Proceedings of the I·R·E

Ref Stee

January 1944

FEBRUARY, 1944

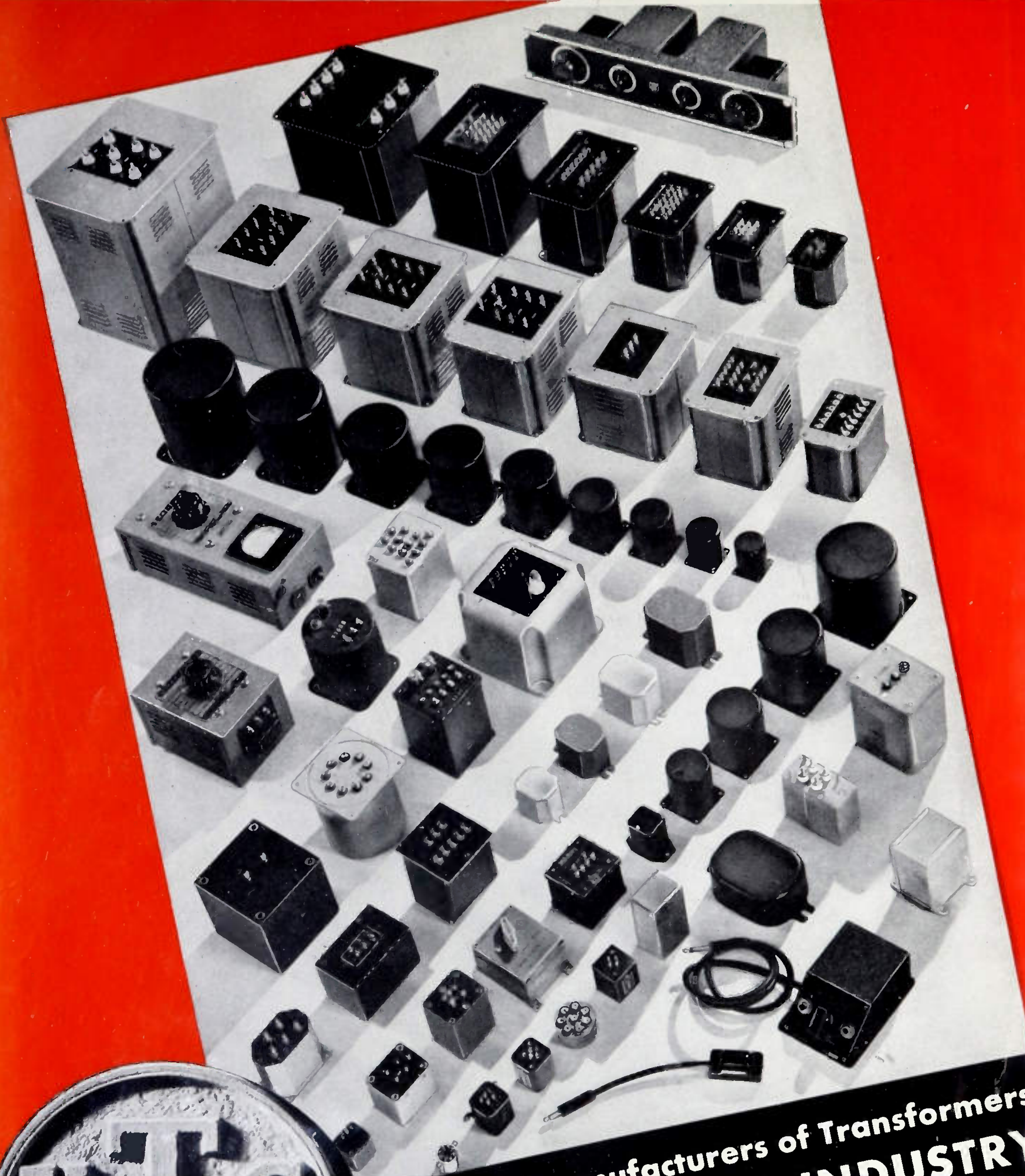
VOLUME 32 NUMBER 2

Electro-Voice Manufacturing Company, Inc.

LIP MICROPHONE

Clear speech transmission despite battle conditions

Institute of Radio Engineers



Proceedings *of the* I·R·E

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 32 — ***February, 1944*** — NUMBER 2

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

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What's Wrong With This Picture?

ARMY-NAVY "E" WITH STAR
Awarded All Four Divisions of Raytheon for continued excellence in production

RAYTHEON

RAYTHEON MANUFACTURING COMPANY
Waltham and Newton, Massachusetts

The thing that is wrong about this picture is that radio engineers have been doing such a bang-up job meeting and anticipating the vast needs of our military services that not enough good things can be said about them by those engaged in the field of electronics.

Seven days a week and night after night, the radio engineers are working out the multitudinous problems of design required to give our Allies the most of the best electronic equipment in the world.

Raytheon is proud of its part in furnishing electronic tubes and equipment that meet engineering requirements of stamina, high quality and complete dependability under the most severe wartime demands.

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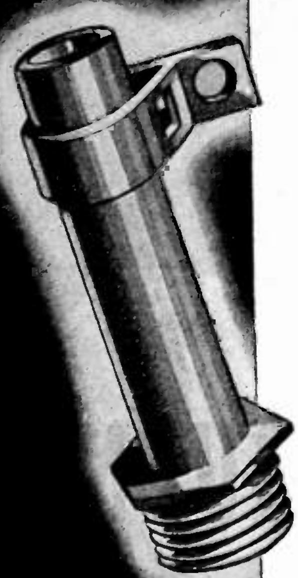
Centralab offers

Type 817-001 55MMF ± 10%
Neg. Temp. Coefficient
—.00052 MMF/MMF °C
Test voltage is 2000 V. D. C.
working voltage 1000 V. D. C.

Type 817-002
Mechanically as above
Capacitance 15 MMF ± 20%
Sketch is TWICE actual size.



Type 814-078 300 MMF ± 10%
Neg. Temp. Coefficient
— .00075 MMF/MMF °C
Test voltage is 1400 V. D. C.
working voltage 500 V. D. C.
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BUSHING MOUNTED CAPACITORS
for special applications

Both types are used in high frequency circuits where a capacity ground to the chassis and a "lead through" is desired.

The ceramic capacitor tube is plated internally and externally with silver and then with copper. The tube is snug fit in the brass bushing and the external capacitor plate is soldered to the bushing.

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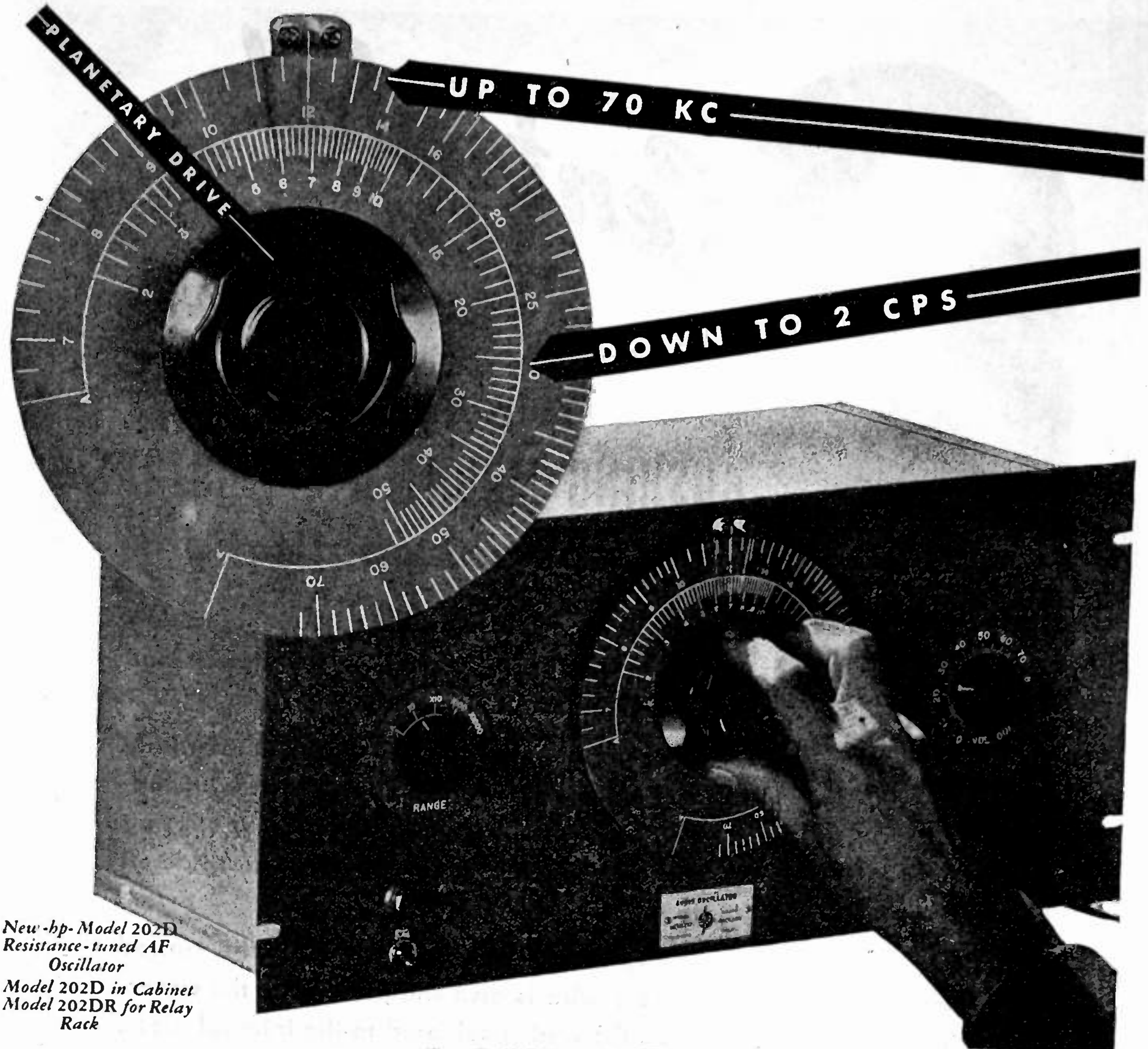

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SENSITIVITY 3 TIMES THAT OF RCA-931

SIGNALS are amplified up to 200,000 times and more in this new RCA multiplier phototube—over three times the amplification possible with the famous RCA-931—because manufacturing techniques have been improved materially by RCA engineers.

This really amazing sensitivity is made possible by the skillful use of secondary emission as cathode electrons are impelled against 9 successive dynodes before they reach the plate. At each dynode, secondary electrons are produced to multiply the electron current enormously.

Because this high amplification is accomplished within the phototube itself, extremely low light levels will produce high outputs without the high-gain amplifier stages required with conventional phototubes.

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HIGH SENSITIVITY. The 931-A operated at 100 volts per stage has a sensitivity of 2 amperes per lumen; or over 3 times that of the superseded 931 at the same voltage per stage.

CIRCUIT SIMPLICITY. Where light signals are very small and high gain is needed, the 931-A provides a simpler circuit than that for a conventional phototube and its accompanying high-gain amplifier stages; also when the 931-A is used as a d-c amplifier, its zero-reading has excellent stability, and there is no problem of circuit feedback.

COLOR SENSITIVITY. The 931-A—like the 931—employs the S-4 photosurface which is highly sensitive to blue light; peak sensitivity is at about 3750 A°. Frequency response is flat up to a limit determined by transit-time effects, well above 10 million cycles per second. Dark current is very low. Full details are available on request (see coupon).

APPLICATIONS. A typical application of the 931-A is in quantitative spectrographic analysis. The 931-A and its associated circuits are substituted for the photographic plate commonly used in such analyses. This method is speedy, and results can be observed with excellent accuracy. This method of spectrographic analysis is widely used in vitamin measurements.

Cross section of the 931-A, showing electron paths in red.

RCA application engineers will be glad to help you apply the RCA-931-A—or other RCA electron tubes—to the solution of your design problems. Write, outlining your problem, to Commercial Engineering Section, RCA, 586 South Fifth Street, Harrison, N. J.

TECHNICAL DATA. Nine multiplier stages. Cathode photosurface, S-4. Max. seated height, 3⅛". Max. diam., 1-5/16". Base, small shell submagnal 11-pin. Mounts in any position.

MAXIMUM RATINGS (Absolute values): Plate volts (d-c or peak a-c), 1250. Volts between dynode No. 9 and anode, 250. Plate current, 2.5 milliamperes. Plate dissipation. 0.5 watt.

CHARACTERISTICS:

Volts per stage	75	100	Volts
Luminous sensitivity	0.3	2.0	µAmp./µLumen
Current amplification	30,000	200,000	
Sensitivity at 3750 Angstroms	270	1800	Amp./Watt

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FREE!

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☐ 4-page technical data sheet giving description, ratings, characteristics, typical circuit diagrams, performance curves, and typical circuits for the RCA-931-A.

☐ 16-page booklet entitled "RCA Phototubes," chock-full of valuable tube data, application notes, etc., for RCA phototubes.

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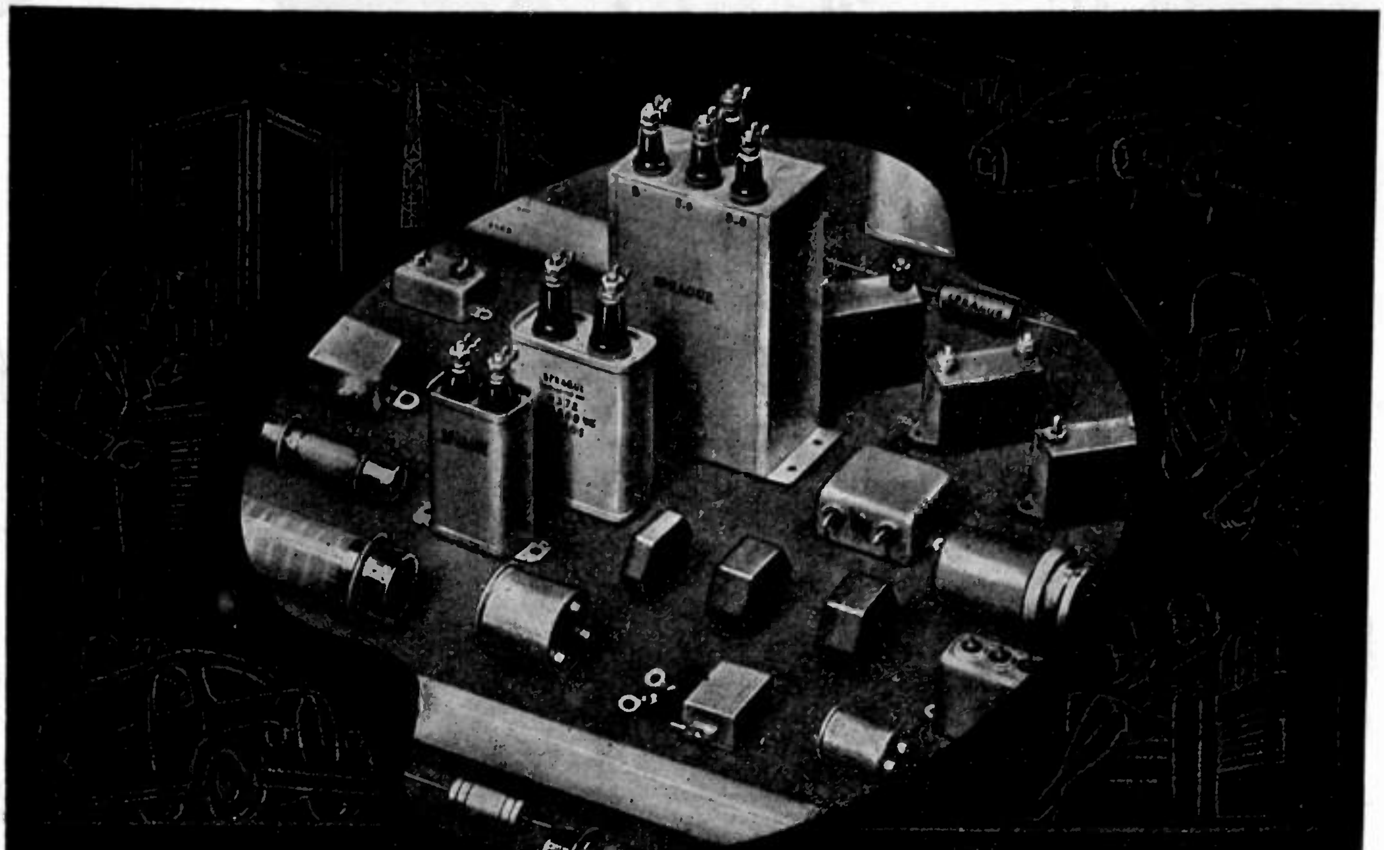


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Rolling all over the world. Hitting the enemy where it hurts him the most, covering infantry, scouting, fighting. Fighting and talking. Talking by radio to coordinate all in a pattern of Victory.
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Our mechanized Army must have brains, but brawn still counts. The big fellow wrestling interminably with 155 millimeter shells serves his greedy howitzer with the broad back developed by endless months of bone-tiring drill.

If it cannot take the jolts, vibrations, concussions, and extreme atmospheric variations of mechanized global war, the best electronic fighting equipment in the world is useless. Hearts of this combat equipment — electronic tubes — have two strikes against them from the start. Inherently delicate and fragile by nature, still they must be as rugged as the men who depend upon them.

Bump, vibration, immersion, life, and other punishing tests prove the mettle of Hytron tubes before they leave the factory. More important still, results of these tests form the basis for continual improvements in construction and processing. Throughout manufacture — in stem, mount, sealing-in, exhaust, aging, basing, and test departments — engineers, foremen, and skilled operators are ceaselessly striving to achieve in Hytron tubes not only the tops in electronic performance, but also the peak of dependable stamina which combat demands.

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Electronic's engineers have specialized for years in the technique of vibrator power supplies. They have conducted the most extensive research ever known on power supply circuits. They have extended the practical application of vibrator-type power supplies far beyond previous conceptions.

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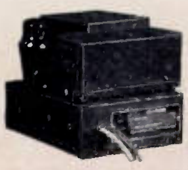

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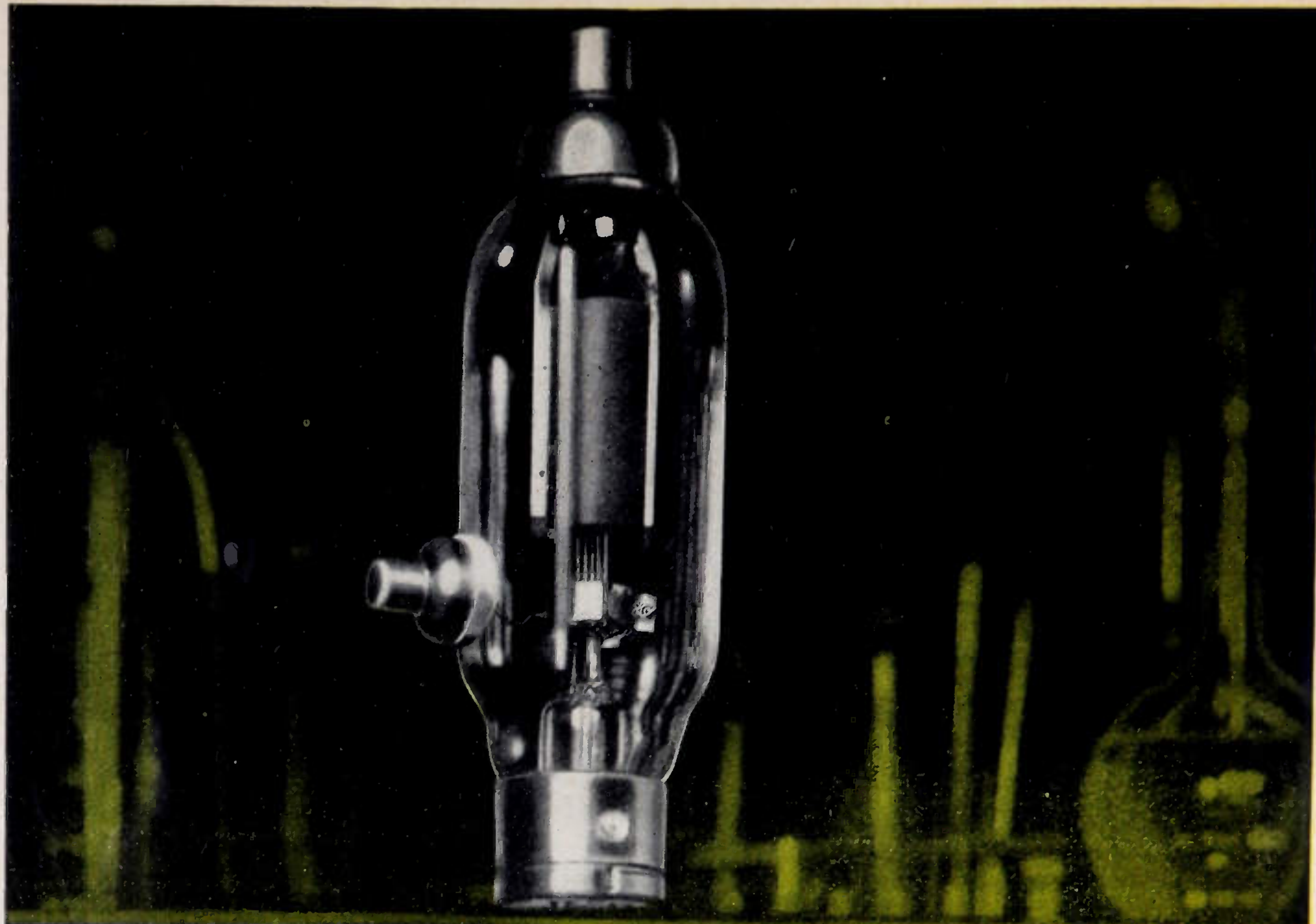

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The mechanism that operates this control passes through the main rotor hub.

It is the heart of the helicopter.

And you will find this heart fastened safely and securely with Elastic Stop Nuts.

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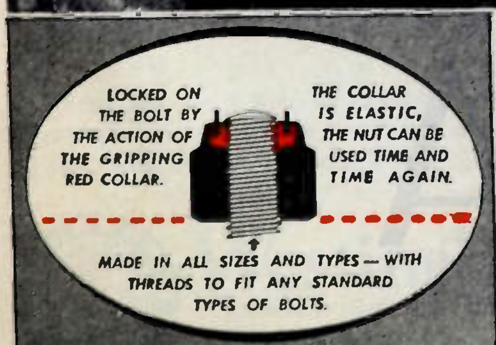

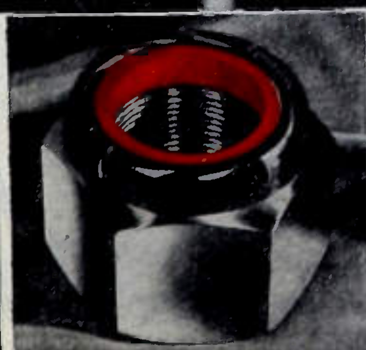

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"Somebody better hurry up with a stick of chewing gum before our left wing leaves us!"

* * *

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It is to these superlative standards that Rola builds equipment for the Army-Navy Air Forces . . . highly specialized transformers and coils, supersensitive headphones, and other electronic parts having to do with Communications. And it is to these same standards that Rola will build its after-the-war products, whatever they may be. The ROLA COMPANY, Inc., 2530 Superior Avenue, Cleveland 14, Ohio.

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Here again the specialized skill of DuMont engineers and craftsmen is in evidence. While adhering rigidly to standard specifications governing this popular type, DuMont has introduced its own refinements, improvements and double-check inspection for more rugged, longer-lasting cathode-ray tubes.

Always remember, when it bears the DuMont seal it is a product of *the pioneer in the commercialized cathode-ray tube field.*

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5" electrostatic deflection and focus tube. Intensifier feature for maximum deflection sensitivity and brilliance.

Choice of four fluorescent screens: Green Medium (5CP1); Green Long (5CP2); White Medium (5CP4); Blue Short (5CP5).

Bulged envelope for greater mechanical strength. Tube base design provides adequate insulation between electrode leads for high-altitude installations.

Heater Voltage 6.3. Intensifier Electrode Potential 4400 v. max. Focusing Electrode Potential 1100 v. max. Accelerating Electrode Potential 2200 v. max.

Deflection Factor: D_1D_2, 36.5 d.c. volts/kv inch, plus-minus 20%; D_3D_4, 32.0 d.c. volts/kv inch, plus-minus 20%.

Spot is to be positioned within a 25 mm. square, with Eb_2=2000 v., under RMA standard test conditions.

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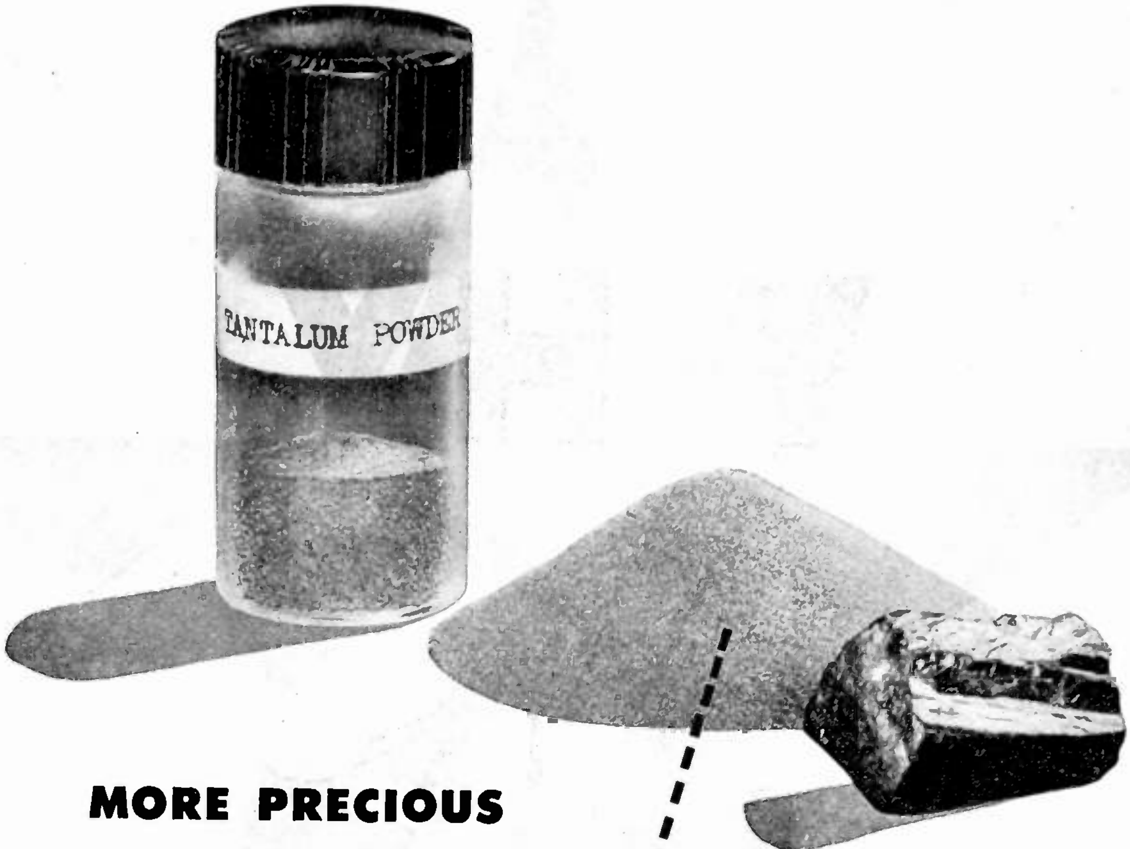

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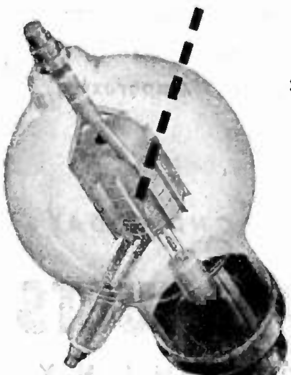
One of the reasons you will find so many Gammatrons in use where dependability is essential, is that all Gammatrons have tantalum plates and grids. They can and do take heavy overloads safely—punishment which would cause any other type of tube to cease functioning.



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But another resource is now available. A new source of wealth and well-being has been developing gradually and almost unnoticed which is tremendously important today and of still greater importance for tomorrow.

This new resource is the research laboratory.

Today, in hundreds of industrial and college laboratories, trained minds are expanding the world's knowledge, and applying the results of research to industry and to war.

In the Bell System, research has always been a fundamental activity.

The telephone was invented in a research laboratory. And for years Bell Telephone Laboratories has been the largest industrial laboratory in the world.

Underlying modern research is the realization of vast latent values in nature. Although the lone genius does from time to time bring to light some part of these hidden values, only organized scientific research can assure the thorough exploration that will render the full measure of use for human welfare.

Research means imagining and experimenting. It means the searching out and bringing together of facts. It means clear statements of problems, precise measurements and keen analysis. It means tenacious following along unexpected paths.

These are the procedures of research. Its consummation is the grasping by subtle minds of relationships in nature no one has previously known. And on the basis of the broader knowledge so established are built new materials, new methods and new structures to serve the people of America.

The Bell Telephone Laboratories has now concentrated its efforts on communication systems and equipment for the armed forces. When the war is over its researches in communication will again be applied to an ever-improving telephone service in America.

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HALLICRAFTERS WAS READY!
Under the abnormal climatic and operating conditions of war, the Signal Corps SCR-299 communications truck, built by Hallicrafters, is providing peak performance for the Allied armed forces, fighting throughout the world.
Hallicrafters peacetime communications equipment is meeting the wartime qualifications and demands of the Military!
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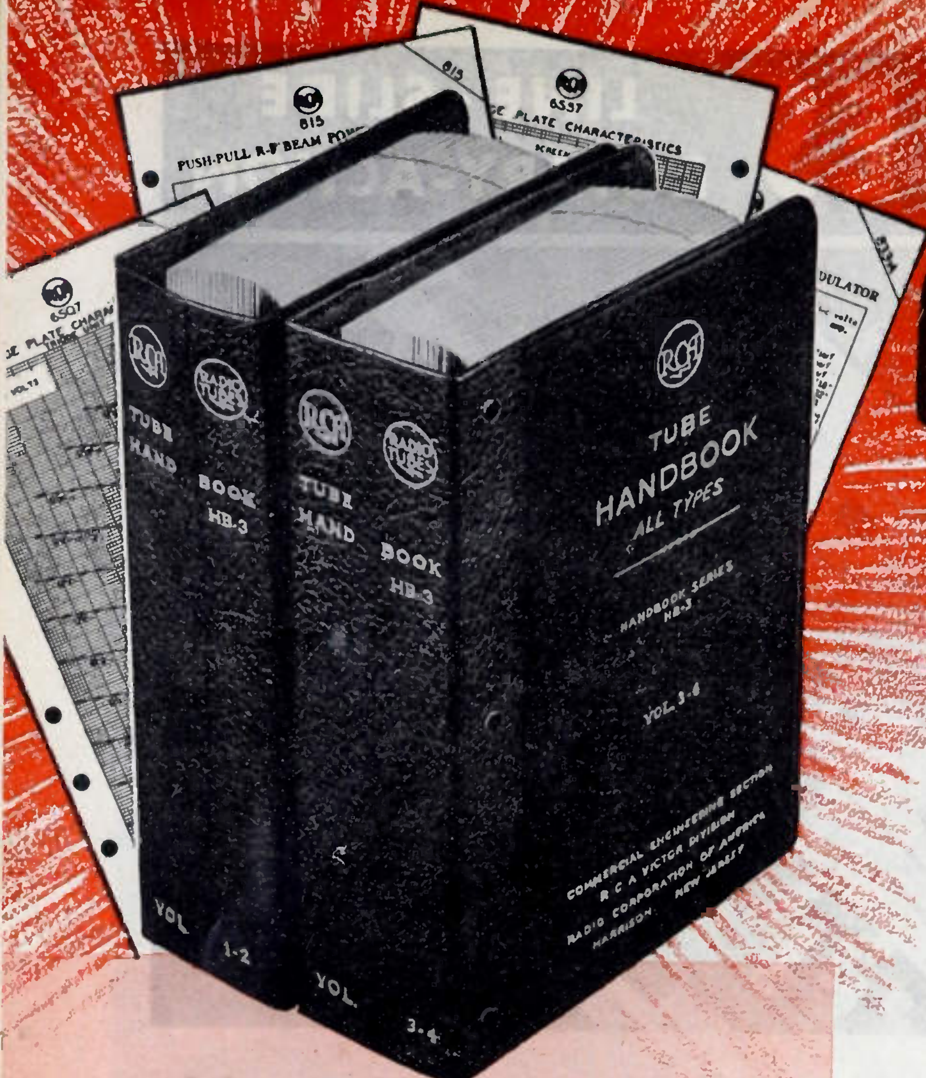
BUY MORE BONDS
World's largest exclusive manufacturer of short wave radio communications equipment...First exclusive radio manufacturer to win the Army-Navy Production Award for the third time.


These are OUR weapons



How to keep posted on Tubes!

WITH the RCA TUBE HANDBOOK on your desk, you have available at all times, and in concise, handy form, the most up-to-date technical data for each tube in RCA's comprehensive tube line...plus a wealth of detailed information such as the Army-Navy Preferred Tube list; definitions of general terms; tube ratings and their significance; types of cathodes and their use; outline-, base-, and cap-drawings; classification charts of receiving, power, and cathode-ray tubes; resistance-coupled amplifier chart; list prices; etc.

The $6.00 subscription price* brings you not only the two 7⅜" x 5" binders finished in black fabricoid and imprinted in gold, complete with data sheets, but *also* additional sheets containing new and/or revised data as issued during the year! Annual service fee thereafter, $1.50.

Fill out (or copy) Subscription Order Form shown below, and mail it with your check today. This is the *only* way for you to make sure of having the same completely up-to-the-minute and authoritative information about RCA tubes . . . exactly when and where you want it...that thousands of other electronics engineers are using every day. RADIO CORPORATION OF AMERICA, *Commercial Engineering Section, Harrison, New Jersey.*

**This price applies only in the United States and its possessions. For other countries, subscription price will be furnished on request. Orders will be accepted subject to our ability to make deliveries.*

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Vice-President, I.R.E.

Since the turn of the century and the birth of the radio art, much progress has been recorded.

Prior to the war great strides had been made in the radio art by the pioneers of wireless and radio communication, including of course broadcasting and such allied services as frequency modulation and television.

The advances have been made possible by the combined efforts of the research workers, scientists, and physicists in the field of pure science, and the engineers who, in the field of applied science, have been responsible for the application and the large-scale production of radio equipment for the benefit of mankind.

The radio industry can be justly proud that many of the developments and the production of radio communication and electronic equipment play an important role, if not a deciding factor, in the ultimate success of the United Nations war effort.

The physicists and the engineers whose joint efforts have made possible the use of radio-and-electronic equipment for war purposes must in the postwar period use their knowledge and skill to create a better world by making available radio-and-electronic devices to serve the peoples of the world in the fields of safety, public service, education, and entertainment.

To achieve these hoped-for results there must be complete co-operation between the various groups representing Government, the industry, and the public.

The members of The Institute of Radio Engineers have played an important part in the development of the radio art. It is their duty and responsibility to take an active role in shaping the world of tomorrow by becoming more active in the work of the Institute, in the general advancement of the engineering profession, in business management, as well as in the fields of government and political influence which affect or control the art, profession, or industry in which we are so vitally interested.

The value of engineering services in managerial and industrial positions is becoming more widely recognized. Engineers in the past have not taken a sufficiently vocal part in public affairs and therefore the qualities of leadership residing within the profession have definitely been neglected.

Let us plan for the future with confidence realizing that we have a duty and a responsibility in assuring full postwar employment and the wider distribution of modern, well-designed, high-quality products.

Ralph A. Hackbusch

1944 Vice-President, I.R.E.

Ralph A. Hackbusch was born in Hamilton, Ontario, Canada, on September 18, 1900. He was educated at Hamilton Collegiate Institute and later specialized in electrical engineering. For thirteen years Mr. Hackbusch was with the Canadian Westinghouse Company, Ltd., before becoming affiliated with the Canadian Brandes Company (Kolster Radio). In 1930 he became Chief Engineer and Factory Manager of the Stromberg-Carlson Telephone Manufacturing Company at Toronto and in 1940 he became its Vice-President and General Manager. That same year he was requisitioned by the Canadian Government and placed in charge of the Radio Division of the government-controlled Research Enterprises, Ltd., at Toronto, later being elected its Vice-President in Charge of Radio and Director of the Radio Division. He has since resigned this post and has returned to Stromberg-Carlson at Toronto as its Managing Director.

Mr. Hackbusch was a member of the Canadian Electrical Code Committee, the main committee of the Canadian Engineering Standards Association, the Radio Manufacturers Association General Standards Committee, chairman of the Toronto Section of the Institute of Radio Engineers in 1933, a member of the Sections Committee in 1936, Board of Directors in 1940. He was Director of Engineering for the RMA of Canada for ten years and guest member of the Joint Coordination Committee of the Edison Electric Institute, the National Electrical Manufacturers Association, and the Radio Manufacturers Association. He has for many years been a guest member of the General Standards Committee of the Radio Manufacturers Association of the United States.

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Mr. Hackbusch was elected an Associate member of the Institute of Radio Engineers in 1926, transferred to Member grade in 1930, and to Fellow grade in 1937. He was recently elected Vice-President of the Institute of Radio Engineers for the year 1944.

Electronic Tin Fusion*

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Summary—**Application of electronic methods to the fusion of electrolytic tin plate is an important recent technological advance in the steel industry. It has played a prominent part in the conservation of tin, whereby tin plate is now produced using considerably less tin than heretofore. Handling operations have been greatly reduced, and the rate of production of tin plate speeded up to 500 feet per minute with some installations being made in anticipation that in the near future these tin lines may be operated at 1000 feet per minute. One tin mill alone has 3730 kilowatts installed capacity of electronic high-frequency generators operating at 200 kilocycles, which is comparable to the total installed power of all the broadcast stations in the United States. This pioneering example of high-frequency electronic equipment of substantial power-handling capacity at work in industry undoubtedly is the forerunner of many processes where high-frequency heating has as yet been unthought of but where, if applied, would produce a superior product, speed up operation, and effect operating economies.**

PRIOR to December 7, 1941, steel strip was being tin-plated for the canning and food packing industries by the well-known "hot-dip" process. Thickness of this tin coating was such that about $1\frac{1}{2}$ pounds of tin were required for each 100 pounds of tin plate. The tin plate was processed as individual sheets rather than as a continuous strip. Steel mills were considering a new electrolytic process which would permit thinner tin coatings at greatly increased speed of production as well as producing the tin plate as a continuous strip. However, electrolytically deposited tin has to be heated to its melting temperature and "flowed" to obtain improved corrosion resistance and a satisfactory method for accomplishing this at high strip speeds did not exist. After Pearl Harbor, need to conserve tin and to increase production of tin plate for the canning industry so accelerated investigation of suitable methods for flowing electrolytic tin that research work which would normally require two years was accomplished in less than six months with the result that today continuous electrolytic tinning lines in several steel mills are in production using between one third and two thirds of the amount of tin which would have been required by the old hot-dip method, and the tin is "flowed" with power from electronic-type generators of 200-kilowatt output capacity.

To understand the contribution made by the electronic flowing process, it will be of interest first to consider briefly the electrolytic-plating process itself. A typical tinning and tin-fusion line is shown schematically in Fig. 1. The steel strip to be tinned as delivered to the entry end of the tinning line is 28 to 36 inches wide, 0.010 inch thick, about three miles long, and wound in a tight coil. This coil is placed on an uncoiler, threaded through a welder, and enters the first pinch roll. The welder is used to attach the end of one coil to the beginning of the succeeding coil, and the pinch roll maintains tension against the drag of the uncoiler. Beyond the pinch roll is a looping pit for the purpose of accumulating a length of strip which is taken up when the strip is stopped at the welder to attach the end of a coil to the beginning of a fresh coil. Considering the speed of the line and the length of time needed to weld the beginning of a new roll of strip to the end of the old, it is not practicable to accumulate enough length of strip in the looping pit to enable the speed of the line to be maintained during the welding operation. The line has to be slowed down and automatic controls are provided for both the tinning and flowing apparatus so that thickness of the tin coating and reflow temperature are maintained essentially constant with variations in line speed. After passing through a drag unit the strip enters the plating tanks where it is coated with tin one side at a time. Desired thickness of coating is obtained by adjustment of the plating current which is of the order of 60,000 amperes at 12 volts direct current. The strip passes from the plating tank to a rinse (not shown on Fig. 1) which removes any plating solution still adhering to the strip. This completes the plating operation and the tinned strip next enters the inductor-heating coil where it is heated to a temperature corresponding to the melting point of tin and the tin flows to produce the mirrorlike finish expected of tin plate. The strip is then immediately quenched to preserve this surface.

Following this, the strip either may be coiled or continued directly to a shearing unit to be cut into appropriate sizes to meet the requirements of the canning industry. A shearing line does not operate faster than about 700 feet per minute without danger of excessive buckling of the sheets so that provision is made for coiling the finished strip and later shearing it into sheets as a separate operation. To shear accurate lengths of strip it is essential there be no tension in the strip as it enters the shear which might cause slippage of the strip on the feed rolls, and for this reason the strip enters the shear from a loop. Just before the strip enters the shear it is usually passed through a pinhole detector and a flying micrometer (not shown on Fig. 1). A suitable memory device operates a gate in a classifier to shunt out rejects.

For a half-pound coating of tin per 100 pounds of strip the coating on each side of the steel strip is only 30 millionths of an inch, or less than 1/100 the diameter of a human hair. This unflowed surface has a dull matte appearance which under a microscope is seen as a succession of steep pinnacles and deep valleys and tests show that such a tinned surface does not have a satisfactory corrosion resistance. Upon heating the tin to its melting point the hills and valleys level off and a superior corrosion resistance obtained. Various methods

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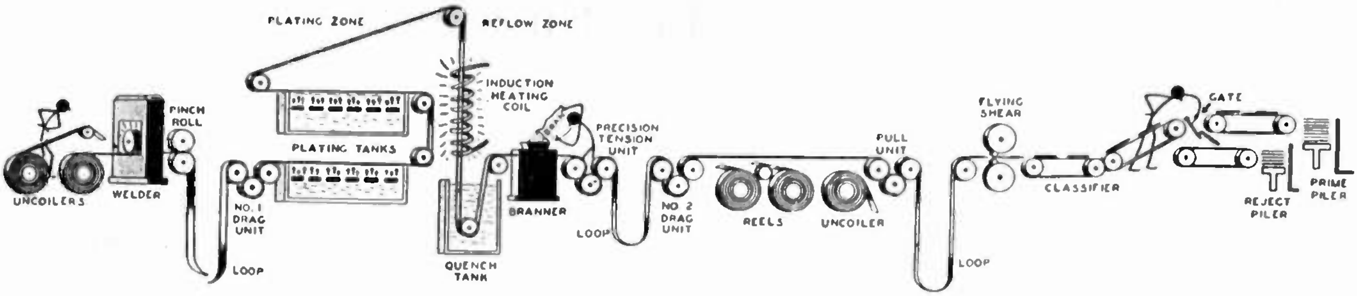

Fig. 1—Typical tinning and fusing line.

for heating the tin strip were considered but high-frequency induction heating using vacuum-tube oscillators as the source of high-frequency energy was chosen by some operators for the following reasons:

(a) No physical contact is needed between the moving strip and the electrical circuit.
(b) Work may be kept at ground potential.
(c) Lends itself to rapid changes in speed of the tin line such as occur when the line is slowed down while the end of one piece of work is welded to the next following.
(d) Provides uniform heating of the tin strip avoiding overheated spots or areas insufficiently heated to flow the tin.
(e) Permits rapid rate of heating and eliminates difficulties attendant on those heating processes which depend upon heating the tin strip by passing it through a furnace which tends to store a large quantity of heat.
(f) Makes it possible to combine the plating and fusion lines, eliminating the handling, delay, and expense where the fusion process is such that its speed cannot be matched to the plating line and it is necessary to conduct the two operations separately.

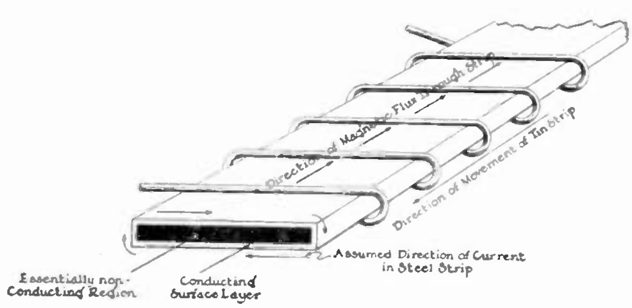

Fig. 2—Arrangement of inductor coil with respect to tin strip.

To flow the tin it must be heated to about 450 degrees Fahrenheit. To accomplish this the strip is threaded through a solenoidal coil carrying high-frequency current and the energy in the electromagnetic field surrounding this electrical circuit is transferred into the moving strip where it is converted into heat. The energy required for a given strip and line speed is determined from the well-known expression for the quantity of heat which must be added to M pounds of a substance of given specific heat to increase its temperature ΔT degrees. For steel strip, values of 0.13 may be assigned for specific heat and 7.82 grams per cubic centimeter for density and the following expression obtained for the power required to heat tin strip to 450 degrees Fahrenheit.

$$KW = 3 \times S \times b \times t. \tag{1}$$

KW = kilowatts power required.
S = speed of line in feet per minute
b = width of steel strip in inches
t = thickness of steel strip in inches.

For speeds of 500 and 1000 feet per minute energy has to be transferred at quite a rapid rate and considerations of first importance are to establish whether a practical inductor-coil design is possible and to select an appropriate frequency which will produce the desired rate of heating.

The arrangement of the strip with respect to the inductor coil is shown in Fig. 2. It is convenient to assume the steel strip as a single-turn conductive sheet linking with the flux from the inductor coil. The reasonableness of this assumption is evident, since at the high frequencies involved, skin effect will concentrate any current flowing around the strip in a thin surface layer, and for all practical purposes, the main part of the steel strip will behave as an electrical nonconductor. According to Faraday's law of induction, the voltage around a circuit through which magnetic lines of force vary is equal to the time rate of change of this flux; i.e., the voltage induced in the steel strip will be proportional to (a) the frequency of variation; and (b) the total flux linking the strip with the inductor coil. It is evident that if the total flux through the work were to remain constant, an increase in frequency would always produce more voltage around the strip and, hence, more heating current. However, for a given magnetizing force in the inductor coil, because of skin effect, the alternating flux density at the center of the strip becomes less and less with increase of frequency, and the higher the frequency the more pronounced this effect, until at very high frequencies there is alternating flux only within a skin layer at the surface of the strip. As frequency is increased, this restriction of flux results in a decreasing number of flux linkages between the strip and inductor coil, and this tends to offset the advantage otherwise gained by increasing frequency. Exact mathematical expressions for the effect of frequency on rate of heating are available[1-2] but the difficulty of assigning appropriate values to some of the variables makes it seem preferable to use an approximate relationship given by Rosenberg[3] for eddy-current loss.

$$P = 2 \times \sqrt{\rho f N^3 B} \times 10^{-4}. \tag{2}$$

P = eddy-current loss in iron mass-watts per square centimeter of surface
ρ = resistivity ohm-centimeters
f = frequency—cycles
N = root-mean-square ampere turns per centimeter—magnetizing force at surface of the iron mass
B = saturation flux density for the iron mass—gausses

This expression is applicable for any cross section if (a) the thickness of the conducting skin layer is small compared to the total thickness of the iron mass; (b) in this depth the magnetic-flux density is substantially uniform; (c) the current density in the conducting skin

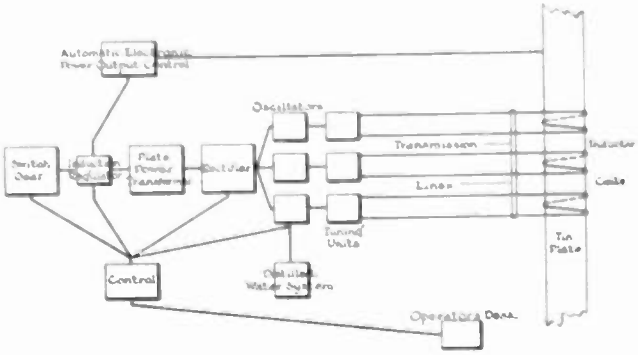

Fig. 3—Block schematic for a high-frequency induction-heating equipment for electrolytic tinning lines.

layer varies linearly from a maximum at the outside surface to zero at a depth corresponding to the thickness of the conducting skin layer. For tin flowing, these conditions can be approximated if (a) frequency is made sufficiently high that skin depth of current flow is small compared to thickness of the tin strip; (b) the magnetizing force at the surface of the tin strip is well in excess of that necessary to produce saturation flux density. Current density will vary in the conducting skin layer according to the usual exponential expression but for practical purposes a sufficiently close approximation is obtained by assuming that current varies in a straight-line relationship from a maximum at the outside surface to zero at the inside boundary of the conducting skin layer.

For practical purposes it is desirable to know the rate of heating per unit volume of tin strip and by assigning a value of 18,000 gausses as saturation flux density for the steel strip, converting centimeters to inches, and assuming a resistivity of 15×10^{-6} ohm-centimeter. The rate of heating for tin strip to its flowing temperature of 450 degrees Fahrenheit is

$$W = 3.307 \times 1/t \times \sqrt{fN^3} \times 10^{-4} \tag{3}$$

W = rate of heating tin strip in the temperature range between room temperature and 450 degrees Fahrenheit—watts per cubic inch
t = thickness of tin strip in inches
N = root-mean-square ampere turns per inch magnetizing force at surface of tin strip
f = frequency—cycles per second

It is seen that the rate of heating of the tin strip within the temperature range under consideration for tin flowing is inversely proportional to thickness of the tin strip, proportional to square root of frequency, and to the three-halves power of the ampere turns per inch in the inductor coil through which the strip passes. It is interesting to note that for a given frequency, if the root-mean-square current in the inductor coil is held constant, the rate of heating per unit volume for a given thickness of tin strip is independent of the strip width.

The product of the volume of strip within the inductor coil and the rate of heating per unit volume must equal the total power required as determined from (1), and for heating tin strip to flowing temperature the length of the inductor coil will therefore be

$$l = 3 \times S/W \times 10^3 \tag{4}$$

l = length of inductor coil in inches
S = speed of strip in feet per minute
W = rate of heating in watts per cubic inch

Fig. 3 shows the block schematic for a high-frequency induction-heating equipment for electrolytic-tinning lines. It comprises the following major units:

Rectifier	Metal-clad switchgear
Oscillators	Induction regulator
Tuning units	Plate-power transformer
Vacuum tubes and rectifier tubes	Control and metering unit
Transmission lines	Cooling water heat exchanger and water pumps
Inductor coils	Electronic regulator

Fig. 4 is a much simplified schematic diagram omitting metering, control, and protective and interlock circuits. The circuit is a self-excited push-pull oscillator with a three-phase full-wave rectifier. The equipment operates from the steel mill's service bus; 6900 volts, 3-phase, 60 cycles being typical. In steel mills, because of the capacity of the feeder circuits to which the induction heating equipment is connected, the circuit breaker which provides alternating-current protection

[1] C. R. Burch and N. Ryland Davis, "Theory of Eddy Current Heating," Ernest Benn, Ltd., London, England.
[2] A. G. Warren, "Mathematics Applied to Electrical Engineering," D. Van Nostrand Co., New York, N.Y., 1940, pp. 243–264.
[3] E. Rosenberg, "Eddy currents in iron masses," *The Electrician*, August 24, 1923.

for the high-frequency electronic equipment usually has an interrupting capacity of about 250,000 kilovolt-amperes. An air circuit breaker is used in preference to an oil breaker because unlike usual circuit-breaker installations the main breaker for tin-line oscillators is expected to be opened and closed repeatedly at frequent intervals as it must open or close the plate-power circuit for every "on-off" operation at the flow-line operator's control desk.

It is beyond the scope of this paper to discuss the control and interlock circuits in detail but the more important features provided will be mentioned. The main plate-power breaker is equipped with an automatic recloser circuit so that if a fault occurs which is only temporary in nature, such as a tube flashover, the breaker will immediately reclose automatically. If the fault persists the breaker will trip out after one reclosure and remain locked out. In case of an interlock circuit opening up such as might happen if a cubicle door is opened, water does not flow properly to the tube water jackets, cooling water temperature becomes dangerously high or tube filament circuits are open, the plate-power circuit breaker will open and remain locked out and automatic reclosing does not occur when the breaker is tripped from any one of these causes. To extend tube life, time-delay relays automatically insure that there is the necessary time interval after starting the tube filaments and before applying plate voltage to the oscillator tubes.

To avoid current surges into the plate-power transformer, step-start resistors together with a short-circuiting contactor are inserted between the main plate-power breaker and the induction regulator in the plate-power circuit. The short-circuiting contactor acts to short out the starting resistances within a few cycles after the plate-power circuit is closed. Considering the large kilovolt-ampere ratings of the circuit to which the plate-power equipment is connected, the starting resistance short-circuiting contactor must be rugged enough and of sufficient interrupting capacity to open in the event of a short circuit beyond it.

