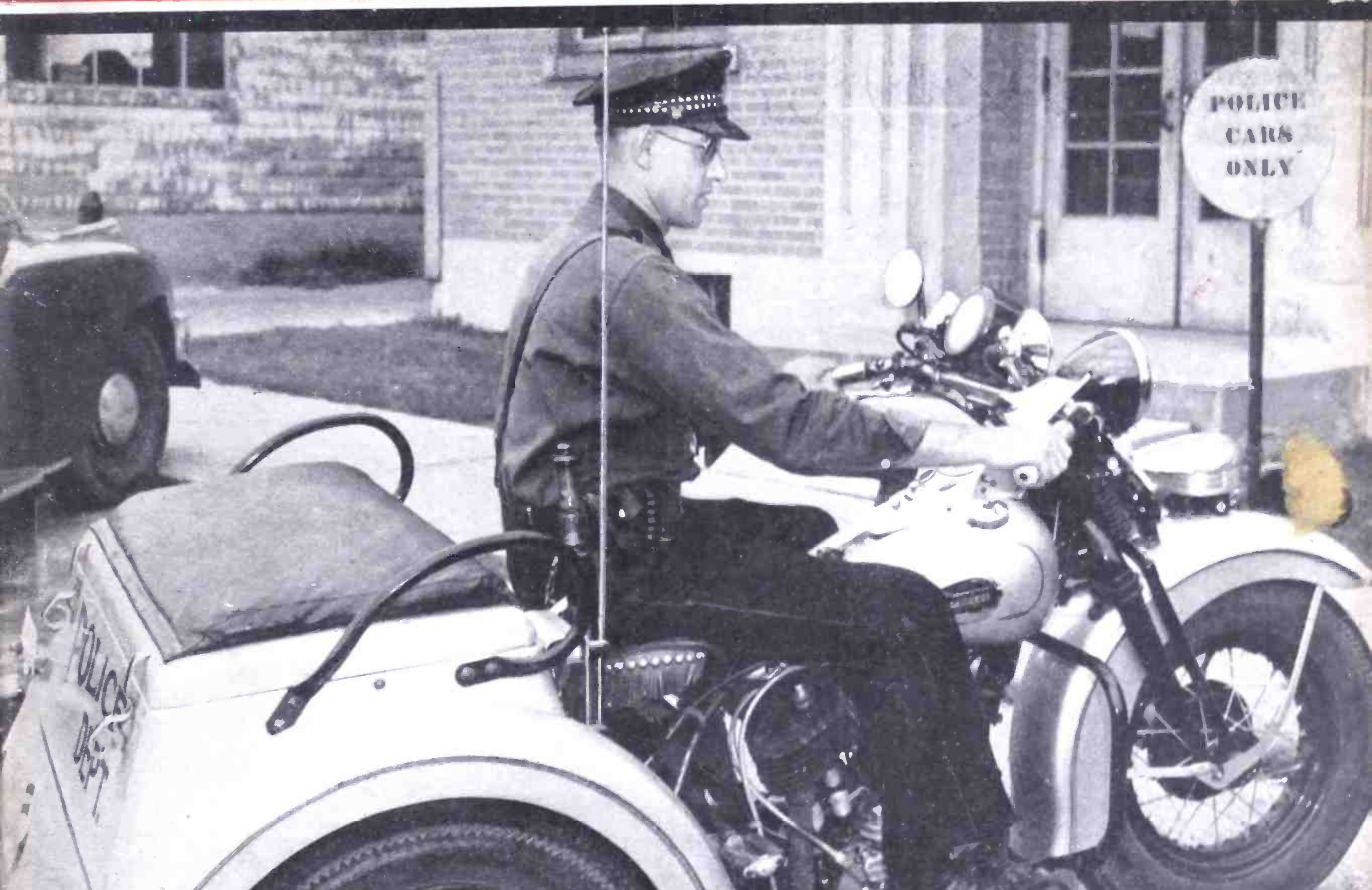
COMMUNICATIONS

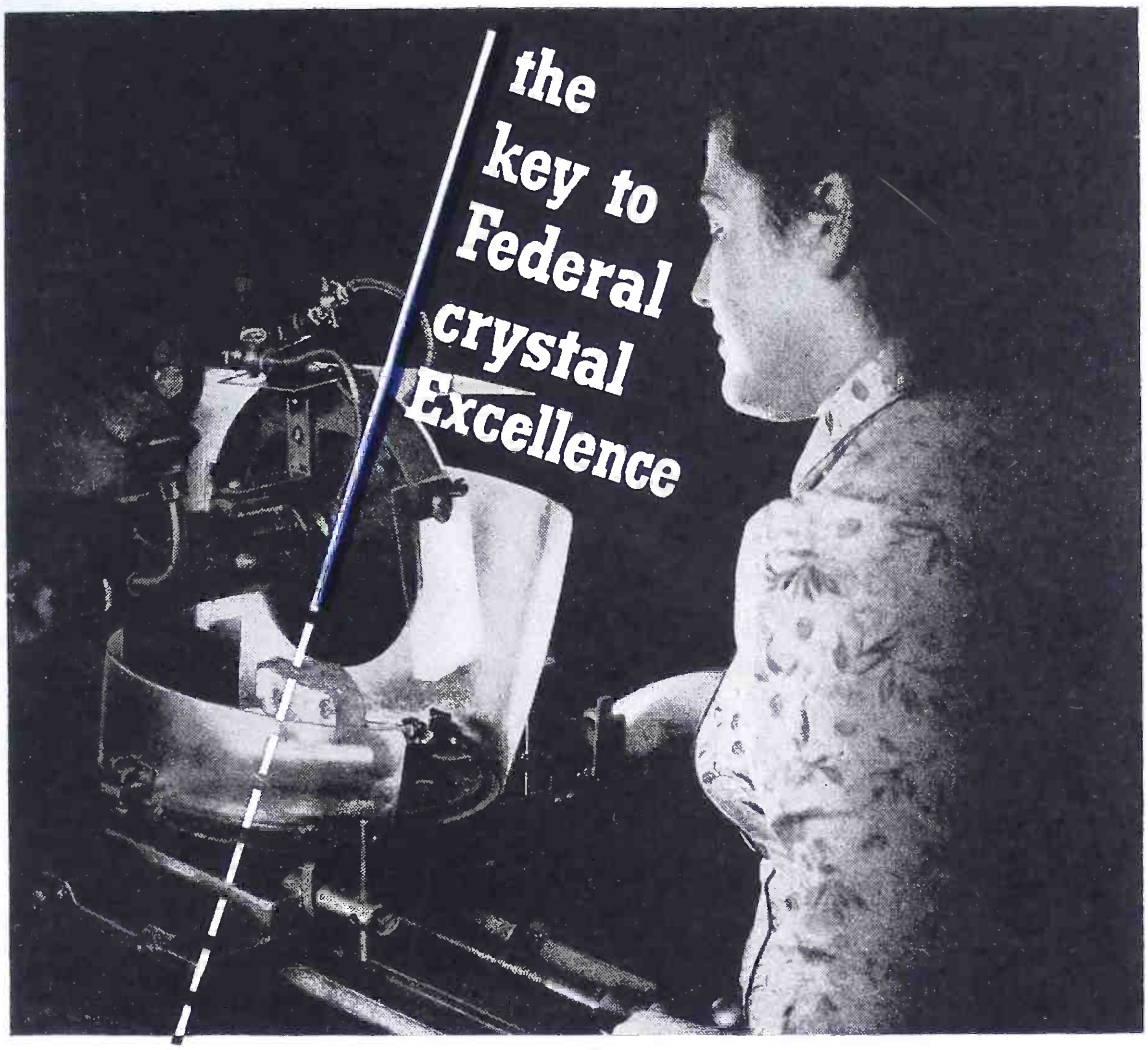


MARCH

★ RADIO ENGINEERING
★ RAF RADIO EQUIPMENT
★ COAXIAL CABLE LOSS AT U-H-F
★ NEGATIVE CAPACITANCE
★ HIGH Q AUDIO REACTORS
★ AIRCRAFT COMMUNICATIONS

1944

INSPECTION



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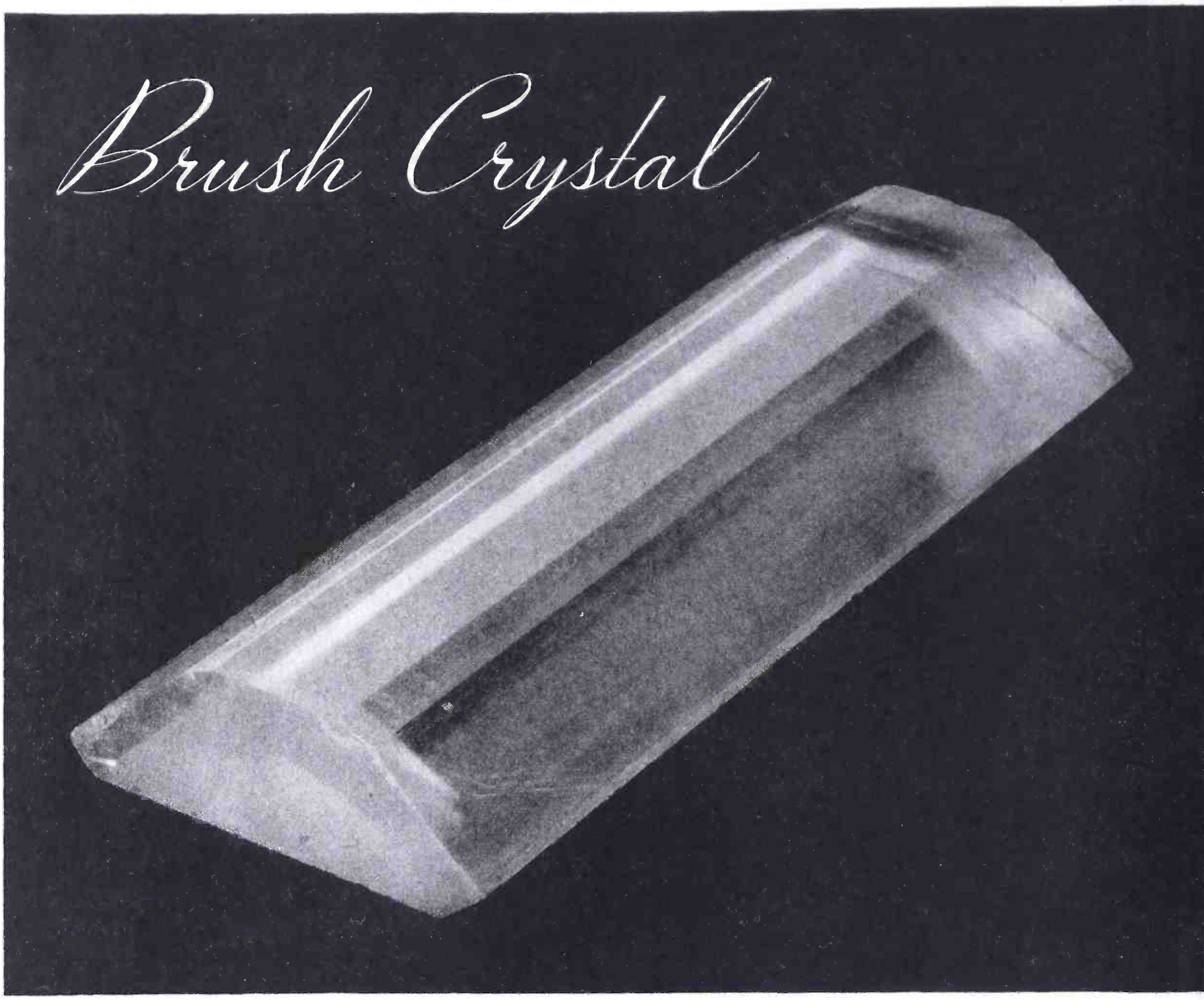
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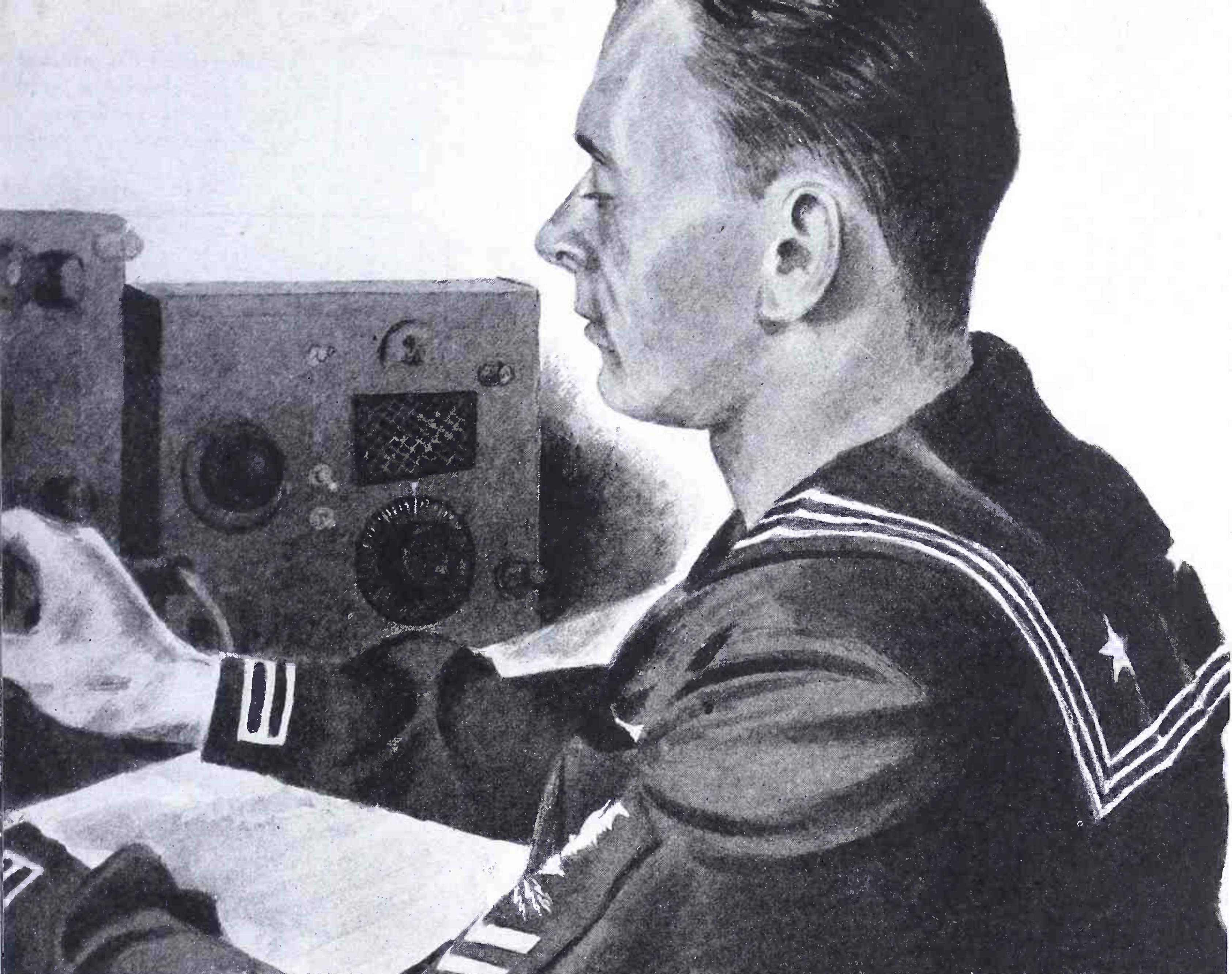

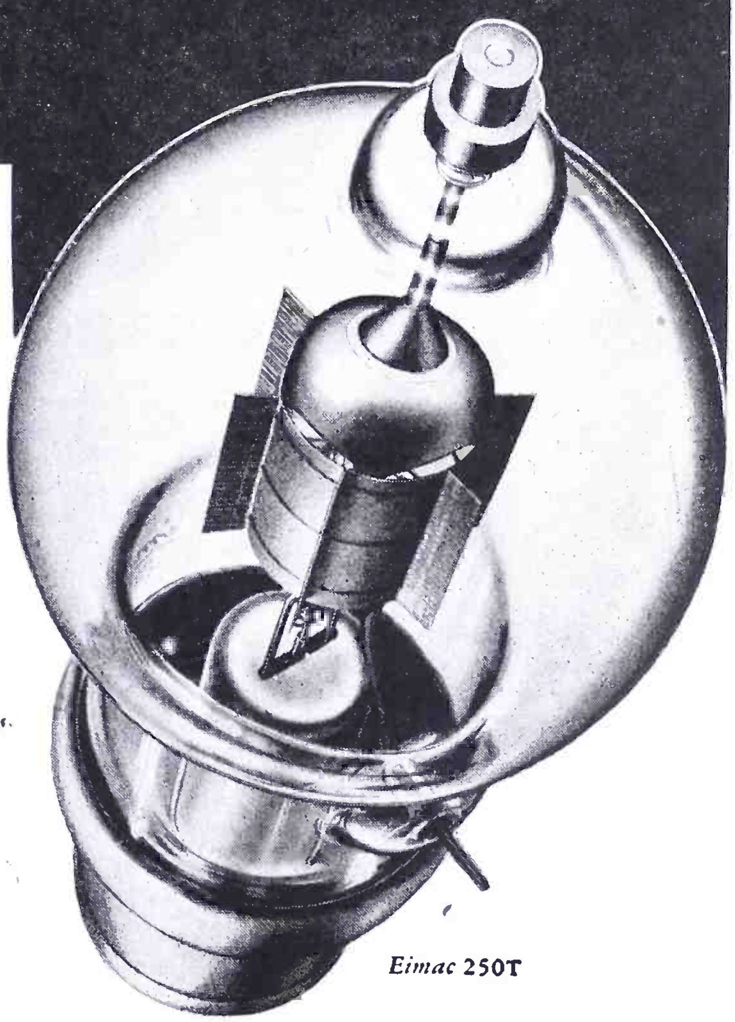

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help keep the communications

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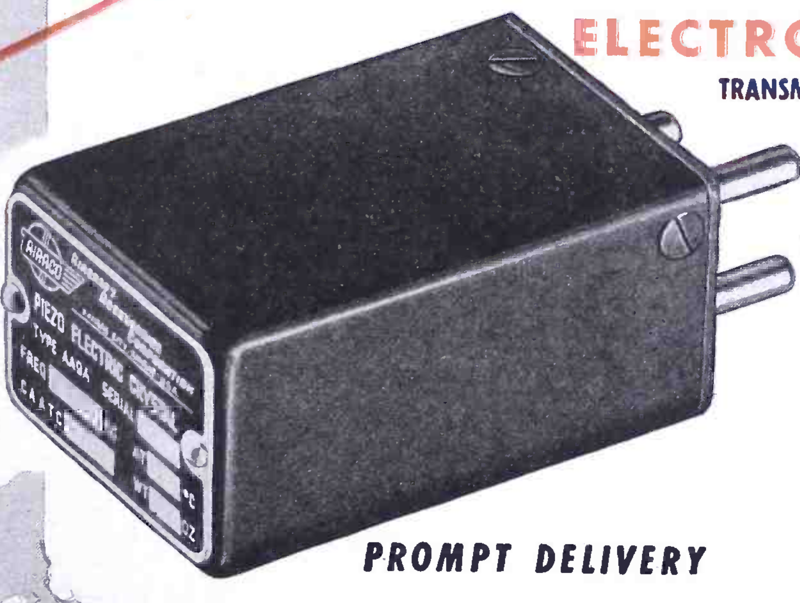

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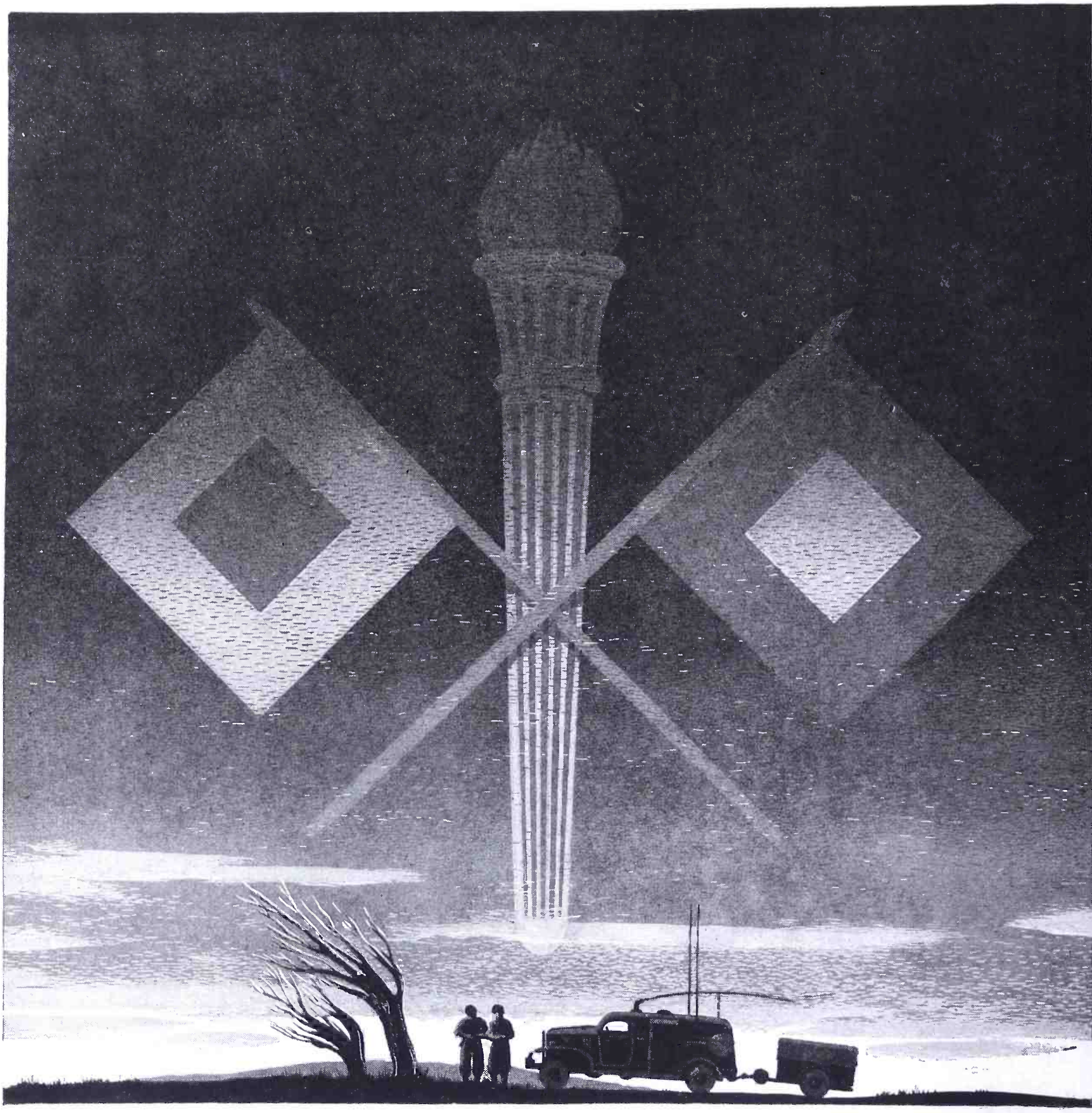

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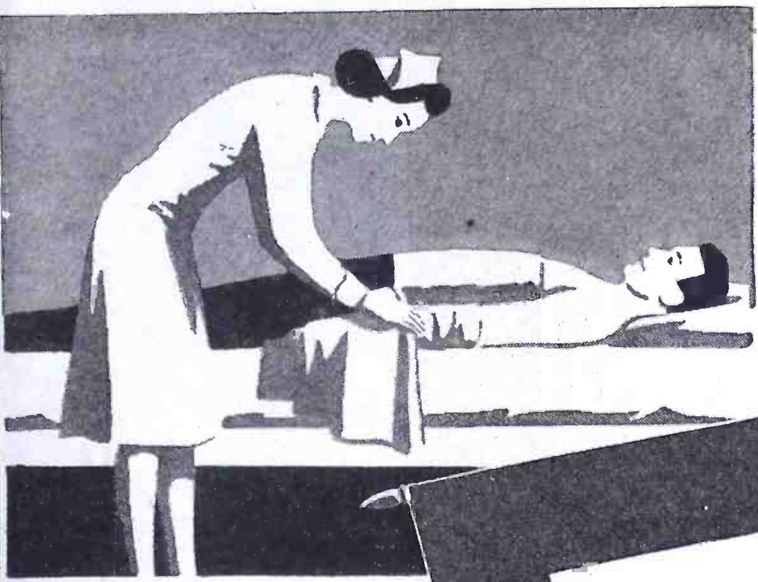

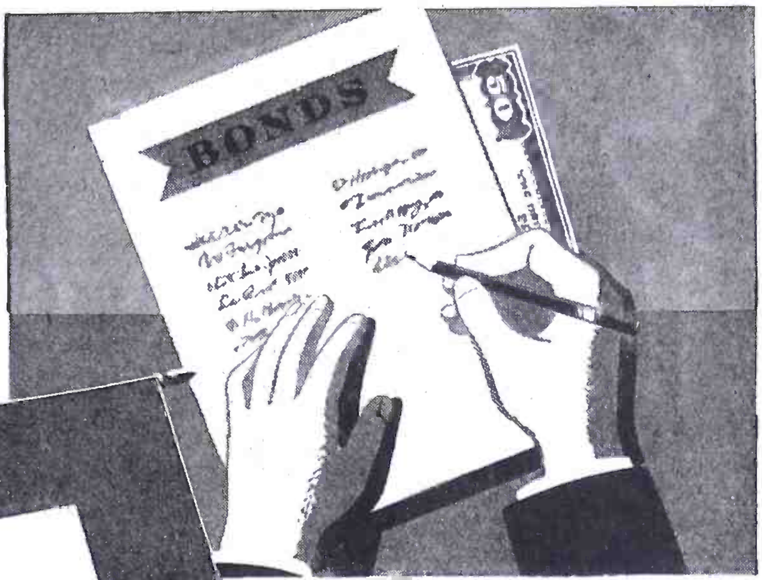

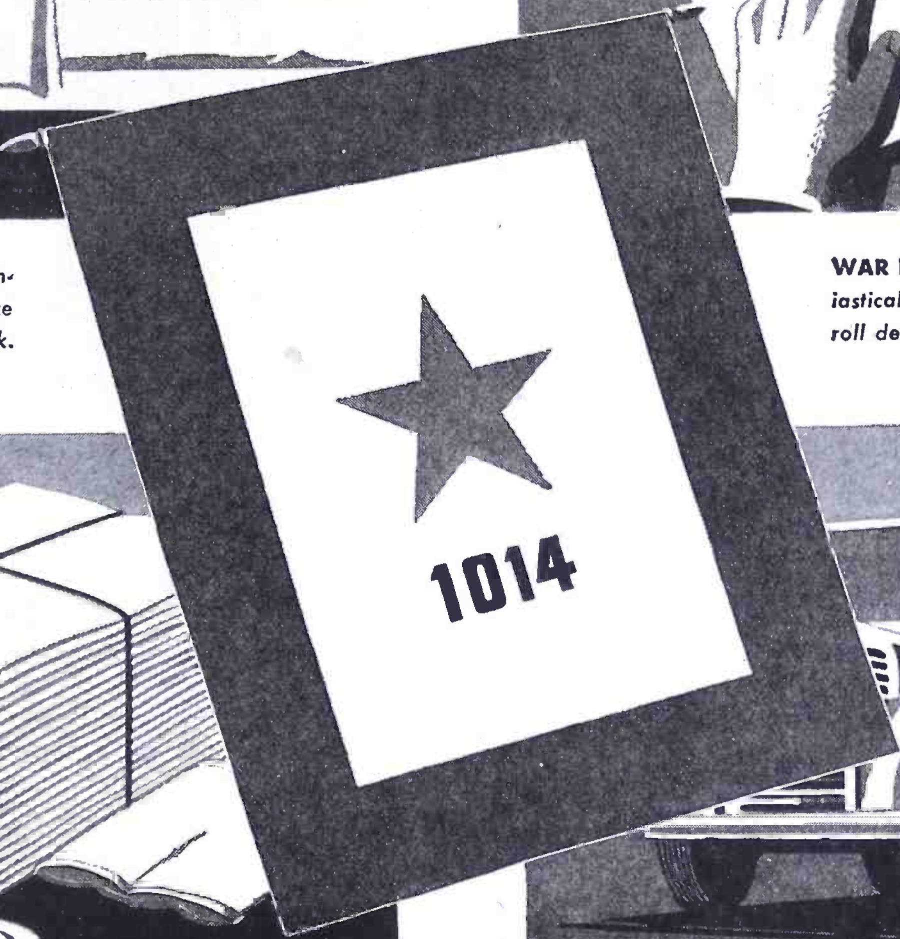

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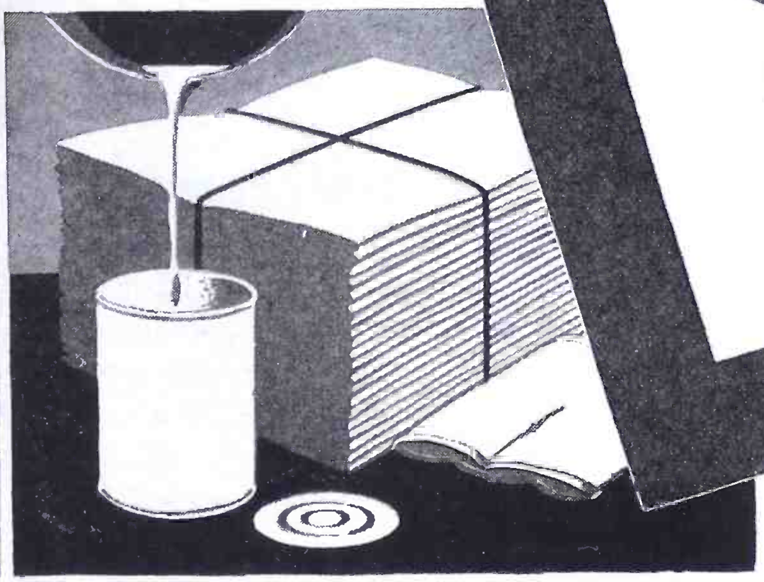

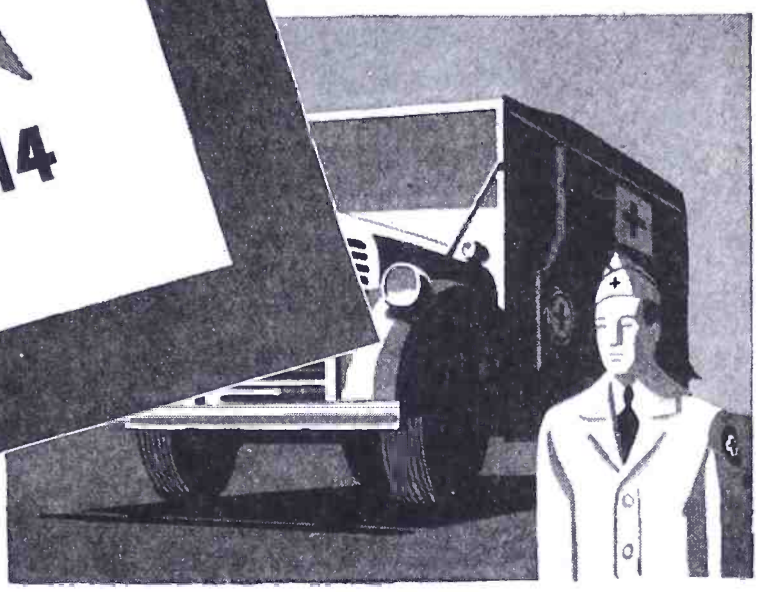

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THERE IS a Third Front, here at home, on which the men and women of Breeze are fighting. Putting 10% and more of their pay in war bonds, giving blood regularly to the Red Cross Blood Bank, cooperating to the fullest extent in civilian defense activities and government war campaigns, Breeze workers are a part of the great team that is backing up the boys at the front. Without this teamwork, the efforts of our fighting men might well be wasted.

