIR is attempting to augment the international exchange of radio program material through the medium of mechanical disc and magnetic tape transcriptions. The chief problem is to complete the necessary field of agreement for recording and reproducing standards among the various CCIR nations. It was indicated that, through Mr. McNaughten's efforts, a large part of the United States magnetic tape recording standard was accepted by the international group. The remaining objective is international standardization in the mechanical recording field, and in the field of methods and procedures for the measurement of the characteristics of recorded information of both mechanical and magnetic transcriptions. Standardization also is necessary on the recording amplitude as a fluctuation of frequency. Mr. McNaughten talked widely on the various practical problems of his ventures into these fields of international standardization. His remarks contributed much to the committee members' knowledge of CCIR problems, and it was agreed to set up under the Sound Recording and Reproducing Committee a CCIR Task Group. Mr. Roys will serve as Chairman. The other members will be A. W. Friend, L. Thompson, R. M. Fraser, W. S. Bachman, and R. C. Moyer. Mr. Roys announced that he would proceed, with the assistance of the various Task Group members, to prepare as much material as possible relating to the present state of the mechanical and magnetic recording arts in the United States. This material will be submitted to Mr. McNaughten for his use in the forthcoming CCIR meeting.

IRE People

Leo L. Beranek (S'36–A'41–SM'45–F'52) has been made a Fellow in the American Academy of Arts and Sciences.



L. L. Beranek

Dr. Beranek was born in Solon, Iowa, on September 15, 1914. He received the B.A. degree in 1936 from Cornell College, and the M.S. and D.Sc. degrees from Harvard University in 1937 and 1940, respectively, and an honorary D.Sc. degree from Cornell College in 1946.

Dr. Beranek was associated professionally with Harvard University, and served successively as a research assistant instructor, director of sound-control research, and director of the Electracoustic and Systems Research Laboratories from 1937 to 1946. In 1947 he was appointed to his present position of associate professor of communications engineering and technical director of the Acoustics Laboratory at the Massachusetts Institute of Technology.

Dr. Beranek is a fellow of the American Physical Society, a fellow and associate editor of the Acoustical Society of America, and a member of the American Association for the Advancement of Science, Sigma Xi, and Eta Kappa Nu. He served as the National Chairman of the IRE Professional Group on Audio, as chairman of the Acoustics Section Z-24 of the American Standards Association, and is the chairman of the Acoustics Panel, Research and Development Board, Department of Defense.

❖

Edward Stanko (A'27) has been appointed manager of engineering, technical products division, RCA Service Company, Inc. He will direct specialized training of field personnel, preparation of technical information, and development of new and improved methods for installation and servicing of RCA technical products.

Mr. Stanko was born in Hastings, Pa., and attended the National Radio Institute, the University of Buffalo, and the College of South Jersey.

Beginning his career as a ship's radio operator in 1920, for fifteen years he held various positions as announcer, transmission engineer, and chief engineer with broadcasting companies in Buffalo, N. Y. He served as both commentator and engineer on a night-time musical program over Station WEBR, Buffalo, one of the first radio stations in the United States to broadcast continuously for 24 hours each day. In 1928, Mr. Stanko brought television to Buffalo for the first time, using a crude "scanning disc" set to receive a picture of a girl transmitted from an experimental station at Schenectady, 200 miles away.

Joining RCA as a sound engineer in 1937, Mr. Stanko later worked as a senior engineer on specialized projects for the government. In 1943 he was appointed manager of the RCA Company's technical products and district operations groups until his recent promotion.

Mr. Stanko is a member of the Society of Motion Picture and Television Engineers, and has a commercial radio operator's license. He is an active radio "ham" operating station W2RHT.

❖

A. B. Hunt (SM'43) has been elected president and chairman of the board of the Radio-Television Manufacturers Association of Canada, for the 1952–1953 year term.

Mr. Hunt, a native of London, Ontario, received his B.S. degree from the University of Toronto in 1928, and joined the Northern Electric Company Ltd., as a manufacturing methods engineer in connection with theatre sound systems and vacuum tube production. In 1933, he was appointed special products

(Continued on page 1258)

IRE People

manufacturing superintendent and, in 1935, took over the added duties of radio receiver engineering. He was later made a special products superintendent and then electronics manager of a separate division of Northern Electric. In 1950, Mr. Hunt was made assistant manager of Northern's telephone division in Montreal and later was appointed to his present position of manager of the communications division.

Mr. Hunt was winner, in 1946, of the R. A. Rose Medal for his paper, "The Future of Radio Communications in Canada." For many years a director of the RTMA of Canada, his affiliations include memberships in the Corporation of Professional Engineers of Quebec, Engineering Institute of Canada, Canadian Manufacturers Association, and the Canadian Industrial Preparedness Association. He also holds the rank of Lieutenant Colonel as former commanding officer of a Signals Regiment.

❖

Alvin J. Zink, Jr. (M'44–SM'51) has been appointed general manager of the Tru Connector Corporation of Lynn, Mass., manufacturers and designers of microwave fittings.

Mr. Zink was born in Methuen, Mass., in 1914, and received his B.S. degree from the University of North Carolina in 1936.

Beginning his career with the National Company, Inc., Mr. Zink formed a private consulting firm in the field of industrial, marine, and municipal communication systems. In 1943 he became a staff member of the Radiation Laboratory at the Massachusetts Institute of Technology as a design and production engineer for microwave components. Until his recent appointment, Mr. Zink was associated with the research division of the United Shoe Machinery Corporation, Beverly, Mass., designing and developing industrial electronic equipment.

❖

Claud E. Cleeton, (A'41–SM'46), head of the Security Systems Branch of Radio I Division, has been appointed superintendent of Radio I Division, National Research Laboratory.

C. E. Cleeton

A native of Missouri, Dr. Cleeton received his B.S. degree in 1928 from the Northeast Missouri State Teachers College, the M.S. degree from the University of Missouri in 1930, and the Ph.D. degree in physics from the University of Michigan in 1935. He was an instructor in physics at the Teachers College from 1930–1931, Moberly Junior College, Missouri, in 1931–1935, and the University of Michigan from 1935–1936.

Joining NRL as a physicist, Dr. Cleeton did pioneer research work in microwave communications and electronic switching circuits from 1936–1940. During 1941–1942, he directed the research and development of electronic identification systems, radio control, guided missile electronics and countermeasures, and speech privacy systems. In 1942 he was appointed head of a combined research group from the United States, United Kingdom, and Canada to develop a new and uniform radar identification and recognition system for use by the allied armies, navies, and air forces. For this activity, he was awarded the Meritorious Civilian Service Award in 1947.

Dr. Cleeton is a member of the American Physical Society.

❖

Axel G. Jensen (A'23–M'26–F'42) has been awarded the G. A. Hagemann Gold Medal, by the Royal Technical University of Denmark, in recognition of his contributions to fundamental television research.

A. G. Jensen

Mr. Jensen graduated from the university in 1920 and acted as an instructor and scientific assistant there for two years. He then came to the United States as a fellow of the American Scandinavian Foundation and soon entered the services of the Bell Telephone Laboratories, Inc. He recently became director of television research there.

❖

Robert J. Bibbero (M'49–SM'50) has been appointed head, servomechanisms section, of the newly formed guided missiles division of Republic Aviation Corporation, New York, N. Y. He was formerly servomechanisms group leader with the Bell Aircraft Corporation, Buffalo, N. Y.

R. J. Bibbero

Mr. Bibbero was born in San Francisco, Calif., and received the B.S. degree in chemistry from the University of California in 1938. During 1938–1939, he did chemical engineering graduate work at the University of Michigan on a fellowship.

Until 1942, Mr. Bibbero was a research engineer of Moore Business Forms, Inc., when he was called to active duty as ordnance officer with the United States Navy. After training at Bowdoin College and Massachusetts Institute of Technology Radar School, he served as electronics officer in several naval shipyards and as ordnance officer in the Pacific Fleet. Released from the Navy in 1946 as a lieutenant commander, Mr. Bibbero joined the research staff of Linde Air Products Company.

In 1948 Mr. Bibbero attended the University of Buffalo where he received the M.S. degree in electrical engineering and worked on his Ph.D. degree in chemistry.

Mr. Bibbero is a senior member of the American Chemical Society, a member of the American Association for the Advancement of Sciences, and a licensed professional engineer in New York and California. He was recently elected Secretary-Treasurer of the Buffalo-Niagara IRE Section.

❖

Wilbur S. Hinman (SM'46) has been appointed associate director for ordnance development of the National Bureau of Standards, Washington, D. C. He will co-ordinate NBS work in the field of ordnance research and development.

W. S. Hinman

Mr. Hinman was born in Washington, D. C., and attended the Virginia Military Institute, receiving the B.S. degree in electrical engineering in 1926. From 1926–1928, he was a student and radio engineer for Westinghouse Electric Corporation. In 1928, Mr. Hinman joined the staff of NBS as a member of the radio section.

Mr. Hinman has achieved outstanding recognition as an authority on the radio proximity fuze and radio meteorography. He is co-inventor with the late Harry Diamond of a basic design utilized in the radio proximity fuze, the radio sonde, and the automatic weather station. He has also conducted extensive research on upper-air wind-measuring equipment, direction finders, automatic volume control for aircraft radio receivers, radio compasses, and blind landing aids.

In recognition of his contributions during World War II, Mr. Hinman received the Presidential Certificate of Merit, the Ordnance Development Award, the Army's Citation for Exceptional Service, and Department of Commerce Medal for Meritorious Service.

Mr. Hinman has served for some years as chairman of the Warhead and Fuzes Panel of the Guided Missile Committee of the Research and Development Board.

Raymond E. Drake (SM'45), staff engineer of the electronic subdivision Air Materiel Command, Wright Field, died recently.

Mr. Drake was born in Indiana on June 10, 1907, and received the B.S.E.E. degree from Purdue University in 1929.

From 1929–1930, Mr. Drake was a student engineer at Westinghouse, East Pittsburgh, Pa., and from 1930–1935, he served as a radio engineer at RCA Victor, Camden, N. J. He then became associated with the Aircraft Radio Laboratories, Wright Field, Dayton, Ohio, where he worked on communication equipment for the Army Air Force, subsequently transferring to the position he held at the time of his death.

Books

Introduction to Electronic Circuits by R. Feinberg (4359)

Published (1952) by Longmans, Green and Company, Inc., 55 Fifth Avenue, New York, N. Y. 161 pages+2-page index+xiv pages. 121 figures. 5¼×8½. $3.50.

R. Feinberg is a research engineer and former lecturer at the University of Manchester, England.

According to the preface, this book is based on a lecture and laboratory course given to students in the Honours Schools of electrical engineering and physics at the University of Manchester. It is intended as an introductory course in electronics for university undergraduates and may be helpful to research workers as a review.

The chapter headings offer a reasonable outline of the contents; these are: Thermionic vacuum valves I; Thermionic vacuum valves II; Fundamentals of alternating-current amplification; Nonlinearity effects of vacuum valve characteristics; Sinusoidal oscillators; Relaxation oscillators; Thermionic gas-filled valves; Cold-cathode valves, photo-valves, and mercury-pool valves.

The general application emphasizes the fundamental physical principles that are pertinent to the situation. Each chapter is provided with a set of numerical problems, usually comparatively simple, with answers. Twenty-four laboratory experiments, well chosen for an introductory treatment, are interspersed throughout the text. The experiments, described in reasonable detail, employ English valve types. It is unfortunate that it is almost as hard to replace a valve by a tube in a laboratory experiment as in a radio or radar chassis, a circumstance which puts a significant limitation on the book's general utility in this country.

The mathematical treatment, though fairly full and carefully done, is vaguely unsatisfying even for an introductory text in that it does not seem to go quite far enough. The expressions for gain as a function of frequency are given but have not been exploited in detail. There is no treatment of the relation between gain and bandwidth that is so basic in wide-band circuits. However, there are useful points about some of the less common circuits.

There are a few errors that are noticeable on casual reading. On page 47, the condition for a stationary Lissajous figure is incorrectly stated; on page 67, it is apparently stated that the gain of a cathode follower is always much less than unity. On page 150, in a discussion of cold-cathode gas-filled tubes it is implied that electron emission from the cathode is due to the electric field there, rather than the result of positive ion bombardment and photoelectric effect.

The references are valuable as a further search of the literature, including many of the older articles where various points are first described. On the whole, the book is likely to be moderately useful as a companion to other treatments, but it is not likely to replace other sources.

S. N. Van Voorhis
M.I.T.
Cambridge, Mass.

Receiving Problems in the Ultra-High-Frequency Range (Empfangsprobleme im Ultrahochfrequenzgebiet) by Herbert F. Matare (4360)

Published (1951) by R. Oldenburg Publisher, Munich, Germany. 261 pages+3-page index. 190 figures. 6¾×9½.

Herbert F. Matare is a physicist at Westinghouse, Paris, France.

This book presents primarily a mathematical treatment of the sensitivity of uhf and microwave receivers and contains valuable design information for input stages of such receivers. The presentation is concentrated on the basic elements of electron tubes and crystal detectors, their performance and noise characteristics for various operating conditions.

The first half of the book considers input stages with conventional electron tubes; a very complete coverage of the problem of mixing with diodes and triodes is included in this section.

The second half is concerned with crystal detector circuits. The topics include: the physics of the crystal detector and its characteristics, impedance measurements, a discussion of the equivalent circuit, frequency conversion, rectification, a comparison between the performance of tube and crystal detector input stages, instructions for the choice of crystals with the most appropriate characteristics for various applications, and examples for the design of mixer input circuits. Also included is a short chapter on the measurement technique of receiver sensitivity.

The theoretical results are reduced to simple formulas which are supplemented by a large number of graphs. Examples of measurements are quoted to justify the validity of the assumptions in the theory.

The book is largely based on material which was compiled in Germany during the war and which, in part, has not been published before; new literature is considered.

The author, an expert in receiver and detector problems, has made valuable contributions to the literature on these subjects. This fact, or limited space, perhaps, may be the reason for the omission of background information helpful to understanding of the contents by a reader not skilled in this field. However, one specializing in the design of high quality receivers will find much useful information in this book.

Georg Goubau
Signal Corps Engineering Labs.
Fort Monmouth, N. J.

Principles of Radio, Sixth Edition by Keith Henney and Glen A. Richardson (4361)

Published (1952) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 641 pages+13-page index+vii pages. 404 figures. 5¼×8½. $5.50.

Keith Henney is consulting editor of *Electronics*, New York, N. Y. Glen A. Richardson is assistant professor of electrical engineering, Iowa State College, Ames, Iowa.

This sixth edition of Henney and Richardson's "Principles of Radio" has been issued ten months after the ninth printing of the fifth edition; an indication of the acceptance and popularity of this book which first appeared in 1929.

The book is aimed at "those who must learn radio without the help of a teacher." It uses little algebra and, for the most part, is a verbal and extremely elementary description of circuits and systems. A great deal of the book is devoted to the explanation of common terms.

The new edition represents a thorough revision and an appreciable expansion of the preceding edition, but in substance and character the book remains much the same. Rather than acknowledge the many details of the revision, we will proceed to take a critical look at this text as a mature unit.

Basically and with due consideration for the level of the book, it can be asked whether the material is sufficiently comprehensive and well correlated, whether the book shows those elements of editing which might be more important here than at other levels, and similar questions.

Rather surprisingly, there seems to be many points for criticism. The authors' insistence on using "flow" with current whenever possible makes the distinction between current and charge almost disappear in the introductory discussion. Polarity and polarity marks enter almost surreptitiously, are used liberally, but are not included in the index. Potential is in somewhat the same situation. Phase, phase angle, and phase difference appear in confusing, if not contradictory, relation to one another. Terminology and notation are not completely consistent throughout the book, and the discussion of radiation, of which the authors endeavor to ward off adverse criticism by a footnote, should still be noted; it should not be present at all in the displayed form. Occasional references to engineers and engineering, for example, "Engineering the Voltage Divider," tend to give the reader an impression of a level considerably removed from the one before him.

The reviewer concludes that age has not brought complete maturity to this popular book.

J. G. Brainerd
Moore School of Elec. Eng.
Philadelphia, Pa.

Radio and Television Receiver Troubleshooting and Repair by Alfred A Ghirardi and J. Richard Johnson (4362)

Published (1952) by Rinehart Books, Inc., 232 Madison Ave., New York 16, N.Y. 795 pages+26-page index+xxiv pages. 416 figures. 6×9. $6.75.

Alfred A. Ghirardi is a radio and electronic engineering consultant and technical writer and editor, Darien, Conn. J. Richard Johnson is the technical editor of Rinehart Books, Inc., New York, N.Y.

Based on the success of Ghirardi's original "Modern Radio Servicing" textbook, the present book was written to provide an enlarged and more comprehensive up-to-date treatise on troubleshooting and repairing radio and TV equipment. The new book assumes the reader's familiarity with "Radio and Television Circuitry and Operation," by the same authors.

Books

The book is written on the nonmathematical technican level for school courses and practicing service technicians, although the authors recognized that service work is beyond the scope of a single practical volume. About 110 pages are devoted exclusively to television; however, much other information on AM and FM receivers is applicable to television.

Well written, the book generally follows in the logical order. Each chapter is followed by a summary and a set of review questions. A conscientious study of the book by the technician should bring about increased technical understanding and ability.

Considering the class of individuals to which the book is directed and the rapid advancement of the art, the book is timely.

ALOIS W. GRAF
135 South LaSalle St.
Chicago, Ill.

The Recording and Reproduction of Sound by Oliver Read (4363)

Published (1952) by Howard W. Sams & Company, Inc., 2201 East 46 st., Indianapolis 5, Ind. 708 pages+70-page appendix+10-page index+xv pages. 690 figures. 6×9. $7.95.

Oliver Read is the editor of *Radio and Television News* and *Radio-Electronic Engineering*, Chicago, Ill.

Almost every communications engineer of professional competence needs to defend himself from time to time against naive inquiries that begin, "How do I go about designing an equalizer (or filter, or loudspeaker enclosure, or preamplifier, etc.)?" Here, at last is an encyclopedic audio manual to which such inquirers can be referred. This shunt-path solution of the problem is not wholly without negative feedback hazards, however, for while there is a wealth of practical and useful information packed into the 800 pages of this fat volume, the portions of it devoted to the scientific foundations of the audio art are lamentably weak.

On the black-ink side of the ledger is the completeness of coverage afforded. Division of the subject matter into 29 chapters and an appendix provides for the separate treatment of almost every category of activity or equipment that concerns the audio practitioner. Constructional details, specifications, and complete wiring diagrams are given for a substantial fraction of the commercial equipment in current use, and while this feature may date the handbook within a few years, it also constitutes one of its valuable features. The NARTB Recording and Reproducing Standards are presented *in toto* and the appendix collects for the laboratory technician a further glossary of terms and miscellaneous reference data ranging from Ohm's law through color codes, tube date, and computation nomograms.

On the other hand, the author's attempt to translate acoustical science into semitechnical language falls far short of achieving success (and in some cases constitutes a positive menace to understanding). One of the features that distinguishes science from pseudoscience is the accurate and careful usage of technical terminology. The practical man to whom this manual is addressed might have been able to make substantial gains in understanding with the help of a good glossary if the author had taken more pains to avoid wrong, ambiguous, and nonspecific use of technical terms. Unfortunately, there is much loose scientific talk in this manual. For example, no glossary will straighten out, "all sound waves are composed of frequency, intensity, periodicity, and wave form" (page 2). Neither are the quantitative concepts of acoustics clarified very much by "the decibel is a ratio . . . not a quantity" (page 2), nor by "Intensity is the amplitude or power of vibration" (page 2). It is also a little startling to read that "a sound having 10 watts of power is 10 times as loud to the ear as a sound of 1 watt . . . a sound of 100 watts only 20 times as loud" (page 124). How much simpler life would be for psycho-acousticians if subjective loudness could be computed in this way.

Almost no branch of the audio art is more beset with old-wives' tales than is the recording and reproduction of disk records. The various chapters devoted to different aspects of this art assemble an impressive body of practical information, but the author has not succeeded in freeing the subject from the technical nonsense that clings tenaciously to it. As a result, these sections do not fill the long felt need for a truly discriminating survey of this field. A good many examples of this loose handling of the facts could be exhibited, but a few will suffice. Translations loss and pinch effect are treated as synonymous in one passage (p. 90), while elsewhere a "knee-action" playback needle is suggested as a palliative for the pinch effect; also, among the first requirements specified for a good phonograph stylus is that "it must be kind to the ear . . . " (page 170). This reviewer also takes an especially dim view of the author's perpetuation of the dimensional inconsistency involved in repeated reference to pickup "stylus pressure" in weight units. The inconsistency is particularly glaring when in successive sentences there is reference to a stylus contact pressure of 25,000 pounds per square inch and to an assumption that the point is operating at a pressure of 1.5 ounces (page 168). There is urgent need for reform of this terminology since the 1-mil stylus of a modern light-weight microgroove pickup actually operates at higher bearing pressures than prevail for 1-ounce pickups using a 2.85-mil stylus.

Engineers whose professional occupation is not primarily concerned with audio will find this manual a useful reference book, since they can make good use of the collected information about available equipment and will have enough technical sophistication to take or leave the science. Engineers who are primarily concerned with audio will welcome this collection of data into a single volume and will enjoy comparing their professional prejudices with the author's. The lay technician will only welcome the practical guidance and reference data set forth here, but he is likely to be seduced into believing that this is the real scientific lowdown on his art.

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Proceedings of the National Electronics Conference, Volume VII (4364)

Published (1952) by the National Electronics Conference, Inc., 852 E. 83 St., Chicago 19, Ill. 607 pages +14-page index+xii pages. 477 figures. 6×9. $5.00.