Power output from the oscillator and rate of heating is adjusted by varying the alternating voltage to the rectifier. A 100 per cent buck or boost motor-driven remote-controlled induction regulator is used so that for full power twice the service-bus voltage is applied to the primary of the plate-power transformer and when the induction regulator is in its 100 per cent buck position, it bucks out the primary line voltage so that there is zero voltage on the primary of the plate-power transformer. Control circuits are so arranged that whenever the main breaker is tripped open the induction regulator starts toward its 100 per cent buck position so that when the main breaker is closed after having been open for any appreciable time interval an initially low or zero voltage will be applied to the rectifier. Upon closing of the main plate-power breaker, if automatic control of power output is in use, the induction regulator will automatically adjust itself to give the required power output. When manual control of power output is in use the induction regulator will remain in the position it had assumed just prior to closing of the main breaker.

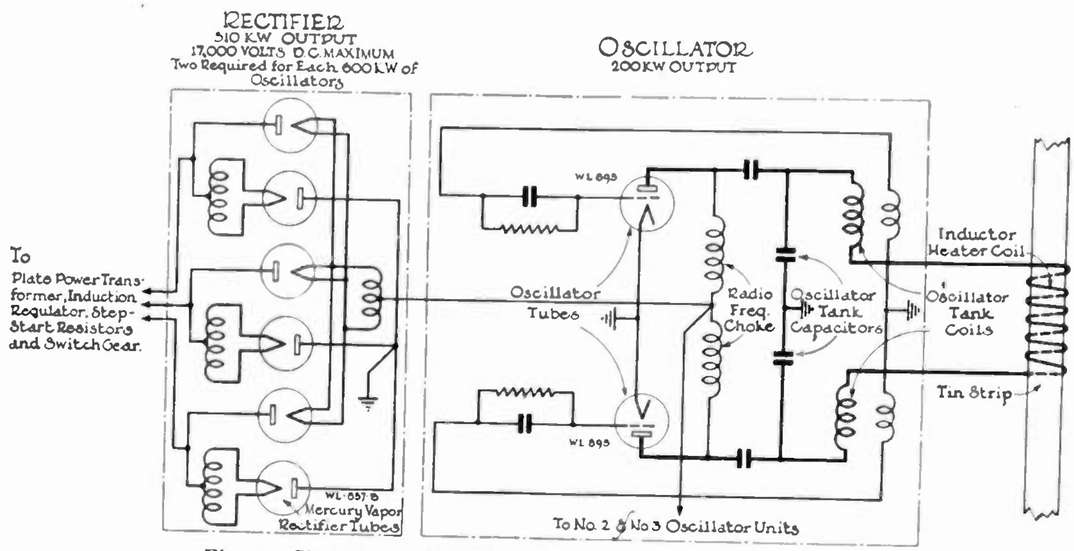

Fig. 4—Simplified schematic industrial electronic-type high-frequency generator, 200 to 600 kilowatts.

The rectifier, assembled on a steel chassis and cubicle construction (Fig. 5), is a conventional 3-phase, full-wave type using 12 WL-857-B mercury-vapor rectifier tubes. It delivers a maximum of 1020 kilowatts at 17,000 volts direct current which is sufficient to supply the needs of three 200-kilowatt oscillator units. Each oscillator unit (Fig. 6) comprises two water-cooled type WL-895 vacuum tubes and in conjunction with the associated tuning cubicle (Fig. 7) is capable of continuously supplying 200 kilowatts of power output at a frequency of 200,000 cycles. The tuning-unit cubicle houses the tank inductance, and capacitors which, in combination with the inductor coil, determine operating frequency. The tank-circuit inductance is water-cooled to help carry away the heat produced by coil losses and to minimize the resistance of the tank circuit inductance. Shielded transmission lines connect each 200-kilowatt

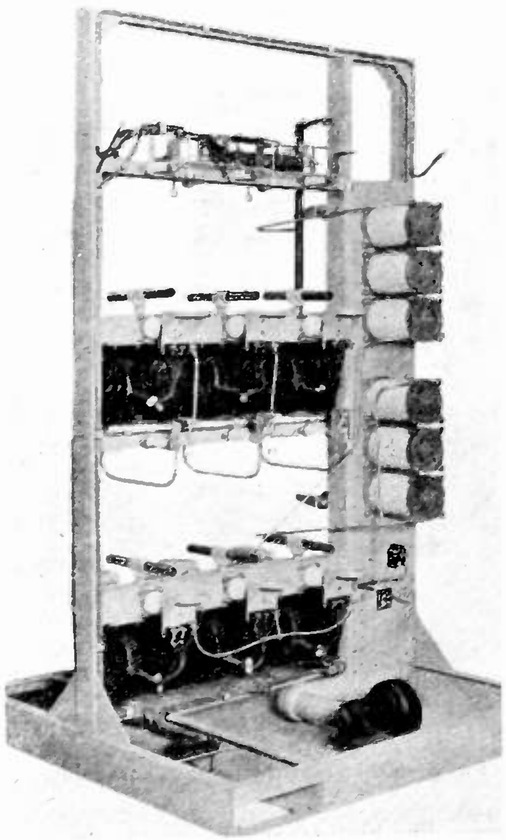

Fig. 5—1020-kilowatt rectifier (tubes and cover removed) is a conventional 3-phrase, full-wave type.

oscillator-tank circuit to its inductor coil at the tin-flow line.

Since the oscillator tubes are water-cooled, the usual distilled-water closed circulating system with pumps and distilled-water storage tank forms a part of each installation. All piping and tanks in the distilled-water system including the pumps are of noncorrosive metal to prevent contamination of the water, minimize scale deposit on the tube plates, and maintain the nonconducting properties of the distilled water. Heat dissipated in the tubes, tank inductances, and inductor coils and carried away in the cooling water is transferred through a water-to-water heat-exchanger, and eventually carried away by the steel mills raw-water system.

Construction of the inductor coils is shown in Fig. 8. The important factors influencing choice of the particular construction used were maximum freedom from tin-dust accumulation, suitable shielding, provision for use of controlled atmosphere, accessibility for maintenance, ability to substitute different-sized inductor coils in the shortest possible time (the inductor coils are water-cooled), and guarding against possibility of bowed or warped strip accidentally making contact with turns of the inductor coil.

Excepting such few instruments as are mounted with the metal-clad switchgear all meters, controls, timing relays, small contactors, etc., are mounted in a separate control cubicle and on an operator's control desk. This permits flexibility of control arrangement without sacrificing advantage of quantity production of a standardized 200-kilowatt electronic-type generator design. In this control cabinet and on the operator's control desk are such meters, relays, and contactors, as alternating voltage at rectifier, inductor-coil current, oscillator, tube plate and grid currents, filament voltages, oscillator direct-plate-current overload relays, grid overcurrent relays, relays for over- and undervoltage of filament supply, water-pump overload relays, thermal overload relays for induction-regulator driving motor and water-pump motor, contactors for induction-regular driving

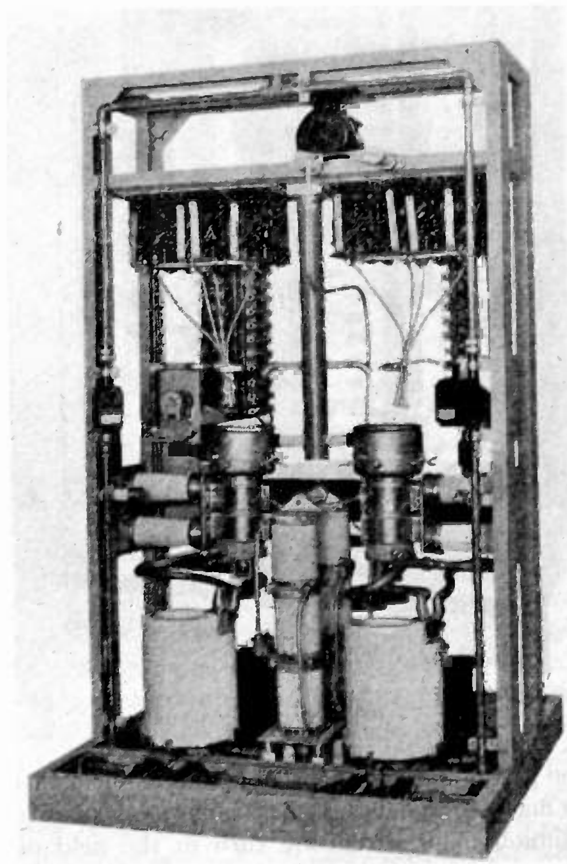

Fig. 6—200-kilowatt oscillator (tubes and cover removed).

motor, oscillator and rectifier filaments, water pump, and blower motors.

Probability of considerable fluctuation in the steel-mill's service-bus voltage and the importance of maintaining constant-filament voltage on the oscillator and rectifier tubes is provided for by connecting filaments and miscellaneous control circuits to the service bus through an auxiliary transformer and small automatic line-voltage induction-type regulator. This auxiliary circuit is connected to the service bus ahead of the main plate-power circuit breaker so that the filament and auxiliary power circuit are independent of operation of the main plate-power circuit breaker. The auxiliary circuit is protected by its own current-limiting disconnect fuses.

All cubicles are of steel construction as are the inductor-coil shields, and precaution has been taken to guard against losses due to the strong electromagnetic fields which are present particularly in the tuning-unit cubicles and at the inductor coils. Where possible, construction is made to avoid the steel framework constituting a continuous short-circuiting turn in the field of the tank inductance or inductor coil; and the inside of the cubicles is given a thin highly conducting coating of copper. This layer although relatively thin is effective in minimizing losses because the high-frequency currents by reason of skin effect will flow mostly only on the surface of the sheet-steel cubicles and shields, and only on the side adjacent to the inductor coils or tank inductances. The high-frequency electronic equipment, transmission lines, and inductor coils are shielded as a precaution against interference to radio services. Controls or switches manipulated during operation are of the dead-front or low-voltage type. Interlocks are provided on doors to all compartments in which dangerous voltages are present.

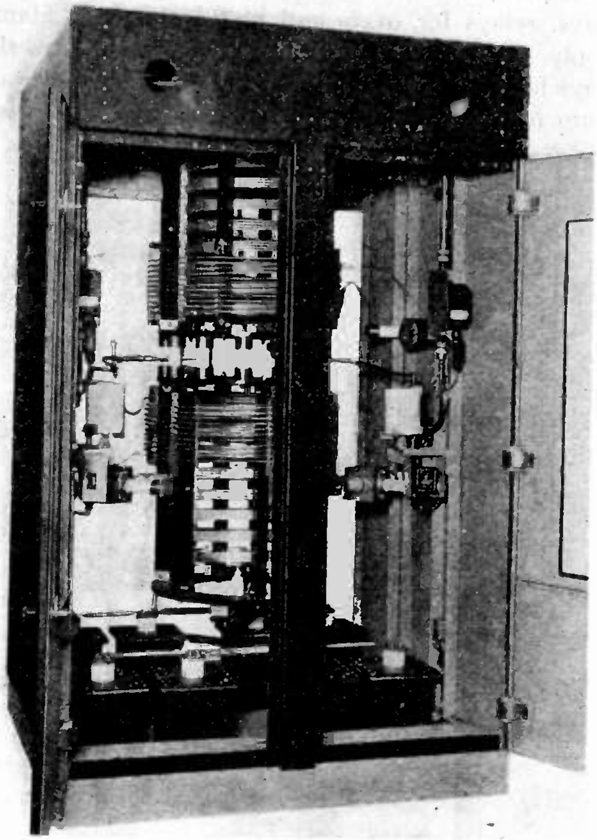

Fig. 7—200-kilowatt tuning unit.

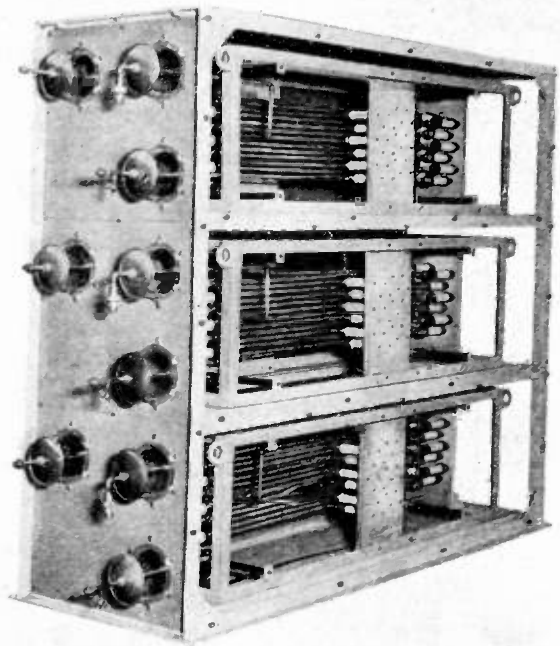

Fig. 8—600-kilowatt inductor-coil assembly.

Although these steel-mill applications pioneer in the industrial use of high-frequency high-power electronic equipment, they are but a variation of high-power vacuum-tube equipment as developed for communication applications and therefore already developed to a high state of dependability. They are consistent in performance, stable in operation, capable of remote control, rapid adjustment of power output, and automatically protected against overload or damage to equipment from improper operation.

Correction

C. A. Hultberg, whose paper "Neutralization of Screen-Grid Tubes to Improve the Stability of Intermediate-Amplifiers" appeared in the December, 1943, issue of the PROCEEDINGS on pages 663 to 666, has brought to the attention of the editors an error in the last sentence on page 665. The sentence should read ". . . again provide *sufficient* reaction to make the application of Hazeltine's principles advantageous" instead of ". . . again provide *insufficient* reaction to make the application of Hazeltine's principles advantageous."

Flexibility in the Design of Military Radio Apparatus*

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A Proposal to Increase Military Radio Production

Summary—Numerous factors limited the production of military radio and electronic apparatus at the inauguration of the Defense Program. Corrective action was taken to eliminate them as progress was made on the Defense, and later, the War Program. With their elimination, new factors became limiting. The major factor which establishes the current industry production ceiling is the supply of critical materials and components. The application of alternates for both has materially reduced the effects of the present limitations. Much more can be accomplished by concentration on the problems and this paper describes measures which have been taken and suggests procedures for future action.

Early Limitations on Production

SINCE the inauguration of the Defense Program there have been numerous factors which limited the production of military radio equipment to something less than a desirable maximum. As progress was made, corrective action modified these factors and those which limit us today are quite different from those of 1939. In the beginning, the greatest limitation was the lack of buildings and facilities—factories. It was not material then, nor manpower. It was not engineering since a standardized design for almost every kind of military communication service was available and promptly frozen. Through, and since, the last World War a few companies had maintained organizations for the design and production of military radio apparatus. In co-operation with the Government radio laboratories they had produced the apparatus which filled the requirements of the Military over the intervening years. It was a small business, perhaps 15 million dollars a year in a highly competitive market, and at times a discouraging business because of the wrong guesses fostered by the overconfidence of young people to meet the increasingly difficult requirements of the Military Services. At all times, however, it was a period of progress. The job was mainly one of engineering with the engineering cost a disproportionate share of the total as measured by standards for other types of electrical apparatus. Consequently, the factory area per cost dollar was small by comparison with the same standards.

When the Army and Navy were given the green light, both were "ready to go" in radio communication equipment. The immediate requirements were greater than the capacity of available facilities and the first bottleneck was created. The Government and industry co-operated to provide facilities for the prime contractors and their suppliers, and in a short time a number of new plants and buildings were under construction. Then, with the freezing of home-receiver and other commercial radio activities, additional facilities were made available, and this limitation became progressively less important.

There were periods when the most limiting item varied between machine tools, jigs and fixtures, steatite parts, aluminum, and engineering talent. These things, however, are of historical interest, and the matter of greatest present interest is: What are the limitations today and what can we do about them?

Factors That Limit Production Today

Today, opinion may vary in the industry as to what the most important single factor is, but concurrence would be quite general on what the important factors are. Manpower is one, and a cause for concern by some, but it seems doubtful that this is of first importance. A military transmitter or receiver may be considered as a structure rather than a machine, because in general there is no relative motion between component members other than that due to the elasticity of the materials of construction. Each structure may and often does contain several examples of machines, such as relays, meters, gears, and levers, but a major share of the structure can be fabricated by the semiskilled. If a basic force of skilled workers can be retained to provide the tools and fixtures for the unskilled and make the few precision parts which seem to be beyond the capabilities of those without long experience, it should not be too difficult to take care of the rest of the job even if it must be done 100 per cent by unskilled workers, predominantly females.

Another factor to be considered is standardization. The lack of it will be classed by some as a most serious handicap. Granted that standardization is of importance and that any improvement is desirable, just how vital is it? There is more than one radio set which is used by both the Army and Navy. There is another which is used in airplanes, in ground stations, and in tanks. For years there have been Government standards on various components. Therefore, we do have standardization; not enough perhaps, but if we could improve it greatly overnight it is probable that an increased output of the industry would be noticeable only after a considerable period of time.

Some would class the need for design simplification as the major need. This objective is worthy, as always, but how much can be accomplished? The radio engineers who have been at this job for 20 years in a highly competitive market have not been using two vacuum tubes

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† Designing Engineer, Transmitter Division, General Electric Company, Schenectady, New York.

where one would suffice, and the Government laboratories, with the problem of wartime maintenance and service in mind, have permitted few complexities which could be avoided. It would be ridiculous to maintain that no simplifications can be effected, but it is doubtful, unless the service requirements are changed, that their magnitude would be substantially reflected in increased production.

Production methods are a factor which rate consideration. The job of ordering, routing, planning, making the utmost of available facilities, can account for weeks saved or lost. The expansion of the organization, which bridges the gap between the completed drawing and the delivery of the finished article to test, has been one of the greatest problems. The job requires specialists trained in the system of their particular company, and the absence of draft deferments for this type of man has not made the job any easier. A good production man may be worth more to the war effort than two good toolmakers, but not in the opinion of most draft boards. Considering the handicaps, production organizations generally, if not beyond criticism, are doing a good job.

To determine the importance of the factors mentioned let us group them. Let us suppose that we could further simplify and standardize our designs and hand them over to an ideal production organization which had adequate manpower at its disposal. By imagination we have eliminated four of the handicaps to production which are regarded as substantial obstacles to greater output. As a consequence we have a right to expect a prompt reflection of the action in increased shipments. If we could perform this magic today, would we guarantee more production? Would we even guarantee a continuation of the present going rate? In today's situation, insofar as the prime contractors and major subcontractors are concerned, the answer is "no".

Materials the Real Bottleneck

Current production is limited chiefly by the supply status of critical materials, and the quantity of apparatus delivered is governed by the allocation of these materials and certain components. It is not always evident whether the shortage is basic—raw material, or is due to insufficient facilities—rolling mills, etc., but the reason is unimportant because, in a large measure, the situation is beyond the control of the radio industry.

There are about 23 materials in group I of the critical metals list. About 14 of these are used in radio apparatus, but the requirements for 2 or 3 of them are almost insignificant. For the remaining 11 or 12, it is impracticable to obtain an authoritative listing of the materials in the order of scarcity. If this were feasible it is probable that the list would be subject to frequent revisions due to unpredictable conditions. The knowledge of what materials cannot be obtained in sufficient quantities, all too often comes by the hard way—the notification that the requirements for a given period cannot be filled. The problem to be solved by industry, if output is to increase or hold its own, is the material problem.

Substitute Materials Can Relieve Shortages

The obvious way to relieve the critical-material situation is by a program of substitution. However, because of the uncertainty of forecasts it should not be a program of outright substitution. Rather it should be the provision of alternates or the establishment of flexibility to the greatest feasible extent. In lieu of drawing changes the designer should provide an alternate list and allow his production and planning departments a free choice, dependent upon their current knowledge of material availability. They can use the optional material, go back to the original if conditions change, or use both.

Facts the Designer Must Keep in Mind

In considering alternate materials the designer must keep a few fundamentals in mind. The apparatus must be reliable and perform satisfactorily under severe conditions. The requirements written into the specifications on temperature range, altitude, humidity, etc., can be checked any day in the newspapers. The vital necessity of reliable equipment for fighting a war is self-evident. Much is heard about "20 to 40 hours combat life" but who has seen a directive or specification change to this effect? Every departure from past practice which is authorized by a military agency is with the reservation that it must not cause any degradation in performance or reliability. Authorizations for alternate materials or components cannot be given lightly. Thorough engineering consideration, and perhaps a trial run or lot, must first provide the assurance that reliability and performance will be maintained.

In selecting alternate materials the obvious thing to do is apply group III and group II materials if possible. Of some eight metals in group III we find only two, gold and lead, with possibilities. Among 17 in group II we find cast iron, platinum, silver, and paradoxically—because copper is in group I—we find beryllium copper. Three of the seven mentioned possibilities in groups II and III—cast iron, lead, and silver—are finding increasing applications. It does not require great consideration, however, to determine that much relief cannot be obtained from the less critical groups. Likewise there is some, but inadequate, relief in the less critical miscellaneous materials such as: some plastics, glass, ceramics, plywood, etc. Unfortunately the best substitutes for critical materials are other critical materials and the indicated action is to allow, insofar as feasible, a free choice dependent upon conditions at any given time.

Examples of Flexibility in Design and Materials

Considering the group I critical list it appears, from current procurement experience, that zinc is more plentiful than nickel or cadmium, common steel can be obtained more readily than the alloy steels or aluminum, and aluminum sheet is more readily obtainable than aluminum bar, rod, or extruded stock. Although we do not have any guarantee that the situation will be the same several months from now, the course of action for the present is clear. The included photographs serve to illustrate flexibility of design.

Steel Instead of Aluminum

Figs. 1, 2, and 3 show components of a high-production aircraft transmitter as made of aluminum and steel. The steel parts are completely interchangeable with their aluminum counterparts and weigh no more, thus complying with the basic requirement on alternates for aircraft applications. The use of steel in these components, plus two others not shown, conserves 2,088,000 pounds of aluminum on current orders.

Many more photographs could have been provided to illustrate alternate designs involving other metals than

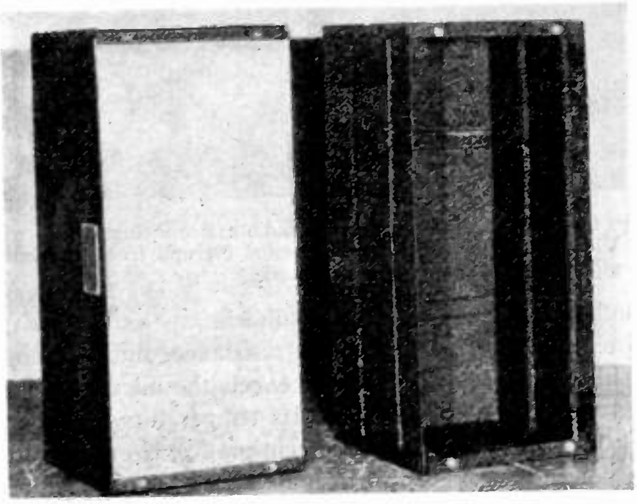

Fig. 1—Transmitter tuning unit cases for aircraft radio transmitting equipment. View showing aluminum-alloy construction (at left) and alternate steel construction (at right).

steel and aluminum. These would show parts made of beryllium copper, spring steel, and phosphor bronze; aluminum, zinc and plastics; copper and copper-weld; brass and steel; and others. Likewise many photographs of aluminum parts made from extrusions and sheet, from rod and sheet, and from rod and by die casting could be provided. In all cases tools are provided for the alternates and the selection of the material is by the production and planning groups.

To be fully effective, the plan for flexibility must include certain manufactured components. It is well known that the supply of some components is so inadequate as to subject them to allocation. In these cases the difficulty may be due to raw materials or insufficient facilities. Let us consider two of these: steatite and mica capacitors.

Substitutes for Steatite

Although the supply of steatite has been greatly increased during the past months, there are still numerous instances where insulators of this material are a definite bottleneck. While the major increase in the steatite output was due to new facilities, the generous application of alternates has helped considerably. Plastics, mycalex, porcelain, and glass have been applied where feasible. (See Figs. 4 and 5.) In most instances the application of these materials involved some loss, inasmuch as none of them is equivalent to steatite, but it was a loss which could be tolerated. The most satisfactory answer to the steatite problem is more steatite. If this answer requires more facilities than now operated by the ten or more

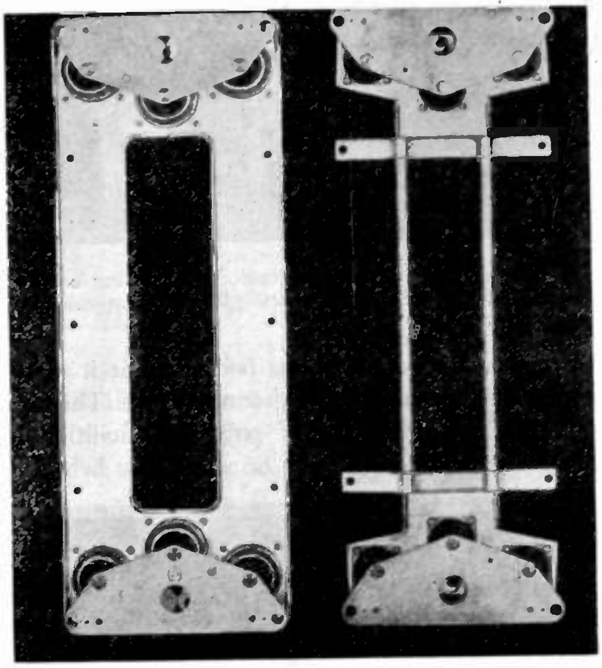

Fig. 2—Transmitter mounting bases for aircraft radio transmitter equipment. View showing aluminum-alloy construction (at left) and alternate steel construction (at right).

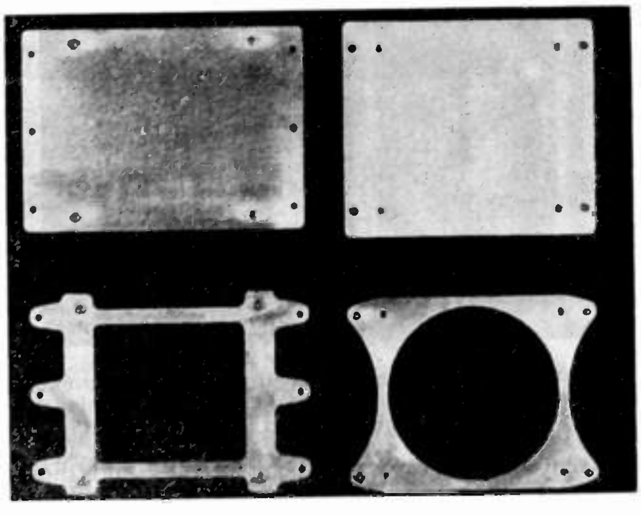

Fig. 3—Antenna tuning unit and dynamotor unit mounting bases for aircraft radio transmitting equipment. View showing aluminum-alloy construction (top row) and alternate steel construction (bottom row).

suppliers of the material the outlook would not be encouraging, because it is probable that the Army, Navy, and War Production Board would not support further expansion. What is needed is an equivalent material which can be produced without new facilities, and a body which meets these requirements will be available soon. By talc content it is a steatite. Its electrical and physical properties are slightly better than average of the materials in use. Its shrinkage is the same as one steatite in general use, satisfactory parts having been pressed in the same molds. However, this body contains no feldspar. The frit is a synthetic compound which is subject to precise control and is, therefore, uniform. The most

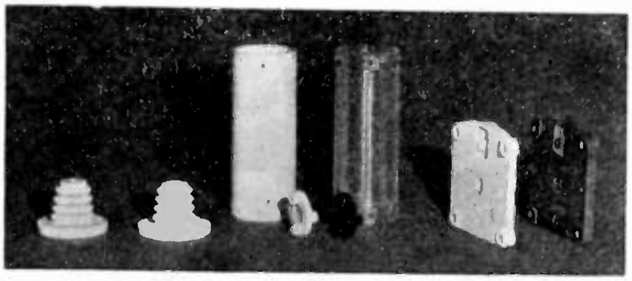

Fig. 4—Components for aircraft radio transmitting equipment. View showing porcelain, plastics, glass, and mycalex alternates for steatite.

important characteristic of the body is that it matures at cone 10, the porcelain firing temperature. This opens up possibilities for the use of porcelain facilities, particularly the tunnel kilns, few of which are being used

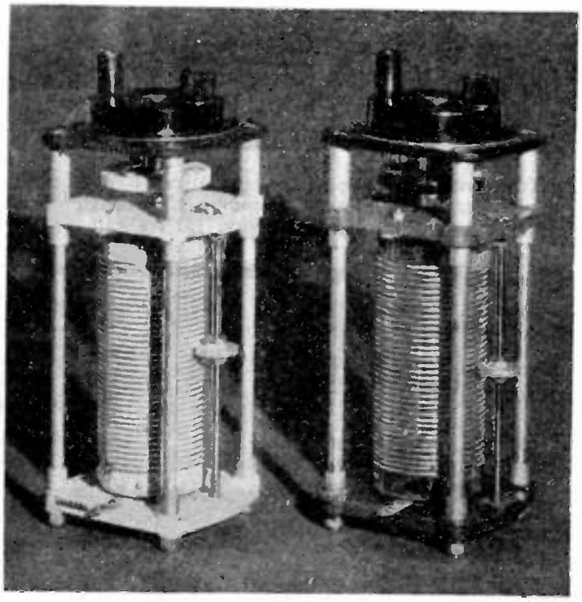

Fig. 5—Antenna tuning inductor for aircraft radio transmitting equipment. View showing (at right) plastics, mycalex, and glass alternates for similar steatite parts (at left).

at present to full capacity. Tools—molds and dies—must be provided for porcelain makers but the facilities and technique are available. At this time the material is in the proving grounds. A production run has been made and the parts are in various stages of subassembly, final assembly, and test. If this double and triple checking bears out the results obtained with test specimens, steatite will be definitely on the way out of the highly critical list of components.

Alternates for Mica Capacitors

Probably the current number 1 item in the list of critical components is the mica capacitor. The need for developing alternates for this item is evident to the Government and to industry as well. A WPB-sponsored program on the application of oil-impregnated, hermetically sealed, paper capacitors is well under way. This undoubtedly will provide some relief, but the limitations of the paper capacitors of the types now being manufactured will greatly restrict their application. A better answer is needed and, in particular, one which meets the requirements for flexibility by providing a capacitor which is strictly interchangeable, mechanically and electrically, with its mica counterpart. Fig. 6 shows a casing 9, American War Standard CM56, 400-micromicrofarad, 2500-volt capacitor. Twelve of this type and rating are used in one Army equipment for which unfilled orders require about 800,000 units. There is also shown

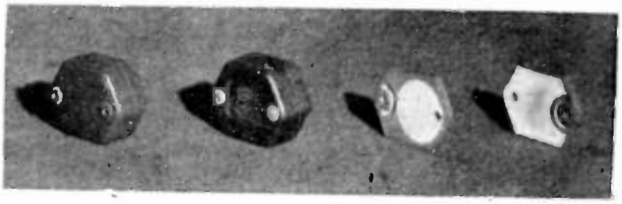

Fig. 6—Components for aircraft radio transmitting equipment. View showing typical mica capacitor (second from left) and alternate interchangeable capacitors.

another capacitor of identical dimensions and capacity. In temperature rating, leakage resistance, and operation in high humidity, it equals or excels the mica capacitor and yet it contains no mica. Its temperature coefficient (+0.05 per cent per degree centigrade) is greater and its power factor at low voltage (0.5 per cent to 1.5 per cent) is poorer, but for blocking and by-pass, where the twelve capacitors per equipment are used, these values can be tolerated. The dielectric used is a material which, as far as we know, has not been used heretofore for such a purpose. It is this material which provides the characteristics which compare so favorably with mica, and an adequate supply of it appears to be available. If proving runs and Government laboratory tests are successful, as preliminary tests indicate, there will be an alternate for CM56 micas which will help the general situation materially.

Fig. 6 also shows another alternate for the mica capacitor. It is made of a rutile body, titanium-dioxide with sufficient plastic materials to make feasible its pressing in thin intricate sections. It is fired at porcelain temperatures and silvered as necessary to provide the electrodes. Two plates of this size will make a capacitor of 400 micromicrofarads, 2500 volts rating. For blocking and by-pass applications the temperature coefficient will not degrade the transmitter performance materially. The plate has the same mounting dimensions as the CM56 mentioned heretofore and two of them together have the same over-all dimensions as the CM56. The 400-micromicrofarad assembly is, therefore, readily interchangeable and meets the requirements of complete flexibility. It is anticipated that production quantity tests will be complete and Government approval available in a short time. This ceramic capacitor, being a product of the porcelain plant, will impose no additional burden on the steatite maker and has possibilities for materially reducing the mica-capacitor requirements. Directly interchangeable ceramic capacitors for alternate use in frequency-determining circuits are not impracticable. These will be made of a magnesium orthotitanate body with its lower power factor, better than mica, and negligible temperature coefficient. This will take more time but undoubtedly such capacitors will come into use in the future.

An Alternate for the Porcelain-Tube Resistor

The two units shown by Fig. 7, while not by any means an answer to the wire-wound resistor problem,

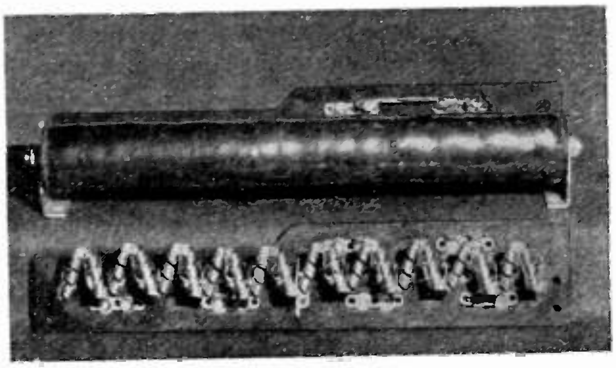

Fig. 7—Components for aircraft radio transmitting equipment. View showing (at bottom) alternate for conventional porcelain-tube-type resistor (top).

show substantial relief in an isolated case. One unit is a 1½-ohm tapped resistor, of conventional design, for the filament circuit of a transmitter. The alternate unit utilizes nichrome ribbon, welded to phosphor-bronze wires which serve as taps and supports. It has proved satisfactory on vibration and salt-spray tests. An incidental advantage is the substantial weight reduction. Several thousand of these resistors are required for one type of transmitter.

The provision of alternate materials and components is more feasible in frozen designs which are in production. Therefore, their greatest application so far has been on communication apparatus. Any alternate components which are available should, of course, be applied, but the provision of new alternates should not be allowed to deter the release of manufacturing information on the basic design. This task can be taken care of after engineering instructions are complete.

A Proposal for Increasing Production

The photographs illustrate the plan of giving a manufacturing and production group wide latitude in the utilization of critical materials and components. They are a representative few, selected to demonstrate an idea, and many more which achieve the same end could have been provided. If it had not been for these accomplishments by the General Electric Company, their output of military radio apparatus to date would have been considerably less and future schedules would predict lower quantities than at present. Undoubtedly this statement with respect to one company can be applied to the industry as a whole. Other companies have had to produce alternatives and substitutes to cope with the difficulties of the material situation. However, the total

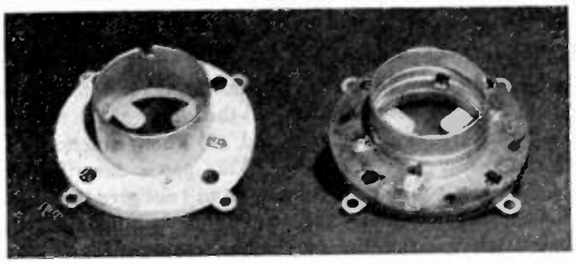

Fig. 8—Component for aircraft radio transmitting equipment. View showing (at left) tube socket of steatite and aluminum-alloy construction and alternate mycalex and steel construction (at right).

achievements of the industry are probably only a small percentage of what can be done, and the most important task ahead of us today is to finish the job. The output of military radio apparatus, while many times greater

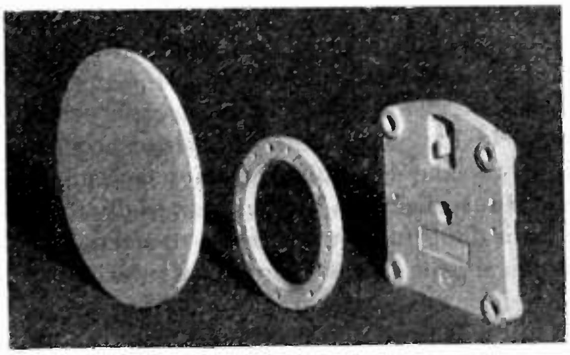

Fig. 9—Components for aircraft radio transmitting equipment. Samples of steatite which mature at porcelain firing temperature.

than in peacetime, has not been good enough. The output in the months to come will not be good enough, and it will be far less satisfactory if we stand by and let allocation take its course. There is not enough of the most desirable materials and components to carry the minimum schedules, and through ingenuity and resourcefulness we must compensate for this fact. The material situation is the major obstacle to increased production and, as such, it rates preferred attention from the engineers of the industry.

While this paper was not begun with the idea of advocating a program of joint action, such a conclusion appears logical upon its completion. It seems definite that the most good can be accomplished in the shortest time by a co-ordinated plan, between industry and the Government laboratories, which will provide for a free interchange of ideas and avoid duplication of effort. It would be well if such a plan were adopted promptly and followed aggressively because, if we do not make the most strategic use of critical materials and components, one war which will not be won is the War of Radio Production.

Acoustical Design and Treatment for Speech Broadcast Studios*

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Summary—**Three small speech studios have recently been completed by WOR in New York City. The factors taken into consideration are: (1) the reverberation time, (2) the shape of the reverberation-time-versus-frequency curve, and (3) the reduction of the standing sound waves in the studio. This paper describes how a studio designed for good music conditions will not give the best conditions for speech; particularly speech of the news broadcast type. By lowering the reverberation time at the low frequencies and allowing it to rise at the high frequencies, the effect is achieved of making the speaker in the studio sound as though he were actually in the home where he is being heard.**

The types of materials used in correcting broadcast studios are described. Graphs of the calculated time-frequency curves and as determined with a high-speed level recorder are shown. This paper explains the method of testing with the high-speed level recorder and also the discrepancies which occur between the caclulated and actual graphs. It explains the method by which the standing waves in the room are broken up to such an extent as not to be objectionable.

THE object of this article is to illustrate why ideal studio acoustical characteristics for music are not satisfactory for the transmission of speech. Potwin and Maxfield[1] have stated that there is still considerable question as to the desirable frequency characteristic of the time of reverberation in rooms especially for speech. In fact, it is doubtful if it is possible to arrange adjustable acoustical panels so as to provide a sufficient amount of change in acoustical characteristics to provide optimum conditions for both music and speech.

In the early days of broadcasting, it was realized that some kind of acoustical treatment was necessary for studios as broadcasts originating in untreated rooms sounded as though they were being given from a large barn. It was found that by hanging drapes, the reverberation of the studios could be decreased. There was still something wrong, so more drapes were put up. This did reduce the reverberation time at the middle and higher frequencies but, to such an extent however, that many of the upper harmonics in music were lost.

It was about this time that acoustical engineers realized that the curve of reverberation time versus frequency of a studio should have a definite shape. Materials were then tested and new materials developed which would permit the acoustical engineers and constructors to predetermine the shape of this curve as well as to control the amount of reverberation.

This was not enough, however, as opposing parallel surfaces caused standing waves and interference patterns in the studios to such an extent that it was difficult, if at all possible, to find good locations for the microphones where these interferences did not cause serious distortion to the sound pickup. Because of this, engineers then began to find effective means to break up these standing waves. This breakup of standing waves has been accomplished so effectively that, today, there are practically no bad spots in studios which have been properly treated and microphone placement has been greatly simplified.

It can be seen then that the governing factors for good acoustics are:

1. The reverberation time. This is the amount of reverberation and is measured in the time required for a sound, when suddenly interrupted, to die away or diminish to a value 60 decibels lower than the original sound.

2. The shape of the reverberation-time-versus-frequency curve. This curve shows the reverberation time for different frequencies in the studio spectrum.

3. The standing sound waves in a room. The standing sound waves in a room are caused by resonant conditions in the room. These resonant conditions, due to opposing parallel surfaces such as ceiling and floor, and opposing parallel walls, cause serious peaks in the reverberation-time-versus-frequency curve. These peaks cannot be eliminated by the use of acoustically absorbent materials. Therefore, some other method must be used to eliminate this condition.

The reverberation time can be controlled by the amount of sound-absorbing materials used. The greater amount of material, the shorter will be the reverberation time. These materials are applied to, or built into the walls and ceilings of the room which is being treated. The people in the studios must be taken into consideration, as each person represents a fairly definite amount of sound absorption. The shape of the reverberation-time-frequency curve can be controlled by selecting the proper amounts of the different available sound-absorbing materials and to a certain extent by the method in which some of these materials are used. Manufacturers of the materials available maintain testing laboratories and their products are tested at periodic intervals. This assures the constructor that his calculations and estimated results will be reasonably accurate.

The standing waves can be effectively reduced and practically eliminated by the use of surfaces placed at slight angles from the walls in such a manner as to eliminate large-area reflecting surfaces. This eliminates the large-order reflections by breaking them into a

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[1] C. C. Potwin and J. P. Maxfield, "Modern concept of acoustical design," *Jour. Acous. Soc. Amer.*, vol. 11, pp. 48–55; July, 1939.

larger number of smaller reflections. The absorbent material also should be applied in a large number of smaller sections rather than in a small number of large sections. If these angled surfaces and sections of acoustical treatment are placed in such positions that the reflected sound waves strike the absorbent sections more often, the large-order reflections will be more effectively broken up. This is achieved empirically and only through the use of knowledge gained from practical experience.

Optimum Reverberation Time

The optimum or most desirable reverberation time of a broadcast studio varies 1. with the size of the studio and 2. the uses of the studio, such as music, speech, etc.

The optimum reverberation time for music studios has been fairly well determined by Potwin.[2] (See Fig. 1.)

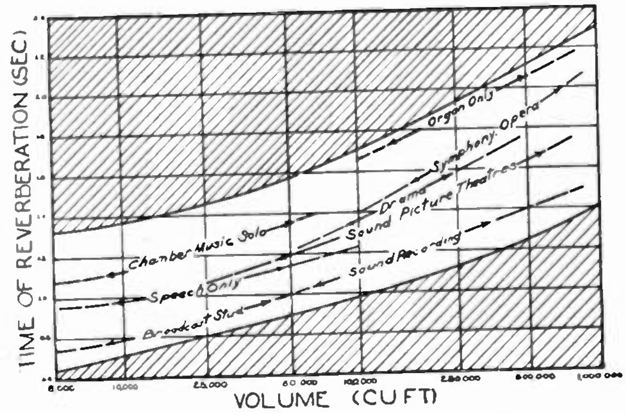

Fig. 1—Optimum reverberation time, in seconds, for various uses and for structures of various sizes.

It will be noticed that the reverberation time at 512 cycles, varies from 0.7 second for broadcast studios of 5000 cubic feet to 1.5 seconds for studios of 1,000,000 cubic feet.

Shape of Optimum Reverberation Curves

The optimum reverberation curves, for music and speech, in auditoriums have been published by MacNair.[3] These curves are shown in Fig. 2 and are based on the threshold of hearing for average ears. With reverberation in accordance with these curves the musical and speech sounds in an auditorium all die away, in the human ear, at the same instant. The reverberation time at low frequencies is about 60 per cent higher than at 512 cycles; the reverberation time at 1000 to 2000 cycles is shortest as this is the region of greatest sensitivity of the ear. The time at 8000 cycles rises again to a point 8 to 10 per cent greater than that at 512 cycles. Remember these are optimum reverberation curves for a large auditorium or legitimate theater. Orchestral music sounds best under these acoustical conditions.

[2] C. C. Potwin "Architectural accoustic design," *Arch. Forum*, September, 1939.

[3] W. A. MacNair, "Optimum reverberation time for auditoriums," *Jour. Acous. Soc. Amer.*, vol. 1, p. 242; 1930.

Broadcasting, however, presents an entirely different problem as both music, from orchestras, large or small and from individual instruments, and speech must be reproduced by the loudspeakers of the radio receivers in the listeners' homes.

At this point let us stop and consider, basically, what is to be achieved. Nothing can be done about treating the home acoustically, but the studio for music can be treated to provide the same characteristics as for the auditorium and effectively transfer the auditorium acoustical characteristics into the home by means of the loudspeaker. Under these conditions the music will be reproduced under the best conditions, as the reverberation time in the home is so short that the musical characteristics will not be materially affected.

For speech, however, these characteristics are not desired, as the speech, when reproduced in the home, should not sound as though the person speaking is in an auditorium or large hall. What effect then, is desired? The speaker's voice, in the home, should sound as though the person speaking were actually in that home. Only in this manner can the "intimacy," so desired by producers be achieved; that is, without the breath and lip sounds which are heard when a person speaks directly into a microphone from a distance of three or four inches.

This can be accomplished by treating a speech studio in such a manner that the acoustical characteristics of the room neither add nor detract anything from the tones of the speakers' voice. The optimum cnaracteristic

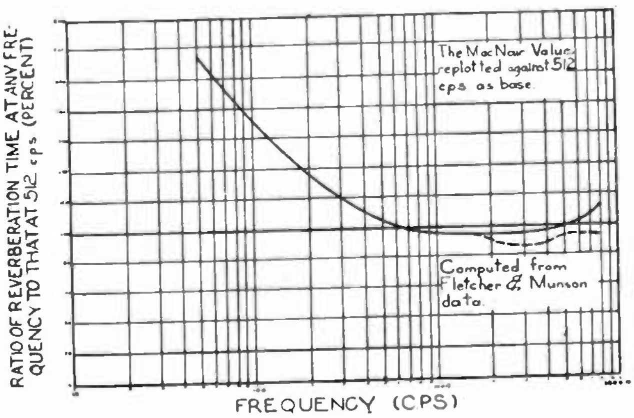

Fig. 2—Optimum ratio of reverberation time at any frequency to that at 512 cycles per second, in per cent, for auditoriums.

for a speech studio then, is one having a very low reverberation time which is fairly uniform throughout the lower and middle audio frequencies. The reverberation time at the high frequencies may be allowed to be 20 to 25 per cent greater than at 512 cycles. This rise at the higher frequencies provides greater intelligibility and also allows for the presence of one or two extra persons in the studio without materially affecting the reverberation-time–frequency curve.

Harvey Fletcher, in his articles on high-fidelity transmission, has demonstrated that the intelligibility of speech sounds is translated through the higher tones, those above 500 or 1000 cycles, while the apparent intensity or volume of speech sounds is provided by the tones having frequencies lower than these values.

A preponderance of low bass reverberation tends to make the voice sound "boomy," and impairs the intelligibility of speech, rather than to improve the quality of reproduction.

Three such speech studios have been built at station WOR. These vary in volume from 1000 to 1600 cubic feet. The calculated reverberation time at frequencies from 128 to 2048 cycles varies from 0.4 to 0.5 second and at 4096 cycles is approximately 0.6 second. The low bass resonance so frequently heard on speech broadcasts is entirely lacking in broadcasts from these studios.

As stated previously, the principal objective in the design of these studios was expressed: "Should sound as though the person speaking in the studio were actually in the room in the home." This result has been attained by adjusting the studio characteristics so that they do not change the quality of the speaker's voice materially. The remarkable fact is that proper acoustical characteristics of the studio improve the reception by all types and grades of radio receivers, regardless of their frequency characteristics. This paper will endeavor to show how, through the use and arrangement of acoustical materials, this objective was accomplished.

About six years ago the Johns-Manville research department, at their acoustical laboratory, started work on the development of special sound-absorbing treatments to be used primarily in broadcast studios, audition rooms, and similar spaces. This research resulted in the development of three types of acoustical treatments known as high-frequency, low-frequency, and triple-tuned elements. These three treatments are all composed of rock wool of different densities and thicknesses, used in conjunction with diaphragms of various densities and weights, and all in turn covered with perforated Transite. The Transite is used as a protective covering material, presenting a hard finished surface, readily capable of maintenance, decoration and repair. Perforated Transite also has the acoustical property of proportionately greater reflection of sound for the frequencies above 2000 cycles, as compared to the frequencies below 2000 cycles. The actual laboratory sound-absorbing values of these three treatments as determined at the Johns-Manville acoustical laboratory at Manville, New Jersey, are given in Table I.

A better comprehension of the sound-absorbing values of these materials can probably be shown by means of graphs.

Fig. 3 shows graphically the sound-absorbing characteristics of these three materials. The low-frequency element has a sharply descending curve as the frequency rises, the high-frequency element has a sharply ascending curve as the frequency rises, and the triple-tuned element has, for a commercial acoustical material a curve which has a fairly uniform or flat characteristic over the frequency range. It might be considered as the product obtained by combining the other two treatments.

TABLE I

	Frequency in Cycles Per Second					
	128	256	512	1024	2048	4096
	Percentage—Sound Absorption					
High-frequency element	20	46	55	66	79	75
Low-frequency element	66	60	50	50	35	20
Triple-tuned element	66	61	80	74	79	75

By adjusting the amounts of these three materials on the interior surfaces of any studio, it is readily seen that the contour of any time-frequency curve can be controlled or adjusted to any desired condition.

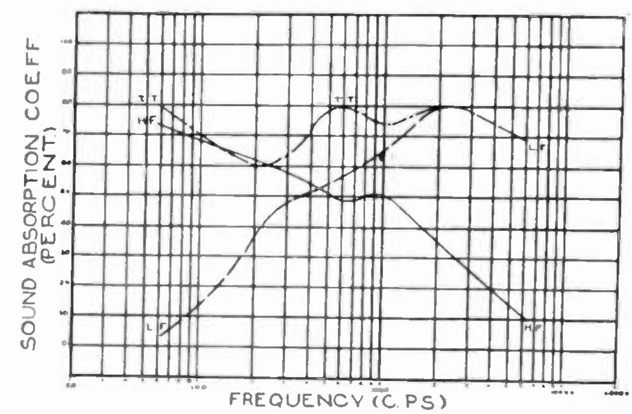

Fig. 3—Sound-absorption coefficients, in per cent, of:
T.T.—J-M triple-tuned element
L.F.—J-M low-frequency element
H.F.—J-M high-frequency element

The studios referred to are WOR studios 8, 9, 10, and 11, on the 24th floor of the building at Broadway and 40th Street, New York, N. Y. In estimating the desirable time-frequency curve, it was originally believed that the studio should have fairly flat absorption characteristics up to about the 1000- to 2000-cycle range with a rise in reverberation above that range not to exceed about 25 per cent of the actual reverberation time. It was believed that a small drop in the time at the low-frequency end would be permissible and might, in fact, even be desirable. It was afterwards found that this was very desirable and it is believed that it is due to the extremely low reverberation time below the 500-cycle range that the studio gets away from the "boominess" of the male voice which is so evident in most speakers' studios.

All calculations for reverberation were made using the Eyring formula which is a modification of the original Sabine reverberation formula. Sabine[4] showed "the general applicability of the hyperbolic law of inverse proportionality" between reverberation time and absorbing power as given by the relation: $T = KV/a$ where T is the duration of the residual sound, V is the volume

[4] Wallace C. Sabine, "Collected papers on Acoustics," Harvard University Press, 1900.

of the room, and a is the absorbing power of the surfaces. Standardizing reverberation time as the time required for the intensity of sound to drop one millionth of its original intensity, Sabine determined the constant K and obtained in English units the equation: $T=0.05V/S\alpha_a$ where S is the surface of the room and α_a is the average coefficient of absorption.

Eyring[5] showed that the Sabine formula did not hold in extremely dead rooms where the average sound absorption coefficient exceeded 0.05. He developed the formula: $T=KV/-S\log_e(1-\alpha_a)$ in which the symbols are the same as noted above.

The absorption coefficients for surfaces, people, and furnishings were taken from the tables given by Knudsen.[6] The tables in his book were from data published by Knudsen, Sabine, F. R. Watson, Bureau of Standards and other recognized laboratories.

A short description of the studios being discussed will now be given in order to explain more clearly the actual construction and layout.

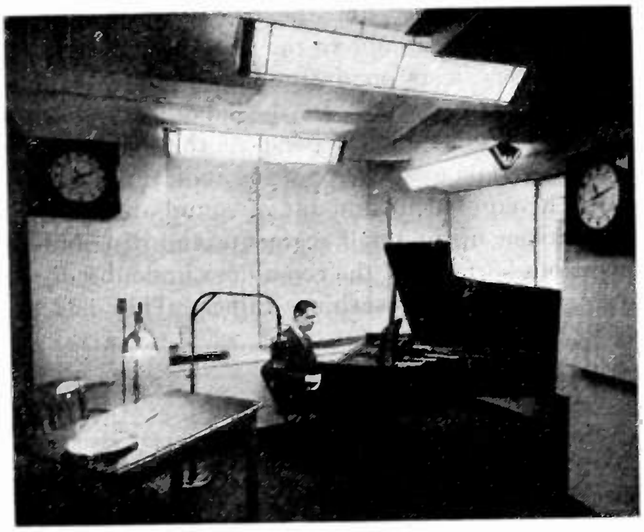

Fig. 4—WOR studio 8 showing sizes and placement of sound-absorbent panels.