And in addition to their outside work, the men and women of Breeze are on the job day and night, turning out in tremendous quantities the well-known Breeze products which are serving America today on fighting fronts the world over.

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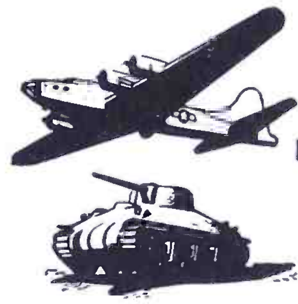

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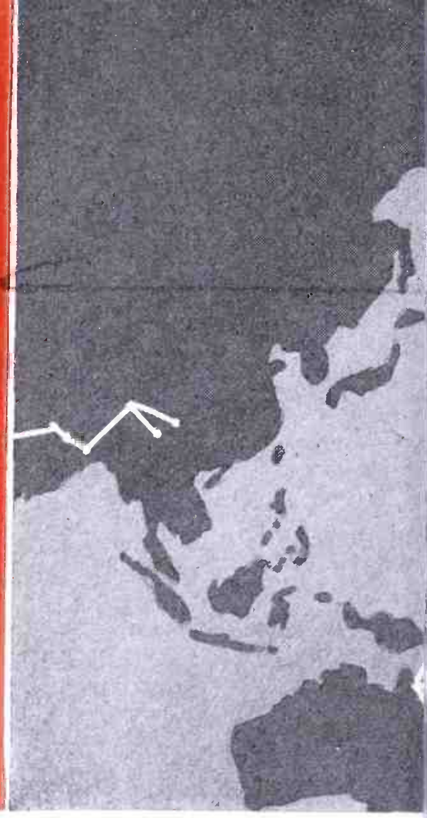
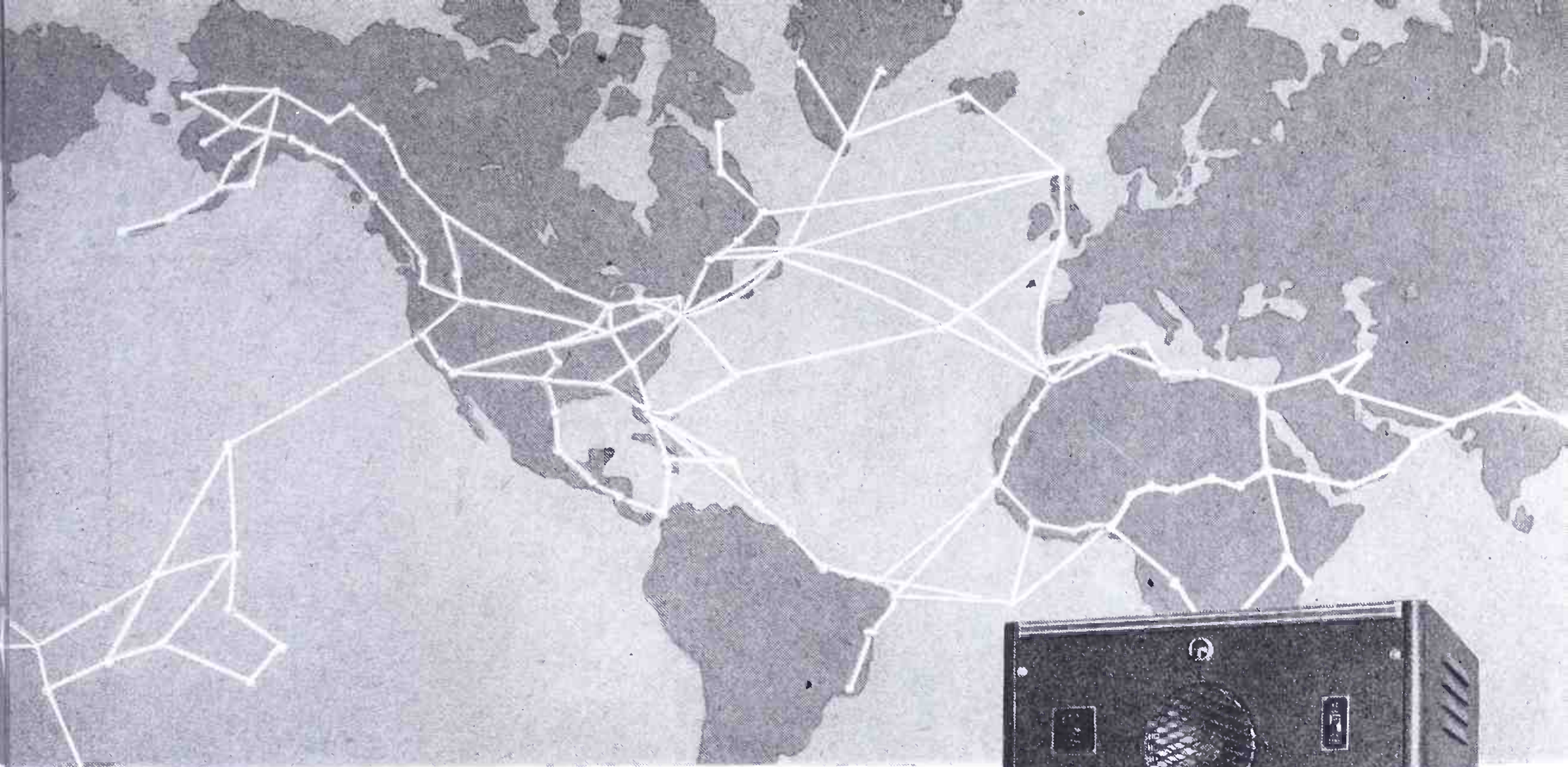


Signal Corps Photo



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Airports are more than bricks and mortar. Our non-technical booklet, "Highways of the Air," explains the importance of radio to aviation. Write for your free copy . . . Desk R-3

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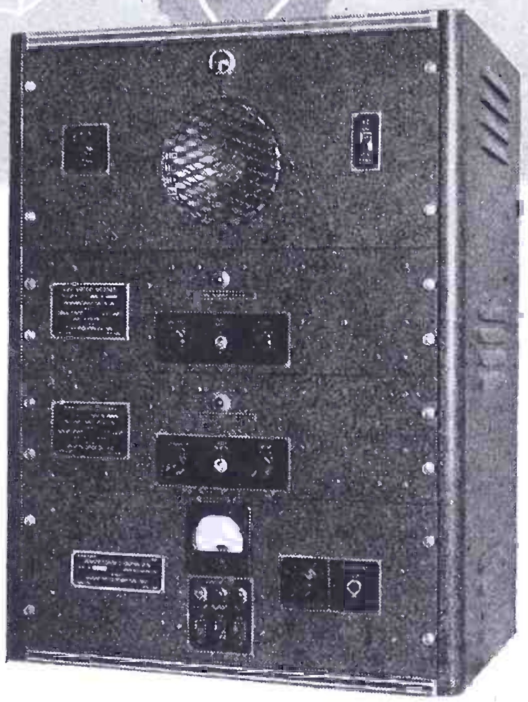
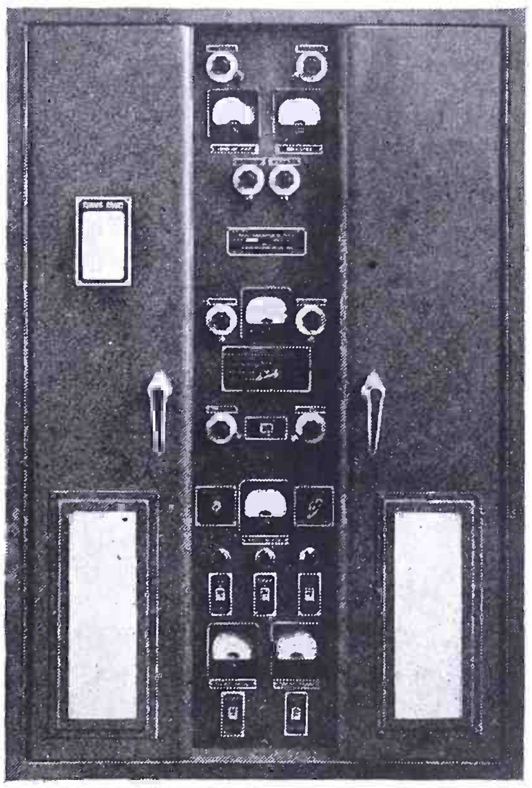


Control Cabinet Assembly. Consists of transmitter remote control unit, loud speaker and two fixed frequency airport receivers.



Transmitter for Airport Traffic Control. Output rating 50 watts, Frequency range 116 to 126.25 megacycles.

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ON A FINE JOB, WELL DONE!

THE Treasury "Star" Flag—the bond-buying counterpart of the Army-Navy "E"—marks plants with at least 90% of personnel participating in the Payroll Savings Plan to at least 10% of gross payroll, *and also having reached, or topped, a War Loan Drive quota!*

The successful close of the 4th War Loan Drive finds many more "Star" Flags than ever before flying over the industrial plants of America. To all these, go the heartiest thanks of the nation, and the deep appreciation of the Treasury Department for a great job! And to those who may not quite have qualified for the "Star," go equally sincere thanks—and the confidence that soon they, too, will join the ranks of the "Star" fliers.

One thought that many concerns have found helpful in stepping up the intake from their Payroll Savings Plans is this. In many cases the Treasury Representative in a plant has been able to point out the fact that during *Loan Drive periods* the employees have found it possible to spare much more than they had counted on when setting up their original subscription, and that—*when properly approached*—a very substantial fraction of such employees will decide they can well afford a distinct increase in their current Payroll Savings Plan.

Talk this over with your Treasury Representative—it offers important possibilities when correctly handled. And again accept the Treasury Department's congratulations for your fine work in helping to put over the 4th War Loan.

LET'S **ALL** *KEEP BACKING THE ATTACK WITH* **WAR BONDS**

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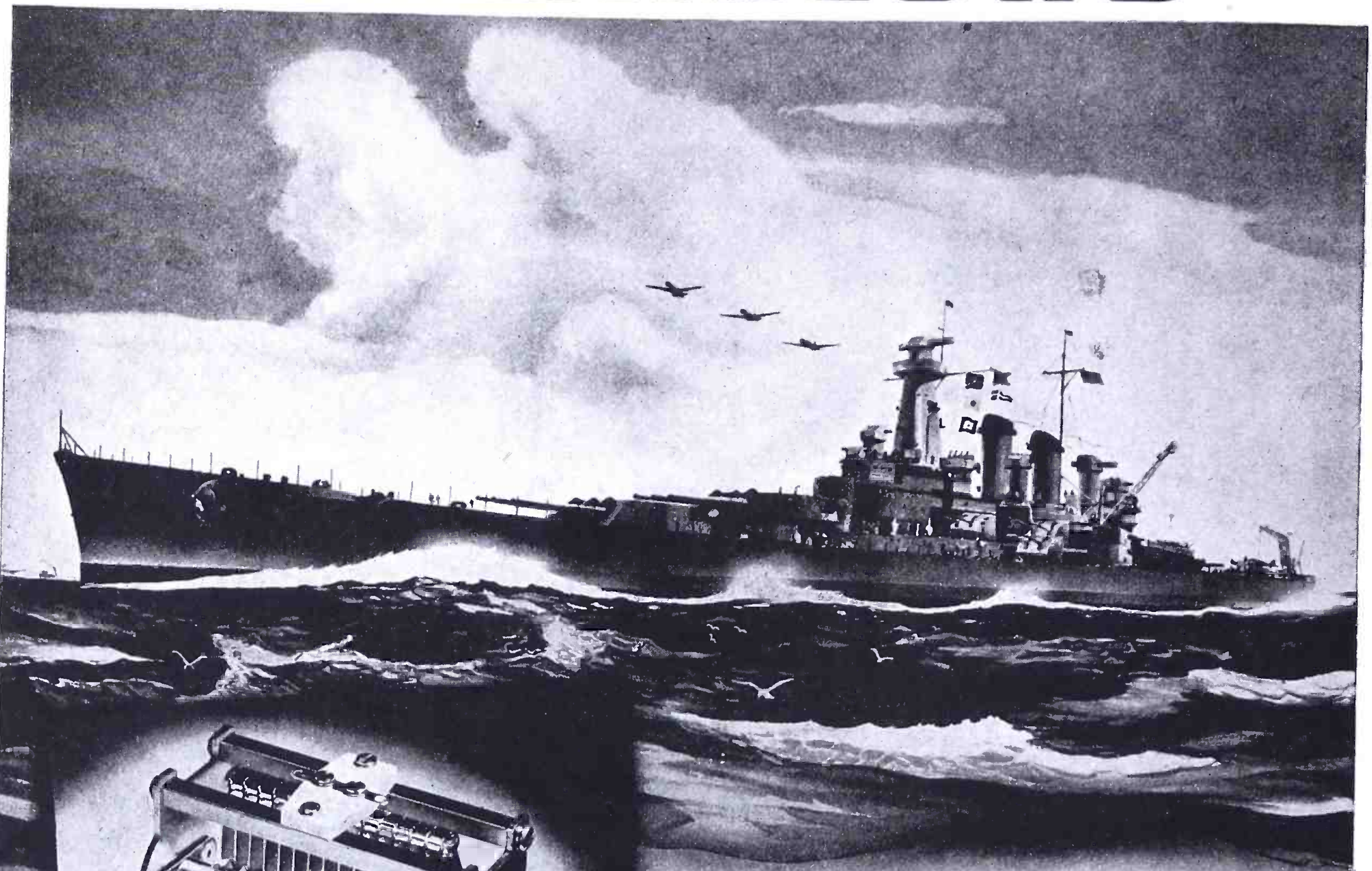

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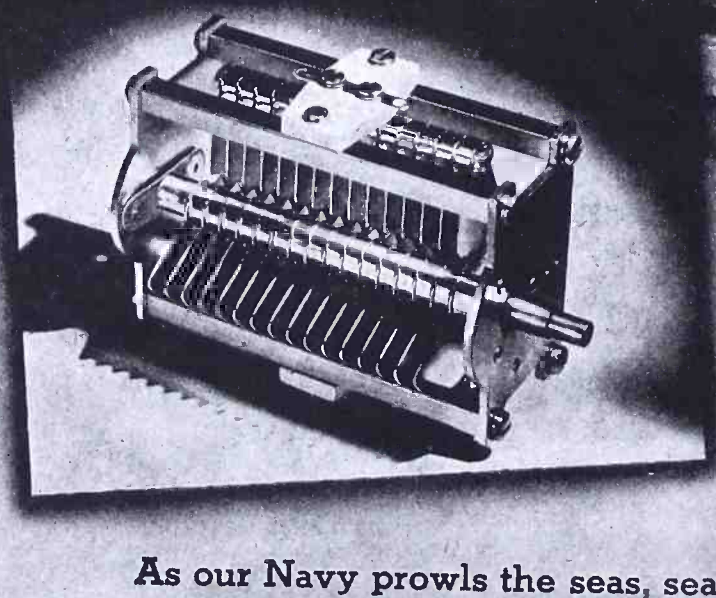

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As our Navy prowls the seas, searching out and destroying the enemy, Hammarlund Radio products help guide our great ships to certain victory. In commercial type marine equipment too, you'll find Hammarlund products are outstanding for their record of service.

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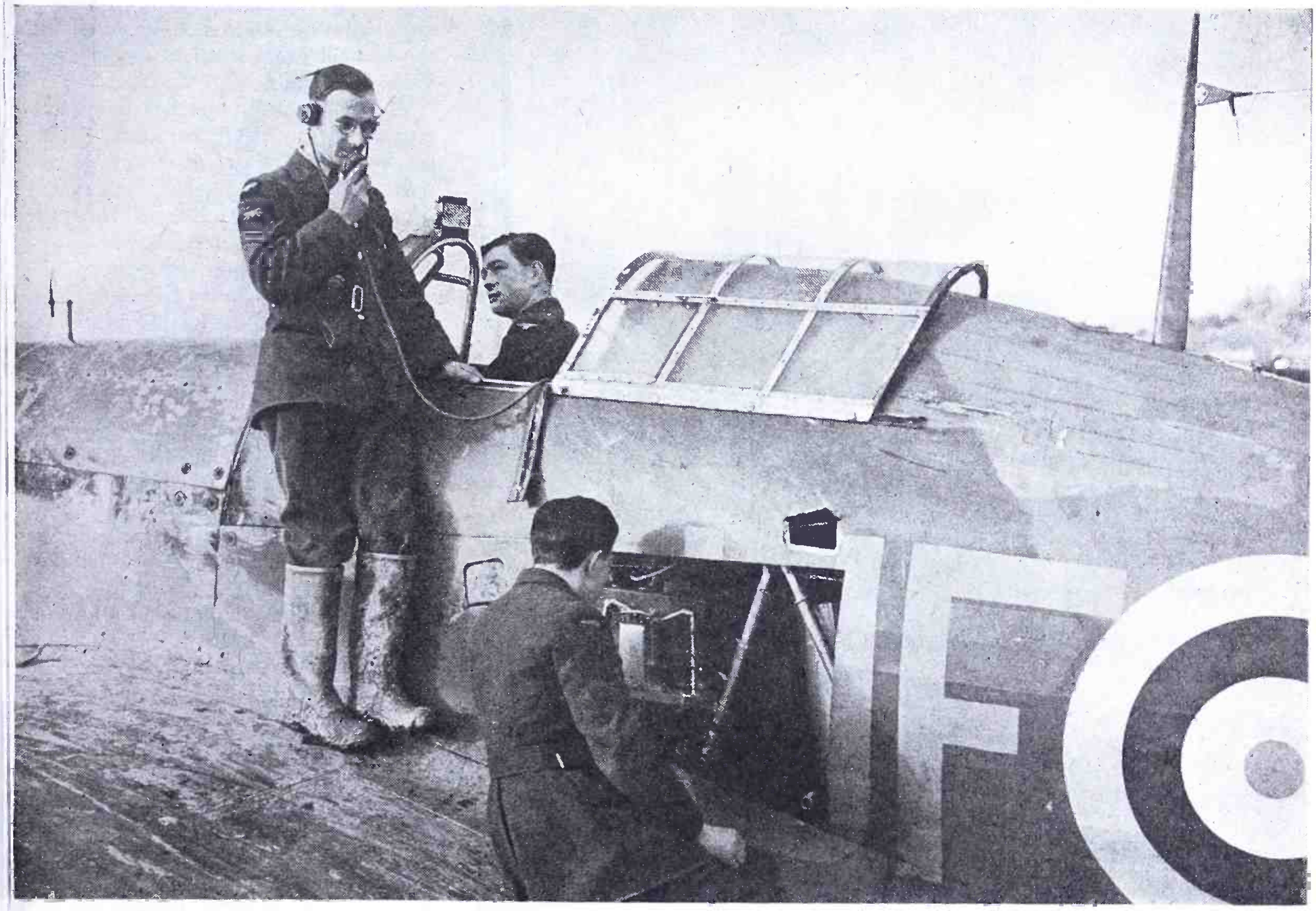
COMMUNICATIONS

LEWIS WINNER, Editor

* * MARCH, 1944 * *



RAF COMMUNICATIONS

by ROY NORRIS

Editor, Electrical Trading and Radio Marketing, London

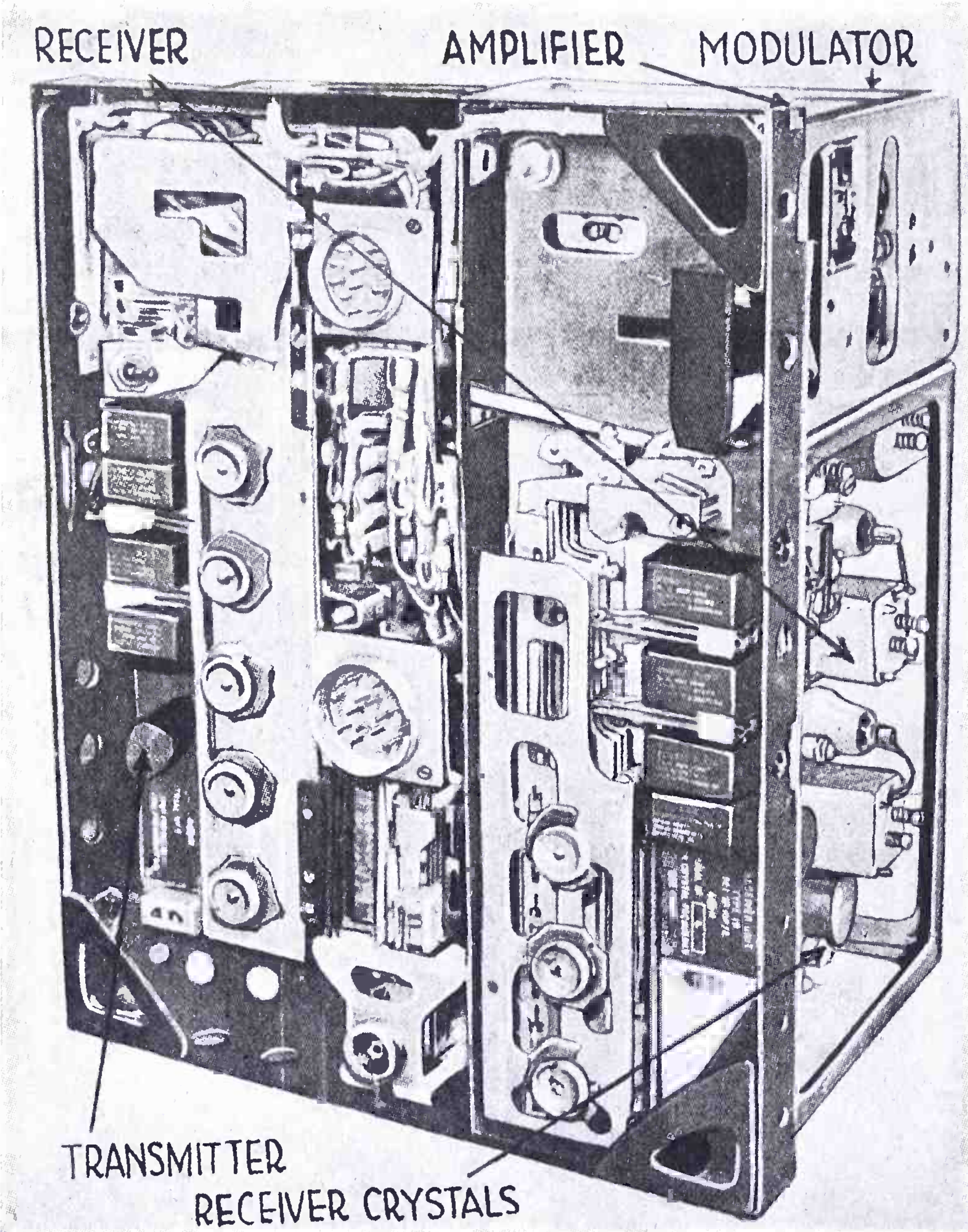
IN Britain the Royal Air Force and Army are two separate organizations. The RAF, therefore, has its own world-wide radio network for administrative and operative communication. Traffic runs in several million words daily and it has to be handled with the greatest reliability and accuracy as a whole campaign may, at some point, hang on a single vital message.

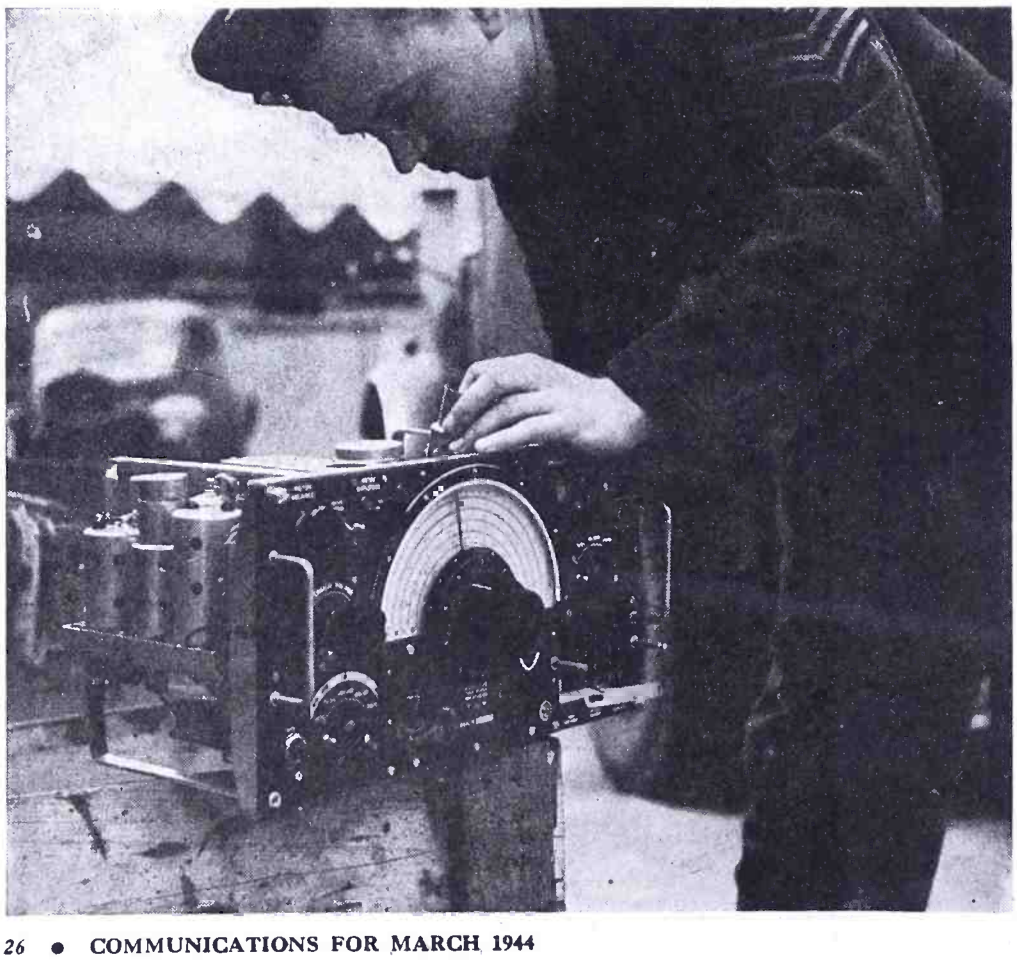
For communications between London headquarters and the Overseas Commands, and between the Commands themselves, a high-speed code system is used, the equipment being basically similar to that of big commercial communication networks.

Figure 1 (above)
Installing communications equipment in a RAF fighter.

All smaller radio gear within this main communication framework is portable—mobility being an outstanding feature of Royal Air Force radio. The *secondary* links between the headquarters of Commands and their Groups and Wings are provided by manually-operated code transmitters, which can be truck-mounted.

This apparatus works mainly in the 3-20 mc band. It is rugged, accessible, simple to operate and is suitable for all extremes of climate. The transmitter is





Figures 2 (left) and 3 (below)
Figure 2, standard v-h-f fighter communications unit. Figure 3, a receiver used in RAF bombers for long distance code reception.

applicable for both crystal and master oscillator drive, and normally runs from Britain's standard supply of 230 volts, 50 cycles, 3-phase. Receivers work from either the power lines or batteries.

Mobile equipments are available for high-speed automatic code transmission powers of 10 and 5 kilowatts. All the airfield radio services can be provided *on wheels*.

A familiar sight in Britain's countryside is the RAF *pack set*. This can be carried short distances by two men and it gives reliable communications of up to 100 miles without the erection of a tuned aerial array.

Apart from medium frequencies, the RAF uses two bands; h-f and v-h-f. Broadly speaking the h-f band is 3 to 8 mc and the v-h-f band is from 100 to 124 mc.

At an airfield, the radio system provides: *(1)*—h-f code transmission on medium power for signalling to aircraft at long range, *(2)* h-f direction-finding station, *(3)*—several h-f channels for instructing aircraft coming in to land, *(4)*—several v-h-f channels for the same duties, *(5)*—a v-h-f direction-finding station, and *(6)*—a beam approach system for assisting aircraft to approach and land in poor visibility.

On the larger type of aircraft the radio installation includes a general purpose code set manipulated by the radio man; a pilot-operated h-f telephony set; a beam-approach receiver used by the pilot; a direction-finding loop operated by the radio man; and an intercom amplifier linking up the crew.

With this equipment an aircraft can communicate by Morse with its base and can talk by phone to other aircraft in the formation, escort fighters or to control-tower when coming in to land. It can *home* on any medium band transmission or obtain a *fix* from the high or medium frequency ground organizations. It can come in and land almost *blind*.

Standardization of sets has been carried a long way to speed production and simplify maintenance. A standard high frequency code set is made in great quantities to use in Britain and overseas for all bomber, general reconnaissance and training aircraft.

For telephony there are two main

(*Continued on page* 75)

Diameter	Temper	K_1	K_2	Maximum recommended frequency in megacycles
1/4"	Soft	1.59	.014	1200
0.286"	Soft	1.23	.012	1000
3/8"	Soft	0.95	.0084	740
5/8"	Soft	0.53	.0095	430
0.643"	Soft	0.51	.0095	430
7/8"	Soft	0.36	.0056	340
7/8"	Hard	0.35	.0024	120
7/8"	Hard	0.37	.0073	450
1 3/8"	Hard	0.21	.0024	70
1 5/8"	Hard	0.18	.0024	60
2 1/8"	Soft	0.15	.0085	155
2 1/8"	Hard	0.13	.0020	48
2 5/8"	Hard	0.11	.0026	47
4 1/8"	Hard	0.07	.0026	47

LOSS IN COAXIAL CABLES AT U-H-F

by DR. VICTOR J. ANDREW

Andrew Company

THESE losses are based on termination of the cable in a pure resistance equal to the characteristic impedance of the cable. Moderate mismatches increase the loss only slightly.

The frequency range on the large diameter cables can be extended by using smaller bead spacings.

Loss in decibels per 1,000 feet = loss in conductors *plus* loss in insulation

$$\text{db}/1{,}000 \text{ ft.} = K_1\sqrt{f} + K_2 f$$

where f = frequency in megacycles.

Example: To determine loss in 350′ of 2⅛" hard temper cable at 46 mc: Loss = 350/1,000 ($.13\sqrt{46}$ + .002 × 46) = 0.34 db = 7.5% of power into line.