This bound volume of some 600 pages presents the Proceedings of the National Electronics Conference, held in October, 1951, in Chicago, Ill. Included are over 70 papers, a few of which are reported in abstract form. The general coverage of the conference is similar to that of previous years, and as customary, the term "electronics" is interpreted in the broad sense. The Proceedings include papers on servomechanisms, information theory, signal detection, medical applications, as well as electronic tubes and circuits.

One general comment is that the papers are not of uniform quality. A good number of papers are well written, thoroughly documented, and represent a significant advance in, or report of, their respective fields. On the other hand, there are a few papers whose quality is so distinctly inferior, it is doubted that they would have been accepted for publication if subjected to a critical review. However, in spite of this criticism, the reader will find a wealth of reference material in this volume and will find it well worth his while to obtain or have access to a copy.

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Most Often Needed 1952 Television Services Information Compiled by M. N. Beitman (4365)

Published (1952) by Supreme Publications, 3727 W. 13 St., Chicago 23, Ill. 190 pages of diagrams, models, and figures. 2-page index. 8½×10½. $3.00.

In this new "1952" manual, are included the circuit diagrams and the essential service facts on every popular television receiver currently produced by some thirty manufacturers. From the favorable acceptance of his previous volumes, the compiler has had good reason to believe that his selection and editing of the factory-supplied service and adjustment procedures fills a definite and timely need for this information on new receivers. The material presented is selected carefully and edited concisely so as to provide, in a compact and handy format, most of the information needed to identify a particular make and model, and to service and align its circuits properly.

Essential portions of the manufacturers service notes and alignment operations are cleanly and clearly reproduced by the planograph process; forty-one complete circuit diagrams, supplemented by the compiler's notes, are extended to cover some six hundred types or chassis numbers of the current years' models. The editor of this half-inch manual has performed an excellent job of tying together these "most often needed" service notes in order that they may be of greater assistance to the television serviceman.

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10.2 Dielectric Measurements

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10.4 Methods of Measurement

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15.2 Radio Telegraph Transmitters up to 50 Mc

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15.3 Double-Sideband AM Transmitters

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15.4 Pulse-Modulated Transmitters

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15.6 Television Broadcast Transmitters

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17.8 Task Group to Review I.E.C. Proposal on Standardization of Measurements of Receivers for AM Broadcast Transmissions

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25.5 Video-Frequency Measurements

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Abstracts and References

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The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February 1952, through January 1952, may be obtained for 2s8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London, S.E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

COUSTICS AND AUDIO FREQUENCIES

6:534 2406

References to Contemporary Papers on coustics—R. T. Beyer. (*Jour. Acous. Soc. mer.*, vol. 24, pp. 234–243; March, 1952.) ontinuation of 1797 of August.

4.2 2407

Transient Phenomena in Sound Transmison—A. Darré. (*Frequenz*, vol. 6, pp. 65–71; arch, 1952.) Discussion of phenomena conrned in the production of linear distortion ily, with particular reference to frequency sponse, phase relations and group transit me. The characteristics of moving-coil and rn-type loudspeakers are considered and also e effect of room reverberation on the response rve of a loudspeaker.

14.213.4–13 2408

The Propagation of Sound through Gases ntained in Narrow Tubes—L. E. Lawley. *Proc. Phys. Soc.*, vol. 65, pp. 181–188; March 1952.) Results for air, O, H and N at frequenes between 60 and 150 kc indicate a vissity/thermal-conductivity constant 5 per nt above the theoretical value.

14.23:534.321.9 2409

Transmission of Ultrasonic Waves through Thin Solid Plate at the Critical Angle for the ilatational Wave—K. R. Makinson. (*Jour. cous. Soc. Amer.*, vol. 24, pp. 202–206; March, 952.) The transmission through an isotropic ate immersed in a liquid is examined experientally and theoretically. Total reflection curs near the critical angle for a considerable nge of thickness of plate. It is therefore ossible to measure the velocity of the dilataonal wave in a solid even when only thin ecimens are available.

34.23–16 2410

Spherical-Wave Propagation in Solid Media -F. G. Blake, Jr. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 211–215; March, 1952.) An analysis shows that when an impulsive pressure is generated in a spherical cavity in an infinite solid medium, a damped oscillatory wave train is radiated which differs essentially from the form of the original pressure pulse.

534.23–16 2411

Transmission of Sound through Plates—A. Schoch. (*Acustica*, vol. 2, no. 1, pp. 1–17; 1952. In German.) Theory for plane waves and limited beams is developed in a form clearly showing the connection with free waves in plates. Cremer's concept of total transmission as "coincidence" of the incident wave with a free wave in the plate is critically discussed. Experiments with Al plates and ultrasonic waves give results in good agreement with theory.

534.231 2412

A New Expansion for the Velocity Potential of a Piston Source—A. H. Carter and A. O. Williams, Jr. (*Jour. Acous. Soc. Amer.*, vol. 24, p. 230; March, 1952.) Correction to paper abstracted in 1815 of 1951.

534.232 2413

Radiation Loading of a Piston Source in a Finite Circular Baffle—R. B. Watson. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 225–228; March, 1952.) Experimental results are given for baffle dimensions of the order of a wavelength. Examination of the results shows the lack of suitable expressions for calculations.

534.24:534.321.9 2414

Lateral Displacement of a Totally Reflected Ray at Ultrasonic Frequencies—A. Schoch. (*Acustica*, vol. 2, pp. 18–19; 1952. In German.) Schlieren photographs of 5.5-mc and 16-mc waves reflected from an Al plate in xylol clearly show this displacement, which occurs at the angle of incidence for which a Rayleigh wave is excited in the solid.

534.26 2415

The Diffraction of a Plane Sound Pulse Incident Normally on a Regular Grating of Perfectly Reflecting Strips—E. N. Fox. (*Proc. Roy. Soc. A*, vol. 211, pp. 398–417; March 6, 1952.) The general methods previously described (2417 of 1949) are used to find the pressure on both sides of gratings whose aperture areas are $\frac{1}{2}$, $\frac{1}{3}$, $\frac{1}{4}$ of the total grating area. Both the exact solution and an asymptotic solution for use in the later stages, when exact calculation becomes too laborious, are discussed. The results of both solutions are shown graphically. The analysis can be extended simply to rectangular pulses of finite duration.

534.321.9:534.232].047 2416

The Biological Effect of High-Level Complex Noise (Ultrasonic Region of the Spectrum)—P. Bugard. (*Ann. Télécommun.*, vol. 7, pp. 139–143; March, 1952.) Pen and cro recordings of the output of a Hartmann whistle for different adjustments of the compressed-air jet are discussed and the corresponding auditory sensations are noted. Two distinct types of output, depending on the jet adjustment, are identified: (*a*) a fairly pure high-level ultrasonic wave; (*b*) white noise of higher mean level.

534.41 2417

A Detector of Transients and its Applications to the Study of Music and Speech Signals—A. Moles and G. Corsain. (*Radio franç.*, no. 3, pp. 1–7; March, 1952.) Discussion of equipment designed to isolate the transients from the envelope of the energy spectrum and to effect their summation. Preliminary results obtained on signals derived from speech in different languages, orchestral, piano and violoncello music, and logatoms formed by coupling selected consonants and vowels, are shown and discussed.

534.7 2418

Recovery of the Auditory Threshold after Strong Acoustic Stimulation—I. J. Hirsh and W. D. Ward. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 131–141; March, 1952.) The elevation of the threshold after fatigue by pure tones and white noise was measured. Recovery to about the normal value usually occurs during the first minute, followed by an increase reaching a maximum about two minutes after the cessation of the fatiguing tone. In some cases a minor maximum occurs about five minutes later.

534.75 2419

Auditory-Psychological [hörpsychologische] Acoustics—P. Burkowitz. (*Funk u. Ton*, vol. 6, pp. 136–140; March, 1952.) Summary of the essentials of a new theory of the basic principles of single-channel transmission and reception of sound.

534.845 2420

Comparative Reverberation-Room Measurements of the Absorption Coefficient of Sound-Absorbent Materials—G. Venzke. (*Tech. Hausmitt. Nordw Dtsch. Rdfunks*, vol. 4, pp. 1–3; January/February, 1952.) The results of measurements in eight laboratories of the absorption coefficient of hallonit, a rock-wool material made in slabs 40×40×3 cm, are shown graphically. The size of the test surface used had no appreciable effect on the results, and measurement accuracy was about the same for warble-tone and white-noise sources. No dependence of the results on the shape and size of the test chamber could be found. Two sets of measurements by a Kundt's-tube method (perpendicular incidence) gave values of the maximum absorption coefficient about 15–20 per cent less than the mean value given by the reverberation measurements, in which surface areas of 7.5 m^2 or 15 m^2 were used, in general distributed on two adjacent walls and the floor of the test chamber.

534.846 2421

Acoustics of the Remodeled House and Senate Chambers of the National Capitol—P. E. Sabine. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 121–124; March, 1952.) Details are given of the materials and dispositions of sound-absorbing surfaces used, together with the results of articulation tests showing the excellent performance obtained.

534.85 2422

Thorn Needles—S. Kelley: A. M. Pollock. (*Wireless World*, vol. 58, pp. 243–244; June, 1952.) Further discussion on 2089 of 1951 (Pollock) and author's reply.

534.851 2423

Phonograph Needle-Drag Distortion—J. Rabinow and E. Codier. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 216–225; March, 1952.) Analysis indicates that tangential motion of the tip of a pickup needle will occur and may cause distortion of the output signal. Attempts to detect such distortion were unsuccessful owing to the presence of other distortions which masked it.

534.861.2 2424

Helmholtz Resonators in the Acoustic Treatment of Broadcasting Studios—C. L. S. Gilford. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 86–92; March, 1952.) "A theory of the action of Helmholtz resonators as sound absorbers is presented, covering both the isolated resonator and regular arrays. Experiments in reverberation rooms and acoustically treated studios are described and general recommendations for design are given. Regular arrays are preferable to single resonators, openings being made more resistive by covering with a fabric. It is concluded that great variations in design to suit architectural requirements may be made without loss of effectiveness, and the widths of the frequency band over which absorption takes place may be varied between wide limits."

621.395.623.8 2425

Centralized Public-Address System—(*Telefunken Ztg*, vol. 25, pp. 68–70; March, 1952.) Recent investigations indicate that for large audiences better results are obtained by using one or two vertical arrays of loudspeakers at a central point than by means of many loudspeakers distributed over the area to be covered. For vertical arrays the sound pressure increases with distance up to a certain point. With two vertical arrays (each with 48 loudspeakers) of over-all length 7 m and raised 6 m above the ground, a 300-w amplifier sufficed to give good sound distribution to a crowd of 300,000 people assembled in the holy place Fatima, Portugal.

621.395.625.2 2426

New Sound Reproducer for Engraved-Tape Records—P. Hémardinquer. (*TSF et TV*, vol. 28, pp. 113–114; March, 1952.) Description of commercially available equipment using a piezoelectric head with sapphire needle for reproduction from records on wax-coated tape.

621.395.625.3 2427

An Investigation into the Mechanism of Magnetic-Tape Recording—P. E. Axon. (*Proc. IEE*, Part III, vol. 99, pp. 109–124; May, 1952. Discussion, pp. 124–126.) Asymmetry of hysteresis loops is found to give distinctive properties to recording and distortion characteristics of unbiased recording. The properties have been confirmed experimentally. The mechanism of recording using high-frequency bias is examined. Predictions are made concerning transitions to be expected in the characteristics as the high-frequency bias field is increased from zero to saturation value; these are experimentally confirmed. Adequate high-frequency bias eliminates the asymmetry of the af intensity variation. The effect of a bias-leakage field outside the recording gap is discussed with reference to the coercivity of the tape material.

621.395.625.3 2428

A New Recording Medium for Transcribed Message Services—J. Z. Menard. (*Bell Sys. Tech. Jour.*, vol. 31, pp. 530–540; May, 1952.) A magnetic recording medium composed of rubber impregnated with magnetic oxide and lubricant, and used in the form of moulded bands, is found particularly suitable for applications involving repetition of short messages.

ANTENNAS AND TRANSMISSION LINES

621.315.212 2429

Elements with Rotational Symmetry for Coaxial-Cable Junctions—H. Meinke and A. Scheuber. (*Fernmeldetech. Z.*, vol. 5, pp. 109–114; March, 1952.) General explanation of the properties of units with sudden changes of conductor diameters or uniform variation of diameter and change of dielectric. The design of such units for reflection-free connection of cables with different characteristic impedances is outlined. A simple conversion rule is given which enables the calculations to be applied to lines of different characteristic impedance.

621.392:621.396.67 2430

Anti-Resonant H. F. Transmission Lines, Input Impedance Characteristics—H. M. Barlow. (*Wireless Eng.*, vol. 29, pp. 145–147; June, 1952.) Treatment of the short-circuited $\lambda/4$ line, deriving convenient expressions for the maximum value of the resistive component R_s and of the reactive component X_s of the input impedance. $R_{s,max} \approx 2X_{s,max}$. The lengths of line at which these maxima occur are slightly different.

621.392.2:621.396.67 2431

A Cage Type of Feeder—A. Schweisthal. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, pp. 45–47; March/April, 1952.) Description of feeder lines for broadcasting antennas. At Coblenz, Saarbrücken, Bad Dürrheim and Ravensburg the lines are 200 m long; that for the Rhine transmitter has a length of 500 m. The central conductor is a Cu tube (2-cm diameter) with joints hard soldered and is supported by an internal steel-wire rope under 500 kg tension. The outer screen consists of twelve 3-mm Cu wires evenly spaced round a circle of diameter 50 cm and with a similar total tension. Insulated supports, carried on poles 6 m high, are provided at 12-m intervals for both core and wires, the latter having short-circuiting rings at the ends. The characteristic impedance is about 205 Ω. Attenuation is considerably less than for a 9.5/36 cable and total cost much less.

621.392.21 2432

Note on the Variations of Phase Velocity in Continuously-Wound Delay Lines at High Frequencies—I. A. D. Lewis. (*Proc. IEE*, Part III, vol. 99, pp. 158; May, 1952.) Discussion on 2637 of 1951.

621.392.26 2433

The Completeness of the System of *E*- and *H*-Type Waves in Waveguides—E. Ledinegg and P. Urban. (*Arch. elekt. Übertragung*, vol. 6, pp. 109–113; March, 1952.) It is proved mathematically that the solutions correspondind to the *E*-type and *H*-type waves represent all the possible solutions of Maxwell's equations, and that the plane waves in waveguides constitute a complete system in this sense. An essential element in the proof is the assumption of a finite value for the em field at infinity.

621.392.26 2434

The Impedance of Unsymmetrical Strips in Rectangular Waveguides—L. Lewin. (*Proc. IEE*, Part III, vol. 99, pp. 167–168; May, 1952.) Summary only. Formulas are derived for the impedance of inductive and capacitive strips situated either centrally or unsymmetrically in a waveguide.

621.392.43 2435

The Use of Directional Couplers in Aerial-Matching Problems—S. Gratama. (*Tijdschr. ned. Radiogenoot.*, vol. 17, pp. 85–102; March, 1952.) The operation of the coaxial-line reflectometer is described; this incorporates two directional couplers, measuring the intensity of the original and reflected waves respectively. Advantages of this instrument over the standing-wave indicator include wide frequency range, absence of moving parts, rapid operation. Calculations are made for an experimental model for the frequency band 5–500 mc. A method is described for obtaining a cro indication of the bandwidth of an antenna.

621.396.67 2436

A Note on Booker's Extension of Babinet's Principle—R. S. Elliott. (Proc. I.R.E., vol. 40, p. 729; June, 1952.) Discussion showing that Booker's extension of Babinet's principle (1335 of 1947) is only applicable to a restricted class of apertures with symmetry about the polarization axis of the primary source.

621.396.67 2437

A New Solution for the Current and Voltage Distribution on Cylindrical, Ellipsoidal, Conical or Other Rotationally Symmetrical Aerials—O. Zinke. (*Frequenz*, vol. 6, pp. 57–65; March, 1952.) The method is based on the solution of the static potential equation $\Delta\phi = 0$. For rotationally symmetrical antennas the solution can be effected either by means of electrolyte-tank measurements, or graphically, or by Southwell's relaxation methods. The es field strength normal to the metal surface is thus known and the charge per unit length of the surface contour is deduced. Only in exceptional cases, such as homogeneous cables or ellipsoidal antennas, is the charge per unit length independent of position. In the dynamic case the current and voltage distributions are sinusoidal only in these special cases. Constant charge per unit length in the static case thus corresponds to sinusoidal current distribution in the dynamic case, and nonuniform static charge density to nonsinusoidal current distribution. The scalar potential along cylindrical transmission lines and antennas is sinusoidal; on circular plates it is given by Bessel's functions. The theory is applied to the determination of the charge, current and voltage distributions along a cylindrical antenna for the cases of current resonance ($l \sim \lambda/4$) and voltage resonance ($l < \lambda/2$).

621.396.67 2438

Theory of Multiple-Feed Aerials—R. Walter. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, pp. 12–16; January/February, 1952.) Analysis shows that a current distribution corresponding to the function e^{-z^2} gives a good directional characteristic for a broadcasting antenna, but that even better characteristics can be obtained with multiple-feed arrangements. Vertical directional characteristics are shown for a system of two antennas of heights λ and $\lambda/2$, independently fed at the foot, the ratio of the currents having the values 1.5, 2, 2.5, 3 and 4. Corresponding characteristics for single homogeneous antennas of heights 0.56λ and 0.625λ are included for comparison. The current ratio 4 gives the closest approximation to the Gaussian curve, but in practice a ratio of 2.5 or 3 is preferred. The 0.625λ single antenna has a null point at 37 degrees, but a side lobe with an amplitude 31 per cent of that of the horizontal radiation. Any further increase of height is consequently not permissible. With the double antenna system, improvement of the directional characteristic by increase of antenna height is practicable up to 1.2λ, while for a triple-antenna system with triple feed, improvement of the characteristic is theoretically possible up to a height of about 1.8λ.

621.396.67 2439

Experience with Double-Feed Medium-Wave Aerials—A. Schweisthal. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, pp. 52–59; March/April, 1952.) The antennas at Bad Durrheim and Ravensburg are essentially similar and consist of square-section lattice masts 120 m high, insulated at the foot and with an insulating section at a height of 75 m. The lower section forms a 175-Ω transmission line, with a galvanized-iron tube as core, for feeding the upper section. Copper tubes connect the tubular core and the foot of the lower section with the antenna tuning network used for adjusting the location of the potential node. The observed vertical radiation diagram is shown together with the calculated one. For the Rhine transmitter, the required 2:1 or 3:1 concentration of the radiation in the NW-SE direction is effected by means of two 150-m antennas, divided at a height of 80 m and located 100 m (about λ/3) apart on a NW-SE line. These are fed in antiphase with a prescribed power ratio. A diagram shows the horizontal radiation pattern (*a*) with one mast fed at the foot and the other earthed, (*b*) with double feed for both masts. A method of measuring the potential distribution on a transmitting antenna is outlined in an appendix.

621.396.67:621.397.62 2440

Television Receiving Aerials—F. R. W. Strafford. (*Wireless World*, vol. 58, pp. 213–218 and 264–267; June and July, 1952.) The characteristics of simple dipole and multi-element types of antenna are discussed; calculated and measured values of impedance are plotted against frequency for two types of dipole for channel 4. In receiving antennas, losses due to mismatching are less important than those due to feeder attenuation. With present British television standards there is no advantage in using folded dipoles, though the greater bandwidth corresponding to the French 819-line standard does require their use. Problems involved in making measurements on antennas are examined, and some aspects of indoor antennas and the mechanical design of outdoor antennas are discussed.

621.396.671:537.311.5:538.569 2441

On the Current Induced in a Conducting Ribbon by the Incidence of a Plane Electromagnetic Wave—E. B. Moullin and F. M. Phillips (*Proc. IEE*, Part III, vol. 99, pp. 165–166; May, 1952.) Summary only. The analysis in terms of Mathieu functions given by Morse and Rubenstein (905 of 1939) for the diffraction of plane waves by ribbons and slits is extended, and the distribution of current density in ribbons of widths λ/π, $2\lambda/\pi$ and $4\lambda/\pi$ is evaluated. The results show that, in the range of widths examined, the distribution of both the in phase and the quadrature component near the edge depends very little on the width of the ribbon. The distribution is practically the same as that near the edge of a half-plane, for which a solution has previously been given. The disturbed density is largely concentrated in a region very near the edge and can be replaced by an equivalent filament for the purpose of predicting the polar diagram. A practical treatment is thus available which does not involve laborious mathematics. A curve is given showing the strength of the echoed field as a function of ribbon width; this is valid down to zero width.

621.396.677 2442

An Annular Corrugated-Surface Antenna—E. M. T. Jones. (Proc. I.R.E., vol. 40, pp. 721–725; June, 1952.) Analysis is presented for an antenna system which is the axially symmetrical counterpart of the rectangular antenna discussed by Reynolds and Lucke (922 of May). The surface wave is easily excited from the end of a coaxial line, with the center conductor extending λ/4 above the surface of the antenna. The far-zone radiation pattern is uniform in the azimuthal direction and polarized in a direction perpendicular to the surface. The major lobe is directed slightly above the plane of the antenna. Experimental results for an antenna operating at a wavelength of about 4 cm, in a finite ground plane, are in good agreement with theory.

621.396.677 2443

A New Type of U.H.F. Lens [aerial]—J. C. Simon. (*Onde élect.*, vol. 32, pp. 181–189; April/May, 1952.) Theory shows that the usual defects of uhf antenna systems are not due to diffraction effects difficult to calculate, but to phase aberrations analogous to those met with in optical instruments. For this reason, in the antenna systems for the Paris-Lille link phase correction methods are adopted. The antennas consist of two elements, one being a small waveguide horn radiating towards a concave hyperbolic reflector of diameter 150 cm, which is rigidly attached, by means of a conical metal structure, to the second element, a special lens of variable index of refraction, constructed from metal plates drilled with circular holes of wavelength dimensions and with an aperture of 7 m². Phase correction is applied to both elements of the system, and it is possible to correct local phase defects. The resulting beam has a half-power width of 1.7 degrees, and the first ring at 4 degrees from the axis is 20 db below the main lobe. The measured gain is about 38.5 db for the 7-m² area and the swr is <1.12 in a band of 300 mc centered on 3.64 kmc. See also 970 of May and 820 of 1951 (Ortusi and Simon).