Fig. 4 is a photograph of WOR studio 8. This studio is approximately 14×15×8 feet, 6 inches high. As it is placed in the corner of an irregularly shaped building, it is fortunate in not having any two parallel walls of any size. Furthermore, the ceiling is quite broken up to conceal ventilating ducts. The volume of studio 8 is approximately 1600 cubic feet.

All studios discussed are finished with plaster walls and ceilings, linoleum floor, speaker's table, two microphones, and two chairs. Studio 8 had, in addition at the time the tests were made, a studio grand piano.

Fig. 5 shows the calculated time-frequency curve and the actual time-frequency curves which were obtained by tests with a high-speed level recorder.

Curve C-C is the calculated design curve. Curves $A1$ and $A2$ are the conditions as determined by two test runs made with the high-speed level recorder. Curve $A1$ was obtained when the microphone was placed directly in front of the speaker source, face to face, and about two feet away from it, both near the center of the room. Curve $A2$ was obtained when the microphone and speaker were placed back to back about two feet apart and facing opposing walls.

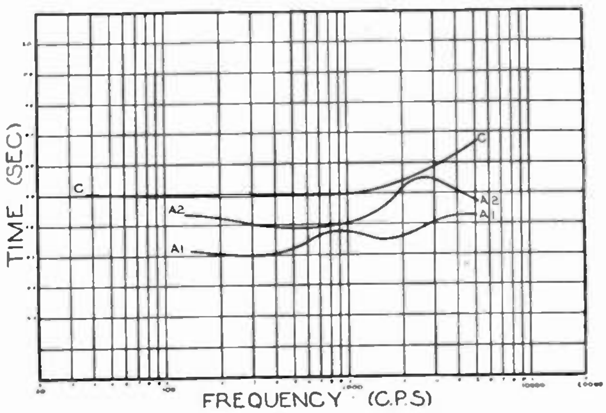

Fig. 5—C—calculated reverberation-time—frequency curve for studio 8. $A1$ and $A2$—curve of measured reverberation time, at various frequencies, of studio 8.

The test procedure for making the reverberation time runs was made in the following manner: Decay-time records were obtained with a high-speed level recorder manufactured by Georg Neumann and Company, Berlin, Germany, Type Pz-No. 62. This instrument is assembled for a 50-decibel range with attenuator box Type PS. No. 40. The sound sources were Bell Telephone Laboratories records BTL 108, BTL 175. These are warble records over the following six ranges, in cycles per second: 160 to 250, 260 to 500, 500 to 1000, 1000 to 2000, 2000 to 3000, and 3100 to 5400. The warble was rapid, approximately seven warbling cycles per second. Regular studio loudspeakers and pickup microphones were used, with their associated amplifiers. The signal was broken by quickly pulling out a lead plug in the loudspeaker circuit. Three decays were taken at each frequency range and averaged for the data given.

The values shown are reverberation times, that is, time in seconds for the sound-pressure level to drop 60 decibels from its initial steady-state condition. A straight-line slope was drawn on the record, using the first 40 decibels of the decay. The tape speed was 50 millimeters per second. From these the reverberation times were computed.

The microphone-speaker-position relation was chosen and a record taken over the frequency range. Another microphone-speaker-position relation was then taken, and a second record was run. Except in studio 8, these relations were so chosen as to avoid directing the speaker into the microphone, in order that the record would give random sound energy.

Studios 9 and 11 are twins but not quite duplicates. They are twins as far as their size goes, each being 8 feet, 4 inches×15 feet, 9 inches×8 feet, 3 inches high and,

[5] Carl F. Eyring, *Jour. Acous. Soc. Amer.*, vol. 1, p. 217; 1930.
[6] Vern O. Knudsen, "Architectural Acoustics," John Wiley and Sons, New York, N. Y., 1932.

having a volume of approximately 1050 cubic feet. They are not duplicates because one is right hand and the other reversed. Building considerations made the arrangements of acoustical materials, doors, etc. somewhat different, but the amounts of acoustical treatment and the furnishings of both studios are the same. Fig. 6 is a photograph of studio 11 and Fig. 7 is the calculated and actual time-frequency curves for studio 11.

Fig. 6—View of studio 11.

Through the middle range, the actual and calculated time-frequency curves came fairly close together, but at the low frequencies the actual time-frequency curve was considerably less than the calculated. Fig. 8 shows the calculated and actual results in studio 9.

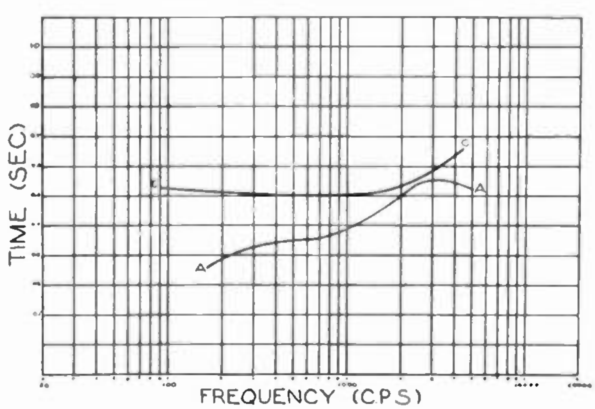

Fig. 7—*C*—calculated reverberation-time—frequency curve of Studio 11. *A*—curve of measured reverberation of studio 11.

Here again the actual and calculated time-frequency curves came reasonably close in the middle range, but again the actual time-frequency curve was much lower than the calculated time-frequency curve for frequencies below 2000 cycles.

As noted before, in the calculated results the coefficients of the sound-absorbing materials used were taken

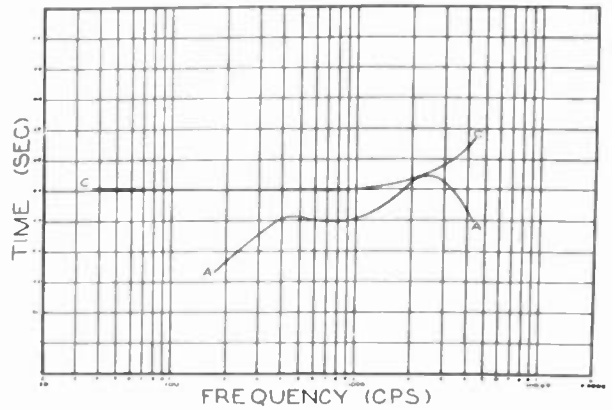

Fig. 8—*C*—calculated reverberation time and *A*—measured reverberation time of studio 9.

from laboratory values. It is generally recognized that all acoustical measurements made electrically with high-speed recording devices usually indicate reverberation times considerably below the calculated results. We believe this is largely due to the method used in making the laboratory tests on the sound-absorbing values of the acoustical materials. In the laboratory, tests are made on one solid unit or panel of the material to be tested, placed on one surface of the cubical proportioned room. The equivalent amount of sound-absorbing material broken up in small segments and distributed on the various surfaces of the room does undoubtedly give higher unit sound-absorbing values. This has been

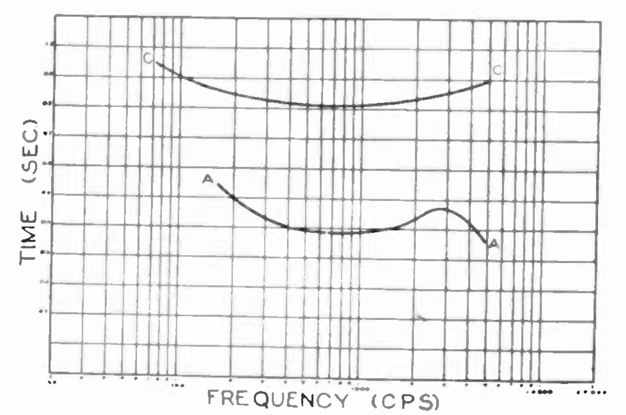

Fig. 9—Calculated and measured reverberation time of studio 10 as previously treated. The acoustical treatment of this studio now is similar to studios 8, 9, and 11.

proved by Parkinson.[7] As the material used in all these studios was predominantly low-frequency sound-absorbing elements, the studios undoubtedly had considerably more low-frequency sound-absorbing value than had been calculated. However, as was mentioned before, the very low reverberation time at the low frequencies certainly gives a much better tonal quality to the studio when used for the male voice.

[7] John S. Parkinson, "Area and pattern effects in the measurement of sound absorption," *Jour. Acous. Soc. Amer.*, vol. 2, p. 112; 1930.

As a comparison with the three studios described above, reference is made to studio 10 which is one of this same group of studios. This studio was built long before the other three studios; in fact, it antedates Pearl Harbor and the demand for small speaker studios was not so urgent at that time. Studio 10 has approximately the same proportions, dimensions, and furnishings as 9 and 11. It was originally designed to follow the standard practice for a speakers' studio, using the accepted curve for music or speaking and bringing the reverberation time down to adjust it to the volume. It is interesting to compare, in Fig. 9, the actual and calculated time-frequency curves of this studio with the three studios previously discussed.

The actual reverberation times, as shown by the high-speed level recorder, are considerably less than the calculated times over the entire range. The contour of the graph of the actual times is entirely different from that of the other three studios. It shows high reverberation times at the low frequencies and after vacillating around the middle range, drops down at the high frequencies. The quality of speech from this studio is distinctly inferior to that of the other three studios, and the only possible cause for this is the difference in acoustical adjustment. These comments on this studio have been borne out by various commentators who have used all of the studios. They uniformly are loud in their praises of the first three studios, and all prefer any one of the three studios to studio 10 for their work. Since this paper was written the acoustical material and arrangement of this studio have been made to conform to the other three.

It will be noticed from observing the photographs that the acoustical treatment is broken up into small panels and these panels are distributed in irregular arrangement on walls and ceilings. This tends to break up the large-order reflections. In addition to this, the surface of nearly all the acoustical panels are sloped. This degree of slope varies in the different panels, being between 5 and 10 degrees. This slope tends to break up the standing sound waves in the room, and in this manner minimize the serious peaks that often occur in the time-frequency reverberation curve. Then, too, there was absolutely no symmetry about the arrangement, size, or slope within the limits mentioned for the acoustical panels. The sloping of the panels and degree of slope were studied from a common-sense point of view, care being taken that the slopes on opposing parallel surfaces were placed in different planes so as to obtain as much as possible three-dimensional dispersion of the surface-reflected sound. This has been accomplished very well as the decay-time curves from the high-speed level recorder show a very straight drop for the first 40 decibels with no large pronounced peaks or valleys. The last 10-decibel drop was slightly wavy but this was probably due to extraneous sounds picked up by the microphone, and probably not due to decay of sound in the room.

The three studios have received nothing but favorable commendation from all speakers who have used them. When a person speaks in one of these studios, the natural tonal quality of the voice is very noticeable. Several news commentators have remarked at the ease of speaking in these studios, as their voices in these studios have a natural ring and not the blanketed effect that comes from a predominance of high-frequency sound-absorbing material. The net result to the auditor in his home with a fairly good commercial receiving set and speaker seems to be very similar to the effect of a person talking in the room without any appreciable loss due to transmission and reproduction facilities.

Polydirectional Microphone*

HARRY F. OLSON†, ASSOCIATE, I.R.E.

***Summary*—This paper describes a polydirectional microphone consisting of a single ribbon, the back of which is coupled to a damped folded pipe and an inertance in the form of an aperture. A single infinity of directional characteristics, ranging from bidirectional, through all variations of unidirectional to nondirectional, may be obtained by simply varying the size of the aperture.**

Introduction

From the inception of sound reproduction it was apparent to those associated with the problems of sound pickup that some form of directivity was desirable in the sound-collecting system to improve the ratio of direct to reflected sounds and thereby improve the reverberation characteristics and otherwise discriminate against undesirable sounds. Horns and reflectors were used for the early directional sound-collecting systems. As the fidelity of reproducing systems was improved, it became apparent that considerable distortion in the form of frequency discrimination in both the direct and reflected sounds was introduced because the directional characteristics of the horn and reflector systems varied with frequency. About ten years ago the velocity directional microphone[1] was developed which exhibited uniform directional characteristics over the entire audible spectrum. This microphone established

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[1] H. F. Olson, "Mass controlled electrodynamic microphone: The ribbon microphone," *Jour. Acous. Soc. Amer.*, vol. 3, pp. 56–68; July, 1931.

the usefulness and superiority of a sound-collecting system with uniform directional characteristics.

The conventional velocity microphone consists of a single mass-controlled ribbon with both sides freely accessible to the medium. The many desirable performance characteristics exhibited by this microphone may be attributed to the obvious simplicity of the vibrating system. The constants of the system may be chosen so that response and directional characteristics will be uniform over the entire audible frequency range. The nonlinear distortion for the intensity range of the human ear is a small fraction of a per cent. The light-mass-controlled system insures good transient response.

The polar directional characteristic of the velocity microphone is bidirectional. For certain sound pickup problems, unidirectional characteristics are more desirable. Accordingly, shortly after the development of the velocity microphone, a unidirectional microphone[2,3] with uniform directional and frequency response over a wide frequency range was developed. The conventional unidirectional microphone consists of the combination of a pressure element and a velocity element. The original unidirectional microphone employed ribbon elements for both the velocity and pressure elements. Employing ribbon elements makes it possible to maintain uniform phase relations between the velocity and pressure elements without resorting to correcting networks. The acoustic fidelity of the unidirectional microphone is essentially the same as that of the velocity microphone.

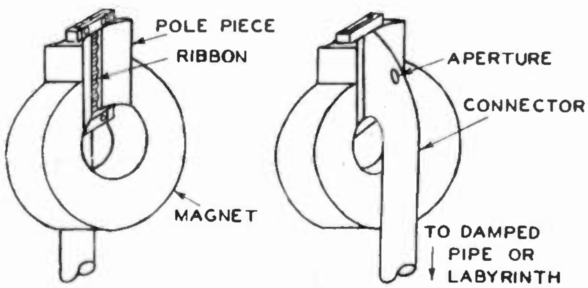

Fig. 1—The elements of a single-ribbon polydirectional microphone.

[2] H. F. Olson, "A uni-directional ribbon microphone," *Jour. Acous. Soc. Amer.*, vol. 3, p. 315; January, 1932.

[3] H. F. Olson and J. Weinberger, U. S. Patent, Reissue. 19115.

In view of the importance of directional microphones and the high fidelity of ribbon transducers, work has been continued on these systems with the object of increasing the scope and simplifying the vibrating system. As a result of this work a polydirectional microphone, known as the Varacoustic microphone, has been developed consisting of a single ribbon, in which it is possible to obtain any type of limacon[4,5] directional characteristic. It is the purpose of this paper to describe a single-ribbon polydirectional microphone.

Description

The single element unidirectional microphone is shown in Fig. 1. The ribbon is located in the air gap formed by the pole pieces. A permanent magnet supplies the flux to the air gap. The entire one side of the ribbon is covered by the labyrinth connector. The connector, in turn, is coupled to a damped pipe or labyrinth. The type of directional characteristic is governed by the size of the aperture in the labyrinth connector.

The action of this microphone can be obtained from Fig. 2 which shows the schematic view of the microphone and the acoustic circuit. The sound pressure acting on the open side of the ribbon may be written

$$p_1 = p_{01}\epsilon^{j(\omega t+\phi_1)} \tag{1}$$

where p_{01} = amplitude of the pressure

$\omega = 2\pi f$

f = frequency

t = time, and

ϕ_1 = phase angle with respect to a reference point.

The sound pressure acting on the aperture in the labyrinth connector may be written

$$p_2 = p_{02}\epsilon^{j(\omega t+\phi_2)} \tag{2}$$

where p_{02} = amplitude of the pressure, and

ϕ_2 = phase angle with respect to a reference point.

[4] A limacon is a curve defined by $e=a+b\cos\theta$. When $a=0$, $e=b\cos\theta$, a bidirectional characteristic. When $b=0$, $e=a$, a nondirectional characteristic. When $a=b$, $e=a+a\cos\theta$, a cardioid characteristic. For other values of a and b any type of characteristic of this family may be obtained.

[5] See H. F. Olson, "Elements of Acoustical Engineering," D. Van Nostrand Co., New York, N. Y., 1940.

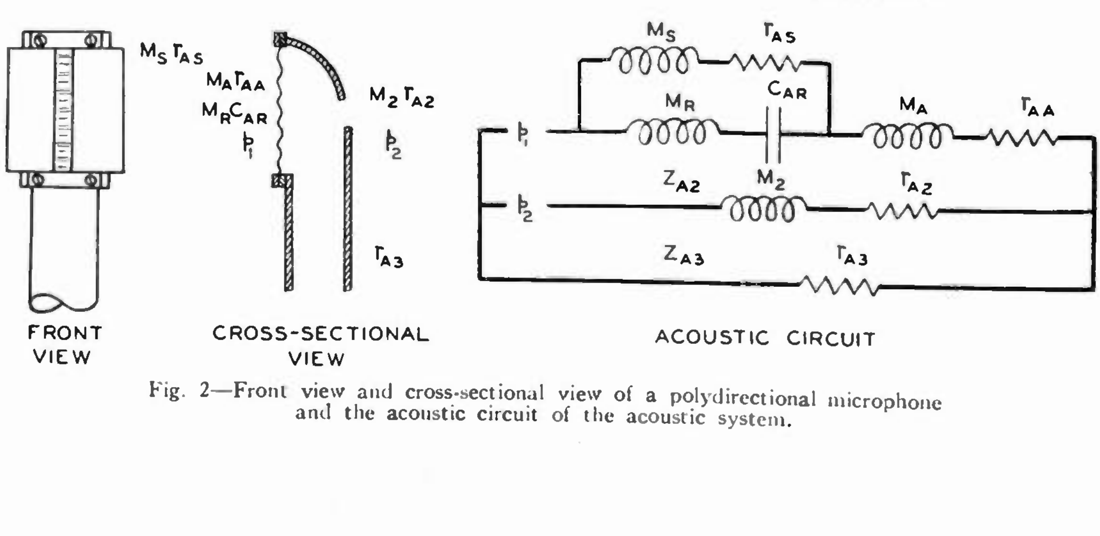

Fig. 2—Front view and cross-sectional view of a polydirectional microphone and the acoustic circuit of the acoustic system.

The reference point for the phase may be changed so that

$$p_1 = p_{01}\epsilon^{j(\omega t)} \tag{3}$$

and

$$p_2 = p_{02}\epsilon^{j(\omega t+\phi_3)}. \tag{4}$$

The phase angle ϕ_3 is a function of the angle of the incident sound as follows:

$$\phi_3 = \phi \cos \theta \tag{5}$$

where θ = angle between the normal to the surface of the ribbon and the direction of the incident sound, and

ϕ = constant phase angle.

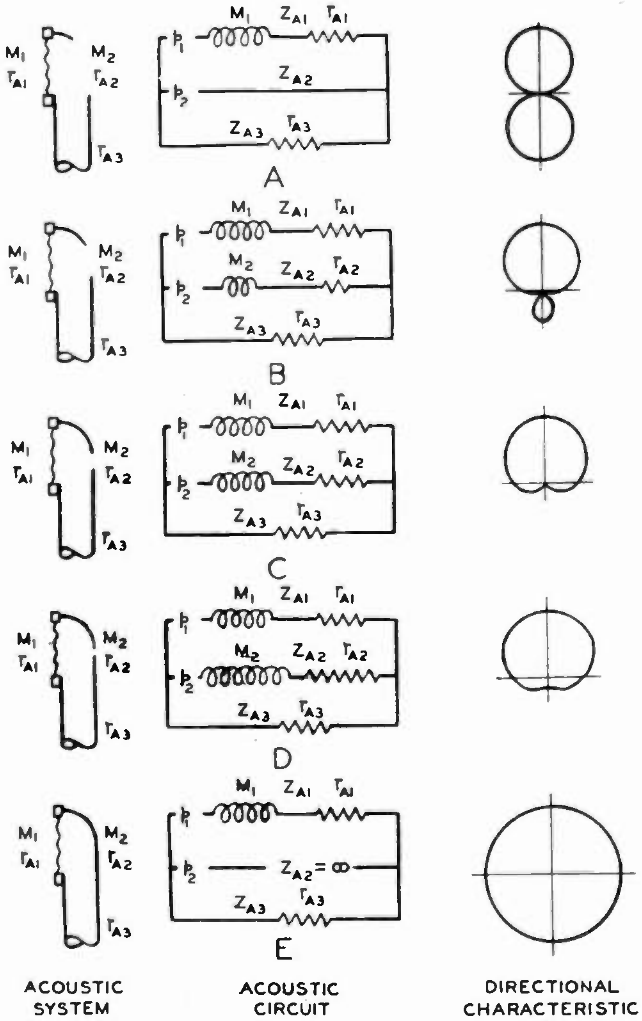

Fig. 3—The acoustic system, the acoustic circuit, and the directional characteristics for various values of the aperture M_2, r_{A_2}.

The volume current of the ribbon due to the pressure p_1 is

$$\dot{X}_1 = \frac{p_1(z_{A2} + z_{A3})}{z_{A1}z_{A2} + z_{A1}z_{A3} + z_{A2}z_{A3}} \tag{6}$$

where $z_{A1} = r_{AA} + j\omega M_A$

$$+\frac{r_{AS} - \omega^2 r_{AS} M_R C_{AR} + i\omega M_S - j\omega^3 M_S M_R C_{AR}}{1 - \omega^2 C_{AR}(M_R + M_S) + j\omega r_{AS} C_{AR}}$$

$z_{A2} = r_{A2} + j\omega M_2$

$z_{A3} = r_{A3}$

r_{AA} = acoustic resistance of the air load on the ribbon

M_A = inertance of the air load on the ribbon

r_{AS} = acoustic resistance of the slit between the ribbon and the pole pieces

M_S = inertance of the slit between the ribbon and the pole pieces

M_R = inertance of the ribbon

C_{AR} = acoustic capacitance of the ribbon

r_{A2} = acoustic resistance of the aperture

M_2 = inertance of the aperture, and

r_{A3} = acoustic resistance of the damped pipe.

Since the acoustic impedance due to the inertance and resistance of the slit between the ribbon and pole pieces is very large compared to the acoustic impedance of the ribbon, these two elements may be neglected. Further, since the resonance of the ribbon is placed below the

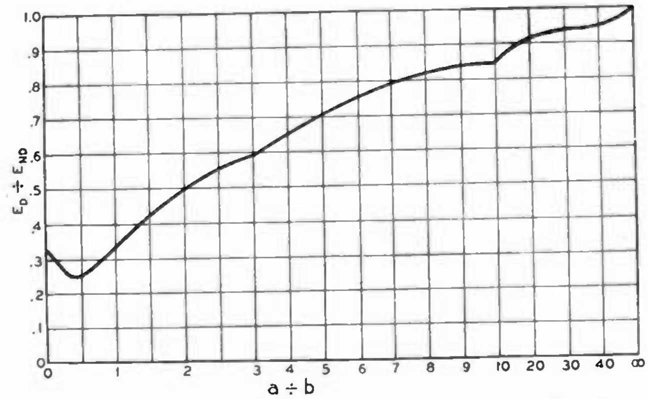

Fig. 4—The ratio of the energy response of the polydirectional microphone having a directional pattern defined by $e = a + b \cos \theta$ as compared to a nondirectional microphone as a function of the ratio $a \div b$. E_{ND}, energy response of a nondirectional microphone. E_D, energy response of the polydirectional microphone.

audible range, the acoustic impedance due to the acoustic capacitance of the ribbon may be neglected for the audible frequency range. Then,

$$z_{A1} = r_{AA} + j\omega M_A + j\omega M_R = r_{A1} + j\omega M_1.$$

The volume current of the ribbon due to the pressure p_2 is

$$\dot{X}_2 = \frac{p_2(z_{A3})}{z_{A1}z_{A2} + z_{A1}z_{A3} + z_{A2}z_{A3}}. \tag{7}$$

The resultant volume current $\dot{X}_R$ of the ribbon is the difference between (6) and (7).

$$\dot{X}_R = \dot{X}_1 - \dot{X}_2. \tag{8}$$

The value of the phase angle ϕ can be determined from the geometry of the microphone. The values of the impedance can be determined from the mass and dimensions of the ribbon, the area of the damped pipe or labyrinth, and the diameter of the aperture in the labyrinth connector.

The directional characteristics of the microphone are controlled by varying the area of the aperture in the labyrinth connector. The effect of varying the aperture can be obtained from the schematic view and the equivalent circuit. See Fig. 3.

In Fig. 3A the aperture is so large that the back of the ribbon is effectively open to the atmosphere. In this

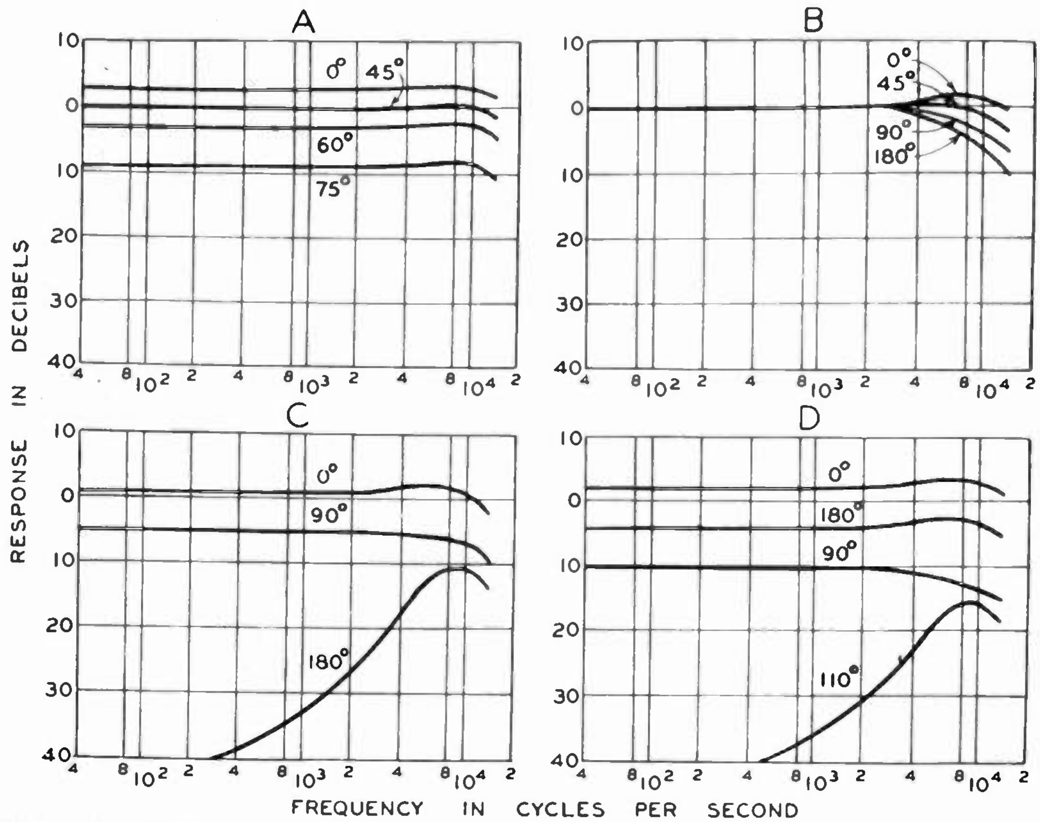

Fig. 5—Theoretical response-frequency characteristics of the polydirectional microphone for aperture setting to yield: *A*, bidirectional; *B*, nondirectional; *C*, cardioid; and *D*, limacon of maximum discrimination.

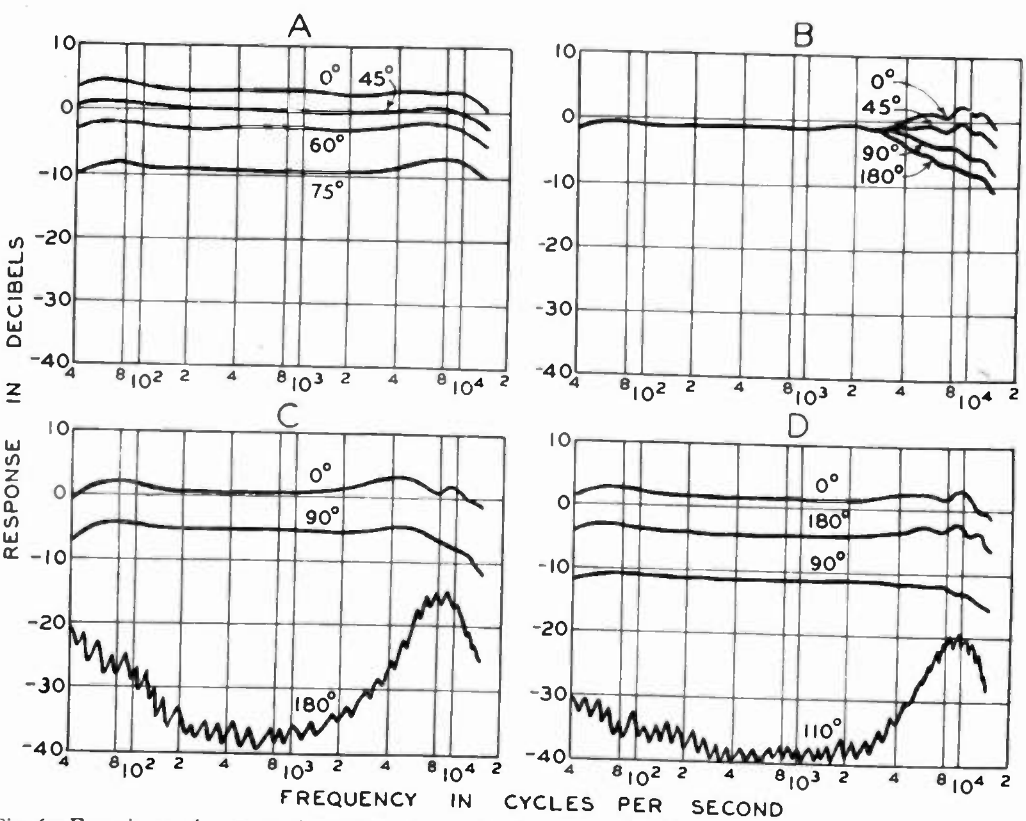

Fig. 6—Experimental response-frequency characteristics of the polydirectional microphone for aperture settings to yield: *A*, bidirectional; *B*, nondirectional; *C*, cardioid; and *D*, limacon of maximum discrimination. Note: In the case of the bidirectional unit the unit is symmetrical, that is the response at 0 and 180 degrees is the same. The response at 90 degrees is so far down that it cannot be measured.

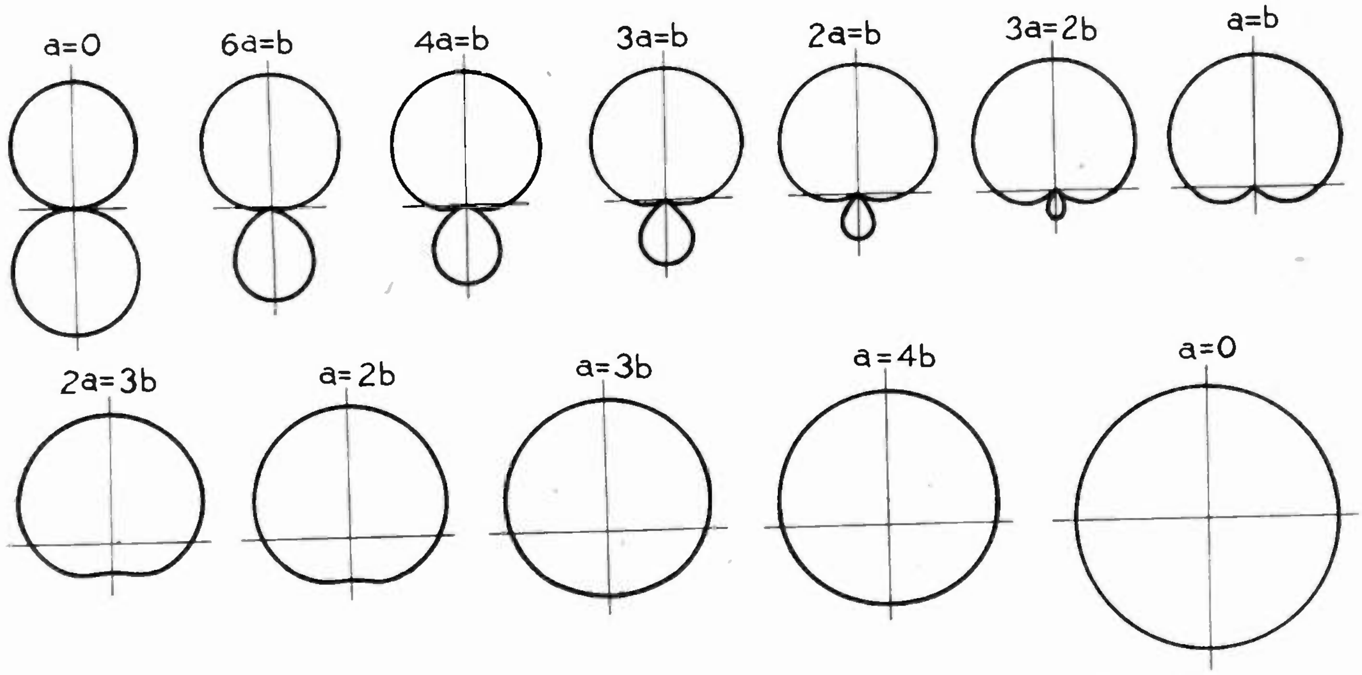

Fig. 7—A few of the single infinity of directional characteristics obtainable with the polydirectional microphone.

case the acoustic impedance z_{A2} is zero. Therefore, the resistance r_{A3} of the labyrinth is effectively short-circuited. The action then is exactly the same as that of the velocity microphone. From (6), (7), and (8) the volume current of the ribbon is

$$\dot{X}_R = \dot{X}_1 - \dot{X}_2 = \frac{(p_1 - p_2)}{z_{A1}}. \tag{9}$$

If the amplitudes of p_1 and p_2 are equal, then

$$\dot{X}_R = \frac{p_{01} - p_{01}\epsilon^{j\phi \cos \theta}}{z_{A1}}. \tag{10}$$

If the angle ϕ is small

$$\dot{X}_R = \frac{p_{01}\phi}{z_{A1}} \cos \theta = \frac{\Delta p \cos \theta}{z_{A1}} \tag{11}$$

where $\Delta p = p_{01}\phi$ the difference in pressure between the two sides of the ribbon.

Equation (11) will be recognized as that of the velocity microphone. The directional characteristic is bidirectional.

In Fig. 3E the aperture is closed. In this case the acoustic impedance z_{A2} is infinite. Under these conditions the pressure p_2 is ineffective. From (6), (7), and (8) the volume current of the ribbon is given by

$$\dot{X}_R = \dot{X}_1 - \dot{X}_2 = \dot{X}_1 = \frac{p_1}{z_{A1} + z_{A3}}. \tag{12}$$

Equation (12) will be recognized as that of the pressure ribbon microphone. The directional characteristic is nondirectional

Using an aperture which may be varied, it is possible to obtain any limacon characteristic between the cosine bidirectional Fig. 3A and the nondirectional characteristic Fig. 3E, as depicted by Fig. 3, parts B, C, and D. The directional characteristic of Fig. 3C is given by

$$e = a + a \cos \theta. \tag{13}$$

This is a cardioid characteristic which is obtained in the two-element unidirectional microphone by making the output of the bidirectional element equal to the non-

Fig. 8—Varacoustic-microphone.

directional unit. The directional characteristic of Fig. 3B is given by

$$e = a/2 + 3a/2 \cos \theta. \tag{14}$$

For a wider directional pickup angle the characteristic of Fig. 3D may be more desirable. This characteristic is given by

$$e = 8a/7 + 6a/7 \cos\theta. \qquad (15)$$

The energy response to random sounds as compared to that of a nondirectional microphone is $\frac{1}{3}$ for the bidirectional characteristic, Fig. 3A, and the cardioid characteristic, Fig. 3C. The energy response for the characteristic of Fig. 3B is $\frac{1}{4}$. This is the maximum value of discrimination obtainable in this microphone. That is, the energy response varies from $\frac{1}{3}$ to $\frac{1}{4}$ and back again to $\frac{1}{3}$ in going from the bidirectional characteristic, Fig. 3A, to the cardioid characteristic of Fig. 3C. The energy response of the characteristic of Fig. 3D given by (15) is 0.39. The energy response varies from $\frac{1}{3}$ to 1 in going from the cardioid characteristic of Fig. 3C to the nondirectional characteristic of Fig. 3E. The general expression for the directional characteristics obtainable with this microphone is

$$e = a + b\cos\theta. \qquad (16)$$

The ratio of the energy response[1,6] of this microphone as compared to a nondirectional microphone for any ratio of a to b is shown in Fig. 4.

As mentioned in the early part of this paper, an important requirement of a sound-collecting system is a directional characteristic which is independent of the frequency. Small deviations from a uniform directional characteristic, particularly in the angular range of high attenuation, are not serious because this region is used for discrimination and not pickup. The theoretical response-frequency characteristics of the polydirectional microphone for aperture settings to yield the following characteristics; bidirectional, nondirectoinal, cardioid, and limacon of maximum discrimination characteristics, are shown in Fig. 5. The experimentally determined characteristics for the same aperture setting are shown in Fig. 6. As in the case of the conventional velocity microphone, the directional characteristics for the bidirectional setting are independent of the frequency, Fig. 6A. For the nondirectional position, see Fig. 6B, the pickup is independent of frequency up to 3000 cycles. The maximum deviation of 10 decibels occurs between 0 and 180 degrees. However, from 0 to 45 degrees the maximum deviation is less than 3 decibels. In the case of the cardioid, Fig. 6C, the shape of the response characteristics in the front hemisphere is practically independent of the angle of the incident sound. The discrimination for 180 degrees is more than 30 decibels in the important range of 100 to 4000 cycles. From 40 to 100 cycles the average attenuation is more than 25 decibels. From 4000 to 14,000 the average attenuation is more than 20 decibels. In the case of the limacon, $a/2+3a/2\cos\theta$, the shape of the response characteristics in the front hemisphere is practically independent of the angle of the incident sound. The attenuation for 110 degrees is more than 30 decibels over the range from 40 to 4000 cycles. This attenuation is somewhat greater than that of the cardioid. It has been found that for the change from the bidirectional characteristic to the cardioid characteristic, the attenuation is progressively greater in the zone of minimum reception as the characteristic recedes from the cardioid form and advances toward the bidirectional characteristic form. However, the attenuation for the angle of 180 degrees in the cardioid position is greater than can be usefully employed in practice. An important factor is uniform response in the useful angular pickup zone. This precludes discrimination in the direct sound. Another important factor is uniform response to random sounds. In all the types of directional characteristics obtainable with this microphone the energy response is practically independent of the frequency. This precludes discrimination in the reverberant or generally reflected sound.

A few of the single infinity of directional characteristics obtainable with this type of microphone are shown in Fig. 7.

A photograph of the commercial design of the polydirectional microphone known as the Varacoustic microphone is shown in Fig. 8. This design was developed by L. J. Anderson.

Representation of Impedance Functions in Terms of Resonant Frequencies*

S. A. SCHELKUNOFF†, ASSOCIATE, I.R.E.

Summary—**The conventional extension of Foster's reactance theorem to electric circuits with an infinite number of degrees of freedom (sections of transmission lines and cavity resonators) leads to series which converge so slowly that often seemingly natural approximations make the series actually divergent. There exist, however, modified expansions which are suitable for numerical calculations and which admit of an attractive physical interpretation. Similar expansions can also be obtained for the transfer impedance. The method of approach is function-theoretic and is based on the assumption that the driving-point impedance and the transfer impedance are analytic functions of the oscillation constant. When these functions are single-valued, they may be represented as certain series of partial fractions or series of functions analogous to partial fractions. In the case of pure reactances each term of these series corresponds to a resonant circuit coupled to the input element or a resonant transducer. The results are approximately true for slightly dissipative systems.**

I. Introduction

IT HAS been shown by Foster[1] that the input impedance (or the "driving-point" impedance) of a network composed of a finite number of self-inductances, mutual inductances, and capacitances is equal to the impedances of properly constructed networks of types shown in Figs. 1 and 2. The number of parallel circuits in Fig. 1 is equal to the number of natural frequencies of the given network when the input terminals are open. The number of series combinations in Fig. 2 is equal to the number of natural frequencies of the network when its input terminals are short-circuited.

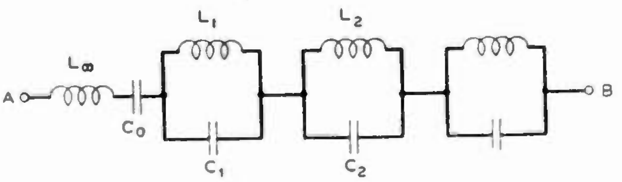

Fig. 1—Representation of a reactive network.

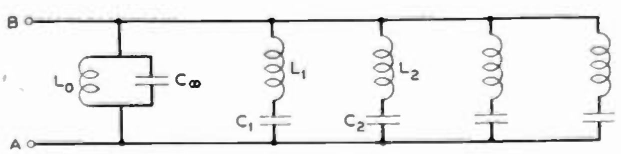

Fig. 2—Representation of a reactive network.

In principle there is no difficulty in extending Foster's theorem to include systems with an infinite number of natural frequencies either by representing certain solutions of Maxwell's equations in terms of appropriate orthogonal functions[2] or by function-theoretic methods.[3] The only question is that of convergence of corresponding analytic expansions. In some instances (in the case of transmission lines, for example), the series are known to converge. Perhaps they always converge if the coefficients are given their exact values; but it is also true that these series often fail to converge if certain seemingly natural approximations in the values of the coefficients are made. Even the expansions which are known to be convergent, converge slowly. In this paper we obtain modified expansions which converge more rapidly and in which reasonable approximations can be made without introducing serious errors. It turns out that these new expansions possess an attractive physical interpretation. Finally, we obtain similar expressions for the transfer impedance.

In view of the practical importance of these analytic representations of impedance and admittance functions, we shall consider the entire problem anew, starting with simple systems but in a way which permits desired generalizations and modifications. In order to avoid approximations that would tend to obscure certain facts which we are anxious to bring out, we shall start with nondissipative systems; then we shall extend the results to dissipative systems.

II. The Admittance of a Simple Series Circuit

Consider a circuit consisting of an inductance L and a capacitance C connected in series as shown in Fig. 3.

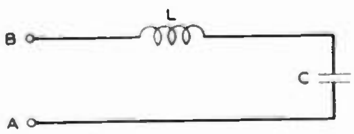

Fig. 3—Parallel-resonant circuit.

Its impedance and admittance across the input terminals A, B are

$$Z = i(\omega L - 1/\omega C), \quad Y = 1/Z = i\omega C/(1 - \omega^2 LC), \quad (1)$$

where ω is the frequency in radians per second. The impedance vanishes and the admittance becomes infinite when

$$\omega = \pm \hat{\omega}, \quad \hat{\omega} = 1/\sqrt{LC}; \quad (2)$$

$\hat{\omega}$ is the natural frequency of the circuit with its input terminals short-circuited as in Fig. 4. Expressing the

* Decimal classification: R140. Original manuscript received by the Institute, June 7, 1943; revised manuscript received, August 9, 1943.

† Bell Telephone Laboratories, Inc., New York 14, N. Y.

[1] Ronald M. Foster, "A reactance theorem," *Bell Sys. Tech. Jour.*, vol. 3, pp. 259–267; April, 1924.

[2] E. U. Condon, "Forced oscillations in cavity resonators," *Jour. Appl. Phys.*, vol. 12, pp. 129–132; February, 1941.
A derivation similar to Condon's is given also in an unpublished report by J. C. Slater.

[3] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Company, Inc., New York, N. Y., 1943.

admittance in terms of $\hat{\omega}$, we have

$$Y = i\omega\hat{\omega}^2 C/(\hat{\omega}^2 - \omega^2) = i\omega/L(\hat{\omega}^2 - \omega^2). \quad (3)$$

The energy content of the circuit at resonance is

$$\mathcal{E} = \tfrac{1}{2}LI^2, \quad (4)$$

where I is the amplitude of the current; therefore

$$L = 2\widehat{\mathcal{E}}, \quad (5)$$

where $\widehat{\mathcal{E}}$ *is the energy stored in the circuit at resonance when the input terminals are short-circuited and the amplitude of the current passing through them is unity.* The admittance may then be expressed as

$$Y = i\omega/2\widehat{\mathcal{E}}(\hat{\omega}^2 - \omega^2). \quad (6)$$

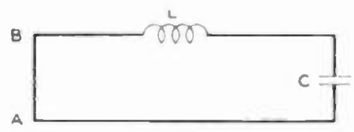

Fig. 4—Series-resonant circuit.

III. The Impedance of a Simple Parallel Circuit

Consider now a parallel combination of an inductance and a capacitance as shown in Fig. 5. Across the terminals A, B the admittance and impedance of the circuit are

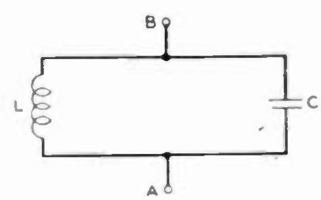

Fig. 5—Series connection of an inductance and a capacitance.

$$Y = i(\omega C - 1/\omega L), \qquad Z = i\omega L/(1 - \omega^2 LC). \quad (7)$$

The admittance vanishes and the impedance becomes infinite when

$$\omega = \pm\bar{\omega}, \qquad \bar{\omega} = 1/\sqrt{LC}; \quad (8)$$

$\bar{\omega}$ is the natural frequency of the circuit when the input terminals are left open. Expressing the impedance in terms of $\bar{\omega}$, we have

$$Z = i\omega/C(\bar{\omega}^2 - \omega^2). \quad (9)$$

The energy of the circuit at resonance is

$$\overline{\mathcal{E}} = \tfrac{1}{2}CV^2, \quad (10)$$

where V is the voltage amplitude across the terminals; therefore

$$C = 2\overline{\mathcal{E}}, \quad (11)$$

and

$$Z = i\omega/2\overline{\mathcal{E}}(\bar{\omega}^2 - \omega^2), \quad (12)$$

where $\overline{\mathcal{E}}$ *is the energy stored in the circuit at resonance when the input terminals are open and the amplitude of the voltage across them is unity.*

IV. Construction of Impedance and Admittance Functions Having Prescribed Simple Poles and Given Residues at These Poles

For a number of parallel circuits connected in series as shown in Fig. 6 the impedance function is simply the

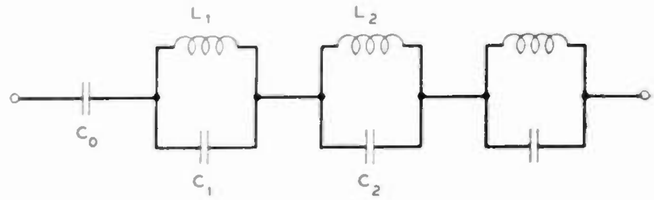

Fig. 6—Impedance with prescribed poles.

sum of impedance functions of the type (9) or (12); thus

$$Z = \sum_n i\omega/C_n(\bar{\omega}_n^2 - \omega^2) = \sum_n i\omega/2\overline{\mathcal{E}}_n(\bar{\omega}_n^2 - \omega^2), \quad (13)$$

where

$$\bar{\omega}_0 = 0, \pm\bar{\omega}_1, \pm\bar{\omega}_2, \cdots, \pm\bar{\omega}_m, \quad (14)$$

are the given simple poles and

$$1/iC_0 = 1/2i\overline{\mathcal{E}}_0, \quad \pm 1/2iC_1 = \pm 1/4i\overline{\mathcal{E}}_1, \cdots, \quad \pm 1/2iC_m = \pm 1/4i\overline{\mathcal{E}}_m \quad (15)$$

are the residues. The isolated capacitor is a "parallel circuit" with an infinite inductance. The point $\omega = \infty$ is a zero of Z.

Now it is well known in the theory of functions of a complex variable that any single-valued function which has the same poles and residues as Z can differ from Z only by an entire function.[4] It is easy to prove that the impedance of a finite network varies ultimately as $i\omega L$, where L may be zero. Thus any other impedance function which has the same singularities in the finite part of the plane as Z can differ from it only by[5] $i\omega L$. Hence if we add a series inductance to the network in Fig. 6, we can say that it has the same impedance as any other network having the same poles and residues.

The coefficient L is twice the maximum energy $\widehat{\mathcal{E}}$ stored in the circuit at infinite frequency when the amplitude of the current through the input terminals is unity. Hence our general impedance function having a finite number of simple poles may be expressed as

$$Z = 2i\omega\widehat{\mathcal{E}} + \sum_n i\omega/2\overline{\mathcal{E}}_n(\bar{\omega}_n^2 - \omega^2). \quad (16)$$

A positive reactive network can have only simple poles and zeros and these must be on the real frequency axis. For if some $\bar{\omega}_n$ has an imaginary part, the amplitude of the corresponding natural oscillation will either increase or decrease indefinitely; this cannot happen unless there is either a concealed source of energy in the network or a resistance. And if any zero or pole is of the order higher than unity, then the impedance $Z(p)$, where $p = \xi + i\omega$, will be negative at some points in the right

[4] A function such as a polynomial or an exponential function which has no singularities in the finite part of the plane.

[5] A constant term is ruled out because the impedance of a purely reactive network is an odd function of ω.

half of the p plane and this also would require concealed sources of power. Thus, for systems with a finite number of degrees of freedom, (16) and the representation in Fig. 1 are quite general.

The poles of Z are the zeros of Y and it is easy to show from (16) that

$$Y'(\bar{\omega}_n) = iB'(\bar{\omega}_n), \quad B'(\bar{\omega}_n) = 4\bar{\mathcal{E}}_n, \text{ if } n \neq 0, \tag{17}$$

and
$$Y'(0) = iB'(0), \quad B'(0) = 2\bar{\mathcal{E}}_0. \tag{18}$$

Thus most of the coefficients in (16) are related to the slopes of the susceptance at its zeros. The first coefficient is related to the ultimate slope of the reactance function.

What we have said about the impedance function of networks in Figs. 1 and 6 can be repeated for admittance functions of the networks shown in Figs. 2 and 7.

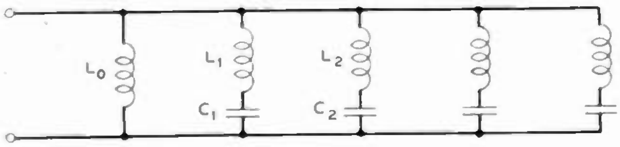

Fig. 7—Impedance with prescribed zeros.

Thus we write

$$\begin{aligned} Y &= i\omega C + \sum_n i\omega/L_n(\hat{\omega}_n^2 - \omega^2) \\ &= 2i\omega\bar{\mathcal{E}} + \sum_n i\omega/2\hat{\mathcal{E}}_n(\hat{\omega}_n^2 - \omega^2). \end{aligned} \tag{19}$$

$$\begin{aligned} Z'(\hat{\omega}_n) &= iX'(\hat{\omega}_n), \quad X'(\hat{\omega}_n) = 4\hat{\mathcal{E}}_n, \quad \text{if} \quad n \neq 0, \\ Z'(0) &= iX'(0), \quad X'(0) = 2\hat{\mathcal{E}}_0. \end{aligned} \tag{20}$$

V. A Simple Illustration

As a simple illustration of how the foregoing expansions can be used let us obtain the input impedance of the network shown in Fig. 8. If the input terminals are

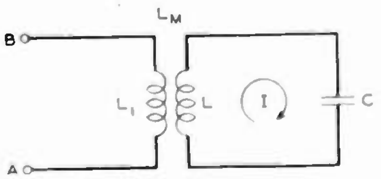

Fig. 8—Inductance coupled to a resonant circuit.

open, the network has only one natural frequency and this is unaffected by the coupling; thus

$$\bar{\omega} = 1/\sqrt{LC}. \tag{21}$$

Next we compute the energy of the system on the assumption that the voltage between A and B is unity. To begin with we express the energy in terms of the current amplitude I; thus

$$2\bar{\mathcal{E}} = LI^2. \tag{22}$$

The voltage across AB is $\bar{\omega}L_M I$ and if this voltage is to be unity, I must equal $1/\bar{\omega}L_M$; hence

$$2\bar{\mathcal{E}} = L/\bar{\omega}^2 L_M^2. \tag{23}$$

Next we calculate the maximum energy $\hat{\mathcal{E}}$ of the circuit at infinite frequency on the assumption that the terminals A, B are short-circuited and the amplitude of the current through them is unity. At infinite frequency the impedance of the capacitor is zero, the impedance of the secondary circuit is $i\omega L$, the impedance coupled into the primary circuit is $\omega L_M^2/iL$, the effective inductance of the primary circuit is $L_1 - L_M^2/L$, and

$$2\hat{\mathcal{E}} = L_1 - L_M^2/L. \tag{24}$$

Substituting (23) and (24) in (16), we have

$$Z = i\omega(L_1 - (L_M^2/L) + i\bar{\omega}^2 L_M^2 \omega/L(\bar{\omega}^2 - \omega^2). \tag{25}$$

For future reference let us note that this equation may be written in the following form:

$$Z = i\omega L_1 + \frac{\bar{\omega}^2 L_M^2}{L}\left[\frac{i\omega}{\bar{\omega}^2 - \omega^2} - \frac{i\omega}{\bar{\omega}^2}\right], \tag{26}$$

that as the frequency approaches zero the contribution from the last terms becomes negligible as compared to that of the first term, and that consequently

$$Z \to i\omega L_1. \tag{27}$$

VI. Impedance and Admittance Functions of Systems With an Infinite Number of Natural Frequencies

If the oscillating system has an infinite number of natural frequencies we have no *a priori* assurance that expansions (16) and (19) will converge. The convergence of these series has nothing to do with the actual existence of functions Z and Y but depends on their analytic properties.