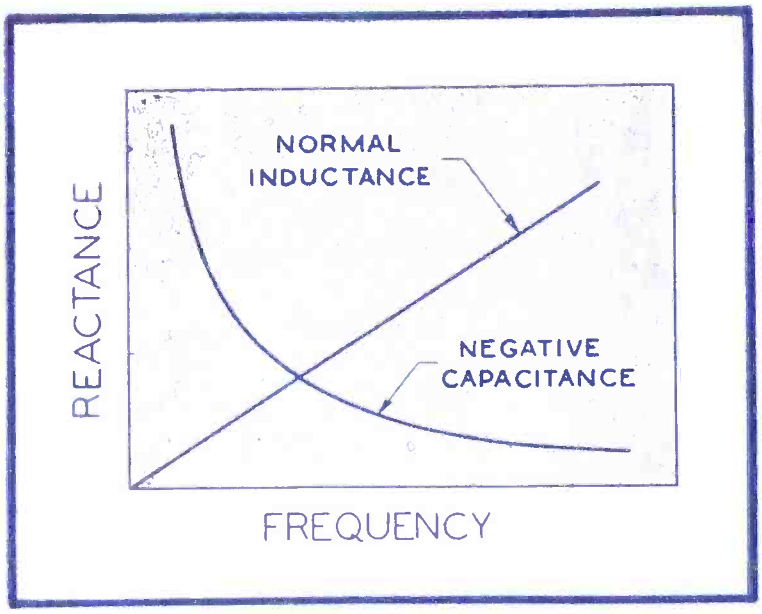


Figure 1
Comparison of negative capacitance with normal inductance.

NEGATIVE

by CLEDO BRUNETTI

Senior Radio Physicist, National Bureau of Standards, Washington, D. C.
Formerly Assistant Professor of Electrical Engineering, Lehigh University

WHEN the output of an ordinary two-stage amplifier is coupled back to the input through a condenser, it is possible, by proper adjustment of the amplification, to secure an input impedance that behaves as a negative capacitance[1]. A negative capacitance is a reactive circuit element whose reactance varies inversely as the frequency, but is positive in sign (Figure 1). At any given frequency a negative capacitance will present the same type of reactance as an inductance. However, if the frequency is increased, the reactance of the inductance will increase in value while that of the negative capacitance will decrease.

Negative Capacitance Uses

A practical negative capacitance will be found useful in improving the frequency response of amplifiers by neutralizing tube input and output capacitances, in improving the Q of tuned circuits by neutralizing shunting capacitances, in extending the range of measurement of Q meters and in many similar applications. Other uses are to

This paper is based on work carried on by the authors at Lehigh University.

be found in special networks such as filters and corrective networks, in transmission line problems[2], and in obtaining impedances proportional to the Nth power of the frequency where N may be any positive or negative integer[3].

A negative capacitance obtained in the manner described above can be made to be substantially independent of amplifier tube and supply voltage variations and of frequency by employing inverse feedback in the amplifier.

In Figure 2 is shown an amplifier with an impedance Zpf connected between input and output terminals, 1 and 3 respectively, of the two-stage amplifier. The equivalent circuit, assuming the impedance of the grid circuit of the first stage of the amplifier to be infinitely large, is shown in Figure 3. The input current is found to be

$$I_i = \frac{E_i - AE_i}{Zpf} \qquad (1)$$

Where A is the amplification and E_i is the input voltage, the input impedance is

$$Z_n = \frac{E_i}{I_i} = Zpf \frac{1}{1 - A} \qquad (2)$$

Equation *2* shows that if A is real (i.e. has no phase shift) and greater than unity, Z_n will be a negative multiple of Zpf. If Zpf is a condenser of value C, equation *2* yields

$$Z_n = \frac{1}{jwC_{pf}(1 - A)} = \frac{1}{jwC_n} \qquad (3)$$

where $C_n = C_{pf}(1 - A)$.

The input impedance is that of a capacitance of value $C_{pf}(1 - A)$ which, for A greater than unity and real in value, is negative.

Inverse Feedback

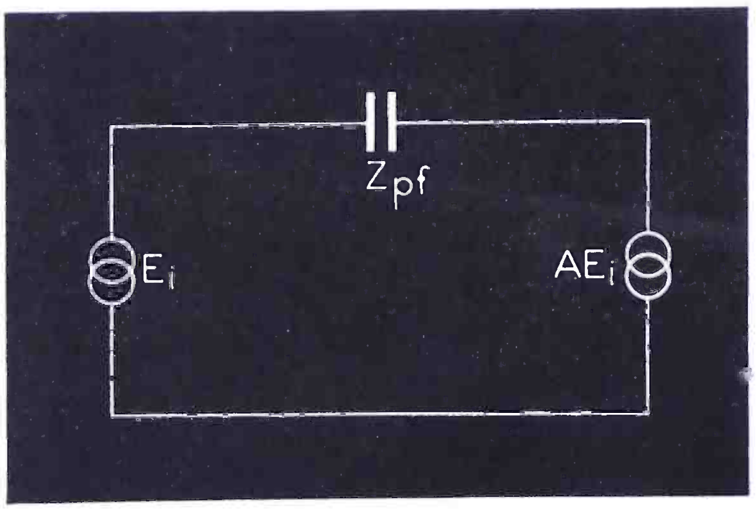
In the equivalent circuit of Figure 3 it was assumed that effective internal impedance in the amplifier output circuit was negligible in comparison with Zpf. When this is not so, the internal

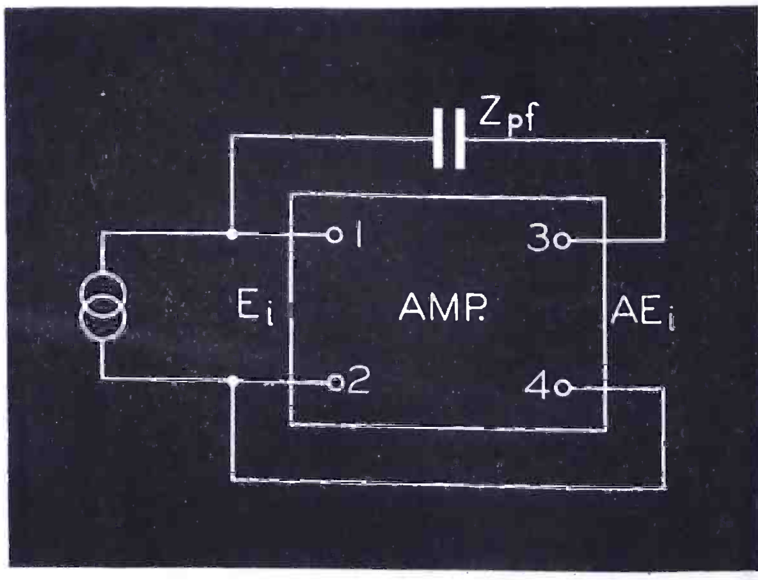
[1] C. Brunetti and L. Greenough, *Some Characteristics of a Stable Negative Resistance*, Proc. IRE, vol. 30, pp. 542-546; Dec., 1942.

[2] H. Mouradian, *Long Distance Transmission Problems*, Jour. Frank. Inst., vol. 207, No. 2; Feb., 1929.

[3] L. C. Verman, *Negative Circuit Constants*, Proc., IRE, Vol. 19, pp. 676-681; April, 1931.

Figure 2 (right) and 3 (below). In Figure 2 appears a method of obtaining negative capacitance. Figure 3 is a circuit equivalent of Figure 2.





CAPACITANCE

and LEIGHTON GREENOUGH

Research Engineer, RCA Mfg. Co., Princeton, N. J.

npedance must be added to Zpf and ntroduces what is often an undesirable egative resistive component in series vith the negative capacitance. Emloyment of inverse feedback helps maerially to reduce the amplifier output mpedance and to yield a negative apacitance with a very low power actor, as will be illustrated later.

Negative Capacitance Circuits

A practical circuit for obtaining negative capacitance is shown in Figure 4. It is a slight modification of a circuit used to obtain negative resistance. Some performance characteristics of this type of circuit have already een reported by the authors[1]. Figure shows the typical relation of negaive capacitive reactance, as a function f frequency, obtained from the circuit f Figure 4. For negative capaciances up to -0.5 mfd, the maximum eviation in capacitance from a fixed alue was 0.2% over the range 250 to ,500 cps.

The method employed to measure he negative capacitance is somewhat he same as that used previously by he authors for other negative elenents.[1] A positive impedance is conected in parallel with the unknown to ring the real part of the impedance of he combination into the positive region where it can be measured by ordinary neans. The technique of carrying out he measurements, however, requires areful attention. A satisfactory circuit or the measurement is shown in Figre 6. It is a modified Maxwell substitution bridge with a Wagner ground nd a headphone detector.

Obtaining Balance

The procedure recommended in obaining a balance is as follows:

(1)—With the negative capacitance in the circuit, the bridge is first balanced by varying only R_p and C_p. Rough adjustment is provided by L_s and R_1 or R_2. The Wagner ground is then balanced. C_{p1} and R_{p1} are recorded. The subscript *1* indicates the first reading.

(2)—The negative capacitance is removed from the circuit and the bridge again balanced. C_{p2} and R_{p2} are read.

The equations for finding C_e and R_e, the equivalent parallel negative capacitance and resistance respectively, are

$$C_e = -(C_{p1} - C_{p2})$$

$$R_e = \frac{-R_{p1} R_{p2}}{R_{p2} - R_{p1}}$$

The equivalent series negative capacitance is given by

$$C_s = C_e \frac{1 + R_e^2 w^2 C_e^2}{R_e^2 w^2 C_e^2}$$

$$R_s = \frac{R_e}{1 + R_e^2 w^2 C_e^2}$$

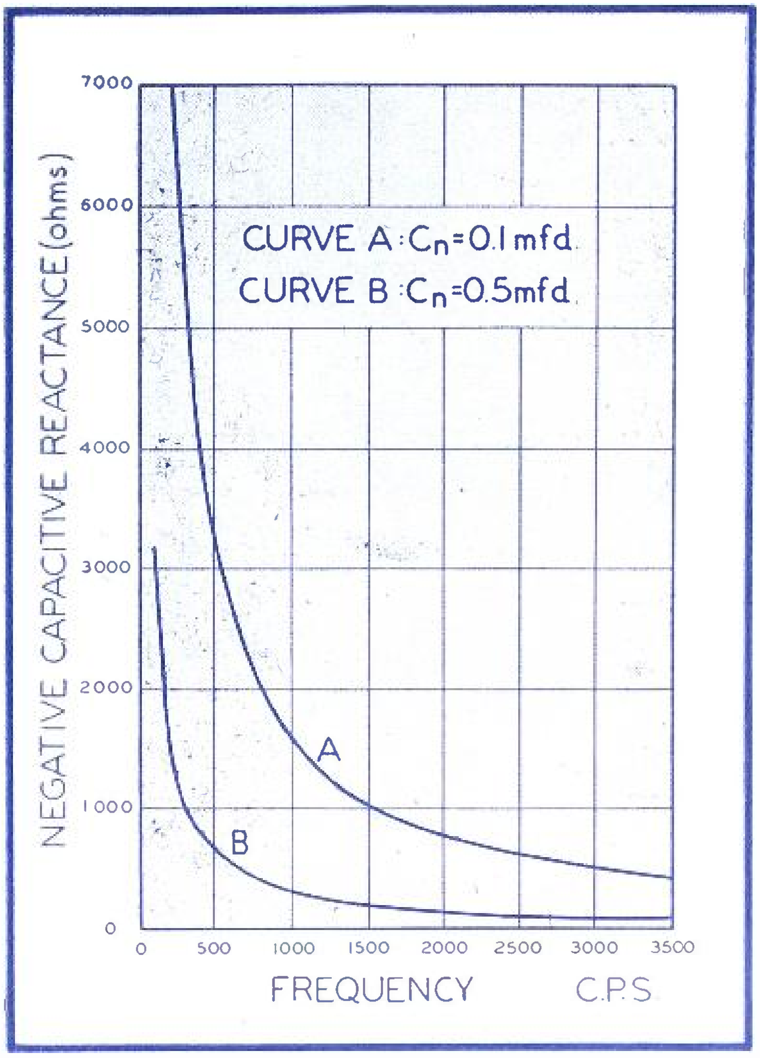


Figure 5

Negative capacitive reactance as a function of frequency.

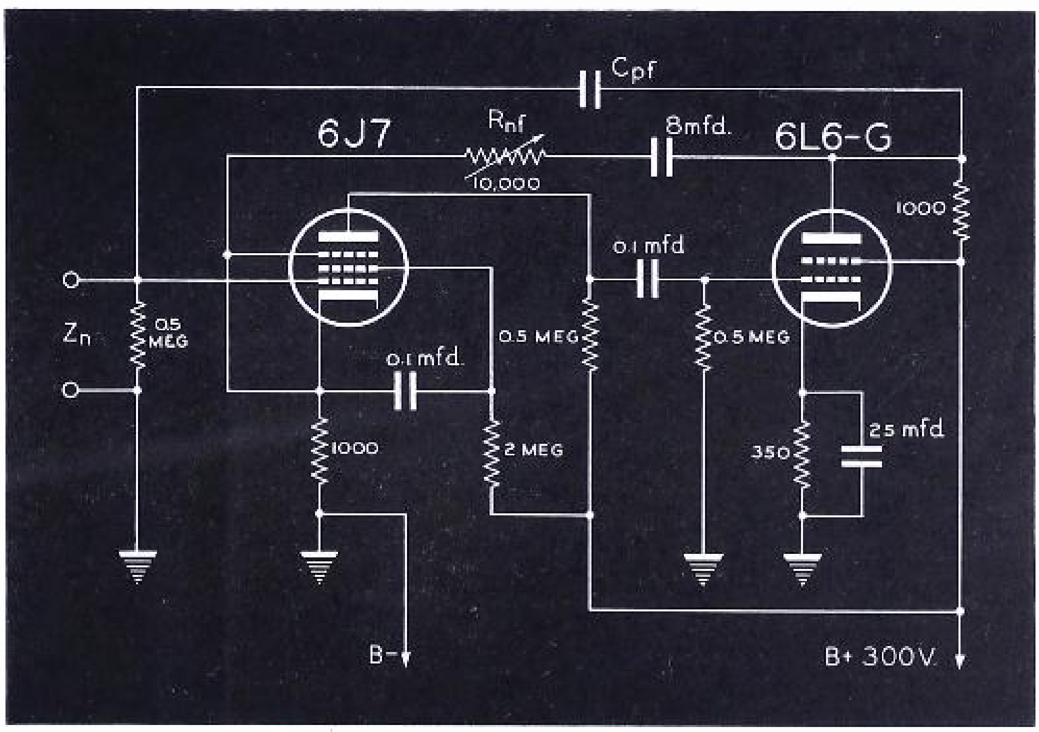


Figure 4

Circuit for obtaining negative capacitance.

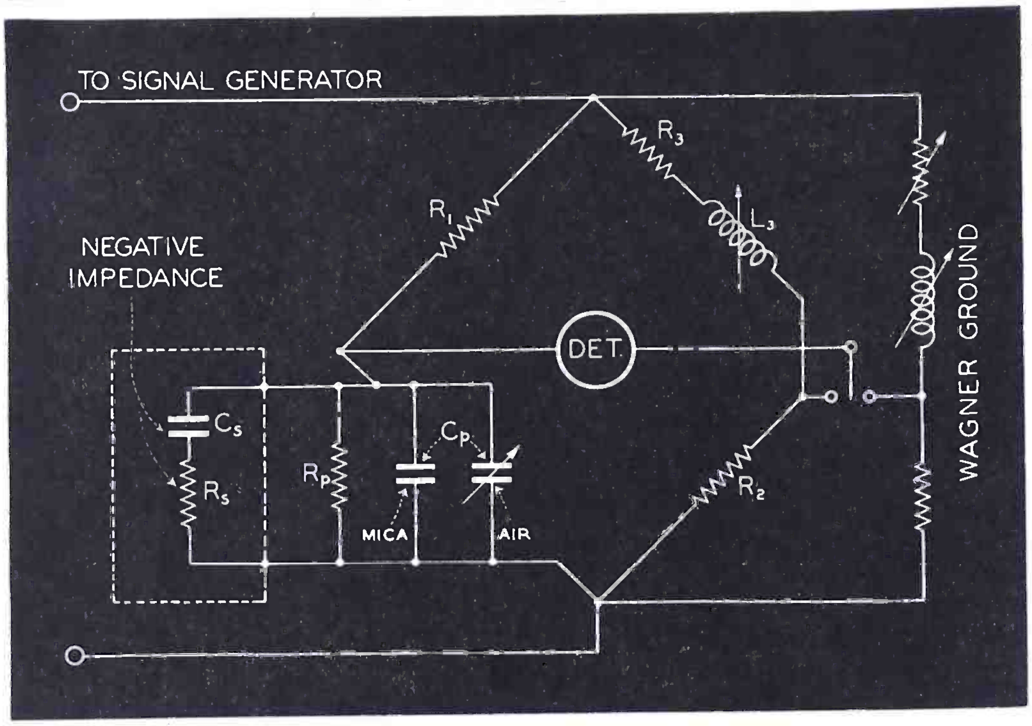


Figure 6 (above)

Bridge used to measure negative impedance. This is a modified Maxwell substitution bridge with a Wagner ground and a headphone detector. In operation a positive impedance is connected with the unknown to bring the real part of the impedance of the combination into the positive region where it can be measured by standard means.

This method was found to be the most satisfactory of a large number tried. One of its chief advantages is that the difference quantity (R_{p2} — R_{p1}) does not depend upon two calculated values based on measurements of other components in the bridge. Such a difference would be subject to considerable error if slight inaccuracies were to occur in the calibration of R_1 and R_2. In the above method R_p may first be set at R_{p1} and the variation in this quantity provided by means of another low range decade resistance box in series with R_p. Such a step is necessary because a setting of as close as 0.2 ohm in 5,000 may be required

Range of Negative Capacitance

Figure 7 shows the range of negative capacitance obtained from the circuit of Figure 4 for various values of C_{pf} as the amount of inverse feedback was varied by means of R_{nf}. These values were measured at a frequency of 1,000 cps. The relation between R_{nf} and the negative capacitance is seen to be linear. Figure 8 shows the relationship between amplification *A* and inverse feedback resistance R_{nf}. As the amplification is, within reasonable limits, a linear function of R_{nf} over the range shown in the Figure, the linearity of the graphs of Figure 7 would be expected.

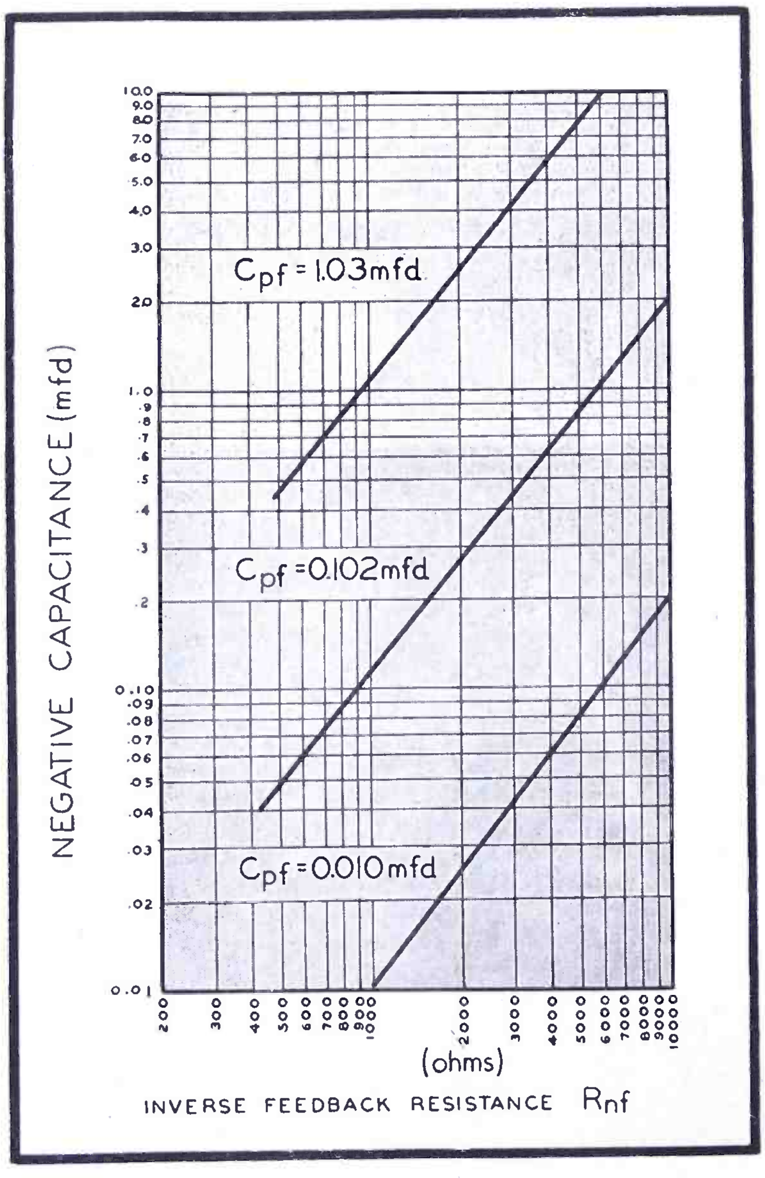


Figure 7

Negative capacitance as a function of the inverse feedback resistance R_{nf}. The range of negative capacitance shown here was obtained from the circuit of Figure 4. These values were measured at 1000 cps. It will be noted that the relation between R_{nf} and the negative capacitance is linear.

Power Factor Considerations

In general the impedance is not a pure negative capacitance but may contain a small negative resistance component. The power factor is therefore negative. If a value of C_{pf} of 0.01 mfd or less is used, the absolute magnitude of the power factor is less than 0.5%; if C_{pf} is in the range 0.01 mfd to 0.10 mfd, the negative power factor increases to approximately 1%. For larger values of C_{pf} the absolute magnitude of the power factor increases somewhat proportionally; also the losses in C_{pf} contribute materially to the power factor. If paper capacitors are used, care should be exercised to select those with low losses. The use of mica capacitors is recommended.

Negative Capacitance Stability

The negative capacitance is practically independent of small variations in supply voltage. The stability in this respect is comparable to that of the negative resistance obtainable from the same amplifier [1]

Applications

The negative capacitance describe

erein has proven useful in extending he frequency response range of amplifiers by a factor of as much as 5 to 0 times. As pointed out earlier, this erformance is obtained simply by hunting the input, interstage or output capacitance of the amplifier with a egative capacitance.

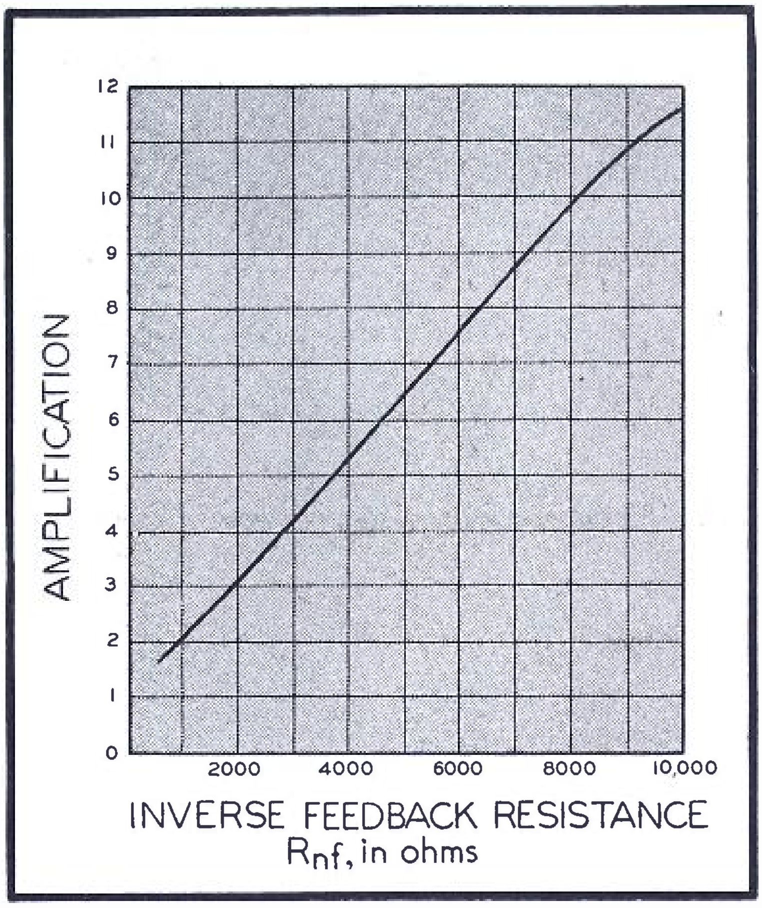


Figure 8

The relationship between amplification and the inverse feedback resistance is shown in this graph. We note that the amplification is within reasonable limits a linear function of R_{nf}.

ses in V-T Voltmeters

Another useful application is in reucing the input capacitance of vacum tube voltmeters or other instruents fed by long cables with high apacity to ground. Insertion of the egative capacitance at the input or utput terminals of the cable makes it ossible to neutralize the shunting efect of the cable and raise its input npedance at high frequencies appreiably.

-F Transformer Application

The high-frequency response of udio-frequency transformers is limted by the distributed capacity of the vindings. This response may be improved materially by means of negative capacitance. For example, a negaive capacitance connected across the econdary of a transformer, whose ormal output voltage with constant nput was down to 85% of maximum t 2,000 cps, increased the range of peration for the same output voltage erformance to 12,000 cps.

pplications in Q Meters

Q meters, as a rule, are limited in ange of measurement by the residual apacitance of the instrument. Figure shows a typical *Q*-meter circuit. The nknown inductance L_x is measured in erms of the magnitude of the standard apacitance C_s required to seriesesonate it. The minimum value of C_s obtainable is limited by the residual apacitance of the instrument. Introuction of a negative capacitance in arallel with C_s, as shown by the otted lines, removes this limitation nd extends the range of measureient.

Brief consideration of the possibilies will lead to other applications of imilar or varied nature.

In using negative capacitance to ancel or neutralize the positive capaciince of a circuit, care should be exrcised to see that too much cancellaon is not attempted. As mentioned bove, the source of negative capaciince may contain a negative resisince component and complete concelion of the capacitance across the outut terminals will present a circuit in hich oscillations may be sustained. hese unwanted oscillations may make the use of the negative capacitance impractical unless they should occur at frequencies well beyond the working range for which the device is intended and provided they do not produce undesirable secondary effects.

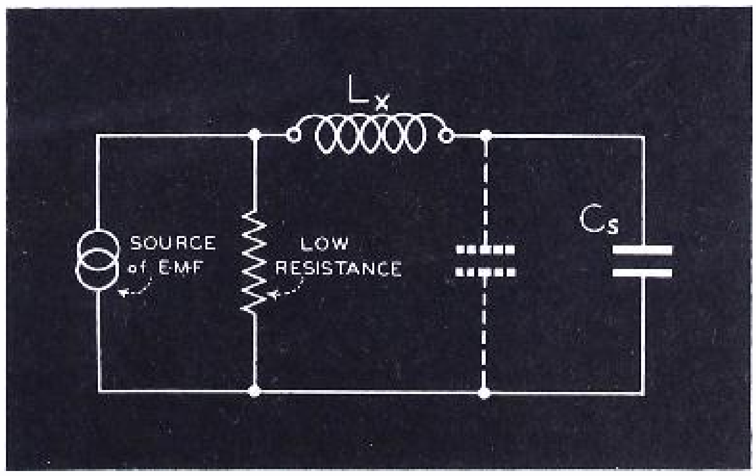
Cancellation of the capacity across a tuned parallel circuit was attempted by connecting a negative capacitance across the terminals. The tuned circuit consisted of a coil having an inductance of 0.153 henry and an 0.09-microfarad capacitor. The normal frequency of resonance of this circuit was approximately 1400 cycles per second. By means of the negative capacitance it was possible to raise the resonance frequency to 10,000 cycles per second without introducing oscillations. The resulting equivalent capacitance was approximately 1.66×10^{-3} microfarad. This represents a reduction of fifty-four to one in capacitance. Reductions of ten to one may be obtained with very stable performance.

The major factor contributing to the input capacitance of a triode amplifier usually is the grid-plate capacitance. A large current flows through this capacitance as the latter appears at the input terminals magnified by an amount $1 + A \cos \phi$ where A is the voltage amplification and ϕ is the phase angle of the load impedance in the plate circuit. For large amplification this capacitance is approximately proportional to the amplification and is much greater than the interelectrode capacities themselves. With screen-grid and pentode tubes the grid-plate capacitance may be negligible and the input capacitance is principally the sum of the grid-screen and grid-cathode capacitances.

Capacitance neutralization in amplifiers may be obtained in a variety of methods other than with the use of negative capacitance. Typical of such methods are the Hazeltine neutrodyne and Rice systems of neutralization, the inverted neutrodyne and the bridge circuits of Ballantine and Hull.

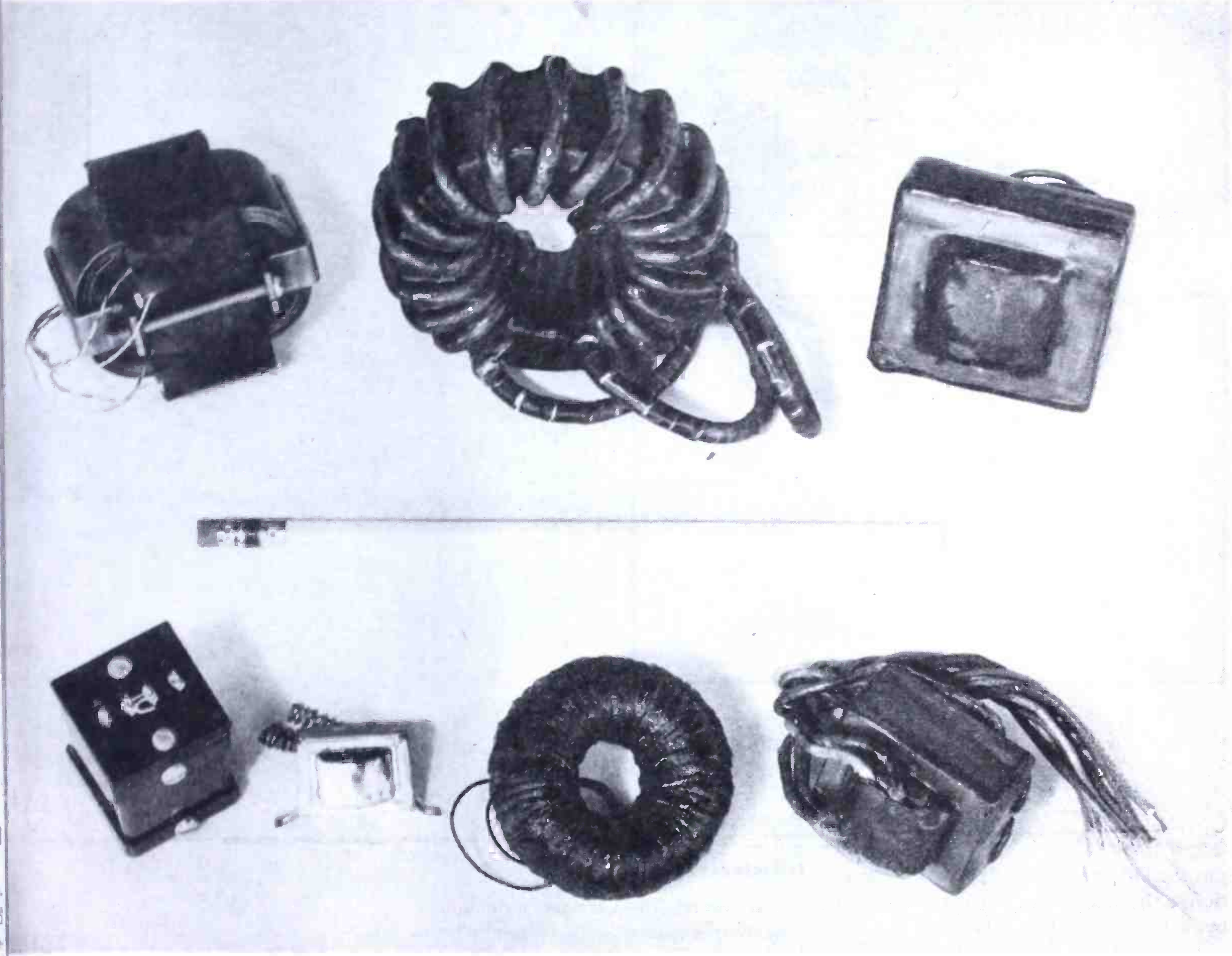
Figure 9

A typical *Q*-meter circuit wherein the unknown inductance L_x is measured in terms of the magnitude of the standard capacitance C_s required to series-resonate it.