621.396.677:537.226 2444

Isotropic Artificial Dielectric—Corkum. (See 2523.)

621.396.677:621.396.9 2445

A Family of Designs for Rapid-Scanning Radar Antennas—R. F. Rinehart. (Proc. I.R.E., vol. 40, pp. 686–688; June, 1952.) Analysis resulting in the design of lenses with smaller feed circles than those of the lenses previously described (1593 of 1949), thus permitting rapid rotation of the source.

621.396.677.012.71+621.396.81 2446

Aerial Measurements in the Microwave Range—J. M. G. Seppen. (*Tijdschr. ned. Radiogenoot.*, vol. 17, pp. 63–83; March, 1952. Discussion, p. 84.) Energy radiation and collection in the wavelength range 3–10 cm are discussed generally. Various methods of obtaining radiation characteristics are mentioned, and an account is given of the method used for measurements over the path from DenHelder to Tessel, readings being taken at successive 10′ antenna rotation angles. The equipment is described, with special attention to the attenuator. Tidal effects over the sea path, and atmospheric scattering and attenuation are taken into account.

621.396.677.029.64 2447

A Practical Method for the Design of Parabolic Aerials for Microwaves—J. Deschamps and G. G. Esculier. (*Onde élect.*, vol. 32, pp. 209–213; April/May, 1952.) Description of a simple method for determining the contour of a reflector excited by a waveguide horn, and for evaluating the horn dimensions, for specified radiation characteristics. No elaborate mathematics is required.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.013.5:621.3.011.23:621.314.2.045 2448

Calculation of the Magnetic Field created by Conductors of Rectangular Cross-Section in a Slot, and its Application to the Determination of the Reactance of Transformer Windings—E. Billig. (*Rev. gén. Élect.*, vol. 61, pp. 135–149; March, 1952. Correction, *ibid.*, vol. 61, p. 196; April, 1952.) French version of paper abstracted in 602 of April.

621.3.015.7:621.387.4 2449

A Single-Channel Pulse-Amplitude Analyser for Measurement of Coincident Pulses—R. Wilson. (*Jour. Sci. Instr.*, vol. 29, pp. 70–72; March, 1952.)

621.314.3†:[621.396.615.17+621.396.619.2 2450

The Use of Saturable Reactors as Discharge Devices for Pulse Generators—W. S. Melville. (*Proc. IEE*, Part III, vol. 99, pp. 156–157; May, 1952.) Discussion on 2362 of 1951.

621.316.8.029.53/.55 2451

Survey of Radio-Frequency Resistors with Kilowatt Ratings—D. R. Crosby. (*RCA Rev.*, vol. 12, pp. 754–763; December, 1951.) Discussion of design and operating characteristics of resistors used for communication and RF-heating equipment in the frequency range 300 kc–30 mc. Resistance values may be from a few tenths of an ohm to about 600 Ω, and the power range is 5–100 kw. The three basic types used are metal-wire, carbon-film and water-column resistors.

621.316.842(083.74) 2452

A Design for Standard Resistance Coils—C. R. Barber, A. Gridley and J. A. Hall. (*Jour. Sci. Instr.*, vol. 29, pp. 65–69; March, 1952.) Details are given of a method of construction specially suitable for resistance coils used in bridges. A strain-free helix of minalpha (manganin) wire is supported in a spiral groove cut in a perspex disk, and hermetically sealed between perspex cover plates. The heat treatment of the wire to obtain good stability is described. Coils of 1,000 Ω resistance, made from about 42 m of 0.006-inch wire, showed a rise of resistance of the order of 2 parts per million per month during the first few months after winding.

621.316.87:541.18:537.311.35 2453

Polaresistivity and Polaristors—Hollmann. (See 2538.)

621.318.42.018.78 2454

Harmonic and Combination Oscillations in Ferromagnetic Materials—G. Hoffmann. (*Arch. elekt. Übertragung*, vol. 6, pp. 99–108; March, 1952.) Distortion effects at very low frequencies in coils with ferromagnetic cores are investigated theoretically, particularly for the case of simultaneous excitation by two sinusoidally varying fields of different frequencies. The relation between the complex-permeability curve and the combination frequencies is derived for the Rayleigh region (where the branches of the hysteresis loop are parabolic). For magnetically stable (carbon-free) materials the calculated values are well supported by measured values over a wide range. For magnetically unstable materials, the calculation provides, in conjunction with distortion measurements, a possible method of investigating the creep effect.

621.392 2455

Introduction to Formal Realizability Theory: Part X—B. McMillan. (*Bell Sys. Tech. Jour.*, vol. 31, pp. 541–600; May, 1952.) Discussion of conditions to be satisfied for a network to realize a given positive real impedance matrix. Part 1: 2138 of September.

621.392 2456

The Synthesis of RC Networks to have Prescribed Transfer Functions—H. J. Orchard (Proc. I.R.E., vol. 40, pp. 725–726; June 1952.) Discussion on 2371 of 1951.

621.392.5 2457

The Iterated Network and its Application to Differentiators—M. C. Pease. (Proc. I.R.E., vol. 40, pp. 709–711; June, 1952.) A compact and convenient expression is derived for the transmission matrix of any iterated structure in terms of the number of units in the structure and of the transmission matrix of the unit element. The expression is used for analysis of the operation of an iterated RC differentiating network, which is shown to have an effective time constant much less than that of the unit element.

621.392.5 2458

Image Impedances of Active Linear Four-Terminal Networks—H. Sutcliffe. (*Wireless Eng.*, vol. 29, pp. 169–170; June, 1952.) A matrix treatment showing that the formula expressing the image impedance of a passive quadripole in terms of open-circuit input impedance and short-circuit admittance applies to an active linear network.

621.392.5 2459

The Gyrator—G. W. O. H. (*Wireless Eng.*, vol. 29, pp. 143–145; June, 1952.) Comment, with further analysis, on the special 4-terminal network conceived by Tellegen. See 301 of 1951 and back references.

621.392.5 2460

Operational Analysis of Variable-Delay Systems—L. A. Zadeh. (PROC. I.R.E., vol. 40, pp. 564–568; May, 1952.) The output of a variable-delay system is related to its input by a delay operator which has the usual exponential form, but differs from conventional (time-invariant) delay operators in that the time delay is a function of time. Systems in which the variation of delay is due to motion of the receiver (R) or source (S) or both (RS) are analyzed in general terms. An operational relation is obtained for the correlation function of the output of a type-R system, and is applied to the determination of the correlation function of a FM sound wave.

621.392.5:517.56 2461

The Approximation with Rational Functions of Prescribed Magnitude and Phase Characteristics—J. G. Linvill. (PROC. I.R.E., vol. 40, pp. 711–721; June, 1952.) A method of successive approximations is applied to the selection of network functions having desired magnitude and phase variation with frequency. Adjustment of the magnitude and phase characteristics is effected simultaneously.

621.392.5:534.321.9:534.133 2462

Performance of Ultrasonic Vitreous-Silica Delay Lines—M. D. Fagen. (*Tele-Tech*, vol. 11, pp. 43–45, 144; March, 1952.) The electrical performance of an ultrasonic delay line is analyzed in terms of its equivalent circuit; insertion loss and bandwidth are investigated in relation to the parameters of the piezoelectric transducer, the acoustic medium and the electrical termination. Results of tests at 10 and 60 mc and with resistive terminations of 75 to 1,000 Ω are shown. With low terminal impedance a large bandwidth is obtained but insertion loss is relatively high.

621.392.52 2463

Electrical Separating Networks with Series-Resonance Circuits as Blocking Elements—R. Becker. (*Telefunken Ztg.*, vol. 25, pp. 33–40; March, 1952.) Discussion of devices enabling a single antenna to be used with two transmitters or receivers simultaneously, or with a transmitter and a receiver. General design formulas are derived and applied to the design of a unit for use with two 1-kw transmitters with frequencies of 46.4 and 49.1 mc respectively. Calculated values of attenuation for the two frequencies were in good agreement with measured values (~93db) for practical equipment, which is described. The construction is also described of a transmitter-receiver type for the range 45–75 mc and of a low-power unit for beam R/T on 80 mc.

621.392.52 2464

On the Theory of Filtration of Signals—L. A. Zadeh. (*Z. angew. Math. Phys.*, vol. 3, pp. 149–156; March 15, 1952. In English.) An outline of the theory of linear variable filters. See also 2147 of September.

621.392.52 2465

Theory of Transmission Time and Build-Up Time in Electrical Filters with Phase Distortion—T. Laurent. (*Arch. elekt. Übertragung*, vol. 6, pp. 91–98; March, 1952.) The theory is based on the frequency-transformation method developed previously (471 of 1937). New definitions of transmission time and build-up time are derived which are valid for filters with phase distortion and which enable values to be calculated easily from attenuation and phase shift.

621.392.52.015.7 2466

Pulsed Circuits. Transmission Function. Problem of Isomorphic Transmission—H. Borg. (*Ann. Télécommun.*, vol. 7, pp. 115–126; March, 1952.) A definition of an ideal filter, based on the symbolic expression for the transient response of a passive linear system, implies certain conditions of amplitude, phase and pass band. Echo phenomena in actual filters are studied by different series developments of the transmission function and by analysis based on Laplace transforms. An analytical expression for the transmission characteristic of a passive circuit is derived in terms of pulse amplitude, duration and delay time; this determines the conditions under which no alteration of the pulse shape occurs. 39 references.

621.392.54.012.3:621.392.26 2467

Chart for the TE_{11} Mode Piston Attenuator—C. M. Allred. (*Bur. Stand. Jour. Res.*, vol. 48, pp. 109–110; February, 1952.) An abac for determination of attenuation as dependent on frequency and on conductivity and radius of a cylindrical-waveguide attenuator.

621.395.645:621.395.97 2468

Radio-Diffusion Amplifiers for Standard [telephone] Circuits—J. Jacot. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, vol. 30, pp. 81–87; March 1, 1952. In French.) Developments in Switzerland up to 1938 are reviewed and descriptions are given of two types of amplifier designed to meet C.C.I.F. requirements for program transmission on standard telephone circuits. The two types are similar in principle, both having two coupled stages, the essential difference being in the feedback circuit used to correct the gain for the very low and the high frequencies.

621.395.661.1 2469

Study of, Tests on, and Suggestions for Acceptance Standards for Repeater Coils used on Lines for Musical-Programme Transmission—R. Salvadorini. (*Poste e Telecommunicazioni*, vol. 20, pp. 115–130; March, 1952.) The characteristics of repeater coils are examined, and results of tests on six samples reported. Minimum performance requirements in respect of insulation, distortion, losses, frequency characteristics, transients, crosstalk, and dc tests are listed, as a basis for acceptance standards.

621.396.611.1 2470

Free Oscillations in n-Mesh Networks with Varying Parameters—W. Haacke. (*Arch. elekt. Übertragung*, vol. 6, pp. 114–119; March, 1952.) The system is represented by a matrix of n linear second-order differential equations with periodically varying coefficients; the individual equations are separated by a linear transformation and solved by Erdélyi's method (1934 Abstracts, p. 436), and the complete solution is obtained by combining the individual solutions. Circuits with capacitances varying periodically about a mean value are mainly considered; the calculation is similar for the case of varying inductances. The analysis applies only to cases where all the parameter variations obey the same law. Where the solutions of any of the n equations are unstable, the corresponding fundamental frequencies may disappear completely, to be replaced by multiples of half the variation frequency.

621.396.611.1:517.51 2471

Resonant Circuit with Periodically Varying Parameters—P. Bura and D. M. Tombs. (*Wireless Eng.*, vol. 29, pp. 95–100 and 120–126; April and May, 1952.) Theoretical and experimental investigation of the circuit with periodically varying resistance. The variation is achieved by applying a voltage of given frequency to the grid of a dynatron, thus varying the magnitude of the negative resistance which the dynatron presents to the oscillatory circuit. A steady-state solution of Mathieu's equation is obtained by means of integral equations, and for the oscillatory regime a solution of an extended Hill's equation is obtained by a method similar to that of Ince. The steady-state response to an applied alternating voltage, of frequency approximately the same as the resonance frequency, exhibits multiple resonance effects, each maximum corresponding to detuning equal to an integral multiple of the frequency of the resistance variation. As in the case of oscillation excitation by variation of inductance or capacitance, the frequency at which oscillations are most easily excited by resistance variation is double the resonance frequency of the circuit. If the alternating voltage applied to the dynatron grid is increased gradually from zero, at a certain critical voltage oscillations suddenly commence, with frequency exactly half that of the grid voltage, even if the circuit is detuned. Experimental results are shown graphically.

621.396.611.21.029.3 2472

Quartz Vibrators for Audio Frequencies—J. E. Thwaites. (*Proc. IEE*, Part III, vol. 99, pp. 158–159; May, 1952.) Summary only. Results obtained for silver-plated wire-mounted +5 degrees X-cut bars show that the frequency f of flexural vibrations is given by the formula $f=5,740\ w/l^2$, where f is in kc and w and l are respectively the width and length of the bar in mm. The frequency/temperature characteristics of such bars are approximately parabolic, with a vertex of maximum frequency at some temperature between −10 degrees and +50 degrees C, depending on the value of w/l. For bars of ring form with a small gap, $f=4,350\ w/l^2$ where w is the width in the plane of flexure and l is the mean arc length. Approximate relations between dimensions, frequency and temperature are shown graphically for straight bars and for gapped rings. Both types have two nodes and can be supported by four wires in a glass envelope. In straight bars the distance of the nodes from each end is 0.224 l; in gapped rings the nodes subtend an angle of 66 degrees at the center of curvature. A simple method of determining the parameters of the equivalent circuit for such vibrators is described.

621.396.611.39:621.392.26 2473

Microwave Coupling by Large Apertures—S. B. Cohn. (PROC. I.R.E., vol. 40, pp. 696–699; June, 1952.) A frequency-correction factor is proposed for Bethe's small-aperture coupling relation for a transverse diaphragm in a rectangular waveguide. Experimental results for apertures of many different shapes and sizes show this correction factor to be accurate up to slightly above the resonance frequency of each aperture. Approximate formulas are given for the resonance length of a narrow rectangular aperture and for the Q of a resonant iris loaded by matched waveguide. The effect of wall thickness is also considered.

621.396.615.001.8 2474

The Action of Locked Oscillators—E. Roessler. (*Fernmeldetech. Z.*, vol. 5, pp. 97–100; March, 1952.) Fixing of the starting point of oscillations by means of a locked oscillator is used at present for three purposes: (*a*) for the suppression of transmitter noise in multi-channel p.phm transmission on dm-wave links, (b) for the suppression of receiver noise in superregenerative reception, (c) for the production of a phase-related pulse for utilizing the Doppler effect in radar measurements. The action of the locked oscillator in these three cases is explained and further possible applications are considered.

621.396.615.029.3:621.396.621.53 2475
Beat-Frequency Tone Source. Mathematical Theory of Mixing—C. G. Mayo. (*Wireless Eng.*, vol. 29, pp. 148–155; June, 1952.) Three systems of mixing are discussed: (a) multiplication in a square-law device, (b) addition followed by linear rectification, (c) addition of one component to a square wave synchronous with the other component, followed by linear rectification. Method (c) is analyzed in detail and the distortion likely to occur in practical circuits is evaluated. With regard to output, distortion and noise, method (c) is by far the best.

621.396.615.17 2476
Production of Very Short Pulses by Means of a Pulse-Excited Oscillatory Circuit Shunted by a Germanium Crystal—H. Mayer. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1131–1133; March 10, 1952.) The circuit described by Reiffel (2375 of 1951) is adapted for actuation by pulses from a multivibrator. The optimum operating conditions are calculated and a comparison is made between theoretical and experimental results.

621.396.615.17:621.3.087.4:551.510.535 2477
A Timebase Circuit for a High-Precision Ionospheric Sounding Apparatus—B. M. Banerjee and R. Roy. (*Indian Jour. Phys.*, vol. 24, pp. 411–419; September, 1950.) Description of a circuit which produces a 10-line raster on the screen of the cro. Each line of the raster takes 333μs, a time which corresponds to 50 km of ionosphere height. Marker pips indicate 5 km intervals and an intensifying pulse is applied to any one of the ten lines selected by an adjustable trigger circuit. Fully detailed circuit diagrams of the different units of the equipment are given.

621.396.615.17:621.396.615.141.2 2478
Magnetron Harmonics at Millimeter Wavelengths—J. A. Klein, J. H. N. Loubser, A. H. Nethercot, Jr., and C. H. Townes. (*Rev. Sci. Instr.*, vol. 23, pp. 78–82; February, 1952.) Harmonics present in the output of magnetrons are isolated (a) by using a tapered waveguide section to cut off at wavelengths shorter than the fundamental, (b) by using a diffraction grating. A Golay cell or a Si-crystal rectifier is used as detector. The tenth harmonic ($\lambda=1.25$ mm) has been obtained from K-band tubes and the third harmonic ($\lambda=1.1$ mm) from 3.3-mm tubes, with an estimated peak power of a few hundred microwatts. Data are included on the harmonic spectra of various types of magnetron.

621.396.616.015.7:621.316.546 2479
Nonelectronic Rectangular-Wave Generator—B. Bederson and M. Silver. (*Rev. Sci. Instr.*, vol. 23, p. 133; March, 1952.) A circuit including a mercury relay tube in series with dc supply and resistive load produces square pulses of peak power 250 w and duration from 10 ms to 2.5 seconds.

621.396.645 2480
Cathode-Coupled Amplifier—I. F. Macdiarmid. (*Wireless Eng.*, vol. 29, p. 169; June, 1952.) Discussion on 1559 of June (Lyddiard). An alternative equivalent circuit and convenient method of calculating harmonic distortion are described.

621.396.645 2481
Cathode-Follower Operation—A. J. Shimmins. (*Wireless Eng.*, vol. 29, pp. 155–163; June, 1952.) The response of cathode-follower circuits to pulse and sawtooth signals has been considered previously (846 of 1951). Three methods of improving the transient and steady-state performance of cathode-follower circuits with a capacitive load are studied: (a) simple series-inductance compensation, (b) use of a low-pass filter as the cathode load, (c) increase of the capacitance between grid and cathode. Method (a) offers considerable advantages; an inductance value of 0.5 $C_LR_0^2$ is found suitable, where C_L is the load capacitance and R_0 the output impedance.

621.396.645 2482
Amplifier Frequency Response—D. A. Bell. (*Wireless Eng.*, vol. 29, pp. 118–119; May, 1952.) Discussion of the effect of feedback, including consideration of (a) faults that cannot be corrected by use of feedback, (b) cases in which the desired result can be achieved equally well by other means.

621.396.645:621.396.822 2483
Background Noise in Amplifiers—Its Reduction—Application in Physiology—B. Bladier. (*Acustica*, vol. 2, no. 1, pp. 23–34; 1952. In French.) Experiments with a 1–200-cps amplifier for encephalography showed that the background noise consisted mainly of thermal agitation noise in the input resistance and in the load resistance of the first tube, with a smaller contribution due to shot effect. By reducing the input resistance and using a diode as the load resistance, the noise was reduced by two thirds. Type-4673 and Type-1603 tubes were found to be best suited for the particular purpose.

621.396.645.018.7 2484
Distortion of N-Shaped Signals in RC Amplifiers—H. Oertel. (*Funk u. Ton*, vol. 6, pp. 123–129; March, 1952.) Analysis indicating that with a single-stage amplifier, less than 10 per cent distortion of an N-shaped pulse is only to be expected when $\omega_u \cdot T<0.1$ and $\omega_o \cdot T>100$, T being the signal duration and ω_u and ω_o the angular velocities corresponding to the lower and upper limiting frequencies respectively.

621.396.645.029.3 2485
High-Quality Amplifier Modifications—D. T. N. Williamson. (*Wireless World*, vol. 58, pp. 173–176; May, 1952.) Describes how the original circuit (3101 of 1949) should be modified for use with long-playing records, and gives details of circuit changes necessary for the direct connection of high-impedance pickups. Published articles will be reprinted in a revised edition of the "Williamson Amplifier" booklet.

621.396.822:621.396.615:529.786 2486
Effect of Background Noise on the Frequency of Valve Oscillators—A. Blaquière. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1140–1142; March 10, 1952.) In most practical cases the primary changes of phase and amplitude produced by noise pulses (see 2162 of September) are accompanied by a secondary effect due to the dependence of oscillation frequency on amplitude; the frequency fluctuation is related to noise power. The magnitude of this secondary effect is the best criterion of quality of an electronic clock. The extension of Nyquist's theory previously developed (335 of March) is used to classify oscillators from this point of view.

GENERAL PHYSICS

535.42:538.566 2487
Huyghens' Principle in Diffraction Problems—J. P. Schouten and A. T. de Hoop. (*Tijdschr. ned. Radiogenoot.*, vol. 17, pp. 45–62; March, 1952.) Using a method due to Clavier (*Elec. Commun.*, vol. 25, p. 148; 1948), Huyghens' Principle is derived directly from Maxwell's equations, without introducing fictitious magnetic charges and currents. In the limiting case when the surface over which the integration is extended is an infinite plane, the expressions obtained coincide with those derived by Bethe (706 of 1945) and Smythe (*Phys. Rev.*, vol. 72, p. 1066; 1947).

537.311.31 2488
On the Theory of Electrical Conductivities of Monovalent Metals—A. B. Bhatia. (*Proc. Phys. Soc.*, vol. 65, pp. 188–191; March 1, 1952.)