Mittag-Lefler's theorem in the theory of functions of a complex variable states that it is always possible to construct a function which is analytic in the finite part of the plane except at given simple poles and that any other function having the same poles and the same residues at these poles differs from the original function by an additive entire function. Consider an infinite sequence of poles arranged in the order of increasing magnitude

$$\bar{\omega}_1, \bar{\omega}_2, \bar{\omega}_3, \cdots \tag{28}$$

and assume that we found an integer p such that the series

$$\sum_{n=1}^{\infty} \frac{1}{\bar{\mathcal{E}}_n \bar{\omega}_n^{2p}} \tag{29}$$

converges. Next we consider the following expansion

$$\frac{1}{\bar{\omega}_n^2 - \omega^2} = \frac{1}{\bar{\omega}_n^2} + \frac{\omega^2}{\bar{\omega}_n^4} + \frac{\omega^4}{\bar{\omega}_n^6} + \cdots \frac{\omega^{2(p-1)}}{\bar{\omega}_n^{2p}} + \cdots. \tag{30}$$

It can then be shown that the series

$$Z_1 = \sum_1^{\infty} \frac{1}{2\bar{\mathcal{E}}_n}\left[\frac{i\omega}{\bar{\omega}_n^2 - \omega^2} - \frac{i\omega}{\bar{\omega}_n^2} - \frac{i\omega^3}{\bar{\omega}_n^4} - \cdots - \frac{i\omega^{2p-3}}{\bar{\omega}_n^{2(p-1)}}\right] \tag{31}$$

is convergent if $\omega \neq \bar{\omega}_n$.

We believe that in the case of impedance functions encountered in practice the series corresponding to $p = 2$

$$Z_1 = \sum_1^{\infty} \frac{1}{2\bar{\mathcal{E}}_n}\left[\frac{i\omega}{\bar{\omega}_n^2 - \omega^2} - \frac{i\omega}{\bar{\omega}_n^2}\right] \tag{32}$$

will be convergent; but naturally the proof must be supplied in each case when the information regarding the sequence of poles and residues is available.

Since (32) may be written in the form

$$Z_1 = \sum_{n=1}^{\infty} \frac{i\omega^3}{2\mathcal{E}_n \bar{\omega}_n^2(\bar{\omega}_n^2 - \omega^2)}, \tag{33}$$

it is evident that

$$Z_1/i\omega \to 0 \quad \text{as} \quad \omega \to 0, \tag{34}$$

provided of course that (32) converges. Hence if at vanishingly small frequencies the system has an inductance L, then the complete impedance function is

$$Z = i\omega L + \sum_1^{\infty} \frac{1}{2\mathcal{E}_n}\left[\frac{i\omega}{\bar{\omega}_n^2 - \omega^2} - \frac{i\omega}{\bar{\omega}_n^2}\right]. \tag{35}$$

Similarly for the admittance function, we have

$$Y = i\omega C + \sum_1^{\infty} \frac{1}{2\hat{\mathcal{E}}_n}\left[\frac{i\omega}{\hat{\omega}_n^2 - \omega^2} - \frac{i\omega}{\hat{\omega}_n^2}\right]. \tag{36}$$

Here C is the capacitance at vanishingly small frequencies or a measure of the electric energy stored in the circuit when the input terminals are *open* and the voltage across them is unity.

Compare now (35) with (26) which was obtained for the circuit shown in Fig. 8. As we should have expected expansions (16) and (35) are the same if both are convergent. Furthermore, if we write the impedance of the circuit directly in terms of the impedance of the primary and the impedance coupled from the secondary, we have

$$\begin{aligned} Z &= i\omega L_1 + \omega^2 L_M^2/i(\omega L - 1/\omega C) \\ &= i\omega L_1 + i\omega^3 L_M^2/L(\bar{\omega}^2 - \omega^2). \end{aligned} \tag{37}$$

Comparing this equation with (33), we have

$$L_M^2/L = 1/2\mathcal{E}\bar{\omega}^2. \tag{38}$$

Hence we may rewrite (35) as follows

$$\begin{aligned} Z &= i\omega L + \sum_1^{\infty}\left[\frac{i\omega\bar{\omega}_n^2 L_{M,n}^2}{L_n(\bar{\omega}_n^2 - \omega^2)} - \frac{i\omega L_{M,n}^2}{L_n}\right] \\ &= i\omega L + \sum_1^{\infty} \frac{i\omega^3 L_{M,n}^2}{L_n(\bar{\omega}_n^2 - \omega^2)} \end{aligned} \tag{39}$$

and interpret it as the input impedance of the network shown in Fig. 9. Here L is the inductance of the primary circuit which consists of a loop coupled to a succession of independent simple circuits; L_n is the inductance of a typical secondary circuit and $L_{M,n}$ is the mutual inductance between it and the primary circuit.

If the series formed by the first terms in the brackets of (39) is convergent, then (39) may be rewritten as an expansion of the type (16); thus

$$Z = i\omega(L - L') + \sum_1^{\infty} \frac{i\omega\bar{\omega}_n^2 L_{M,n}^2}{L_n(\bar{\omega}_n^2 - \omega^2)}, \tag{39'}$$

$$L' = \sum_1^{\infty} \frac{L_{M,n}^2}{L_n}.$$

Of course, in the immediate vicinity of any natural frequency one particular term is so much larger than the rest that it does not matter whether Z is written in the form (39) or (39′); but off resonance, particularly in the lower-frequency range the situation is quite different. For instance, consider two resonant circuits for which

$$L_1 = L_2 = L \qquad C_1 = C, \qquad C_2 = 0.01C$$
$$\bar{\omega}_1 = \bar{\omega} = 1/\sqrt{LC}, \qquad \bar{\omega}_2 = 10\bar{\omega},$$

and let the self-inductances of the coupling coils and the mutual inductances be equal to $\hat{L}$ and L_M, respectively; then by (39) we have

$$Z = 2i\omega\hat{L} + i\omega^3 L_M^2/L(\bar{\omega}^2 - \omega^2) + i\omega^3 L_M^2/L(100\bar{\omega}^2 - \omega^2).$$

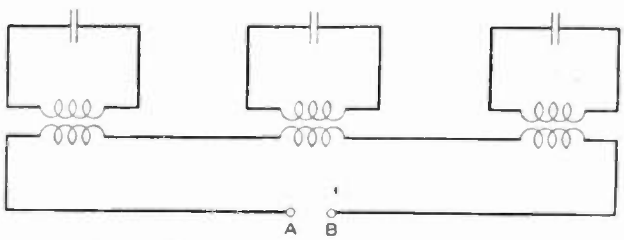

Fig. 9—Representation of a reactive network.

If $\omega = 2\bar{\omega}$, the last term contributes an inductance equal to $0.04L_M^2/L$. This represents a very small effect and our formula is in agreement with our physical intuition which tells us that we are so far below the higher resonant frequency that the effect of the second resonant circuit should be small. On the other hand if we write Z in the form (39′), then we have

$$\begin{aligned} Z = {} & i\omega(2\hat{L} - 2L_M^2/L) + i\bar{\omega}^2\omega L_M^2/L(\bar{\omega}^2 - \omega^2) \\ & + 100 i\bar{\omega}^2\omega L_M^2/L(100\bar{\omega}^2 - \omega^2). \end{aligned}$$

In this case when $\omega = 2\bar{\omega}$ the contribution of the last term exceeds the contribution of the second term.

VII. A Simple Example of an Oscillating System with an Infinite Number of Natural Frequencies

Before passing to applications of the foregoing results to cavity resonators, it will be instructive to consider a case of an oscillating system with an infinite number of

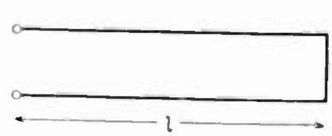

Fig. 10—Section of a transmission lines short-circuited at the far end.

natural frequencies in which the expansion given by (16) indubitably converges. Even in this case the expansion given by (35) is preferable on account of its more rapid convergence. Let us take an ordinary transmission line of length l and let it be short-circuited at one end, Fig. 10. If x is the distance from the open end, then the voltage for a typical oscillation mode is

$$V(x) = \cos\left((2n+1)\pi x/2l\right), \quad n = 0, 1, 2, 3, \cdots. \tag{40}$$

The amplitude has already been adjusted so that $V(0)=1$. The corresponding natural frequencies and wavelengths are

$$\bar{\omega}_n = (2n+1)\pi/2l\sqrt{\bar{L}\bar{C}}, \quad \bar{\lambda}_n = 4l/(2n+1), \tag{41}$$

where $\bar{L},\bar{C}$ are, respectively, the inductance and capacitance per unit length. The energy associated with a typical mode is

$$2\bar{\mathcal{E}}_n = \bar{C}\int_0^l [V(x)]^2 dx = \tfrac{1}{2}\bar{C}l. \tag{42}$$

Therefore by (16) we have

$$Z = \sum_{n=0}^{\infty} \frac{2i\omega}{\bar{C}l(\bar{\omega}_n^2 - \omega^2)} \cdot \tag{43}$$

The direct-current inductance of the line is $\bar{L}l$; hence by (35) we have

$$Z = i\omega\bar{L}l + \sum_{n=0}^{\infty} \frac{2}{\bar{C}l}\left[\frac{i\omega}{\bar{\omega}_n^2 - \omega^2} - \frac{i\omega}{\bar{\omega}_n^2}\right]$$

$$= i\omega\bar{L}l + i\sum_{n=0}^{\infty} \frac{8\bar{L}l\omega^3}{(2n+1)^2\pi^2(\bar{\omega}_n^2 - \omega^2)} \cdot \tag{44}$$

Since (43) and (44) are convergent, we must have

$$\bar{L}l = \sum_{n=0}^{\infty} \frac{2}{\bar{C}l\bar{\omega}_n^2} \cdot \tag{45}$$

This is actually the case.

Let us now obtain (44) directly from (39) in accordance with which we interpret the transmission line as a simple loop having inductance $L=\bar{L}l$ and which is coupled to its own oscillation modes. If the maximum current amplitude is unity, the current distribution in the line for a typical oscillation mode is

$$I_n(x) = \sin((2n+1)\pi x/2l). \tag{46}$$

The inductance for this mode is then

$$L_n = 2\mathcal{E}_n = \bar{L}\int_0^l [I_n(x)]^2 dx = \tfrac{1}{2}\bar{L}l. \tag{47}$$

The mutual inductance is equal to the flux coupled with the loop; hence

$$L_{M,n} = \bar{L}\int_0^l I_n(x)dx = \frac{2\bar{L}l}{(2n+1)\pi} \cdot \tag{48}$$

Substituting from the above equations in (39) we obtain (44).

Similarly we can obtain the following formulas for the admittance function of the transmission line shown in Fig. 10

$$Y = \frac{1}{i\omega\bar{L}l} + \sum_{n=1}^{\infty} \frac{2i\omega}{\bar{L}l(\hat{\omega}_n^2 - \omega^2)}, \quad \hat{\omega}_n = \frac{n\pi}{l\sqrt{\bar{L}\bar{C}}}; \tag{49}$$

$$= \frac{1}{3}i\omega\bar{C}l + \frac{1}{i\omega\bar{L}l} + \sum_{n=1}^{\infty} \frac{2}{\bar{L}l}\left[\frac{i\omega}{\hat{\omega}_n^2 - \omega^2} - \frac{i\omega}{\hat{\omega}_n^2}\right]$$

$$= \frac{1}{3}i\omega\bar{C}l + \frac{1}{i\omega\bar{L}l} + i\sum_{n=1}^{\infty} \frac{2\bar{C}\omega^3 l}{n^2\pi^2(\hat{\omega}_n^2 - \omega^2)} \cdot \tag{50}$$

It is noteworthy that the first two terms of this equation represent an excellent low-frequency approximation to the admittance of a short-circuited transmission line.

VIII. The Input Impedance Across the Terminals of a Loop in a Cavity Resonator

Let us now consider a cavity resonator with a perfectly conducting boundary and a perfectly conducting loop ACB, Fig. 11. In practical calculations it is difficult to avoid approximations; but before making them let us consider an exact mathematical problem so that later we could appraise the nature of our approximations. In order to make the problem definite we assume that the electric charge is transferred from B to A and back by applied forces concentrated in a tubular region which in Fig. 11 is indicated by dotted lines. The impedance of the loop is then the ratio of the applied electromotive force whose positive direction is from B to A, let us say, to the current at A flowing into the loop. The applied electromotive force is equal to the electromotive force of the field taken from A to B along the boundary of the "electric generator."

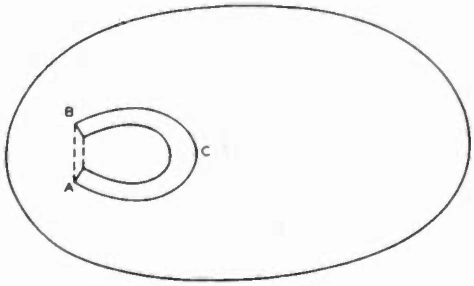

Fig. 11—Loop inside a cavity resonator.

In accordance with the preceding section this impedance may be expressed in the form (35) where $\bar{\omega}_n$ is a typical natural frequency of the system obtained on the assumption that the terminals of the loop are electrically open, $\bar{\mathcal{E}}_n$ is the corresponding energy content when the voltage from A to B along the boundary of the electric generator is unity, and L is the "direct-current inductance" of the loop as affected by the cavity resonator. In computing the latter the generator is to be short-circuited; in the present case this means that on the boundary of the generator we must suppose a perfectly conducting sheet. The terminals of our generator are "electrically open" if there is no current flowing inside the tube occupied by the generator and that consequently the generator is surrounded by a sheet of infinite impedance[6] so that the magnetic intensity tangential to it vanishes. Thus we have a complete set of boundary conditions for obtaining the natural frequencies of the system:

1. The tangential electric intensity vanishes on the boundaries of the cavity resonator and the loop,

2. The tangential magnetic intensity vanishes on the boundary of the "generator region."

For obtaining L the secondary boundary condition is replaced by: The tangential electric intensity vanishes on the boundary of the generator region and the particular solution we are interested in corresponds to $\omega=0$.

[6] Zero conductivity and infinite permeability.

A variant method of expressing the impedance of the loop in a cavity resonator is based on (39). The boundary value problem to be solved is the same as in the preceding method.

The lowest natural frequencies of the complete system may be divided roughly into two groups: 1) the natural frequencies of the cavity as affected by the presence of the loop and 2) the natural frequencies of the loop as affected by the cavity. Naturally such separation is impossible when a natural frequency of the loop is nearly equal to a natural frequency of the cavity. For higher-oscillation modes such proximity is unavoidable and then we can speak only of the natural frequencies of the entire system.

If now the loop is small, the lowest natural frequencies of the cavity are affected but little by the loop and the effect of the natural frequencies on the impedance of the loop is small. This is certainly the case when the length of the loop is so small that the current in it is distributed substantially uniformly. Under these conditions the input impedance of the loop is given by (39) where

$$L_n = \mu \iiint H_n \cdot H_n^* dv,$$

$$L_{M,n} = \mu \iint H_{n,n} dS. \tag{51}$$

The volume integral is extended over the entire cavity and the surface integral over a surface subtended by the loop; H_n is the magnetic intensity for the n^{th} oscillation mode, H_n^* is the conjugate of it and $H_{n,n}$ is the normal component of H_n.

On account of rapid convergence of (39) a good approximation for the input impedance of the loop for frequencies not exceeding the second natural frequency is

$$Z \simeq i\omega L + i\omega^3 L_{M,1}^2/L_1(\bar{\omega}_1^2 - \omega^2) + i\omega^3 L_{M,2}^2/L_2(\omega_2^2 - \omega^2). \tag{52}$$

In the vicinity of the first natural frequency only the first two terms will usually suffice.

IX. The Input Impedance of a Slightly Dissipative System

The foregoing method of expressing impedance functions in terms of the natural frequencies of the system is exact when the system is nondissipative. If the system is slightly dissipative, then we obtain an approximate expression based on retaining the physical picture corresponding to (39); thus we have

$$Z = R + i\omega L + \sum_{n=1}^{\infty} \frac{i\omega^3 L_{M,n}^2}{L_n(\bar{\omega}_n^2 - \omega^2 + i\delta_n\bar{\omega}_n\omega)}, \tag{53}$$

where δ_n is the relative bandwidth for the n^{th} oscillation mode or the reciprocal of the Q

$$\delta_n = \Delta\omega/\bar{\omega}_n = 1/Q_n. \tag{54}$$

X. The Transfer Impedance Across a Cavity Transducer

The foregoing results can readily be extended to cavity transducers, Fig. 12, which may be represented diagrammatically as shown in Fig. 13 with the understand-

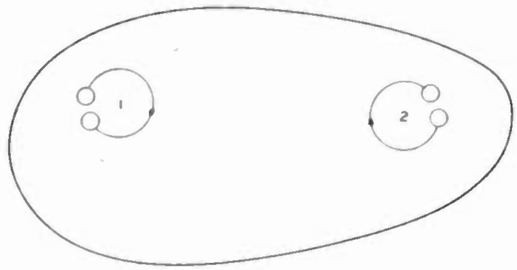

Fig. 12—Two loops inside a cavity resonator.

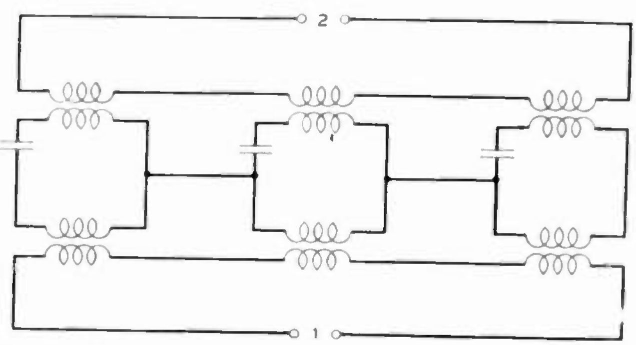

Fig. 13—Representation of two loops inside a cavity resonator.

ing that the number of simple resonant circuits is infinite. Thus the transfer impedance is

$$Z_{12} = i\omega L_{12}^0 + \sum_{n=1}^{\infty} \frac{i\omega^3 L_{1n} L_{2n}}{L_n(\bar{\omega}_n^2 - \omega^2 + i\delta_n\bar{\omega}_n\omega)}, \tag{55}$$

where L_{12}^0 is the low-frequency mutual inductance between the loops in the cavity, L_{1n} is the mutual inductance between the first loop and the cavity, and L_{2n} is the mutual inductance between the second loop and the cavity. If we are interested in frequencies which are of the order of magnitude of the first resonant frequency or higher, then L_{12}^0 is computed for frequencies which are small compared with the first resonant frequency and yet are sufficiently large to make the cavity a good shield. The field is then confined to the interior of the cavity. At really low frequencies L_{12}^0 is affected by the field spreading outside the cavity.

The ratio $L_{M,n}^2/L_n$ is the inductance coupled into the primary from the n^{th} mode at frequencies very much higher than the corresponding resonant frequency and may be referred to as high-frequency coupled inductance. Similarly the ratio $L_{1n} L_{2n}/L_n$ is the high-frequency transfer inductance.

XI. Examples

Consider a parallelepipedal cavity the dimensions of which are a, b, h, and let h be small. Let the inner conductor of a coaxial pair extend into the cavity as shown in Fig. 14. At frequencies for which h is substantially smaller than a quarter wavelength the current in the wire is practically uniform and hence in the direction parallel to h the field is substantially uniform. Hence we need to consider only those oscillation modes for which the electric intensity parallel to h is

$$E_z = \sin(m\pi x/a)\sin(n\pi y/b). \tag{56}$$

When referred to a unit maximum E_z, the energy content $\mathcal{E}$ of the cavity for a typical oscillation mode is

$$2\mathcal{E} = \tfrac{1}{4}\epsilon abh = L_{m,n}{}^S. \tag{57}$$

The amplitude of the voltage induced in the wire is

$$\omega_{m,n}L_{m,n}{}^M = h\sin(m\pi/2)\sin(n\pi\bar{y}/b), \tag{58}$$

where $\bar{y}$ is the distance between the axis of the coupling wire and the nearest cavity wall. The natural frequencies are

$$\omega_{m,n} = \pi v\sqrt{m^2/a^2 + n^2/b^2}, \tag{59}$$

where v is the velocity of light.

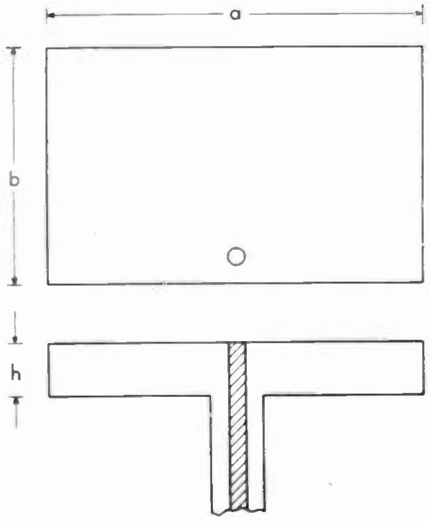

Fig. 14—Top and side views of a parallelepipedal cavity resonator.

From the above equations we have

$$[\omega_{m,n}L_{m,n}{}^M]^2/L_{m,n}{}^S = 4h\sin^2\frac{m\pi}{2}\sin^2\frac{n\pi\bar{y}}{b}\Big/\epsilon ab. \tag{60}$$

Using (16) or the equivalent equation[7] (39′), we obtain

$$Z = \sum_{m,n}^{\infty}\sum \frac{4hi\omega\sin^2\dfrac{m\pi}{2}\sin^2\dfrac{n\pi\bar{y}}{b}}{\epsilon ab(\omega_{m,n}{}^2 - \omega^2 + i\delta_{m,n}\omega_{m,n}\omega)}. \tag{61}$$

If $a=b$, then with respect to convergence this series behaves as

$$\sum\sum 1/(m^2+n^2) \tag{62}$$

which is evidently divergent. It is quite likely that using the exact values for $\omega_{m,n}$ and $L_{m,n}{}^M$ obtained by replacing the conducting wire with a wire of infinite impedance we should obtain a convergent expansion; but this is not practicable.

On the other hand from (39) we have

$$Z = i\omega L + \sum_{m,n}^{\infty}\sum \frac{4hi\omega^3\sin^2\dfrac{m\pi}{2}\sin^2\dfrac{n\pi\bar{y}}{b}}{\epsilon ab\omega_{m,n}{}^2(\omega_{m,n}{}^2-\omega^2+i\delta_{m,n}\omega_{m,n}\omega)}, \tag{63}$$

[7] Including, however, the effect of dissipation.

where L is the direct-current inductance of the wire (of radius r) and is approximately[8]

$$L = (\mu h/4\pi)\cosh^{-1}(2(\bar{y}^2/r^2) - 1). \tag{64}$$

The series is now convergent and is not sensitive to small changes in the coefficients of higher modes.

In the case of a cylindrical cavity (Fig. 15) of radius a and height h, small compared with a, the fields of the lowest oscillation modes are given by

$$E_z = J_n\left(k_{n,m}\frac{\rho}{a}\right)\cos n\phi, \qquad J_n(k_{n,m}) = 0. \tag{65}$$

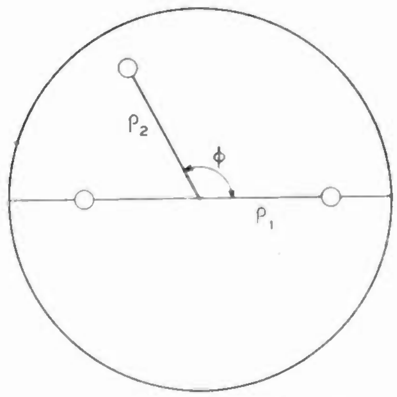

Fig. 15—Top view of a cylindrical cavity resonator.

For the natural wavelengths and frequencies we have

$$\lambda_{n,m} = 2\pi a/k_{n,m}, \qquad \omega_{n,m} = 6\pi\times 10^8/\lambda_{n,m} \tag{66}$$

where

$$\begin{aligned} &k_{0,1} = 2.40, && k_{0,2} = 5.52, && k_{0,3} = 8.65, \cdots\\ &k_{1,1} = 3.83, && k_{1,2} = 7.02, && k_{1,3} = 10.17, \cdots\\ &k_{2,1} = 5.14, && k_{2,2} = 8.42, && k_{2,3} = 11.62, \cdots\\ &k_{3,1} = 6.38, && k_{3,2} = 9.76, && k_{3,3} = 13.02, \cdots \end{aligned} \tag{67}$$

The energy associated with each mode is

$$\begin{aligned} 2\mathcal{E} &= \pi\epsilon ha^2J_1{}^2(k_{0,m}), && \text{if} \quad n = 0,\\ &= \tfrac{1}{2}\pi\epsilon ha^2[J_n'(k_{n,m})]^2, && \text{if} \quad n \neq 0; \end{aligned} \tag{68}$$

and the voltage induced in the coupling wire is

$$V = hJ_n(k_{n,m}\rho_1/a), \tag{69}$$

where ρ_1 is the distance between the axes of the wire and the cavity. Hence the high-frequency coupled inductance is

$$\begin{aligned} \bar{L}_{n,m} &= \frac{(L_{n,m}{}^M)^2}{L_{n,m}{}^S} = \frac{\mu h\lambda_{0,m}{}^2}{4\pi^3a^2}\left[\frac{J_0(k_{0,m}\,\rho_1/a)}{J_1(k_{0,m})}\right]^2, && \text{if} \quad n = 0,\\ &= \frac{\mu h\lambda_{n,m}{}^2}{2\pi^3a^2}\left[\frac{J_n(k_{n,m}\,\rho_1/a)}{J_n'(k_{n,m})}\right]^2, && \text{if} \quad n \neq 0. \end{aligned} \tag{70}$$

The direct-current inductance of the coupling wire is

$$L = (\mu h/2\pi)\cosh^{-1}[(a^2 - \rho_1{}^2 + r^2)/2ar]. \tag{71}$$

If $s_1 = a - \rho_1$ is small compared with a, then approximately

$$\begin{aligned} \bar{L}_{n,1} &= \mu hs_1{}^2/\pi a^2, && n = 0,\\ &= 2\mu hs_1{}^2/\pi a^2, && n \neq 0. \end{aligned} \tag{72}$$

[8] Ignoring the effect of the far walls of the cavity.

A typical high-frequency transfer inductance is

$$\bar{L}_{n,m}{}^{t}=\frac{\mu h\lambda_{0,m}{}^2 J_0(k_{0,m}\,\rho_1/a)\,J_0\,(k_{0,m}\,\rho_2/a)}{4\pi^3a^2[J_1(k_{0,m})]^2},\quad \text{if}\quad n=0,$$
$$=\frac{\mu h\lambda_{n,m}{}^2 J_n(k_{n,m}\,\rho_1/a)\,J_n\,(k_{n,m}\,\rho_2/a)\cos n\phi}{2\pi^3a^2[J_n{}'(k_{n,m})]^2},\ \text{if}\ n\neq 0. \tag{73}$$

For low-frequency transfer inductance we obtain

$$\bar{L}^t=\frac{\mu h}{4\pi}\log\frac{(\rho_1\rho_2/a)^2-2\rho_1\rho_2\cos\phi+a^2}{\rho_1{}^2-2\rho_1\rho_2\cos\phi+a^2}. \tag{74}$$

There exists also a transfer resistance associated with each oscillation mode; this resistance is negligible however in its effect on coupling between the coupling elements.

At resonance the principal terms in the expansions (53) and (55) are

$$Pr(Z)=\bar{\omega}_n\bar{L}_n/\delta_n=\bar{\omega}_nQ_n\bar{L}_n,$$
$$Pr(Z_{12})=\bar{\omega}_nQ_n\bar{L}_n{}^t. \tag{75}$$

There exists also the following relationship between high-frequency inductances coupled into two coupling elements and high-frequency transfer inductance

$$\bar{L}_n{}^t=\sqrt{\bar{L}_n{}'\bar{L}_n{}''}. \tag{76}$$

The quantities $\bar{\omega}_n$, Q_n, $\bar{\omega}_nQ_n\bar{L}_n$ are measurable.

XII. Cavity Resonators as Sections of Wave Guides

Cavities considered in the preceding section may be regarded as sections of wave guides. For example, a coupling wire in a cylindrical cavity of radius a, Fig. 16, sends outward a cylindrical wave which impinges on the lateral wall of the cavity and is reflected from it. For the

Fig. 16—Side view of a cylindrical cavity resonator.

purposes of computing the reflected wave it is convenient to regard the incident wave as emerging from a virtual source along the axis of the cavity. Thus assuming again that the current distribution in the wire is uniform, we have

$$E_z{}^i=-(i\omega\mu I/2\pi)K_0(i\beta\rho')$$
$$=\tfrac{1}{4}\omega\mu I[-J_0(\beta\rho')+iN_0(\beta\rho')], \tag{77}$$
$$\beta=2\pi/\lambda=\omega/v,$$

where ρ' is the distance from the axis of the wire. This function can be expressed in terms of ρ and ϕ, where ρ is the distance from the axis of the cavity and ϕ is the angle measured from the plane passing through the axes of the cavity and the wire; thus we have (for $\rho>\rho_1$)

$$E_z{}^i=\tfrac{1}{4}\omega\mu I\sum_{n=0}^{\infty}\epsilon_nJ_n(\beta\rho_1)[-J_n(\beta\rho)+iN_n(\beta\rho)]\cos n\phi, \tag{78}$$

where

$$\epsilon_0=1,\quad \epsilon_n=2\quad \text{if}\quad n\neq 0. \tag{79}$$

For cavities with perfectly conducting walls, the reflected intensity is obtained from the condition that the total intensity should vanish on the boundary $\rho=a$; thus we have

$$E_z{}^r=-\tfrac{1}{4}\omega\mu I\sum_{n=0}^{\infty}\epsilon_nJ_n(\beta\rho_1)[-J_n(\beta a)+iN_n(\beta a)]\frac{J_n(\beta\rho)}{J_n(\beta a)}\cos n\phi. \tag{80}$$

Hence the total field is

$$E_z=\tfrac{1}{4}i\omega\mu I\left[N_0(\beta\rho')-\sum_{n=0}^{\infty}\epsilon_n\frac{N_n(\beta a)J_n(\beta\rho_1)}{J_n(\beta a)}J_n(\beta\rho)\cos n\phi\right]. \tag{81}$$

Thus we have the following expressions for the input impedance as seen from the wire at $(\rho_1,0)$ and the transfer impedance between this wire and another wire at (ρ_2,ϕ)

$$Z=\tfrac{1}{4}i\omega\mu h\left[-N_0(\beta r)+\sum_{n=0}^{\infty}\epsilon_n\frac{N_n(\beta a)J_n{}^2(\beta\rho_1)}{J_n(\beta a)}\right],$$
$$Z_{12}=\tfrac{1}{4}i\omega\mu h\left[-N_0(\beta\rho_{12})+\sum_{n=0}^{\infty}\epsilon_n\frac{N_n(\beta a)J_n(\beta\rho_1)J_n(\beta\rho_2)}{J_n(\beta a)}\cos n\phi\right]. \tag{82}$$

We note that while the expressions in the preceding section are double series, the present ones are simple series. If we were to take into account the nonuniformity of current distribution in the wire or to consider small coupling loops, we should arrive at the results expressed as double series by the method of this section and as triple series by the method of the preceding section.

In the above expressions the effect of dissipation has been excluded and the formulas are applicable to actual cavities only sufficiently off resonance. Equation (82) could be generalized but there is no practical need for it because near a resonance only one term in the expansions of the type (53) and (55) is needed; it is only off resonance that we may be interested in expansions (82).

In computing (82) it should be remembered that for large n we have approximately

$$-N_n(\beta a)J_n(\beta\rho_1)J_n(\beta\rho_2)/J_n(\beta a)=(1/\pi n)(\rho_1\rho_2/a^2)^n. \tag{83}$$

Furthermore

$$\sum_{n=m}^{\infty}\frac{1}{n}x^n\cos n\phi=-\tfrac{1}{2}\log(1-2x\cos\phi+x^2)$$
$$-\sum_{n=1}^{n=m-1}\frac{1}{n}x^n\cos n\phi. \tag{84}$$

The Dependence of Interelectrode Capacitance on Shielding*

LEONARD T. POCKMAN†, NONMEMBER I.R.E.

Summary—Although it has been known for many years that the interelectrode capacitances of a vacuum tube with a glass envelope depend upon the shielding of the tube, no quantitative information on the extent and character of this dependence has been published. In the present work the general theory is developed with the help of Maxwell's coefficients of capacity and induction. A simple mathematical transformation shows why these cofficients may be used for practical measurements in which all potentials are measured with respect to the earth rather than infinity. The relation between the Maxwell coefficients and the interelectrode and electrode-to-earth capacitances is also developed. The effect of changes in shielding is not necessarily small. The experiments reported show that C_{pf} for the particular triode studied can be made to vary from 0.16 micromicrofarad to 0.41 micromicrofarad by changes in the geometry of the external environment. Furthermore, ungrounding the shield with which C_{pf} was 0.16 micromicrofarad made the "effective" C_{pf} 1.40 micromicrofarad.

ALTHOUGH in 1929 Loughren and Parker[1] pointed out that reproducible results could be obtained in the measurement of interelectrode capacitances only by the use of a standardizable fixture or shield, they did not report any measurements to show how changes in the environment of the tube were correlated with the changes in interelectrode capacitances, nor did they point out the theoretical reasons for the dependence of the interelectrode capacitances on the external environment.

Since that time, no further work has been published on this aspect of the subject of interelectrode capacitance.

The IRE Standards on Electronics, 1938, point out on page 39 that "For most precise results, it is necessary to mount the vacuum tube in a specified way as regards shielding." This statement, though quite correct, nevertheless gives the impression that the effect of changes in shielding on interelectrode capacitance is a small one. Nor, does this statement and the accompanying discussion emphasize the fact that these "most precise results" are not precise in an absolute sense, but only precise for precisely the kind of environment seen by the tube during the measurements. More recently,[2] Barco has touched upon the subject of standardization in the mechanical construction of tube adapters, but gives no quantitative data. The question of the standardization of the total environment of the tube is not discussed.

The lack of quantitative information on the dependence of interelectrode capacitance on the total environment of the tube electrodes, has resulted in considerable confusion on the subject of interelectrode capacitance measurement among practising radio engineers, and is also reflected in the Government specifications on the subject.

The measurements to be reported here have all been carried out on one particular triode. It is not the intention of the writer to give the impression that the large variations found with this triode are necessarily typical of all triodes with glass envelopes, but rather to present the measurements as a striking example of how strongly the interelectrode capacitances of a particular tube may depend upon the external tube environment. The approximate physical dimensions of the tube and the circuit layout will be given later, and from a comparison of these with those of any other tube and with the help of the general theory to be presented, an estimate can be made of the relative magnitudes of the variations to be expected.

On the theoretical side, it is illuminating to analyze the problem with the help of the Maxwell[3,4] coefficients of capacity and induction. If this is done, the analysis is simple and straightforward from a mathematical point of view and brings out sharply the physical aspects of the problem. And this approach, through its emphasis on the actual geometrical system involved, avoids the blurring of important physical details. On the other hand, an electrically equivalent network, unless applied with full knowledge of its limitations is apt to give the false impression that the interelectrode capacitances are independent of the capacitances to earth.

The results of this analysis may then be easily transformed into a mathematically equivalent network of six 2-plate condensers[5] (for a triode) which are needed for the application of the results in circuit analysis. The correlation of the equivalent network of condensers and the Maxwell coefficients also will make it possible to take advantage of a theorem about the coefficients which has been proved by Maxwell, and which turns out to be of singular importance for the present problem.

Unfortunately, as usually presented, these coefficient are treated much more abstractly than need be, by talking about "n" conducting bodies isolated in free space with their potentials all measured with respect to infinity. Thus the impression that these coefficients are of little earthly use is easily obtained. Or, if an

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[3] James Clerk Maxwell, "Electricity and Magnetism," vol. 1, Clarendon Press, Oxford, England, 1892, chapter 3.

[4] L. Page and N. Adams, "Principles of Electricity," D. Van Nostrand Company, New York, N. Y., 1931, p. 63.

[5] E. T. Hoch, "A bridge method for the measurement of interelectrode admittance in vacuum tubes," Proc. I.R.E., vol. 16, pp. 487–493; April, 1928.

application of the coefficients is made, the potential of the earth is quite arbitrarily taken to be zero. In order to get rid of both the abstract and arbitrary aspects of these coefficients, it is only necessary to measure the difference between the potentials of the "n" bodies under consideration and the potential of the earth (let the earth be the $(n+1)$ st body) and to forget about the absolute charge on the $n+1$ bodies and think only about the *change* in the charge on each body corresponding to the changes of potential of the "n" bodies with respect to the potential of the earth. In actual experiments, of course, these changes in charge and differences of potential are what are actually measured.

In order to understand this use of the coefficients with potentials measured with respect to the earth and charges measured as the difference between the charges on the bodies when earthed and the charges on the same bodies at potentials different from earth potential, it is first necessary to understand how they are used when potentials are measured with respect to infinity and the charges are the absolute or net charges on each body.

The main ideas can be introduced without loss of generality by considering the system of four conducting bodies formed by the elements of a triode and the earth. The charge on each of the four conducting bodies can be written as a linear function of the potentials of each body with respect to infinity. If the cathode, grid, plate, and earth are denoted respectively by the subscripts 1, 2, 3, and e, the sytem is described by the following equations:

$$\begin{aligned} Q_1 &= C_{11}V_1 + C_{12}V_2 + C_{13}V_3 + C_{1e}V_e \\ Q_2 &= C_{21}V_1 + C_{22}V_2 + C_{23}V_3 + C_{2e}V_e \\ Q_3 &= C_{31}V_1 + C_{32}V_2 + C_{33}V_3 + C_{3e}V_e \\ Q_e &= C_{e1}V_1 + C_{e2}V_2 + C_{e3}V_3 + C_{ee}V_e \end{aligned} \tag{1}$$

where Q_1 = net charge on body 1; Q_2 = net charge on body 2; etc.

V_1 = potential of body 1 with respect to infinity; V_2 = potential of body 2 with respect to infinity; etc.

The coefficients having identical subscripts are called "coefficients of capacity" and those having two differing subscripts, "coefficients of induction." Both kinds of coefficients are constants independent of charges and potentials and depend only on the geometry of the system and the dielectric constant of the medium.

The physical meaning of the coefficients may be seen by supposing, for example, that the system is charged in such a way that all potentials are zero except V_1 which is greater than zero. Lines of force will then start from body 1, with some going to infinity and the remaining going to the other three bodies. Since lines of force start on positive charge and terminate on negative charge, this means that Q_1 is positive and Q_2, Q_3, and Q_e are negative.

From these considerations it follows that

$$C_{11} = Q_1/V_1 \text{ with } V_2 = V_3 = V_e = 0.$$

Or in words, C_{11} is equal to the charge on body 1 divided by the potential of body 1 when the other bodies of the system are at zero potential. Since $Q_1>0$ and $V_1>0$, $C_{11}>0$. Similarly, $C_{21}=Q_2/V_1$ with $V_2=V_3=V_e=0$. Or in words C_{21} is equal to the charge on body 2 divided by the potential on body 1 when the potentials of all bodies in the system are zero except that of body 1. Since $Q_2<0$ if $V_1>0$, then $C_{21}<0$.

Thus, it can be shown quite generally that all coefficients of capacity (C_{ii}'s) are greater than zero and that all coefficients of induction (C_{ij}'s) are equal to[6] or less than zero. It can also be proved quite generally that $C_{ij}=C_{ji}$.

Maxwell has shown that for a general "n" body system the coefficient of capacity of any given body is larger than or equal to the sum of the absolute values of the coefficients of induction of this body with respect to the other $n-1$ bodies of the system. Thus, in the case of C_{11} in the above four-body system, this means that $C_{11} \geqq |C_{12}| + |C_{13}| + |C_{1e}|$.

By inspection of (1), it can be seen that, physically, this inequality means that if the system is charged in such a way that $V_1>0$ and $V_2=V_3=V_e=0$ then, $Q_1 \geqq -(Q_2+Q_3+Q_e)$, i.e., in all but special cases some of the lines which start from Q_1 fail to terminate on $(Q_2+Q_3+Q_e)$ and go off to a negative charge at infinity.

In all practical problems bodies 1, 2, and 3 are very small compared to the earth and are located at a distance from the earth's surface which is also small compared to the radius of the earth. Because of this fact, C_{11} is very nearly equal to $|C_{12}| + |C_{13}| + |C_{1e}|$. In fact by a simple application of the theory of electrostatic images it can be shown that C_{11}, C_{22}, or C_{33} is larger than $|C_{12}| + |C_{13}| + |C_{1e}|$, $|C_{21}| + |C_{23}| + |C_{2e}|$, or $|C_{31}| + |C_{32}| + |C_{3e}|$, respectively, by a factor of the order of magnitude $(R+L)/R$ where R = radius of earth and L = means distance of body from surface of earth.

This means that the following equalities obtain with a far higher accuracy than that of the best possible capacitance measurements:

$$\begin{aligned} C_{11} &= |C_{12}| + |C_{13}| + |C_{1e}| \\ C_{22} &= |C_{21}| + |C_{23}| + |C_{2e}| \\ C_{33} &= |C_{31}| + |C_{32}| + |C_{3e}|. \end{aligned} \tag{2}$$

For practical problems the exact magnitude of C_{ee}, the coefficient of capacity of the earth, is of no importance and unlike the three other coefficients of capacity is tremendously larger than the sum of the absolute values of its three coefficients of induction.

If the filament, grid, and plate are all earthed so that $V_1=V_2=V_3=V_e$, no lines of force join the four bodies, but all of them send lines to infinity. These lines correspond to the fact that the net charge of the earth is not zero.

Although the net charge on the earth is certainly of no significance as far as ordinary capacitance measurements are concerned, this does not mean that these

[6] The coefficient of induction between any two bodies is obviously zero if a third conducting body completely surrounds one of them.

Maxwell equations must be discarded as of merely academic interest. They can be brought down to earth very simply. The equation for Q_1 can be rewritten as follows:

$$\begin{aligned} Q_1 = {} & C_{11}(V_1 - V_e) + C_{12}(V_2 - V_e) + C_{13}(V_3 - V_e) \\ & + C_{1e}(V_e - V_e) \\ & + [C_{11}V_e + C_{12}V_e + C_{13}V_e + C_{1e}V_e]. \end{aligned}$$

This equation is identically the same as the original equation for Q_1. Because it may be safely assumed that V_e is not sensibly altered[7] by the slight redistribution of charge corresponding to the changing of filament, grid, and plate from earth potential to potentials differing from that of the earth by $(V_1 - V_e)$, $(V_2 - V_e)$, and $(V_3 - V_e)$, the terms in the square brackets are evidently equal to the charge on the filament when filament, grid, and plate are earthed. If this charge be denoted by Q_{1e}, the equation can then be written

$$\begin{aligned} Q_1 - Q_{1e} = {} & C_{11}(V_1 - V_e) + C_{12}(V_2 - V_e) \\ & + C_{13}(V_3 - V_e). \end{aligned}$$

Now if q_1 is taken to denote the change in the charge on body 1 caused by changing bodies 1, 2, and 3 from earth potential by amounts $V_1 - V_2$, $V_2 - V_e$, and $V_3 - V_e$ respectively and v_1, v_2, and v_3, are taken to denote these differences in potential, the equation may be written

$$q_1 = C_{11}v_1 + C_{12}v_2 + C_{13}v_3$$

and similarly:

$$\begin{aligned} q_2 &= C_{21}v_1 + C_{22}v_2 + C_{23}v_3 \\ q_3 &= C_{31}v_1 + C_{32}v_2 + C_{33}v_3 \\ q_e &= C_{e1}v_1 + C_{e2}v_2 + C_{e3}v_3. \end{aligned} \tag{3}$$

It should be noted that this transformation does not require that the potential of the earth be known or that it be arbitrarily taken as zero. Furthermore, the coefficients of capacity and induction retain all of the properties which they enjoy when the potentials are measured with respect to infinity and the charges are taken to be the absolute or net charges. Likewise the charges q_i and the voltages v_i, though not absolute, are precisely what are measured in all practical circuit problems.

A case of particular importance for practical measurements of interelectrode capacitance can be nicely treated with the aid of the Maxwell equations as reformulated above. It is the case in which the earth completely surrounds the other three bodies; i.e. the case in which an earthed conducting shield completely surrounds the tube and that portion of the tube lead wires lying within the main body of the shield. In this very important special case it can be seen with the help of Faraday's ice-pail experiment[4] that q_1, q_2, and q_3 are not only the change of charge on bodies 1, 2, and 3 respectively caused by the changes in potential of v_1, v_2, and v_3, but are also the absolute or net charges on each of the three bodies. Likewise q_e, while not the absolute or net charge on the earth is the absolute or net charge on the "inside surface of the earth." Also, for this special system, the relations between the coefficients of capacity and the coefficients of induction, given by (2), hold exactly, because no lines of force can travel from any of the enclosed bodies to infinity. Furthermore, the earth seen by the tube in this case is ideal in the sense that it is conducting and therefore equipotential. Also, the geometry of the system as a whole is entirely insensitive to disturbances on the outside of the shield and this is of particular moment for purposes of standardization as will be illustrated by the experiments to be reported.

With the aid of the foregoing, the relation between the Maxwell coefficients and the six two-plate capacitances of the equivalent capacitance network shown in Fig. 1 can now readily be obtained. Suppose that bodies

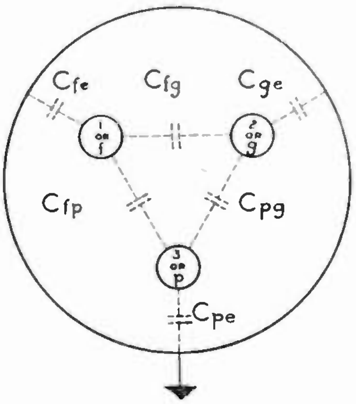

Fig. 1—Equivalent six-condenser network of a triode.

1, 2, and 3 are charged to potentials v_1, v_2, and v_3 with respect to earth. Then, in terms of the equivalent network,

$$\begin{aligned} q_1 &= v_1C_{fe} + (v_1 - v_2)C_{fg} + (v_1 - v_3)C_{fp} \\ q_2 &= v_2C_{ge} + (v_2 - v_1)C_{fg} + (v_2 - v_3)C_{pg} \\ q_3 &= v_3C_{pe} + (v_3 - v_1)C_{fp} + (v_3 - v_2)C_{pg} \\ -q_e &= v_1C_{fe} + v_2C_{ge} + v_3C_{pe}. \end{aligned} \tag{4}$$

Taking (3) for the q's and substituting for the C_{ii}'s in terms of the C_{ij}'s with the aid of (2), gives the following set of equations for the q's in terms of the Maxwell coefficients:

$$\begin{aligned} q_1 &= v_1(-C_{1e}) + (v_1 - v_2)(-C_{12}) + (v_1 - v_3)(-C_{13}) \\ q_2 &= v_2(-C_{2e}) + (v_2 - v_1)(-C_{21}) + (v_2 - v_3)(-C_{23}) \\ q_3 &= v_3(-C_{3e}) + (v_3 - v_1)(-C_{13}) + (v_3 - v_2)(-C_{32}) \\ -q_e &= v_1(-C_{1e}) + {}_2(v - C_{2e}) + v_3(-C_{3e}). \end{aligned} \tag{5}$$

By comparing this set of equations for the q's with the preceding set of equations, it can be seen that the coefficient of induction between two conducting bodies is equal to the negative of the equivalent capacitance between the bodies.[8] Thus, for the case of a triode and earth,

[7] With the help of the theory of electrostatic images, it can be shown that if a charge is transferred from the earth (of radius R) to another conducting sphere (of radius r) with its center at a distance L from the earth's surface, then

$\dfrac{\text{change in } V_e}{\text{change in } V_r} = \dfrac{rL}{R^2}$, i.e., for all practical purposes, V_e does not change.

[4] *Loc. cit.*, pp. 7 and 8.

[8] This relation was first pointed out by G. A. Campbell in his classic paper on "Direct capacity measurement," *Bell Tech. Jour.*, vol. 1, p. 18; July, 1922.

$$C_{21} = C_{12} = -C_{fg} \qquad C_{e1} = C_{1e} = -C_{fe}$$
$$C_{31} = C_{13} = -C_{fp} \qquad C_{e2} = C_{2e} = -C_{ge} \qquad (6)$$
$$C_{32} = C_{23} = -C_{gp} \qquad C_{e3} = C_{3e} = -C_{pe}$$

and, although C_{1e}, C_{2e}, and C_{3e} cannot be measured directly by any feasible experiment, they can be obtained in terms of measured coefficients with the help of (2). Consequently,

$$C_{fe} = C_{11} + C_{12} + C_{13}$$
$$C_{ge} = C_{22} + C_{21} + C_{23} \qquad (7)$$
$$C_{pe} = C_{33} + C_{31} + C_{32}.$$

Combining these relations with those above gives the following relations between the coefficients of capacity and the equivalent capacitances of the six-condenser network:

$$C_{11} = C_{fe} + C_{fg} + C_{fp}$$
$$C_{22} = C_{ge} + C_{gf} + C_{gp} \qquad (8)$$
$$C_{33} = C_{pe} + C_{pf} + C_{pg}.$$

The foregoing relationships between the interelectrode and electrode to earth capacitances and the coefficients of capacity and the coefficients of induction make it possible to take advantage of a theorem proved by Maxwell[3] which is of singular importance for the understanding of the variations of interelectrode capacitance found in the measurements to be reported. The theorem states that the addition of a new conducting body to a system of conducting bodies increases all of the coefficients of capacity of the original system and decreases all of the coefficients of induction. The proof is not only general and rigorous, but is beautifully simple. In order to follow the argument, however, it is necessary to recall that the coefficient of capacity of a body is equal to the charge on the body divided by its potential when the potentials of the other bodies are all zero. It is therefore numerically equal to its charge when its potential is unity and the potentials of all the other bodies are zero. Similarly, the coefficients of induction between the bodies at zero potential and the body at unit potential are numerically equal to and have the same sign as their respective charges. Maxwell's theorem follows:

"Let us suppose that A (any body in original system) is at potential unity and all the rest at potential zero. Since the charge of the new conductor is negative, it will induce a positive charge on every other conductor, and will therefore increase the positive charge of A, and diminish the negative charge of each of the other conductors."

Since placing a shield around a tube is equivalent to adding a new body to the original system, it follows, therefore, that the interelectrode capacitances of a shielded tube are smaller than those of the same tube unshielded. Likewise, decreasing the size of the shield which surrounds the tube is equivalent to keeping the old shield and adding a new body (the smaller shield) between the original shield and the tube. Consequently, if one shield, which completely surrounds the tube, is replaced by another shield, which completely surrounds the tube and whose inner surface lies everywhere within the inner surface of the original shield (or in part coincides with it), the interelectrode capacitances will be smaller with the smaller shield. And since C_{11}, for example, will be increased and as shown above $C_{11} = C_{fe} + C_{fg} + C_{fp}$ it follows that the increase in the capacitance of the filament to earth more than offsets the sum of the decreases in the interelectrode capacitance between the filament and grid and plate, respectively. The corresponding statement may also be made about the increase in C_{ge} and C_{pe}.

The foregoing considerations are quite general in the sense that they apply to the measurement of interelectrode capacitance irrespective of the particular circuit that happens to be used.

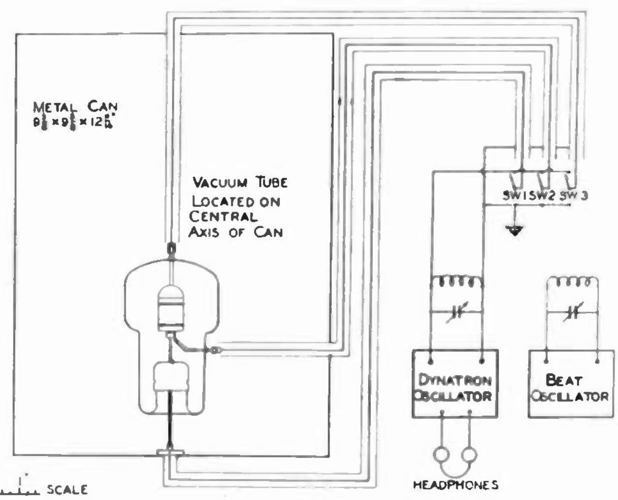

Fig. 2—Scale drawing of tube and shielding used in first series of measurements.

The present measurements of capacitance have been made on one and the same triode surrounded with various kinds of shielding. The method of measurement was first used in this laboratory by P. A. Ekstrand[9] for the determination of interelectrode capacitance and for the simultaneous determination of the capacitance of each electrode to earth. This method of measurement is a substitution method in which a radio-frequency oscillator, adjusted to a particular frequency with the tube out, is brought back to the initial frequency with the tube in, by reducing the capacitance of a variable standard condenser. The return to the initial frequency is determined by listening to the beat frequency between this oscillator and another oscillator loosely coupled to it. In Fig. 2 the circuit diagram is shown. Six measurements are required for the calculation of the six constants of the system. From the same set of measurements, the Maxwell coefficients can also be determined. As a method for the determination of the three interelectrode capacitances only, it is considerably more tedious than

[3] *Loc. cit.*, p. 118.