HIGH-Q AUDIO REACTOR

Design and Production



Top row, at extreme left, a hipersil core: $f_{(max\,Q)}$ = 1400 cycles, L = 2 henrys, Q_{max} = 80; used for tank circuit, 400-cycle oscillator. Center, top, a powdered core: $f_{(max\,Q)}$ = 10,000 cycles, L = 33 microhenrys, Q_{max} = 40; used to damp a light valve. Right, top, an E-I core: $f_{(max\,Q)}$ = 1000 cycles, L = 3.3 millihenrys, Q_{max} = 35; used in dividing network for two-way horn system. In the bottom row we have, at extreme left, a dust core: $f_{(max\,Q)}$ = 10,000 cycles, L = 2 henrys, Q_{max} = 12; used in 10-kc filter of radio tuner. Center, at left in bottom row, is an F core: $f_{(max\,Q)}$ = 1000 cycles, L = 16 henrys, Q_{max} = 16; for equalizing low level sound circuits. At right center, bottom, is a powdered core: $f_{(max\,Q)}$ = very broad, L = 100 millihenrys, Q_{max} = 22; used where constant Q is necessary. At extreme right, bottom, we have an E-I core: $f_{(max\,Q)}$ = 50, L = 880 henrys with taps; Q_{max} = 12; used in geophysical circuits.

by COLIN A. CAMPBELL

Plant Engineer

Altec-Lansing Corporation

THE introduction of a magnetic core into a coil is not as innocent as it might seem for it exacts a certain toll or compensation for the benefits derived. The purpose of this paper is to, first, analyze the effect of the magnetic core, and second, show commercial practice of inductance manufacture.

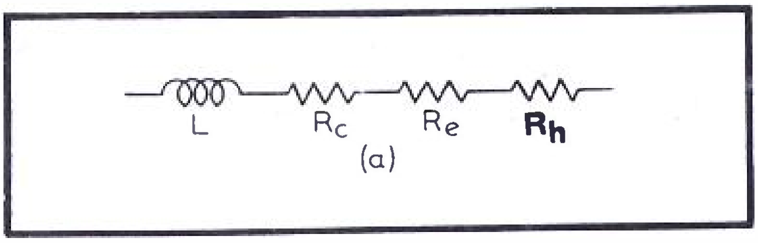
A commercial power transformer might be considered as a special case of an *inductor* or *reactor*. The heat dissipated in the core is quite evident; this is commonly called *core loss*. This loss has two components called hysteresis loss and eddy current loss. Hysteresis loss is determined by the chemical and physical structure of the core itself. The loss is usually expressed by the formula

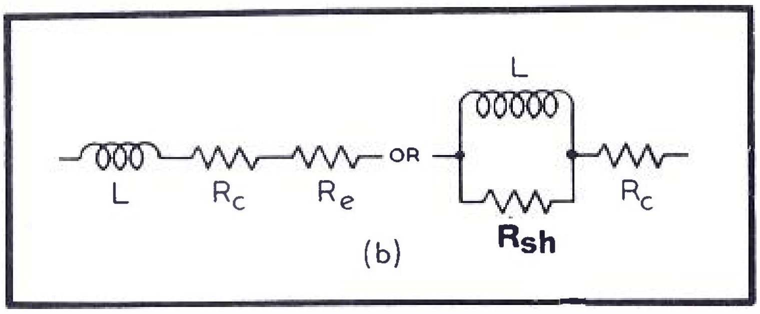
$$\text{Watts loss} = k\,f\,(\beta_{max})^n, \text{ where}$$

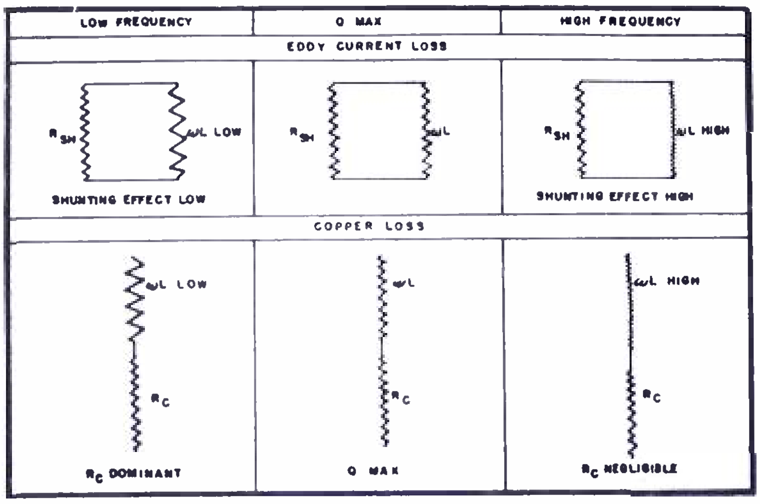
k = a constant for the particular material.
f = frequency in cycles per second.
β_{max} = maximum instantaneous flux density.
n = Steinmetz exponent, usually slightly less than 2.

This formula may be interpreted in the following way. If the flux density is kept constant, the loss will vary in proportion to the frequency, or if the terminal voltage is kept constant and the frequency is varied, the loss will vary approximately inversely to the frequency.

The other principal core loss is due







Figures 1a (upper left), 1b (center left), and 1c (lower left).
In Figure 1a appears a circuit which represents the losses in an inductance. In Figure 1b, R_c is the winding resistance and R_e the eddy current loss resistance. In Figure 1c we have the equivalent circuits that represent loss due to R_c and R_{sh}.

to eddy currents and is expressed as

$$\text{Watts loss} = k_2 f^2 (\beta_{max})^2 t^2$$

Obviously, the loss is independent of frequency. The symbol *t* represents the thickness of the individual laminations; the loss is very greatly reduced by using thin laminations.

In considering an inductance for audio frequencies, it is confusing to think of *loss* in terms of *watts*—rather it is more convenient to visualize a resistor as representing a loss.

In Figure 1a appears a circuit which represents the losses in an inductance. In this circuit L = pure inductance; R_c = winding resistance (ohmic), that is, copper loss not affected by frequency; R_e = eddy current loss; R_h = hysteresis loss.

In order to simplify the analysis the voltage applied to the inductance will be assumed to be very low so that the hysteresis loss will be low compared with the eddy current loss. With this restriction the circuit may be represented as follows:

In Figure 1b, R_c is the winding resistance and R_e the eddy current loss resistance, which should be thought of as being in *shunt* with the resistance. That is, we should think of a transformer which has a secondary winding terminated with a resistance. Resistance is independent of frequency, but its effect is not independent of frequency.

Inductance Qualities

The merit of various inductances may be compared on a common basis known as *Q*.

$$Q = \frac{2 \pi f L}{R}$$

For the purpose of circuit analysis the shunt resistance, R_{sh}, may be equated to a series resistance by the following manipulation. The impedance of the inductance and resistances in parallel, may be expressed algebraically as

$$Z = \frac{j R_e \omega L}{R_e + j \omega L}$$

Conforming to the general rule of parallel impedances

$$(1)\quad Z_o = \frac{Z_1 Z_2}{Z_1 + Z'_2} = \frac{R_e j \omega L}{R_e + j \omega L}$$

(2) Multiplying each term by

$$R_e - j \omega L$$

$$(3)\quad Z = \frac{R_e j \omega L (R_e - j \omega L)}{R_e + j \omega L (R_e - j \omega L)} = \frac{R_e (\omega L)^2 + j (R_e)^2 \omega L}{(R_e)^2 + (\omega L)^2}$$

(4) Separating real and imaginary terms

$$\frac{R_e (\omega L)^2}{(R_e)^2 (\omega L)^2} + \frac{j (R_e)^2 \omega L}{(R_e)^2 + (\omega L)^2}$$

(5) And considering only the real term

$$\frac{R_e (\omega L)^2}{(R_e)^2 + (\omega L)^2}$$

(6) Then factoring the denominator

$$\frac{R_e (\omega L)^2}{(R_e)^2 \left[1 + \left(\frac{\omega L}{R_e}\right)\right]}$$

(7) If $\frac{\omega L}{R_e}$ is small compared with 1

the expression becomes $\frac{(\omega L)^2}{R_e}$ which

being a series resistance may be added directly to R_c making the total loss

$$R = R_c + \frac{(\omega L)^2}{R_e}$$

which gives for *Q*

$$Q = \frac{\omega L}{R}$$

R = total resistance of circuit

Simplifying

$$Q = \frac{\omega L}{R_c + \frac{(\omega L)^2}{R_e}} = \frac{1}{\frac{R_c}{\omega L} + \frac{\omega L}{R_e}}$$

$$= \frac{1}{\frac{R_e R_c + (\omega L)^2}{R_e \omega L}} = \frac{R_e \omega L}{R_e R_c + (\omega L}$$

Now, to find when *Q* is a maximu[m] we have only to differentiate *Q* wi[th] respect to f.

R_c, R_e and $2 \pi f$ are to be co[n]sidered constants. Applying t[he] formula for the differentiation of [a] quotient: let $a = R_c$
$b = R_e$
$c = 2 \pi f$

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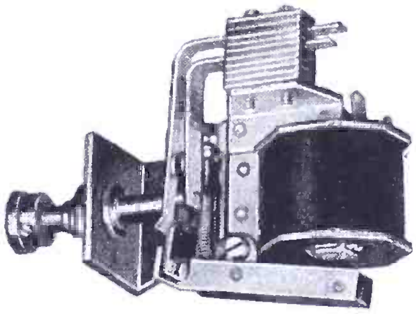

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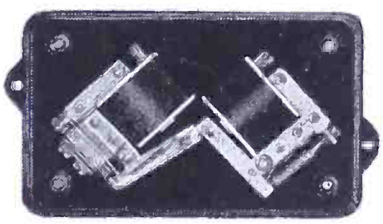

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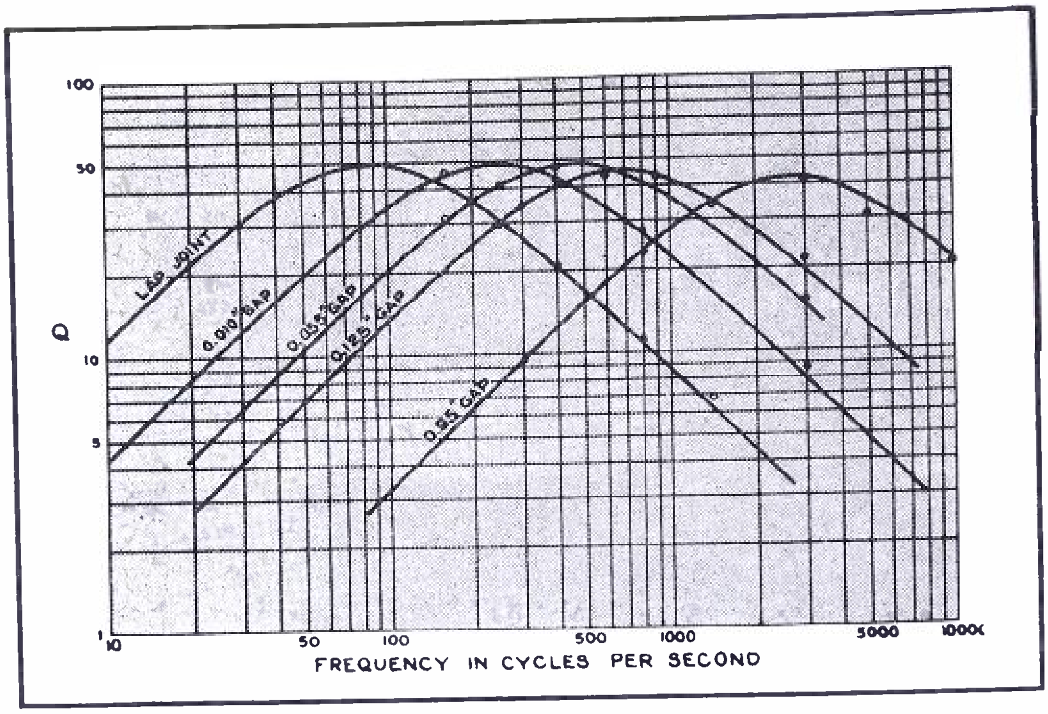


Figure 2

A graphical explanation of how and why Q changes with frequency, and why it goes through a maximum.

(Courtesy General Radio Experimenter)

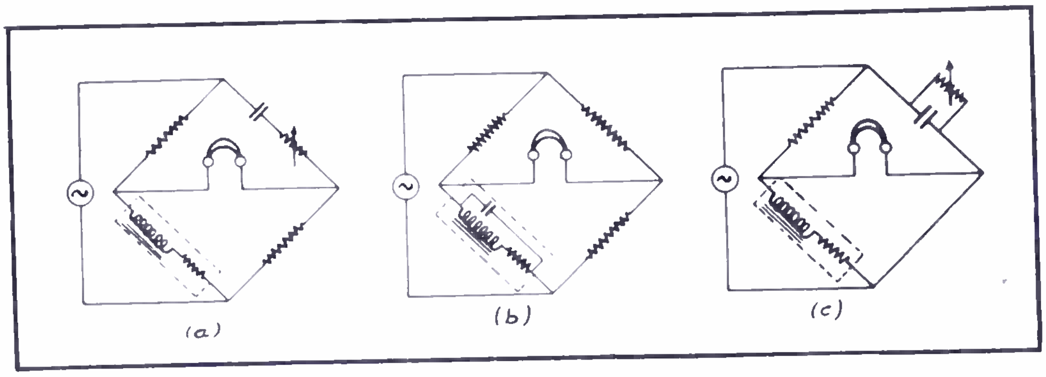


Figure 3

Bridge circuits for measurements of Q. At (a) appears a Hay bridge; at (b), a resistance bridge operated at resonant frequency of tuned circuit; at (c), Maxwell bridge.

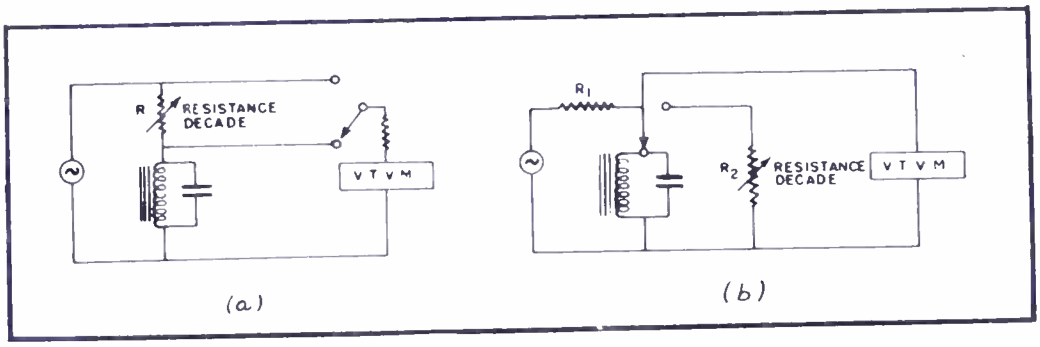


Figure 4

Methods of measuring Q. At (a) is a 6-db drop method, and at (b) appears a resistance substitution method. In (a), $Q = \omega L/R$ when 6-db voltage difference exists across switch. In (b), R_1 is selected so that small changes in Z of tuned circuit can be observed on V.T.V.M.

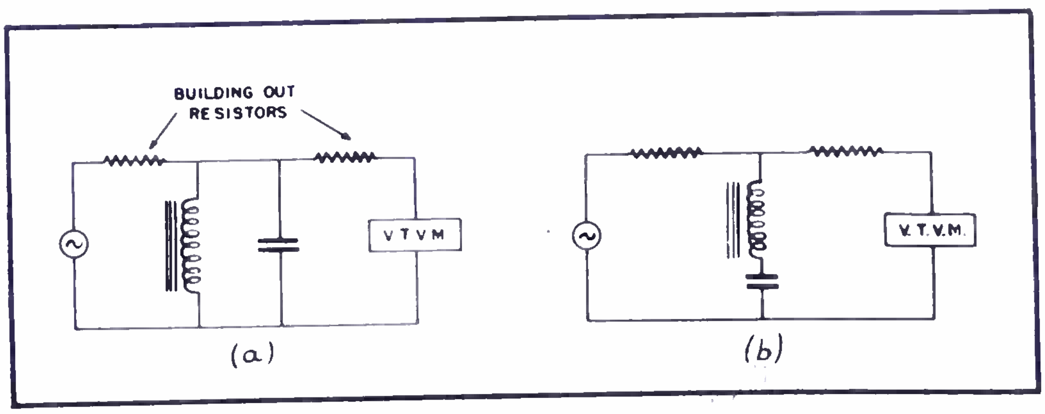


Figure 5

In (a) and (b) appear frequency variation methods. Similar methods can be used with series-resonant circuits.

$$\frac{dQ}{df} = \frac{[ab + (cf)^2]bc - [0 + 2ef] \times c^2bf}{(ab + c^2f^2)^2}$$

$$= \frac{ab^2c + c^3bf^2 - 2c^3f^2b}{(ab + c^2f^2)^2} = \frac{ab^2c - bf^2c^3}{(ab + c^2f^2)^2}$$

$$= \frac{cb\,(ab - f^2c^2)}{ab + c^2f^2}$$

(each term can be set equal to zero)

$$ab - f^2c^2 = 0$$

$$f^2 = \frac{ab}{c^2} \qquad f = \frac{\sqrt{ab}}{c} = \frac{\sqrt{R_e R_c}}{2\pi L}$$

Similar equations are solved to give the following expressions

$$Q_{max} = \frac{\pi f_{(of\ max\ Q)} L}{R_c}$$

$$Q_{max} = \tfrac{1}{2}\sqrt{\frac{R_e}{R_c}}$$

A graphical analysis is helpful in showing how and why Q changes with frequency and why it goes through a maximum (Figure 2).

A typical air core coil has only copper loss, so the Q is expressed $Q = \frac{\omega L}{R_c}$. A graph of Q plotted against frequency is a straight line starting at the origin, with a positive slope of *1*. Now, the effect of a shunt resistance is just the opposite; a line with a slope of *—1*. The frequency of intersection of these two lines is the frequency of maximum Q. The sum of the two lines at any frequency is the Q for that frequency. (The sum in this case is not the arithmetical sum but rather the reciprocal of the sum of the reciprocals; think of *D*, dissipation factor!)

An air core coil of *1* henry with 50 ohms of resistance would have the following Q:

100 cycles	1.256
1,000 cycles	12.56
10,000 cycles	125.6

Now let us visualize a shunt resistance of 80,000 ohms to represent eddy current losses. ωL at 100 cycles is 628 ohms, so 80,000 ohms in shunt would have a very small effect. At 10 kc, ωL = 62,800, so 80,000 ohms in shunt would have a pronounced effect. The frequency at which Q is a maximum is

$$f_{max} = \frac{\sqrt{R_e R_c}}{2\pi L}$$

$$= \frac{\sqrt{500 \times 80{,}000}}{(6.28)\ (1)} = 1{,}000 \text{ cycles}$$

The actual Q would be

$$Q_{max} = \tfrac{1}{2}\sqrt{\frac{80{,}000}{500}}$$

$$= \tfrac{1}{2}\sqrt{160} = \frac{12.6}{2} = 6.3$$

In all the above expressions L is varied by adjusting the air gap of the magnetic structure.

For the foregoing discussion the flux density has been considered as being very small, so small that the hysteresis loss is negligible. The question might be raised: "What good is such an analysis since it does not represent working conditions?" Since the Steinmetz exponent is somewhat less than 2, it follows that hysteresis loss is almost constant with frequency, decreasing slowly as the frequency is increased, provided a constant voltage is maintained at the inductance terminals. As the voltage is increased the hysteresis loss increases very rapidly, masking, at high voltage, the effect of R_c and R_e in determining the frequency of Q_{max}. However, since this loss resistance varies so slowly with frequency, it may be considered as being independent of frequency. The most important thing about the analysis is that it provides the engineer with explanatory data.

In order that an inductance have high Q the three losses, R_c, R_e and R_h must be low.

R_c is kept low by using a magnetic structure that has a large winding space and by filling this space as nearly as possible with copper wire.

Eddy current loss may be kept low by: (a) using thin laminations; (b) using laminations which have a high specific resistance; (c) keeping the leakage flux from metallic objects, especially iron; (d) preventing eddy currents from flowing from sheet to sheet of the laminations, and (e) operating the core at low flux density.

R_h may be kept low by operating at low flux density and by using low loss core material.

Figure 6

The sharpness of resonance of a circuit indicated in Figure 5 is divided by the change in frequency necessary to produce a certain change in current in the tuned circuit, as we see in this table.

(Courtesy Terman's Radio Engineers Handbook, McGraw-Hill Book Co.)

FRACTIONAL DETUNING (= CYCLES OFF RESISTANCE / RESONANT FREQUENCY)	ACTUAL CURRENT / CURRENT AT RESONANCE
$\frac{1}{6Q} = \frac{1}{3} \times \frac{1}{2Q}$	0.95
$\frac{1}{4Q} = \frac{1}{2} \times \frac{1}{2Q}$	0.90
$\frac{1}{2Q} = 1 \times \frac{1}{2Q}$	0.707
$\frac{1}{Q} = 2 \times \frac{1}{2Q}$	0.447
$\frac{2}{Q} = 4 \times \frac{1}{2Q}$	0.242
$\frac{4}{Q} = 8 \times \frac{1}{2Q}$	0.124

How to Measure Q

The simplest methods of measuring Q is to use an impedance bridge such as a GR 650A. After the balance is obtained, the Q may be read directly from the DQ or Q dial. Bridges may be made from laboratory parts. The Hay bridge is most convenient if the inductance under test has a high Q;

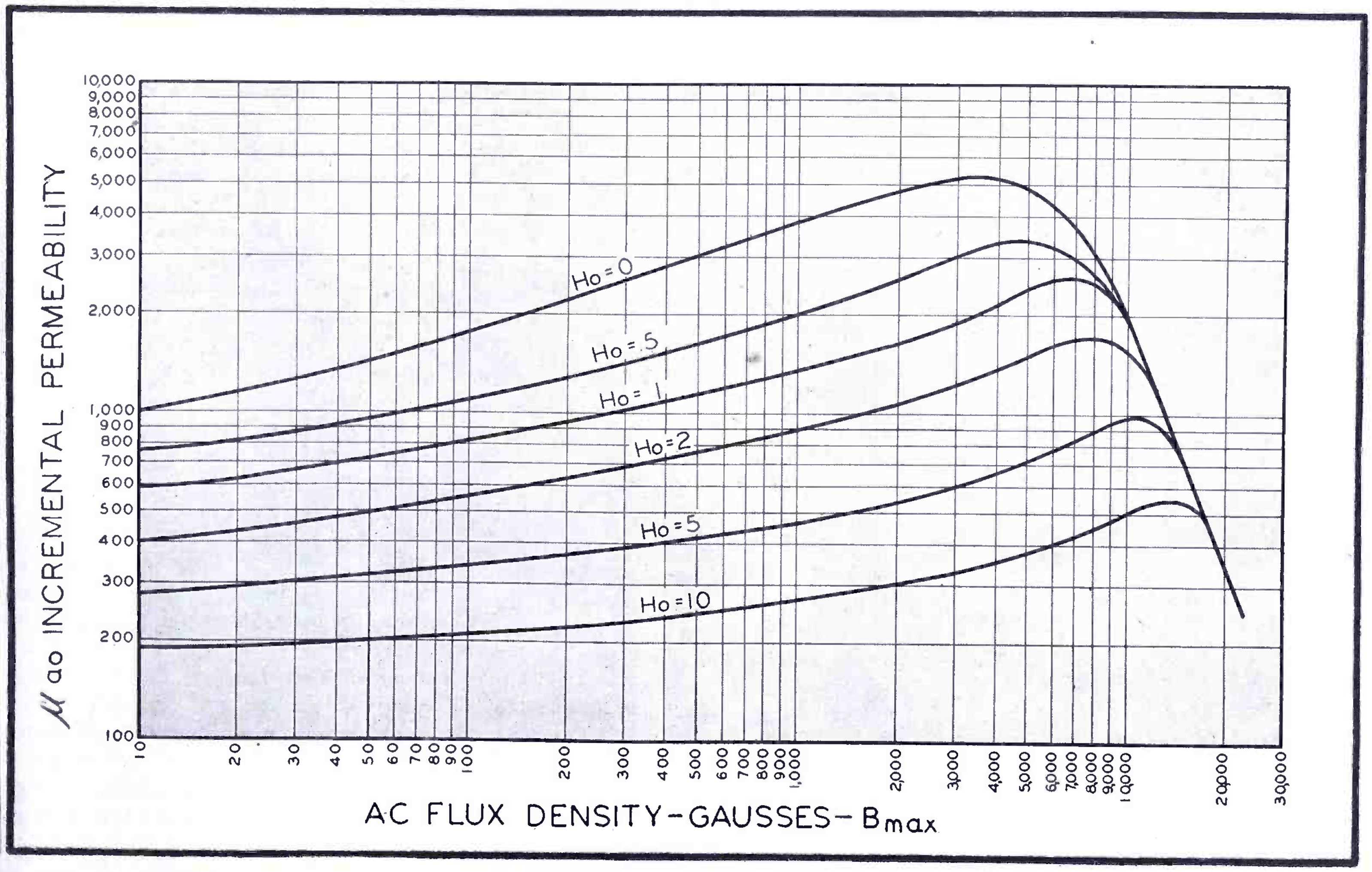


Figure 7

60-cycle incremental permeability characteristics of Allegheny audio transformer "A" sheet steel (No. 29 U. S. Gauge). *(Courtesy Allegheny Steel Company)*

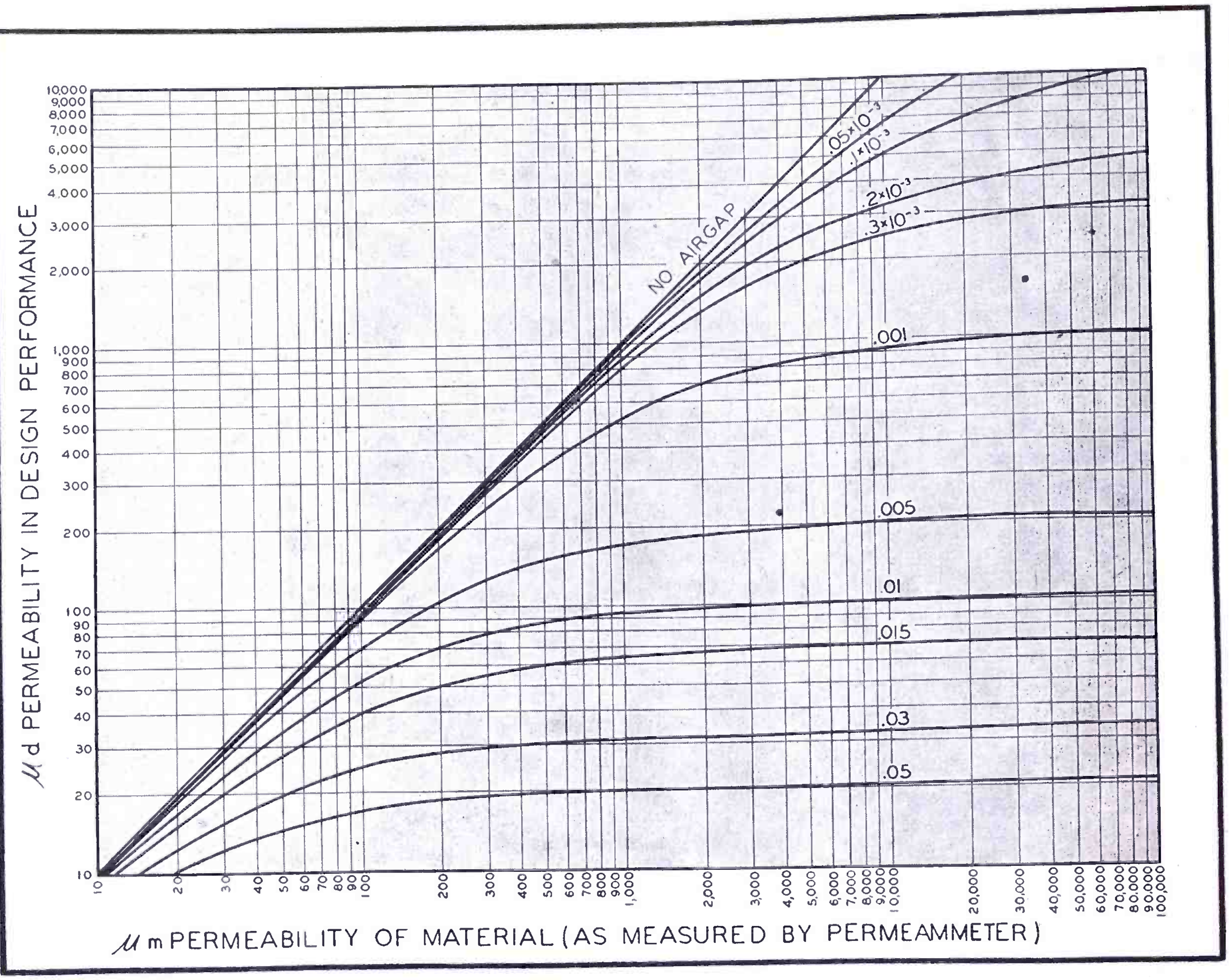


Figure 8

Permeability of design (μ_d) resulting from known permeability values (μ_m), corresponding to various ratios of airgap to total length of magnetic path (a/l).

the Maxwell bridge is most convenient for low Q.

A different method is to tune the inductance to resonance by use of a suitable condenser, and then measure the impedance of the tuned circuit by any of several different methods, (Figures 3 and 4).

For parallel resonance the impedance (for Q = 10 or more) is $Z = \omega L \times Q$.

$$Z = R = \omega LQ$$

$$Q = \frac{R}{\omega L}$$

The formula in Figure 4 is not correct for parallel resonance, but is correct for series resonance.

A still different method is the *frequency variation method*, (Figure 5). Here the sharpness of resonance is indicated by the change in frequency necessary to produce a certain change in current in the tuned circuit. (Figure 6).