538.221 2489
Some Magnetic Properties of Metals: Part 1—General Introduction, and Properties of Large Systems of Electrons—R. B. Dingle. (*Proc. Roy. Soc. A*, vol. 211, pp. 500–516; March 20, 1952.) The Schrödinger equation is solved for an unbounded system, and the limiting case of a system much larger than the electronic orbits is considered. Expressions for the density of states and the free energy of the system are derived, and the magnetic susceptibility is evaluated assuming the thermodynamic potential per electron is constant. Explicit formulas are given for the temperature dependence of the field-independent term in the susceptibility. Corrections for electron spin are applied to the results obtained.

538.221 2490
Some Magnetic Properties of Metals: Part 2—The Influence of Collisions on the Magnetic Behaviour of Large Systems—R. B. Dingle. (*Proc. Roy. Soc. A*, vol. 211, pp. 517–525; March, 20, 1952.)"A discussion of the effect of collisions on the magnetic properties of a large system of free electrons shows that the nonperiodic term in the susceptibility is hardly affected, but that the periodic terms are reduced in magnitude by a factor exp $(-hp/\iota\beta H)$, where p is the harmonic considered, ι is the mean collision time, and $\beta=eh/2\pi mc$."

538.521 2491
Effect of Torsion on a Longitudinally-Magnetized Iron Wire—G. W. O. H. (*Wireless Eng.*, vol. 29, pp. 115–117; May, 1952.) An account of effects observed by W. V. Dromgoole, New Zealand. An emf is induced in a ferromagnetic wire on twisting one end of it in an alternating axial magnetic field, due to the circular component of the alternating magnetic flux; the emf depends on the angle of twist and the permeability of the material and is large over the frequency range 300 cps to 20 kc. Several practical applications are suggested.

538.569.4 2492
On the Absorption of U.H.F. Radio Waves by Opalescent Binary Liquid Mixtures—A. Choudhury. (*Indian Jour. Phys.*, vol. 24, pp. 507–512; November, 1950.) Investigations of nitrobenzene-hexane and aniline-cyclohexane mixtures in the range 300–510 mc show that in both cases a new absorption peak appears on the lf side of the peak found for one of the pure constituents having polar molecules, the original peak being much reduced.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5:621.396.9 2493
Characteristics of Radio Echoes from Meteor Trails: Part 3—The Behaviour of the Electron Trails after Formation—J. S. Greenhow. (*Proc. Phys. Soc.*, vol. 65, pp. 169–181; March 1, 1952.) It is suggested that long-duration meteor echoes are due to reflection from trails with very high electron density. Further evidence is adduced to show that the amplitude fluctuations observed are caused by the influence of atmospheric turbulence on the meteor trail. Winds with velocities of the order of 20 m/s at heights between 80 and 100 km are inferred. See also 2782 of 1948 (Lovell and Clegg) and 359 of 1951 (Greenhow).

523.72:539.1 2494
Emission of Corpuscles from the Sun—K. O. Kiepenheuer. (*Jour. Geophys. Res.*, vol.

57, pp. 113–120; March, 1952.) Arguments in favor of the existence of solar corpuscles are reviewed and the geomagnetic action of such particles is analyzed. The streams producing moderate disturbances of the earth's magnetic field are identified with invisible extensions of the solar streamers.

523.746 "1951.10/.12" 2495
Provisional Sunspot-Numbers for October to December 1951—M. Waldmeier. (*Jour. Geophys. Res.*, vol. 57, p. 138; March, 1952.) See also *Z. Met.*, vol. 6, p. 58; February, 1952.

621.396.11:523.78 2496
Effect of the Annular Eclipse of March 7, 1951, on Radio-Wave Propagation—(See 2575.)

523.8:621.396.822 2497
A New Radio Interferometer and its Application to the Observation of Weak Radio Stars—M. Ryle (*Proc. Roy. Soc. A.*, vol. 211, pp. 351–375; March 6, 1952.) The reception pattern of a pair of in-phase spaced antennas is $A(\theta)$. If a $\lambda/2$ length of feeder is inserted in the line to one of them the pattern in which the maxima and minima are interchanged, becomes $A^2(\theta)$. A switch changes the pattern rapidly from A_1 to A_2 so that the output contains an alternating component which is proportional to (A_1-A_2) and to the power flux received from a point source. An amplifier discriminates in favor of the alternating component and allows the output due to a point source to be observed without interference from steady sources or random noise. The method can be used for the detection of weak point sources but has other important advantages, including high accuracy of position finding.

550.372:551.311.234.5:621.3.029.62 2498
The Electrical Constants of Soil at Ultra-High Frequencies—P. M. Sundaram. (*Indian Jour. Phys.*, vol. 24, pp. 469–478; November, 1950.) A Lecher-wire system, completely buried in the soil for an adjustable length, was used to investigate the variation of the dielectric constant κ and conductivity σ of a clay soil as a function of frequency (43–74 mc), sand admixture and moisture content. Results are shown in tables and curves. A peak in the curves for κ occurs at about 47 mc. At all the frequencies used, κ increases with moisture content up to 10 per cent or 12 per cent and then decreases. A maximum value of σ was found for moisture content of 8–10 per cent.

550.38 2499
The Earth's Magnetism and its Changes—S. Chapman. (*Proc. Indian Ass. Cult. Sci.*, vol. 33, 16 pp.; 1950; *Indian Jour. Phys.*, vol. 24, 16-page insert between pp. 420–421; September, 1950.) Text of Ripon Professorship Lecture, Calcutta, January 1949, reviewing present knowledge and discussing various theories.

550.38:523.75:551.510.535 2500
Characteristics of the Solar Flare Effect (Sqa) on Geomagnetic Field at Huancayo (Peru) and at Kakioka (Japan)—T. Nagata. (*Jour. Geophys. Res.*, vol. 57, pp. 1–14; March, 1952.) The transient characteristics of the geomagnetic-field variations due to solar-flare effects are examined statistically, taking into account induction effects in the ionosphere and in the earth. An estimated value of $6-7\times10^{-8}$ emu is found for the integrated conductivity of the ionosphere over Huancayo, Kakioka and Watheroo.

550.38"1951.07/.09" 2501
International Data on Magnetic Disturbances, Third Quarter, 1951—J. Bartels and J. Veldkamp. (*Jour. Geophys. Res.*, vol. 57, pp. 135–137; March, 1952.)

550.38"1951.10/.12" 2502
Cheltenham [Maryland] Three-Hour-Range Indices *K* for October to December, 1951—R. R. Bodle. (*Jour. Geophys. Res.*, vol. 57, p. 138; March, 1952.)

550.384/.385 2503
Geomagnetic Field Variations at Kodaikanal—M. V. Sivaramakrishnan. (*Nature* (London), vol. 169, pp. 409–410; March 8, 1952.) Anomalies in sudden commencements, and geomagnetic effects of solar flares, are reported for the period 1949–1951.

550.385 2504
On the Theory of the First Phase of a Geomagnetic Storm: a New Illustrative Calculation Based on an Idealized (Plane not Cylindrical) Model Field Distribution—V. C. A Ferraro. (*Jour. Geophys. Res.*, vol. 57, pp. 15–49; March, 1952.) Extension of discussion presented previously by Chapman and Ferraro (14 of 1941).

550.385"1951.07/.12" 2505
Principal Magnetic Storms [July-Dec. 1951]—(*Jour. Geophys. Res*, vol. 57, pp. 139–141; March, 1952.)

551.510.3 2506
The Pressure, Density, and Temperature of the Earth's Atmosphere to 160 Kilometers—R. J. Havens, R. T. Koll and H. E. LaGow. (*Jour. Geophys. Res.*, vol. 57, pp. 59–72; March, 1952.) From rocket measurements made at White Sands, New Mexico, and at the equator the following values are deduced: pressure at 160 km, 2×10^{-6} mm Hg; density at 160 km, 1.5×10^{-6}g/m³: temperature passes through a maximum of 270°K at 50 km and a minimum of 190°K at 80 km, increasing to about 500 K at 160 km.

551.510.535 2507
Limitations on the Calculation of Expected Virtual Height for Specific Ionospheric Distributions—J. Shmoys. (*Jour. Geophys. Res.*, vol. 57, pp. 95–111; March, 1952.) Virtual height is defined (*a*) on a geometrical-optics basis and (*b*) in terms of the frequency derivative of the phase of the reflection coefficient; the former definition can be derived from the latter by use of the phase-integral method. The two definitions are compared for the cases of linear, rectangular, Epstein and parabolic charge distributions. When the reflected wave contains more than one pulse the relation between virtual height and frequency derivative of phase is not valid; in this case the frequency derivative of phase cannot be interpreted as the time delay of any one of the pulses.

551.510.535 2508
The Reflection Coefficient of the Exponential Layer—J. Shmoys. (*Jour. Geophys. Res.*, vol. 57, pp. 142–143; March, 1952.) Discussion on 132 of February (Mitra).

551.510.535 2509
Movements of the Sporadic-*E* Layer of the Ionosphere—N. C. Gerson. (*Z. angew. Phys.*, vol. 4, pp. 81–82; March, 1952.) Report of observations made in North America on 16th and 17th June 1949. Mean drift velocities of 130 and 300 km/hr were deduced. See also 2426 and 2998 of 1951.

551.510.535.525.624 2510
Tides in the Ionosphere—A. P. Mitra. (*Indian Jour. Phys.*, vol. 24, pp. 387–404; September, 1950.) A connected account of the results of various recent investigations, both theoretical and experimental, of tidal effects in the ionosphere, and discussion of Martyn's electrodynamic theory of such effects. Results of observations at Calcutta, Delhi and Chunking during 1946–1948 are presented; the curves showing the average diurnal variations of the height of the F_2 layer have two maxima, one about noon and the other about midnight.

551.510.535:551.594.5 2511
The Association of Absorption and E_s Ionization with Aurora at High Latitudes—J. P. Heppner, E. C. Byrne and A. E. Belon. (*Jour. Geophys. Res.*, vol. 57, pp. 121–134; March, 1952.) An analysis based on the coexisting auroral conditions is made of nocturnal E_s ionization and "no echo" occurrences as observed from *hf* records obtained at College, Alaska, during the period 8th Sept. 1950–16th of April 1951. In general, (*a*) E_s ionization increases at successively greater heights as aurora approaches the zenith from the north; (*b*) in the presence of different nonpulsating auroral forms the E_s ionization varies with changes in auroral form in the same way as luminosity varies, and variations in the height of maximum ionization follow variations in auroral height; (*c*) complete absorption is only slightly more frequent during nonpulsating aurora than during absence of aurora, but prevails in the presence of pulsating aurora. Geomagnetic influences are discussed.

621.396.9 2512
Technique for Measurement of Radar Characteristics of Targets—R. G. Peters. (*TV Eng.* (N.Y.), vol. 2, pp. 10–11, December, 1951; and vol. 3, pp. 26–27; January, 1952.) Description of a method using models, the reflections from which cause unbalance of a hybrid junction in the waveguide feeding a horn antenna, the amount of unbalance being a function of the echoing area of the model. A 9.8-kmc frequency-stabilized transmitter was used, and balance stability was improved by constructing the hybrid junction and horn of invar. Suitable suspension arrangements for the models were determined by experiment. Calibration was effected by use of spheres, of diameter ranging from 5 to 10 inches.

621.396.9:[523.7+523.4 2513
On the Possibility of obtaining Radar Echoes from the Sun and Planets—F. J. Kerr. (PROC. I.R.E., vol. 40, pp. 660–666; June, 1952.)

621.396.9:621.396.8 2514
Fluctuations of Ground Clutter Return in Airborne Radar—T. S. George. (*Proc. IEE*, Part III, vol 99, pp. 160–161; May, 1952.) Summary only. Two types of fluctuations are analyzed: that in a single-range sweep at fixed azimuth where clutter is assumed to be due to a large number of small reflectors randomly situated on the ground, and that which occurs between video pulses at the same range due to the relative motion of the aircraft and ground.

621.396.932./.933].1+621.396.97 2515
Common-Wave Broadcasting and Hyperbolic Navigation—M. Pohontsch. (*Telefunken Ztg*, vol. 25, pp. 27–32; March, 1952.) Discussion of the connection between the problems of frequency control of common-wave transmitters and those of Decca and similar navigation systems. To be continued.

621.396.932.2 2516
Requirements for Modern Radio Direction Finders and Means for their Fulfilment—W. E. Steidle. (*Telefunken Ztg*, vol. 25, pp. 12–15; March, 1952.) Discussion of sensitivity and sharpness of direction indication, with comparison results for Telefunken ship equipment used with different antenna systems. The use of an iron-cored goniometer in conjunction with a high-gain superheterodyne receiver has increased the sensitivity of the Telegon df equipment, using crossed screened frame coils of area 0.95 m², up to that of equipment on the German hydrographic survey vessel *Gauss* which used crossed stretched-wire loops of area 9 and 52 m², respectively.

621.396.932.2:621.396.677.5 2517
Comparison between Frame-Coil and Stretched-Wire-Loop Aerials for Ship Direction Finders—H. Gabler, G. Gresky and W. Runge. (*Telefunken Ztg*, vol. 25, pp. 5–11; March, 1952.) Measurements showed that when using the Telegon direction finder with its crossed coils of area 0.95 m², the width of the minimum was 2.4 times greater than with

stretched-wire loops of area about 9 m². The direction-finding sensitivity of the Telegon equipment was found equal to that of goniometer direction finders using stretched-wire loops of area about 10 m². Theoretical investigations of the effect of the geometry and electrical data of a coil on the sharpness of the df indication were confirmed experimentally.

621.396.033 2518

An Aided Layer for Shoran—R. C. Richardson (*Aust. Jour. Appl. Sci.*, vol. 3, pp. 16–24; March, 1952.) Description of a unit in which a motor, whose speed is controlled through a velodyne [962 of 1948 (Williams and Uttley)] by rotation of the shoran handwheel, is geared differentially to the handwheel shaft and thus facilitates its rotation when the range is varying rapidly. The output drive from the unit to the display and other associated units is through an Admiralty M type electrical transmission system.

621.396.933.2 : 621.396.677.6 2519

Recent Developments in Short-Wave Adcock Direction Finders—A. Troost. (*Telefunken Ztg.*, vol. 25, pp. 16–27; March, 1952.) The relative merits of U-type and H-type antennas are discussed. For transportable equipment U-type antennas are preferable. Systematic errors can be reduced by increasing the number of antennas used; this also results in increased sensitivity. Subdivision of antennas gives improvement as regards night effect, suppression of resonance errors, and a cosine shape of vertical characteristic for single antennas throughout the frequency range. A 6-mast system is considered the best. With eight or more masts the slight increase of sensitivity is offset by increased liability to polarization and resonance errors. A description is given of equipment using six 8.5 m telescopic masts at the corners of a hexagon 8 m across. The goniometer is of the iron ring type with es screen. The use of a push-pull wide-band amplifier gives a sensitivity of 0.15–0.3 μV/m, depending on the frequency, the range covered in four bands being 4.5–25 mc.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.541.56 2520

The Measurement of Extremely Low Pressures below 10⁻⁸ mm Hg by means of an Adsorption Manometer—M. Seddig and G. Haase (*Z. angew. Phys.*, vol. 4, pp. 105–108; March, 1952.) Description of an experimental procedure by which pressure changes after gettering and sealing off are determined from the slope of the work function time characteristic of a purified tungsten emitting surface in the evacuated space. The work function is determined by photoelectric measurement.

535.5 2521

The Production of Very High Vacua by the Use of Getters—S. Wagener. (*Proc. IEE*, Part III, vol. 99, pp. 135–147; May, 1952.) "The characteristics of getters are defined and correlations between them are derived. Methods for measuring the characteristics are described and values of these for different getters and gases are given. Two main mechanisms of the gettering process are distinguished, namely contact gettering and discharge gettering, of which the latter, on account of the higher gettering rates attainable, is found to be of chief importance for the production of very high vacua. The processes contributing to discharge gettering are investigated in detail, and an attempt is made to determine the parts played by adsorption, diffusion and chemical reaction in different systems of getters and gases. The efficiency of flash getters is compared with that of coating getters, particularly thorium. It is found that gases like oxygen and carbon dioxide are taken up irreversibly during discharge gettering with thorium. The pressures attainable in tubes using flash and coating getters are measured, and the influence of some parameters such as baking time on the pump or pressure during sealing off is investigated."

535.215.4 : 537.311.33 : 546.86 2522

Properties of Films of Non-Metallic Antimony—T. S. Moss (*Proc. Phys. Soc.*, vol. 65, pp. 147–148, February 1, 1952.) Consideration of the periodic table of elements in the light of recent investigations of photoconductivity indicates that there should be a semiconducting form of Sb with activation energy between 0.37 and 0.1 ev and with photoconductive properties for wavelengths near 8 μ. Such layers were obtained experimentally by evaporation on to substrates at 195°K or 90°K. Resistance/reciprocal temperature and sensitivity/wavelength curves obtained from measurements are shown.

537.226 : 621.396.677 2523

Isotropic Artificial Dielectric—R. W. Corkum (Proc. I.R.E., vol. 40, pp. 574–587; May, 1952.) Theoretical and experimental investigations of media consisting of a cubic lattice of metal or dielectric spheres are described. Expressions are derived for the index of refraction, dielectric constant, and magnetic permeability of such media. These quantities are independent of frequency if the size and spacing of the spheres are small compared with the wavelength in the medium. Sample media were produced with steel or fused-quartz spheres embedded in styrofoam, waveguide measurements of their dielectric properties at 5 kmc are in good agreement with theoretical values.

537.311.33 2524

Impurity Scattering in Oxide Semiconductors—F. W. J. Mitchell (*Proc. Phys. Soc.*, vol. 65, pp. 154–161, February 1, 1952.) "It is suggested that electron scattering by neutral impurity centres makes an important contribution to the resistivity of these materials. The familiar relation $\log \sigma_0 = \alpha + \beta e$ together with the theory given by Busch are discussed and an alternative explanation involving impurity scattering and a dependence of ϵ on N is given. This is compared with experimental results for several oxides, including measurements by the author on the $Fe_2(Ti)O_3$ system." See also 907 of 1951 (Erginsoy).

537.311.33 2525

Study of the Conductivity of Semiconductors and Application to Oxide Mixtures—C. Guillaud and R. Bertrand (*Jour. Rech. Centre nat. Rech. Sci.*, no. 18, pp. 118–130; March, 1952.) Investigations are described of oxide mixtures of composition $xFe_3O_4(1-x)MO$, where M represents either Zn, Mg or Mn and x is the molecular proportion of Fe_3O_4 stoichiometric composition resulting in ferrites. Particular difficulties encountered in resistivity (ρ) measurements on semiconductors, such as the contact between the electrodes and the material, are discussed. Curves are given for the three types of mixture, showing log ρ for values of x from 0 to 1 and for temperatures of 20°, 100° and 200°C. The curves all indicate minimum resistivity for values of x near that for stoichiometric composition. In the case of the ZnO mixtures, a maximum of the resistivity is found for $x = 0.65$, decreasing to a low value for pure ZnO. Curves are also given showing ρ_{20}/ρ_{100} and ρ_{100}/ρ_{200} as functions of x, all exhibit sharp maxima with values about 40. Materials with compositions corresponding to such values can be used with advantage in the production of thermistors with very high sensitivity.

537.311.33 : 546.289 2526

Probing the Space-Charge Layer in a *p-n* Junction—G. L. Pearson, W. T. Read and W. Shockley. (*Phys. Rev.*, vol. 85, pp. 1055–1057; March 15, 1952.) Theory indicates that for a *p-n* junction the capacitance (C) per unit area is $C = K/(4\pi W)$, where K is the dielectric constant and W the thickness of the space-charge layer. The Ge *p-n* junction previously used by Goucher *et al.* (1669 of 1951), a specimen of square cross section, was investigated by measuring the zero-current potential at a number of points across the edge of the space-charge layer on one face of the specimen, using a finely pointed tungsten probe. Since the value of K is 16.1, fringing effects should be small, so that the potential distribution across the edge of the space-charge layer will be nearly the same as in the interior of the specimen. Measured potentials for three different reverse-bias voltages are plotted against distance of probe travel. The corresponding thicknesses deduced for the barrier layer are respectively 3.7, 2.9 and 2.4×10^{-3} cm if the curves are taken to be cubics. Similar results were obtained on the other three faces of the specimen. Since, however, the curves can be represented equally well as parabolas or cubics, it appears that the junction studied has neither a linear hole and electron concentration gradient [164 of February (McAfee *et al.*)] nor an abrupt transition from *n* to *p* type [379 of 1950 (Shockley)]. This introduces an uncertainty of about 30 per cent in the estimates of layer thickness.

537.311.33 : 621.396.822 2527

A Unidirectional Generator of Electrical Fluctuations—G. Wlerick (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1260–1262; March 17, 1952.) Observations on thin layers of PbS indicate that in some cases the additional noise fluctuations produced by passage of current may depend on the direction of the current, though the resistance is otherwise the same for both directions. It is shown that this does not violate the laws of thermodynamics. Fluctuation phenomena are capable of revealing directional differences small enough to escape detection by methods such as measurement of conductivity or photoconductivity.

538.221 2528

An Interpretation of the Magnetic Properties of Some Iron-Oxide Powders—W. P. Osmond (*Proc. Phys. Soc.*, vol. 65, pp. 121–134, February 1, 1952.) "The observed magnetic properties of fine dispersed powders of the ferromagnetic iron oxides are discussed in the light of modern theories of the magnetization of ferrites and of ferromagnetics containing nonmagnetic inclusions."

538.221 2529

Investigation of the Temperature Variations of Thin Mumetal Tapes subjected to Weak Alternating Fields—C. Abgrall and I. Épelboin. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1265–1267; March, 1952.) The calculated value of eddy-current losses for a homogeneous magnetic material accounts for only a part of the experimentally observed losses. The variation of the discrepancy over a wide temperature range is investigated; this enables the influence of magnetic after-effect to be separated from that of macroscopic structure. Experimental results are given for tapes of thickness 10μ and 59μ. See also 1933 of 1951 (Épelboin).