[9] Now with the Navy Department.

any of the methods described in the IRE Standards,[10] but it makes possible the determination of the six Maxwell coefficients or of the equivalent network by a minimum number of measurements.

A seventh measurement serves as an internal check on the readings and the computations.

The shield and the triode are both drawn to the scale indicated in Fig. 2. The shield shown is one of the shields used in the measurements to be reported—a standard five-gallon gasoline can, cut so that the top is removable.

When the oscillator was adjusted to zero beat with the tube out, the portion of the lead wires within the can and, of course, their coaxial shields, were placed so as to be in the same position with respect to each other and the main body of the shield, i.e., the inner surface of the can, as they would be when connected to the tube. The three single-pole double-throw switches facilitate the seven distinct connections used for the seven measurements. As pointed out in the theoretical discussion, the shielding or lack of shielding of these switches and lead wires lying outside of the grounded equipotential surface which surrounds the tube, have no influence on the coefficients of induction and of capacity of this closed system. These coefficients and hence the interelectrode capacitances and the earth-electrode capacitances are completely determined by the geometry and the dielectric constant of the closed system. The grounded equipotential surface which surrounds the tube in Fig. 2 is evidently that surface formed by the inside of the five-gallon can and the outside of that portion of the lead-wire shields lying within the can.

It should be noted, however, that if the circuit wires lying outside of the tube shield are not shielded by grounded conductors, the operator must be careful that the frequency is not altered by the motion of his body. Care was always taken to see that this was the case. For routine work, of course, such shielding of external leads is a necessary convenience.

If the seven directly measured capacitances are denoted by $C_1, C_2, \cdots C_7$ the connections corresponding to each can be obtained from Table I.

TABLE I

	Elements above Ground Potential	Elements Grounded
C_1	p, g, f (or 3, 2, 1)	
C_2	p	g, f
C_3	g	p, f
C_4	f	g, p
C_5	p, g	f
C_6	g, f	p
C_7	p, f	g

Seven simultaneous equations, of which any six are independent, may be set up by equating the C's to the appropriate combination of Maxwell coefficients or of the six equivalent capacitances.

[10] IRE Standards on Electronics, 1938, pp. 38–44.

When set up in terms of the Maxwell coefficients, the equations assume an extremely simple form and may be solved by inspection of the last six equations: the first equation then provides an over-all check.

Thus, for example, the first equation is obtained by noting that $C_1=(q_1+q_2+q_3)/v$.

Substituting for the q's with the help of (3), and noting that in the measurement of C_1, $v_1=v_2=v_3=v$, leads to

$$C_1 = C_{11} + C_{22} + C_{33} + 2C_{12} + 2C_{13} + 2C_{23}.$$

Similarly,

$$\begin{aligned} C_2 &= C_{33} \\ C_3 &= C_{22} \\ C_4 &= C_{11} \\ C_5 &= C_{22} + C_{33} + 2C_{23} \\ C_6 &= C_{11} + C_{22} + 2C_{12} \\ C_7 &= C_{33} + C_{11} + 2C_{13}. \end{aligned} \tag{9}$$

Solving for the six coefficients in terms of the measured capacitances and making use of the relations between the coefficients of induction and the interelectrode capacitances given in (6) leads to the following equations for the coefficients of induction and the interelectrode capacitances in terms of the measured capacitances:

$$\begin{aligned} C_{23} &= -C_{gp} = \frac{C_5 - (C_2 + C_3)}{2} \\ C_{21} &= -C_{gf} = \frac{C_6 - (C_3 + C_4)}{2} \\ C_{31} &= -C_{pf} = \frac{C_7 - (C_2 + C_4)}{2}. \end{aligned} \tag{10}$$

Similarly, the following expressions for the three capacitances to earth are obtained by combining (7), (9), and (10):

$$\begin{aligned} C_{fe} &= \frac{C_6 + C_7 - (C_2 + C_3)}{2} \\ C_{ge} &= \frac{C_5 + C_6 - (C_2 + C_4)}{2} \\ C_{pe} &= \frac{C_5 + C_7 - (C_3 + C_4)}{2}. \end{aligned} \tag{11}$$

As mentioned earlier, the measurements reported in Table II have all been carried out on one and the same triode by the method described above. In the table, each row of data was obtained from measurements made on this triode surrounded by the particular physical environment described in the first column. The tube and its physical environment in the first series of measurements are shown to scale in Fig. 2. The second series was made with the same orientation of lead wires as on the interior of the can in Fig. 2, but with a one-gallon benzene can ($4\times6\frac{3}{4}\times9\frac{1}{4}$ inches) instead of the five-gallon gasoline can. The tube was also located at the same distance from the bottom of the can and on the central axis of the can. The bakelite adaptor was used for the filament prongs. In one of the series of measurements it was possible to eliminate completely all of the

TABLE II

CAPACITANCE MEASUREMENTS ON ONE AND THE SAME TRIODE: TUBE COMPLETELY SURROUNDED BY A GROUNDED METAL SHIELD

"Earth" seen by triode	C_{11}	C_{22}	C_{33}	C_{fe}	C_{ge}	C_{pe}	$C_{gp}=-C_{23}$	$C_{fg}=-C_{12}$	$C_{pf}=-C_{13}$	C_{gf}/C_{pf}
1. Inside surface of five-gallon can plus those parts of lead wire shields on interior of can (see Fig. 2)	5.34	6.84	4.52	1.29	0.70	1.61	2.50	3.64	0.41	8.9
2. Same as "1" but with a one-gallon can (see text)	5.54	6.91	4.80	1.61	0.87	2.01	2.45	3.59	0.34	10.6
3. Mainly the inside surface of a close fitting shield (see Fig. 3)	7.15	7.78	5.93	3.67	2.29	3.60	2.17	3.32	0.16	20.7
"EFFECTIVE" CAPACITANCE MEASUREMENTS ON SAME TRIODE: TUBE ENVIRONMENT HETEROGENEOUS										
4. Exactly as in "3" but with shield electrically floating	5.98	7.31	4.91	0.48	0.35	0.65	2.86	4.10	1.40	—
5. Grounded metal plate (7×9 inches) about 1 inch below glass envelope plus grounded lead-wire shields, plus general environment of laboratory	5.33	6.80	4.42	1.23	0.61	1.45	2.53	3.66	0.44	—
6. Same as "5," but with lead-wire shields connected to respective lead wires	5.18	6.68	3.69	0.71	0.01	0.37	2.76	3.91	0.56	—

bakelite without otherwise changing the system. A direct comparison of measurements made with and without the bakelite revealed no difference greater than 0.01 micromicrofarad, the approximate limit of error in all the measurements.

The third series of measurements was made with a close-fitting grounded shield essentially like that shown in Fig. 3. Since the tube had to be put into and removed

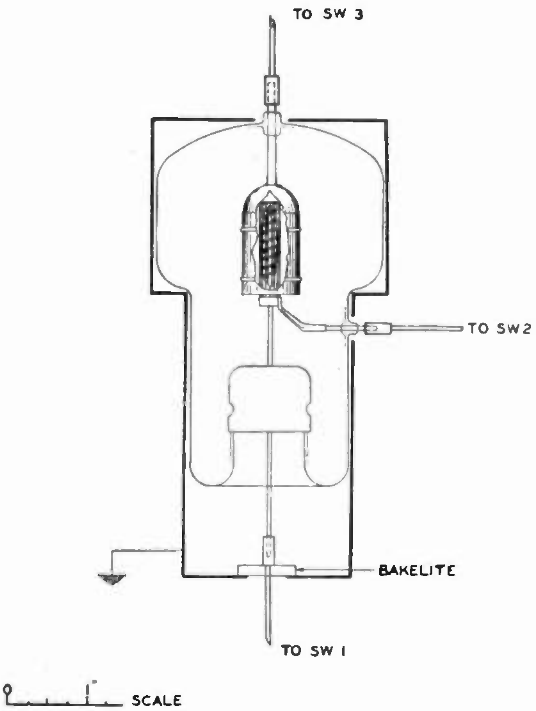

Fig. 3—Tube with close-fitting shield used in third and fourth series of measurement.

from the shield, the actual shield was made up of several parts which could be separated and reassembled quickly. The details of the tube construction are also shown to scale in this figure. The stem shield was of course electrically connected to one of the filament leads.

These first three measurements are grouped together under one main heading because only for these three cases can it be said that the measurements determine the coefficients of capacity and induction as defined above. For these three cases the geometry and the dielectric constant of the system enclosed within the shields completely determine capacitative constants of the system.

In the fourth case, with the close-fitting shield electrically floating, the six measurements do determine a set of six possibly useful capacitative constants for describing this particular system with a floating shield but no longer are these constants determined solely by the geometry and dielectric constant of the internal system but are determined also by the geometry and dielectric constant of the external environment as well as by whether or not the conducting bodies in this external environment are grounded or floating, and by whether or not the lead wires are shielded or unshielded.

For purposes of standardization it would be possible in principle to specify so completely both the precise internal and external environment of the shield that reproducible measurements of these "effective" capacitative constants determined with the shield floating could be obtained, but clearly it is far simpler for purposes of standardization to obtain reproducible results by grounding the shield. This also retains the original simple definitions of the Maxwell coefficients and their related capacitances.

Furthermore, there is a very practical reason for using a tube with all of its shields (including any metal-base shell) grounded and not floating. Thus, with the help of the Maxwell coefficients it is easy to show quite generally that a floating conductor in the neighborhood of the tube will always give "effective" interelectrode capacitances larger than the interelectrode capacitance measured with these bodies or shields grounded.[11]

Similarly, it can also be shown that the "effective" capacitances of the electrodes to earth will be decreased

[11] Measurements made at Heintz and Kaufman on the feedback capacitance of the HK-257-B by E. L. Ung show that the feedback capacitance with base shell floating is nearly twice as large as with the base shell grounded, the geometry being the same in the two cases.

by ungrounding a near-by body. This, however, is an advantage for which an unduly high price is paid in terms of the corresponding increase in the interelectrode capacitance. This is forcefully illustrated by a comparison of the third and fourth series of measurements.

In the fifth series of measurements, the tube was mounted above the center of a grounded metal plate approximately 8 inches square. The same bakelite socket was used for the filament prongs as that shown in Figs. 2 and 3. The three lead wires were arranged in such a way that their geometry with respect to each other and with respect to the tube was approximately the same in the near neighborhood of the tube as the corresponding geometry of the lead wires on the inside of the five-gallon can. From the axis of the tube to the nearest part of the set of three single-pole double-throw switches (see Fig. 2) the shortest distance was 12 inches. The grounded shields of the lead wires extended all the way to these switches. That the presence of these unshielded switches at this distance caused no important change in the measurements was established by running a series of measurements with the set up exactly as in this fifth series except that the switches and high-voltage lead wires were surrounded by some grounded metal screens. No significant change was caused in the interelectrode capacitances and the capacitances to earth were increased by only a few per cent.

Furthermore, the various and sundry floating bodies in the general environment of the laboratory are at such large distances from the tube compared to the dimensions of the electrodes that the "effective" constants determined by this series of measurements differ by a probably negligible amount from measurements which could be made with a shield two or three times as large as the five-gallon shield. This view is supported by the small differences between the results of the first series of measurements and this fifth series. Also these differences are in such a direction as to support this view further; i.e., the interelectrode capacitances in the fifth series are all slightly larger and the capacitances to earth somewhat smaller than in the first series.

A comparison of the fifth and sixth series shows that the "effective" interelectrode capacitances measured with grounded lead-wire shields are all significantly smaller and the "effective" capacitances to earth all significantly larger than the corresponding capacitances measured with lead-wire shields connected to their respective lead wires. A detailed analysis made with the help of the Maxwell coefficients shows that the "effective" interelectrode capacitance, such as that between grid and plate, determined from measurements made with shields connected to lead wires is related in the following way to the corresponding capacitance measured with the lead-wire shields grounded:

$$C_{gp} \text{ (shields connected to lead wires)} =$$
$$C_{gp} \text{ (shields grounded)} + C_{p'g} + C_{g'p} + \Delta C_{p'g'} \tag{12}$$

where

$p' \sim$ plate lead shield

$g' \sim$ grid lead shield

$\Delta C_{p'g'} = C_{p'g'}$ (tube in) $- C_{p'g'}$ (tube out).

Since $C_{p'g'}$ is greater with the tube out, $\Delta C_{p'g'}$ is less than zero. But as the measurements show this is more than offset by the positive sum of $C_{p'g}$ and $C_{g'p}$, so that the "effective" C_{gp} is greater with the shields connected to the leads than with the shields grounded.

Similar analysis of the capacitance of grid to earth shows that:

$$C_{ge} \text{ (shields connected to leads)}$$
$$= C_{ge} \text{ (shields grounded)} - (C_{gg'} + C_{gp'} + C_{gf'}) + \Delta C_{g'e}. \tag{13}$$

And neglecting the relatively small term $\Delta C_{g'e}$ this means that with the shields grounded the shields form part of the "earth" whereas when connected to the lead wires, they do not. And because the mean distance of the lead-wire shields from the tube is small compared to the mean distance of the rest of "earth," this explains why the "effective" capacitances to earth are so very much smaller in the *sixth* series than in the fifth series of measurements.

It might be supposed that the large influence of the unshielded leads can be mainly attributed to the large diameter of the leads in the *sixth* series of measurements. But that this is not so was proved by substituting No. 25 wire, unshielded, and in the same relative position, for the large lead wires. The results werre intermediate between those of the fifth and sixth series, but considerably closer to those of the sixth

The results of the first three series of measurements strikingly verify Maxwell's theoretical prediction[12] about the variation o the coefficients of capacity and the coefficients of induct on with the shielding of the system. They, of course, also show the corresponding dependence of the interelectrode capacitances and the electrode-earth capacitances on shielding. Clearly the variations are significantly large, the change from the five-gallon shield to the close-fitting shield, to take the extreme case, causing a two and a half fold change in C_{pf}. Furthermore, it becomes quite clear that it is meaningless to speak of one set of interelectrode capacitances as being more nearly the "true" set than any of the others. And this means that the wide variations in the reported interelectrode capacitances of a given tube can only be avoided by strict standardization of the type of shielding to be used in the measurements. If this is done concordant results will be obtained by different workers, regardless of which of the many adequate methods of measurement happens to be used.

The column showing the ratio of C_{gf} to C_{pf} is of peculiar interest if the ratios and the trend of these ratios is

[12] See discussion following equation (8) for details.

compared with the measured μ of the triode, which is 28.5. Thus Chaffee,[13] following van der Bijl and using the Maxwell coefficients, shows that

$$\mu = C_g/C_p \qquad (14)$$

or, in the notation used here,

$$\mu = C_{12}/C_{13}. \qquad (15)$$

Hence, using the relations established earlier between the Maxwell coefficients and the interelectrode capacitances, this leads to

$$\mu = C_{gf}/C_{pf}. \qquad (16)$$

Of course, this is true only for the idealized case in which all of the lines of force corresponding to C_{pf} and C_{gf} terminate on the active portion of the filament. In the case of the triode measured here this ideal condition is never realized because some of the lines terminate on the nonactive filament lead wires and the stem shield. Nevertheless this ideal condition is more closely realized the smaller the shield. This can be seen qualitatively by inspection of the geometry of the system and noting that as the grounded shield approaches the tube it will have a much more pronounced effect in reducing the number of stray lines of force from plate or grid (for a given potential) to the inactive parts of the filament, than in reducing the number of lines to the helix, which is much better screened from the influence of the shield than are the inactive parts of the filament. Also, of course, the percentage decrease in C_{pf} will be much greater than in C_{gf} because these strays form a much larger portion of C_{pf} than of C_{gf}. This accounts for the rapid rise of the ratio toward the μ of the tube.

[13] E. Leon Chaffee, "Theory of Thermionic Vacuum Tubes," McGraw-Hill Book Co., New York, N. Y., 1933, p. 146.

Acknowledgment

The writer would like to acknowledge the valuable assistance which he has received throughout the course of this work from Mr. W. G. Wagener. It is also a pleasure to thank Heintz and Kaufman, Ltd., for its splendid co-operation on a project which the author hopes will be of some value for the problem of standardization in the interest of the war effort.

Equivalent Circuits for Discontinuities in Transmission Lines*

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***Summary*—Exact equivalent circuits for several types of discontinuities in parallel-plane transmission lines are obtained by Hahn's method of matching electromagnetic-wave solutions. Values of lumped elements to be used in these are given in curve form, with rules for using results in the corresponding coaxial-line problems. Experimental checks are reported, which verify results of the calculations and stress the importance of the discontinuity capacitances appearing in the equivalent circuits.**

I. Introduction

In 1941, Hahn[1] presented a description of a new approach toward matching of electromagnetic-wave solutions across discontinuities, with application to the problem of cavity resonators as an example. In much unpublished work, Hahn has shown that the method when applied to transmission systems with discontinuities leads to simple equivalent circuits with lumped-circuit constants representing the effect of these discontinuities. The equivalent circuits are simple, exact, and very useful for engineering thinking or calculation. It is our purpose in this paper to present this point of view with the following goals in mind:

1. To show the exact equivalent circuits for several representative transmission-line discontinuities.
2. To present curves showing numerical values to be used in these equivalent circuits.
3. To discuss the physical pictures inherent in the method and results, and the application to practical problems.
4. To show, in enough detail to be readily followed, the mathematical steps for one simple example.

It should be pointed out at the outset that the point of view in the equivalent circuits is identical with that used by Schelkunoff[2] in radiation problems, in which he has shown that the "complementary" end-effect waves of an antenna may be included in a principal wave equivalent circuit as a lumped admittance at the end of the antenna. Similarly, we find equivalent circuits to be used with the principal wave on a transmission system, the effect of local waves at a discontinuity being included in lumped admittances.

In this paper, numerical values are given only for parallel-plane transmission lines. Although it is planned to present corresponding results for coaxial lines in a succeeding paper, instructions are given in Part III for application of present parallel-plane results to coaxial lines. The experimental checks described were made on coaxial lines. These not only check the calculations, but emphasize the importance of knowing values for the discontinuity capacitances at high frequencies.

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† Electronics Laboratory, General Electric Company, Schenectady, New York.

[1] W. C. Hahn, "A new method for the calculation of cavity resonators," *Jour. Appl. Phys.*, vol. 12, pp. 62-68; January, 1941.

[2] S. A. Schelkunoff, "Theory of antennas of arbitrary size and shape," Proc. I. R. E., vol. 29, pp. 493-521; September, 1941.

II. The Results

In Figs. 1 to 13, inclusive, several typical discontinuities that may arise in parallel-plane transmission lines are sketched, together with the exact equivalent circuits derived in Part VI and the Appendixes.

The equivalent circuits may be used with any termination or excitation, subject only to certain cautions listed in Part III. The terminating conditions are simply placed at the corresponding z co-ordinate of the equivalent circuit. Electromagnetic behavior between the exciting source and the termination is then calculated by means of ordinary transmission-line equations for the transmission-line sections with known characteristic impedances Z_0. The shunt admittances included at the junction points between lines account for the local fields of the discontinuities and may consequently be called "discontinuity" admittances.

Engineering values for the lumped discontinuity capacitances appearing in the equivalent circuits may be found from a single curve for all cases, Figs. 1 to 9. This curve, Fig. 14, is the exact discontinuity capacitance for the single-step discontinuity, Fig. 1. It is applied to the other cases, Figs. 2 to 9, by combining in a straightforward manner the lumped capacitances for the corresponding simple steps.

It might be expected that field distortions from the combinations of steps would be great enough to make the combination of simple-step results a poor approximation, but accurate curves of the discontinuity capacitances for several of these complex cases are plotted in the appendixes (Fig. 23 to 25), justifying this approximation. In the double-step discontinuities of Figs. 2 to 9, discontinuity fields from the separate steps are in each case fairly well shielded from one another, as contrasted to those of Figs. 10 and 11, where fields from one step are definitely influenced by the other. Fig. 15 must then be used in combination with Fig. 14 to give the discontinuity capacitances for the two cases of Figs. 10 and 11, as these capacitances are definite functions of d/a.

If transverse dimensions are comparable to a half wavelength, it is necessary to correct the values of capacitance taken from Fig. 14. Correction curves showing the proper value of susceptance to place at the discontinuity as a ratio to that calculated from the capacity of Fig. 14, are plotted in Fig. 16 versus (largest transverse dimension/wavelength). These are calculated for the single-step discontinuity, but may be applied approximately to the cases of Figs. 2 to 9.

The discontinuity, Fig. 12, is represented by the three transmission lines in series with a shunting capacitance placed across each line at the junction. Values for these capacitances are obtained approximately from Fig. 17. Fig. 13 is a special case of Fig. 12, with one line short-circuited so that it represents an inductance of given value. The lumped capacitances may be appreciably affected in this case by the proximity of the short-circuiting end if l/a is less than 0.5. Discontinuity capacitances should then be multiplied by a proximity factor given in Fig. 18.

III. Application to Practical Problems

It is not our purpose to list all applications to practical problems, since many such applications will be evident to engineers once the validity of the equivalent circuits is recognized. However, the systems used for experimental checks will serve as example. A few general instructions and cautions are also in order.

A. Coaxial-Line Discontinuities

The parallel-plane transmission lines analyzed here may be considered as coaxial lines with $r_0/r_i \rightarrow 1$. (r_0 = outer radius; r_i = inner radius). The total lumped capacitance at a discontinuity is then

$$C_d = 2\pi r C_d' \tag{1}$$

where C_d' is discontinuity capacitance per unit width obtained from the corresponding parallel-plane case of Figs. 1 to 13 and r is a representative radius (which is easy to select since r_i is close to r_0).

The present results can also be used for approximate, though very good, results for coaxial lines of *any ratio* r_0/r_i with the following rules:

(a) For a discontinuity in the outer line, as in Fig. 19 (a), use r_1 in (1).
(b) For a discontinuity in the inner line, as in Fig. 19 (b), use r_3 in (1).
(c) For a discontinuity in both inner and outer conductors, as in Fig. 19 (c), add capacitances from the two types of discontinuities.

$$C_d = 2\pi r_2 C_d'(a/b) + 2\pi r_3 C_d'(a/c). \tag{2}$$

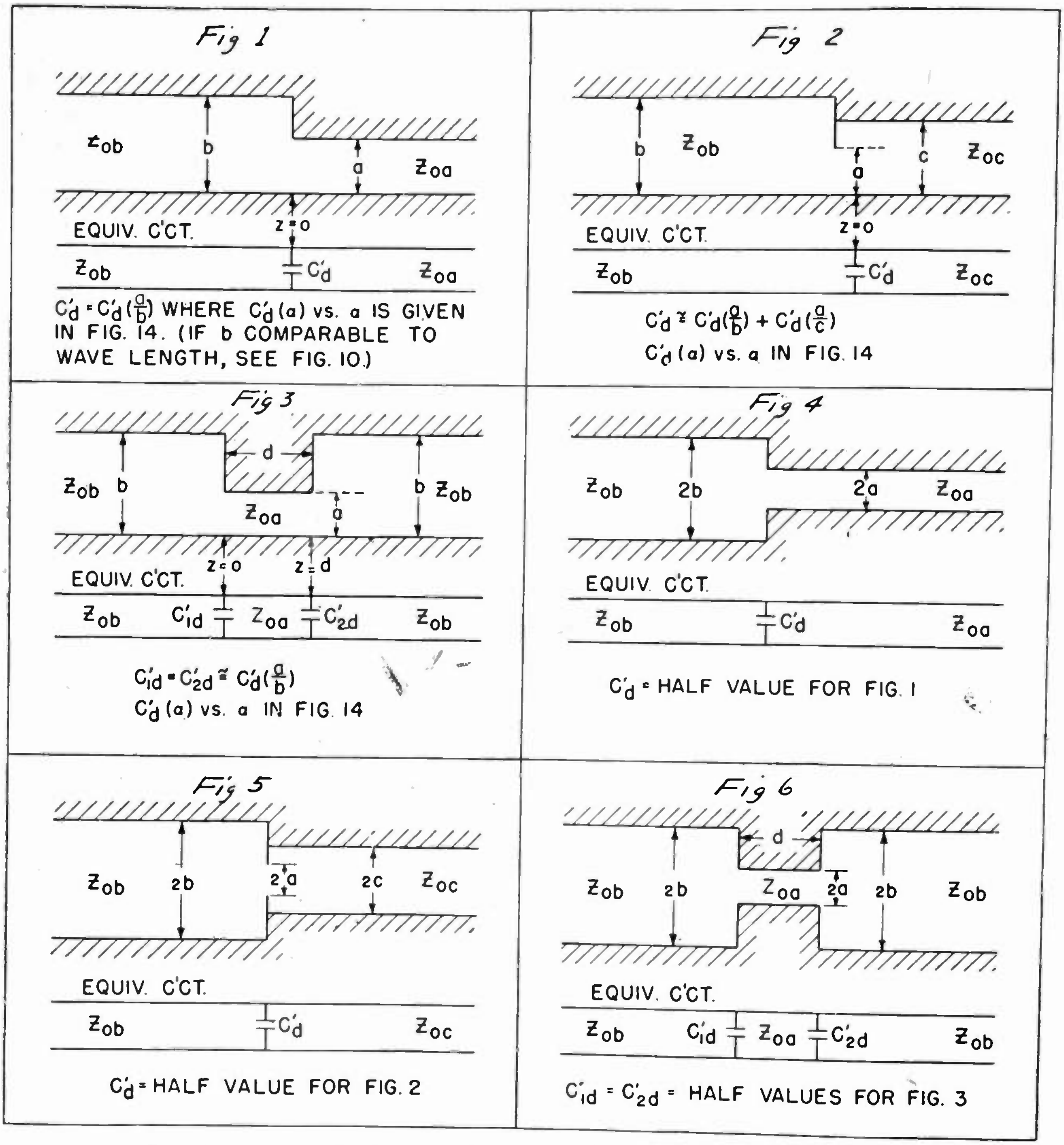
Justification for these rules will be given in the paper on coaxial lines, where it is shown that such approximate results do not differ from the exact by more than about 10 per cent for values of r_0/r_i up to 5.

B. Cavity Resonators

Application of the equivalent circuits to many resonant-cavity types with discontinuities is at once evident. It is only necessary to place short circuits corresponding to the closed ends at the corresponding points on the equivalent circuit, and calculate resonance by simple circuit equations. However, we must recognize that this requires that the local field of the discontinuity be negligible at the point where the short-circuiting ends are placed, so that these are truly short-circuiting the transmission-line wave, and affecting only slightly the local waves. This is not true in many resonators, and if accurate results are required in such cases, one must include the complete dimensions of the cavity in the matching operation, as in Hahn's original paper.

C. High-Frequency Filters

The configuration of Fig. 12 appears in certain high-frequency filter combinations and may be used for



Discontinuities in parallel-plane transmission (Figs. 1 to 13) lines with equivalent circuits.

analysis of these with the cautions mentioned elsewhere in this part of the paper.

D. Cautions

We have mentioned previously these cautions:

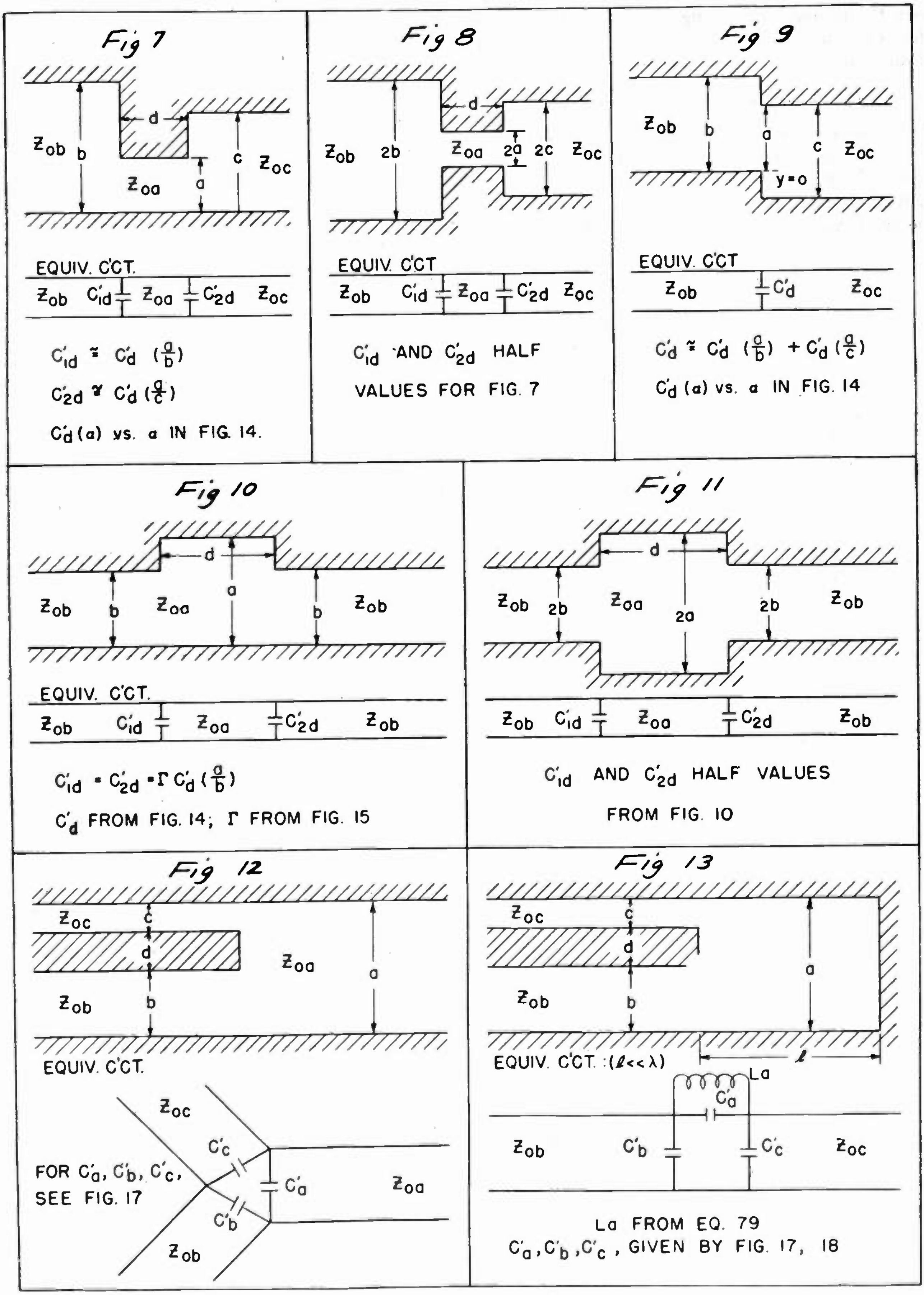
(1) Proper radii must be used in applying results to coaxial lines of large ratio r_0/r_i.

(2) Results must be corrected if terminations are close to the discontinuity (an axial distance comparable to the transverse dimension between conductors being used as a criterion of closeness at low frequencies).

(3) Present results are limited to cases with largest transverse dimension not greater than a half wavelength.

(4) Caution must be used if multiple discontinuities closely affect one another, as in the cases of Figs. 10 and 11.

We should also caution against interpreting results for taper sections from these cases which apply to abrupt discontinuities. It may at first appear that the major difference is in the magnitude of the lumped capacitance but this is only part of the story. For example, if the problem is one of the mismatch between the lines A and B of Fig. 1, much of this mismatch comes from the difference between Z_{0A} and Z_{0B}, and in most cases a smaller



Discontinuities in parallel-plane transmission (Figs. 1 to 13) lines with equivalent circuits.

amount from the lumped capacitance C_d'. Tapering decreases the mismatch in going from Z_{0A} to Z_{0B}, as is evident from nonuniform line theory or the appropriate field solutions.[3]

IV. Experimental Verification

To check the validity of the theory by experiment, two sets of tests were performed. In the first, the resonant wavelength of the coaxial line system in Fig. 20 (a) (which maintains constant-characteristic impedance across the step, and would resonate at $\lambda = 4l$ except for

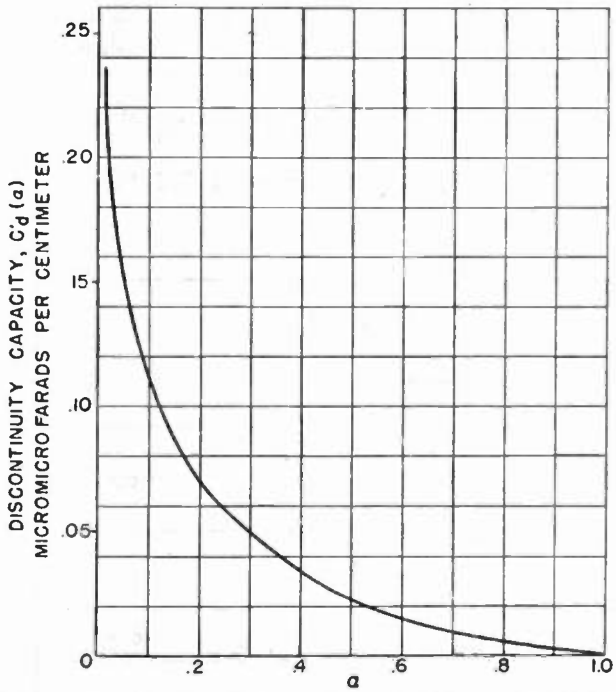


Fig. 14—Exact discontinuity capacitance for Fig. 1. Curve is used also for approximate results in Figs. 2 to 9. Results given for air dielectric; for other dielectrics multiply by dielectric constant.

the discontinuity capacitance) was measured and compared with the value calculated from the equivalent circuit, using previous curves and the rules of Part III, *A*, to find the value of discontinuity capacitance. In the second, standing waves were measured in (b) of Fig. 20 (b) at a frequency corresponding to $l = \lambda/2$ and with the line C terminated in its characteristic impedance. Under these conditions, the generator should see Z_{0C} paralleled with $2C_d$, the discontinuity capacitance calculated from the rules of Part III, *A*, and would be matched perfectly if it were not for C_d. Measured standing-wave ratio was compared with that calculated from ordinary transmission-line theory assuming the line terminated in an admittance $Y_{0C} + j2\omega C_d$. The two sets of tests will be outlined in more detail to serve as actual numerical examples and to indicate the order of magnitude of the discontinuity capacitance encountered in ordinary circuits and of the resulting effects on the electrical properties of a given system.

[3] See, for example, J. C. Slater, "Microwave Transmission," McGraw-Hill Book Co., New York 18, N. Y., 1942, sections 6 and 25.

Case A: Calculation of resonant λ for line of Fig. 20 (a). Physical dimensions of the resonator in inches are

$$r_1 = 0.217 \qquad c = 0.283$$
$$r_2 = 0.400 \qquad a = 0.100$$
$$r_3 = 0.500 \qquad b = 0.520$$
$$r_4 = 0.920$$
$$a/c = 0.353 \qquad a/b = 0.192$$
$$2l = 1.969 \text{ inches} = 10 \text{ centimeters.}$$

The resonance requirements are $Y_A + Y_B = -j\omega C_d$ where C_d is the appropriate discontinuity capacitance. Then

$$\frac{2}{Z_0 \tan \beta l} = \omega C_d = \frac{2\pi(3 \times 10^{10})}{\lambda} C_d. \quad (3)$$

Equation (2) used with Fig. 14 gives C_d for this type of discontinuity

$$2\pi r_2 C_d'(a/b) = 0.469 \text{ micromicrofarad}$$
$$2\pi r_3 C_d'(a/c) = 0.311 \text{ micromicrofarad}$$
$$C_d = 0.780 \text{ micromicrofarad}$$

Using this value of C_d in (3) and solving graphically for λ, it is found that $\lambda = 12.27$ centimeters (calculated). By a resonance measurement, accurate to about ± 0.005 centimeter, $\lambda = 12.34$ centimeters (measured).

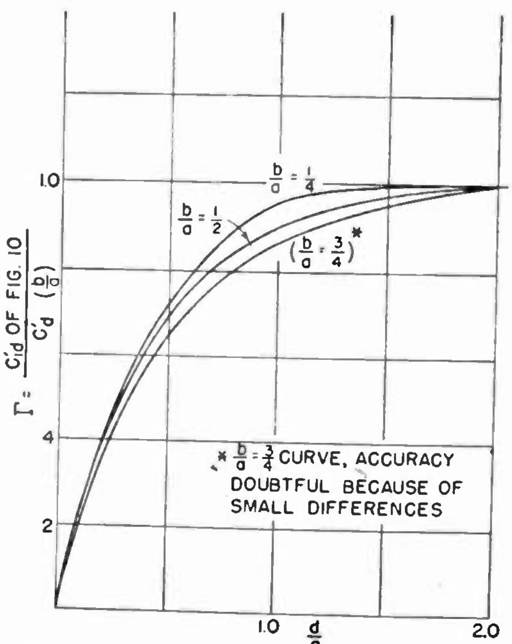


Fig. 15—Proximity effect for Figs. 10 and 11.

Case B: Calculation of reflection coefficient for line of Fig. 20(b). Three such discontinuities were checked having $(a/b) = 0.25$, 0.5, 0.75. The one with $a/b = 0.5$ will be chosen as a typical example.

Dimensions

$r_1 = 0.125$ inch $\quad a = 0.156$ inch $\quad l = \lambda/2 = 1.930$ inch
$r_2 = 0.281$ inch $\quad a/b = 0.50 \quad Z_{0B} = 75$ ohm
$r_3 = 0.437$ inch $\quad b = 0.312$ inch

From transmission-line theory the reflection coefficient is given in terms of the characteristic admittance and terminating admittance as

$$R = \left| \frac{Y_0 - Y_R}{Y_0 + Y_R} \right| = \left| \frac{-jZ_0\omega C_d}{1 + jZ_0\omega C_d} \right|. \tag{4}$$

Now from the curve of Fig. 14, $C_d'(1/2) = 0.024$ micromicrofarad per centimeter and $C_d = 2\pi r_3 C_d' = 0.167$ micromicrofarad. Solving (4) for R using this value of C_d, we find $R = 23.45$ per cent (calculated). The value found by measurement of standing waves was $R = 23.5$ per cent (measured).

The results of calculations and measurements (with probable accuracy as noted) on the three discontinuities may be summarized here

(a/b)	C_d micro-microfarads	$R_{calculated}$ per cent	$R_{measured}$ per cent
0.75	0.105	7.55	7.4 ± 0.5
0.5	0.334	23.45	23.5 ± 0.5
0.25	0.815	50.8	$58 \quad \pm 2.0$

The results in these two cases verify the use of the parallel-plane equations, and corresponding rules for selection of radii in coaxial-line calculations. Remembering that the rules given in III, *A*, give only an approximation to the exact results for coaxial systems, the accuracy of the calculations is certainly within the limits expected. The importance of the discontinuity capacitance is emphasized by noting that in the case of the resonant line, $4l = 10$ centimeters, the resonant wavelength is changed from that of a smooth line by 2.3 centimeters (23 per cent) and reflection coefficients of as high as 50 per cent are present in lines that by ordinary transmission-line treatment, neglecting C_d, would be assumed perfectly matched.

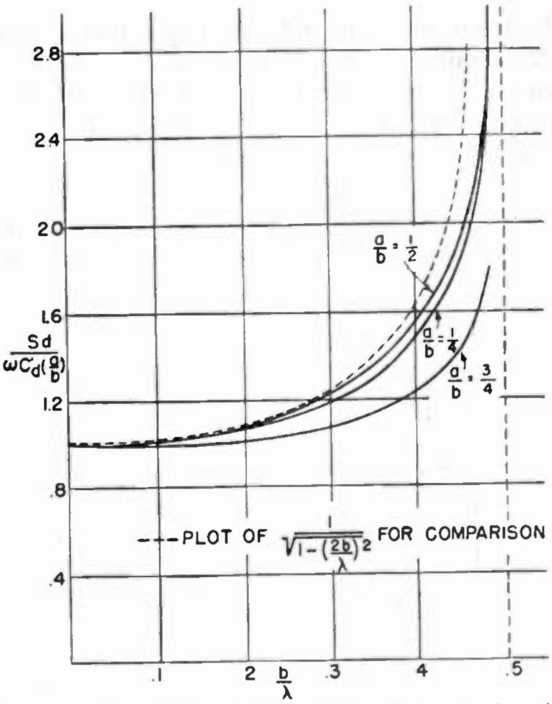


Fig. 16—Effect of dimensions comparable to wavelength; S_d = discontinuity susceptance.

V. The Physical Pictures and Interpretations

A. The Concept of Higher-Order Waves

In the modern use of wave guides, it has come forcibly

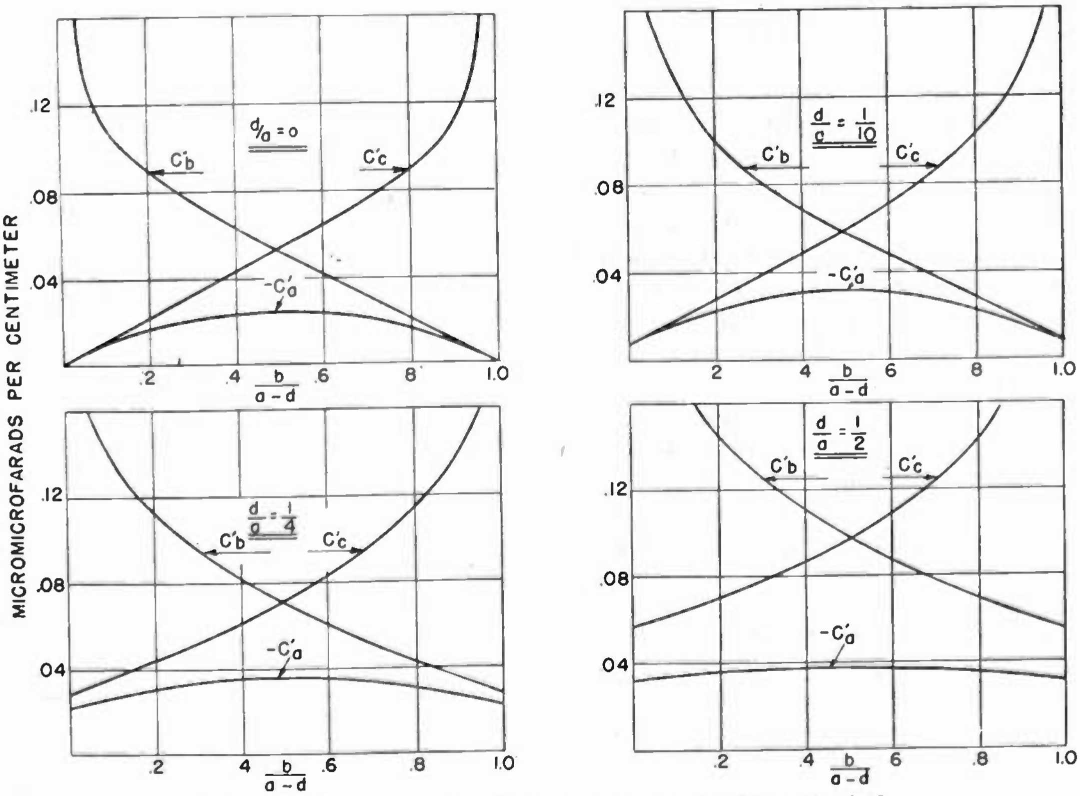


Fig. 17—Approximate discontinuity capacitance for Figs. 12 and 13.

to the attention of engineers that any guiding system for electromagnetic-energy transmission is capable of supporting not one but many wave types (actually an infinite number). An equivalent circuit showing a given inductance, capacitance, series resistance, and shunt conductance per unit length, and resulting in a given characteristic impedance and propagation constant, can be drawn only for a given wave type of the transmission

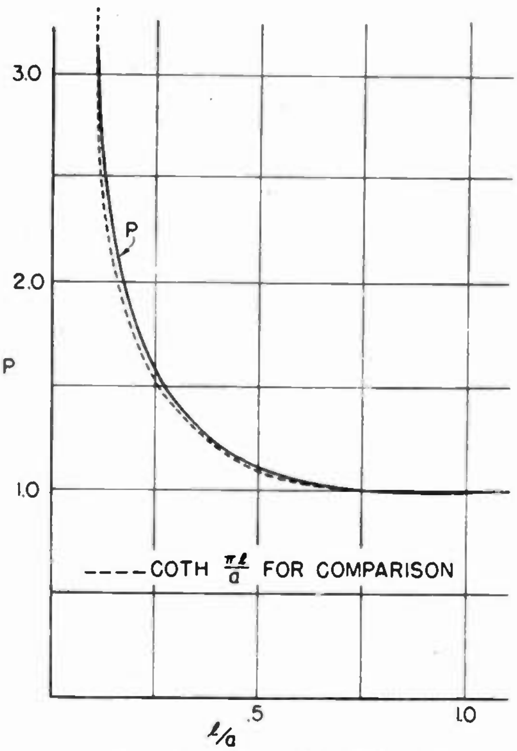


Fig. 18—Proximity factor for Fig. 13. (Multiply values from Fig. 17 by P to obtain C_a', C_b', C_c' for Fig. 13.)

system and not for the transmission system itself. Fortunately, the simple equivalent circuit arrived at by circuit reasoning which considers differential elements in classical transmission-line theory has been shown by rigorous field theory to represent the single-wave type which is most important in such transmission lines, the transverse electromagnetic wave, principal wave, or transmission-line wave, as it is variously called. This representation is exact for a uniform line of zero series resistance (although conductance need not be zero) and is an exceedingly good approximation for lines with finite series resistance, provided that this is not so great as to make the transmission line inefficient for energy transfer.[4,5]

Now it is known that practically any physical means of exciting a wave on a transmission line excites not one but many of the possible wave types. Mathematically, we state this by noting that we must add a series of the wave solutions, not one alone, to match the boundary conditions imposed by the exciting source. There is, however, no need to worry about the equivalent circuit not accounting for these in most practical cases, since a study of these "higher-order" or "complementary" wave types shows that they attenuate exponentially at an extremely rapid rate unless the transverse dimensions of the transmission line are appreciable compared to wavelength. As a rough rule we may think of these waves disappearing in a longitudinal distance comparable to the transverse dimensions of the line, since they merely add to represent the fringing field predicted by static-field considerations if transverse dimensions are truly small compared to wavelength. The name "local waves" is thus excellent to describe the waves when below cutoff. The attenuation of the local waves is not a dissipative attenuation however, but is a reactive attenuation as in conventional filters in the cutoff region, and mainly adds a certain reactive component to the driving source. Thus it is usually sufficient to note that there is a slight end effect, but over the main body of the uniform line there is only the single transmission-line wave which may be represented by the transmission-line equations.

Similarly, any discontinuity in the transmission line will excite certain of the higher-order waves. Mathematically, again, we say that the matching conditions at the boundary require a series of waves, not one wave alone. This requirement is particularly clear in the simple problem of the step which we use as the mathematical example later. (Fig. 21(a).) The transmission-line wave between parallel planes has only field components E_y and H_x, and these do not vary in the y direction. Consider such a wave progressing down the line in the z direction to the right, confronting the step discontinuity at $z=0$. The conducting portion from $y=a$ to $y=b$ requires that E_y be zero over this region (assuming perfect conductivity in the first approximation). Thus if there were only this single-wave type, E_y would have to be zero everywhere at the plane $z=0$, since there are no field variations with y for this wave type. The simple step would then act like a complete short circuit reflecting all energy (since no energy could pass if E_y and hence the Poynting vector were zero everywhere across the plane). This is ridiculous, physically, and the trouble is cleared up at once when we allow other higher-order waves to be added so that E_y in the transmission-line wave need not be zero; total E_y, summing up the contribution from all waves, will be zero from $y=a$ to $y=b$ but not from $y=0$ to $y=a$. These higher-order waves again attenuate in a distance comparable to the transverse dimensions if these are small compared to wavelength and contribute to a fringing field such as could be drawn from a static field map, Fig. 21(b). We would, of course, never make the mistake described above, but as a first-order approximation would draw the equivalent circuit of Fig. 21(c) connecting the line of characteristic impedance Z_{0B} directly to the line of characteristic impedance Z_{0A}, using ordinary

[4] John R. Carson, "The guided and radiated energy in wire transmission," *Jour. A. I. E. E.*, vol. 43, pp. 906–913; October, 1924.
[5] S. A. Schelkunoff, "The electromagnetic theory of coaxial transmission lines and cylindrical shields," *Bell Sys. Tech. Jour.*, vol. 8, pp. 532–579; October, 1934.

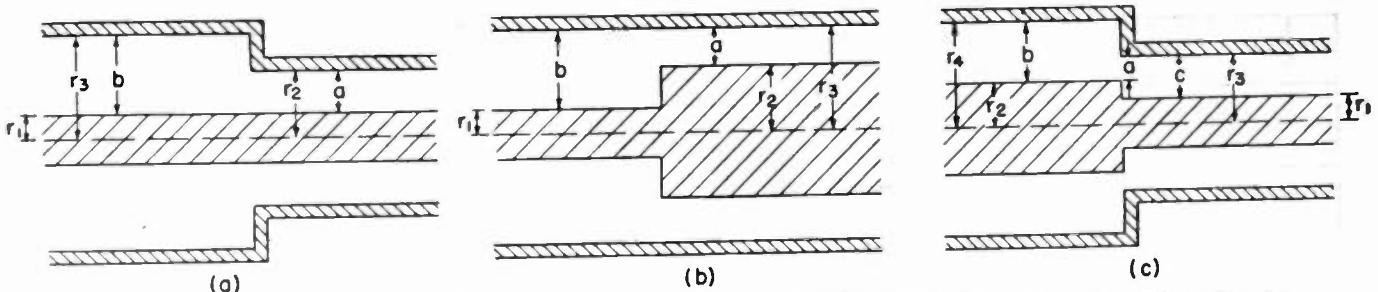


Fig. 19—Discontinuities in coaxial lines for which approximate discontinuity capacitances are obtainable from Fig. 14.

transmission-line equations, with the assumption of continuous voltage and continuous current in the transmission-line wave across the discontinuity. We recognize that the picture is incomplete since it nowhere accounts for the presence of the higher-order waves in the equivalent circuit which applies only to the transmission-line wave; we only hope, if we draw such a circuit, that effects from these local waves will be small.

B. Local E Waves Represent Electric Energy, Local H Waves, Magnetic

Of the many possible ways of classifying the higher-order-wave types, one of the most convenient divides waves into two classes: (1) E waves or transverse magnetic having electric field in the direction of propagation, magnetic field confined to the transverse plane, (2) H or transverse electric waves with magnetic field in the direction of propagation, electric field confined to the transverse plane. This is a convenient separation for our purposes since often only one of these classes will be excited by a given discontinuity. For example, between parallel planes uniform in the x direction, the E waves contain components E_y, H_x, and E_z only, the H waves contain H_y, E_x, and H_z. A study of the picture Fig. 21(b), shows that the former will certainly be required in the local field, but no reason for any of the components of the latter can be detected. This type of discontinuity thus excites E waves but not H waves.

The field equations will show that the energy stored in E waves below cutoff is a net energy in electric fields;[6] in H waves it is a net energy in magnetic fields. Thus we should expect local E waves to enter in reflection effects as a capacitive reactance, and local H waves should enter somehow as an inductive reactance.

C. Voltage in Transmission-Line Wave is Continuous at Discontinuities, Current is Not

The first approximate equivalent circuit for the step discontinuity (Fig. 21(c)) was written from the belief that voltage and current must be continuous across the discontinuity. This is true if we are talking of total voltage and current, but in the equivalent circuit we are not; we are only talking of voltage and current in the transmission-line wave, to which must be added those in the local waves near the discontinuity. In Schelkunoff's notation[2] $V(z)=V_0(z)+\overline{V}(z)$ and $I(z)=I_0(z)+\overline{I}(z)$.

$V(z)$ and $I(z)$ represent total voltage and current as functions of z, V_0 and I_0 represent those in the transmission line wave, $\overline{V}$ and $\overline{I}$ those in the local waves.

If voltage is defined as the line integral of electric field between planes in the equiphase XY plane, we can check from the distribution of all higher-order waves that these have just as much negative as positive electric field components, so that their net contribution to the voltage integral is zero. $\overline{V}(z)=0$ or $V(z)=V_0(z)$. That is, total voltage when defined in this fashion, is exactly that in the transmission-line wave; since total voltage must be continuous across the discontinuity, so must that in the transmission-line wave. Say for a discontinuity between A and B regions at $z=z_1$,

$$V_{0A}(z_1) = V_{0B}(z_1). \tag{5}$$

Since contribution to current from the higher-order waves $\overline{I}(z)$ is not zero, the complete expression for total current $I(z)$ must be used. Continuity of this total current at $z=z_1$, $I_{0A}(z_1)+\overline{I}_A(z_1)=I_{0B}(z_1)+\overline{I}_B(z_1)$ allows a discontinuity in transmission-line wave currents

$$I_{0A}(z_1) - I_{0B}(z_1) = \overline{I}_B(z_1) - \overline{I}_A(z_1). \tag{6}$$

D. Effect of Local Waves at a Discontinuity May Be Represented by Lumped Admittance Exactly at Discontinuity

It follows at once that a continuity of voltage but discontinuity of current at $z=z_1$ may be taken care of in the transmission-line equivalent circuit by lumping an admittance[7] across the lines at this discontinuity, of magnitude

$$Y_d = \frac{I_{0B}(z_1) - I_{0A}(z_1)}{V_0(z_1)} = \frac{-\overline{I}_B(z_1) + \overline{I}_A(z_1)}{V_0(z_1)}. \tag{7}$$

It is not obvious at once that this admittance will be as generally useful as we would wish, since it might conceivably turn out to be a function of the impedance termination of the line and its method of excitation. Fortunately, it turns out to be independent of these, at least if the exciting means and the terminations are removed far enough from the discontinuity so that the local fields of these are not mutually linked to the local fields of the discontinuity. We will assume this in most of our analyses.