Manufacturing Practices for High Q Inductances

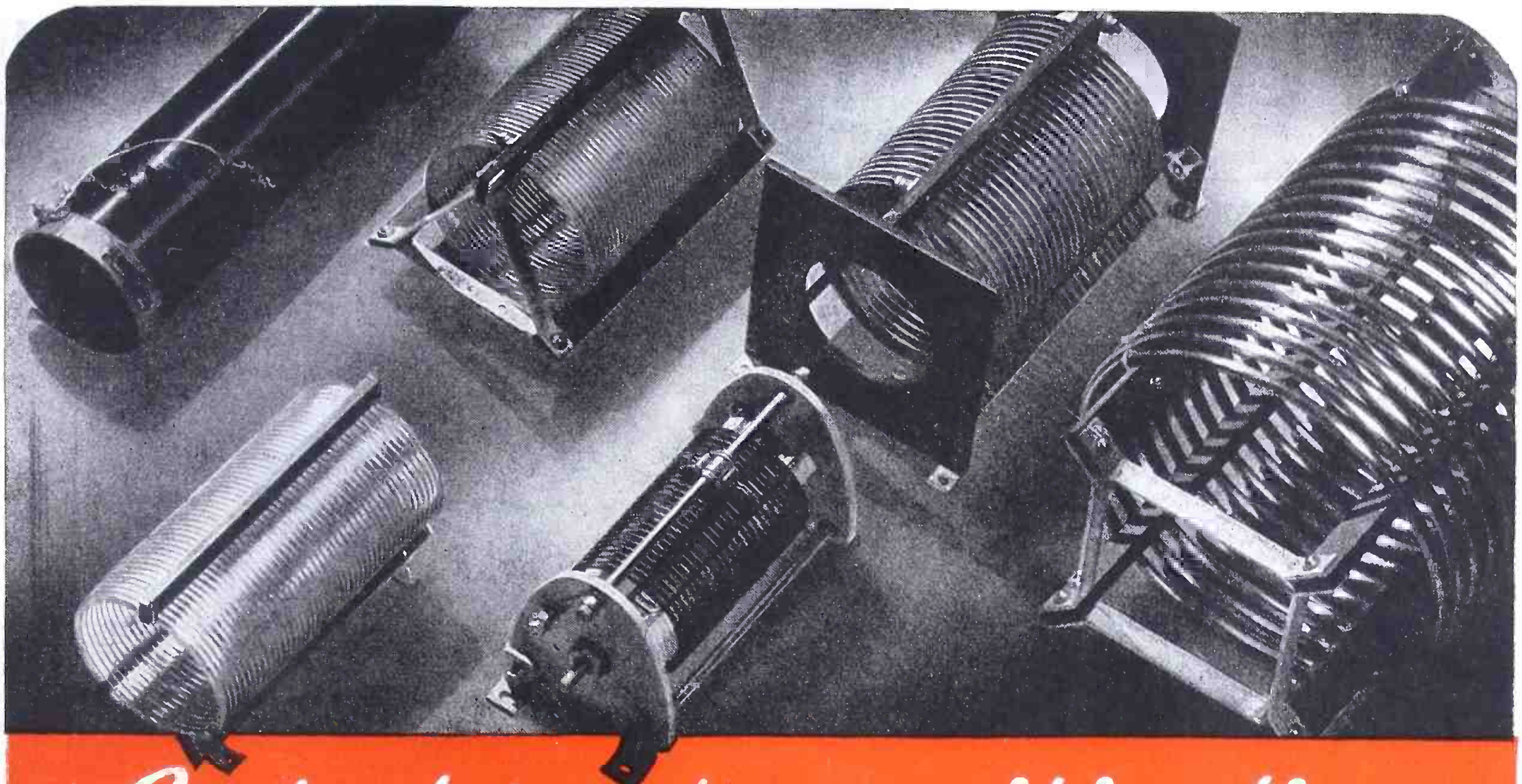
The conventional transformer, such as an output transformer, or a microphone transformer, has characteristics which are not at all suitable for inductances. To begin with, a transformer usually shows a very poor Q, that is, if each winding is considered as an inductance.

A conventional transformer has another poor characteristic, that is, the inductance of a given winding does not remain constant. A graph which shows the relationship between a-c permeability and flux density shows clearly that if the applied voltage is increased from a small value, the a-c permeability increases. This results in an increase in inductance (Figure 7). The flux density is proportional to the applied voltage.

For example, silicon steel laminations which are 100% interleaved, show a 4 to 1 increase in inductance when the flux density changes from 10 gausses to 1000 gausses corresponding to a 100 to 1 voltage increase. If this applied voltage is still further increased, the inductance will also increase up to a critical value, at which point the core material is said to exhibit its maximum permeability. This is around 4,000 to 5,000 gausses. If the applied voltage is still increased, the inductance decreases very rapidly as the core material approaches saturation. The curve mentioned above, up to the point of maximum permeability, will show at a positive slope. At maximum permeability the slope is zero, and at a still higher value of flux density the slope is negative.

This fact can be more easily observed by using an inductance bridge such as the GR 650-A, and instead of using the source of voltage within the bridge, an audio oscillator with an attenuator can be used, so that widely different voltages can be applied to the bridge. The highest voltage would be determined by the maximum allowable voltage which the bridge could withstand without damage, and the lowest voltage by the amount of amplification available for determining the null point of the bridge. Using this method, voltages through a range of 1,000 to 1 can be applied to the coil under test, noting

(*Continued on page* 87)



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IMPEDANCE TRANSFORMATION

(PART TWO OF A TWO-PART PAPER)

by PAUL J. SELGIN

Instructor, Electrical Engineering, Polytechnic Institute of Brooklyn

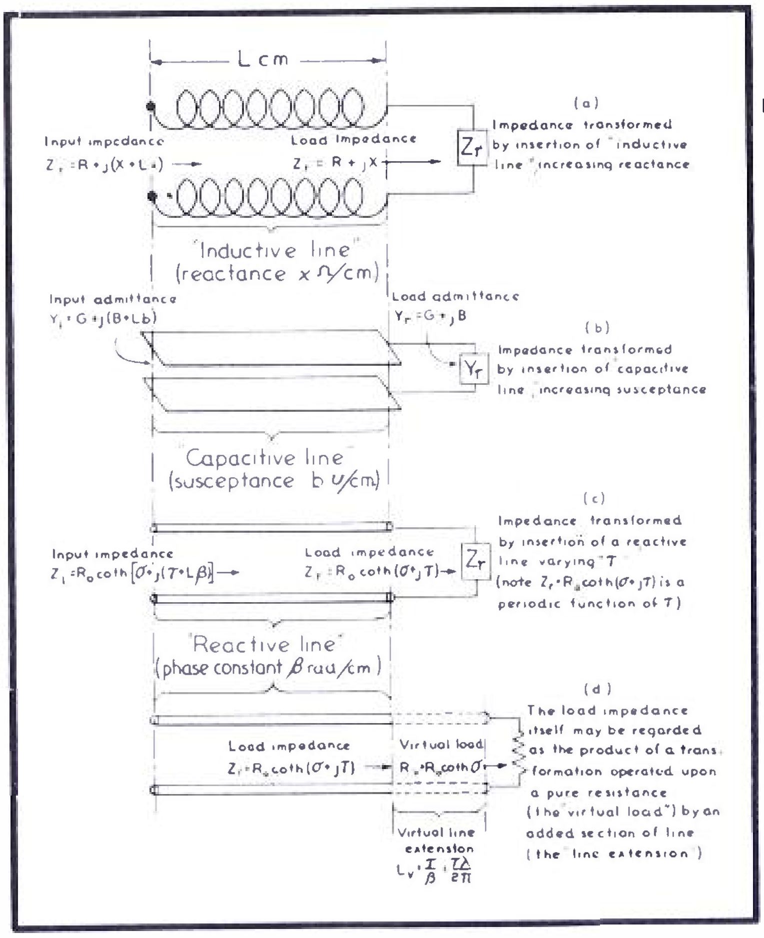


Figure 6

Illustrating the impedance transforming action of a reactive line.

IN the first part of this paper, the **principle** of impedance trans**formation** was considered in gen**eral**; impedance transformation by **means** of the addition of series and **parallel** branches, and its geometric representation, were taken up in greater detail. It appeared that every single-frequency impedance transformation problem could be solved, *on paper*, by the use of *two-element* networks, comprising a perfect coil and a perfect condenser (except two coils or two condensers).

Unfortunately, however, this solution can be used only up to a value of frequency which cannot be set once for all, but depends on the values of load and generator impedance and the allowable dissipation.

Deviation of Coil Reactance

As the frequency is increased, the true reactance of a coil deviates more and more from the *theoretical* value based on the *inductance* of the coil (a parameter which, strictly speaking, has a meaning only at d-c). The device of linking the same current many times with the same magnetic path, used so successfully at low frequency, must be discarded when the displacement current between turns grows to high values. At very high frequencies, there must be some relationship between the voltage and the distance between two points.

These and other considerations point to transmission lines as the answer. The operation of transmission lines in contrast to that of coils, becomes more straightforward as the frequency increases (up to the point where the wavelength is comparable to the lateral dimensions).

Let us consider the input impedance of a section of transmission line (Figure 6 (c)) when the load impedance (output impedance) has some value Z_r. We cannot conveniently think of the line as a combination of series and parallel branches, which would transform the load impedance by successive additions similar to those considered in the first portion of this paper. We can, however, think of the line as a single *symmetrical network*. As a matter of fact, the transmission line, at sufficiently high values of the frequency, is an exceptionally simple network. Its *characteristic impedance* is a pure resistance, of value

$$R_o = \sqrt{l/c} \qquad (1)$$

where l is the distributed inductance, c the distributed capacity. Its transfer constant is an imaginary, of value

$$\theta L = j\frac{2\pi L}{\lambda} \qquad (2)$$

where L is the length of the line in meters, λ the wavelength. The number $\frac{2\pi}{\lambda}$ is also called the *phase constant* of the line, and $\frac{2\pi L}{\lambda}$ is often called the *line angle*.

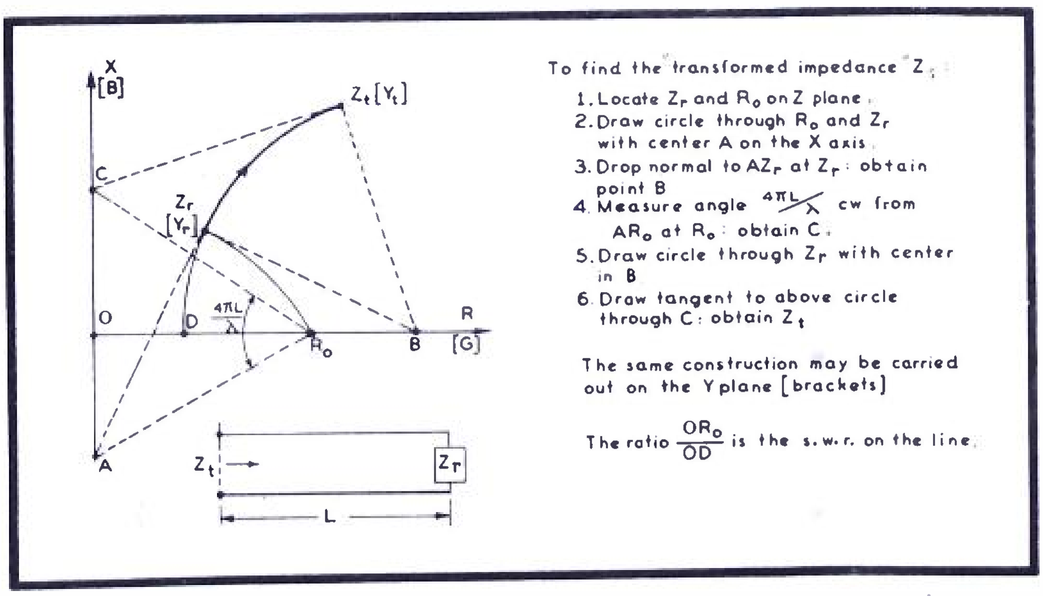


Figure 7

Impedance transformation by a reactive line.

Graphical Methods of U-H-F Transmission Line Analysis

The problem of determining the *impedance transforming action* of a line of this type (which we will call *reactive line*), reduces to the following:

(1)—Given the load impedance, find the corresponding *reflection factor* ρ. This may be done analytically, by expanding the equation

$$Z_L = R_o \coth \rho \qquad (3)$$

expressing the general relation between load impedance and reflection factor. Since R_o is known (equation *1*), the above may be solved for ρ, which will appear, in general to be complex. It is often more convenient to find ρ by chart methods or by construction. The hyperbolic tangent map (see the first part of this paper) may be used; or we may use constructions and charts discussed in *2* and *3*.

(2)—Given ρ, add to this number the *transfer constant* of the line, given by *2*. Thus we find the complex number:

$$\rho' = \rho + j\frac{2\pi L}{\lambda} \qquad (4)$$

(3)—Repeating step *(1)* in the opposite direction, find:

$$Z_t = R_o \coth \rho'$$

which is the required *transformed* impedance. Whatever method has been used in step (1) may be used here again, of course.

Figure 6 illustrates the analogy between an *inductive line* (two long solenoids with negligible mutual and inter-turn capacitances), a *capacitive line* (two flat parallel strips with negligible inductance) and an ordinary transmission line with appreciable values of both l and c. The three lines add imaginary terms to the values of Z, Y and ρ respectively, in proportion to their length. The concepts of *virtual load* and *virtual line extension* are also illustrated.

Determining Line Performance by Ruler and Compass

Constructions using only a ruler and a compass are very desirable, as they are infinitely less laborious than the corre-

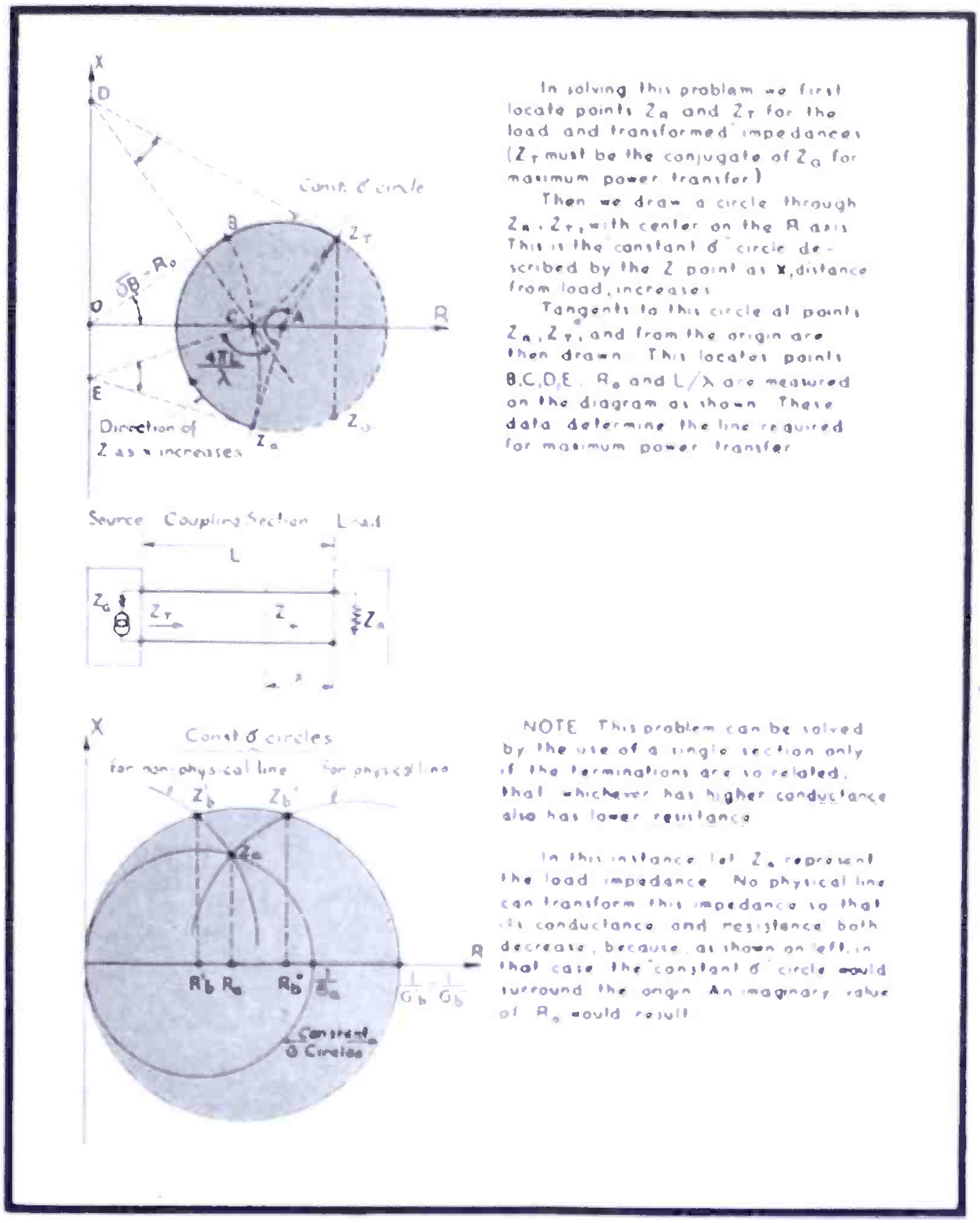
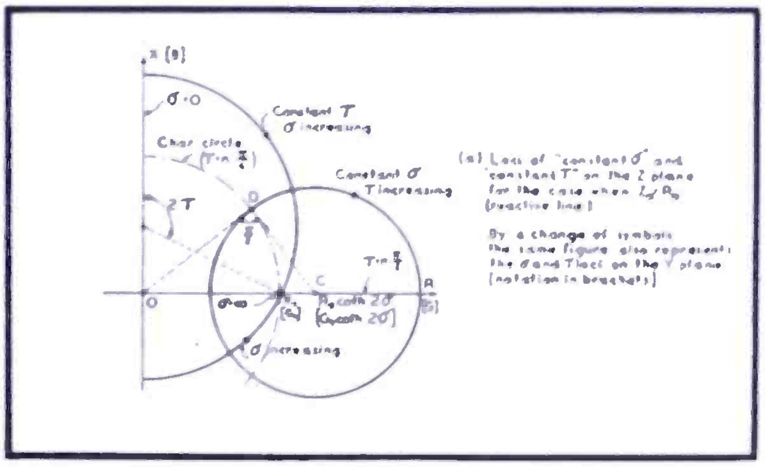


Figure 9

Solution of the coupling problem by means of a section of reactive line.

Figure 8

Impedance transformation by a network insertion.



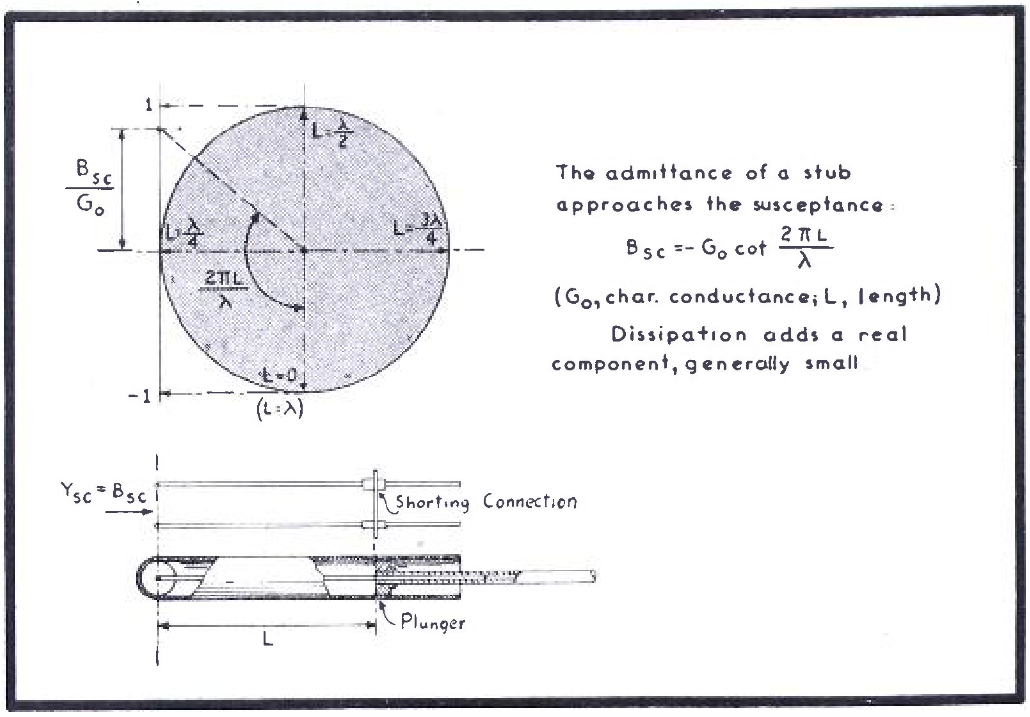


Figure 10
Short circuited line section (stub). Upper left shows construction for B_{SC}. Lower left shows parallel and coaxial stubs.

sponding computations, and more accurate and flexible than most chart methods. The problem of finding the input impedance of a reactive line, and hence any other data pertaining to the line in question, is solved by the construction of Figure 7.

Essentially, this construction locates the *point* corresponding to the required input impedance on the *impedance plane*, by tracing out the path described by this point, as an increasing length of line is inserted between the load and the input terminals. This path must correspond to a variation of the imaginary part of the reflection factor, as we have seen. Writing this factor in the form

$$\rho = \sigma + j\tau \qquad (5)$$

the path in question may be considered as a *locus of constant* σ, because a variation of τ does not change σ, or vice versa (just as a variation of reactance does not affect resistance).

The loci of constant σ on the *Z* plane, and also on the Y plane, are circles, as shown in Figure 8. The loci of constant τ are also circles, intersecting the former at right angles. This double family of circles is a map of the function

$$\rho = \sigma + j\tau$$

on the *Z* plane. σ and τ are sometimes called *bipolar coordinates* because, oddly enough, the identical family of circles also represents the magnetic and electric flux lines in a *bipolar* system of two parallel cylindrical conductors, which is one of the two types of transmission line systems in common use. [The other is the coaxial line, with concentric conductors of which the outer one is tubular.]

Line Couplings and Use of Stubs

Returning to the construction of Figure 7, the inverse construction is also evidently possible. That is, if we are given the two *end points* of the impedance transformation, Z_r and Z_t, we may by construction determine the parameters of a reactive line capable of effecting the transformation. Figure 9 shows this construction in detail. If we carry it out in a number of cases, we find that a solution is not always possible. The required value of R_o may turn out to be negative, which of course is not true of any physical line. The problem may be solved only if that termination for which the *conductance* is higher, also has the lower value of *resistance*.

For this reason, it is often necessary to use couplings which include, in addition to the coupling section of line, one or two *parallel* susceptances shunting this line at the ends or at an intermediate point. These susceptances are supplied by added sections of transmission line, connected across the coupling section at one end, while the other end is shorted (Figure 10). These additional *dead-end* line sections are called *stubs*. An example of coupling including two stubs, and the corresponding *Y plane* diagram, is shown in Figure 11.

An important reason for using stubs (aside from the impossibility of obtaining a conjugate match with a simple line coupling in particular cases) is that the susceptance of stubs can easily be adjusted, simply by moving the shorting connection. Such adjustments are very necessary in practice, not only when the transmitted frequency is varied, but also because the terminal impedances Z_r and Z_t are never accurately known. The setting of the stubs which permits maximum power flow is usually found by *tuning* for maximum power, as in the case of a tuned amplifier. Because two separate conditions must be met for maximum power transfer, and one condition is required for maximum voltage, coupling systems in general must have two adjustments and cannot be tuned

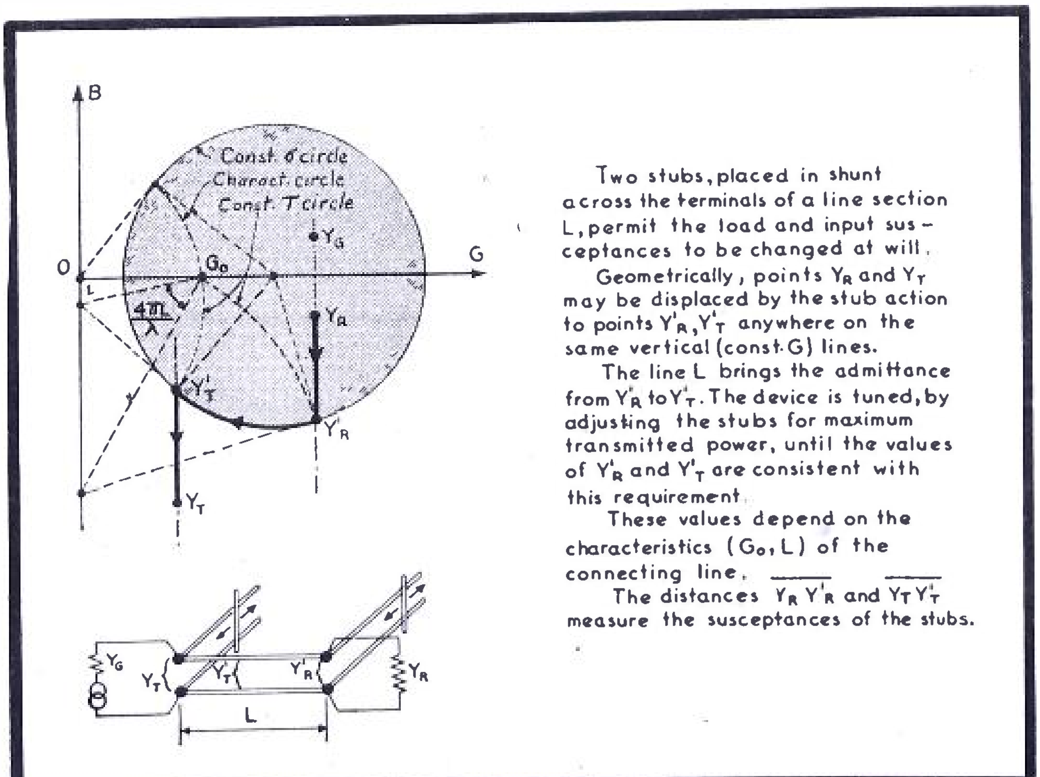
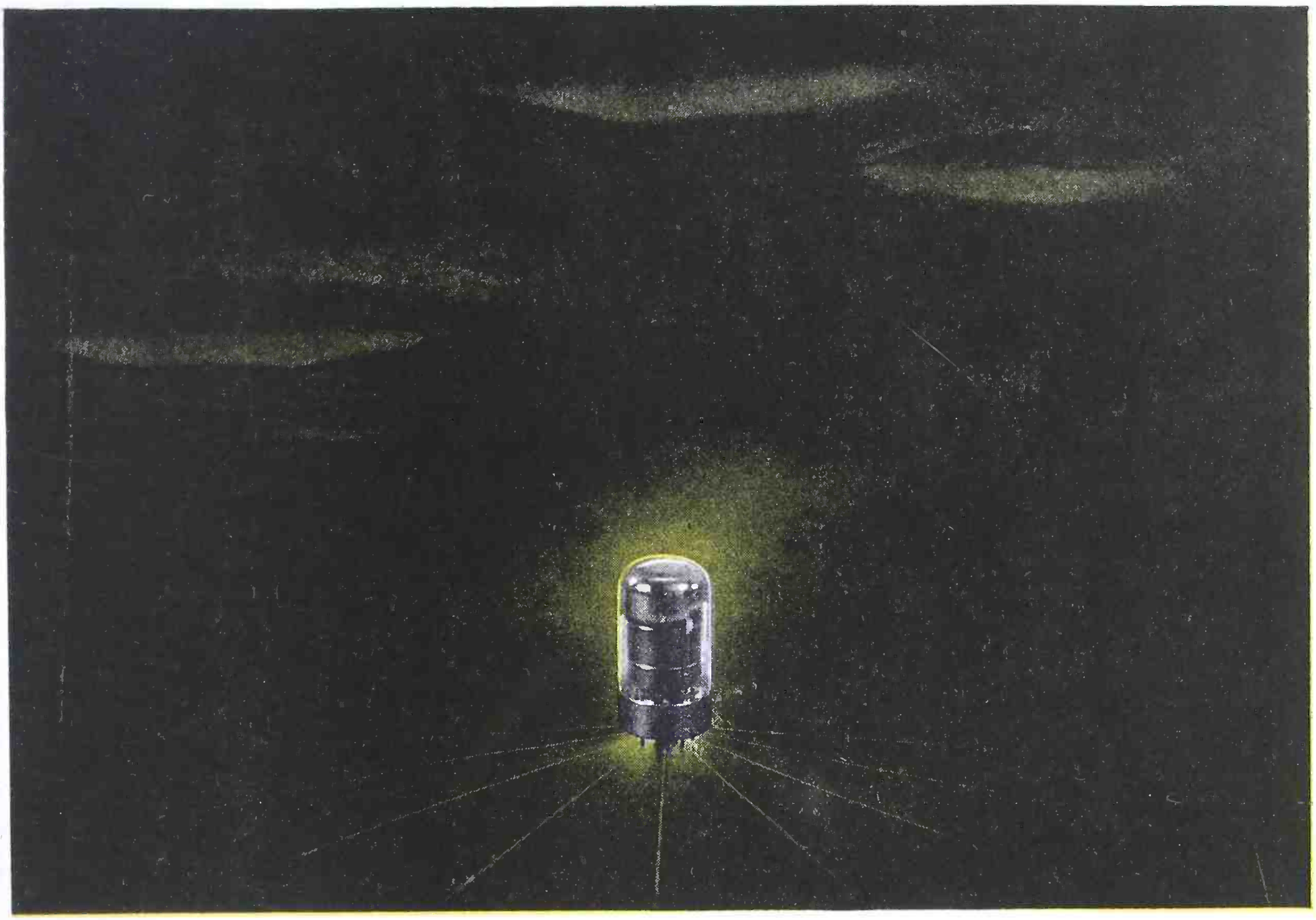


Figure 11
A double shunt tuner.