538.221 2530

Physical Structure and Magnetic Anisotropy of Alnico 5: Part 1—R. D. Heidenreich and E. A. Nesbitt. (*Jour. Appl. Phys.*, vol. 23, pp. 352–365; March, 1952.) Results of electron microscope and diffraction investigations indicate that the high coercive force and anisotropy of alnico 5 are due to a finely divided precipitate produced by the permanent-magnet heat treatment. Part 2: 2531 below.

538.221 2531

Physcial Structure and Magnetic Anisotropy of Alnico 5: Part 2—E. A. Nesbitt and R. D. Heidenreich. (*Jour. Appl. Phys.*, vol. 23, pp. 366–371; March, 1952.) Measurements of the magnetic anisotropy and coercive force of single crystals are described, and also the results of heat treatment on the physical structure. The present physical picture of the alloy is one of single domains of precipitate material in parallel with single domains of matrix material, the observed coercive force being the result of this combination.

538.221 2532
Domain Structure of Perminvar having a Rectangular Hysteresis Loop—H. J. Williams and M. Goertz. (*Jour. Appl. Phys.*, vol. 23, pp. 316–323; March, 1952.)

538.221 2533
Study of Imperfections of Crystal Structure in Polycrystalline Materials: Low-Carbon Alloy and Silicon Ferrite—J. J. Slade, Jr. and S. Weissmann. (*Jour. Appl. Phys.*, vol. 23, pp. 323–329; March, 1952.)

538.221:538.652 2534
Magnetostriction of Various Ferrites Oriented while Hot—L. Weil (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1351–1352; March 24, 1952.) Measurements were made on Co ferrite and on solutions of Co ferrite in Mg or Ni ferrites oriented by application of a field while cooling from 850°C to room temperature. Results are shown graphically.

538.221:[621.317.335.2+621.317.411 2535
Dielectric Constant and Permeability of Various Ferrites in the Microwave Region—T. Okamura, T. Fujimura and M. Date. (*Phys. Rev.*, vol. 85, pp. 1041–1042; March 15, 1952.) Results are tabulated of cavity-resonator measurements at 6.6-cm wavelength of polycrystalline Mg, Cu, Co, Ni and Mn ferrites.

538.221:621.318.2 2536
Torque Curves and Other Properties of Permanent-Magnet Alloys—K. Hoselitz and M. McCaig. (*Proc. Phys. Soc.*, vol. 65, pp. 229–235; March 1, 1952.) Further investigations of alcomax III and related alloys. See also 426 of March.

539.23:[546.28+546.57+546.621 2537
Electrical Properties of Very Thin Films of Silver, Aluminium and Silicon—A. Blanc-Lapierre, M. Perrot and J. P. David. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1133–1135; March 10, 1952.) The conductivity at ordinary temperature of films of equivalent thickness <10 mμ, evaporated at a pressure of 10^{-6} mm Hg, was investigated experimentally. Field strengths up to 8 kV/cm were used; large deviations from Ohm's law were observed, together with hysteresis effects in some cases. Results are tabulated and shown as I/V characteristics measured at different times after deposition.

541.18:537.311.35:621.316.87 2538
Polaresistivity and Polaristors—H. E. Hollmann. (Proc. I.R.E., vol. 40, pp. 538–545; May, 1952.) If semiconductive particles are suspended in an insulating fluid, the conductivity of the suspension becomes nonlinear as soon as the fibration at a certain critical value of the applied field reaches the point at which the semiconductive fibres bridge the gap between the electrodes. Particular fluid carriers can be transformed into a solid state, so forming what are termed "polaristors." The current I and voltage V for such resistors are connected by the relation $I=V(1+kV^2)/R_0$, where R_0 is the initial resistance and k a constant for any given material. Solid polaristors are produced by use of thermo-setting or cold-setting plastics. The degree of fibration in the forming process can be checked by means of cro I/V characteristics, or by IF oscillograms similar to modulation trapezoids. Polaristors are very sensitive to temperature and to mechanical stress and may consequently have a wide range of applications. See also 2247 of 1950.

546.23:546.49 2539
The Effect of Mercury on Selenium—H. K. Henisch and E. W. Saker. (*Proc. Phys. Soc.*, vol. 65, pp. 149–154; February 1, 1952.) "The interaction between crystalline selenium and liquid mercury or mercury vapor results in the formation of mercuric selenide. When produced in this form, this material is an excess semiconductor of high conductivity. The associated volume and surface diffusion have been investigated using radioactive ^{203}Hg. Mercury added to selenium before the crystallization process results in material of high resistivity. This is ascribed to a reduction of the positive hole mobility."

546.23.02 2540
A Note on the Structure of Selenium—E. Billig. (*Proc. Phys. Soc.*, vol. 65, pp. 216–221; March 1, 1952.) "The recently established identity of the width of the (optical) energy gap, 2.05 ev, for both amorphous and crystalline selenium, suggests that the chain structure of crystalline selenium is maintained also in the amorphous phase. The low electrical conductivity of amorphous selenium is ascribed to potential barriers existing between individual chains."

546.28+546.289]:541.26 2541
Lattice Parameters, Coefficients of Thermal Expansion, and Atomic Weights of Purest Silicon and Germanium—M. E. Straumanis and E. Z. Aka. (*Jour. Appl. Phys.*, vol. 23, pp. 330–334; March, 1952.)

546.3-74-56+546.3-98-57].02/.03 2542
Electronic Structures Physical Properties in the Alloy Systems Nickel-Copper and Palladium-Silver—B. R. Coles. (*Proc. Phys. Soc.*, vol. 65, pp. 221–229; March 1, 1952.)

547.476.3:548.55 2543
Growth-Rates of Single Crystals of Ethylene Diamine *d*-Tartrate: Flamed Growth and its Inhibition by Boric Acid—A. H. Booth and H. E. Buckley. (*Nature* (London), vol. 169, pp. 367–368; March 1, 1952.)

621.396.611.21 2544
Single Response Thickness-Shear Mode Resonators using Circular Bevelled Plates—R. Bechmann. (*Jour. Sci. Instr.*, vol. 29, pp. 73–76; March, 1952.) The effectiveness of bevelling for removing unwanted resonances is illustrated by response curves for an AT-cut quartz disk of diameter 27 mm. thickness 5.28 mm and frequency 350 kc, taken after applying one, two, three and four bevels respectively. Where there is a clear, single-frequency response, the series-resonance resistance of the crystal is almost constant over a wide temperature range; this affords a practical criterion of quality.

MATHEMATICS

519.24 2545
Properties of Certain Statistics involving the Closest Pair in a Sample of Three Observations—J. Lieblein. (*Bur. Stand. Jour. Res.*, vol. 48, pp. 255–268; March, 1952.) Analysis is presented which indicates that rejection of an outlying high or low reading from a triad not only tends to overestimate considerably the precision of the experimental procedure, but also gives less accurate results.

681.142 2546
A Sampling Analogue Computer—J. Broomall and L. Riebman. (Proc. I.R.E., vol. 40, pp. 568–572; May, 1952.) Description of a computer comprising an algebraic unit and a pulse converter. Only 15 tubes are used. Accuracy is within 1 per cent of full scale.

681.142:621.396.812.3:519.272 2547
A Computer for Correlation Functions—F. E. Brooks, Jr., and H. W. Smith. (*Rev. Sci. Instr.*, vol. 23, pp. 121–126; March, 1952.) Data relating to fluctuating phenomena, such as recordings of radio signals or temperatures, are transcribed in the form of pwm signals on to magnetic tapes. These are run through a computer which computes the difference between the integrals of two products in two separate channels and hence plots the required auto- or cross-correlation function. The apparatus is described and its operation is illustrated by application to the study of atmospheric turbulence.

MEASUREMENTS AND TEST GEAR

621.3.018.41+531.764.5](083.74)(431.55) 2548
Standard Frequencies and Standard Time of the "Funkamt" (FuA), Berlin, January and February 1952—F. Conrad. (*Frequenz*, vol. 6, p. 88; March, 1952.) Two graphs show comparisons between the time markings of the FuA standard-frequency equipment (Rhode and Schwarz quartz clock) with GBR, GIC and WWV signals and with the transmissions from the Hydrographic Institute, Hamburg, and from München-Ismaning.

621.317.3:621.396.611.21 2549
A Simple Method of Measurement of Quartz Crystal Characteristics—J. Coulon. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1269–1271; March 17, 1952.) The equivalent-circuit parameters of the crystal are found from (*a*) the curve for the variation of oscillator frequency with variation of a control capacitance in series with the crystal, (*b*) the curve for the voltage at the oscillator tube grid, (*c*) a capacitance measurement and (*d*) an impedance measurement. The circle diagram introduced previously (2550 below) is used.

621.317.3:621.396.611.21 2550
Plotting the Resonance Curve of a Piezo-electric Quartz Crystal, using a Circle Diagram—J. Coulon. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1138–1140; March 10, 1952.) The plotting of the resonance curve is facilitated by constructing a circle diagram whose co-ordinates are related to the parameters of the equivalent circuit of the crystal.

621.317.3:621.396.65 2551
Measuring Techniques for Broad-Band, Long-Distance Radio Relay Systems—W. J. Albersheim. (Proc. I.R.E., vol. 40, pp. 548–551; May, 1952.) Methods are outlined for investigation of (*a*) the transient response of the system to step voltages or square waves, (*b*) the frequency characteristics of gain, phase, impedance, and their frequency derivatives, (*c*) the amplitude characteristics of output nonlinearity and of intermodulation products

621.317.31/.32].029.3:537.324 2552
Thermal Converters as A.C.-D.C. Transfer Standards for Current and Voltage Measurements at Audio Frequencies—F. L. Hermach. (*Bur. Stand. Jour. Res.*, vol. 48, pp. 121–128; February, 1952.) An account of thermoelements and associated equipment used at the National Bureau of Standards primarily for the standardization of ac ammeters and voltmeters submitted for test. An accuracy to within 0.01 per cent in the ac-dc transfer can be achieved for currents of 1 ma–50 a and voltages of 0.2–750 v at frequencies from 25 cps to 20 kc. Factors limiting the transfer accuracy are analyzed and approximate solutions are given of nonlinear differential equations governing the heating of a conductor by an electric current.

621.317.324(083.74) 2553
Development of V.H.F. Field-Intensity Standards—F. M. Greene and M. Solow. (Proc. I.R.E., vol. 40, p. 573; May, 1952.) Abstract of U.R.S.I.-I.R.E. Meeting paper, Washington, 1949. See 3090 of 1950.

621.317.336:621.315.212 2554
Reflections in a Coaxial Cable due to Impedance Irregularities—G. Fuchs. (*Proc. IEE*, Part III, vol. 99, pp. 161–164; May, 1952.) Summary only. Theoretical relations between impedance irregularities, input-impedance deviation and pulse echo are derived and experiments are described which confirm the theory.

621.317.42 2555
A New Device for Measuring the Strength of a Magnetic Field—R. Birebent. (*Compt. Rend. Acad. Sci.* (Paris) vol. 234, pp. 1135–1136; March 10, 1952.) A straight conductor, carrying an alternating current, is arranged perpendicular to the lines of force of the field to be measured. The mechanical force acting on the conductor is applied to a piezoelectric crystal, and the ac output of the crystal is amplified.

621.317.73.011.21 2556
Admittance Meter for High Frequency—J. Neumann. (*Funk u. Ton*, vol. 6, pp. 113–122; March, 1952.) Description of a Rohde and Schwarz instrument for the range 0.1–10 mc. A substitution method is used, resonance measurements being made on an LCR circuit with and without the element under test connected in parallel, readings being taken on the calibrated variable capacitor used for tuning. Eleven coils are provided for coverage of the total frequency range with adequate overlaps; a diode voltmeter is used as resonance indicator.

621.317.733.029.62 2557
A Bridged-T Impedance Bridge for the V.H.F. Band—R. F. Proctor. (*Proc. IEE*, Part III, vol. 99, p. 105; March, 1952.) Summary only. Description of three bridges designed for the measurement, at 50–100 mc, of complex impedances with resistances of 40–200 Ω and reactances between +200 and −200 Ω. Two of the bridges are suitable for measurement of the impedance between the two live terminals of a 3-terminal network; the third bridge is suitable for measurement of impedances balanced to earth.

621.317.755:621.314.7 2558
Apparatus for Testing Transistors—P. J. W. Jochems and F. H. Stieltjes. (*Philips Tech. Rev.*, vol. 13, pp. 254–265; March, 1952.) Description of equipment for cro display of various transistor characteristics and for direct indication of transducer gain on a meter with decibel scale.

621.317.755:621.397.62 2559
Television Oscilloscope—W. Tusting. (*Wireless World*, vol. 58, pp. 233–236 and 280–283; June and July, 1952.) Television-receiver waveforms to be examined have repetition frequencies of 50 cps–10.125 kc. Design requirements for oscilloscopes are considered in terms of "sag" and rise time, which depend respectively on the low-frequency and high-frequency response of the circuits used; 5 per cent and 1 μs are taken respectively as upper-limit values, using a 5-inch tube. A description, with complete circuit diagram, is given of equipment built to these requirements.

621.317.789:621.396.822 2560
The Design of an Equipment for Measuring Small Radio-Frequency Noise Powers—K. E. Machin, M. Ryle and D. D. Vonberg. (*Proc. IEE*, Part III, vol. 99, pp. 127–134; May, 1952.) The fundamentals of noise-power measurement are considered, the minimum detectable power being determined by the receiver noise and by the ratio of input and output bandwidths of the receiver. Inherent practical difficulties have been overcome by the design of self-balancing equipment in which a locally generated noise power is continuously adjusted to equality with the incoming power. The equipment is described, some sections in detail. The performance is analyzed in terms of its accuracy, its response to an input step function and the fluctuations of its output indication; experimental performance figures compare reasonably well with theory.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.232:621.315.612.4:620.178 2561
New Techniques for Measuring Forces and Wear in Telephone Switching Apparatus—W. P. Mason and S. D. White. (*Bell. Sys. Tech. Jour.*, vol. 31, pp. 469–503; May, 1952.) Very rapid wear tests are made using the $BaTiO_3$ transducer described by Mason and Wick (1818 of 1951) to produce a normal or a tangential wearing force. The force is measured by inserting a $BaTiO_3$ ceramic plate at the point of application of the force and observing on a cro the piezoelectric voltage generated in the plate.

537.533:535.417 2562
Electron Interferometer—L. Marton. *Phys. Rev.*, vol. 85, pp. 1057–1058; March 15, 1952.) Discussion of the basic principles of an interferometer operating with electron beams.

621.3.012.8:629.11.012.8 2563
Application of the Methods of Electromagnetic Analogy to the Study of Motor-Car Suspension Systems—G. Cahen. (*Onde élect*, vol. 32, pp. 89–90; March, 1952.) Comment on article noted in 1060 of May (Lansard).

621.365.55† 2564
Temperature Distribution with Simultaneous Platten and Dielectric Heating—H. M. Nelson. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 79–86; March, 1952.)

621.38.001.8:543/545 2565
Recent Developments in Electronic Instrumentation for Chemical Laboratories—F. Gutmann. (*Jour. Brit. IRE*, vol. 12, pp. 161–180; March, 1952.) The account previously given (710 and 2308 of 1950) is brought up to date. 98 references.

621.384.611.1† 2566
A Coil System for an Air-cored Betatron—R. Latham, M. J. Pentz and M. Blackman. (*Proc. Phys. Soc.*, vol. 65, pp. 89–93; February 1, 1952.) Description of the coil for a small betatron producing 300-kev electrons, and of measurements of the field configuration.

621.384.612.1† 2567
A Fixed-Frequency Cyclotron with 225-cm Pole Diameter—H. Atterling and G. Lindström. (*Nature* (London), vol. 169, pp. 432–434; March 15, 1952.) Designed to produce deuterons of energy 25 mev, this unit is now in operation at the Nobel Institute for Physics, Stockholm. See also 2254 of 1951 (Atterling).

621.384.612.1† 2568
The University of Birmingham Cyclotron—(*Nature* (London), vol. 169, pp. 476–477; March 22, 1952.) Deuterons of 20-mev energy are obtained from this 10.24-mc fixed-frequency unit.

621.385.833 2569
Magnetic Electron-Microscope Projector Lenses—G. Liebmann. (*Proc. Phys. Soc.*, vol. 65, pp. 94–108; February 1, 1952.) Theory previously given [1701 of July (Liebmann and Grad)] is applied to projector and similar lenses. The calculated results are in good agreement with measurements by Ruska.

621.385.833 2570
The Magnetic Electron Microscope Objective Lens of Lowest Chromatic Aberration—G. Liebmann. (*Proc. Phys. Soc.*, vol. 65, pp. 188–192; March 1, 1952.)

621.387.424 2571
Toroidal Geiger Counters—N. G. Trott. (*Jour. Sci. Instr.*, vol. 29, pp. 87–88; March, 1952.) Two new types are described, in one, the cathode consists of a set of parallel wires, in the other of a wire helix.

PROPAGATION OF WAVES

538.566 2572
Universal Wave Polarization Chart for the Magneto-ionic Theory—W. Snyder and R. A. Helliwell. (*Jour. Geophys. Res.*, vol. 57, pp. 73–84; March, 1952.) The expression for complex polarization given by the magneto-ionic theory is plotted on the complex plane, using normalized parameters related to ionization density and collision frequency. A further chart gives ellipticity and tilt angle related to the same normalized parameters. A chart is included for converting the normalized parameters to the corresponding values of ionization density and collision frequency. See also 1471 of 1951 (Scott).

621.396.11 2573
North Pacific Radio Warning Service—(*Tech. Bull. Nat. Bur. Stand.*, vol. 36, pp. 33–34; March, 1952.) Twice-weekly forecasts of propagation conditions for local and long-distance communication circuits are issued by the N.B.S. center at Anchorage, Alaska. 24-hour operation is envisaged.

621.396.11 2574
Comparison of Ionospheric Radio Transmission Forecasts with Practical Results—A. F. Wilkins and C. M. Minnis. (*Proc. IEE*, Part III, vol. 99, pp. 148–154; May, 1952.) Discussion on 2520 of 1951.

621.396.11:523.78 2575
Effect of the Annular Eclipse of March 7, 1951, on Radio-Wave Propagation—(*Nature* (London), vol. 169, pp. 361–362; March 1, 1952.) Summary of paper by L. H. Martin, presented at the N.Z. Geophysical Conference. Observations made at Wellington on 11.75-mc signals from GSD and 10-mc signals from WWVH are reported; measurements were also made of received radio noise on frequencies ranging from 1.4 to 30 mc. Effects due to the eclipse were difficult to segregate because it occurred soon after sunrise and because a magnetic storm of moderate intensity commenced on the same day. The results confirm those of previous observers in indicating that density of ionization and values of critical frequencies for the ionized layers decrease during the eclipse. The noise measurements are not conclusive; greater attention to this aspect should be given during future eclipses.

621.396.11:551.510.52 2576
Average Radio-Ray Refraction in the Lower Atmosphere—M. Schulkin. (PROC. I.R.E., vol. 40, pp. 554–561; May, 1952.) Corrections hitherto applied for atmospheric refraction in the calculation of radio field strengths are reviewed with particular regard to their application in ray-bending computations. A practical scheme is presented for calculating the atmospheric refraction of rf rays numerically from radiosonde data. Ray-bending computations are made for a range of climatological conditions for rays passing entirely through the atmosphere and departing or arriving tangentially at the earth's surface. About 90 per cent of the ray bending occurs in the lowest 10 km of the atmosphere. Uncertainties in the determination of atmospheric refractive indices from meteorological sounding data are discussed.

621.396.11.029.51 2577
The Ionospheric Propagation of Radio Waves with Frequencies near 100 kc/s over Short Distances—K. Weekes and R. Stuart. (*Proc. IEE*, Part III, vol. 99, pp. 99–102; March, 1952.) Summary only. Investigations of downcoming waves, isolated by the use of suitable antenna systems, are described. The transmitters used were those of the British chain of Decca navigation stations, which are all <150 km from the receiving sets in Cambridge. Diurnal and seasonal variations of the conversion coefficient (ratio of amplitude of abnormal component of downcoming wave to that of incident wave) are shown graphically. The average decrease in the apparent height of reflection around sunrise was 7–8 km, but the value was very variable from day to day. The effect of a sudden ionospheric disturbance is to decrease the amplitude of the downcoming wave very greatly and to decrease the apparent height of reflection by an amount which may be as large as 10 km and is the same on all frequencies from 16 to 113 kc.

results of heat treatment on the physical structure. The present physical picture of the alloy is one of single domains of precipitate material in parallel with single domains of matrix material, the observed coercive force being the result of this combination.

538.221 2532
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538.221 2533
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538.221:538.652 2534
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538.221:[621.317.335.2+621.317.411 2535
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538.221:621.318.2 2536
Torque Curves and Other Properties of Permanent-Magnet Alloys—K. Hoselitz and M. McCaig. (*Proc. Phys. Soc.*, vol. 65, pp. 229–235; March 1, 1952.) Further investigations of alcomax III and related alloys. See also 426 of March.

539.23:[546.28+546.57+546.621 2537
Electrical Properties of Very Thin Films of Silver, Aluminium and Silicon—A. Blanc-Lapierre, M. Perrot and J. P. David. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1133–1135; March 10, 1952.) The conductivity at ordinary temperature of films of equivalent thickness <10 mμ, evaporated at a pressure of 10^{-6} mm Hg, was investigated experimentally. Field strengths up to 8 kV/cm were used; large deviations from Ohm's law were observed, together with hysteresis effects in some cases. Results are tabulated and shown as I/V characteristics measured at different times after deposition.