[6] In the sense of net electric energy for a series L-C circuit below resonance, where net reactance is capacitive.

[2] *Loc. cit.*, p. 499.

[7] It is also proper, and sometimes helpful, to think of this discontinuity admittance as an infinite number of admittances in parallel, one for each of the local waves.

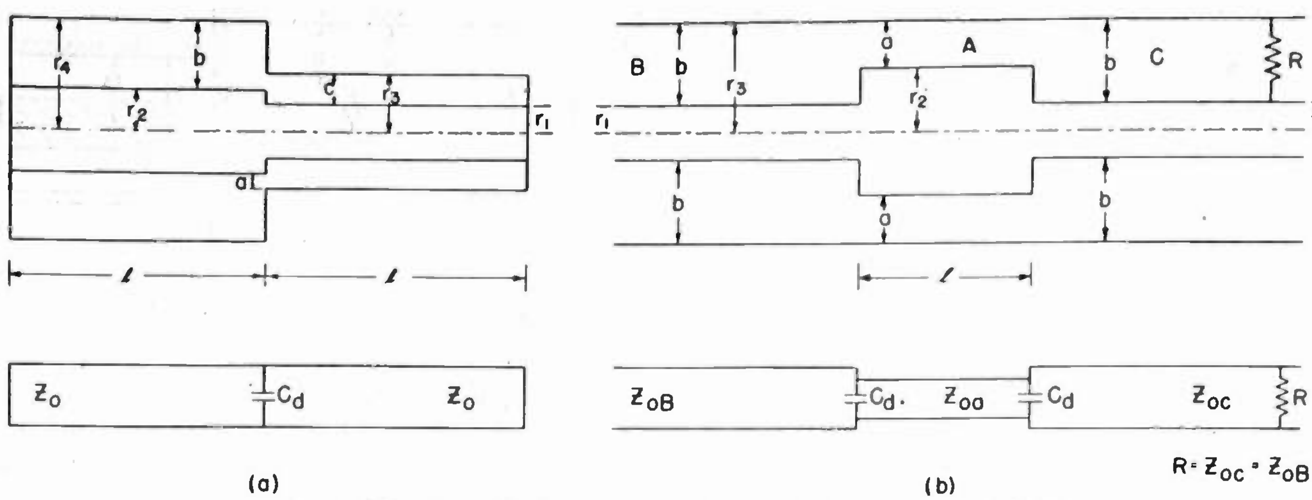


Fig. 20—Coaxial-line discontinuities used for experimental checks, with equivalent circuits.

There may at first be a feeling that the placing of this lumped susceptance is a little arbitrary. We have agreed from our reasoning of Part V, C, that the local waves should enter as a reactive effect in some manner, but this effect is clearly spread over a certain finite region, not concentrated exactly at $z=z_1$. This objection is answered by the separation between total quantities and those of the transmission-line wave to which the equivalent circuit applies. In the latter, we ask only about discontinuity in current in this single wave, and a review of the above steps shows that, for the equivalent circuit, it is correct to take care of this discontinuity exactly at $z=z_1$. Physically, the local waves do add to this current over a finite region to insure continuity of total current, as sketched qualitatively in Fig. 21(e).

E. Lateral Dimensions Approaching Half Wavelength

The curves, Fig. 16, show that as the dimension b of the stepped transmission line approaches a half wave, the capacitive susceptance shunted across the discontinuity approaches infinity. This simply means that the lowest-order local wave (one which has half-sine variations of fields in the y direction) is approaching cutoff. For $b=\lambda/2$, it is exactly at cutoff, which is another way of saying that it is exactly at resonance for propagation of the energy of this wave in a completely transverse direction between planes. We would then expect some resonance phenomenon to appear, and it does, corresponding to a series resonance in the shunting circuit at the discontinuity. The cutoff factor for the lowest order local wave is $1/\sqrt{1-(2b/\lambda)^2}$. This is plotted in Fig. 16, and is a fair approximation to $S_d'/\omega C_d'$ for all except values of a/b approaching unity. Such an approximation assumes that all the frequency change is occurring in the capacity associated with the first-order local wave.

F. Reduction to the Static-Field Problem at Low Frequencies

If frequency is low enough so that wavelength is huge compared to transverse dimensions of the line, it is evident that the local disturbance of fields extends only over a region comparable to the transverse spacing, and therefore over a region negligibly small compared to wavelength. The equivalent circuit derived in this report would then be arrived at by physical reasoning, the discontinuity capacitance being the difference between total capacitance of a section which includes all appreciable effects of the discontinuity, and capacity of the corresponding section calculated on the basis of field lines passing straight across (that is, capacitance from the simple parallel-plane formula). One would not worry about the exact placing of this extra capacitance in such reasoning, since $2\pi z/\lambda$ is the variable of transmission-line equations, and for this reasoning we are postulating that z/λ varies by a negligible amount over the region of the discontinuity. (Such general reasoning should not mask the result from the mathematical analysis that the placing of the capacitance exactly at the discontinuity is rigorous.)

The excess discontinuity capacitance in the static case may be found by matching of series solutions to Laplace's equation in a manner similar to that described in Part VI for the series-wave solution. We have carried through such solutions, and results of course agree with those calculated from the wave solutions with $(2b/\lambda)^2$ negligible compared to unity.

Results for the static case might also be obtained from a carefully made flux plot, and the discontinuity of Fig. 1 may also be analyzed by a Schwarz-Christoffel transformation. Utilizing results of such analyses,[8] we find the discontinuity capacitance, $C_d'(\alpha)$ to be expressible in closed form

$$C_d'(\alpha)=\frac{\epsilon}{\pi}\left(\frac{\alpha^2+1}{\alpha}\cosh^{-1}\frac{1+\alpha^2}{1-\alpha^2}-2\ln\frac{4\alpha}{1-\alpha^2}\right) \qquad (8)$$

$\alpha=a/b$ for the step of Fig. 1.

Although the curve of $C_d'(\alpha)$ of Fig. 14 was calculated from the series result, (16), the above equation checks these values to the degree of accuracy of the calculations.

[8] The problem is worked, for example, in Miles Walker "Conjugate Functions for Engineers," Oxford University Press, London, England, 1933, pp. 53 to 65.

Although the closed form may seem to have advantages, we believe that the series-wave picture is superior, particularly for analysis of the more complicated discontinuities, and for information on the effects of frequency.

VI. Example of Step in Parallel-Plane Transmission Line

To demonstrate the method of obtaining the equivalent circuits presented in Part II, as well as to illustrate the general technique in the use of Hahn's functions for wave-matching problems, consider the simple case of a

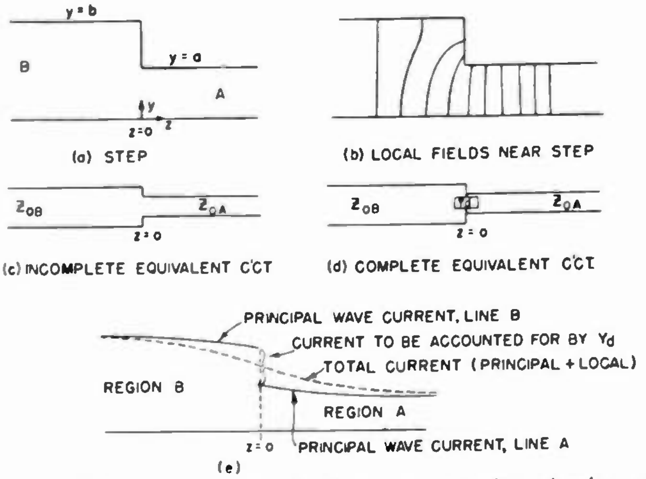


Fig. 21—Illustrations of field distributions, equivalent circuits, and current components for step discontinuity in a parallel-plane transmission line.

single step in a parallel-plane transmission line, located at $z=0$. The line may be terminated at any point far removed from the step, on either the A or B side, and will be assumed to be excited in some manner with a transmission-line wave.

The general attack of the problem will be as follows. (1) Write expressions for the total field components in region B as an infinite sum of wave types with arbitrary constant coefficients. (2) Write similar expressions for region A. (3) To fulfill the required conditions of continuity of tangential electric- and magnetic-field components across the boundary, match E_y components in B region to A region over the range $0<y<a$, $E_y=0$ in B region for $a<y<b$. (4) Match H_{zA} to H_{zB} for $0<y<a$. (5) Eliminate between the equations resulting from (3) and (4), to evaluate the arbitrary constants. (6) Finally, obtain the equivalent admittance to represent the discontinuity in the transmission-line wave current at the discontinuity.

A. Wave Solutions and Matching

Having assumed the line excited from either end with a transverse electromagnetic wave, we shall represent the total amplitude of this wave in region A as A_0, and in region B as B_0 at $z=0$. At the discontinuity, there must arise components of E in the z direction to account for the fringing field, while no z components of H are required. Thus the E (or transverse magnetic) waves will originate at the discontinuity. These waves will be distributed in such a manner as to satisfy Maxwell's equations in the space bounded by the perfectly conducting infinite planes. Assuming uniformity in the x direction, the distribution equations in rational m.k.s. units for an nth-order wave are[3]

$$E_{z_n} = \sin(n\pi y/b)\left[B_n e^{-\gamma_n z} + B_n' e^{\gamma_n z}\right] \tag{9}$$

$$E_{y_n} = (\gamma_n b/n\pi)\cos(n\pi y/b)\left[-B_n e^{-\gamma_n z} + B_n' e^{\gamma_n z}\right] \tag{10}$$

$$H_{z_n} = j(\omega\epsilon b/n\pi)\cos(n\pi y/b)\left[B_n e^{-\gamma_n z} + B_n' e^{\gamma_n z}\right] \tag{11}$$

where $\gamma_n = \sqrt{(n^2\pi^2/b^2) - \omega^2\mu\epsilon} = (n\pi/b)\sqrt{1-(2b/n\lambda)^2}$.

The equations for the A space are similar to these. Now define a wave admittance Y_n for a positive exponential term.

$$H_{z_n}/E_{y_n} = Y_n = j(\omega\epsilon b/n\pi K_n) = j(\omega\epsilon/\gamma_n) \tag{12}$$

where $\quad K_n = \sqrt{1-(f/f_{cn})^2} = \sqrt{1-(2b/n\lambda)^2}. \quad (13)$

For a negative exponential term, the sign of γ_n is opposite. Expressions for total E and H in regions A and B may now be written as the sums of the principal and higher-order wave components. Since we shall equate E_y components at $z=0$ and similarly equate H_z components, we shall confine our attention to these and write

$$E_{yA}\big|_{z=0} = A_0 + \sum_{m=1}^{\infty} A_m \cos\frac{m\pi y}{a} \tag{14}$$

$$H_{zA}\big|_{z=0} = Y_{A_0}A_0 + \sum_{m=1}^{\infty} Y_{A_m}A_m \cos\frac{m\pi y}{a} \tag{15}$$

and $\quad Y_{A_m} = -j(\omega\epsilon a/m\pi K_m)$

for the A region. Similarly, the expressions for E_y and H_z for the B region are

$$E_{yB}\big|_{z=0} = B_0 + \sum_{n=1}^{\infty} B_n \cos\frac{n\pi y}{b} \tag{16}$$

$$H_{zB}\big|_{z=0} = Y_{B_0}B_0 + \sum_{n=1}^{\infty} Y_{B_n}B_n \cos\frac{n\pi y}{b} \tag{17a}$$

and $\quad Y_{B_n} = j(\omega\epsilon/\gamma_n) = j(\omega\epsilon b/n\pi K_n). \quad (17b)$

We have here assumed that terminations are far enough removed from the discontinuity so that only positive exponentials for the local waves are required in region

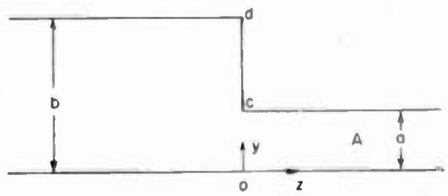


Fig. 22—Step discontinuity in parallel-plane transmission line, showing co-ordinate system.

$B(z<0)$ and negative exponentials in region $A(z>0)$. Hence the opposite signs for the Y's in the two regions. The Y_0's are the transmission-line wave admittances whose values depend upon the line terminations.

The coefficients are now to be related through the boundary conditions at $z=0$.

[3] See for example Slater, *loc. cit.*, where these waves are given in slightly different notation by his equation 14.13.

$$E_{yB}\big|_{z=0} = E_{yA}\big|_{z=0} \qquad 0 < y < a \tag{18a}$$
$$E_{yB}\big|_{z=0} = 0 \qquad a < y < b$$
$$H_{xA}\big|_{z=0} = H_{xB}\big|_{z=0} \qquad 0 < y < a. \tag{18b}$$

Consider for a moment the expression (16) for $E_{yB}\big|_{z=0}$. It is evident that this equation may be considered as a Fourier expansion for E_{yB} over the region $0<y<b$. Further, we recognize from the boundary conditions (18a) that E_{yB} is defined as a function of y over the region $0<y<b$. From $y=0$ to $y=a$, it is given as the series function of (14), and is zero from $y=a$ to $y=b$. The B_0 and B_n coefficients are then obtained from the usual equations for determining Fourier coefficients

$$B_0 = \frac{1}{b}\int_0^b E_{yB}\big|_{z=0} dy$$
$$= \frac{1}{b}\int_0^a \left(A_0 + \sum_m A_m \cos\frac{m\pi y}{a}\right) dy$$
$$B_0 = (a/b)A_0. \tag{19}$$

Similarly,

$$B_n = \frac{2}{b}\int_0^b E_{yB}\big|_{z=0} \cos\frac{n\pi y}{b}\, dy$$
$$= \frac{2}{b}\int_0^a \left(A_0 + \sum_m A_m \cos\frac{m\pi y}{a}\right)\cos\frac{n\pi y}{b}\, dy$$
$$B_n = \frac{2A_0}{n\pi}\sin\frac{n\pi a}{b} + 2\frac{n}{\pi}\frac{a^2}{b^2}\sin\frac{n\pi a}{b}\sum_m \frac{A_m(-1)^m}{m^2\{(n^2a^2/m^2b^2)-1\}}. \tag{20}$$

We may also obtain an expression for A_m as a function of B_n from the boundary condition (18b), equating tangential components of H. Equations (15) and (17a) combine to give

$$Y_{A_0}A_0 + \sum_m Y_{A_m}A_m \cos\frac{m\pi y}{a}$$
$$= Y_{B_0}B_0 + \sum_n Y_{B_n}B_n \cos\frac{n\pi y}{b} \qquad 0 < y < a.$$

In the usual manner the average value, or $Y_{A_0}A_0$, is found from

$$Y_{A_0}A_0 = \frac{1}{a}\int_0^a \left(Y_{B_0}B_0 + \sum_n Y_{B_n}B_n \cos\frac{n\pi y}{b}\right) dy$$
$$Y_{A_0}A_0 = Y_{B_0}B_0 + \sum_n \frac{b}{n\pi a} Y_{B_n}B_n \sin\frac{n\pi a}{b}. \tag{21}$$

The $Y_{A_m}A_m$ coefficients are evaluated from

$$Y_{A_m}A_m = \frac{2}{a}\int_0^a \left(Y_{B_0}B_0 + \sum_n Y_{B_n}B_n \cos\frac{n\pi y}{b}\right)\cdot\cos\frac{m\pi y}{a}\, dy$$
$$Y_{A_m}A_m = \frac{2}{a}\sum_n Y_{B_n}B_n \frac{na^2}{m^2\pi b}\frac{(-1)^m \sin(n\pi a/b)}{\{(n^2a^2/m^2b^2)-1\}}. \tag{22}$$

B. The Equivalent Circuit

Before considering the evaluation of the B_n and A_m coefficients from (20) and (22), let us inquire into the meaning of the "average value" terms obtained. If we define voltage between conducting planes as

$$V = -\int \overline{E}\cdot\overline{dl}$$

then for the A space

$$V_A\big|_{z=0} = -\int_0^a A_0 dy - \int_0^a \sum_m A_m \cos\frac{m\pi y}{a}\, dy.$$
$$= -aA_0.$$

Similarly, for the B space

$$V_B\big|_{z=0} = -\int_0^b B_0 dy - \int_0^b \sum_n B_n \cos\frac{n\pi y}{b}\, dy.$$
$$= -bB_0.$$

Note now that by equation (19) $V_A = V_B$. By the above, it is also evident that the total voltage at any point is that in the principal wave, since $\int\overline{E}\cdot\overline{dl}$ for the higher-order waves will always be zero. This is consistent with the boundary conditions, and establishes the continuity of voltage at $z=0$. Further, note that the current in the top conductor due to the principal wave J_z is given by $+H_x$, and so current in the principal waves for the two spaces,

$$\left.\begin{aligned} J_{0A}\big|_{z=0} &= Y_{A_0}A_0 \\ J_{0B}\big|_{z=0} &= Y_{B_0}B_0 \end{aligned}\right\} \text{ amperes per meter width.}$$

and

However, (21) indicates that $J_{0A}\big|_{z=0} \neq J_{0B}\big|_{z=0}$, but differs by a factor as follows:

$$J_{0A} = J_{0B} + \sum_n \frac{b}{n\pi a} Y_{B_n}B_n \sin\frac{n\pi a}{b}.$$

This discontinuity of current may be represented by a lumped admittance at $z=0$. Rewrite the foregoing equation as $J_{0A} = J_{0B} - j\omega C_d' V$ where V is voltage at $z=0$, or $V\big|_{z=0} = -aA_0 = -bB_0$. Then $j\omega C_d'$ is

$$j\omega C_d' = \sum_n \frac{b}{n\pi a^2 A_0} Y_{B_n}B_n \sin\frac{n\pi a}{b}. \tag{23}$$

This capacitance C_d' will be useful if we can show that it is independent of both termination and voltage, for we may then represent the effect of the discontinuity by a lumped admittance at this point, having an exact equivalent circuit for all conditions of line termination. We must then show that neither Y_{B_n} nor B_n/A_0 are functions of the voltage or the principal wave admittances, the Y_0's. If the termination is not so close to the discontinuity as to require both positive and negative exponential solutions for the higher-order waves, then (12) holds and shows that Y_{B_n} at least is not a function of termination or of voltage, that is, of Y_{A_0}, Y_{B_0}, A_0, or B_0. Then substituting for Y_{B_n} from (17b) and B_n from (20)

$$C_d' = \frac{2\epsilon}{\pi^3}\left[\sum_n \frac{\sin^2 n\pi(a/b)}{K_{B_n} n^3(a^2/b^2)} + \sum_n \frac{\sin^2 n\pi(a/b)}{K_{B_n} n}\sum_m \frac{(-1)^m(A_m/A_0)}{m^2\{(n^2a^2/m^2b^2)-1\}}\right]. \tag{24}$$

Let us define two of the summations

$$T_{0B}(\alpha) = \sum_n \frac{\sin^2 n\pi\alpha}{K_{nB}n^3\alpha^2} \tag{25}$$

$$T_{mB}(\alpha) = \sum_n \frac{\sin^2 n\pi\alpha}{K_{nB}n\{(n^2\alpha^2/m^2 - 1)\}}. \tag{26}$$

By comparison with the Hahn functions given in the Appendix E, we see that the above may be written

$$T_{0B}(\alpha) = S_0(\alpha) + \sum_n \left(\frac{1}{K_{nB}} - 1\right)\frac{\sin^2 n\pi\alpha}{n^3\alpha^2} \tag{27}$$

$$T_{mB}(\alpha) = S_m(\alpha) + \sum_n \left(\frac{1}{K_{nB}} - 1\right)\frac{\sin^2 n\pi\alpha}{n\{(n^2\alpha^2/m^2) - 1\}}. \tag{28}$$

This separation is made since the latter series converge very rapidly, and values for the S functions are available; values of T's may then be calculated readily from (27) and (28). In fact, if transverse dimensions are small compared to wavelength, $K_n \cong 1$ (equation (13)) and the T functions reduce to the S functions, values of which are given in the Appendix E.

Reversing the order of summation in the last term of (24), this equation may now be written

$$C_d' = \frac{2\epsilon}{\pi^3}\left[T_{0B}(a/b) + \sum_m \frac{(-1)^m A_m/A_0}{m^2} T_{mB}(a/b)\right]. \tag{29}$$

Thus the first term in this expression is not a function of A_0, B_0, Y_{A0}, Y_{B0}. We shall show later that it is the most important term, and approximate values of C_d' may be calculated using only this term. However, we may also show that the ratio A_m/A_0 may be evaluated, so that the entire expression fulfills requirements for a useful equivalent circuit.

C. Determination of Wave Amplitudes.

To evaluate the ratio A_m/A_0, consider for the moment (22). Let us confine attention to some particular A_m term, say the pth term to avoid confusion in the summation. Then

$$Y_{A_p}A_p = \frac{2a(-1)^p}{p^2\pi b}\sum_n \frac{nY_{Bn}B_n \sin(n\pi a/b)}{\{(n^2a^2/p^2b^2) - 1\}}.$$

Substitute Y_{Bn} from (17b) and B_n from (20),

$$Y_{A_p}A_p = \frac{4a(-1)^p j\omega\epsilon A_0}{p^2\pi^3}\sum_n \frac{\sin^2(n\pi a/b)}{nK_{nB}\{(n^2a^2/p^2b^2) - 1\}}\left[1 + \frac{a^2n^2}{b^2}\sum_m \frac{(-1)^m A_m/A_0}{m^2\{(n^2a^2/m^2b^2) - 1\}}\right]. \tag{30}$$

The first summation of the foregoing equation is recognized as the $T_{pB}(a/b)$ series from (26). Consider the second term separately, reversing the order of summation.

$$\sum_m \frac{(-1)^m(A_m/A_0)}{m^2} \cdot \sum_n \frac{n^2(a^2/b^2)\sin^2(n\pi a/b)}{nK_{nB}\{(n^2a^2/p^2b^2) - 1\}\{(n^2a^2/m^2b^2) - 1\}}. \tag{31}$$

Through a partial fraction expansion, the following identity is obtained:

$$\frac{n^2(a/b)^2}{\{(n^2a^2/p^2b^2) - 1\}\{(n^2a^2/m^2b^2) - 1\}} = \frac{p^2m^2}{p^2-m^2}\left[\frac{1}{\{(n^2a^2/p^2b^2)-1\}} - \frac{1}{\{(n^2a^2/m^2b^2)-1\}}\right] p \neq m. \tag{32}$$

Then (31) may be written

$$\sum_{\substack{m \\ m\neq p}} \frac{(-1)^m(A_m/A_0)p^2}{p^2 - m^2}\left[\sum_n \frac{\sin^2(n\pi a/b)}{nK_{nB}\{(n^2a^2/p^2b^2) - 1\}} - \sum_n \frac{\sin^2(n\pi a/b)}{nK_{nB}\{(n^2a^2/m^2b^2) - 1\}}\right] + \frac{(-1)^pA_p}{A_0}\sum_n \frac{n(a^2/b^2)\sin^2(n\pi a/b)}{p^2K_{nB}\{(n^2a^2/p^2b^2) - 1\}^2}. \tag{33}$$

The last term in which m is set equal to p in (31) must be included since the partial fraction expansion does not hold for $m = p$. Two of the series are of the form (26). For the last, define

$$V_{pB}(\alpha) = \sum_n \frac{n\alpha^2 \sin^2 n\pi\alpha}{p^2K_{nB}\{(n^2\alpha^2/p^2) - 1\}^2}. \tag{34}$$

This is related to Hahn's U function, (82) of Appendix E.

$$V_{pB}(\alpha) = U_p(\alpha) + \sum_n \left(\frac{1}{K_{nB}} - 1\right)\frac{n\alpha^2 \sin^2 n\pi\alpha}{p^2\{(n^2\alpha^2/p^2) - 1\}^2}. \tag{35}$$

Thus with the double summation written in the form (33), and the definitions (26) and (34), (30) becomes

$$-(-1)^p\left[\frac{\pi^2}{pK_{pA}4} + \frac{V_{pB}(a/b)}{p^2}\right]\frac{A_p}{A_0} = \frac{1}{p^2}T_{pB}(a/b) + \sum_m \frac{(-1)^m(A_m/A_0)}{(p^2 - m^2)}[T_{pB}(a/b) - T_{mB}(a/b)]. \tag{36}$$

The above equation represents an infinite number of simultaneous equations, one for each value of p, in the infinite number of A_m's. In matrix form,

$$[g_{pm}][A_m/A_0] = [h_p] \tag{37}$$

where

$$\begin{aligned} g_{pm} &= [(-1)^m/(p^2 - m^2)][T_{pB}(a/b) - T_{mB}(a/b)] \quad (m \neq p) \\ g_{pp} &= (-1)^p\left[\frac{V_{pB}(a/b)}{p^2} + \frac{\pi^2}{4pK_{pA}}\right] \\ h_p &= -(1/p^2)T_{pB}(a/b). \end{aligned} \tag{38}$$

Formally, the ratios A_m/A_0 are completely determined by (37) in terms of the ratios a/b and b/λ, and are not functions of the Y_0's, proving our earlier statement. Practically, we may solve for a finite number of these by retaining that number of simultaneous equations. We shall demonstrate this by a numerical example.

D. Numerical Example

Let us consider the case of $a/b = 1/2$ at low frequencies ($b \ll \lambda$) so that the T functions reduce directly to Hahn's S functions. Using values of these from Appendix E, the coefficients (38), retaining four of the equations (37), are

$$\begin{bmatrix} -4.695 & -0.110 & +0.066 & -0.0453 \\ +0.110 & +2.403 & +0.04 & -0.0292 \\ +0.0663 & -0.045 & -1.61 & -0.021 \\ +0.045 & -0.029 & +0.021 & +1.212 \end{bmatrix} \begin{bmatrix} A_1/A_0 \\ A_2/A_0 \\ A_3/A_0 \\ A_4/A_0 \end{bmatrix} = \begin{bmatrix} 1.00 \\ 0.333 \\ 0.170 \\ 0.105 \end{bmatrix}$$

A method of successive approximations, using first the main diagonal terms only, and successively correcting by resubstituting first approximations for other than the main diagonal terms in the equations, leads to values of A_m/A_0 very quickly, and is applicable to any number of equations. For the above,

$A_m/A_0=$	$m=1$	$m=2$	$m=3$	$m=4$
	-0.219	0.151	-0.1192	0.1003

Substituting in (29)

$$C_d'=(2\epsilon/\pi^3)[4.21-0.219-0.050-0.020-0.011].$$

It is seen that four A_m terms in the summation were sufficient.

$$C_d'=0.253\epsilon.$$

For air or space, $\epsilon=1/36\pi\times10^{-9}$ farad per meter $=$ 0.0885 micromicrofarad per centimeter so $C_d'=0.0224$ micromicrofarad per centimeter.

Although they are not required for the equivalent circuit, it is interesting to note the wave coefficients in the B space, calculated from (20).

$B_n/B_0=$	$n=1$	$n=2$	$n=3$	$n=4$	$n=5$	$n=6$	$n=7$	$n=8$
	1.16	-0.220	-0.532	0.152	0.344	-0.119	-0.200	$+0.100$

E. Approximate Expressions

The above numerical example is typical, demonstrating that the major part of the capacitance arises from the first term in (29). For a first approximation to capacitance, we should then not need to calculate the coefficients A_m/A_0 from the matrix (37),

$$C_d' \cong (2\epsilon/\pi^3)T_{0B}(a/b) = (2\epsilon/\pi^3)S_0(a/b),\ (b/\lambda) \ll 1. \quad (39)$$

Values of $S_0(a/b)$ are tabulated over a wide range in Appendix E. For small values of a/b, the approximate value of the S_0 function may be used. The resulting equation, including the effect of the coefficients A_m/A_0, is then

$$C_d' \cong \epsilon[(2/\pi)\ln(b/a) - 0.268]. \quad (40)$$

Equation (40) differs from (29) by less than 10 per cent if a/b is less than 0.3.

For values of capacitance when transverse dimensions approach a half wavelength, the K's depart from the value of unity and some correction terms in (27), (28), and (35) must be calculated. To the extent that only one correction term is required, (meaning that the approach of the $n=1$ E wave to cutoff is of the most importance) the ratio of susceptance to that calculated from the low-frequency value of C_d' is

$$\frac{S_d'}{\omega C_d'} \cong \frac{1}{\sqrt{1-(2b/\lambda)^2}}. \quad (41)$$

This approximate relation has been compared with complete calculations for $S_d'/\omega C_d'$ in the range of b from 0 to $\lambda/2$ in Fig. 16.

It is interesting to note that the approximate result (39) could be obtained by the first step in a converging step-by-step method sometimes used in the analysis of discontinuities. In this method the principal wave only is considered in the region A, then the first approximation to E-wave components in B is obtained by expanding these in Fourier series, then the E-wave components in region A are obtained approximately by the discontinuity in E_z, etc. The method followed in this report has included simultaneously all of these steps.

Appendix A

Analysis for Discontinuities Figs. 2 to 8

(1) It is apparent that Fig. 7 is the general case covering all cases, Figs. to 1 to 8. Following the analysis for the step given in Part VI, we may write

$$\begin{aligned} E_{yB}|_{z=0} &= B_0 + \sum_n B_n \cos\frac{n\pi y}{b} \\ H_{xB}|_{z=0} &= Y_{B_0}B_0 + \sum_n Y_{B_n}B_n \cos\frac{n\pi y}{b} \end{aligned} \quad (42)$$

$$\begin{aligned} E_{yC}|_{z=d} &= C_0 + \sum_q C_q \cos\frac{q\pi y}{c} \\ H_{xC}|_{z=d} &= Y_{C_0}C_0 + \sum_q Y_{C_q}C_q \cos\frac{q\pi y}{c} \end{aligned} \quad (43)$$

$$\begin{aligned} E_{yA} &= [A_0e^{-jkz}+A_0'e^{jkz}] \\ &\quad + \sum_m [A_me^{-\gamma_m z}+A_m'e^{\gamma_m z}]\cos\frac{m\pi y}{a} \\ H_{xA} &= -1/\eta[A_0e^{-jkz}-A_0'e^{jkz}] \\ &\quad + \sum_m Y_{A_m}[A_me^{-\gamma_m z}-A_m'e^{\gamma_m z}]\cos\frac{m\pi y}{a} \end{aligned} \quad (44)$$

$$\begin{aligned} Y_{B_n} &= \frac{j\omega\epsilon b}{n\pi K_{B_n}} \quad Y_{C_q} = -\frac{j\omega\epsilon c}{q\pi K_{C_q}} \quad Y_{A_m} = -\frac{j\omega\epsilon a}{m\pi K_{A_m}} \\ K_{B_n} &= \sqrt{1-(2b/n\lambda)^2} \qquad K_{C_q} = \sqrt{1-(2c/q\lambda)^2} \\ K_{A_m} &= \sqrt{1-(2a/m\lambda)^2}. \end{aligned} \quad (45)$$

The matching operations, as explained in detail in Part VI, follow these steps

(a) Obtain equations of B's in terms of A's from continuity condition $E_{yB}|_{z=0}=E_{yA}|_{z=0}$, $0<y<a$; $E_{yB}|_{z=0}=0$, $a<y<b$.

(b) Obtain equations of C's in terms of A's from condition $E_{yC}|_{z=d}=E_{yA}|_{z=d}$, $0<y<a$; $E_{yC}|_{z=d}=0$, $a<y<c$.

(c) Obtain equations of A's in terms of B's from condition $H_{xA}|_{z=0}=H_{xB}|_{z=0}$, $0<y<a$.

(d) Obtain equations of A's in terms of C's from condition $H_{xA}|_{z=d}=H_{xC}|_{z=d}$, $0<y<a$.

If we define $F_m=A_m+A_m'$ $F_m'=A_m-A_m'$

$$G_m = A_me^{-\gamma_m d} + A_m'e^{\gamma_m d} \quad G_m' = A_me^{-\gamma_m d} - A_m'e^{\gamma_m d}. \quad (46)$$

The resulting equations are

$$aF_0 = bB_0 \tag{47}$$

$$aG_0 = cC_0 \tag{48}$$

$$-\frac{F_0'}{\eta} = Y_{B_0}B_0 + \sum_n \frac{Y_{B_n}B_n b}{n\pi a}\sin\frac{n\pi a}{b} \tag{49}$$

$$-\frac{G_0'}{\eta} = Y_{C_0}C_0 + \sum_q \frac{Y_{C_q}C_q c}{q\pi a}\sin\frac{q\pi a}{c} \tag{50}$$

$$B_n = \frac{2F_0}{n\pi}\sin\frac{n\pi a}{b}\left[1+\frac{a^2n^2}{b^2}\sum_m \frac{(-1)^m F_m/F_0}{m^2\{(n^2a^2/m^2b^2)-1\}}\right] \tag{51}$$

$$C_q = \frac{2G_0}{q\pi}\sin\frac{q\pi a}{c}\left[1+\frac{a^2q^2}{c^2}\sum_m \frac{(-1)^m G_m/G_0}{m^2\{(q^2a^2/m^2c^2)-1\}}\right] \tag{52}$$

$$Y_{A_m}F_m' = \frac{2a(-1)^m}{m^2\pi b}\sum_n \frac{nY_{B_n}B_n\sin(n\pi a/b)}{\{(n^2a^2/m^2b^2)-1\}} \tag{53}$$

$$Y_{A_m}G_m' = \frac{2a(-1)^m}{m^2\pi c}\sum_q \frac{qY_{C_q}C_q\sin(q\pi a/c)}{\{(q^2a^2/m^2c^2)-1\}}. \tag{54}$$

The above equations (47) and (48) again show that voltage is continuous in the principal wave, but (49) and (50) show that current for the principal wave is discontinuous by the factor of the summation. This discontinuity in current can be written in terms of lumped capacitances. For a unit width,

$$C_{1d}' = \frac{[J_{0B}-J_{0A}]_{z=0}}{j\omega(-aF_0)} = \sum_n \frac{Y_{B_n}bB_n}{j\omega n\pi a^2F_0}\sin\frac{n\pi a}{b}.$$

Substitute Y_{B_n} from (45), B_n from (51)

$$C_{1d}' = \frac{2\epsilon}{\pi^3}\left[\sum_n \frac{\sin^2(n\pi a/b)}{K_{B_n}n^3(a/b)^2} + \sum_n \frac{\sin^2(n\pi a/b)}{K_{B_n}n}\sum_m \frac{(-1)^m(F_m/F_0)}{m^2\{(n^2a^2/m^2b^2)-1\}}\right].$$

By the definitions (25) and (26)

$$C_{1d}' = \frac{2\epsilon}{\pi^3}\left[T_{0B}(a/b)+\sum_m \frac{(-1)^m(F_m/F_0)}{m^2}T_{mB}(a/b)\right]. \tag{55}$$

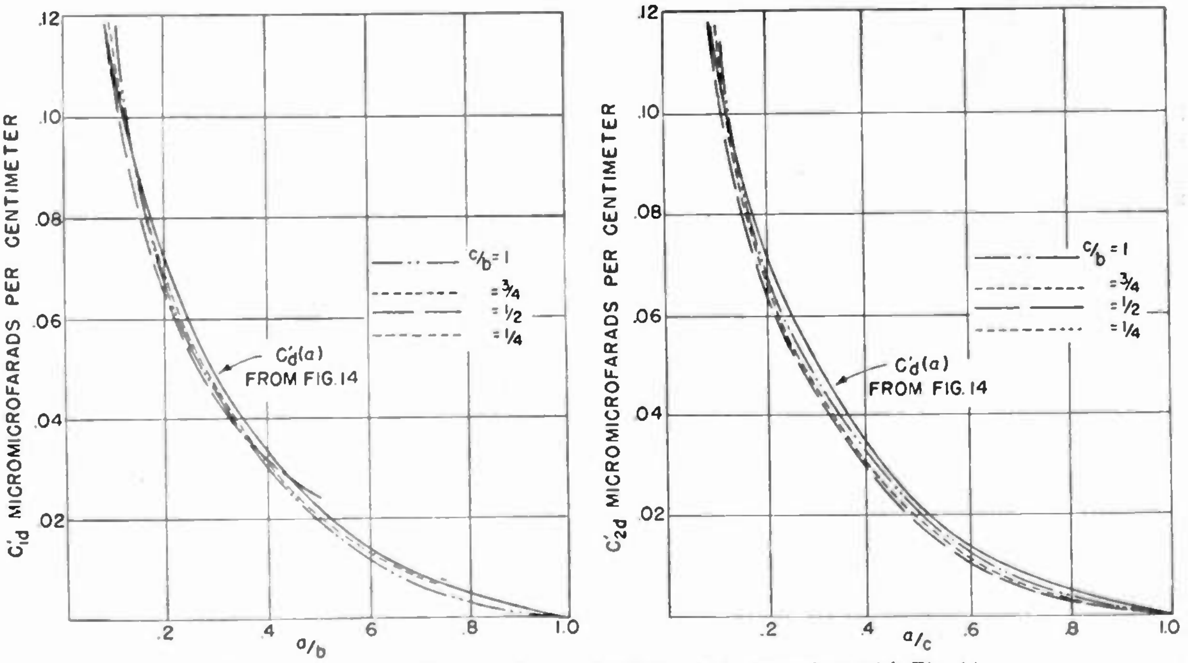


Fig. 23—Discontinuity capacitances for Fig. 2 and comparison with Fig. 14.

Similarly,

$$C_{2d}' = \frac{[J_{0A}-J_{0C}]_{z=d}}{j\omega(-aG_0)}$$

$$C_{2d}' = \frac{2\epsilon}{\pi^3}\left[T_{0C}(a/c)+\sum_m \frac{(-1)^m(G_m/G_0)}{m^2}T_{mC}(a/c)\right]. \tag{56}$$

The expressions (55) and (56) show that at least to the extent that the summation terms represent the usual order of correction to T_{0C}, the lumped capacitances to place at the two discontinuities are those for the corresponding steps,

$$C_{1d}' \cong C_d'(a/b) \quad \text{and} \quad C_{2d}' \cong C_d'(a/c) \tag{57}$$

where $C_d'(\alpha)$ is defined by (29), and plotted in Fig. 14. We should, of course, expect this result if d is large compared to a, but not necessarily when d/a is small. We may check the conclusion on two special cases.

(2) Special Case of $d=0$ (Fig. 2).

If $d=0$, it follows from (46) that $F_m=G_m$, $F_m'=G_m'$. Then (53) and (54) may be equated, with B_n substituted from (51), C_q from (52). The reversal of summation, breaking into partial fractions, and identification of the Hahn functions leads to an infinite matrix of equations as demonstrated in detail for the corresponding step in Part VI, leading to (38)

$$[g_{pm}][F_m/F_0] = [h_p] \tag{58}$$

where

$$\underset{m\neq p}{g_{pm}} = \{(-1)^m/(p^2-m^2)\}[T_{pB}(a/b)+T_{pC}(a/c) - T_{mB}(a/b) - T_{mC}(a/c)]$$

$$g_{pp} = \{(-1)^p/p^2\}[V_{pB}(a/b)+V_{pC}(a/c)]$$

$$h_p = -(1/p^2)[T_{pB}(a/b)+T_{pC}(a/c)].$$

By retaining only a finite number of the equations defined by (58) (usually 4 are sufficient) that number of values of F_m/F_0 are obtained by simultaneous solution, and these may be substituted in (55) and (56) to correct the first approximation to C_{1d}' and C_{2d}'. The curves of Fig. 23 show such values, assuming $b \ll \lambda$, for several values of b/c. It is seen that the approximate results (57) are in excellent agreement with the more accurately determined curves, for all values of b/c.

(3) Special case of $b=c$, $d \neq 0$.

This represents another special case in which the matrix equation may be readily obtained and solved, at least for low frequencies. Results for the corrected capacitances $C_{1d}' = C_{2d}' = C_d'$ are plotted versus a/b for several d/a ratios in Fig. 24. Comparing with $C_d'(a/b)$,

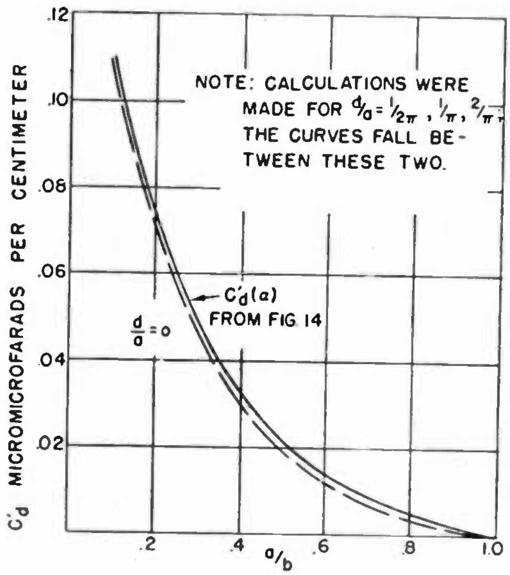


Fig. 24—Discontinuity capacitances for Fig. 3 and comparison with Fig. 14.

it is seen that the latter is a good approximation to the actual C_d' for all values of d/a.

Appendix B

Analysis for Discontinuity of Fig. 9

For this problem, we employ a small A region of infinitesimal length between the B and C regions, extending from $y=0$ to $y=a$. At $z=0$, series forms for E_y and H_z in the B and A regions may then be written as in equations (14), (15), (16), (17). For the region C at $z=0$,

$$E_{yC}|_{z=0} = C_0 + \sum_q C_q \cos\frac{q\pi y'}{c} \qquad (59)$$

$$H_{zC}|_{z=0} = Y_{C0}C_0 + \sum_q Y_{C_q}C_q \cos\frac{q\pi y'}{c}$$

where $y' = y + (c-a)$

$$Y_{B_n} = (j\omega\epsilon b/n\pi K_{B_n}) \qquad Y_{C_q} = -(j\omega\epsilon c/q\pi K_{C_q}).$$

The matching operations are these: (1) Obtain B's in terms of A's by matching $E_B = E_A$, $0<y<a$; $E_{yB}=0$, $a<y<b$. Results are the same as (19) and (20). (2) Obtain C's in terms of A's by matching $E_{yC}=0$, $0<y'<(c-a)$; $E_{yC}=E_{yA}$, $(c-a)<y'<c$.

$$cC_0 = aA_0 \qquad (60)$$

$$C_q = \frac{2A_0(-1)^q}{q\pi}\sin\frac{q\pi a}{c} \cdot \left[1 + \frac{q^2a^2}{c^2}\sum_m \frac{A_m/A_0}{m^2\{(q^2a^2/m^2c^2)-1\}}\right]. \qquad (61)$$

(3) Obtain A's in terms of B's by matching $H_{zA}=H_{zB}$, $0<y<a$. Results are the same as equations (21) and (22). (4) Obtain A's in terms of C's by matching $H_{zA}=H_{zC}$, $c-a<y'<c$.

$$Y_{A_0}A_0 = Y_{C0}C_0 + \frac{c}{\pi a}\sum_q \frac{Y_{cq}C_q(-1)^q}{q}\sin\frac{q\pi a}{c} \qquad (62)$$

$$Y_{A_m}A_m = \frac{2a}{m^2\pi c}\sum_q \frac{(-1)^qY_{C_q}C_q q\sin(q\pi a/c)}{\{(q^2a^2/m^2c^2)-1\}}. \qquad (63)$$

Note that in carrying through steps (2) and (4) it is convenient to use a substitution, $y''=a-y=c-y'$.

Study of the zero-order terms reveals the usual continuity of voltage but discontinuity of zero-order current $Y_{B0}B_0 - Y_{C0}C_0$ which may be accounted for by a capacitance

$$C_d' = \frac{b^2\epsilon}{\pi^2a^2}\left[\sum_n \frac{B_n\sin(n\pi a/b)}{n^2K_{B_n}A_0} + \frac{c^2}{b^2}\sum_q \frac{(-1)^qC_q\sin(q\pi a/c)}{q^2K_{C_q}A_0}\right].$$

Substituting the results for B_n and C_q from (20) and (61), and identifying the functions defined by (25) and (26),

$$C_d' = \frac{2\epsilon}{\pi^3}\left\{\left[T_{0B}(a/b) + \sum_m \frac{(-1)^mA_m/A_0}{m^2}T_{mB}(a/b)\right] + \left[T_{0C}(a/c) + \sum_m \frac{A_m/A_0}{m^2}T_{mC}(a/c)\right]\right\}. \qquad (64)$$

Again, subject to more accurate evaluation of the summation, we may write

$$C_d' \cong C_d'(a/b) + C_d'(a/c). \qquad (65)$$

To obtain the more accurate values, we must find the matrix for A_m/A_0. Equate results for $Y_{A_m}A_m$ from (22) and (63), substitute (20) and (61), and separate in the manner detailed in Part VI. The matrix is

$$[g_{p_m}][A_m/A_0] = [h_p] \qquad (66)$$

$$g_{p_m}\underset{m\neq p}{=} \{1/(p^2-m^2)\}\{(-1)^{p+m}[T_{pB}(a/b) - T_{mB}(a/b)] + [T_{pC}(a/c) - T_{mC}(a/c)]\}$$

$$g_{p_p} = (1/p^2)[V_{pB}(a/b) + V_{pC}(a/c)]$$

$$h_p = -(1/p^2)[(-1)^pT_{pB}(a/b) + T_{pC}(a/c)].$$

In Fig. 25 values from the complete equation are plotted versus a/b for the case $a/c=1/2$, and compared with results from the approximate equation (65). Agreement is again good.

Appendix C

Analysis for Discontinuity of Fig. 12

In Fig. 12 we may write solutions for the components E_y, and H_z in the three regions A, B, and C in the forms

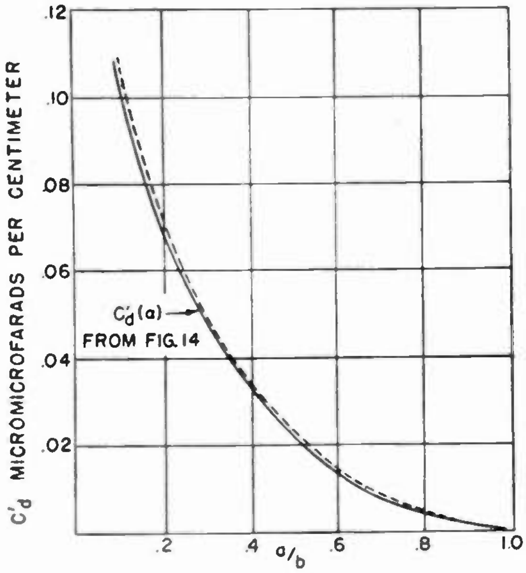


Fig. 25—Comparison of discontinuity capacitance for one case of Fig. 9 with values from Fig. 14.

of (14), (15), (16), (17), and (59), where $y'=y-f$ and $f=b+d$.

$$Y_{A_m} = -(j\omega\epsilon a/m\pi K_{A_m}) \qquad Y_{B_n} = (j\omega\epsilon b/n\pi K_{B_n}) \qquad Y_{C_q} = (j\omega\epsilon c/q\pi K_{C_q}). \tag{67}$$

The matching operations follow in this manner: (1) Obtain A's in terms of B's and C's by matching $E_{yA}=E_{yB}$, $0<y<b$; $E_{yA}=0$, $b<y<f$; $E_{yA}=E_{yC}$, $f<y<a$. (2) Obtain B's in terms of A's by matching $H_{zB}=H_{zA}$, $0<y<b$. (3) Obtain C's in terms of A's by matching $H_{zC}=H_{zA}$, $f<y<a$. Resulting equations are

$$aA_0 = bB_0 + cC_0 \tag{68}$$

$$A_m = \frac{2B_0}{m\pi}\sin\frac{m\pi b}{a}\left[1+\frac{m^2b^2}{a^2}\sum_n \frac{(-1)^n B_n}{n^2B_0\{(m^2b^2/n^2a^2)-1\}}\right] - \frac{2C_0}{m\pi}\sin\frac{m\pi f}{a}\left[1+\frac{m^2c^2}{a^2}\sum_q \frac{C_q}{q^2C_0\{(m^2c^2/q^2a^2)-1\}}\right] \tag{69}$$

$$Y_{B_0}B_0 = Y_{A_0}A_0 + \frac{a}{\pi b}\sum_m \frac{Y_{A_m}A_m}{m}\sin\frac{m\pi b}{a} \tag{70}$$

$$Y_{C_0}C_0 = Y_{A_0}A_0 - \frac{a}{\pi c}\sum_m \frac{Y_{A_m}A_m}{m}\sin\frac{m\pi f}{a} \tag{71}$$

$$Y_{B_n}B_n = \frac{2b(-1)^n}{n^2\pi a}\sum_m \frac{mY_{A_m}A_m\sin(m\pi b/a)}{\{(m^2b^2/n^2a^2)-1\}} \tag{72}$$

$$Y_{C_q}C_q = -\frac{2c}{q^2\pi a}\sum_m \frac{mY_{A_m}A_m\sin(m\pi f/a)}{\{(m^2c^2/q^2a^2)-1\}}. \tag{73}$$

Equation (68) shows that voltage across lines B and C must equal that across line A, suggesting a series connection. However, equations (70) and (71) show that zero-order current is not continuous, requiring additional shunting elements at the junction. If (67) and (69) are substituted in (70) and (71), these may be written

$$J_{0B} = J_{0A} + j\omega C_B'V_{0B} + j\omega C_A'V_{0A}$$
$$J_{0C} = J_{0A} + j\omega C_c'V_{0C} + j\omega C_A'V_{0A} \tag{74}$$

where

$$C_A' = -\frac{2a^2\epsilon}{\pi^3bc}\sum_m \frac{\sin(m\pi b/a)\sin(m\pi f/a)}{m^3K_{Am}} \cdot\left[1+\frac{m^2c^2}{a^2}\sum_q \frac{C_q}{q^2C_0\{(m^2c^2/q^2a^2)-1\}}\right] \tag{75}$$

$$C_B' = \frac{2a^2\epsilon}{\pi^3b^2}\sum_m \frac{\sin^2(m\pi b/a)}{m^3K_{Am}} \cdot\left[1+\frac{m^2b^2}{a^2}\sum_n \frac{(-1)^nB_n}{n^2B_0\{(m^2b^2/n^2a^2)-1\}}\right] - C_A \tag{76}$$

$$C_C' = \frac{2a^2\epsilon}{\pi^3c^2}\sum_m \frac{\sin^2(m\pi f/a)}{m^3K_{Am}} \cdot\left[1+\frac{m^2c^2}{a^2}\sum_q \frac{C_q}{q^2C_0\{(m^2c^2/q^2a^2)-1\}}\right] - C_A. \tag{77}$$

Equations (74), combined with (68), lead to the equivalent circuit given in Fig. 12, with values of capacitance at the junction defined by (75), (76), and (77). To complete the problem, we should show that C_q/C_0 and B_n/B_0 are independent of the terminations so that (75) to (77) give values of capacitance independent of terminations. The matrix equation is obtained conveniently only for the case $d=0$, and the proof for that case is complete. The proof seems more difficult for the general case of $d\neq 0$, but a few numerical checks support our belief that the major part of the contributions to capacitance arise from the first summations in (75) to (77), which are independent of B_n/B_0 and C_q/C_0, as we have found in previous examples. Curves of Fig. 17 were calculated on the basis of this assumption, showing C_A', C_B', and C_C' as functions of $b/(a-d)$ for different values of d/a.

Note from the curves that (75) gives a negative value for C_A'. Detailed study of certain special cases shows that C_a should be negative if the equivalent circuit is correct. Two of these cases are:

1. The case of either line B or line C short-circuited at $z=0$; the sum of the remaining two discontinuity capacitances should then equal that for the corresponding step.

2. The case of $d=0$, lines B and C perfectly terminated, and a wave traveling to the left from line A; the three capacitances should then combine so that there is no reflection.

Appendix D

The Discontinuity of Fig. 13

Fig. 13 is obviously a special case of Fig. 12, with the A line terminated by a short circuit at $z=l$. If this termination is close to $z=0$, it will be necessary to include positive as well as negative exponentials in the A space. The resulting combination of positive and negative exponentials, with $E_y=0$ at $z=l$ for all waves, gives the following value for Y_{Am}

$$Y_{Am} = -(j\omega\epsilon a/m\pi K_{Am})\coth(m\pi K_{Am}l)/a. \tag{78}$$

By using this value in equations (68) to (73), the appropriate values of C_A', C_B', and C_C' may be calculated. In particular, for the case of $d=0$, $a\ll\lambda$, it was found that all previous values of C_A', C_B', and C_C' may be multiplied by a factor which is only a function of l/a, and is plotted in Fig. 18. This factor may be approximated by coth $\pi l/a$ to the degree shown in the Fig. 18. Note that until l/a is less than 0.5, the correction is less than 10 per cent; at $l/a=1$ it is practically negligible. These results may be used as a guide, showing approximately when any termination is electrically close to the discontinuity.