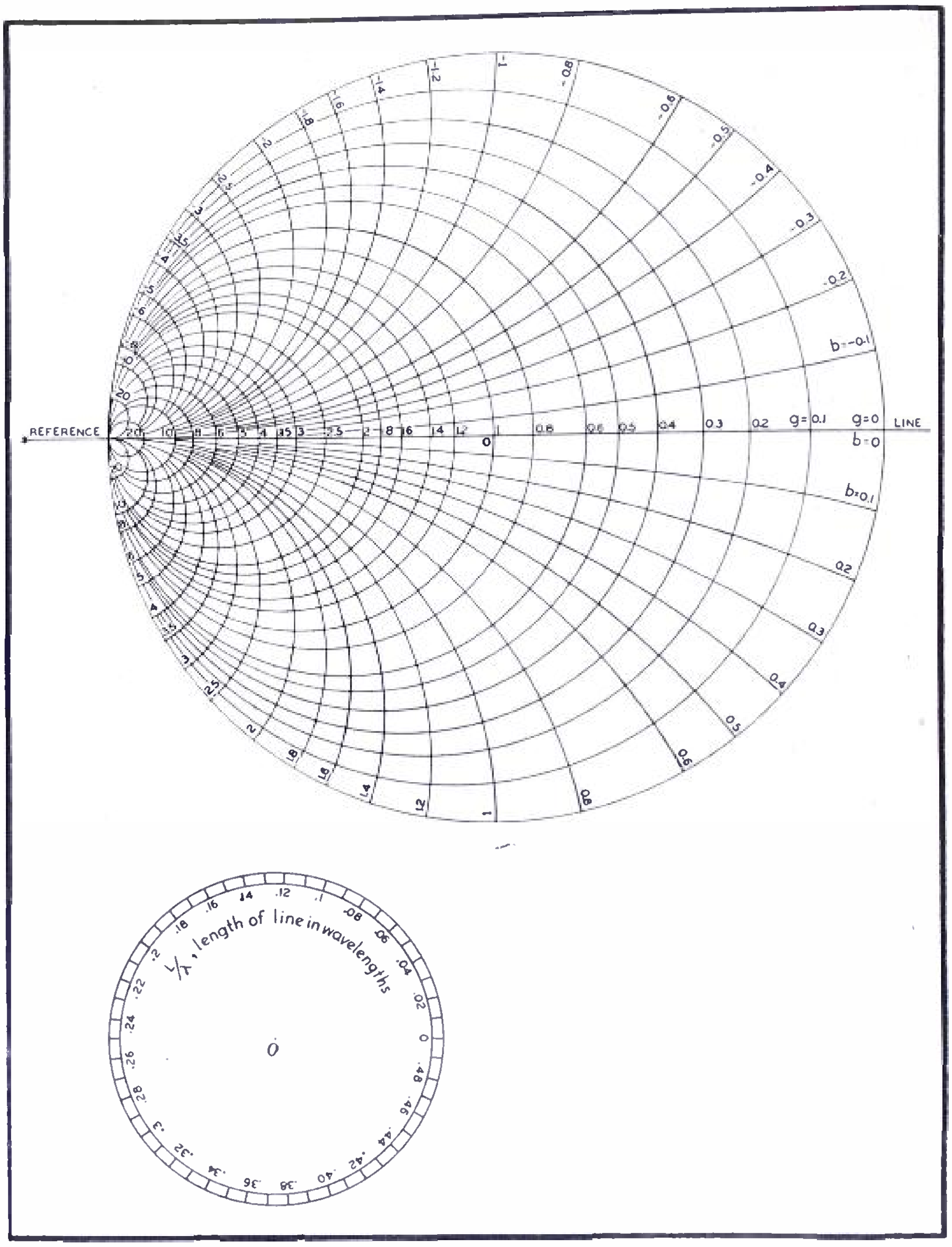


Figure 12

Map of the y function on the k plane ($y = k - 1/k + 1$).

as readily as an amplifier, which has only one.

Relative Advantages of Transmission Line Charts

It was pointed out that the charts used in transmission line computations (at power, telephone or ultrahigh frequencies) are, essentially, tools for obtaining the function ρ from Z (or Y) and vice versa. A number of schemes may be used for this purpose. In all of them a *map* of either function is drawn on the *plane* of the other function. Each map has its particular field of usefulness. The *hyperbolic tangent* map, which appeared in the first part of this paper, is particularly useful for power lines and long telephone lines and cables. On this map, the *point of operation* describes a straight path as the length of the transmission line is varied. The presence of attenuation merely changes the slope of the path. If characteristics of the line or cable are known, it is possible to tell, in case of fault, how far along the line the fault has occurred, and whether it is an *open* or a *short*, simply by measuring the impedance at the *near end* and drawing a straight line with the proper slope on the hyperbolic tangent map.

When attenuation is negligible, as in short lengths of u-h-f line, the map of "ρ" on the Z (or Y) plane, made up of circles like those of Figure 8, has definite advantages. However, in the opinion of the writer it is better not to have the map drawn at all and to draw the required arc of circle by construction (Figure 7). In this way, problems involving successive sections of line with different characteristic impedance values may be solved by repeating the construction, with R_o in a new position for each section. This would be manifestly impossible if the map were pre-drawn.

There is a third type of chart, also extensively used, and possessing definite advantages. It is obtained when we map the impedance (or admittance) function on the plane whose points represent the following function

$$k = e^{-2\rho} = \frac{Z - Z_o}{Z + Z_o} \qquad (6)$$

Considering k as a *vector*, it will be noted that the *angle* of this vector is equal to 2τ (from equation 5). Hence an increase in τ corresponds to a movement of the k *point* on a circle concentric with the origin.

Line Calculator

By following the directions of Figure 12, a useful line calculator may be assembled. Such calculators have

(*Continued on page 87*)

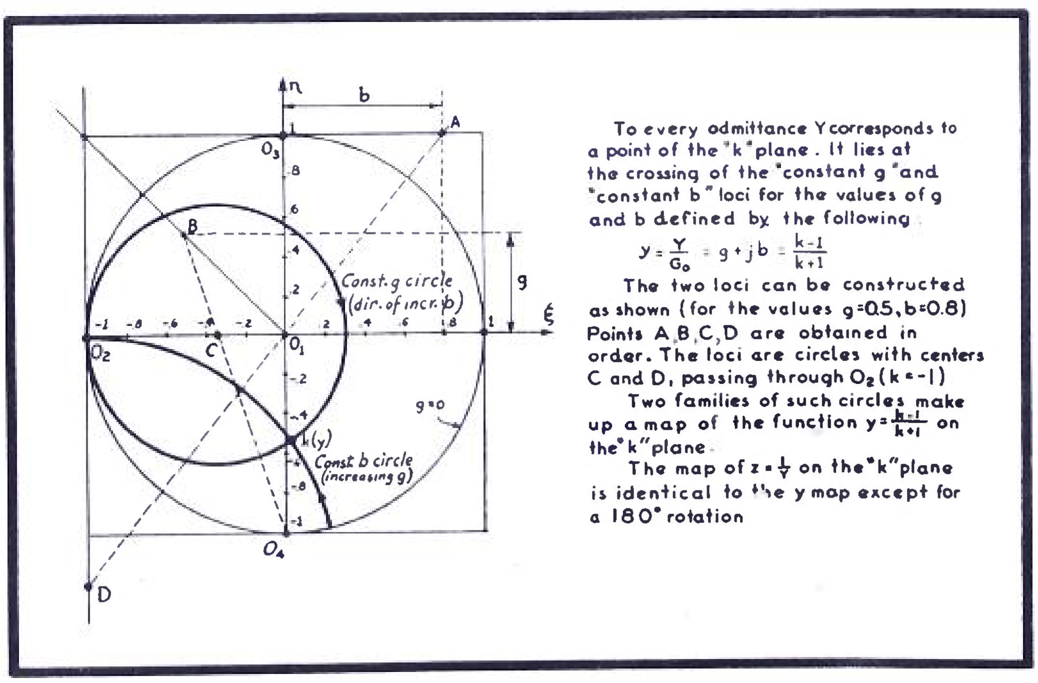


Figure 12(a)

Loci of *constant g* and *constant b* on the *k* plane.

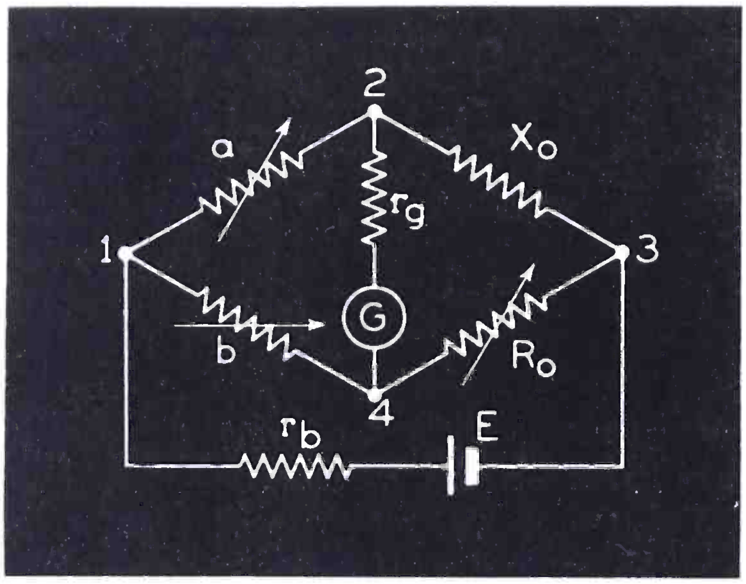
WHEATSTONE BRIDGE

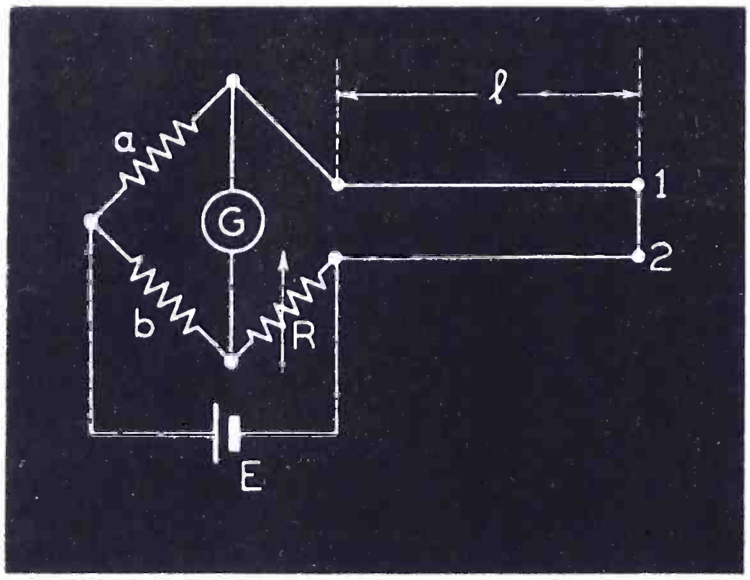
And Its Applications

(PART THREE OF A THREE-PART PAPER)

by PAUL B. WRIGHT

Communications Research Engineer



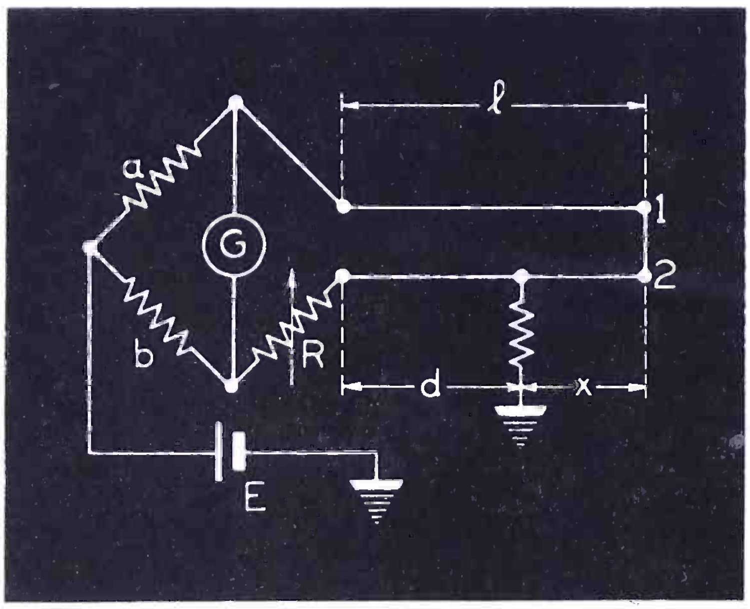


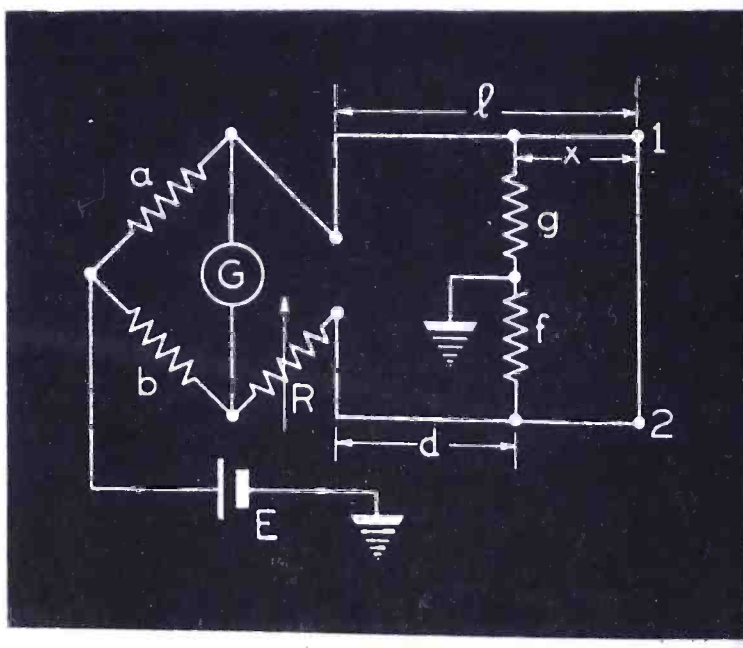
To facilitate reading of this concluding installment, these four illustrations are repeated from previous installments. Figure 3, at upper left, shows a conventional schematic of the Wheatstone bridge circuit. In Figure 6, at left, we have a simple loop measurement of a pair of wires. Figure 7, below left, illustrates a grounded Varley measurement. Figure 10, below right, illustrates a corrected Varley method.

PART I analyzed and considered some of the general aspects of the Wheatstone bridge circuit with respect to accuracy and sensitivity with variations in the element values of the bridge arms. Part II presented an introduction to the theory and application of the bridge for the purpose of locating faults on lines and cables. Formulae and methods of operation were given for a number of the basic applications in fault location work. These were: the simple resistance, line loop resistance, grounded Varley, metallic Varley, corrected Varley, Varley-Fisher or three-wire Varley, Murray loop, and one method of straight resistance measurement. In this, the concluding installment, additional methods of measurement of faults are shown with applications for open location types of measurement by modification of the bridge circuit. Methods of making insulation measurements by means of an auxiliary high resistance voltmeter are also shown. In conclusion, a number of practical considerations are pointed out to assist in evaluating the results obtained.

Straight Resistance

Method 2. This method assumes the same conditions as Method *1* and is subject to the same limitations, but permits measurements to be made from one end only. It thus possesses the advantage of requiring only one bridge. This method may also be used to deter-





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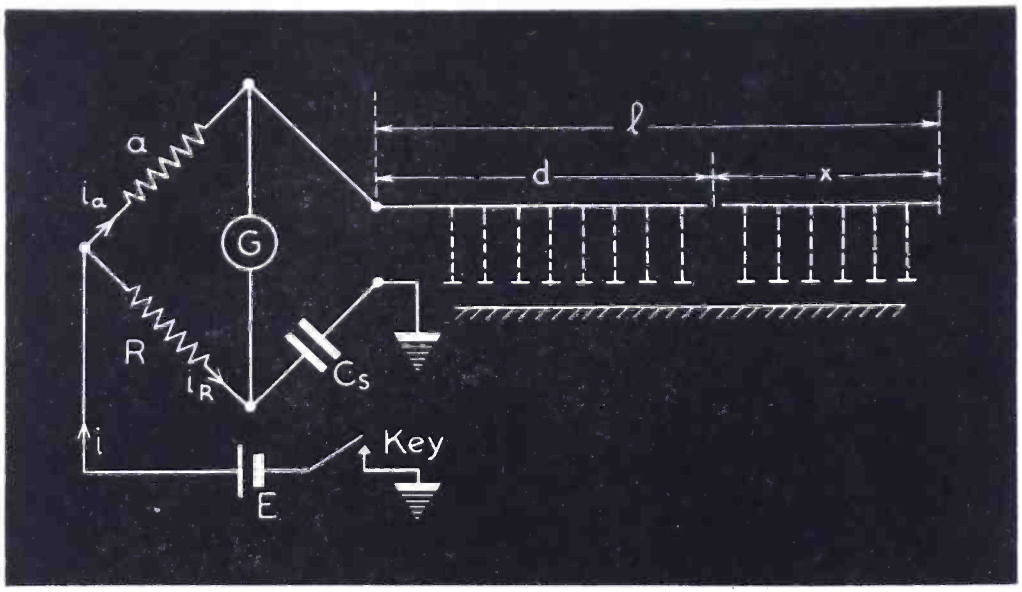
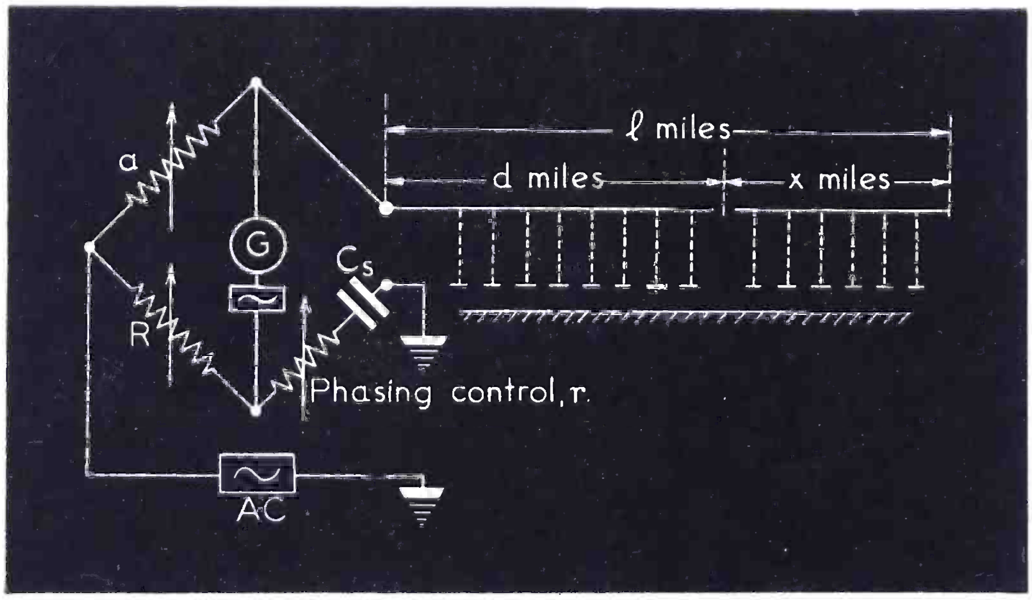
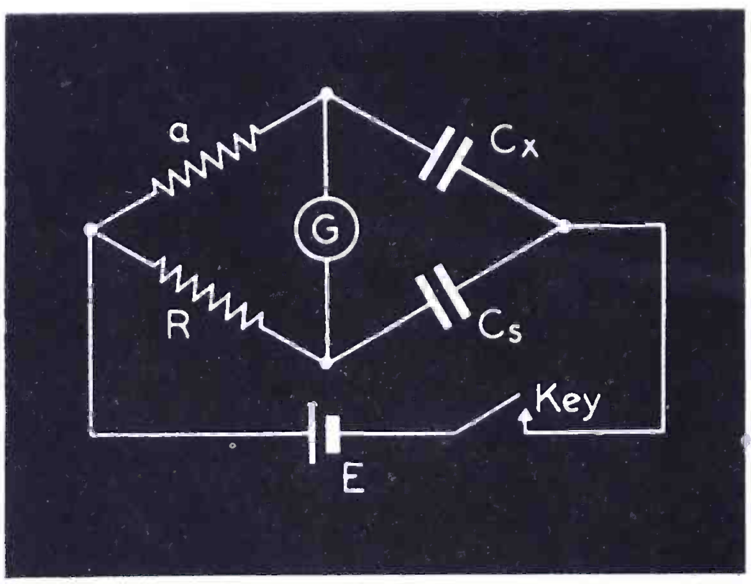


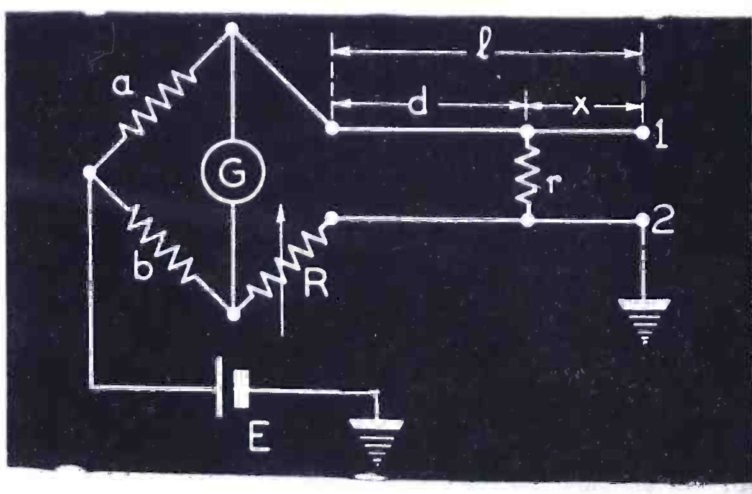
Figure 13a
Measurement of an open fault on a line by deflection balance.

Figure 13b (below)
Measurement of an open fault by means of a modified bridge circuit.



Figures 14 (below, left) and 15 (below, right)
Figure 14, measurement of unknown condenser capacity. Figure 15, a Varley measurement of the resistance in a fault crossing a pair of wires.





mine the value of a high resistance to low resistance cross when a suitable ground connection is not available at the far end. When such ground connections are available, it is much more satisfactory and reliable to use the regular grounded or metallic Varley type of measurement. A simple loop measurement is taken, connected as in Figure 6 with a resistance of unknown value as in Figure 15, or with wires grounded through f and g ohms as in Figure 10. Letting

h = resistance of unknown cross, or sum of the wire resistances to ground.

L = 2 l, total loop resistance, obtained from records, or calculated from known wire constants and temperature.

2 d = L — 2 x = loop resistance of conductors to the fault.

For the far end open, let

$$M = 2d + h \qquad (86)$$

For the far end short circuited, let

$$N = 2d + \frac{h\,(2x)}{h + 2x} \qquad (87)$$

Solving *86* and *87* for h, equating the results, and solving for the resistance from the measuring end to the trouble in terms of *M*, *N* and *L*, noting that $2x = L - 2d$,

$$(2d)^2 - 2N(2d) + (MN - ML + NL) = 0 \qquad (88)$$

which is a quadratic equation in 2 d. By the quadratic formula,

$$2d = N \pm \sqrt{(N-L)(N-M)}\ \text{ohms} \qquad (89)$$

From *87*, $N \geq 2d$, also $L > N$ and $M > N$, hence *89* becomes,

$$2d = N - \sqrt{(L-N)(M-N)}\ \text{ohms} \qquad (90)$$

Using *90* in *86*, the resistance of the short or cross may be ascertained, giving,

$$h = (M-N) + \sqrt{(L-N)(M-N)}\ \text{ohms} \qquad (91)$$

These equations must each be divided by the resistance per loop foot in the

case of cable fault location work, or by the resistance per loop mile for open wire fault location. This gives the distance to the trouble in feet for cable and miles for open wire in equations *89* and *90*, and equivalent distance in equation *91*. If bridges should be available at both ends, the same method applied at both ends should give for each the same value of h, the resistance of the cross. This acts as a useful check upon the method and checks the accuracy of the final measurements. This method should never be used if it is possible to use either the grounded or metallic Varley methods however, since the frequent variation of contact resistance causes h to vary quite widely at times, decreasing its reliability and consequent utility.

Resistance of a Cross or Short

This is a variation of the grounded Varley of Figure 7 and is shown in Figure 15, with one wire grounded at the far end. A Varley measurement is then taken as shown. Since for this method it is assumed that the line wires are balanced, the only unbalance existing is that caused by the unknown short or cross, r. This must be balanced by the resistance R, for ordinarily, a = b, and we have immediately,

$$R = V = r \text{ ohms.} \qquad (92)$$

or the value of the resistance causing the fault is obtained by simply reading the value of R giving the Varley balance. Since R is generally between ten and twelve-thousand ohms, a can in practically all cases be made equal to b. When this cannot be done, the lowest ratio that can be used and is available with a particular bridge should be applied so that the sensitivity of the system will not be impaired unnecessarily. When this is the case, the resistance may be found by applying the previous theory to Figure 15. This gives

$$r = \frac{a}{b}(R + d) - d \qquad (93)$$

where the distance represented by d must be obtained by one of the methods given previously and converted to ohms. This illustrates strikingly the need for using the ratio arms equal to each other if complications are to be avoided in even the simplest Varley measurements in some cases.

The additional Figures of various applications shown by 16, 17, 18 and 19 are specific uses to which the ordinary Varley measurement may be put. In each case, the amount of unbalance is read directly from the value of the adjustable resistance, R, with the ratio arms equal to each other.

Other specific methods of locating series resistance unbalances by bridge measurements and alternating current impedance runs are given by the two papers referred to, in references *6* and *7*. These papers should be consulted by the interested reader who wishes to go further into the subject than is given here.

Open measurements are made upon lines by a modification of the Wheatstone bridge circuit. The ratio arm b is replaced by R, and the location of R in the circuit is replaced by a condenser capacity, usually on the order of 1 microfarad. This is a standard condenser of good quality, and is used for the purpose of comparing the capacity of a line to that of the condenser. This method of locating open faults is quite good when the lines are short and the insulation high. The accuracy goes down with longer lines and poor insulation to ground. This is because the line behaves less like a pure capacity under these conditions and therefore has both a resistive and a reactive component which would

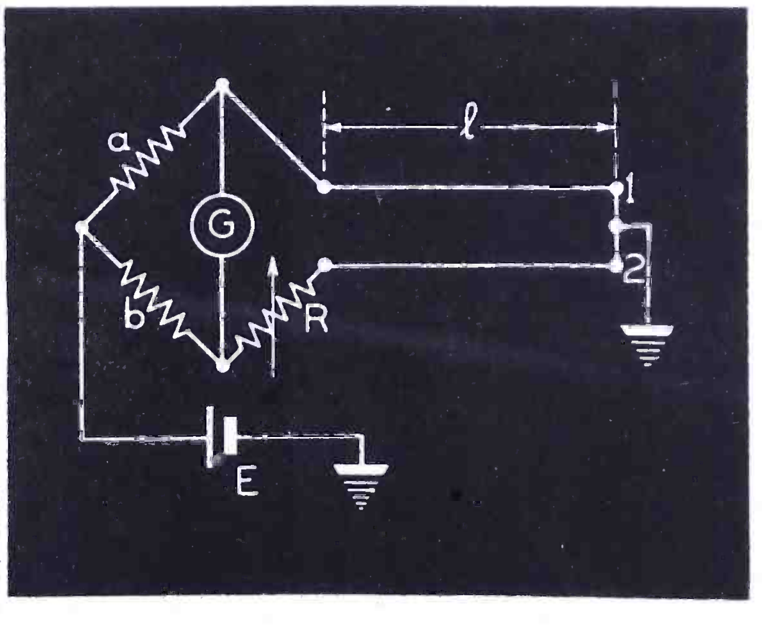


Figure 16
A Varley measurement to determine the unbalance between the wires of a pair.

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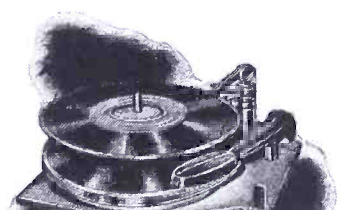

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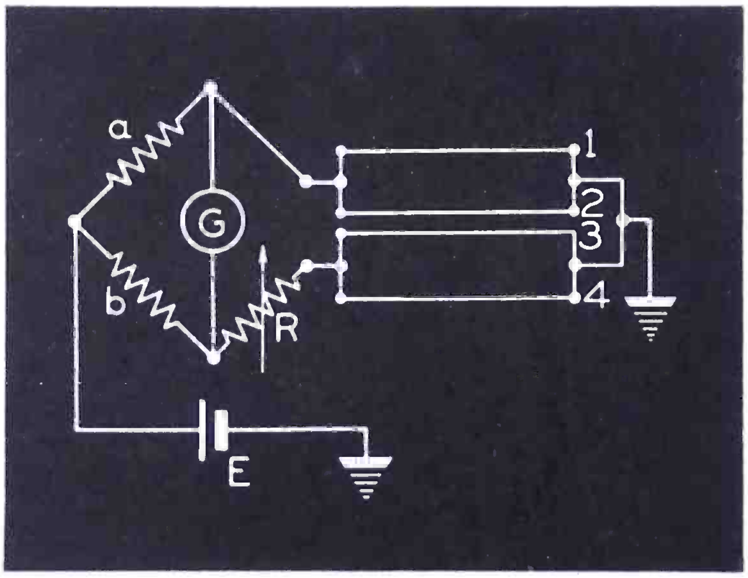
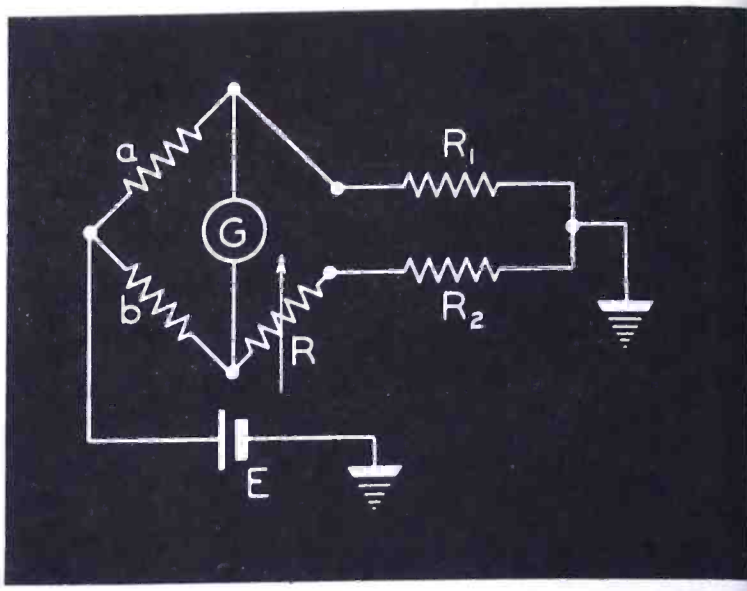




Figures 17 (top, left) and 18 (top, right)

Figure 17, a Varley measurement is used to determine the unbalance existing between the side circuits of a phantom group. Figure 18, resistance unbalance between selected resistors may readily be determined by the Varley method.

have to be balanced for accuracy. Many of the more modern types of bridges do take the resistive component into account, however, by means of a phasing control, which is simply a variable resistance in series with the standard condenser, c_s, as shown by Figure 13(b).

The line may be charged up as shown by the simplified diagram in Figure 11 using a key in the battery lead to ground while adjusting the resistance, R, to give zero deflection on the galvanometer. This very simplified method is not a very accurate one to use because the average throws or deflections on the meter are not zero for the charge and discharge of the line. And when measuring on lines having very short time constants, the discharge takes place so rapidly that the tendency is to attempt to balance the bridge by unbalancing it. Hence, in practice, it is more customary to use an alternating current of very low frequency from 4 to 20 cycles per second, and a reversing relay or commutation device for the galvanometer so that the galvanometer is always connected in the proper direction. Since the unequal charge and discharge currents through the galvanometer are equal for each poling of the alternating current, the average is zero for the separate charges and discharges of the line.