541.18:537.311.35:621.316.87 2538
Polaresistivity and Polaristors—H. E. Hollmann. (Proc. I.R.E., vol. 40, pp. 538–545; May, 1952.) If semiconductive particles are suspended in an insulating fluid, the conductivity of the suspension becomes nonlinear as soon as the fibration at a certain critical value of the applied field reaches the point at which the semiconductive fibres bridge the gap between the electrodes. Particular fluid carriers can be transformed into a solid state, so forming what are termed "polaristors." The current I and voltage V for such resistors are connected by the relation $I=V(1+kV^2)/R_0$, where R_0 is the initial resistance and k a constant for any given material. Solid polaristors are produced by use of thermo-setting or cold-setting plastics. The degree of fibration in the forming process can be checked by means of cro I/V characteristics, or by IF oscillograms similar to modulation trapezoids. Polaristors are very sensitive to temperature and to mechanical stress and may consequently have a wide range of applications. See also 2247 of 1950.

546.23:546.49 2539
The Effect of Mercury on Selenium—H. K. Henisch and E. W. Saker. (*Proc. Phys. Soc.*, vol. 65, pp. 149–154; February 1, 1952.) "The interaction between crystalline selenium and liquid mercury or mercury vapor results in the formation of mercuric selenide. When produced in this form, this material is an excess semiconductor of high conductivity. The associated volume and surface diffusion have been investigated using radioactive ^{203}Hg. Mercury added to selenium before the crystallization process results in material of high resistivity. This is ascribed to a reduction of the positive hole mobility."

546.23.02 2540
A Note on the Structure of Selenium—E. Billig. (*Proc. Phys. Soc.*, vol. 65, pp. 216–221; March 1, 1952.) "The recently established identity of the width of the (optical) energy gap, 2.05 ev, for both amorphous and crystalline selenium, suggests that the chain structure of crystalline selenium is maintained also in the amorphous phase. The low electrical conductivity of amorphous selenium is ascribed to potential barriers existing between individual chains."

546.28+546.289]:541.26 2541
Lattice Parameters, Coefficients of Thermal Expansion, and Atomic Weights of Purest Silicon and Germanium—M. E. Straumanis and E. Z. Aka. (*Jour. Appl. Phys.*, vol. 23, pp. 330–334; March, 1952.)

546.3–74–56+546.3–98–57].02/.03 2542
Electronic Structures Physical Properties in the Alloy Systems Nickel-Copper and Palladium-Silver—B. R. Coles. (*Proc. Phys. Soc.*, vol. 65, pp. 221–229; March 1, 1952.)

547.476.3:548.55 2543
Growth-Rates of Single Crystals of Ethylene Diamine *d*-Tartrate: Flamed Growth and its Inhibition by Boric Acid—A. H. Booth and H. E. Buckley. (*Nature* (London), vol. 169, pp. 367–368; March 1, 1952.)

621.396.611.21 2544
Single Response Thickness-Shear Mode Resonators using Circular Bevelled Plates—R. Bechmann. (*Jour. Sci. Instr.*, vol. 29, pp. 73–76; March, 1952.) The effectiveness of bevelling for removing unwanted resonances is illustrated by response curves for an AT-cut quartz disk of diameter 27 mm. thickness 5.28 mm and frequency 350 kc, taken after applying one, two, three and four bevels respectively. Where there is a clear, single-frequency response, the series-resonance resistance of the crystal is almost constant over a wide temperature range; this affords a practical criterion of quality.

MATHEMATICS

519.24 2545
Properties of Certain Statistics involving the Closest Pair in a Sample of Three Observations—J. Lieblein. (*Bur. Stand. Jour. Res.*, vol. 48, pp. 255–268; March, 1952.) Analysis is presented which indicates that rejection of an outlying high or low reading from a triad not only tends to overestimate considerably the precision of the experimental procedure, but also gives less accurate results.

681.142 2546
A Sampling Analogue Computer—J. Broomall and L. Riebman. (Proc. I.R.E., vol. 40, pp. 568–572; May, 1952.) Description of a computer comprising an algebraic unit and a pulse converter. Only 15 tubes are used. Accuracy is within 1 per cent of full scale.

681.142:621.396.812.3:519.272 2547
A Computer for Correlation Functions—F. E. Brooks, Jr., and H. W. Smith. (*Rev. Sci. Instr.*, vol. 23, pp. 121–126; March, 1952.) Data relating to fluctuating phenomena, such as recordings of radio signals or temperatures, are transcribed in the form of pwm signals on to magnetic tapes. These are run through a computer which computes the difference between the integrals of two products in two separate channels and hence plots the required auto- or cross-correlation function. The apparatus is described and its operation is illustrated by application to the study of atmospheric turbulence.

MEASUREMENTS AND TEST GEAR

621.3.018.41+531.764.5](083.74)(431.55) 2548
Standard Frequencies and Standard Time of the "Funkamt" (FuA), Berlin, January and February 1952—F. Conrad. (*Frequenz*, vol. 6, p. 88; March, 1952.) Two graphs show comparisons between the time markings of the FuA standard-frequency equipment (Rhode and Schwarz quartz clock) with GBR, GIC and WWV signals and with the transmissions from the Hydrographic Institute, Hamburg, and from München-Ismaning.

621.317.3:621.396.611.21 2549
A Simple Method of Measurement of Quartz Crystal Characteristics—J. Coulon. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1269–1271; March 17, 1952.) The equivalent-circuit parameters of the crystal are found from (*a*) the curve for the variation of oscillator frequency with variation of a control capacitance in series with the crystal, (*b*) the curve for the voltage at the oscillator tube grid, (*c*) a capacitance measurement and (*d*) an impedance measurement. The circle diagram introduced previously (2550 below) is used.

621.317.3:621.396.611.21 2550
Plotting the Resonance Curve of a Piezoelectric Quartz Crystal, using a Circle Diagram—J. Coulon. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 1138–1140; March 10, 1952.) The plotting of the resonance curve is facilitated by constructing a circle diagram whose co-ordinates are related to the parameters of the equivalent circuit of the crystal.

621.317.3:621.396.65 2551
Measuring Techniques for Broad-Band, Long-Distance Radio Relay Systems—W. J. Albersheim. (Proc. I.R.E., vol. 40, pp. 548–551; May, 1952.) Methods are outlined for investigation of (*a*) the transient response of the system to step voltages or square waves, (*b*) the frequency characteristics of gain, phase, impedance, and their frequency derivatives, (*c*) the amplitude characteristics of output nonlinearity and of intermodulation products

621.317.31/.32].029.3:537.324 2552
Thermal Converters as A.C.-D.C. Transfer Standards for Current and Voltage Measurements at Audio Frequencies—F. L. Hermach. (*Bur. Stand. Jour. Res.*, vol. 48, pp. 121–128; February, 1952.) An account of thermoelements and associated equipment used at the National Bureau of Standards primarily for the standardization of ac ammeters and voltmeters submitted for test. An accuracy to within 0.01 per cent in the ac-dc transfer can be achieved for currents of 1 ma–50 a and voltages of 0.2–750 v at frequencies from 25 cps to 20 kc. Factors limiting the transfer accuracy are analyzed and approximate solutions are given of nonlinear differential equations governing the heating of a conductor by an electric current.

621.317.324(083.74) 2553
Development of V.H.F. Field-Intensity Standards—F. M. Greene and M. Solow. (Proc. I.R.E., vol. 40, p. 573; May, 1952.) Abstract of U.R.S.I.-I.R.E. Meeting paper, Washington, 1949. See 3090 of 1950.

621.317.336:621.315.212 2554
Reflections in a Coaxial Cable due to Impedance Irregularities—G. Fuchs. (*Proc. IEE*, Part III, vol. 99, pp. 161–164; May, 1952.) Summary only. Theoretical relations between impedance irregularities, input-impedance deviation and pulse echo are derived and experiments are described which confirm the theory.

621.317.42 2555

A New Device for Measuring the Strength of a Magnetic Field—R. Birebent. (*Compt. Rend. Acad. Sci.* (Paris) vol. 234, pp. 1135–1136; March 10, 1952.) A straight conductor, carrying an alternating current, is arranged perpendicular to the lines of force of the field to be measured. The mechanical force acting on the conductor is applied to a piezoelectric crystal, and the ac output of the crystal is amplified.

621.317.73.011.21 2556

Admittance Meter for High Frequency—J. Neumann. (*Funk u. Ton*, vol. 6, pp. 113–122; March, 1952.) Description of a Rohde and Schwarz instrument for the range 0.1–10 mc. A substitution method is used, resonance measurements being made on an LCR circuit with and without the element under test connected in parallel, readings being taken on the calibrated variable capacitor used for tuning. Eleven coils are provided for coverage of the total frequency range with adequate overlaps; a diode voltmeter is used as resonance indicator.

621.317.733.029.62 2557

A Bridged-T Impedance Bridge for the V.H.F. Band—R. F. Proctor. (*Proc. IEE*, Part III, vol. 99, p. 105; March, 1952.) Summary only. Description of three bridges designed for the measurement, at 50–100 mc, of complex impedances with resistances of 40–200 Ω and reactances between +200 and −200 Ω. Two of the bridges are suitable for measurement of the impedance between the two live terminals of a 3-terminal network; the third bridge is suitable for measurement of impedances balanced to earth.

621.317.755:621.314.7 2558

Apparatus for Testing Transistors—P. J. W. Jochems and F. H. Stieltjes. (*Philips Tech. Rev.*, vol. 13, pp. 254–265; March, 1952.) Description of equipment for cro display of various transistor characteristics and for direct indication of transducer gain on a meter with decibel scale.

621.317.755:621.397.62 2559

Television Oscilloscope—W. Tusting. (*Wireless World*, vol. 58, pp. 233–236 and 280–283; June and July, 1952.) Television-receiver waveforms to be examined have repetition frequencies of 50 cps–10.125 kc. Design requirements for oscilloscopes are considered in terms of "sag" and rise time, which depend respectively on the low-frequency and high-frequency response of the circuits used; 5 per cent and 1 μs are taken respectively as upper-limit values, using a 5-inch tube. A description, with complete circuit diagram, is given of equipment built to these requirements.

621.317.789:621.396.822 2560

The Design of an Equipment for Measuring Small Radio-Frequency Noise Powers—K. E. Machin, M. Ryle and D. D. Vonberg. (*Proc. IEE*, Part III, vol. 99, pp. 127–134; May, 1952.) The fundamentals of noise-power measurement are considered, the minimum detectable power being determined by the receiver noise and by the ratio of input and output bandwidths of the receiver. Inherent practical difficulties have been overcome by the design of self-balancing equipment in which a locally generated noise power is continuously adjusted to equality with the incoming power. The equipment is described, some sections in detail. The performance is analyzed in terms of its accuracy, its response to an input step function and the fluctuations of its output indication; experimental performance figures compare reasonably well with theory.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.232:621.315.612.4:620.178 2561

New Techniques for Measuring Forces and Wear in Telephone Switching Apparatus—W. P. Mason and S. D. White. (*Bell. Sys. Tech. Jour.*, vol. 31, pp. 469–503; May, 1952.) Very rapid wear tests are made using the $BaTiO_3$ transducer described by Mason and Wick (1818 of 1951) to produce a normal or a tangential wearing force. The force is measured by inserting a $BaTiO_3$ ceramic plate at the point of application of the force and observing on a cro the piezoelectric voltage generated in the plate.

537.533:535.417 2562

Electron Interferometer—L. Marton. *Phys. Rev.*, vol. 85, pp. 1057–1058; March 15, 1952.) Discussion of the basic principles of an interferometer operating with electron beams.

621.3.012.8:629.11.012.8 2563

Application of the Methods of Electromagnetic Analogy to the Study of Motor-Car Suspension Systems—G. Cahen. (*Onde élect.*, vol. 32, pp. 89–90; March, 1952.) Comment on article noted in 1060 of May (Lansard).

621.365.55† 2564

Temperature Distribution with Simultaneous Platten and Dielectric Heating—H. M. Nelson. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 79–86; March, 1952.)

621.38.001.8:543/545 2565

Recent Developments in Electronic Instrumentation for Chemical Laboratories—F. Gutmann. (*Jour. Brit. IRE*, vol. 12, pp. 161–180; March, 1952.) The account previously given (710 and 2308 of 1950) is brought up to date. 98 references.

621.384.611.1† 2566

A Coil System for an Air-cored Betatron—R. Latham, M. J. Pentz and M. Blackman. (*Proc. Phys. Soc.*, vol. 65, pp. 89–93; February 1, 1952.) Description of the coil for a small betatron producing 300-kev electrons, and of measurements of the field configuration.

621.384.612.1† 2567

A Fixed-Frequency Cyclotron with 225-cm Pole Diameter—H. Atterling and G. Lindström. (*Nature* (London), vol. 169, pp. 432–434; March 15, 1952.) Designed to produce deuterons of energy 25 mev, this unit is now in operation at the Nobel Institute for Physics, Stockholm. See also 2254 of 1951 (Atterling).

621.384.612.1† 2568

The University of Birmingham Cyclotron—(*Nature* (London), vol. 169, pp. 476–477; March 22, 1952.) Deuterons of 20-mev energy are obtained from this 10.24-mc fixed-frequency unit.

621.385.833 2569

Magnetic Electron-Microscope Projector Lenses—G. Liebmann. (*Proc. Phys. Soc.*, vol. 65, pp. 94–108; February 1, 1952.) Theory previously given [1701 of July (Liebmann and Grad)] is applied to projector and similar lenses. The calculated results are in good agreement with measurements by Ruska.

621.385.833 2570

The Magnetic Electron Microscope Objective Lens of Lowest Chromatic Aberration—G. Liebmann. (*Proc. Phys. Soc.*, vol. 65, pp. 188–192; March 1, 1952.)

621.387.424 2571

Toroidal Geiger Counters—N. G. Trott. (*Jour. Sci. Instr.*, vol. 29, pp. 87–88; March, 1952.) Two new types are described, in one, the cathode consists of a set of parallel wires, in the other of a wire helix.

PROPAGATION OF WAVES

538.566 2572

Universal Wave Polarization Chart for the Magneto-ionic Theory—W. Snyder and R. A. Helliwell. (*Jour. Geophys. Res.*, vol. 57, pp. 73–84; March, 1952.) The expression for complex polarization given by the magneto-ionic theory is plotted on the complex plane, using normalized parameters related to ionization density and collision frequency. A further chart gives ellipticity and tilt angle related to the same normalized parameters. A chart is included for converting the normalized parameters to the corresponding values of ionization density and collision frequency. See also 1471 of 1951 (Scott).

621.396.11 2573

North Pacific Radio Warning Service—(*Tech. Bull. Nat. Bur. Stand.*, vol. 36, pp. 33–34; March, 1952.) Twice-weekly forecasts of propagation conditions for local and long-distance communication circuits are issued by the N.B.S. center at Anchorage, Alaska. 24-hour operation is envisaged.

621.396.11 2574

Comparison of Ionospheric Radio Transmission Forecasts with Practical Results—A. F. Wilkins and C. M. Minnis. (*Proc. IEE*, Part III, vol. 99, pp. 148–154; May, 1952.) Discussion on 2520 of 1951.

621.396.11:523.78 2575

Effect of the Annular Eclipse of March 7, 1951, on Radio-Wave Propagation—(*Nature* (London), vol. 169, pp. 361–362; March 1, 1952.) Summary of paper by L. H. Martin, presented at the N.Z. Geophysical Conference. Observations made at Wellington on 11.75-mc signals from GSD and 10-mc signals from WWVH are reported; measurements were also made of received radio noise on frequencies ranging from 1.4 to 30 mc. Effects due to the eclipse were difficult to segregate because it occurred soon after sunrise and because a magnetic storm of moderate intensity commenced on the same day. The results confirm those of previous observers in indicating that density of ionization and values of critical frequencies for the ionized layers decrease during the eclipse. The noise measurements are not conclusive; greater attention to this aspect should be given during future eclipses.

621.396.11:551.510.52 2576

Average Radio-Ray Refraction in the Lower Atmosphere—M. Schulkin. (PROC. I.R.E., vol. 40, pp. 554–561; May, 1952.) Corrections hitherto applied for atmospheric refraction in the calculation of radio field strengths are reviewed with particular regard to their application in ray-bending computations. A practical scheme is presented for calculating the atmospheric refraction of rf rays numerically from radiosonde data. Ray-bending computations are made for a range of climatological conditions for rays passing entirely through the atmosphere and departing or arriving tangentially at the earth's surface. About 90 per cent of the ray bending occurs in the lowest 10 km of the atmosphere. Uncertainties in the determination of atmospheric refractive indices from meteorological sounding data are discussed.

621.396.11.029.51 2577

The Ionospheric Propagation of Radio Waves with Frequencies near 100 kc/s over Short Distances—K. Weekes and R. Stuart. (*Proc. IEE*, Part III, vol. 99, pp. 99–102; March, 1952.) Summary only. Investigations of downcoming waves, isolated by the use of suitable antenna systems, are described. The transmitters used were those of the British chain of Decca navigation stations, which are all <150 km from the receiving sets in Cambridge. Diurnal and seasonal variations of the conversion coefficient (ratio of amplitude of abnormal component of downcoming wave to that of incident wave) are shown graphically. The average decrease in the apparent height of reflection around sunrise was 7–8 km, but the value was very variable from day to day. The effect of a sudden ionospheric disturbance is to decrease the amplitude of the downcoming wave very greatly and to decrease the apparent height of reflection by an amount which may be as large as 10 km and is the same on all frequencies from 16 to 113 kc.

621.396.11.029.51 2578
The Ionospheric Propagation of Radio Waves with Frequencies near 100 kc/s over Distances up to 1000 km—K. Weekes and R. Stuart. (*Proc. IEE*, Part III, vol. 99, pp. 102–105; March, 1952.) Summary only. An account of reception experiments at Swansea, Leeds and Cambridge at distances of about 300 and 900 km from Decca transmitting stations (British and Danish chains). The Hollingworth interference pattern was determined from signal-strength records obtained in an aircraft flying at a height of 2,000 feet. Typical results are shown graphically and discussed. Field-strength minima at 940 km from a 71-kc transmitter are noted and the changes of their times of occurrence throughout a year are shown. Records indicate that during a sudden ionospheric disturbance the amplitude of the downcoming wave is greatly increased. In a large disturbance it was deduced that the reflection coefficient (ratio of amplitude of normal component of downcoming wave to that of incident wave) for 71-kc waves was increased by a factor of nearly 6. On the higher frequencies the increase appeared to be rather smaller.

621.396.11.029.55 2579
Instantaneous Prediction of Radio Transmission Paths—O. G. Villard, Jr, and A. M. Peterson. (*QST*, vol. 36, pp. 11–20; March, 1952.) Report of an investigation of the occurrence of back scatter as a transmission path becomes serviceable. "Scatter-sounding" records made at 5-minute intervals during a 24-hour period are compared with the strength of signals received from some 300 amateur R/T stations during this time in the 14-mc band. Maps show the correlation between scatter areas around the central transmitter and the location of stations heard. The corresponding cro scatter patterns are shown. The results obtained indicate that scatter soundings, which can be made with ordinary amateur transmitting and receiving equipment, show the areas to which radio transmission is possible, via both the *F* and sporadic-*E* layers. As a prediction method for communication, scatter sounding is remarkably sensitive and comparatively simple.

621.396.11.029.62 2580
Abnormal Ranges of Ultrashort Waves and Their Meteorological Causes—B. Abild. (*Tech. Hausmitt. NordwDtsch. Rdfunks.* vol. 4, pp. 4–11; January/February, 1952.) Theory of the effects of variations of temperature and relative humidity, and of inversions of the refractive-index gradient, is briefly reviewed, and a detailed account is given of investigations of the relation between the signal strength at Flensburg of 89.3-mc transmissions from Hamburg and the meteorological conditions over the 150-kb transmission path. In general, a close correlation exists between the signal strength and the difference between the relative humidities at ground level and at 900-mv level. The highest values of signal strength occur when there is a pronounced inversion near the ground. Sudden changes of moisture content at heights of 200–800 m have little effect in producing abnormal ranges, but inversions at heights of about 1 km have a large effect. Low values of signal strength are observed when the lower 2 km of the atmosphere have no pronounced layered structure the strength of the received signal and the gradient of the refractive index at moisture-content discontinuities at heights up to 1.5 km. Other records illustrate the effect of squally conditions, the occurrence of thunderstorms, and the passage of warm fronts.

621.396.11.029.62:551.510.535 2581
A New Kind of Radio Propagation at Very High Frequencies observable over Long Distances—D. K. Bailey, R. Bateman, L. V. Berkner, H. G. Booker, G. F. Montgomery, E. M. Purcell, W. W. Salisbury and J. B. Wiesner. (*Phys. Rev.*, vol. 86, pp. 141–145; April 15, 1952.) The discovery and some results of a preliminary investigation of weak vhf propagation via the ionosphere are reported. Some preliminary speculations suggest that mechanism concerned in such propagation may be scattering caused by ever-present irregularities in the *E* region. An approximate transmission equation is derived in terms of parameters describing inhomogeneities in this region. Experiments carried out on a frequency of 49.8 mc over a test path of 1245 km always showed an observable signal irrespective of season, time of day, or geomagnetic disturbance, although showing dependence of signal strength on these factors and possibly on meteor activity as well. During sudden ionospheric disturbances when hf fade-outs occurred, the signal showed no evidence of weakening and was usually enhanced. For shorter accounts see *Electronics*, vol. 25, pp. 102–103; June, 1952; and *Wireless World*, vol. 58, pp. 273–274; July, 1952.