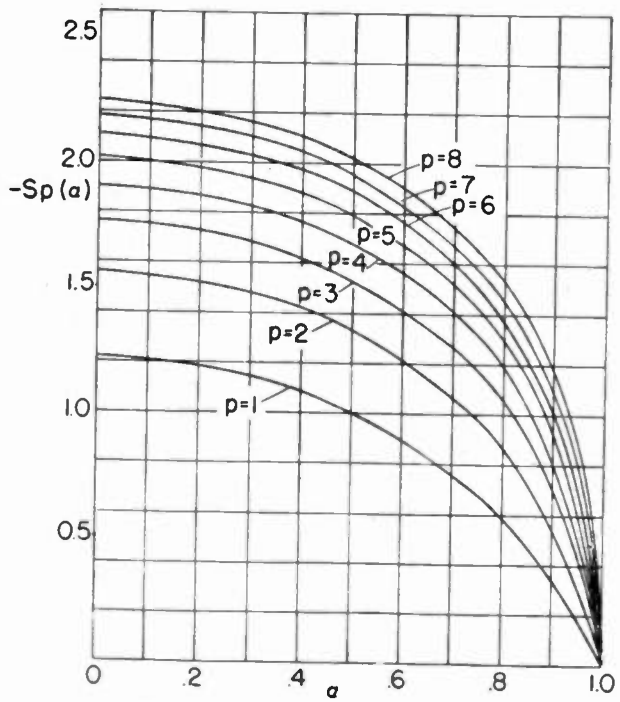


Fig. 26—The Hahn functions $S_p(\alpha)$.

Note that since we have taken $a \ll \lambda$, this case will also infer that $l \ll \lambda$. The short-circuited line A then looks like an inductance $L=\mu la/w$. For air or space, $\mu=4\pi \times 10^{-7}$ henry per meter so

$$L = 4\pi \times 10^{-7}(la/w) \text{ henry (dimensions meters)}$$
$$= 4\pi \times 10^{-9}(la/w) \text{ henry (dimensions centimeters)} \quad (79)$$

w is width of the line.

Appendix E

The Hahn Functions

Hahn has defined and calculated values for the following summations

$$S_0(\alpha) = \sum_{n=1}^{\infty} \frac{\sin^2 n\pi\alpha}{\alpha^2 n^3} \quad (80)$$

$$S_p(\alpha) = \sum_{n=1}^{\infty} \frac{\sin^2 n\pi\alpha}{n\{(n^2\alpha^2/p^2) - 1\}} \quad (81)$$

$$U_p(\alpha) = \sum_{n=1}^{\infty} \frac{n\alpha^2 \sin^2 n\pi\alpha}{p^2\{(n^2\alpha^2/p^2) - 1\}^2}. \quad (82)$$

For small values of α,

$$S_0(\alpha) \cong \pi^2[\ln (1/a) - 0.338]. \quad (83)$$

Hahn[1] has given a quite complete table of values for $S_p(\alpha)$. (Note that all values in his table should be negative.) Curves taken from this table are plotted in Fig. 26.

For values of $U_p(\alpha)$ it is possible to neglect variation of these functions with α in their first approximation (and since they enter only in the second-order term for capacitance, this is good enough for our purposes). Then

$$U_1 \cong 2.23; \quad U_2 \cong 4.69; \quad U_3 \cong 7.12;$$
$$U_4 \cong 9.54; \quad U_p \cong 0.97(p\pi^2/4).$$

For a very complete table of the S_0 summation, we include results calculated by the Mathematical Tables Project of the Works Projects Administration under the direction of Dr. Arnold N. Lowan.

TABLE I

α	$S_0(\alpha)$	α	$S_0(\alpha)$	α	$S_0(\alpha)$	α	$S_0(\alpha)$
0.01	42.117	0.26	10.14	0.51	4.041	0.76	1.0877
0.02	35.276	0.27	9.787	0.52	3.880	0.77	1.0095
0.03	31.276	0.28	9.443	0.53	3.722	0.78	0.93403
0.04	28.439	0.29	9.113	0.54	3.569	0.79	0.86125
0.05	26.924	0.30	8.795	0.55	3.421	0.80	0.79116
0.06	24.442	0.31	8.488	0.56	3.276	0.81	0.72377
0.07	22.924	0.32	8.192	0.57	3.135	0.82	0.65907
0.08	21.611	0.33	7.906	0.58	2.997	0.83	0.59706
0.09	20.453	0.34	7.630	0.59	2.864	0.84	0.53775
0.10	19.418	0.35	7.364	0.60	2.734	0.85	0.48115
0.11	18.483	0.36	7.105	0.61	2.607	0.86	0.42728
0.12	17.630	0.37	6.856	0.62	2.484	0.87	0.37616
0.13	16.847	0.38	6.613	0.63	2.365	0.88	0.32784
0.14	16.123	0.39	6.379	0.64	2.248	0.89	0.28234
0.15	15.450	0.40	6.151	0.65	2.135	0.90	0.23973
0.16	14.822	0.41	5.930	0.66	2.025	0.91	0.20006
0.17	14.232	0.42	5.716	0.67	1.918	0.92	0.16341
0.18	13.678	0.43	5.508	0.68	1.814	0.93	0.12988
0.19	13.154	0.44	5.306	0.69	1.713	0.94	0.099584
0.20	12.659	0.45	5.110	0.70	1.615	0.95	0.072684
0.21	12.188	0.46	4.919	0.71	1.520	0.96	0.049373
0.22	11.741	0.47	4.733	0.72	1.428	0.97	0.029917
0.23	11.315	0.48	4.553	0.73	1.339	0.98	0.014692
0.24	10.907	0.49	4.378	0.74	1.252	0.99	0.0042972
0.25	10.518	0.50	4.207	0.75	1.169	1.00	0

Acknowledgments

The authors wish to express thanks to Mr. G. E. Feiker of the General Engineering Laboratory for measurements reported in Part IV, to Mrs. Theo E. Robbins for her careful work in computation of the curves given, to Dr. Arnold N. Lowan of the Mathematical Tables Project for the calculation of Table I, and to Mr. W. C. Hahn for the inspiration of his original work.

A Dummy Dipole Network*

HANS SALINGER†, ASSOCIATE, I.R.E.

Summary—**Design data are presented for a network which closely simulates the impedance of a dipole antenna in the range from one half to twice its series resonance frequency.**

RADIO transmitters are generally designed for use with a specified aerial, and in the majority of cases they have to operate with the same antenna over a certain range of frequencies. It then is necessary to ascertain whether they deliver sufficient power to the antenna all over this range. As it is inconvenient to perform this test on an actual antenna, a "dummy antenna" is customarily used, i.e., a network the impedance of which approximates that of the actual antenna. In this note, design data will be presented of a simple network to be used as a dummy for a $\frac{1}{2}\lambda$ dipole, the characteristics of which it simulates over a range from about one half to twice the series resonant frequency.

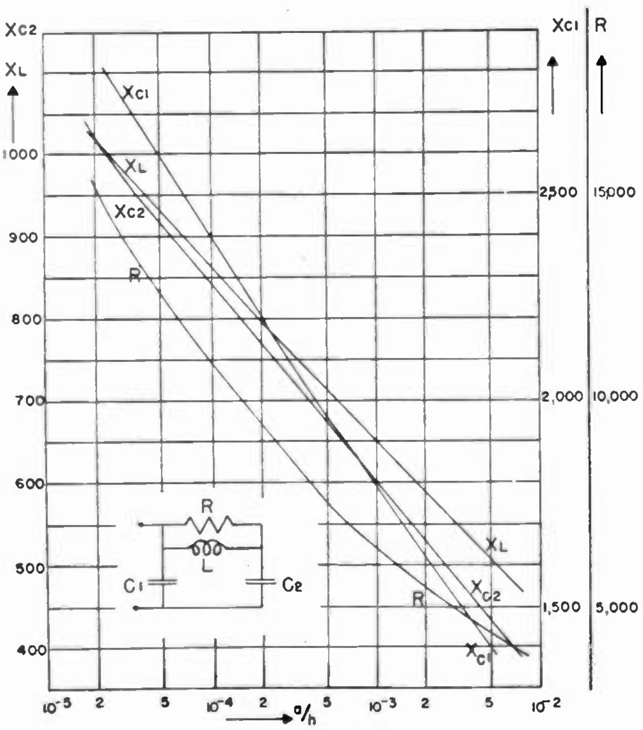


Fig. 1—The network and its design curves.

The design is based on data recently supplied by King and Blake[1] which evidently are very reliable. Their curves contain the parameter a/λ, where a stands for the conductor radius. As in our problem λ is variable whereas h (half the dipole length) remains constant, the curves had to be redrawn for the parameter a/h; the results appear in Table I, for three values of a/h, viz., 4×10^{-5}, 4×10^{-4}, and 4×10^{-3}. For self-supporting short-wave antennas higher values of a/h occur, and it would be desirable to extend the curves of King and Blake correspondingly. But the approximations used by them seem to become invalid in the extended range; therefore, the present investigation has been restricted to the range covered by their curves.

TABLE I
Comparison of Dipole and Network Impedance

a/h	f/f_0	Z (dipole)	Z (network)
4×10^{-5}	$1/\pi=0.318$	$5-j\,2050$	$4.04-j\,1938$
	$2/\pi=0.637$	$22-j\,750$	$19.3-j\,728$
	$3/\pi=0.955$	$61-j\,47$	$59.4-j\,81.8$
	1	$69+j\,36$	$69.2+j\,2.2$
	$4/\pi=1.273$	$157+j\,540$	$182.3+j\,591$
	$5/\pi=1.591$	$550+j\,1650$	$778.5+j\,1792$
	$6/\pi=1.910$	$7040+j\,2160$	$6740+j\,1302$
	2	$5720-j\,3800$	$6180-j\,2790$
4×10^{-4}	$1/\pi=0.318$	$5-j\,1550$	$4.08-j\,1509$
	$2/\pi=0.637$	$22-j\,540$	$18.9-j\,532$
	$3/\pi=0.955$	$60-j\,28$	$58.4-j\,42.2$
	1	$68+j\,34$	$68.1+j\,22.1$
	$4/\pi=1.273$	$155+j\,440$	$180.7+j\,465$
	$5/\pi=1.591$	$540+j\,1250$	$757.5+j\,1340$
	$6/\pi=1.910$	$4740+j\,660$	$4410+j\,273$
	2	$3450-j\,2400$	$3860-j\,1791$
4×10^{-3}	$1/\pi=0.318$	$5-j\,1050$	$3.81-j\,988$
	$2/\pi=0.637$	$22-j\,350$	$18.1-j\,327$
	$3/\pi=0.955$	$58-j\,12$	$56.7+j\,17.4$
	1	$66+j\,32$	$66.0+j\,62.8$
	$4/\pi=1.273$	$149+j\,340$	$175+j\,380$
	$5/\pi=1.591$	$520+j\,800$	$715+j\,944$
	$6/\pi=1.910$	$2600-j\,200$	$2600-j\,241$
	2	$1900-j\,1350$	$2180-j\,1129$

A simple π mesh as shown in the lower part of Fig. 1 was found to simulate the dipole rather closely. The elements R, L, C_1, and C_2 were calculated to give the best fit, and these values may be taken from Fig. 1; the reactances X_L, X_{C1}, and X_{C2} mean $2\pi f_0 L$, etc., f_0 being the frequency at which $4h=\lambda$ (this is not exactly the frequency where the dipole reactance vanishes, as can be seen from the insert of Fig. 2).

Table I compares the dipole and network impedances for different values of f/f_0 and a/h. For the lowest and highest of the a/h values, curves have been drawn; Fig. 3 shows the resistance, Fig. 2 the reactance behavior. Some parts of the curves have been drawn separately on a larger scale.

Discussion

From Table I, it would seem that the reactances do not fit too well at certain frequencies. But the drawings show that this is only due to a horizontal shift of the curves, which means that the network does not tune exactly at the same frequency as the dipole. This is considered as insignificant, inasmuch as the dipole itself is slightly detuned from the $\lambda/4$ value. It is possible to design the elements so that network and dipole tune at the same point, but then the resistance match will not be as good, and it seemed to be more important to match the

* Decimal classification: R320. Original manuscript received by the Institute, June 1, 1943.
† Farnsworth Television and Radio Corporation, Fort Wayne, Indiana.
[1] R. King and F. G. Blake, "The self-impedance of a symmetrical antenna," Proc. I.R.E., vol. 30, pp. 335–349; July, 1942.

resistances, in order to make the network useful as a dummy load for power measurements. With the present values, the resistances fit reasonably well in the whole frequency range, except that the dipole resistance curves show a sag near $f/f_0 = 1.6$ which it was not possible to simulate without impairing the fit at other frequencies.

The network recommends itself by its simplicity. It is to be noted that it contains only one resistor; thus one meter placed in series with it will suffice to read the power absorbed by the network.

As mentioned above, the series-resonance points of dipole and dummy do not coincide. This defect could be removed by placing a small reactance in series with the network, without giving up the advantage that the power can be read with one meter.

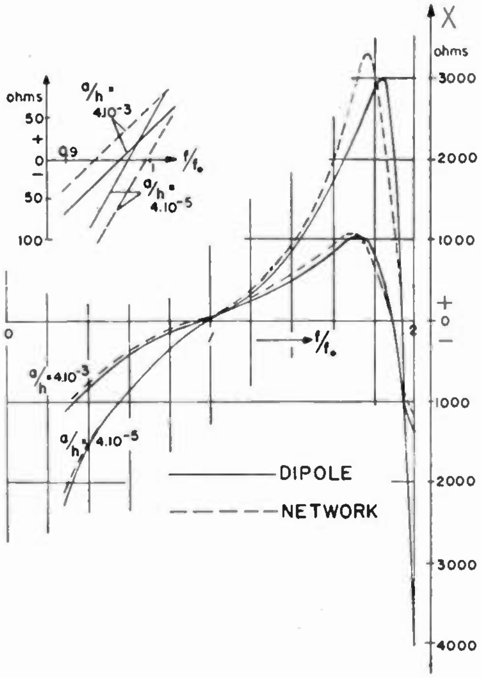


Fig. 2—Reactance of dipole and dummy.

Corrections

The R values shown in Fig. 1 are rather high, and therefore the capacitance of R might cause trouble at frequencies above a few megacycles. Fortunately, R is shunted by the much smaller X_L. If the combined capacitance of R and L is called K, its reactance (at f_0) X_K (we take X_K as positive, as we did with X_{C1} and X_{C2}), then its effect can be taken into account by correcting X_L to $X_L' = X_L/(1+X_L/X_K)$. This will hold as long as $X_K \gg X_L$, and may materially extend the use of our network into the short-wave region.

Also, the loss of L has not yet been considered. Since the Q of the combination R, L $(Q = R/X_L)$ is always less than 15, no trouble is anticipated if a coil of reasonably high Q_L is used; we only have to correct R to R', where $R' = Q_L R/(Q_L - R/X_L')$, subject to the condition $Q_L \gg R/X_L$. If we then want to determine the power P, from the current I flowing in R', we have to put $P = R_{eff} I^2$, where R_{eff} is in first approximation given by $R_{eff} = R(1+(2R/Q_L X_L')) = R'(1+(R'/Q_L X_L'))$. R_{eff} is different from R' because part of the power is dissipated in the X_L branch, and it differs also from R because the current in X_L is not the same as in R'.

It should, however, be understood that these corrections have been computed only for the frequency f_0. The presence of K and the coil loss will also affect the impedance curves of Fig. 2 and 3 to a certain extent. But

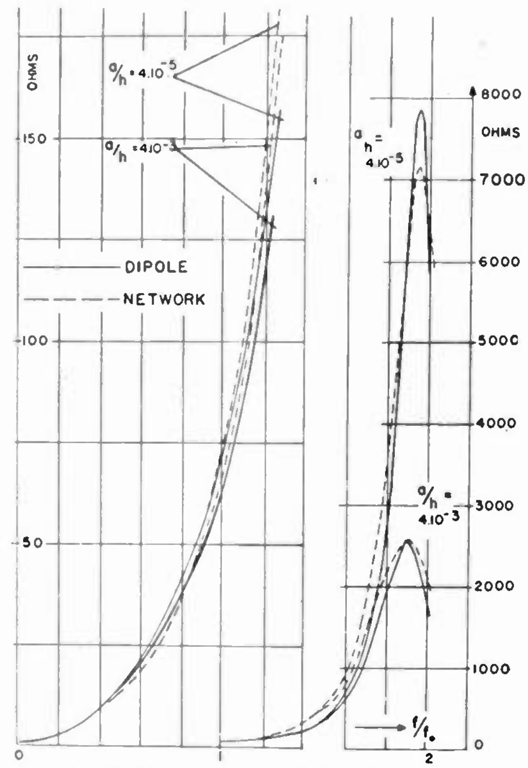


Fig. 3—Resistance of dipole and dummy.

this cannot be worked out quantitatively unless definite assumptions are made on the dependence of the coil losses on frequency.

Example

Take $f_0 = 10$ megacycles, $a/h = 10^{-3}$, which means a conductor diameter of 1.5 centimeters. From Fig. 1, we find $R = 6700$, $X_L = 650$, $X_{C1} = 1800$, $X_{C2} = 607$ ohms, or $L = 10.35$ microhenries, $C_1 = 88.6$, $C_2 = 26.3$ micromicrofarads. Assume $Q_L = 150$, and a coil capacitance of 1 micromicrofarad. Then we should have $X_L/X_K = 0.0408$, and therefore $X_L' = 623.5$ ohms or $L = 9.94$ microhenries. Similarly, R has to be replaced by $R' = 7180$ ohms, while $R_{eff} = 7650$ ohms.

THE INSTITUTE OF RADIO ENGINEERS
INCORPORATED

SECTION MEETINGS

ATLANTA February 18	CHICAGO February 18	CLEVELAND February 24	DETROIT February 18	LOS ANGELES February 21
NEW YORK March 1	PHILADELPHIA March 2	PITTSBURGH March 13	PORTLAND March 13	WASHINGTON March 13

SECTIONS

ATLANTA—Chairman, C. F. Daugherty; Secretary, Ivan Miles, 554—14 St., N. W., Atlanta, Ga.

BALTIMORE—Chairman, G. J. Gross; Secretary, A. D. Williams, Bendix Radio Corp., E. Joppa Rd., Towson, Md.

BOSTON—Chairman, R. F. Field; Secretary, Corwin Crosby, 16 Chauncy St., Cambridge, Mass.

BUENOS AIRES—Chairman, G. J. Andrews; Secretary, W. Klappenbach, *La Nacion*, Florida 347, Buenos Aires, Argentina.

BUFFALO-NIAGARA—Chairman, Leroy Fiedler; Secretary, H. G. Korts, 432 Potomac Ave., Buffalo, N. Y.

CHICAGO—Chairman, A. B. Bronwell; Secretary, W. O. Swinyard, Hazeltine Electronics Corp., 325 W. Huron St., Chicago, Ill.

CINCINNATI—Chairman, Howard Lepple; Secretary, J. L. Hollis, 6511 Betts Ave., North College Hill, Cincinnati, Ohio.

CLEVELAND—Chairman, A. S. Nace; Secretary, Lester L. Stoffel, 1095 Kenneth Dr., Lakewood, Ohio

CONNECTICUT VALLEY—Chairman, W. M. Smith; Secretary, R. F. Shea, General Electric Co., Bridgeport, Conn.

DALLAS-FORT WORTH—Chairman, D. J. Tucker; Secretary, P. C. Barnes, WFAA-WBAP, Grapevine, Texas.

DETROIT—Chairman, R. A. Powers; Secretary, R. R. Barnes, 1411 Harvard Ave., Berkley, Mich.

EMPORIUM—Chairman, H. D. Johnson; Secretary, A. Dolnick, Sylvania Electric Products, Inc., Emporium, Pa.

INDIANAPOLIS—Chairman, A. N. Curtiss; Secretary, E. E. Alden, WIRE, Indianapolis, Ind.

KANSAS CITY—Chairman, A. P. Stuhrman; Secretary, R. N. White, 4800 Jefferson St., Kansas City, Mo.

LOS ANGELES—Chairman, L. W. Howard; Secretary, Frederick Ireland, 1000 N. Seward St., Hollywood, 38, Calif.

MONTREAL—Chairman, L. T. Bird; Secretary, J. C. R. Punchard, Northern Electric Co., 1261 Shearer St. Montreal, Que., Canada.

NEW YORK—Chairman, Lloyd Espenschied; Secretary, J. E. Shepherd, 111 Courtenay Rd., Hempstead, L. I., N. Y.

PHILADELPHIA—Chairman, W. P. West; Secretary, S. Gubin, RCA Victor Division, Radio Corporation of America Bldg. 8–10, Camden, N. J.

PITTSBURGH—Chairman, B. R. Teare; Secretary, R. K. Crooks, Box, 2038, Pittsburgh, 30, Pa.

PORTLAND—Chairman, B. R. Paul; Secretary, W. A. Cutting, c/o U. S. Civil Aeronautics, Box 1807, Portland, Ore.

ROCHESTER—Chairman, O. L. Angevine, Jr.; Secretary, G. R. Town, Stromberg-Carlson Co., Rochester, N. Y.

ST. LOUIS—Chairman, N. J. Zehr; Vice Chairman, C. F. Meyer, KFUO, 801 DeMun Ave., St. Louis, Mo.

SAN FRANCISCO—Chairman, Karl Spangenberg; Secretary, David Packard, Hewlett-Packard Co., Palto Alto, Calif.

SEATTLE—Chairman, L. B. Cochran; Secretary, H. E. Renfro, 4311 Thackeray Pl., Seattle, Wash.

TORONTO—Chairman, T. S. Farley; Secretary, J. T. Pfeiffer, Erie Resistor of Canada, Ltd., Terminal Warehouse Bldg., Toronto, Ont., Canada.

TWIN CITIES—Chairman, E. S. Heiser; Secretary, B. R. Hilker, KSTP, St. Paul Hotel, St. Paul, Minn.

WASHINGTON—Chairman, J. D. Wallace; Secretary, F. W. Albertson, c/o Dow and Lohnes, E Street between 13th and 14th Sts., Washington, D. C.

Institute News and Radio Notes

Board of Directors

At the regular meeting of the Board of Directors, which took place on December 1, 1943, the following were present: L. P. Wheeler, president; H. M. Turner, president-elect; F. S. Barton, vice-president; R. A. Hackbusch, vice-president-elect; S. L. Bailey, E. F. Carter, I. S. Coggeshall, W. L. Everitt, Alfred N. Goldsmith, editor; R. A. Heising, treasurer; F. B. Llewellyn, Haraden Pratt, secretary; B. J. Thompson, H. A. Wheeler, and W. B. Cowilich, assistant secretary.

The actions of the Executive Committee, taken at its meeting on November 2, 1943, were ratified.

The subject of submitting quarterly financial statements to the members of the Board of Directors was discussed.

These applications for membership were approved: for transfer to Senior Member grade, G. C. Bruck, A. F. Pomeroy, and F. H. Speir; for admission to Senior Member grade, J. E. Clegg; for transfer to Member grade, H. S. Benowitz, A. O. Bliss, J. G. Ruckelshaus, and R. D. Stewart; for admission to Member grade, A. C. Davis, M. D. Fagen, H. A. Finke, Edward Iannelli, C. J. Mullin, Ricardo Muniz, Ivar Nelson, A. N. Stanton, and W. E. Whiting; Associate grade, 127; and, Student grade, 103.

The following amended and additional Bylaws were adopted:

Section 45: Add "Investment Committee" to the list of Committees, and add "except as provided in these Bylaws" at the end of the last sentence in the Section.

New Section: "The Investment Committee shall consist of the Treasurer as Chairman, the President, the Secretary, with such additional members as the Board may wish to appoint.

"The Investment Committee is authorized to carry out recommendations received from an Investment Advisor, or Advisors, duly named by the Board to sell or buy securities in the name of the Institute within the limits of the Institute investments, plus any balance of moneys in an Investment Account established by the Board. Such selling or buying of securities shall be by means of an order upon the Custodian of Institute Securities, signed by two members of the Investment Committee authorized to do so by the Board of Directors, directing the Custodian to sell named securities and deposit the proceeds in the Investment Account, or ordering the Custodian to buy named securities and charge the cost to the Investment Account, such account to be maintained in the Custodian's institution. Interest and dividends received from securities shall be deposited in the Investment Account. The moneys in the Investment Account is to be subject to withdrawal by the Board through its regularly constituted officers. The identity of special funds shall be maintained when changes in securities are made. A report on sales and purchases of securities shall be made to the Board by the Investment Committee at the Board's first meeting after each transaction is consummated."

Following the report of Treasurer Heising, as Chairman of the Investment Committee responsible for making amendments to the Institute Certificate of Incorporation, the resolution below was adopted:

"Resolved: That, the notice for a special meeting of the Institute on January 28, 1944, at 11:00 A.M., at the Hotel Commodore, New York City, to amend the charter, be adopted as approved by Mr. Zeamans and as submitted by the Chairman of the Investment Committee, and that the President be instructed to sign the notice, and that copies of the notice be mailed to all members of the Institute between December 9, 1943 and January 18, 1944."

Following the report by Mr. H. A. Wheeler, Chairman of the Institute Committee on Professional Representation, the following resolution was passed:

"Resolved: That, the Board does not favor the formation by the Institute, or otherwise encouraging the formation, of collective bargaining agencies for radio engineers and, therefore, the Board requests the Institute Committee on Professional Representation to report on the practicability of national action, preferably in co-operation with other technical societies, to protect and enhance the professional interests of engineers."

The report on Institute investments, given by Treasurer Heising as chairman of the Investment Committee, was discussed and resulted in a revised definition of investment policy.

Treasurer Heising, as chairman of the Office-Quarters Committee, stated that on the basis of estimates obtained it was found that the rental for a 4000-square-foot area would approximate the cost of operating a small building with considerably larger space. Following a discussion of the situation, approval was given to taking immediate steps toward locating a suitable building or buildings for purchase.

A report was given by Mr. H. A. Wheeler on the progress of the plans for the Winter Technical Meeting, to be held on January 28 and 29, 1944.

The subject of a budget for 1944 was discussed at length.

The report of the Awards Committee, presented by Mr. H. A. Wheeler, chairman of the committee, was received and the following were named for the indicated awards: Medal of Honor for 1944: Haraden Pratt; Morris Liebmann Memorial Prize for 1943: W. L. Barrow; Fellowships: S. L. Bailey, C. R. Burrows, M. G. Crosby, Harry Diamond, C. B. Feldman, Keith Henney, D. O. North, K. A. Norton, S. W. Seeley, D. B. Sinclair, and Leo Young.

President-elect Turner was appointed to serve as Institute representative on the United States National Committee of the International Electrotechnical Commission. Dr. J. H. Dellinger was reappointed Institute representative to serve with the American Documentation Institute for the three-year term, 1944–1946. The appointments of Captain L. D. Prehn and Mr. W. L. Roe to the Facsimile Committee were approved.

Secretary Pratt, as Institute representative on the Radio Technical Planning Board, reported on recent activities of the named agency. It was announced that Editor Goldsmith was recently elected vice chairman of the RTPB.

The resignation of Mr. B. J. Thompson, Institute's alternate on RTPB, was reported by President Wheeler and accepted with regret and an expression of appreciation of services rendered to RTPB. Dr. Barrow was appointed to the indicated office.

A proposed amendment of the "RTPB Organization and Procedure" was approved.

President Wheeler called attention to the registered attendance of 524 at the Rochester Fall Meeting, held on November 8 and 9, 1943, included in the report to the Institute by Mr. L. C. F. Horle.

Secretary Pratt gave a report on the recent developments concerning the Consultative Committee on Engineering, War Manpower Commission, on which committee he is the Institute representative.

An Institute Committee on Education, proposed by Dr. Everitt, was given consideration.

Executive Committee

The Executive Committee meeting, held on November 30, 1943, was attended by L. P. Wheeler, president; H. M. Turner, president-elect; Alfred N. Goldsmith, editor; R. A. Heising, treasurer; F. B. Llewellyn, Haraden Pratt, secretary; H. A. Wheeler, and W. B. Cowilich, assistant secretary.

The following applications for membership were approved for confirming action by the Board of Directors: transfer to Senior Member grade, G. C. Bruck, A. F. Pomeroy, and F. H. Speir; admission to Senior Member grade, J. E. Clegg; transfer to Member grade, H. S. Benowitz, A. O. Bliss, J. G. Ruckelshaus, and R. D. Stewart; admission to Member grade, A. C. Davis, M. D. Fagen, H. A. Finke, Edward Iannelli, C. J. Mullin, Ricardo Muniz, Ivar Nelson, A. N. Stanton, and W. E. Whiting; Associate grade, 127; and, Student grade, 103.

Assistant Secretary Cowilich reported on several office matters, including steps being taken to obtain additional personnel and the extent of overtime work during the month of November.

Consideration was given to a new job classification and to a pension plan for office employees.

The statement of accounts for 1943, giving the comparison of disbursements and receipts for 1943 with the budget for the same period, was presented by Secretary Pratt and discussed at length. Arrangements were made to have the budget for 1944 made available at the next meeting.

Treasurer Heising, as chairman of the Office-Quarters Committee, presented an analysis of costs of leasing and owning larger office quarters, which are urgently needed, and a list of quarters available on a rental basis and of small buildings for sale. It was indicated that the cost to the Institute of renting an area of 4000 square feet, the minimum required, would approximate the expense of operating a small building with considerably more space to allow for expansion. Approval was granted to a recommendation to the Board of Directors relative to the location and possible purchase of a suitable building in New York.

A discussion was held on the report concerning the Institute investment policy, presented by Treasurer Heising, chairman of the Investment Committee.

The proposed revision of Bylaws Section 45 and a new section of the Bylaws, relating to organization and duties of the Investment Committee, were reviewed by Treasurer Heising, Chairman of the Constitution and Laws Committee.

Attention was called by Treasurer Heising to the Special Meeting, to be held at 11:00 A.M. on January 28, 1944 at the Hotel Commodore, New York, for the purpose of amending the Institute's Certificate of Incorporation, and to the draft of the meeting notice being submitted to the Board of Directors for approval.

Editor Goldsmith stated that the Temporary Facsimile Test Standards and another special publication, were in the process of printing and would be distributed to the membership at an early date.

Editor Goldsmith reported that the appeal to the War Production Board for 1944 paper supply for the PROCEEDINGS was being prepared and would be submitted promptly.

Consideration was given to correspondence relating to the list of Institute's Charter Members.

Decisions were made on certain matters concerning the arrangement for handling PROCEEDINGS advertising, reported by Mr. H. A. Wheeler.

The program and the budget for the 1944 Winter Technical Meeting, scheduled for January 28 and 29, 1944 at the Hotel Commodore, New York, were presented by Mr. H. A. Wheeler and given approval.

An appropriate light blue was selected as the official emblem color for the new Member grade.

Recommendations were made to the Board of Directors, concerning the Institute's representation on the United States National Committee of the International Electrotechnical Commission and the American Documentation Institute.

President Wheeler introduced the topic of holding regional meetings of the Institute, and it was decided to include the named activities on the agenda of the annual Sections Committee meeting, to take place on January 27, 1944.

Treasurer Heising, as chairman of the Sections Committee, reported on the forthcoming meeting of the particular committee and indicated topics that would be discussed.

Consideration was given to a letter from the New York Section, concerning the rate of rebates to Sections.

Correspondence

Postwar Civilian-Aircraft-Radio Field

It is the purpose of this communication to call to the Institute's attention certain aspects of the postwar civilian-aircraft-radio field. The favorable development of this field, it is felt, may be materially influenced by the focusing upon it of the attention of the radio engineering fraternity. Reference is made to what appears to be the practically certain tremendous increase in civilian aircraft in use in the United States to be anticipated beginning immediately after V day. The vital importance of radio communication to safety of life in the air is believed to be apparent.

That such an expansion in civilian aircraft may be taken as certain to occur seems to be beyond cavil. There are many reasons why this must be true. Neglecting entirely the popularity of aircraft and flying, the significant part air power is playing in the war, the complete catching of popular fancy, the interest of demobilized military flyers, the strong appeal to American youth, there seem to the writer to be two paramount reasons for anticipating such expansion—and a concomitant opening up on significant scale of a major new field for radio equipment engineering, production, and sale.

The first, and possibly most immediately compelling reason, is that aircraft production is today America's No. 1 industry. This is as it should be during the war. But, with probable rapidly dwindling demand for military aircraft after cessation of hostilities a few years hence, what does America's No. 1 industry then do to utilize its productive plant fully? What does it do to contribute to national welfare through continuing its employment of the large number of employable Americans today engaged in aircraft production? A glance at aircraft figures as of Pearl Harbor, December 7, 1941, is suggestive. Inquiry has revealed that, as of December 31, 1941, there were but approximately 486 commercial air transports and 26,500 civilian, itinerant, or other nonmilitary aircraft in the United States. Believed to be correct, these figures provide food for thought. They suggest, what with government "borrowing" of significant numbers of commercial aircraft to aid the war effort, that the first postwar demand upon the aircraft industry will be for commercial transport and cargo planes.

Offsetting the possible production which this initial demand might create are approximately 15,000 transport or like aircraft now utilized by the government. Should these be released to commercial users in an orderly manner, their number would go far to satisfy initial demand. Released in a disorderly manner, their number could be demoralizing to the industry, could create an "Army-Navy surplus store" situation of distasteful and damaging character. Conversion of these aircraft to a form suitable for commercial use could create a new and temporary business in competition with the original producers, who would seem better advised to devote some of their productive capacity initially to such conversion.

Even without the threat represented by 15,000 government-held transport craft, such demand alone cannot preserve current activity rate in the aircraft industry after V Day. The initial, and continuing, commercial demand may be heavy, but in character it is unsuited to support plants specializing in smaller types of aircraft which may find it not feasible to convert rapidly to production of large transport or cargo aircraft. In any event it is reasonably certain that the demand, domestic and export, for large aircraft cannot keep the entire industry operating indefinitely at anywhere near its present rate in terms of the number of aircraft produced and number of workers employed. Survival alone of the many plants producing smaller aircraft will force intensive cultivation of the civilian market. Popular fancy is ready for such cultivation in terms of conventional aircraft and seems to await anxiously the practical helicopter; and industry need can find long-time relief only in civilian volume-sale possibilities.

It is not proposed to guess or estimate what the total number of aircraft in use at any future date may be. It is believed, however, to have been pointed out that with aircraft production capacity what it is today the continuance of prosperity in, and prosperity from, the aircraft industry makes mandatory the intensive development of the civilian aircraft market by the industry itself.

Should, for unfavorable reasons of cost and convenience which may be envisaged, the private-plane market be unable over time to support the expanded industry, the second reason comes to its aid. This reason is a combination of probably much heavier annual military-aircraft procurement than before the war, and the opportunity inherent in developing government-sponsored "flying clubs." The romance of flying is so great that American youth will dream of being able to fly for years to come—a dream in most cases not to be satisfied for economic reasons. The government will require trained pilots in anticipation of some future conflict, the pilot material is ready and waiting for training. The capacity to produce small trainer planes may, at some future date, even be begging for business. What more obvious than the formation of "flying clubs" all over the country by the government? In purchasing the required planes it supports an industry which it must preserve against future need, preliminarily trains at nominal cost to them the pilots it must also have in reserve, and simultaneously provides an outlet for the air enthusiasm of its youth.

The allocation of radio frequencies to amateur investigation and employment has twice stood Uncle Sam in good stead, when as a result the government has found ready and waiting a substantial body of technically skilled radio amateurs in its hours of need—1917–1918 and 1941–?. Are not the multiple benefits of government "flying clubs" potentially as—or maybe even more—valuable?

Consideration of the foregoing will indicate conclusively the inseparability of aircraft and radio in all the aspects of the former after V Day. With a steadily increasing number of aircraft going into service, communication, which may satisfactorily be only by radio to an aircraft in flight, becomes even more essential to safety of life in the air—yes, on the ground too—than is the motor of an aircraft. A glider out of control-contact can indeed be a lethal weapon in a crowded air.

The following proposition is, therefore, respectfully propounded; viz., in civilian aircraft, radio communication equipment is the first and primary requisite. This may appear to be a bold statement, but a little contemplation of confusion possible at a major airport on a Sunday afternoon with but *one* aircraft bereft of contact with the control tower makes it almost inescapable.

In the supposition that it has been demonstrated that civilian aircraft expansion after V Day will be on a major scale, and that the need of radio communication equipment in operative condition is the essential prerequisite to civilian flight, it may be concluded that the radio industry is in the position of being given a new and sizable outlet for its skill and products, literally upon a "golden platter."

What the industry, and the government, does with this opportunity is the writer's acute concern as a radio engineer, for in this expanded field life itself is the price of inadequate radio equipment. Here the primary justification for radio communication is neither pleasure nor convenience, but the preservation of life. Quality may therefore never be allowed to fall, engineering or productionwise, into the classification of broadcast receivers. Yet aircraft radio equipment is logically to be produced by the makers of broadcast receivers, in terms of the probable considerable output which will be required. The steadily diminishing selling price of broadcast receivers throughout the past fifteen years may be regarded as a sad foretaste of what unregulated aircraft radio equipment production might bring.

Fortunately there is something of a limit upon the sales expansion possible for aircraft radio receivers, transmitters, direction finders, positioning equipment, and the like. The reward for reduced selling price brings, as price reduction seems almost invariably to do, an inacceptable deterioration of quality and reduced sales. Two limits exist: one, that few more aircraft radio equipments can be profitably sold than there are aircraft made to use them; second, that the equipment must be of such quality that it is in operative condition when, properly inevitably, government inspection of aircraft and accessories takes place at frequent periodic intervals.

But, lest the shortsighted radio manufacturer—and who will gainsay that there have been a few—even be tempted to seek a larger percentage of total possible sales through the inducement of reduced price with its almost certain derogation of quality, is it not better to erect at least a partial barrier to such deleterious effort?

Engineering refinement, increased production, skill, diminished distribution costs, these would seem to be the only proper devices to be employed to reduce selling price. If these alone are given free reign, will not the essential of preservation of human life in the air be served better? Will not the so enforced maintenance of fair and reasonable quality standards at one and the same time operate to maintain the air-borne radio-equipment business upon a more profitable basis than would cutthroat price competition?

The writer, for himself, answers these questions in the affirmative. Having done so, he has considered ways and means to their practical realization, and found the basic instrumentalities requisite therefor to exist. He proposes government regulation now, (rather than later and after abuses have made it inescapable and punitive in possible effect) to insure maintenance in all air-borne radio equipment of a minimum standard of quality and serviceability. This may be accomplished through modification of the present Civil Aeronautics Administration "Approved Type Certificate" plan. Under this plan the manufacturer may submit his product to C.A.A. for test and when and if it is found suitable for air-borne use, an Approved Type Certificate is issued indicating its suitability for use by commercial air lines—and civilians, should they be interested.

If it may be made mandatory by Federal law that no aircraft, powered or glider, may take off without radio equipment in operative condition, and that such equipment must have previously measured up to rational minimum standards of technical excellence, then would not a significant service have been rendered to the American people and the radio industry? Should the radio industry itself, through The Institute of Radio Engineers, aid in effecting the enactment of such a Federal law or regulation, would it not have rendered a meritorious service to the country?

A seemingly logical corollary thought would be the establishment of an aircraft radio advisory committee, drawing upon Institute membership for engineering and production skill, to collaborate with the responsible government agency in the formulation and interpretation of technical standards best calculated to serve the flying public, and through it, the radio industry itself.

McMurdo Silver,
Vice President, The Grenby
Manufacturing Company
Plainville, Connecticut

Books

Basic Radio Principles, by Maurice Grayle Suffern.

Published by McGraw-Hill Book Company (1943), 330 West 42nd Street, New York 18, N. Y. 256 pages+15-page index+x pages. 269 figures. $5\frac{3}{4}\times8\frac{1}{4}$ inches. Price, $3.00.

This book is a direct result of the current need for a large number of workers trained in practical radio maintenance. Avoiding mathematics entirely and making no assumptions of background beyond the simplest principles of electricity, it presents an adequate development of radio principles for the training of technicians. The stated purpose is to give the student a knowledge of radio fundamentals, the ability to identify circuit components with their symbols, and to interpret diagrams, and an understanding of the principles of operation of radio equipment. This purpose should be achieved for the intended type of student.

The development is conventional. Beginning with a discussion of basic phenomena, terms, and symbols, the author proceeds to outline the various functions performed in radio receivers and transmitters and the circuits used, ending with a brief description of antennas and test equipment. Each chapter is necessarily much limited as to detail, with all nonessential topics omitted, but more could not be attempted in a brief and elementary text. While the initial chapters may seem repetitious to anyone with previous knowledge, once the basis of fundamentals has been established the development is clear and concise. Occasionally, however, the desire for simplification leads to such unjustifiable statements as "A load resistor is an electrical device into which electrical energy is fed in order to produce electrical power" or "The human ear . . . cannot vibrate at a radio-frequency rate. That is why we have no awareness of the radio waves all around us." A lack of specificity in the use of some terms is another doubtful concession to the reader's limited background.

The omission of frequency modulation and ultra-high-frequency material is in line with the elementary character of the book; these topics appear to be the only ones lacking for up-to-date coverage. Sketches and circuit diagrams are numerous and generally informative, and each chapter ends with a set of simple multiple-choice review questions.

O. L. Updike
University of Illinois
Urbana, Illinois

Contributors

Edward J. Content

Edward J. Content (M'39-SM'43) served with the United States Signal Corps during World War I and was graduated from the radio course at the Signal Corps School at Fort Monmouth in 1921. He entered radio broadcasting in 1922 and for two years was with WEAF. He inaugurated and was chief engineer of WGBS until 1926 when he joined the staff of WOR where he has been assistant chief engineer of this station since that time. During his service in broadcasting, he has specialized in studio acoustical treatment, audio switching systems, and operational engineering management.

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John J. Farrell (A'43) was born at Watervliet, New York, in 1897. He entered the student training course of the General Electric Company at Schenectady, New York, upon graduation from high school in 1913. The years 1913 to 1918 were spent in the General Electric drafting department. From 1918 to 1920, Mr. Farrell engaged in tool-design work with the United States Arsenal, Department of Ordnance, at Watervliet, New York; also with the Studebaker Corporation at South Bend, Indiana; and with the Willys-Overland Company at Toledo, Ohio.

In 1920, he rejoined the General Electric Company as a designing draftsman in the radio division. He was promoted to drafting supervisor of the transmitter division in 1921 and entered the mechanical engineering section in 1926. He became head of the section in 1934 and was appointed designing engineer of the transmitter division in 1938.

As installation engineer, Mr. Farrell was in charge of the placing in satisfactory service of many high-power broadcast stations. Among these were the 50-kilowatt transmitters at KOA, Denver; KPO, San Francisco; and KFI, Los Angeles; and the original 50-kilowatt transmitter at WEAF. Mr. Farrell was the General Electric project engineer on the installation of the 500-kilowatt amplifier at WLW, Cincinnati.

John J. Farrell

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Lonsdale Green, Jr.

Lonsdale Green, Jr., was born at Anniston, Alabama, on December 31, 1889. He received the B.S. degree in mechanical engineering from the University of Illinois in 1912. From 1915 to 1930 he was assistant manager of the general acoustical department of the Johns-Manville Corporation. Mr. Green was in the United States Army Air Service during 1917–1918. Since 1930 he has been president of the Acoustical Construction Corporation in New York City. He is a Fellow and Charter Member of the Acoustical Society of America and since 1941 has been its treasurer.

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Hartley C. Humphrey

Hartley C. Humphrey (A '29) is a native of Tiverton, Rhode Island. He received the B.S. degree in electrical engineering from Worcester Polytechnic Institute in 1917. He was in the traffic department of the Western Union Telegraph Company, 1917 to 1924; radio broadcasting department, American Telephone and Telegraph Company, 1924 to 1925; Bell Telephone Laboratories, 1925 to 1926; assistant chief engineer Warner Brothers-Vitaphone Corporation, 1926 to 1927; recording engineering manager and assistant director of engineering, Electrical Research Products, Inc., 1927 to 1936; and technical director, Western Electric Company, Ltd., London, England, 1936 to 1937. Since 1939 he has been in the research products department of the Westinghouse Electric and Manufacturing Company.

Mr. Humphrey is an Associate of the American Institute of Electrical Engineers. Subsequent to his work with multiplex printing telegraph systems in the traffic department of Western Union and while with the Bell System he pioneered in the engineering development of sound on disk and film recording in Hollywood, New York, and later London, England. Most recently he has been closely identified with the planning and layout of high-power high-frequency industrial electronic equipment.

❖

H. W. Jamieson

H. F. Olson

H. W. Jamieson (A'42) was born in Shreveport, Louisiana, on January 19, 1918. He was graduated from the University of California with a B.S. degree in electrical engineering in 1939. In January, 1940, he was employed by the General Electric Company where he was with the instrument transformer development section until July, 1941; high-frequency development section of the General Engineering Laboratory until August, 1942; and since with the electronics laboratory. He is now in the third year of the advanced course in engineering of the General Electric Company.

H. F. Olson (A '37) was born at Mt. Pleasant, Iowa, on December 28, 1902. He received the B.S. degree in 1924, the M.S. degree in 1925, the Ph.D. degree in 1928, and the E.E. degree in 1932 from the University of Iowa. Dr. Olson was a research assistant at the University of Iowa from 1925 to 1928. From 1928 to 1930 he was in the research department of the Radio Corporation of America; from 1930 to 1932, in the engineering department of RCA Photophone; from 1932 to 1934, in the research department of the RCA Manufacturing Company; from 1934 to 1942, in charge of acoustic research, RCA Manufacturing Company, and since 1942, in charge of acoustic research at the RCA Laboratories, Princeton, New Jersey. He is a member of Tau Beta Pi, Sigma Xi, the American Physical Society and a Fellow of the Acoustical Society of America.

Leonard T. Pockman

Leonard T. Pockman was born on July 26, 1911, in San Francisco, California. He received the A.B. degree in physics from Stanford University in 1933 and the Ph.D. degree in physics in 1938. In 1934 he was Charles Coffin, General Electric, Fellow at Stanford, and teaching assistant from 1935 to 1937. During 1937–1938 he was an instructor in physics at the Massachusetts Institute of Technology, and from 1938 to 1940, instructor in physics at Cornell University. He served as assistant physicist in the Aircraft Radio Laboratory at Wright Field during 1941. Since the start of 1942, he has been a physicist and engineer at Heintz and Kaufman Ltd. He is a member of the American Physical Society, Phi Beta Kappa, and Sigma Xi.

Hans Salinger

Hans Salinger (A'37) was born in Berlin, Germany, on April 1, 1891. He received the Ph.D. degree from the University of Berlin in 1915. From 1919 to 1929 he was a research associate at the Reichpostzentralamt in Berlin and from 1929 to 1935, professor at the Polytechnical Institute and the Heinrich Hertz Institut für Schwingungsforschung in Berlin. From 1936 to 1942 and again from July, 1943, to date Dr. Salinger has been with the Farnsworth Television and Radio Corporation. During 1942–1943 he was a consulting physicist and instructor at Indiana Technical College in Fort Wayne. He is a member of the American Association for the Advancement of Science and an Alumni Member, University of Pennsylvania Chapter of Sigma Xi.

S. A. Schelkunoff

S. A. Schelkunoff (A'40) received the B.A. and M.A. degrees in mathematics from the State College of Washington in 1923, and the Ph.D. degree in mathematics from Columbia University in 1928. He was in the engineering department of the Western Electric Company from 1923 to 1925; the Bell Telephone Laboratories from 1925 to 1926; the department of mathematics of the State College of Washington, 1926 to 1929; and Bell Telephone Laboratories, 1929 to date. Dr. Shelkunoff has been engaged in mathematical research, especially in the field of electromagnetic theory.

J. R. Whinnery (A'41) was born on July 26, 1916, in Read, Colorado. He received the B.S. degree in electrical engineering from the University of California in 1937 and since that time has been with the General Electric Company in Schenectady, New York. In 1940 he was a graduate of the three-year program of the advanced course in engineering of the General Electric Company, and during the following two years supervised the high-frequency section of that program. Since 1942 he has been with the Electronics Laboratory of the Company.

J. R. Whinnery

SECTION MEETINGS

Buffalo-Niagara

"Electronics from the Engineer's Point of View," by C. J. Madsen, Westinghouse Electric and Manufacturing Company; November 18, 1943.

Chicago

"Designing Magnetic Fields," by J. F. Calvert, Northwestern Technological Institute; November 19, 1943.

"Voice-Controlled Circuits," by R. H. Herrick, Automatic Electric Laboratories; November 19, 1943.

"The Psychology of Invention," by H. F. Fruth, Galvin Manufacturing Corporation; December 17, 1943.

"Frequency Modulation and Its Postwar Future," by J. E. Brown, Zenith Radio Corporation; December 17, 1943.

Cincinnati

"Improving Engineering Department Efficiency," by H. D. Sarkis, Crosley Corporation; November 23, 1943.

"WKDU—Cincinnati's Station," by J. L. Hearn, Bureau of Communications, City of Cincinnati; December 14, 1943.

Cleveland

"The High-Speed Photoelectric Recorder," by H. L. Clark, General Electric Company; December 16, 1943.

Election of Officers; December 16, 1943.

Connecticut Valley

"Electronics in Industry," by W. I. Bendz, Westinghouse Electric and Manufacturing Company; December 7, 1943.

Emporium

"Considerations in Oscilloscope Design," by F. L. Burroughs, Sylvania Electric Products, Inc.; November 30, 1943.

"Forecasts of the Postwar Cathode-Ray Tube Market," by W. L. Krahl, Sylvania Electric Products, Inc.; December 17, 1943.

"Problems in Quality Control and Standardization," by J. R. Steen, Sylvania Electric Products, Inc.; December 17, 1943.

Indianapolis

"Four-Terminal Network Design," by R. P. Siskind, Purdue University; November 26, 1943.

Los Angeles

"Principles of Radio-Frequency Induction and Dielectric Heating," by Fred Albin, RCA Victor Division; November 30, 1943.

"Demonstration of RCA 15-Kilowatt Radio-Frequency Heating Unit," by Michael Rettinger, RCA Victor Division; November 30, 1943.

Motion Picture, "Radio-Frequency Heating Developments in RCA Princeton Laboratories." November 30, 1943.

Philadelphia

"Television Station WPTZ," by F. J. Bingley, Philco Radio and Television Corporation; December 3, 1943.

Rochester

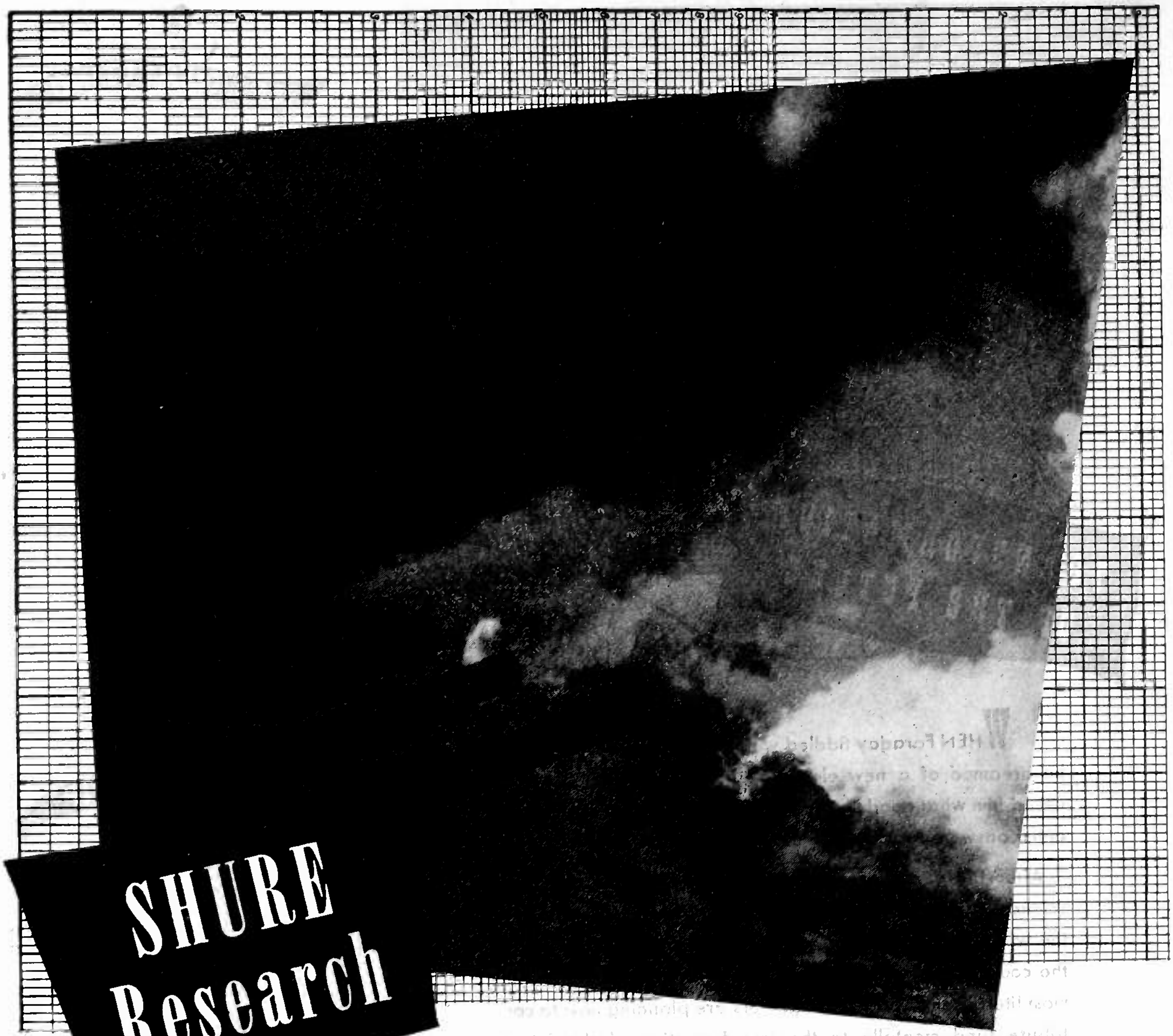
"Supersonics and Its Applications," by H. S. Sack, Cornell University; December 2, 1943.