The necessary conditions of balance for Figures 13(a) and 14 are found by direct application of the previously developed theory, where the capacity of the line is considered as C_x, and the capacity of the standard condenser as C_s. Here, $R_x = 1/j\,\omega\,C_x$ and $X_s = 1/j\,\omega\,C_s$, then

$$a \cdot \frac{1}{j\,\omega\,C_x} = R \cdot \frac{1}{j\,\omega\,C_s} \qquad (94)$$

or

$$C_x = R\,C_s/a \qquad (95)$$

The value of a is most frequently set at 1,000 ohms, although any of the Murray bridge ratios can be used. With C_s at 1 mfd and a at 1,000 ohms, equation *95* becomes

$$C_x = R \times 10^{-3} \text{ microfarads} \qquad (96)$$

When R = 1,000 ohms, the value of the unknown capacity is equal to that of the standard. The reading of R at any value required to obtain a balance is frequently spoken of as having that many *spills*. The methods used to obtain the distance are based upon the proportional part of the number of spills obtained on the faulty wire to the number obtained on a good wire, preferably its mate. This ratio taken, times the total distance then gives the distance to the trouble. With bridges of the older types using 20-cycle reversals, the accuracy of measurement is not as great as with those of the more modern type mentioned above, since these bridges do not have a phasing control to take into account the variable resistance component of the line with variable physical and weather conditions. For average weather conditions with 20-cycle reversals open wire measurements will give about 15 spills per mile and decreases in value as the insulation resistance of the line goes down. In average cables the approximate value of resistance obtained gives about 90 spills per mile, and since the insulation of cables of all kinds is quite high, being better than 500 megohms per mile, the accuracy of results compared to open wire measurements is much greater.

Although the accuracy of open fault location measurements do not nearly approach in general that obtained by the grounded and metallic Varley measurements using direct conducting paths, it has been found in practice to be one of the indispensable types of tests.

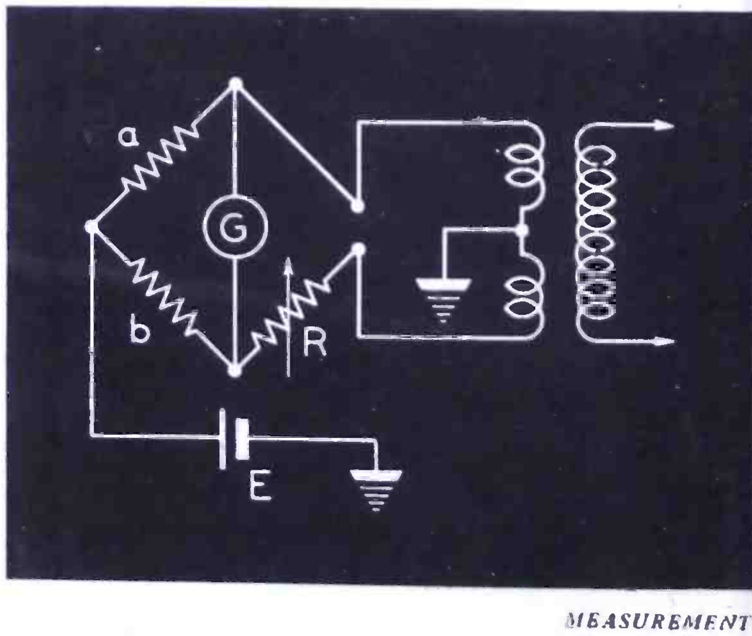
No hard or fast rules can be outlined for all of the conditions which may be encountered in fault location work, but a thorough knowledge of the theory of operation of the Wheatstone bridge circuit will assist in learning the methods of operation through practical experience.

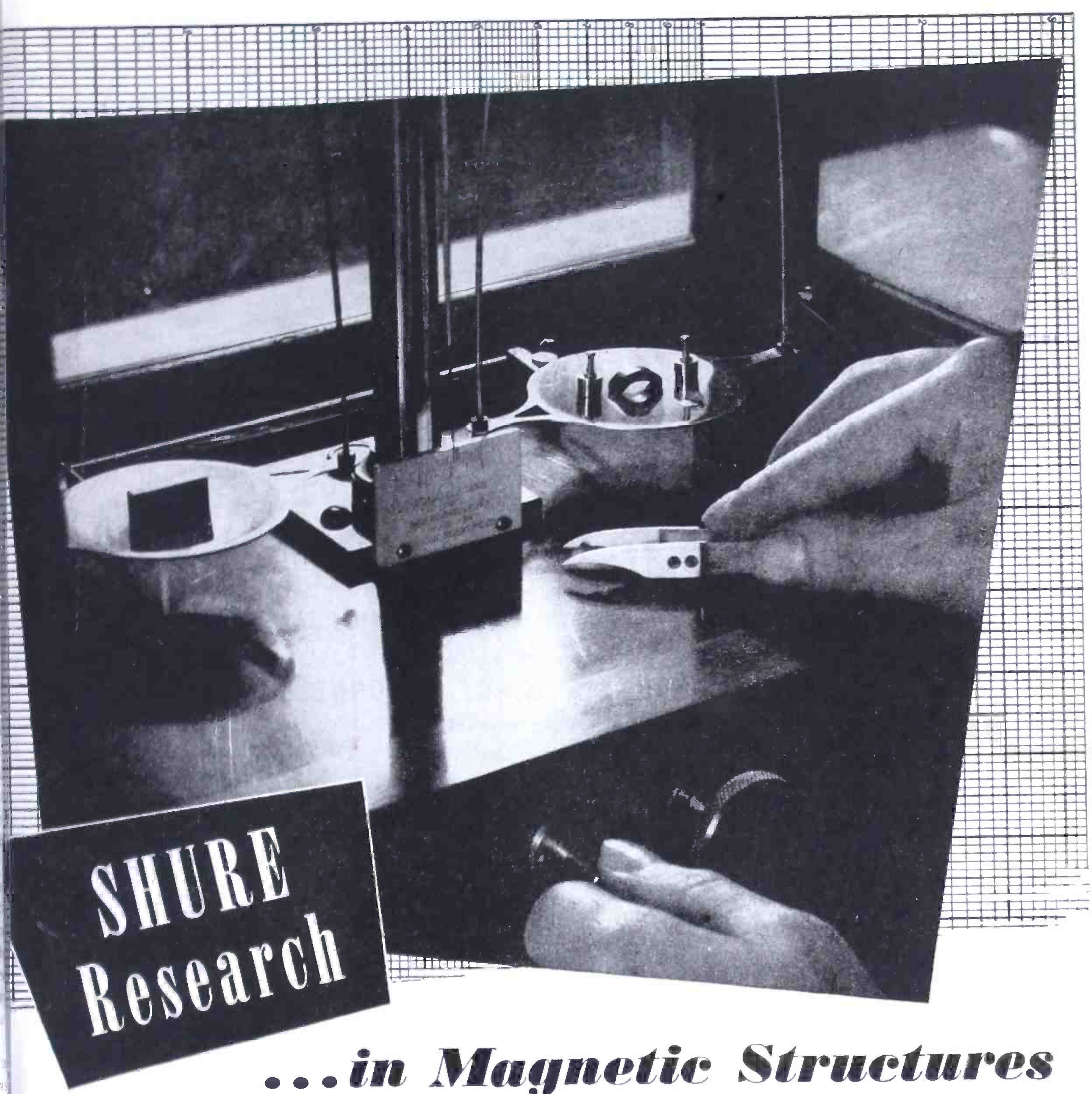
In determining the best method to use to locate a fault, we must learn as exactly as possible the nature of the fault condition. Faults may be classified into many special and specific

(*Continued on page* 70)

Figure 19

The d-c resistance of mid-tapped transformers in pushpull systems is easily obtained by application of the Varley method.





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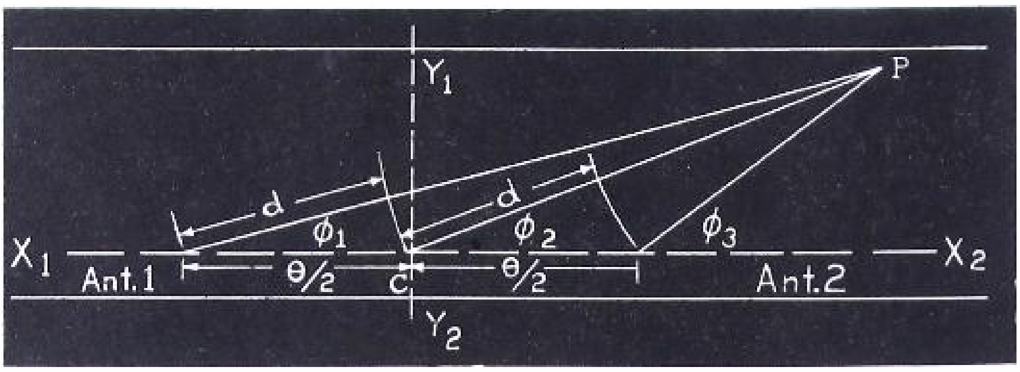

Figure 16

This plot reveals that the difference in t[h]
paths varies as the point P moves about t[he]
station. Assuming that P is at a great d[is-]
tance from the station, its bearing from t[he]
two antennas and from the point C will
equal for all practical purposes.

DESIGN AND OPERATION O[F] RADIO RANGE BEACON[S]

(PART TWO OF A TWO-PART PAPER)

by CAPTAIN W. G. McCONNELL

Army Airways Communications System

Theory, Adjustment and Maintenance

Appendix A

Field Pattern of Two Adjacent Antennas Excited 180° Out of Phase

The radiated field of an antenna can be expressed in the conventional form:

$$E_{t_1} = E_m \text{ sine } wt$$

in which case the field of the second antenna will have the form:

$$E_{t_2} = - E_m \text{ sine } wt$$

since it is 180° out of phase with the first antenna. The sum of the two voltages

$$E_{t_{1-2}} = E_m \text{ sine } wt - E_m \text{ sine } wt = 0$$

if the signal from both antennas reaches the receiving point simultaneously. This is the case on a line perpendicular to the axis of the antennas and passing midway between them.

At all other points, however, the distances from any point P to each of the antennas are not equal, and the difference in distance produces a phase shift so that the voltages do not cancel completely. The residual field is then a function of the electrical spacing of the antennas θ, and the angle ϕ that the point P bears to the axis $X_1 - X_2$ of the antennas.

In order to analyze the resulting field it is convenient to consider the phase of the voltage secured with respect to *C*, the midpoint between the two antennas.

If the spacing in electrical degrees between the antennas is θ, then for a point along the axis of the antennas the field will be

$$E_{t_{1-2}} = E_m \text{ sine } (wt + \theta) - E_m \text{ sine } (wt - \theta)$$

which will no longer equal zero.

By examining Figure 16 it is obvious that the difference in the two paths varies as the point *P* moves about the station. If we assume that *P* is at a great distance from the station its bearing from the two antennas, and from the point *C*, will be equal for all practical purposes. Therefore ϕ_1, ϕ_2 and ϕ_3 will be equal, and the difference will be the distance *d*. Since we have assumed that ϕ_1, ϕ_2 and ϕ_3 are equal, the small triangles will be, for all practical purposes, right triangles and the distance *d* will be equal to

$$\theta/2 \cos \phi$$

The complete expression for the field at any point *P* then becomes

$$E_{t_{1-2}} = E_m \text{ sine } \left(wt + \frac{\theta}{2} \cos \phi \right)$$

$$- E_m \text{ sine } \left(wt - \frac{\theta}{2} \cos \phi \right)$$

The terms of this expression have t[he] form

$$\text{sine } (A \pm B)$$

which can be expanded to

$$\text{sine } A \ \cos B \pm \cos A \ \sin B$$

Making this expansion we secure

$$E_{t_{1-2}} = E_m \text{ sine } wt \cos \left(\frac{\theta}{2} \cos \phi \right)$$

$$+ E_m \cos wt \text{ sine } \left(\frac{\theta}{2} \cos \phi \right)$$

$$- E_m \text{ sine } wt \cos \left(\frac{\theta}{2} \cos \phi \right)$$

$$+ E_m \cos wt \sin \left(\frac{\theta}{2} \cos \phi \right)$$

$$= [2 E_m \cos wt] \left[\text{sine } \left(\frac{\theta}{2} \cos \phi \right) \right]$$

The first part of this expression

$$2 E_m \cos wt$$

determines the magnitude and inst[...]

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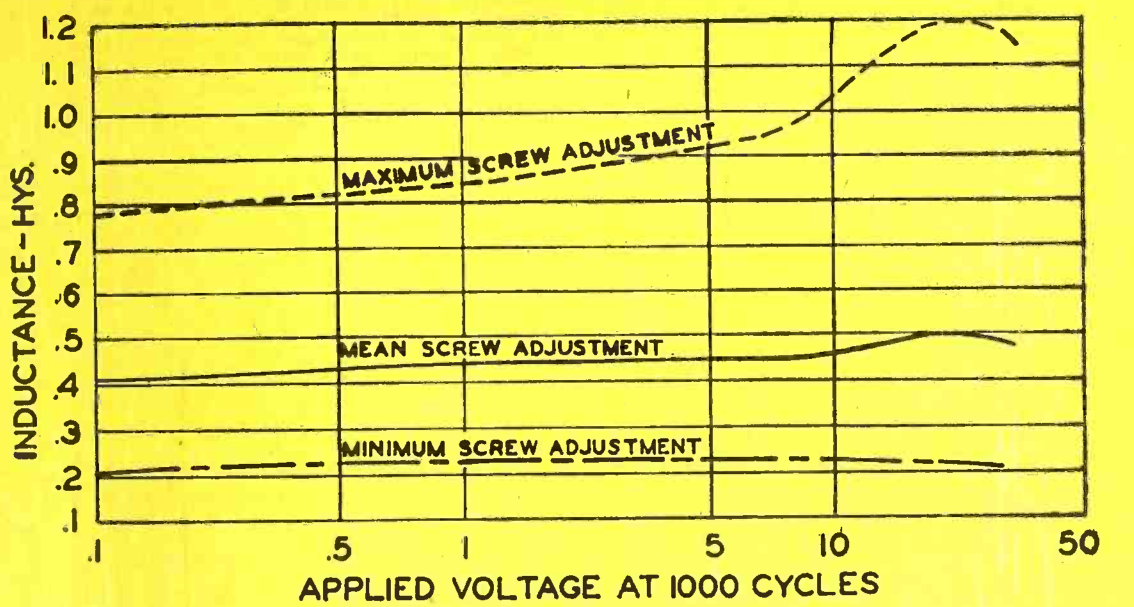

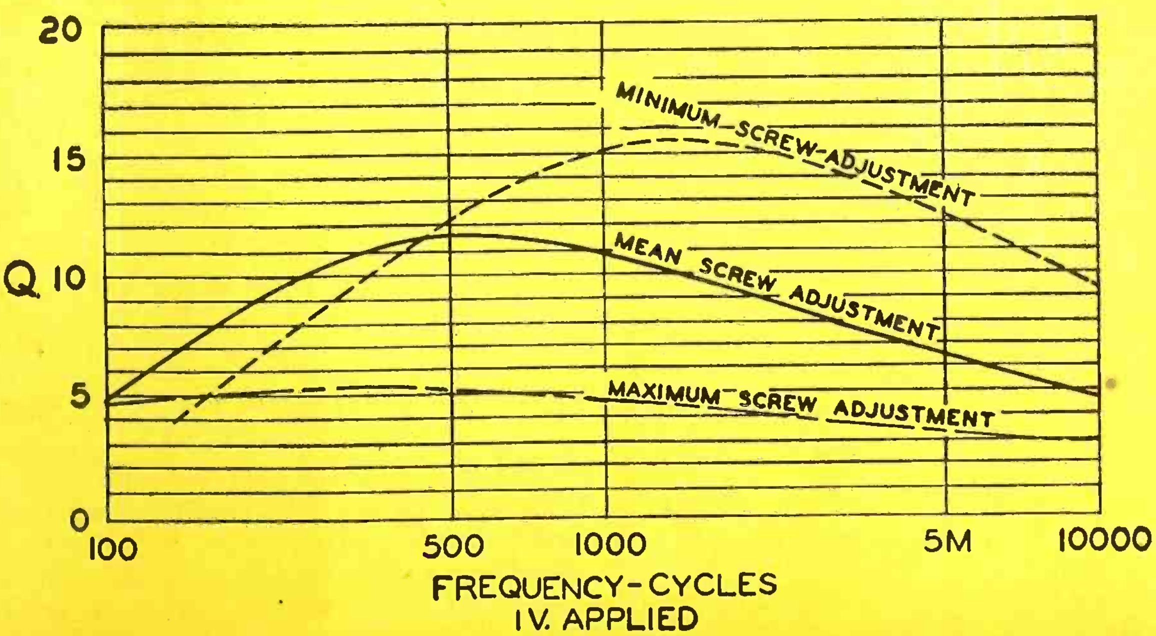

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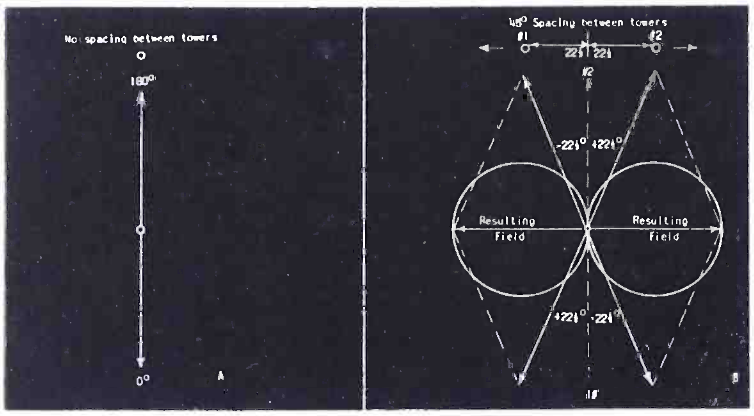
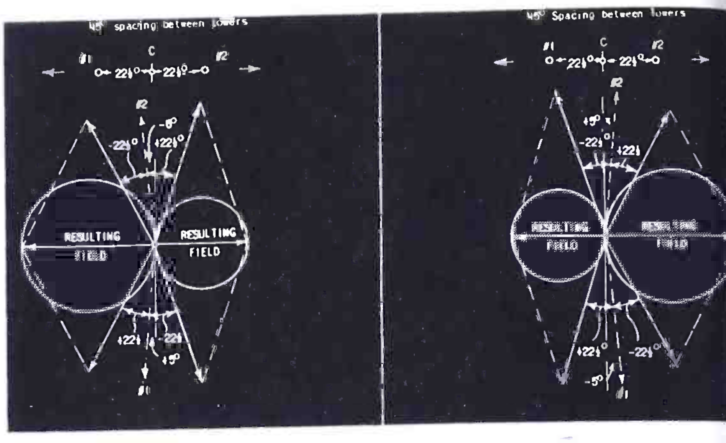




Figures 13 (left, above) and 13a (right, above)

Figure 13, effect of tower spacing in establishing pattern. In A, towers 180° out of phase; vectors cancel, no field. In B, tower 180° out of phase. Tower *1* advanced 22½° in direction of *1*; retarded 22½° in direction of *2*. Tower *2* retarded 22½° in direction of *1*; advanced 22½° in direction of *2*. Pattern is symmetrical. Figure 13a, effect of adding variation of transmission line length to tower to distort pattern. In C, towers 170° out of phase. *1* advanced 5°, *2* retarded 5°. Pattern increased in direction of *1*; decreased in direction of *2*. In D, towers 170° out of phase. *1* retarded 5°; *2* advanced 5°. Pattern increased in the direction of *2*; decreased in direction of *1*.

taneous voltage of the field, and is independent of the spacing and the bearing of the point *P*. It is interesting to note that the phase of the radio frequency voltage has shifted 90°, which would be expected since the maximum voltage would occur when the field of one antenna passes through zero.

The second half of the expression

$$\text{sine}\left(\frac{\theta}{2}\cos\phi\right)$$

determines the field pattern.

Along the axis of the antennas $\cos\phi = 1$, so that the voltage at this point will equal

$$(2\,E_m \cos wt)\left(\text{sine}\,\frac{\theta}{2}\right)$$

For the spacing of 180°

$$\text{sine}\,\frac{\theta}{2} = \text{sine}\,\frac{180°}{2} = \text{sine}\,90° = 1$$

So that the resulting field is equal to

$$2\,E_m \cos wt$$

or twice the voltage due to one antenna with a phase shift of 90°.

For a spacing of 90°, which occurs at 400 kc for a tower spacing of 600 feet

$$\text{sine}\,\frac{\theta}{2} = \text{sine}\,\frac{90°}{2} = \text{sine}\,45° = .707$$

and the field is

$$E_{f_{1\text{-}2}} = 2\,E_m \cos wt \times .707$$

$$= 1.4\,E_m \cos wt$$

If the point *P* is on the line $Y_1 - Y_2$, which is perpendicular to $X_1 - X_2$, and which passes through *C*

$$\cos\phi = \cos 90° = 0$$

and the field strength

$$E_{f_{1\text{-}2}} = 2\,E_m \cos wt\ \text{sine}\left(\frac{\theta}{2}\cos\phi\right) = 0$$

as we stated before. Thus no signal will be received regardless of the spacing θ.

The form $\text{sine}\left(\frac{\theta}{2}\cos\phi\right)$ is not a true figure-of-eight. The lobes tend to bulge at the sides and the larger θ becomes the greater they differ from true circles; until, when $\theta = 360°$ the radiation along the axis becomes zero, and each lobe splits in two with a major axis bearing approximately 45° to the axis $X_1 - X_2$.

If $\theta/2$ is small, however, we can assume that

$$\text{sine}\,\frac{\theta}{2}\cos\phi = \frac{\theta}{2}\cos\phi\ *$$

in which case the equation for the field pattern becomes

$$E_{f_{1\text{-}2}} = 2\,E_m\,\frac{\theta}{2}\cos\phi\ *$$

which is a figure-of-eight with the lobes true circles. Practically, the figure-of-eight diagram can be assumed as a pair of circles up to spacings of 90°.

Appendix B

Course Bending of SRA Stations

Figure 13 illustrates the manner in which the figure-of-eight patterns are distorted by means of vector diagrams.

*In these equations $\frac{\theta}{2}$ must be expressed in radians.

The vectors show the amplitude [of] the pattern along the axis of the [an]tenna only. In the notation of *A[p]pendix A*, ϕ is assumed to be zero, a[nd] $\cos\phi$ is unity. *A* indicates the canc[el]lation of the vectors when θ is ze[ro]. Thus $\theta/2$ is zero and

$$E_m \text{ sine } wt - E_m \text{ sine } wt = 0$$

The two vectors cancel and there [is] no field.

B indicates the value of the fi[eld] vector when θ equals 45°. In this c[ase] we have (see *Appendix A*)

$$E_{f_{1\text{-}2}} = E_m\,\text{sine}\left(wt + \frac{\theta}{2}\right) - E_m\,\text{sine}\left(wt - \frac{\theta}{2}\right)$$

which becomes

$$[2\,E_m \cos wt]\ [\text{sine}\ \ \theta/2]$$

$\cos\phi$ being equal to 1.

$E_m \cos wt$ is the length of [the] vector.

Therefore, the resulting field is

$$F_{1\text{-}2} = (V)\ 2\,\text{sine}\,\frac{\theta}{2}$$

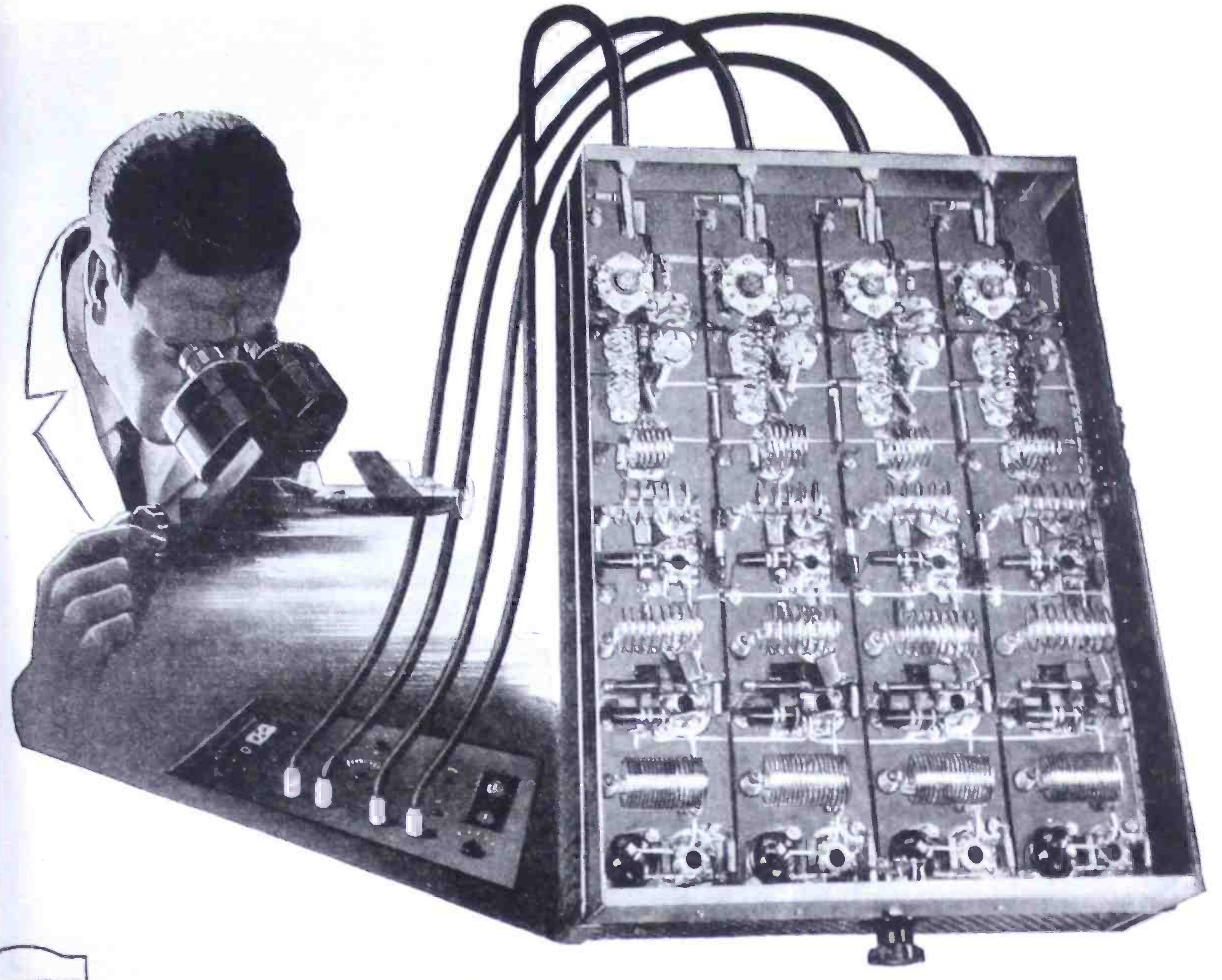
or

$$F_{1\text{-}2} = (V)\ 2\,\text{sine}\,22\tfrac{1}{2}°\,.$$

and the resulting field is equal [to] .77 V.

The vector diagram illustrates [the] manner in which the phase of [each] antenna is retarded for radiation [in] one direction and advanced for rad[ia]tion in the other direction. For rad[ia]tion to the left of the figure, the vec[tor] of tower 1 is rotated clockwise (ad[vanced]) 22½°; the vector of towe[r 2] is rotated counter-clockwise (re[tarded]) 22½°. The resulting field i[s] secured by completing the parall[elo]gram. For radiation to the ri[ght] tower 1 is retarded and tower 2 is a[d]vanced. Thus the resulting field i[s] equal and the pattern is symmetr[ic]al.

C indicates the effect when [the] phase of the radiation of the [two] towers is no longer 180° apart. I[n] this case, the phase of tower 1 a[...]



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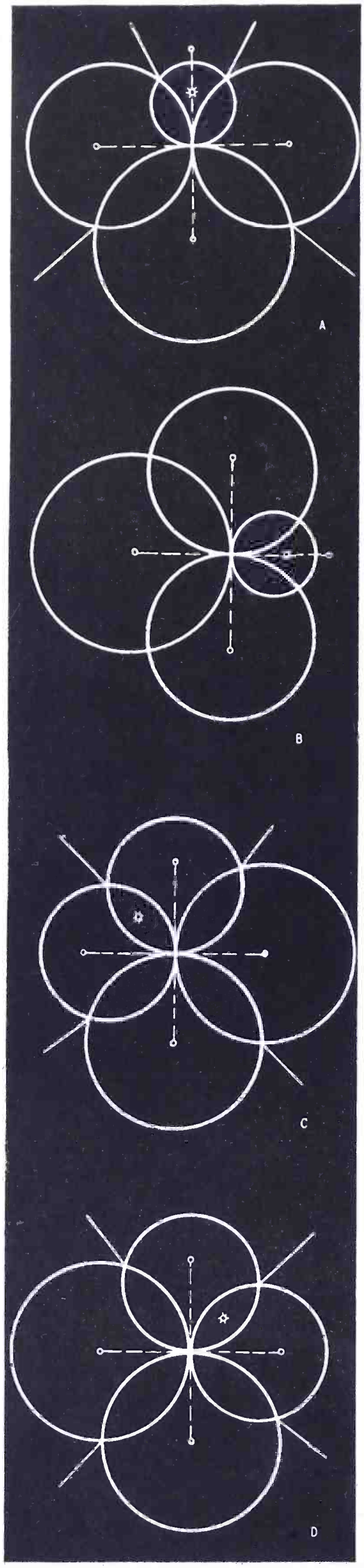
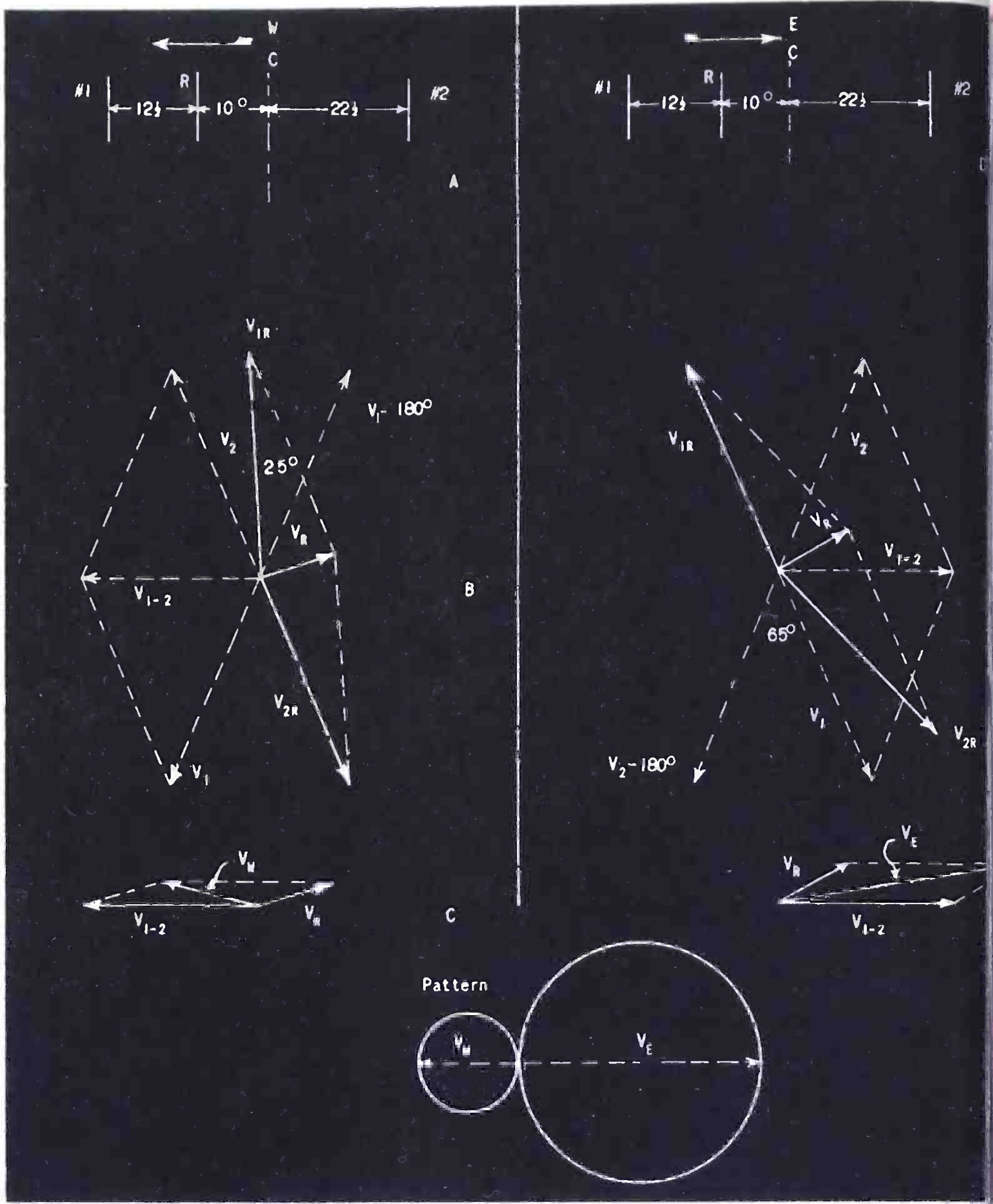




Figures 14 (top) and 15 (left)

Figure 14, effect of course bending antenna (reflector). Figure 15 shows patterns secured due to presence of course bending antenna (reflector). Dashes between small circles indicate loop; asterisks are course bending antenna.

been advanced 5° and the phase of tower 2 has been retarded 5° by transmission line changes.