621.396.8.029.51 2582
Phase Variations with Range of the Ground-Wave Signal from C. W. Transmitters in the 70–130 kc/s Band—A. B. Schneider. (*Jour. Brit. IRE*, vol. 12, pp. 181–194; March, 1952.) Phase-variation/distance curves are plotted for a uniform smooth earth for various values of conductivity, using Norton's (1596 of 1942) and Bremmer's (3242 of 1949) formulas. Confirmation of the curves is provided by phase measurements within the reactive field of a cw radiator and along the base-line extensions of a Decca navigator chain. For the case of an inhomogeneous earth, a simple method of assessing the variation of phase with distance is suggested and is illustrated by an analysis of readings on the Decca navigator system.

621.396.8.029.51 2583
Random Phase Variations of C. W. Signals in the 70–130 kc/s Band—W. T. Sanderson. (*Jour. Brit. IRE.* vol. 12, pp. 195–205; March, 1952.) "The accuracy of cw navigational aids in this frequency band depends fundamentally on the phase stability of the received signals, which in turn is determined by skywave effects. The rms phase errors depend on the relative amplitude of skywave and groundwave, and a method is given for assessing the phase errors in Northern Europe at any given time and range from the transmitter. For the sake of simplicity a number of approximations are made and the resulting limitations are discussed. The predicted and observed errors on the Decca Navigator Chains in England and Denmark are compared, and it is concluded that the method gives sufficiently accurate results for most practical purposes. Some comments are added on the variations of errors with height."

621.396.81.029.62 2584
U.S.W. Field Strength Predictions for Mountainous Terrain—E. Bauermeister and W. Knöpfel. (*Tech. Hausmitt. NordwDtsch. Rdfunks.* vol. 4, pp. 67–73; March/April, 1952.) Field-strength investigations in the Feldberg district of the Black Forest formed the basis of a method for predicting field strengths in the 3-m wavelength region, particularly for places at which multiple refraction and reflection effects are important. The field strength is determined from (*a*) the free-space field strength and (*b*) attenuation factors which take account of the different refractions and reflections that occur. Further measurements in the Moselle valley district confirmed the usefulness of the method.

621.396.81+621.396.677.012.71 2585
Aerial Measurements in the Microwave Range—Seppen. (See 2446.)

621.396.812.029.64 2586
Volume Integration of Scattered Radio Waves—A. H. LaGrone. (PROC. I.R.E., vol. 40, p. 551; May, 1952.) Corrections to paper noted in 1412 of June.

621.396.812.3:519.272:681.142 2587
Computer for Correlation Functions—Brooks and Smith. (See 2547.)

RECEPTION

621.396.621 2588
Linear Rectifiers and Limiters—D. G. Tucker. (*Wireless Eng.*, vol. 29, pp. 128–137; May, 1952.) Theoretical analysis for an applied signal consisting of a carrier (envelope-modulated or unmodulated) accompanied by other tones or noise. The analysis, which is strictly applicable only when the carrier is the predominant input component, is based on a representation of the rectifier or limiter by a switching function whose form is determined by the applied signal and which is expressed in terms of a series of Bessel functions. Consideration is given to the effect on the signal/noise ratio when the detector is used as a frequency multiplier.

621.396.621:621.396.619.13 2589
F.M. Receiver Modification—J. G. Spencer. (*Wireless World*, vol. 58, p. 204; May, 1952.) A note of minor modifications to the receiver previously described (1093 of May) necessary to replace the Type X81 frequency changer, now obsolescent, by a Type-X79 tube. Performance is practically unaffected.

621.396.622:621.396.619.13 2590
Detection of F. M. Waves—Basseras. (*Radio franç.*, no. 3, pp. 21–24; March, 1952.) General description of the method using a limiter and discriminator, with particular reference to the use of an oscillating cricuit as discriminator and to the action of the Travis and Foster-Seeley discriminators.

621.396.621.029.6 2591
Empfangsprobleme im Ultrahochfrequenzgebiet, unter besonderer Berücksichtigung des Halbleiters (U.H.F. Reception Problems, with particular Reference to Semiconductors). [Book Review]—H. F. Mataré. Publishers: R. Oldenbourg, Munich, 1951, 264 pp., 37.80 fr. (*Tech. Mitt. schw.iz. Telegr.-TelephVerw.*, vol. 30, pp. 119–120; March 1, 1952.) A comprehensive treatment, dealing particularly with methods of frequency mixing, amplification, detection, sensitivity and noise, design calculations, and test equipment and methods.

STATIONS AND COMMUNICATION SYSTEMS

621.3.0:5.7(083.7) 2592
Standards on Pulses: Definitions of Terms: Part 2, 1952—(PROC. I.R.E., vol. 40, pp. 552–554; May, 1952.) Standard 52 IRE 20S1. Part 1: see 2826 of 1951.

621.39 2593
Modern Systems of Long-Distance Communication—R. Sueur. (*Bull. Soc. franç. Élect.*, vol. 2, pp. 123–139; March, 1952.) A short review of systems used up to 1940, with a general description of the principles of modern systems using carrier currents, underground balanced-pair or coaxial cables, submarine cables, or radio beam systems for the transmission of telephony, telegraphy, broadcasting or television signals. The economics of the various systems is discussed and basic considerations which should determine future developments are outlined.

621.39.001.11 2594
A Comparison of Signalling Alphabets—E. N. Gilbert. (*Bell. Sys. Tech. Jour*, vol. 31, pp. 504–522; May, 1952.) Two channels are considered, (*a*) a discrete channel transmitting sequences of binary digits, and (*b*) a continuous low-pass channel. The rate of signalling is computed for a large number of simple alphabets; rates near the channel capacity cannot be attained without using complicated alphabets.

621.39.001.11 2595
A Link between Information and Energy—J. H. Felker. (PROC. I.R.E., vol. 40, pp. 728–729; June, 1952.) Discussion of the ultimate limits of low-power operation of switching systems, computers, and other communication machinery. Calculation indicates that for the passage of 10^6 "bits" of information per second the power required is $>2.85\times10^{-15}$ w.

621.395.44 2596
Transmission Characteristics of the V60 Carrier-Frequency Equipment—F. Ring. (*Fernmeldetech. Z.*, vol. 5, pp. 101–108 and 179–186; March and April, 1952.) C.C.I.F. requirements for a 60-channel system are outlined and the specification for the V60 equipment is described. Tests carried out on the Frankfurt-Mannheim cable, with repeaters at Erzhausen, Hahn and Lorsch, gave results in good agreement with theory.

621.395.65:621.318.572 2597
An Experimental Electronically Controlled Automatic Switching System—W. A. Malthaner and H. E. Vaughan. (*Bell. Sys. Tech. Jour.*, vol. 31, pp. 443–468; May, 1952.) An experimental telephone switching system is described; its use enables the number of control and connector circuits to be reduced.

621.395.97:621.395.645 2598
Radio-Diffusion Amplifiers for Standard [telephone] Circuits—Jacot. (See 2468.)

621.396:061.3 2599
Extraordinary Radio-Communication Administrative Conference, Geneva, 1951 (C.A.E.R.)—H. Pressler. (*Fernmeldetech. Z.*, vol. 5, pp. 132–136; March, 1952.) Report of the proceedings.

621.396.018.78:621.396.677 2600
Signal Distortion by Directional Broadcast Antennas—C. H. Moulton. (PROC. I.R.E., vol. 40, pp. 595–600; May, 1952.) Measurements on transmissions from two broadcasting stations with directive antenna systems show that signal distortion occurs and is a function of the direction of reception. Distortion results from changes produced by the directive antenna system in the magnitudes or relative phases of the signal components; it is accentuated by the existence of deep nulls in the radiation pattern, by small antenna band-width, by high audio modulation frequency, and by high degree of modulation.

621.396.619.13:621.396.66 2601
An Aural Monitor for Frequency Modulation—J. L. Hathaway and R. E. Lafferty. (PROC. I.R.E., vol. 40, pp. 545–547; May, 1952.) The aural monitor at a FM station should be responsive to AM to a degree at least equivalent to that of ordinary receivers with little or no limiting. This requirement, as well as others, can be met by proper application and adjustment of a "slope detector." Two units, which are compact and have been found reliable, are described. The simpler unit is suitable for TV sound on one of the lower vhf channels; the other, a heterodyne type, can be used for any TV channel.

621.396.619.13.018.78:621.392.43 2602
Effect of Aerial-Feeder Mismatch on the Distortion in Frequency-Modulation Transmission—E. Kettel. (*Telefunken Ztg.*, vol. 25, pp. 41–50; March, 1952.) Analysis shows that mismatch causes both amplitude and phase distortion owing to reflection effects. For matching conditions attainable in practice it is sufficient to consider only the strongest of such reflections. In the case of receivers using amplitude limiting, only distortion of the instantaneous frequency results from mismatch; this gives a distortion coefficient which increases with the modulation frequency but with normal matching and feeder lengths remains within tolerable limits for broadcast transmissions. For simpler receivers without limiters, using signal-flank demodulation, considerably greater distortion is produced by the AM due to mismatch.

In the case of a carrier-frequency multichannel telephony system, the phase distortion due to antenna mismatch produces nonlinear crosstalk between the individual channels; calculations indicate that in many cases this can only be reduced to a negligible amount by limitation of the feeder length.

621.396.619.16 2603
Delta Modulation, a New Modulation System for Telecommunication—J. F. Schouten, F. de Jager and J. A. Greefkes. (*Philips tech. Rev.*, vol. 13, pp. 237–245; March, 1952.) A form of pulse modulation is described in which the signals transmitted are quantized both in time and in amplitude, so that cumulative interference picked up in the transmission channel may be eliminated at the receiver. The transmitter uses a form of inverse feedback such that any signal transmitted is only a correction of the preceding one; the system thus approximates to the optimum in which no superfluous information is transmitted. Freedom from interference is secured at the expense of bandwidth, but the system described utilizes bandwidth more efficiently than any other in existence except pcm systems, which need very much more complicated apparatus. See also 2330 of September (Libois).

621.396.619.16:621.394/.396 2604
Use of Pulse Technique in the Establishment of Complex Transmission Networks—G. Potier. (*Onde élect.*, vol. 32, pp. 197–201; April/May, 1952.) Discussion of multiplex time-sharing transmission methods. Such systems are technically and economically advantageous and their flexibility renders their application practicable for both simple and complex transmission networks.

621.396.619.16:621.396.4 2605
Use of Pulse Modulation for Transmission in a Group of Telephony Channels of a Carrier-Current System—L. J. Libois. (*Onde élect.*, vol. 32, pp. 190–196; April/May, 1952.) Distortion in pulse modulation due to the sidebands of the modulation spectrum and to the limitation of the transmission channel bandwidth is discussed, and also the nonlinearity of the modulation characteristic. Comparison of pulse-modulation multiplex with time sharing and with frequency sharing is made with regard to the signal/noise ratio for the two methods, which from this point of view are found to be practically equivalent. See also 2331 of September.

621.396.619.16:621.396.41:621.396.822.1 2606
Crosstalk in Time-Division-Multiplex Communication Systems using Pulse-Position and Pulse-Length Modulation—J. E. Flood. (*Proc. IEE*, Part III, vol. 99, pp. 106–108; March, 1952.) Summary only. Some of the results previously obtained for pam systems [2830 of 1951 (Flood and Tillman)] are extended to pulse-position and pulse-length modulation systems. Sets of curves show the crosstalk as dependent on the number of stages of the resistance-coupled amplifier used and on the time constant common to all stages.

621.396.65 2607
Some Data on Two Former Multichannel Beam Links from Athens to Rome and Crete—K. O. Schmidt. (*Telefunken Ztg.*, vol. 25, pp. 64–68; March, 1952.) A few details are given of antennas, relay stations and equipment, installed in 1941, operating on wavelengths in the range 70–80 cm and providing three telephony and twelve voice-frequency telegraphy channels.

621.396.65:621.396.41:621.396.822 2608
Evaluation of the Signal/Noise Ratio for a Multiplex Radio Link—J. Dascotte. (*Onde élect.*, vol. 32, pp. 202–208; April/May, 1952.) A formula is given for the signal/noise ratio which involves the peak power of the carrier wave, the gain of the transmitting and of the receiving antenna relative to a $\lambda/2$ dipole, the losses in the two feeders, the theoretical free-space attenuation between two $\lambda/2$ dipoles for the transmission path and frequency, the noise factor of the receiver, and losses due to the terrain and the heterogeneity of the atmosphere. Use is made of abacs given by Bullington (802 of 1948 and 436 of 1951). Estimates are made of losses due to the terrain for paths within and beyond the optical range.

621.396.65:[621.396.43+621.397.26 2609
The Equipment of the Paris-Lille Link—H. Gutton, J. Fagot and J. Hugon. (*Onde élect.*, vol. 32, pp. 174–180; April/May, 1952.) Discussion of technical considerations which determined the broad lines of development of the C.G.T.S.F. system, operating on 8-cm wavelength, and description of antenna switching arrangements, relays, antennas, filters, and uhf amplifiers using traveling-wave tubes. See also 2028 of August (Marzin).

621.396.65:621.396.5 2610
The Contribution of Radio Beam Links in Field of Telecommunications—R. Cabessa. (*Onde élect.*, vol. 32, pp. 131–151; April/May, 1952.) Radio and line systems are compared from the point of view of transmission quality, noise characteristics, secrecy and cost. Quality requirements for complete circuits, as specified by the C.C.I.F., are correlated with the characteristics of the radio equipment and propagation medium as defined by the C.C.I.R. Short descriptions are given of existing installations in Europe and in the U.S.A.

621.396.65:621.396.5 2611
Radio Beam Links in Modern Telephone Networks—R. Sueur and L. J. Libois. (*Onde élect.*, vol. 32, pp. 121–130; April/May, 1952.) General discussion of the characteristics of beam links and their application for short-distance and long-distance telephone communication and for extension of line systems. A few details are given of the equipment for beam links recently brought into service in France; short-distance links use multiplex pulse modulation, while FM is used for long-distance operation.

621.396.65:621.396.5 2612
Inauguration of the Dijon-Strasbourg Radio Beam Telephony Link—M. Lorach. (*Électronique*) (Paris), no. 64, pp. 6–8, 45; March, 1952.) A short illustrated description of the equipment, which comprises two underground cables linking Dijon with the radio transmitter on Mont Affrique, 7 km away, and beam links thence to Strasbourg, with relay stations on Montfaucon, near Besançon, and on the Ballon de Guebwiller. Wavelengths on the three sections for transmissions from Dijon are respectively 1.25, 1.09 and 1.25 m; in the opposite direction the wavelengths are 1.13, 1.20 and 1.13 m, the changes at the relay stations being made to isolate the incoming from the outgoing signals.

621.396.65:621.396.5 2613
The Equipment of the Dijon-Strasbourg Radio Beam Link—P. Rivère and M. Schwindenhammer. (*Onde élect.*, vol. 32, pp. 163–173; April/May, 1952.) Detailed description of terminal and relay-station equipment, with simplified circuit diagrams. See also 2612 above.

621.396.65.029.62 2614
A 50-Mc/s Beam Link with ±500-kc/s Frequency Swing—H. J. Fründt. (*Telefunken Ztg.*, vol. 25, pp. 51–59; March, 1952.) General description of equipment, including a quartz-controlled 1-kw transmitter and selective receiver, providing multichannel communication between Western Berlin and Torfhaus, in the

Harz mountains. Results are given of measurements of transmitter and receiver distortion, and noise factor.

621.396.712(43-16) 2615
The Technical Installations of the N.W.D.R., 1st January 1952—(*Tech. Hausmitt. Nordw-Dtsch. Rdfunks*, vol. 4, pp. 21–25; January/February, 1952.) A list of the medium-wave and FM usw stations, giving height above sea level, frequency, power, type of transmitter and type of antenna, a few details of the experimental television transmitter and of the monitoring stations at Wittsmoor, Hamburg and Norderney, and particulars of the various broadcasting studios, including size of rooms and their mean reverberation times.

621.396.933 2616
Service Range for Air-to-Ground and Air-to-Air Communications at Frequencies above 50 mc/s—R. S. Kirby, J. W. Herbstreit and K. A. Norton. (PROC. I.R.E., vol. 40, pp. 525–536; May, 1952.) Propagation aspects of communication with aircraft are discussed and contours of equal received signal strength are shown in the form of lobes for various frequencies. For systems with equivalent transmitted power, ground-antenna height, and transmitting- and receiving-antenna gain, the service range decreases with increase of frequency. This is due primarily to decrease of the absorbing area of the receiving antenna and to a larger number of nulls in the lobe structure owing to interference between direct and ground-reflected waves. Ground-station antenna-height diversity and tilted-array ground-antenna systems are discussed as a means of obtaining improved coverage at the higher frequencies.

621.396.97+621.396.932/.933].1 2617
Common-Wave Broadcasting and Hyperbolic Navigation—Pohentsch (See 2515.).

621.396.97:621.316.729 2618
The Radio Common-Wave System of the S.W.F. [Südwestfunk]—A. Kolarz, E. Kniel and K. H. Baer. (*Tech. Hausmitt. NordwDtsch. Rdfunks*. vol. 4, pp. 47–51; March/April, 1952.) In order to improve the frequency stability of the Bad Dürrheim, Ravensburg and Reutlingen transmitters, a radio method of synchronization was developed, utilizing the 200-kc transmissions from Droitwich as a frequency standard. After demodulation and amplification the signals are used to lock the frequency of a 100-kc quartz oscillator from which is derived, by a process of frequency multiplication, division and mixing a frequency of 1,538 kc which locks the frequency of a quartz oscillator used to control the broadcasting transmissions. An independent 1,538-kc quartz oscillator is available for use when the Droitwich transmitter is not operating, or when fading or other causes render the synchronization unreliable.

SUBSIDIARY APPARATUS

621-526 2619
Stability of Control Systems. Methods of Study—J. Kuntzmann, J. Daniel, and Min-Yuan Ma. (*Rev. gen. Elect.*, vol. 61, pp. 149–152; March, 1952.) Two methods of examining the stability of a system are discussed, the method of response curves and one depending on zone separation in the complex plane. The first method is convenient when the effect of only one parameter is to be determined; the second is preferable when two or more parameters are concerned.

621-526 2620
Amplifying Dynamos. Their Use in Servo-mechanisms—G. Lehmann. (*Onde élect.*, vol. 32, pp. 78–88; March, 1952.)

621.314.632.1+621.314.634].011.22 2621
Measurement of the D.C. Resistance of Selenium and Copper Oxide Rectifiers below Room Temperature—D. M. Grimes and S. Legvold. (*Jour. Appl. Phys.*, vol. 23, pp. 312–315; March, 1952.) Results for temperatures (T) from 79 degrees to 300 degrees K show that the logarithm of the resistance is nearly a linear function of 1/T in all cases except for the reverse direction in Se rectifiers, for which the resistance (with 6v applied) has a minimum value below 200 degrees K.

621.314.634 2622
Some Further Observations on the Effect of Bending on Selenium Rectifier Discs—P. Selényi. (*Proc. Phys. Soc.*, vol. 65, pp. 161–162; February 1, 1952. See also 86 of 1948.)

621.316.721 2623
General Theory of Current Stabilizers—J. J. Gilvarry and D. F. Rutland. (*Rev. Sci. Instr.*, vol. 23, pp. 111–114; March, 1952.) The theory of voltage stabilizers previously developed (814 of April) is extended to deal with current stabilizers; the performance of an arbitrary stabilizer is completely specified by four parameters. The effect on stabilizer current of load-resistance variation is evaluated. Two special cases are discussed.

621.351/.355 2624
Modern Batteries and Accumulators used in Telecommunications—J. Pernik. (*Ann. Télécommun.*, vol. 7, pp. 145–148; March, 1952.) Complementary to paper noted in 3120 of 1951. Construction and discharge characteristics of different cells are discussed, including primary cells with air depolarization and Ni-Fe and Ni-Cd alkaline accumulaters.

TELEVISION

621.397.242:621.395.51 2625
The London-Birmingham Television-Cable System—T. Kilvington, F. J. M. Laver and H. Stanesby. (*Proc. IEE*, Part I, vol. 99, pp. 44–58; March, 1952. Discussion, pp. 59–62.) The cable itself has been described by Stanesby and Weston (1279 and 1857 of 1949). An account is here given of the general arrangements providing channels for relaying television signals simultaneously in both directions between central points in the two cities, with extensions to the transmitters at Alexandra Palace and Sutton Coldfield. Details are given of the line amplifiers, which have uniform gain over the working-frequency range 3–7 mc, the cable loss being equalized over the same range. Supervisory and power equipment, and terminal modulation and demodulation equipment, are described, with typical test results showing modulator and demodulator performance on 10-μs pulses and sawtooth signals, over-all vision-frequency characteristics, and test patterns before and after transmission from Alexandra Palace to Birmingham and back.

621.397.26+621.396.43]:621.396.65 2626
The Equipment of the Paris-Lille Radio Link—Gutton, Fagot, and Hugon. (See 2609.)

621.397.26:621.396.65 2627
The Paris-Lille Television Radio Link—Y. Angel and P. Riche. (*Onde élect.*, vol. 32, pp. 152–157; April/May, 1952.) Historical account of developments leading to the establishment of the link and general description of its technical features, with ground-contour diagrams for the three sections of the 218-km path. See also 819 of April.

621.397.26:621.396.65 2628
The Equipment of the Paris-Lille Television Radio Link—J. Laplume, S. Schirman, R. Fraticelli and R. Jeannin. (*Onde élect.*, vol. 32, pp. 158–162; April/May, 1952.) Detailed description of terminal and relay-station equipment developed by the C.F.T.H. and operating on frequencies of 940 and 905 mc. See also 819 of April.

621.397.5 2629
The British Contribution to Television—(*Engineering*, (London), vol. 173, pp. 553 and 603; May 2, and 9, 1952.) *Engineer* (London), vol. 193, pp. 607–608; May 2, 1952.) Report of convention arranged by the Institution of Electrical Engineers in April/May 1952, with summaries of some of the papers presented. These are to be published in full in four special issues of *Proc. IEE*, Part IIIA. (See also *Wireless World*, vol. 58, p. 212, June, 1952; and *Nature* (London), vol. 170, pp. 136–138; July 26, 1952.)