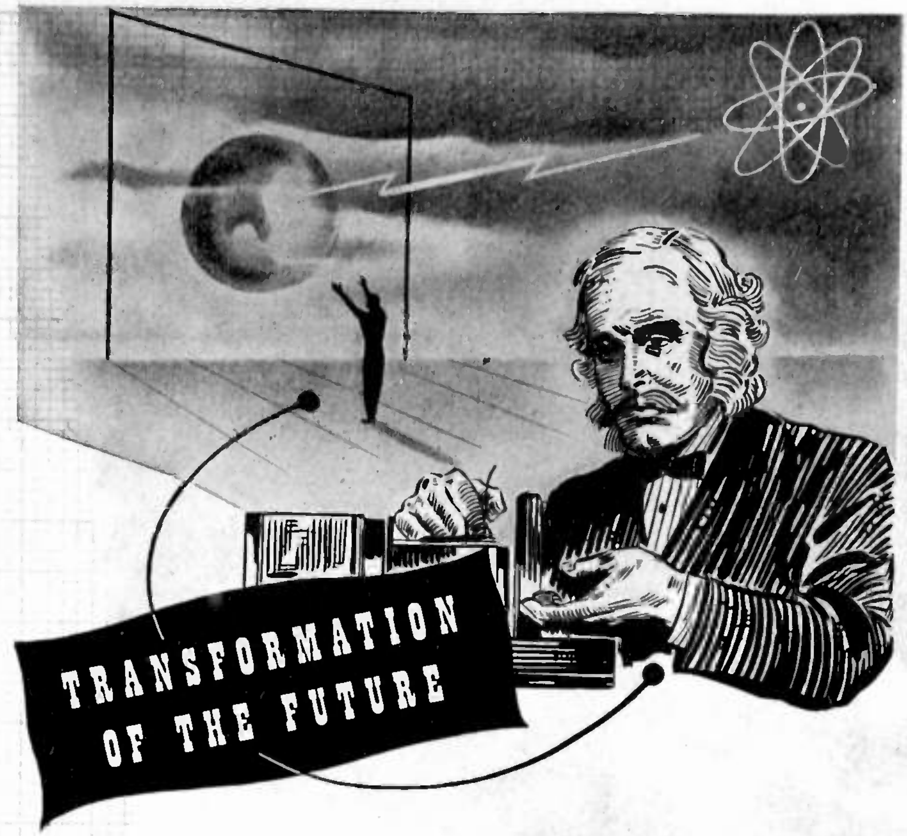
"The Radio Industry Today," by R. H. Manson, Stromberg-Carlson Company; December 7, 1943.

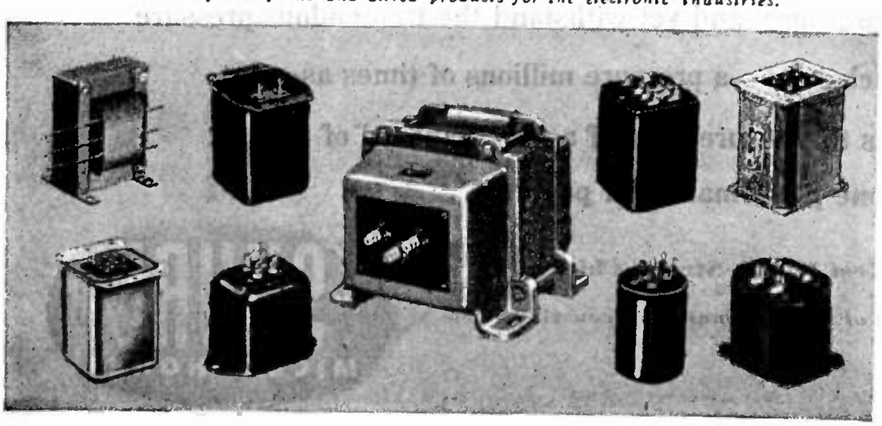
"A Perspective View of Applied Electronics," by M. J. Larsen, Stromberg-Carlson Company; December 14, 1943.

"When December 25 Became Christmas," by C. H. Moehlman, Colgate Rochester Divinity School; December 21, 1943.

(Continued on page 36A)







SECTION MEETINGS

(Continued from page 34A)

SAN FRANCISCO

"Properties of Radio-Frequency Transmission Lines and Stub-Matching Sections," by Lester Reukema, University of California; September 29, 1943.

"Properties of Lines with Inductively Coupled Sections," by Alfred Towne, Associated Broadcasters, Inc.; September 29, 1943.

"Colinear Antennas," by F. R. Brace, Associated Broadcasters, Inc.; October 27, 1943.

"Rhombic Antennas," by G. W. Cattell, Mare Island Navy Yard; October 27, 1943.

ST. LOUIS

"Training of Radiomen at Scott Field," by R. D. Gibson and Henry Spillner, Jr., Department of Communications, Scott Field; November 26, 1943.

TWIN CITIES

"Discussion of the Application of Electronics to Meteorology Through Radio Sonde," by Fred Melius, East High School, Minneapolis; November 17, 1943.

WASHINGTON

"Accelerated Heat Treatment of Metals and Dielectrics by Means of Radio-Frequency Currents," by G. H. Brown, RCA Laboratories; December 13, 1943.

Election of Officers, December 13, 1943.

MEMBERSHIP

The following indicated admissions and transfers of membership have been approved by the Admissions Committee. Objections to any of these should reach the Institute office by not later than February 29, 1944.

Admission to Member

Clarke, F. H., Naval Research Laboratory, Anacostia Station, D. C.
Mauritz, F. E., 6001 Dickens Ave., Chicago, Ill.
Preist, D. H., Naval Research Laboratory, Anacostia Station, D. C.
Sorensen, E. G., 154 Roxbury Rd., Garden City, L. I., N. Y.

Transfer to Member

Chapman, R. W., 2819 Myrtle Ave., N.E., Washington, D. C.
Goldberg, H., 215 Brett Rd., Rochester, N. Y.
O'Shea, J. G., 290 W. State St., Presque Isle, Maine.
Spence, P. W., 463 West St., New York, N. Y.

Transfer to Senior Member

Angevine, O. L., Jr., 284 Brooklawn Dr., Rochester, N. Y.
Dushman, S., General Electric Co., Schenectady, N. Y.
Lohnes, G. M., 4938 Wakefield Rd., Friendship P.O., Washington, D. C.
Shepherd, J. E., 111 Courtenay Rd., Hempstead, L. I., N. Y.
Tholstrup, H. L., 135 Willowbend Rd., Rochester, N. Y.

The following admissions and transfers were approved by the Board of Directors on January 5, 1944.

Transfer to Member

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Ashton, J. O., Hurl Towers Apts., Greenwich, Conn.

(Continued on page 38A)

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(Continued from page 36A)

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Harrison, C. W., Jr., Cruft Laboratory, Harvard University, Cambridge, Mass.
Hidy, J. H., 365 Stewart Ave., A-24, Garden City, L. I., N. Y.

Admission to Senior Member

Attwood, S. S., Rm. 277, West Engineering Building, Ann Arbor, Mich.
Riblet, H. B., 93–18 Lamont Ave., Elmhurst, L. I., N. Y.

Transfer to Senior Member

Atkins, C. E., 1100 Oak Ave., Evanston, Ill.
Brauer, H. H., 76 East Boulevard, Rochester, N. Y.
James, V. N., 2841 Dyer St., University Park, Dallas, Texas.
Martin, H. B., Radiomarine Corporation of America, 75 Varick St., N. Y.
Persons, C. B., 1559—19 St., North, Arlington, Va.
Samuel, A. L., Bell Telephone Laboratories, 463 West St., New York, N. Y.
Sharpe, M. O., 135 N. Park Dr., Arlington, Va.
Town, G. R., 148 Colebourne Rd., Rochester, N. Y.

The following admissions to Associate were approved by the Board of Directors on January 5, 1944.

Abraham, W. G., 129 Second St., Garden City, L. I., N. Y.
Adams, E. W., Jr., 10 Mulberry Ave., Garden City, L. I., N. Y.
Ainlay, A., c/o Inspection Board of U. K. & Canada, Northern Electric Co., Shearer St., Montreal, Que., Canada.
Allen, J. J., 506 N. 63 St., Seattle, 3, Wash.
Allerton, G. L., 47 First St., Yonkers, N. Y.
Almas, S. L., 13200 Strathmoor, Detroit, 27, Mich
Anderson, S. W., 3434 Home St., Fresno, Calif.
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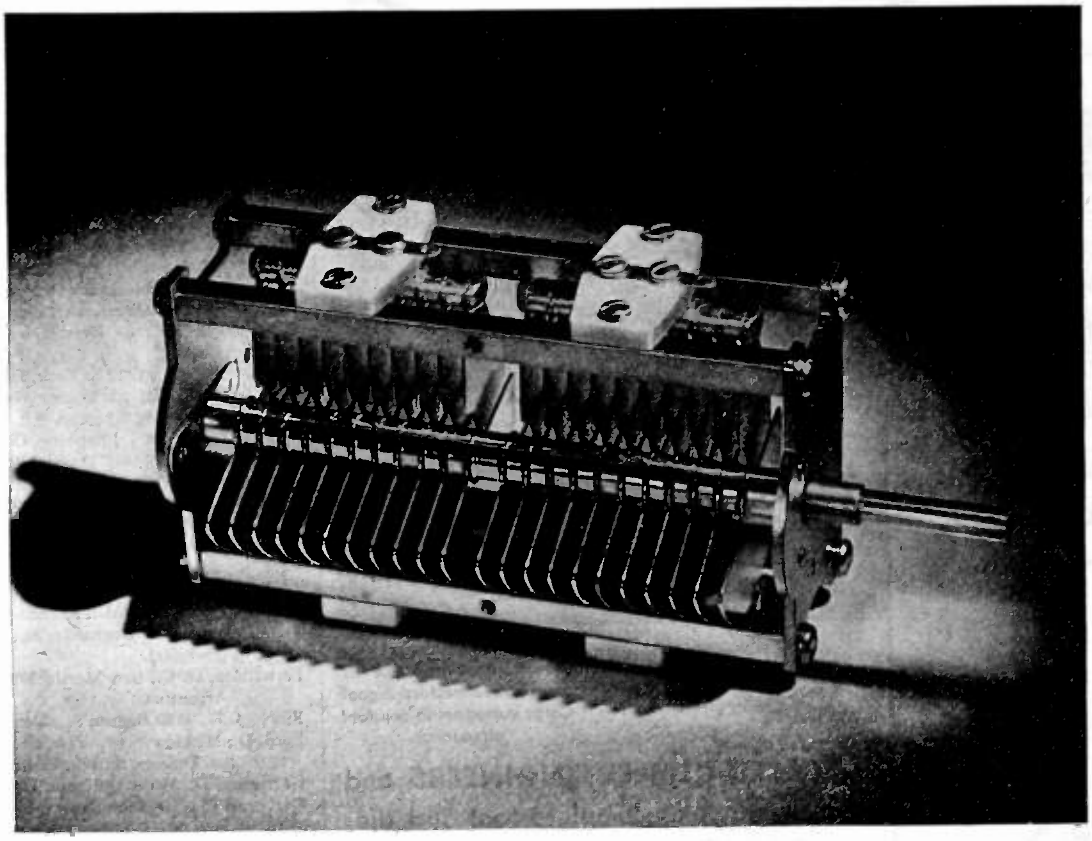
Arnold, J. B., 220 Myrtle Dr., West, Woodlawn Hills, San Antonio, 1, Texas.
Beers, Y., Winthrop Hall, St. John's Rd., Cambridge, Mass.
Bethard, C. T., Civil Aeronautics Administration, 1500 Fourth St., Santa Monica, Calif.
Blackstone, H., 63 Cambridge Ave., Garden City, L. I., N. Y.
Blackman, H. W., 206 Dawson Ave., Boonton, N. J.
Biele, R. J., 432 Gurdon St., Bridgeport, Conn.
Blake, A. F., 1432 Madison St., Denver, 6, Colo.
Bock, A. W., 954 Amelia Ave., Akron, 2, Ohio.
Borders, C. R., Box 722, Sarasota, Fla. (transfer).
Bowman, L. H., 123—15 St., Richmond, Calif.
Braswell, C. P., APO 677, c/o Postmaster, Presque Isle, Maine.
Brock, W. R., 4350 N. Bell Ave., Chicago, Ill.
Broughton, J. B., c/o Fleet Post Office, San Francisco, Calif.
Brueck, J. O., 175 Pinehurst Ave., New York, 33, N. Y.
Brush, L. V., 201 East Platt St., Maquoketa, Iowa (transfer).
Busch, A. E., 5437 Kimbark, Chicago, Ill.
Carlin, B., 439 Marlborough St., Boston, Mass.
Carter, R. W., 4951 Brummel St., Skokie, Ill.
Caspers, G. E., Jr., Box 165, Middletown, N. J.
Cassels, W. H., c/o General Electric Co., Tucuman 117, Buenos Aires, Argentina.
Chasteen, J. W., Jr., 152 N. W. 34 St., Miami, Fla.
Cianela, D. J., 2423 S. 21 St., Philadelphia, Pa.
Clark, J. H., 2005 Laurel Rd., Oakmont, Upper Darby, Pa.
Clark, W. H., Jr., 303 Mountain Way, Rutherford, N. J.
Coleman, G. M., 187 Fayerweather St., Cambridge, Mass.
Collins, J. A., Drummond Court, Apt. 818, 1455 Drummond St., Montreal, Que., Canada.
Collins, E. A., 8747 W. 44 Pl., Lyons, Ill.

(Continued on page 40A)

WHEN PEACE COMES



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Grasser, L. G., VHF Project Engineering Agency, Allenhurst, N. J.
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Harding, R. W., 41 St. Mary's St., Boston, 15, Mass.
Hillier, M., 5 Metford Rd., Redland, Bristol, 6, England.
Hoagland, K. A., 32 Garden Ave., Belleville, 9, N. J.
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Johnson, H. S., 1908 Rosemary St., Denver, 7, Colo.
Johnson, P. B., 806 E. Main St., Belleville, Ill.
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Kelk, G. F., 477 Bayview Ave., Toronto, Ont., Canada.
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Kyland, B. G., 140 Brookdale, San Francisco, Calif.
Langberg, E. I., 5814 N. 16 St., Philadelphia, 41, Pa.

(Continued on page 42A)

A PLANET *Not a Meteor*

Ever notice how a meteor streaks across the heavens in a blaze of fiery splendor? It's a beautiful sight . . . while it lasts. But most meteors burn themselves out long before striking the earth. Not so a planet . . . though much less brilliant, it's there to stay. That's how we like to think of I. C. E. Here to stay . . . Born of the war . . . yes, but acquitting itself well, and all the better to serve you in the post-war future.

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MacDowell, J. E., 7805—19 Dr., Jackson Heights, L. I., N. Y.
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McEwan, T. S., 1046 Dinsmore Rd., Winnetka, Ill.
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Merrill, H. B., Jr., 64 Brookside Rd., Springfield, Delaware Co., Pa.
Millman, A. J., 140 S. Dale Ct., Denver, 9, Colo.
Nienaltowski, W., c/o Northern Electric Co., Special Product Division, Box 369, Montreal Que., Canada.
Oberle, E. J., Arpin Manufacturing Co., Radio Division, 422 Alden St., East Orange, N. J.
OOress, E. C., 25 Park St., Montclair, N. J.
Pverby, H., Box 532, Toronto, Ont., Canada.
kalmer, J. E., 3048 Stuart St., Denver, 12, Colo.
Price, D. R. S., 39 Nelson Rd., Boornemouth, England.
Parr, G., 68 Compton Rd., London, N. 21, England.
Perlman, W., 1630 Gilford Ave., New Hyde Park, L. I., N. Y.
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Pinkerton, P. C., 1542 Gaylord St., Denver, 6, Colo.
Pinter, P. J., 2221 N. Fourth St., Philadelphia, 33, Pa.
Ragsdale, W. H., 4499 Forest Ave., RFD 9, Watkins Lake, Pontiac, Mich.
Provost, P., 3801 Lacombe Ave., Montreal, Que., Canada.
Raleigh, J. H., Box 534, Oak Park, Ill.
Reese, V. N., 16 S. Williams Ave., Dayton, Ohio.
Rehberg, L. H., 43-06—53 St., Woodside, L. I., N. Y.
Reilly, J. A., 2165-C Ala Wai Blvd., Honolulu, 37, T. H.
Rich, A. P., 2419 J Street, San Diego, Calif.
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Rockafellow, S. C., 456 Leta Ave., Flint, Mich.
Roganti, M. S., Paroissien 3906, Buenos Aires, Argentina.
Rogers, R. T., 4110 N. Keystone Ave., Chicago, Ill.
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Rohrer, J. H., 337 Eudora, Denver, 7, Colo.
Rubel, J. H., General Electric Research Laboratory, Schenectady, N. Y.
Schatz, E. R., 624 Clyde St., Pittsburgh, Pa.
Schirk, F. L., 1171 Detroit, St., Denver, 6, Colo.
Schulkin, M., 1620 E St., N. E., Washington, 2, D. C.
Schwartz, H. T., 60 Whalley Ave., New Haven Conn.
Sharpe, W. J., 318 W. Iowa Ave., Denver, 10, Colo.

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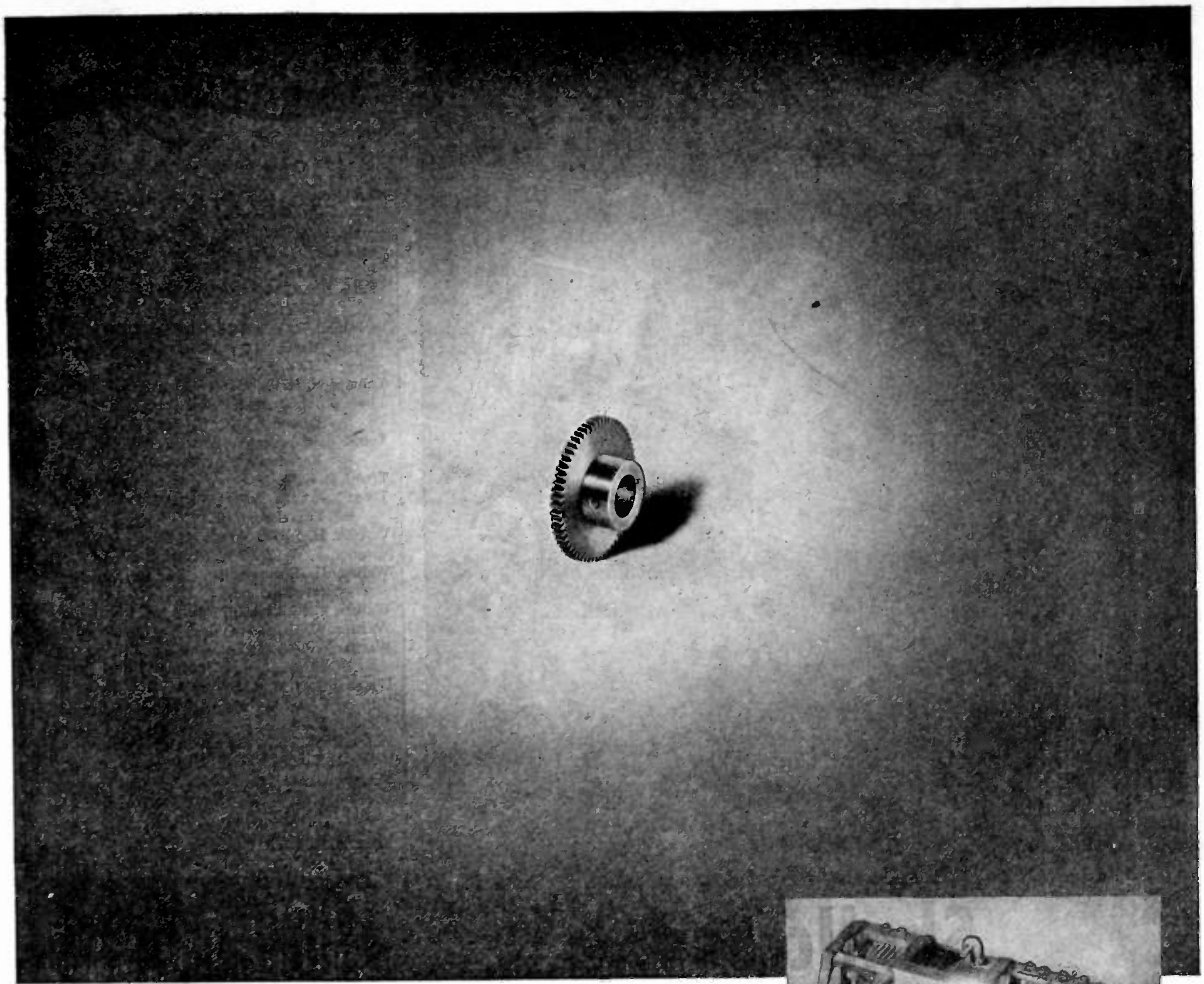
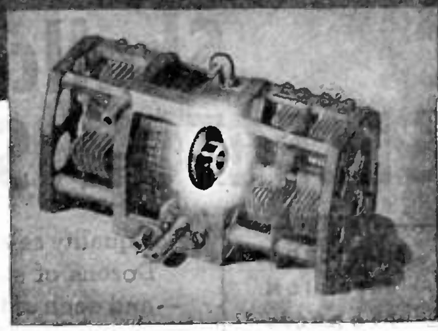




MEMBERSHIP

(Continued from page 42A)

Skroblsch, A., Eastern Air Devices, Inc., 585 Dean St., Brooklyn, N. Y.
Smale, K. E., 421 Wildwood, Jackson, Mich.
Smedstad, H. M., 7036—19 Ave., N. E., Seattle, Wash.
Smith, R. M., 147-53 Sanford Ave., Flushing, L. I., N. Y.
Swanlund, F. A., 2058 S. High St., Denver, 10, Colo.
Tepper, H. H., 42-01—163 St., Flushing, L. I., N. Y.
Thompson, R. C., 2316 Holly St., Denver, Colo.
Thomson, J. R., 112 Tyndall Ave., Toronto, Ont., Canada.
Toms, E. C., c/o Western Electric Co., Inc., 195 Broadway, New York, N. Y.
Troell, R. E., 34 N. Ferry St., Schenectady, 5, N. Y.
Troup, A. B., APO 503, c/o Postmaster, San Francisco, Calif.
Veritch, H. J., 333 Winthrop, Brooklyn, N. Y.
Wagner, F. W., 2555 W. 61 St., Chicago, 29, Ill.
Walker, D., Stranton, 10, Glebe Rd., Cheam, Surrey, England.
Walker, F. B., Lincoln, Kansas.
Warner, W. H., 4544 Quitman St., Denver, Colo.
Warshafsky, H., 1750 Davidson Ave., New York, 53, N. Y.
Weichardt, H. H., 455 W. 34 St., New York, 1, N. Y.
Wilson, T. M., 19 Nassau Ave., Schenectady, 4, N. Y.
Wisenbaker, T. C., 451 Main St., Waltham, Mass.
Wolf, H. A., 29-03—148 St., Flushing, L. I., N. Y.
Yanowitch, M., 1144 E. 18 St., Brooklyn, N. Y.
Yazell, G. W., 6-44B, Naval Research Laboratory, Anacostia Station, D. C.

I. R. E. People

Mr. B. Ray Cummings (A'18–M'20–SM'43), vice president in charge of engineering of the Farnsworth Television and Radio Corporation in Fort Wayne, Indiana, announced recently that Dr. H. Salinger (A'37), mathematical physicist, has returned to active duty in the Farnsworth Research Laboratories after a year's leave of absence, during which time he conducted specialized instructional work in physics and mathematics at various institutions in Indiana.

Dr. Salinger is noted for his work in the fields of telegraphy, telephone, and television, in which sciences he has specialized in original research over a period of many years.

Francois C. Henroteau

Coincidental with the return of Dr. Salinger, Mr. Cummings also announced the appointment of Dr. Francois C. Henroteau to a post on the Farnsworth research staff. Dr. Henroteau, who has a doctor of science degree from the University of Brussels, Belgium, and who was chief of astrophysics division at the Dominion Observatory in Ottawa, Canada, over a period of 14 years, has been an active contributor to the fields of astrophysics and television research in the major countries of Europe and America. At the Farnsworth Corporation he will specialize in the solution of optical problems as related to electronic television.

Future of Television

With thousands of engineers, technicians, mechanics, carpenters, and other skilled laborers required to build television stations and several times that number of people needed to manufacture receivers, there is every reason to believe that television in the post-war period will be a bigger industry than radio ever was, it was predicted by David B. Smith (A'35), director of research for Philco Corporation, in an address on "Electronics" before the Association of Customers' Brokers in New York on December 14, 1943.

"In the postwar television set, the picture will be larger than most of those available today, and some receivers may provide a picture as large as the average road map," Mr. Smith said. "You will probably have the television set in your living room, and you will turn the lights down, but not out,

(Continued on page 57A)

A typical group of H. F. radio coils insulated with Q-Max A-27 Lacquer.

Q-MAX A-27 LACQUER HAS LOW LOSS FACTOR OVER A WIDE FREQUENCY RANGE

The loss factor of Q-Max A-27 Lacquer is very nearly constant as the frequency increases from one megacycle, which is indicative of its excellent performance in the high frequency range. This feature, together with its low dielectric constant and other special characteristics, makes Q-Max A-27 Lacquer an outstanding high frequency coating medium.

Q-Max provides an excellent coating for R. F. solenoid windings and serves as an impregnant on multilayer or star coils. It is used as a tape saturant, a stiffening and strengthening medium, and a surfacer for wood or porous materials. Because of its low dielectric constant and excellent high frequency insulating characteristics, Q-Max is used widely in treating radio frequency coils.

New descriptive booklet sent on request.

Other C. P. products available to the communications industry are: a radiation-free line of copper or aluminum Coaxial Transmission Line, Auto-Dryaire for dehydrating transmission lines, new Sterling Switches, Antennas and Radiating Systems.

Communication PRODUCTS COMPANY, INC.

744 BROAD ST., NEWARK, N. J.
Factory: 346 Bergen Ave., Jersey City, N. J.

Communication Products Company, Inc.
744 Broad Street,
Newark, New Jersey

Send: ☐ Q-Max A-27 Lacquer Booklet

☐ Data on________

*Name*________

*Company*________

*Address*________

ARE YOU IN A HURRY FOR A GOOD TEST INSTRUMENT?

Use This Industrial Emergency Service On Radio and Electronic Supplies

Maintaining a complete stock of Test Instruments has been one of the toughest jobs we've had. The demand has far exceeded the supply. Yet our Industrial Emergency Departments have procured them as if by magic . . . saving many a victory-vital job from prolonged delays. Saving time and trouble getting vital Radio and Electronic Supplies is a highly specialized business with us. Whether you need one or one hundred different items, you'll get ALL of them from this ONE big reliable source faster than you ever thought possible considering wartime scarcities. Your nearest Emergency Service Distributor listed below is specially prepared to lend you every possible aid. Send along your orders or ask for big Reference Book & Buyer's Guide on company letterhead, please!

WRITE OR PHONE YOUR NEAREST DISTRIBUTOR

TERMINAL RADIO CORP.
85 Cortlandt St., Phone WOrth 2-4416 **NEW YORK 7**

WALKER-JIMIESON, INC.
311 S. Western Ave., Phone Canal 2525 **CHICAGO 12**

RADIO SPECIALTIES CO.
20th & Figueroa, Phone Prospect 7271 **LOS ANGELES 7**

NATIONAL INDUSTRIAL EMERGENCY SERVICE

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addrssing reply to company mentioned or to Box No. ....

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.

330 West 42nd Street, New York 18, N.Y.

RADIO ENGINEERS

Graduate engineers with laboratory experience needed for research and development work. Permanent employment with progressive corporation located in small city in central Pennsylvania. Persons now employed at essential activities at their highest skill cannot be considered without a statement of availability. Write for application form to Airplane & Marine Instruments, Inc., Clearfield, Pennsylvania.

RADIO AND ELECTRONICS ENGINEERS

We are seeking the services of one or two trained engineers who have had ample experience in transmitting-tube engineering. The men selected will not only be concerned with current war production, but should eventually fill good positions in postwar operation.

Also, we are looking for a few young engineers with good schooling and background to be trained for transmitting-tube development and production.

This is an excellent opportunity for men who qualify to connect with a progressive, highly regarded manufacturer of transmitting tubes. Many special benefits will be enjoyed in your association with this company.

Please reply in writing, giving complete details, past experience, etc. Interviews will be promptly arranged. Persons in war work or essential activity not considered without statement of availability. Address, Chief Engineer, United Electronics Company, 42 Spring Street, Newark 2, New Jersey.

ELECTRONIC OR AMPLIFIER ENGINEERS

A well-known company is in need of engineers experienced in the design and use of electronic equipment, particularly audio and control amplifiers.

Also, laboratory assistants and junior engineers, who are experienced in testing and laboratory work on electronic apparatus.

Send complete details to Post Office Box 30, Bloomfield, New Jersey.

ENGINEERS AND DRAFTSMEN

EXECUTIVE ENGINEER: Graduate electrical engineer with 15 years' experience in radio research and development, associated with the manufacturing of radio equipment.

DESIGN DRAFTSMEN: Experience in radio communications equipment drafting, as assistant, associate or draftsman.

ENGINEERING ASSISTANTS: May be radio engineers with about five years' experience, preferably in the manufacturing of radio equipment.

Positions offer post-war opportunities. Salaries open. Location, Connecticut. Send full details to Box 312.

RADIO ENGINEER

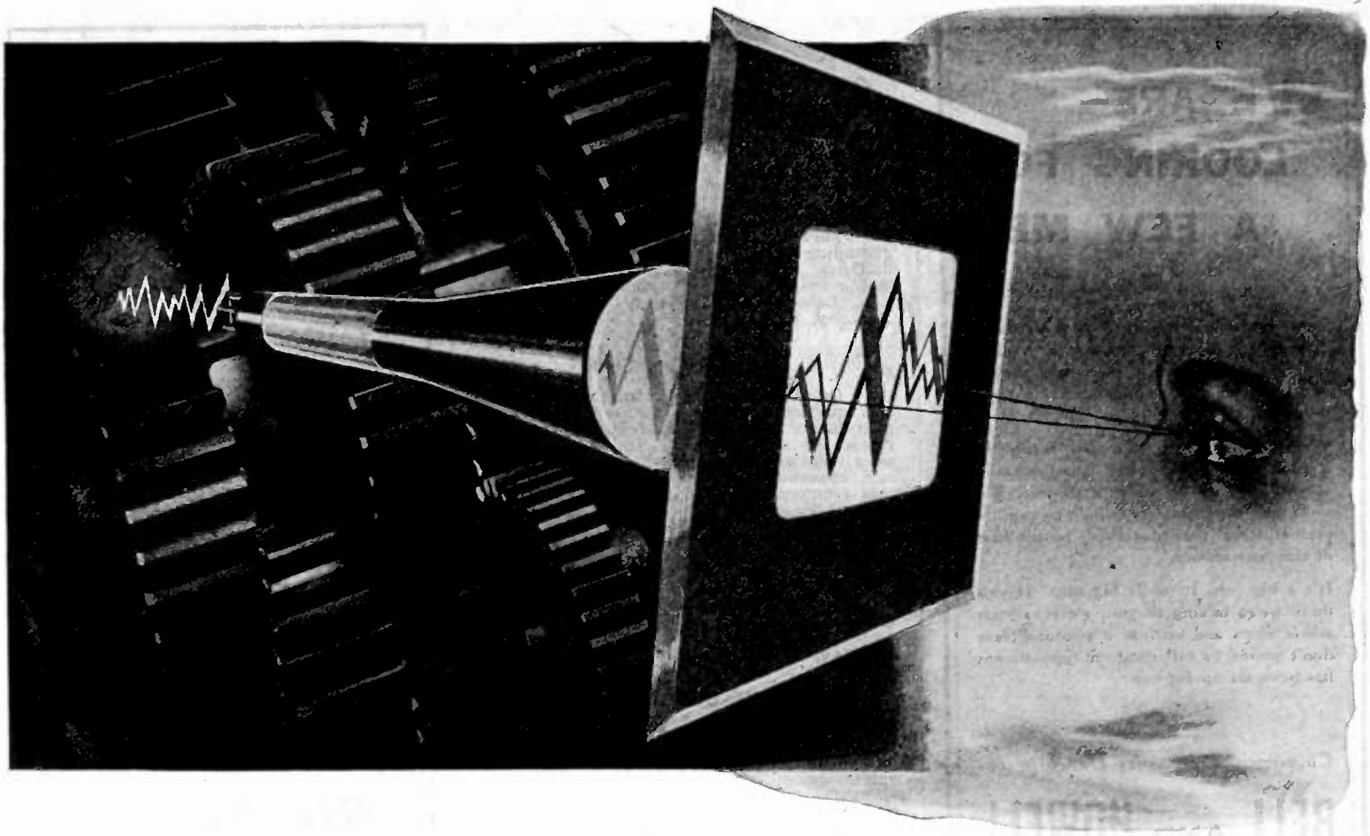
Experienced in the manufacture and testing of ultra-high-frequency apparatus; must be capable of taking complete charge of war projects. Splendid opportunity. War workers at highest skill need not apply.

Inquiries will be kept confidential. Please state age, experience and salary expected. Write Box 288.

ELECTRICAL AND RADIO DESIGN ENGINEERS

Familiar with analysis and design of complex circuits similar to those used in radio transmitter equipment. Should have five years

(Continued on page 52A)



Key to a world within a world

To inspect metal, judge its inner worth with the aid of electronics, is to add a vital chapter to war industry's book of knowledge. More, it is to write a preface to the mightier book of the future.

This same science of electronics, which finds the structural flaw in war metal, holds great possibilities whose commercial use awaits only the welcome day of peace.

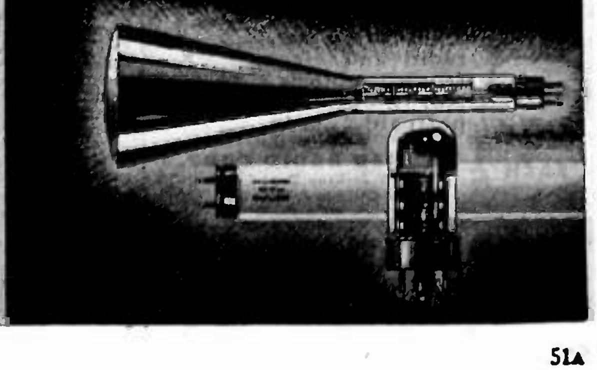
Infinite additions to the knowledge, the safety, the comfort of modern man continuously reveal themselves in the quick flutter of the electronic tubes.

This is an inspiring reason why at Sylvania, in our work with electronics, as in everything else we do to widen the range of the eye and the ear, we set for ourselves a single goal — the highest standard known.

SYLVANIA ELECTRIC PRODUCTS INC. formerly Hygrade Sylvania Corporation
EXECUTIVE OFFICES: 500 FIFTH AVENUE, NEW YORK 18, N. Y.

RADIO TUBES, CATHODE RAY TUBES, ELECTRONIC DEVICES, INCANDESCENT LAMPS, FLUORESCENT LAMPS, FIXTURES AND ACCESSORIES

AIDING THE HOME FRONTS "KNOW-HOW"—Sylvania Fluorescent Lamps and Fixtures give war workers the light they need to produce their armament miracles. Sylvania Radio Tubes bring the news of the world to the American family, keep our people mentally alert. Sylvania Incandescent Lamps economically protect the eyes of the American family. Indeed, the Sylvania name now, as always, means the ultimate in product performance.



POSITIONS OPEN

(Continued from page 50A)

full-time commercial or research experience. Must have B.S. in E.E., or equivalent; thorough grounding in engineering electronics and familiarity with high-voltage rectifier systems. Apply in writing, to Personnel Office, Radiation Laboratory, University of California, Berkeley, California.

RADIO TECHNICIAN

In Brooklyn war plant. Must be able to use test equipment, to set up and use laboratory test instruments and supervise production testing of radio parts and electronic equipment. Will consider men with amateur radio experience. State age, education, experience. Availability certificate required. Write to Box 308.

ELECTRONIC TUBE DESIGN ENGINEER

Experience in actual design and manufacture of large high-vacuum tubes, or electrical and mechanical design, as well as in process, test, and application techniques. Essential workers need release statement. Write to Box 309, giving complete details including salary expected.

ELECTRONIC DESIGN ENGINEERS

One of the largest manufacturers of radio equipment, located in Eastern Massachusetts, has openings for several engineers. Work involves design and development of electronic apparatus having a wide field of application both now and after the war.

A Master's degree, or a B.S. degree in Electrical Engineering with two years' experience in electronic work, would be desirable but not absolutely necessary, depending upon the individual. Those now employed in an essential activity must be able to obtain release. Applicants should submit their qualifications and salary expected to Box 307.

RADIO ENGINEERS AND TECHNICIANS

A progressive company with a sound background in radio and electronics needs, at once, several men with training and experience in any phase of the radio industry. The work open is vital to the war effort but offers a promising post-war future for the right men. College degree or equivalent experience necessary. Men now engaged at highest skill on war production should not apply. Write Box 294.

ELECTRICAL OR CHEMICAL ENGINEER

. . . thoroughly versed in the theory of liquid and solid dielectrics for the position of chief engineer. To direct the research, development and general laboratory on capacitors and capacitor applications. This is an unusual opportunity for a capable engineer interested in his present and postwar future. Write to Industrial Condenser Corp., 1725 W. North Ave., Chicago, Ill.

RADIO ENGINEERS

Transcontinental & Western Air, Inc. has openings at Kansas City for three Radio Engineers in the Communications Department. Applicants should have completed an electrical or radio engineering course, or should have had one to two years of practical experience. These openings are permanent.

For additional details and application forms, write to Personnel Department, Transcontinental & Western Air, Inc., Kansas City, Missouri.

PATENT ATTORNEYS

Patent attorneys, who are electronic physicists and electrical or radio engineering graduates who have maintained contact with the field of high-frequency electronics, radio manufacture, carrier-current telephony, and light-current circuit design and computing, can make a substantial contribution in research or development jobs with one of the National Defense Research Committee laboratories located in the East. The project is secret but is one of the most urgent of all research jobs now under way for the Government.

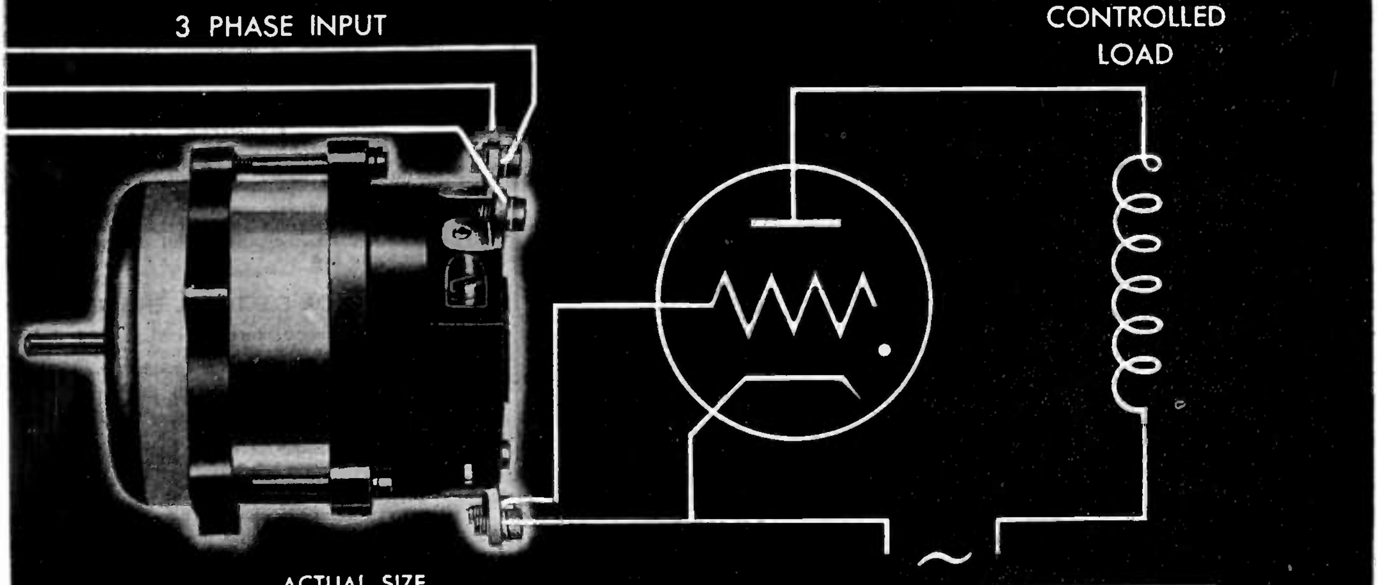
An electrical engineering background in light currents is essential, and amateur radio experience, inventive ability and ingenuity in the design and layout of radio equipment would be of considerable help.

Facilities for specialized refresher training

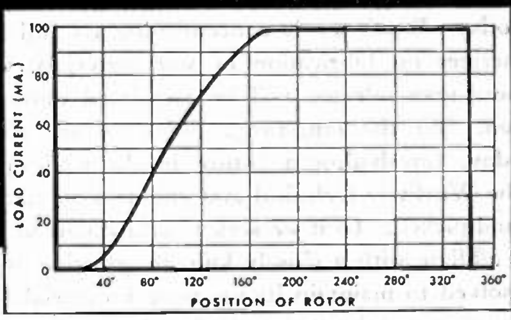
(Continued on page 54A)

Design Engineers!

Here's a versatile unit with many electronic control applications...THE KOLLSMAN CIRCUTROL

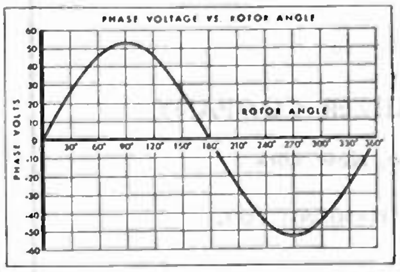


CIRCUTROL UNIT
used as
Thyratron Phase Control



Typical of the many special applications for which design engineers have found the Kollsman Circutrol particularly suited, is phase control of Thyratron type units. In this application the unit offers accurate linear control, as shown by the above graph.

When used as a rotatable transformer, the Circutrol Unit produces a phase voltage which varies sinusoidally with the angular position of the rotor as shown in the graph at right.



Another advantage of the unit as a rotatable transformer is that it is designed to withstand continuous rotation at speeds up to 1800 R.P.M., although many applications require nothing more than positioning of the rotor.

Electrically, the Circutrols are motor-like precision units having high impedance two- or three-phase stator windings and single-phase rotors. Units are available which operate from 32, 115 and 220 volts, 60 cycles, and 110 volts, 400 cycles.

These units may also be used as single or polyphase induction regulators, controllable voltage modulators, single or polyphase alternators or phase shifters.

For complete information about the Kollsman Circutrol write to Kollsman Instrument Division of Square D Co., 80-04 45th Ave., Elmhurst, N. Y.

ELECTRICAL EQUIPMENT · KOLLSMAN AIRCRAFT INSTRUMENTS

SQUARE D COMPANY

ELMHURST, NEW YORK · GLENDALE, CALIFORNIA

POSITIONS OPEN

(Continued from page 52A)

and orientation in the particular field may be available. Anyone who possesses these qualifications and is interested in a vital wartime development job for the duration may get further details on request. All inquiries will be held confidential. Address Box 299.

ELECTRONIC ENGINEER

Electronic engineer with M.A., Ph.D., or the equivalent in physics, for research and design in geophysics. Experience in filter design and sound recording is desirable. Write to Independent Exploration Company, 901 Esperson Building, Houston, Texas.

ENGINEER

AAA-1 eastern manufacturer, over 75 years operation and leader in growing industry, has immediate permanent position for chemical, electrical, electronic or chemical-metallurgical engineer to organize and increase efficiency of production activities. Send full details of experience to Box 301.

RADIO OR ELECTRICAL ENGINEER

Capacitor manufacturer located in New Bedford, Mass., wants an electrical or radio engineer—man or woman—for equipment- and circuit-development work.

Permanent postwar future for right person. This firm has excellent laboratory facilities and is a leader in its field.

Applicant should be college graduate with a degree—or equivalent experience—in radio engineering or electrical engineering.

Interview in Boston, New Bedford or New York can be arranged. Traveling expenses paid to place of interview.

Write fully, giving age, education, experience, etc. Address reply to Box 302.

PHYSICIST OR ELECTRICAL ENGINEER

Leading manufacturer of industrial radio-frequency equipment desires the services of a physicist or electrical engineer to direct developmental and applications laboratory. This field is expanding rapidly and offers excellent opportunities for advancement. Position of a permanent nature. Present activities devoted entirely to the war effort. Address replies to Box 306.

SOUND AND PROJECTION ENGINEERS

Openings exist for sound and projection engineers. Several years experience in the installation and maintenance of 35 mm motion-picture equipment of all types required. Must be draft exempt or over draft age and free to travel anywhere in the United States. Basic starting salary $3200. U. S. Army Motion Picture Service, Engineering and Maintenance Division, 3327-A Locust Street, St. Louis, Missouri.

RADIO ENGINEERS

Permanent radio-engineering position in Southern California for men with creative and design aptitude, especially with UHF circuits. Starting salary and advancement depends upon the engineer's experience and ability.

Applications are solicited from persons that are not using their highest skills in war work.

Write complete qualifying educational training and experience to Chief Radio Engineer, Bendix Aviation, Ltd., in care of The Shaw Company, 816 W. 5th Street, Los Angeles 13, California.

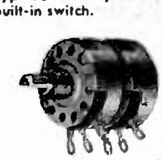
ELECTRONIC ENGINEER

or electrical engineer with high frequency experience, preferably with some background in mechanical engineering. Position with well established company of known reputation in the Middle-West with post-war possibilities in the manufacture of industrial electronic equipment. State education, experience, salary expected, marital and draft status. Write Box 313.

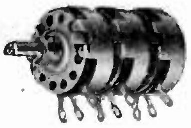
The foregoing positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. ....

"It's this solid molded resistor element that makes the difference"

Type JS Bradleyometer with a built-in switch.



Bradleyometers may be used singly or assembled for dual



— or triple construction to fit any control need.

The resistor element in Allen-Bradley Type J Bradleyometers has substantial thickness (approximately 1/32-inch thick), and in this respect differs from the film, paint, or spray type variable resistors. The Allen-Bradley variable resistor is molded as a single unit with the insulation, terminals, face plate, and threaded bushing. This simple construction does away with rivets, welded or soldered connections, and unreliable conducting paints. The Allen-Bradley type of variable resistor will prove reliable under all extremes of service conditions.

During manufacture, the resistor element may be varied throughout the length of the element to provide practically any resistance-rotation curve. Once the unit has been molded, its performance is not affected by heat, cold, moisture, or hard use. It not only is remarkably quiet when first manufactured but gets even better with age.

Bradleyometers are the only continuously adjustable composition resistors having a two-watt rating with a good safety factor. The Allen-Bradley Bradleyometer is the only commercial type variable resistor that will consistently stand up under the Army-Navy AN-QQ-S91 salt spray test. Write for specifications.

Allen-Bradley Company, 114 W. Greenfield Ave.
Milwaukee 4, Wisconsin

FIXED RESISTORS

½ WATT

Bradleyunits are molded, fixed resistors with lead wires embedded in homogeneous resistor material. They will sustain an overload of ten times rating for a considerable period of time without failing. No special wax impregnation is necessary to pass the salt water immersion test. Available in standard RMA values from 10 ohms to 10 megohms in both the insulated and non-insulated types.

1 WATT

ELEMENT ACTUAL SIZE

A·B ALLEN-BRADLEY

FIXED & VARIABLE RADIO RESISTORS

QUALITY

Future of Television

(Continued from page 48A)

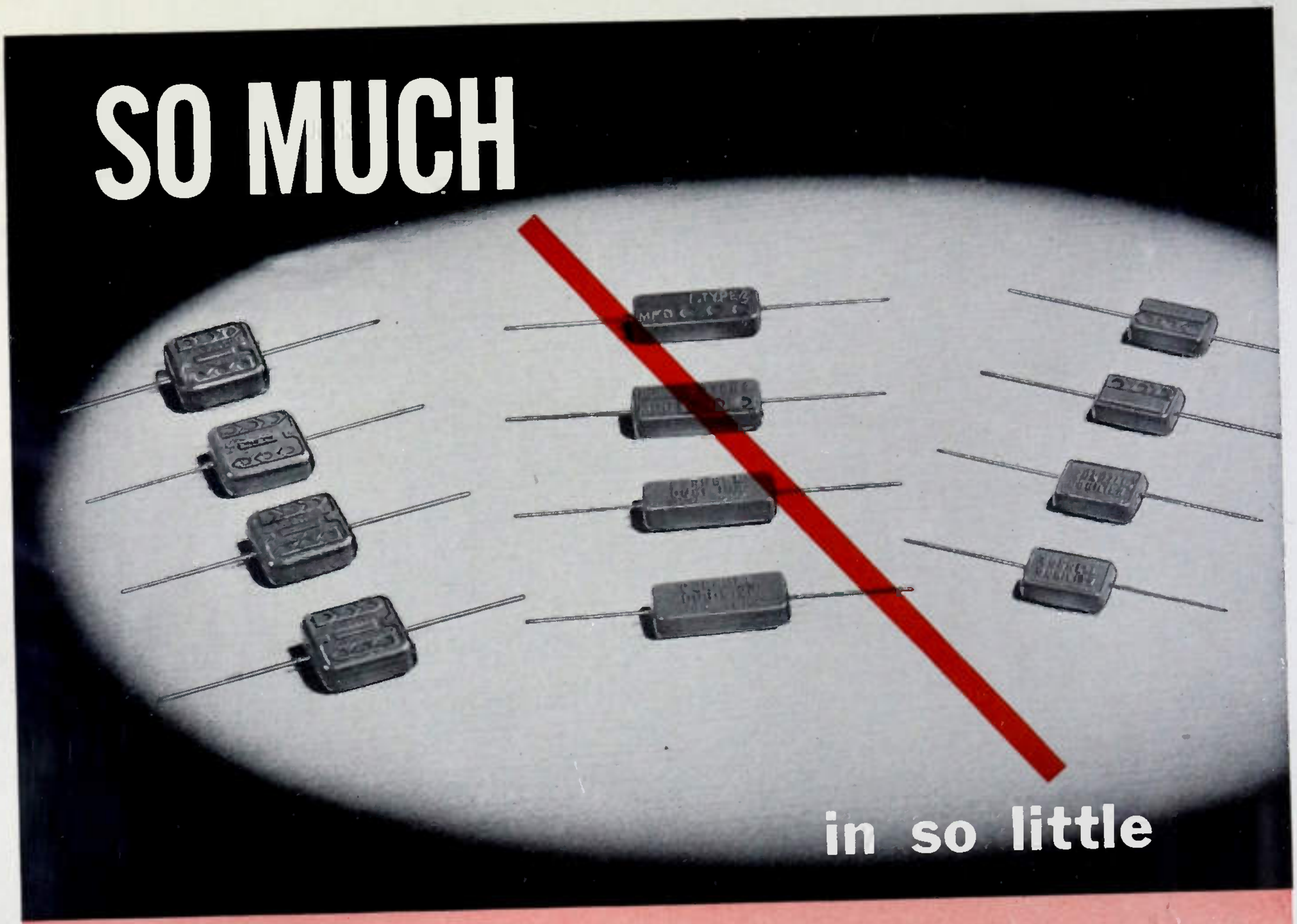
when you look at it. In New York, Philadelphia, and other Atlantic seaboard cities, you may very well have your choice of several programs within a few years. You will see sporting events and scenes from Washington, perhaps the President delivering a fireside chat or sessions of Congress, or variety shows and drama from New York. New kinds of entertainment may be created as a result of television just as the movies changed vaudeville and the theatre."

Rapid expansion of television as soon as the war is over will depend upon the creation of networks linking stations together, so that the best entertainment and news programs can be made available to the viewing audience, Mr. Smith pointed out.

"Sound broadcasting entered its period of greatest growth and expansion when network operation got underway, and the same principles will apply to television," he continued. "Ultra-high-frequency television relay stations have already been developed by Philco Corporation and other research groups to link television transmitters together, and these are in successful operation today. Through these links most of the television audience in the United States made history last night by viewing the official motion pictures of the Cairo and Teheran meetings of President Roosevelt, Prime Minister Churchill, and Marshal Stalin that were put on the air in New York City and relayed to Philadelphia and Schenectady. In the post-war years, through television, people in their own homes will be able to see these historic events as they occur."

INDEX

DISPLAY ADVERTISERS


SO MUCH
in so little