621.397.5 2630
Television-Image Reproduction by use of Velocity-Modulation Principles—M. A. Honnell and M. D. Prince. (PROC. I.R.E., vol. 40, p. 604; May, 1952.) Reply of one of the authors to comment by Thomas (823 of April).

621.397.5(083.74):621.397.335 2631
Variant of the Frame Synchronizing Sequence of the C.C.I.R. 625-Line Standard—H. Laett. (*Tech. Mitt. schweiz. Telegr.-Teleph-Verw.*, vol. 30, pp. 87–90; March 1, 1952. In French and German.) At the plenary session of the C.C.I.R., June 1951, the synchronization signal adopted consisted of three groups of six pulses each. A proposal by the Swiss delegation for the groups to have five pulses was not discussed, owing to lack of time, but has now been approved by correspondence and will be included in the final report of the proceedings. Discussion indicates that while the six-pulse group is very suitable for the American 525-line 60-fields/second system, it is not suitable for a 625-line 50-fields/second system if the synchronization signal is to be derived directly from the mother frequency. The five-pulse signal is shown graphically. The two variants are practically equivalent as regards receiver operation.

621.397.611.2 2632
Improvements in Design and Operation of Image-Iconoscope Type Camera Tubes—J. E. Cope, L. W. Germany and R. Theile. (*Jour. Brit. IRE*, vol. 12, pp. 139–149; March, 1952.) Television camera tubes using high-velocity electrons for scanning are liable to spurious signals and edge flare due to nonuniform redistribution of secondary electrons over the storage surface. A description is given of an improved photicon image iconoscope in which these undesirable effects are reduced by flooding the storage surface with low-velocity electrons from an adjacent annular photo-emission surface. A simplified camera control unit can be used since the "black" picture signal is constant in relation to the level during beam suppression; a simple clamp circuit provides a satisfactory average-brightness component.

621.397.62 2633
Television Developments—a Selection from the I.E.E. Convention of Points Applicable to Receivers—*Wireless World*, vol. 58, pp. 210–211; June, 1952.)

621.397.62 2634
Line Eliminator. "Spot Stretching" as an Alternative to Spot Wobbling—G. N. Patchett. (*Wireless World*, vol. 58, pp. 219–212; June, 1952.) Line visibility in a television receiver is reduced, without impairing horizontal definition, by using an elongated spot, the longer axis being vertical. The required auxiliary focusing device is most conveniently placed between the main focusing device and the deflector coils. Photographs of B.B.C. Test Card C illustrate improvements obtained.

621.397.62 2635
A New U.H.F. Television Converter—H. Hesse. (*Tele-Tech.* vol. 11, pp. 36–39. 118; March, 1952.) Description of a self-contained uhf-vhf conversion unit continuously tunable over the 470–890-band.

621.397.62:621.317.755 2636
Television Oscilloscope—Tusting. (See 2559.)

621.397.621 2637

Workshop Construction of a Scanning Assembly for a Wide-Angle Flat-Faced Tube—M. Duchaussoy and M. Guillaume. (*Télévision*, nos. 22 and 23, pp. 75–79 and 104–108; March/April and May, 1952.) Detailed instructions for snapping the yoke, winding the deflection coils and mounting the whole assembly complete with focusing coil and deflection circuits.

621.397.621:621.385 2638

Line-Scanning Valves and Circuits—B. Eastwood and C. C. Vodden. (*Jour. Brit. IRE*, vol. 12, pp. 150–160; March, 1952.) In high-efficiency line-scanning circuits the tubes have to withstand peak voltages amounting to some thousands of volts during the cut-off part of the cycle, the special design problems of mass-production tubes for use in these circuits are discussed. In pentodes and tetrodes thermionic emission from the screen grid is reduced by suitable processing and by aligning with the control grid. Insulation requirements are examined.

621.397.645 2639

A Study of Grounded-Grid, Ultra-High-Frequency Amplifiers—T. Murakami. (*RCA Rev.*, vol. 12, pp. 682–701; December, 1951.) Though the high cost of uhf amplifiers may prevent their general adoption in domestic television receivers, they are needed for particular applications. The performances of amplifiers using several different types of tube are compared, theoretical curves are given for amplifier gain and for the noise factor of amplifiers with matched or mismatched input or output circuits. Experimentally determined noise factors of amplifiers using types 5876 and 416A triodes are in good agreement with computed values. The use of a grounded-grid amplifier reduces the amount of local oscillator signal passed back to the antenna; experimental curves are given showing the attenuation from output to input of an amplifier using (*a*) a type 5876 triode, and (*b*) a type 416A triode.

621.397.7 2640

Kirk o'Shotts Television Transmitting Station—(*Engineer* (London), vol. 193, pp. 371–373; March 14, 1952.) Further details, including descriptions of the medium-power equipment, the coaxial-cable links between Alexandra Palace, Birmingham and Manchester, and the microwave radio link between Manchester and Edinburgh. See 3143 of 1951.

621.397.82 2641

Interference in Television Pictures. Effect of Line Deflection Circuits—G. Diemer, Z. van Gelder and J. J. P. Valeton. (*Wireless Eng.*, vol. 29, pp. 164–168; June, 1952.) Barkhausen oscillations are discounted as a cause of hf interference effects giving rise to vertical lines on the left-hand side of a television screen. These are ascribed to irregularities above the knee of the I_a/V_a characteristic of the power tube used in the line-deflection circuit. The most important factor influencing the intensity of the interference is the leakage inductance of the transformer.

621.397.82 2642

Relative Magnitudes of Undesired Responses in Ultra-High-Frequency Receivers—Wen Yuan Pan. (*RCA Rev.*, vol. 12, pp. 660–681; December, 1951.) The relative strengths of interfering signals in uhf television receivers using crystal mixers are measured with equipment basically similar to commercially available tuners and converters (see 442 of 1951), except that no selective circuit is connected in front of the mixer. Cases are grouped into single-frequency and two-frequency interference, either dependent on or independent of local-oscillator frequency. The test results are applied in designing the pre-mixer circuits to give required selectivity.

621.397.5 2643

Fernsehen (Television). [Book Review]—F. Kerkhof and W. Werner. Publishers: Philips, Eindhoven, 1951, 506 pp. 28 DM. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, p. 27; January/February, 1952.) An introduction to the physical and technical principles of television technique. Advanced mathematical treatments are avoided as far as possible.

TRANSMISSION

621.396.61:621.396.712 2644

The Ravensburg 2×20-kW Broadcasting Transmitter—A. Kolarz, A. Schweisthal and K. H. Baer. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, pp. 34–42; March/April, 1952.) Special precautions were taken in the development of a 20-kw transmitter using Doherty modulation, to obtain high quality without sacrificing simplicity of construction. The performance obtained was found superior to that of an anode-modulation transmitter of the newest type, the superiority being more evident the higher the quality demanded in the reproduction.

621.396.61:621.396.712 2645

Twin Drive as Active Research for Broadcasting Transmitters—A. Schweisthal. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, pp. 42–45; March/April, 1952.) Discussion of the advantages of an arrangement in which two identical transmitters are connected via a bridge network to a single antenna. In case of failure of either transmitter, the other continues to transmit alone until the fault has been corrected. The arrangement is considered more economical than that in which a stand-by transmitter is provided and seldom used.

621.396.61.029.62:621.396.712 2646

An U.S.W. Frequency Converter—A. Kolarz and B. Pick. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, pp. 62–66; March/April, 1952.) Description of relay station equipment for the frequency range 87–100 mc. A single quartz oscillator is used in conjunction with a local oscillator and three mixer stages. The temperature coefficient of the quartz crystal is so low that a thermostat is unnecessary, its frequency is equal to the frequency difference required between input and output, and the mixing arrangements are such that frequency changes of the local oscillator affect only the IF and have no effect on the output frequency.

621.396.615.14:621.396.619.13 2647

New Master Oscillator of High Frequency Stability for F.M. U.S.W. Broadcasting Transmitters—E. Kettel. (*Telefunken Ztg.*, vol. 25, pp. 60–64; March, 1952.) General description, with outline circuit diagram, of equipment for the 87.7–100 mc range, comprising oscillator and modulator with quartz-crystal control unit giving frequency constancy to within 2 parts in 10^6.

621.396.712:621.396.61:621.398 2648

Problems of Automatic Operation of Transmitter Groups—A. Kolarz and E. Kniel. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, pp. 59–62; March/April, 1952.) The S.W.F. (Südwestfunk) caters for three districts with different types of population and culture. Problems of program switching for the various mf, hf and uhf transmitter groups are discussed and present facilities are described. Monitoring of the transmission quality of all transmitters is carried out by means of a suitably equipped motor van, normally stationed in Baden-Baden.

TUBES AND THERMIONICS

537.533/.534:621.385.2 2649

Ions and Electrons with Uniform Initial Velocities in a Vacuum—F. Wenzl. (*Z. angew. Phys.*, vol. 4, pp. 94–104; March, 1952.) Theoretical analysis for a planar diode system. A solution of the space-charge equation is obtained in terms of elliptic integrals. For the case of a periodic potential distribution, the relation between period and amplitude is investigated; the potential distribution can be represented with fair accuracy by trigonometrical approximations if the amplitudes are not too large. The occurrence of a virtual cathode is investigated for the case of anode emission of ions with zero initial velocity; close qualitative analogies are found with the case of pure electron flow. Questions of stability and the realizability of a periodic potential distribution are discussed.

537.582.004.15 2650

The Efficiency of Thermal Electron Emission—M. J. O. Strutt. (Proc. I.R.E., vol. 40, pp. 601–603; May, 1952. *Bull. schweiz. elektrotech. Ver.*, vol. 43, pp. 350–353; May 3, 1952. In German.) The efficiency is defined as the ratio of the total kinetic energy of the emitted electrons to the heater power applied to a cathode. Assuming losses are solely due to heat radiation, an expression for the optimum efficiency is derived. This efficiency can never be reached with ordinary cathodes. The efficiency determined from experimental results for a tungsten cathode was 0.18 per cent, and for an oxide layer on a Pt-Ir base 3.5 per cent. Comparison with the the theory suggests that higher efficiencies might be obtained with cathodes having a higher index n in the expression c_nT^n for the heat radiation per unit surface area, T being the absolute temperature and c_n a constant depending on the nature of the surface.

621.314.7 2651

Present Status of Transistor Development—J. A. Morton. (*Bell Sys. Tech. Jour.*, vol. 31, pp. 411–442, May, 1952.) An account is given of progress in improving the reproducibility, reliability and performance of transistors. Types discussed include point-contact and *n-p-n*-junction transistors and phototransistors; characteristics are indicated by families of curves. Compared with thermionic tubes, transistors are equal as regards reproducibility and superior as regards length of life and mechanical strength, but inferior as regards temperature effects, the upper limit of operation being 70–80 degrees C, though the effect of this restriction is reduced because power consumption, and hence heating, is often low. As regards miniaturization, transistors are enormously superior. As regards applications, transistors can be considered seriously for pulse systems up to repetition rates of 1–2 mc and for cw transmission at frequencies < 1 mc; for the range 1–100 mc transistors should be considered only where very great importance is attached to smallness and reliability.

621.314.7 2652

A Method of Improving the Electrical and Mechanical Stability of Point-Contact Transistors—B. N. Slade. (*RCA Rev.*, vol. 12, pp. 651–659; December, 1951.) Point-contact transistors embedded in resin suffer practically no change of electrical characteristics when subjected to severe impact and centrifuge tests; they are also largely unaffected by storage at extreme temperatures and by high humidity. Operation at low temperatures is satisfactory, but some changes of electrical characteristics are observed when operating at high ambient temperatures.

621.314.7:546.817.221 2653

Double-Surface Lead-Sulphide Transistor—P. C. Banbury. (*Proc. Phys. Soc.*, vol. 65, p. 236; March 1, 1952.) The transistor voltage gain is found to increase with decreasing crystal thickness for both the type-A and the coaxial electrode configuration.

621.383:546.817.221 2654

Lead Sulphide Rectifier Photocells—A. F. Gibson. (*Proc. Phys. Soc.*, vol. 65, pp. 214–216; March 1, 1952.) The properties of single-crystal cells at 290 degrees, 195 degrees and 90 degrees K, are compared with those of similar PbTe cells (2655 below).

621.383:546.817.241 2655

Single-Contact Lead Telluride Photocells—A. F. Gibson. (*Proc. Phys. Soc.*, vol. 65, pp. 196–214; March 1, 1952.) Of single-crystal cells studied at 90 degrees K, only those of *p*-type show good rectification and marked photo-effects. Photosensitivity and time constant are determined by the total current at the contact point.

621.383.5:546.289 2656

A Large-Area Germanium Photocell—J. I. Pantchechnikoff. (*Rev. Sci. Instr.*, vol. 23, p. 135; March, 1952.) Brief details are given of a cell consisting basically of a Ge disk with a semitransparent conducting metal film deposited on one face. Advantages of the arrangement as compared with the point-contact type cell are indicated and some characteristic data are given for a cell with a gold film on *n*-type Ge.

621.384.5:621.318.572 2657

The Development of a Multi-Cathode Decade Gas-Tube Counter—G. H. Hough. (*Proc. IEE*, Part III, vol. 99, pp. 166–167; May, 1952.) Summary only. As pulses are applied to the counter the glow discharge progresses round the array of ten cathodes, making one complete rotation for every ten pulses. Unidirectional progression is achieved by a special cathode producing asymmetrical priming which reduces the breakdown potential of the preferred adjacent transfer electrode. The tube operates with a supply voltage of 330±20 v, developing at least 40 v across the 15-kΩ cathode resistors and counting aperiodically over the pulse repetition range of 0–20,000/second.

621.385.029.64:168.2 2658

A Symbolism for Microwave-Valve Classification—G. M. Clarke. (*Proc. IEE*, Part III, vol. 99, pp. 98–99; March, 1952.) Summary only, giving an outline of a proposed system.

621.385.032.213 2659

The Plasmatron, a Continuously Controllable Gas-Discharge Developmental Tube—E.O. Johnson and W.M. Webster. (Proc. I.R.E., vol. 40, pp. 645–659; June, 1952.) Full description of the construction and properties of a new type of tube using an independently generated gas-discharge plasma as a conductor between a hot cathode and an anode. See also 2869 of 1951 (Johnson).

621.385.032.213.1.027.5/.6 2660

Use of Thoriated-Tungsten Filaments in High-Power Transmitting Tubes—R. B. Ayer. (Proc. I.R.E., vol. 40, pp. 591–594; May, 1952.) An account of the development and use of Th-W filaments in transmitting tubes such as the RCA-5671, RCA-5770 and RCA-5771, operating with anode voltages>5 kv. Typical self-supported multistrand Th-W filament assemblies are illustrated. Initial tests of type-RCA207 tubes fitted with Th-W filaments showed that grid currents were very sensitive to filament input, best performance being obtained at 517 w instead of the usual 1,144 kw for w filaments, a reduction of 55 per cent. Oxygen-free high-conductivity Cu is used for the anodes of the above-mentioned tubes, and Pt-coated Mo wires for the grids. Tube life in service has proved very satisfactory.

621.385.032.216 2661

Correlation of D.C. and Microsecond Pulsed Emission from Oxide Coated Cathodes—F. A. Horak. (*Jour. Appl. Phys.*, vol. 23, pp. 346–349; March, 1952.) Ba-Sr oxide cathodes were prepared on base metals of pure Ni, Ni with 0.2 per cent and with 4 per cent Si, and Ni with 4.7 per cent W. The results of emission measurements at intervals during zero-emission life tests show that the base metal affects the change of emission with time. The emission from cathodes on the W-Ni base remained nearly constant and was much higher than that for any of the other base metals.

621.385.032.216 2662

The Emission from Oxide-Coated Cathodes in an Accelerating Field—D. A. Wright and J. Woods. *Proc. Phys Soc.*, vol. 65, pp. 134–148; February 1, 1952. "A theoretical treatment based on the existence of a space-charge zone immediately inside the coating. The charge in this zone is shown to vary with applied field and with current density, and with certain coating parameters. The variation in the charge leads to a variation in work function, and thereby to a dependence of emission on field strength, which is to be combined with the normal Schottky effect."

621.385.032.216:539.433.2 2663

Loss of Thermionic Emission in Oxide-Coated Cathode Tubes due to Mechanical Shock—D. O. Holland, I. E. Levy and H. J. Davis. (Proc. I.R.E., vol. 40, pp. 587–590; May, 1952.) Experimental results indicate that while many factors are involved in the effects of shock on cathode activity, evolution of gas from mica spacers is possibly the most important cause of emission reduction after shock. Increase of the number of mica spacers results in greater emission losses; some grades of mica are worse than others in this respect.

621.385.3 2664

Transit Time Oscillations in Triodes—O. H. Critchley and M. R. Gavin. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 92–94; March, 1952.) Parasitic oscillations observed in disk-seal triodes over the wavelength range 6-20 cm are found to depend critically on the value of anode voltage and to occur at frequencies such that the cathode-grid transit time is about five-fourths of the oscillation period. The oscillations are similar to those previously observed by Llewellyn and Bowen in diodes (3155 of 1939).

621.385.83 2665

Axially Symmetric Electron-Beam and Magnetic-Field Systems—L. A. Harris. (Proc. I.R.E., vol. 40, pp. 700–708; June, 1952.) Theory is presented for long high-density beams in axial magnetic fields. Radial oscillations about an equilibrium radius are found to be always stable in the presence of a magnetic field, and can be made stable even without the field. Design formulas are given for two types of cathode, one having the emitting surface in a uniform magnetic field and giving a solid beam, the other having the emitting surface inside a magnetic screen and giving a tubular beam. Limited experimental results confirm most of the theory, and indicate the possibility of focusing without a magnetic field along the whole length of the beam.

621.385.831 2666

Space-Charge Waves in an Accelerated Electron Stream for Amplification of Microwave Signals—Ping King Tien and L. M. Field. (Proc. I.R.E., vol. 40, pp. 688–695; June, 1952.) Exact solutions are given for an idealized electron stream of infinite extent, accelerated or retarded uniformly through a space where dc space-charge effects are assumed to be neutralized by positive ions. The solution indicates that space-charge waves on a retarded stream grow in amplitude and can thus be used for amplifying microwave signals. The theory is relevant to the amplifying tubes discussed previously [2068 of 1951 (Field et al.)]. Three amplifiers of this type have been constructed and are described. Measured values of gain at frequencies of about 3 kmcs are in good agreement with calculated values.

621.385.832 2667

A Novel Type of Monoscope—S. T. Smith. (Proc. I.R.E., vol. 40, pp. 666–668; June, 1952.) Description of a cr tube with a rotationally symmetrical cuspidal target of polished Al shaped so that the variation with beam deflection, of the secondary-emission current received by a conical collector corresponds to the variation of a received radar signal with angular displacement of the radar antenna.

621.385.832 2668

Fundamental Processes in Charge-Controlled Storage Tubes—B. Kazan and M. Knoll. (*RCA Rev.*, vol. 12, pp. 702–753; December, 1951.) A comprehensive analysis is presented of the equilibrium potentials of insulated elements exposed to electron bombardment and to the action of light. The influence of the distribution of secondary-electron velocities is examined. Signal writing, reading and erasing processes are described and the abilities of the different methods to deal with half tones are discussed. Definitions are given of the terms used. 97 references, many of them annotated.

621.396.615.141.2 2669

Theory of the Magnetron Amplifier—F. Lüdi. (*Z. angew. Math. Phys.*, vol. 3, pp. 119–128; March 15, 1952. In German.) Analytical treatment deriving an expression for the amplification factor. In contrast to the case of the conventional traveling-wave tube, amplification is possible when the electron velocity is considerably lower than the velocity of the traveling field. This is of particular interest for extremely short wavelengths.

621.396.615.141.2:621.316.727 2670

R. F. Phase Control in Pulsed Magnetrons—E. E. David, Jr. (Proc. I.R.E., vol. 40, pp. 669–685; June, 1952.) Discussion of magnetron oscillations started in the presence of an external rf exciting signal whose frequency is not greatly different from the steady-state frequency of the magnetron. Two methods of analysis are presented. In the first, quasi-steady-state starting is assumed. Solutions of the corresponding differential equation specify the phase of the oscillations as a function of the time interval after starting. In the second method, the oscillator is represented as a parallel RLC circuit shunted by a negative nonlinear conductance. Approximate solutions of the inhomogeneous van der Pol equation for this system are used to investigate the frequency and phase transients during starting, and also the distortion of the build-up envelope by the exciting signal. The initial conditions are in both cases established in terms of the ratio of exciting signal to preoscillation noise. The results of the two methods of analysis are essentially in agreement.

621.396.615.141.2(083.7) 2671

Standards on Magnetrons: Definitions of Terms, 1952—(Proc. I.R.E., vol. 40, pp. 562–563; May, 1952.) Standard 52 IRE 7S1.

MISCELLANEOUS

621.3.015.7(083.71/.72) 2672

Technical Vocabulary—(*Onde élect.*, vol. 32, pp. 113–114; March, 1952.) A list, prepared by the Vocabulary Commission of the S.N.I.R., of general terms relating to pulse technique, with definitions and in some cases English equivalents.

621.396.6+621.397.6+621.385 2673

New Radio Components in the World Market—M. Alixant. (*Radio tech. Dig., Edn. franç.*, vol. 6, nos. 1–3, pp. 3–62, 89–108 and 139–149; 1952.) Classified review listing the characteristics of new circuit components, loudspeakers, microphones, sound recording apparatus, television equipment and tubes.

621.396 2674

Advances in Electronics, Vol. 3. [Book Review]—L. Marton (Ed.). Publishers: Academic Press, New York, N. Y., 357 pp., $7.50. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 335–336; November, 1951.)