For radiation to the left of the paper, the phase of tower 1 is advanced 5° by its original phase difference. To this is added the 22½° due to the tower spacing, so that this vector is advanced a total of 27½° from the original position of *A*. Tower 2 is retarded 5° due to its original phase difference, and retarded 22½° by the tower spacing. Its vector, therefore, is retarded a total of 27½°. This produces the resultant field to the left, which is greater than the symmetrical fields of *B*. The actual field becomes

$$F_{1-2} = (V)\ 2 \text{ sine } 27\tfrac{1}{2}° = .92\ V$$

For radiation to the right the phase of tower 1 is advanced 5° its original phase difference and tarded 22½° by the tower spac a net retardation of 17½°. The ra tion of tower 2 is retarded 5° by original phase difference and vanced 22½° by the tower spacin net advance of 17½°. This prod the resultant field to the right wl is smaller than the symmetrical fi of *B*. The actual value becomes

$$F_{1-2} = (V)\ 2 \text{ sine } 17\tfrac{1}{2}° = .6\ (V$$

and the resultant figure-of-eight is longer symmetrical.

D illustrates the effect when original phase displacement is versed. In this case the field rever The larger lobe is always in the rection of the tower which has b advanced in phase.

Appendix C

Course Bending of MRL Station

Figures 14 and 15 illustrate effect of the course bending ante at MRL stations.

Figure 14 is a vector diagram the fields due to the vertical sect

(*Continued on page* 74)

NEWS BRIEFS OF THE MONTH . . .

COPPER AND STEEL SUPPLY IMPROVES

The War Production Board conservation division states in its recent report on the *material substitutions and supply list,* that the supply of copper and steel is improving. The report emphasizes, however, that *easing* applies rather to the raw materials than to facilities or manpower for fabricating them.

According to the report, tin continues to be short in supply, an exception to the general improvement in the supply of nonferrous materials. It is recommended that tin be used as little as possible in bronzes and plating.

As groups, chemicals and plastics are somewhat tighter than on the previous listings, while textiles and fibres remain about the same.

Copies of Issue No. 11 of the *material substitutions and supply list* may be obtained by applying in person at any Regional Office of War Production Board, or by writing on official letterhead to the Editor, Conservation Division, Room 2616 Tempo "D", Washington 25, D. C.

* * *

ARMY-NAVY ISSUES NEW PREFERRED TUBE LIST

A preferred list of tubes, that supersedes the Army-Navy preferred list of vacuum tubes dated March 1, 1943, has just been issued.

* * *

DELORAINE ELECTED TO I. T. & T. BOARD

Edmond M. Deloraine, general director of the laboratories division of Federal Telephone and Radio Corporation, was recently elected to serve as a director on the I. T. & T. board of directors.

Mr. Deloraine has specialized in h-f and u-h-f development. Under his direction in 1931 the first demonstrations of single sideband short-wave radiotelephony were carried out between Buenos Aires and Madrid, and between Madrid and Paris. In 1931 and 1933 he established telephone and printer communications across the English Channel on approximately 1,700 megacycles, using very sharp beams, and in 1936 and 1937 made possible the first multi-channel u-h-f telephone link. Later he used ultrahigh frequency in connection with television transmission, including the construction of the station at the Eiffel Tower, providing the highest power ever used.

ASA AIRCRAFT CRYSTAL STANDARDS

A new standard for quartz crystals (C75. 11-1944) used for control of frequency in aircraft radio equipment has just been approved by the American Standards Association.

This standard, for the first time, coordinates British, Canadian, and American practice in the manufacture of aircraft crystal units.

* * *

RADIO CLUB SEES ELECTRON MICROSCOPE DEMONSTRATION

The electronic microscope was demonstrated and described by Dr. C. H. Bachman of G. E. at a recent meeting of the Radio Club of America, at Columbia University.

* * *

TEST INSTRUMENT ADVISORY COMMITTEE FORMED

The War Production Board has announced the formation of a test instrument industry advisory committee.

Committee members include A. H. Hotopp, Jr., Technical Device Corp., Bloomfield, New Jersey; David Newman, Daven Company, Newark, New Jersey; Paul Jackson, Jackson Elec. Instru. Corp., Dayton, Ohio; E. G. Perkins, Supreme Instrument, Greenwood, Mississippi; V. E. Jenkins, Weston Electrical Instru. Corp., Newark, New Jersey; Milton Reiner, Radio City Products Company, New York, New York; J. J. McCarthy, Triumph Mfg. Company, Chicago, Illinois; and A. J. Lush, Rawson Electrical Instru. Co., Cambridge, Mass.

* * *

BONFIG JOINS ZENITH

Henry C. Bonfig has resigned as commercial vice president of the Radio Corporation of America to join Zenith Radio Corporation, as vice president in charge of the household radio sales division.

* * *

WHITE STAR TO UNITED ELECTRONICS

The United Electronics Co., Newark, N. J., has been awarded a white star for their Army-Navy "E" flag.

* * *

DR. DUFFENDACK JOINS N. A. PHILIPS

Dr. Ora Stanley Duffendack, professor of physics at the University of Michigan, has been appointed director of research of North American Philips Company, Inc. Dr. Duffendack will direct a new research laboratory located in the *Richmond Hill* mansion at Irvington, N. Y.

RADIOMARINE COASTAL STATIONS RESUME SERVICE

Two coastal radiotelegraph stations [of] Radiomarine Corporation of Ameri[ca], WNY, New York and WOE, Flori[da] have resumed operation with ships at [sea] by authority of the FCC.

* * *

GENERAL RADIO ANNOUNCES NEW OFFICERS

Errol H. Locke has been elected pre[si]dent of General Radio Company to su[c]ceed Melville Eastham, who howev[er] will continue as chief engineer in char[ge] of research and development. Harold [B.] Richmond was named chairman of [the] board and chairman of the manageme[nt] committee. Henry S. Shaw, form[er] chairman of the board, continues as director of the company, along with M[r.] Eastham, Mr. Locke, and Mr. Richmo[nd]. Frank L. Tucker, formerly comptrol[ler,] was elected treasurer and secretary, [as] well as a member of the board of dir[ec]tors. Arthur E. Thiessen, formerly co[m]mercial engineering manager, was nam[ed] vice president in charge of sales; a[nd] Charles C. Carey, formerly superinten[d]ent, was named vice president in char[ge] of manufacturing. C. E. Hills Jr. w[as] appointed assistant secretary-treasur[er.] He will also retain his post as comm[er]cial manager.

* * *

LAMME MEDAL AWARDED KEHOE

The 1943 Lamme Medal of the AI[EE] has been awarded to Arthur H. Keh[oe,] vice president of Consolidated Edis[on] Company of New York Inc., "for pion[eer] work in the development of alternati[ng] current networks and associated appa[ra]tus for power distribution."

* * *

PFOHL TO MUTER COMPANY

Paul J. Pfohl, western manager of RC[A] tube division since 1931, has become sa[les] manager of the Muter Company, C[hi]cago, Ill.

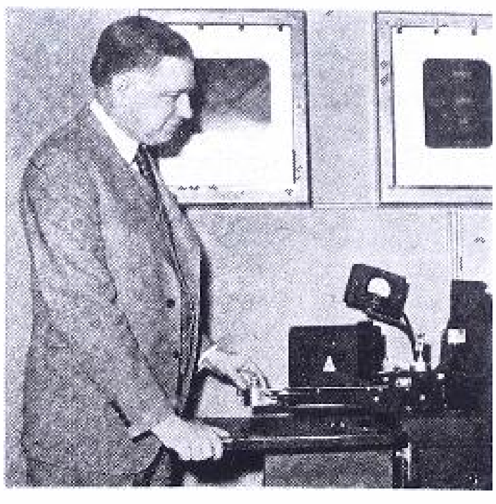
* * *

ANDERSON PROMOTED BY FRANKLIN TRANSFORMER

R. L. Anderson has been appointed ch[ief] engineer of the Franklin Transform[er] Company, Minneapolis, Minn. Mr. A[n]derson was previously associated with [the] Rockefeller Foundation at University

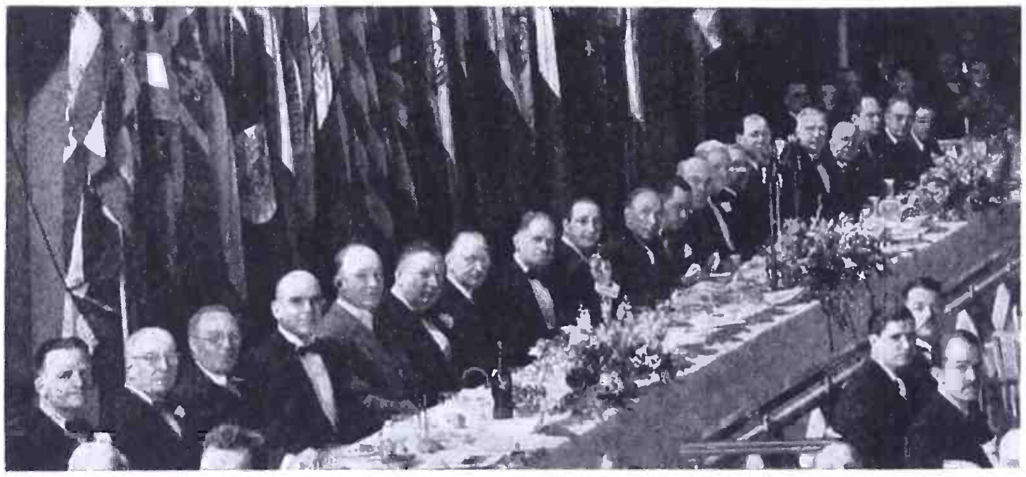
(*Continued on page* 64)

Edmond M. Deloraine



Dr. O. S. Duffendack

R. L. Anderson



At the nineteenth annual dinner-cruise in the Hotel Astor, New York City, left to right: J. R. Poppele, George H. Clark, J. V. L. Hogan, George W. Bailey, J. McWilliams Stone, T. R. McElroy, E. A. Nicholas, Ludwig Arnson, William J. Halligan, Brig. Gen. H. M. McClelland, William J. McGonigle, Rear Adm. Joseph F. Farley, A. J. Costigan, Comdr. Charles W. Horn, Hon. James L. Fly, Major Gen. Joseph O. Mauborgne (USA Ret.), E. H. Rietzke, Lieut. Col. Thomas Mitchell, and Lieut. Comdr. James Flood.

THE nineteenth annual dinner-cruise at the Hotel Astor in New York City, on February 12th, 1944, saw a record attendance. Consular representatives of many of the United Nations and communications specialists from industry and government were present.

Brig. General H. M. McClelland, Air Forces Communications Officer, was present to personally represent General Henry Harley Arnold, Commanding General of Army Air Forces, who received the VWOA Marconi Memorial Commemorative Medal. General Arnold in acknowledging the award, wrote: "The honor bestowed upon me by the awards committee of the Veteran Wireless Operators Association in naming me as the recipient of the Marconi Memorial Commemorative Medal is deeply appreciated. Please convey to your associates that I am most honored that they voted to have the medal struck in my name and that I only regret that I cannot personally express my appreciation to them."

The Marconi Memorial Medal of Achievement was given to T. R. McElroy, president of McElroy Manufacturing Corporation; E. A. Nicholas, president of Farnsworth Television and Radio Corporation; Ludwig Arnson, president of Radio Receptor Company; and William J. Halligan, president of Hallicrafters. Medal presentation ceremonies were broadcast over the NBC network.

Veteran radio men from National Company, Radiomarine, Tropical Radio, Mackay Radio, Hallicrafters, McElroy, Radio Corporation of America, Radio Receptor, Farnsworth, Harvey Radio, Bryan Davis Publishing Company, FCC, W O R-Mutual, National Broadcasting Company, Columbia Broadcasting System, and many others were at the dinner.

Among the pioneers present were: Ben Beckerman of Old Dominion Line fame; E. N. Pickerill, long chief radio officer of the Leviathan and now with RCAC; George McEwen of RCAC; Henry Hayden of Ward Leonard; Sam Schneider, one of our first treasurers and now with the OWI; C. S. Anderson, now retired from Radiomarine, who did so much work on many VWOA Year Books; Ed Content, assistant chief engineer of WOR; Raymond F. Guy, radio facilities engineer of the National Broadcasting Company who was made a VWOA life member; Bill Simon, our very energetic treasurer, who also became a life member; Bryan Davis and Lewis Winner of COMMUNICATIONS; F. P. Guthrie, chairman of our Washington chapter; C. D. Guthrie, supervisor of radio for the Maritime Commission; Haraden Pratt, vice-president and chief engineer of Mackay Radio, who became a life member of VWOA; W. A. Ready, president of the National Company and Arthur Lynch, also of National, who became a life member of VWOA; Guy Entwistle and Walter Butterworth of the Boston chapter of VWOA; John Varian and V. P. Villandre of Radiomarine; Frank Rigby, personnel director of RCA Communications; Lt. Cmdr. A. F. Wallis of the Maritime Service; Charles Maps, wire chief in the New York Telephone Company; Don McNicol, one of the earliest presidents of the IRE; E. H. Price and J. A. Bossen of Mackay Radio; and Frank Orth, technical supervisor of the Columbia Broadcasting System, who became a life member of VWOA; and many, many others.

Music was supplied by a band composed of workers in the McElroy plant.

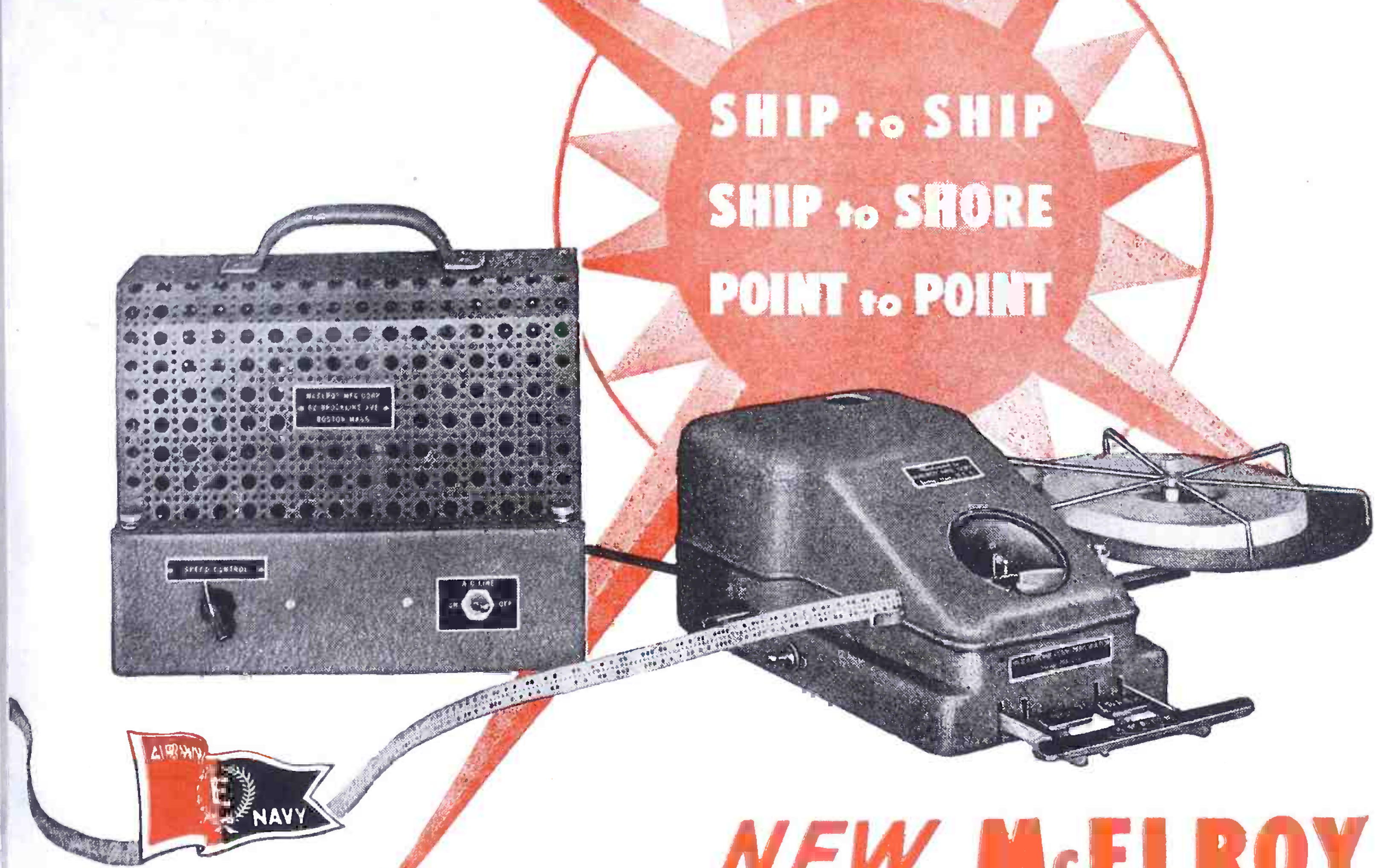
Year Book

COPIES of the 1944 VWOA Year Book are available from headquarters at fifty cents per copy, to cover cost of packaging and mailing. It is the largest issue in our long history and contains some very interesting biographies on the men who received the honors at the dinner-cruise. Interesting facts about pioneers and charter VWOA members also appear. The issue is well worth having and we urge you to get your copy now, while the supply lasts.

VWOA Spring Meeting in New York

ON April 27th members of the VWOA will gather at the 77th Division Club, 28 East 39th Street, New York City, to hear George Clark talk on his famous collection of hieroglyphics.

IMPROVED radiotelegraph communications from



NEW McELROY

ELECTRONIC CODE TAPE PERFORATOR PFR-443-A

Every ship and every marine station should and probably will, eventually, be equipped with perforator and transmitter.

Marine radiotelegraph communications were given a substantial boost in efficiency when Ted McElroy perfected the McElroy Code Tape Perforator PFR-443. This new model . . . the PFR-443-A...represents an even greater advancement in reducing the human margin of error. Entirely mechanical, the PFR-443-A comprises two units: The Keying Unit which is made up of an assembly of two keys, the space bar and punching mechanism; plus the Electronic Unit which relieves the keying contacts of high current and voltage, and provides for utmost ease in operation.

Operation of the PFR-443-A is similar to a semi-automatic ("bug") key excepting that the transmission of dots and dashes is automatic. A light touch actuates the punching mechanism for as long as either the key or space bar is depressed. Experienced operators can maintain, with ease, an accurate speed of more than 40 words per minute . . . in all Morse combinations assigned to the Russian, Turkish, Greek, Arabic and Japanese alphabets and languages, which are not found on the keyboards of standard perforators manufactured in the United States and Great Britain. ***Write for additional information.***

McElroy engineers never imitate...never copy
We CREATE... DESIGN... BUILD

McElroy MANUFACTURING CORPORATION
82 BROOKLINE AVENUE
BOSTON, MASS.

KEEP IT UP...BUY MORE AND MORE WAR BONDS

NEWS BRIEFS

(*Continued from page 60*)

Minnesota, and with CBS in Minneapolis as research specialist on ultrashort-wave experiments.

* * *

MRS. KINZIE NOW ASS'T RTPB SECRETARY

Mrs. Martha Kinzie, in radio since the pioneering days of 1920, has been elected assistant secretary of the Radio Technical Planning Board. Mrs. Kinzie has been Dr. W. R. G. Baker's secretary for many years, and is at present working with him in G. E.'s electronics department at Bridgeport, Connecticut.

* * *

EIGHT NEW I. T. & T. OFFICERS

Four vice presidents and four second vice presidents have been appointed by the board of directors of International Telephone and Telegraph Corporation. The new vice presidents are: H. C. Roemer, vice president and comptroller of Federal Telephone and Radio Corporation; W. H. Freng, assistant general attorney, who has also been appointed solicitor; Charles D. Hilles, secretary of the corporation; and Francis White, vice president of International Standard Electric Corporation. The second vice presidents appointed are: F. F. Davis, H. H. Buttner, G. A. Ogilvie, and Leonard Jacob II.

* * *

SYLVANIA SHIFTS EMPORIUM PLANT

Sylvania Electric Products' industrial apparatus plant at Emporium, Pennsylvania, has been transferred to plant two at Williamsport, Pennsylvania. Virgil M. Graham will continue as manager.

* * *

RADIO CLUB HEARS WHITE ON V-H-F

S. Young White, consulting engineer with the General Communication Company of Boston, spoke recently on v-h-f tuning systems before the Radio Club of America meeting at Columbia University. His paper covered theory and economics of tuning by parallel and concentric lines, cups, sliding contact, coils, core tuning, etc.

* * *

WILSON NOW AAC VICE PRESIDENT

Ira Stuart Wilson has been elected vice president in charge of finance of Aircraft Accessories Corporation. Mr. Wilson will maintain his headquarters at the electronics division of AAC in Kansas City, Kansas.

* * *

ZENITH OPENS N. Y. SUBSIDIARY

Formation of the Zenith Radionics Corporation of New York, a subsidiary of the Zenith Radio Corporation, with offices in the Empire State Building, was recently announced.

H. J. Wines, for the last 17 years general sales manager of the New York and Chicago branches of the Frigidaire division of General Motors Corporation, has been appointed general manager and a director of the New York corporation. E. F. McDonald, Jr., president of Zenith Radio Corporation, will also head the new corporation. Toni Strassman has been appointed sales manager of the Zenith Radionic hearing aid division of the subsidiary.

Ray Hoeffler, for the last two years New York regional manager for Zenith, has been promoted to manager of the Zenith Radio Distributing Corporation, wholesale distributor of Zenith products in the Chicago area.

* * *

BRYANT ELECTRIC WINS "E"

In a dual ceremony at Bridgeport, Conn., two divisions of the Bryant Electric Company were awarded Army-Navy "E" flags.

The wiring device division received its flag from Capt. Carl H. Bushnell, USN; the plastic division was presented with a flag by Col. Arthur H. Rogow, USA.

* * *

AMP SOLDERLESS WIRING DEVICE CATALOG

A 72-page sectionalized catalog of AMP solderless wiring devices, SD-I, has just been published by Aircraft-Marine Products Inc., 1591F North Fourth Street, Harrisburg, Pa.

This manual includes illustrated data on solderless insulation support terminals, standard and flag type terminals, connectors, lighting contacts, cable lugs, wiring plugs, standard and quick disconnect bonding jumpers, splicing terminals, grid clips, etc.

Hand tools and installation presses are also described and illustrated, and complete working instructions for setting up crimping dies are included.

Data, specification and dimensional charts, and Army-Navy and commercial wire sizes are also supplied.

* * *

C. L. PUGH NOW UTAH REPRESENTATIVE

C. L. Pugh has been appointed Utah's wholesale and sound divisions representative in the states of Ohio, West Virginia, and western Pennsylvania. His offices are at 2009 Elmwood, Columbus 8, Ohio. Mr. Pugh was formerly connected with Hughes Peters, Inc., RCA, and the Standard Transformer Company.

* * *

AERO NEEDLE ENLARGES

Aero Needle Company has announced

(Continued on page 66)

NEWS BRIEFS

(*Continued from page 65*)

the completion of their new and larger quarters at 208 East Erie Street, Chicago, Ill.

* * *

O'BRINE JOINS RCA INFORMATION STAFF

Jack O'Brine has joined the RCA information department. Mr. O'Brine was formerly associated with Popular Science Monthly, the New York Herald Tribune, and NBC.

* * *

KELLEY-KOETT'S COVINGTON PLANT ENLARGED

An addition to the Kelley-Koett Manufacturing Company's plant at Covington, Kentucky, has been completed.

A new sales and service office has been opened at 1109 Traction Building, Cincinnati, Ohio. This office will serve as headquarters for Ted Tierney, representative for the Cincinnati area; Floyd Tracht, representative for southern Ohio; and Wayne Kruse, service representative.

* * *

HILLIARD NOW G-M OF BENDIX RADIO

W. P. Hilliard has been named general manager of the radio division of Bendix Aviation Corporation, Baltimore, Maryland and Red Bank, New Jersey. Mr. Hilliard, who has been director of sales and engineering of the radio division since its inception in 1936, succeeds Hugh Benet who will assume other Bendix duties.

* * *

WILLIAM PURVIS TO CBS

William Purvis, who received an honorable discharge from the Navy recently, has joined CBS as a short-wave transmitter technician at the Wayne, New Jersey plant. Mr. Purvis was previously connected with the CAA, WBAX in Wilkes-Barre, Pennsylvania, the American Airlines, and the Radiomarine Corporation of America.

* * *

SENNOTT NEW DETROLA CONTROLLER

John Harmon Sennott, supervising accountant for Ernst & Ernst, Chicago, since 1932, has been appointed controller of the Detrola Radio Division, Detroit.

* * *

KLINGENSCHMITT ELECTED PRESIDENT OF RADIO CLUB OF AMERICA

F. L. Klingenschmitt has been elected president of the Radio Club of America. O. James Morelock was elected vice president; J. J. Stantley is the new treasurer; newly elected corresponding secretary is

M. B. Sleeper; and J. H. Bose is now recording secretary.

* * *

CLARK OF FORMICA ON NACA WOOD-PLASTICS SUBCOMMITTEE

George H. Clark, vice president of Formica Insulation Company, Cincinnati, has been appointed to the Subcommittee on Wood and Plastics for Aircraft of the National Advisory Committee for Aeronautics for 1944.

The NACA was established by Congress as an independent research and fact-finding agency associated with the development of aircraft in this country.

Chairman of the Wood and Plastics Subcommittee is Dr. Gordon M. Kline, National Bureau of Standards, Washington, D.C. Other appointments to this committee for 1944 are: Captain R. M. Houghton, USA, Materials Laboratory, Wright Field; Lieut. Commander S. L. Chisholm, USN, Bureau of Aeronautics; Henry Sang, Sircraft Materials Laboratory, Naval Air Experimental Station, Philadelphia; Paul C. Speiss, CAA, Washington; L. J. Markwardt, Forest Products Laboratory, Madison, Wisconsin; Evan H. Schuette, Lumber Manufacturers Association Laboratories; C. W. Armstrong, Marco Chemicals Inc., Sewaren, New Jersey; and Dr. Harry C. Engel, assistant design engineer, Glenn L. Martin Company, Middle River, Maryland.

George H. Clark

* * *

ARHCO PLANT NOW IN MT. VERNON

A new American Radio Hardware Company plant has been opened at 152 MacQuesten Parkway, South Mount Vernon, New York.

* * *

J. KELLY JOHNSON JOINS HAMMARLUND

J. Kelly Johnson, former production section chief of the Electronics Division, Office of Procurement and Material, Navy Department, Washington D.C., has been appointed executive engineer in charge of all engineering activities at Hammarlund Manufacturing Company Inc., 460 West 34 Street, New York City.

Prior to his Washington post, Mr. Johnson was assistant chief engineer of Silver-Marshall & Company, Chicago; development engineer at Hazeltine Service Corporation, New York; and chief engineer for Wells Gardner & Company, Chicago.

* * *

UNIVERSAL MICROPHONE CO. LTD. NOW A PARTNERSHIP

The Universal Microphone Company Ltd., of Inglewood, California, has dissolved its

(*Continued on page 68*)

NEWS BRIEFS

(Continued from page 67)

corporation and hereafter will be known as the Universal Microphone Company.

James L. Fouch and Cecil L. Sly, president and vice president of the former corporation, have re-organized the company as a partnership.

* * *

CROWLEY RECEIVES WHITE STAR

A white star for the "E" flag has been awarded to Henry L. Crowley & Company, Inc., West Orange, N. J.

* * *

WEBSTER PRODUCTS EXPANDS

Additional floor space will soon be added to the plant of Webster Products, 3825 W. Armitage Ave., Chicago 47, Ill.

* * *

BURNDY CONNECTOR CATALOG

A 64-page catalog, 6050, of indent type electrical connectors has been published by the Burndy Engineering Co., Inc., 107 Eastern Boulevard, New York 54, N. Y.

* * *

CANNON BATTERY CONNECTORS BULLETIN

A 24-page bulletin illustrating and describing battery connectors used with battery carts for engine starting, for the quick disconnect of large storage batteries, general service batteries, and rack battery installations, has been issued by Cannon Electric Development Company, 3209 Humboldt Street, Los Angeles 31, Calif. Application photos and condensed data sheets are also included.

* * *

ALTEC-LANSING TO DISTRIBUTE FOR HANOVIA

The Altec-Lansing Corporation of Hollywood, California, has been appointed exclusive western, Chicago and Detroit distributor of Hanovia Chemical and Manufacturing Company's products.

* * *

"E" AWARDED TO ERCO

The Army-Navy "E" was awarded recently to the Erco Radio Laboratories, Inc., Hempstead, New York. Commander S. J. Singer, USNR, executive officer, Industrial Incentive Division, presented the "E" flag to Edward Ruth, III, secretary-treasurer of Erco.

* * *

DU MONT FIVE-YEAR CLUB INAUGURAL

A dinner-dance was held recently to inaugurate the DuMont Five-Year Club of employees of the Allen B. DuMont Laboratories, Inc., Passaic, New Jersey. Allen B. DuMont, president of the company, awarded five-year pins to all employees with his organization for five or more years.

Ten-year gold pins were presented to A. J. Hinck, Stanley J. Koch, and Harry Gawler, the three men who worked with Mr. DuMont in the latter's home basement in 1932 on the first DuMont tubes. And Mr. DuMont himself received a ten-year pin. Toastmaster for the gathering was Leonard F. Cramer, vice president and general sales manager.

* * *

PHILCO PROMOTES HARDY

Larry F. Hardy has been elected vice president in charge of the home radio division of Philco Corporation. Mr. Hardy, who has been with Philco since 1932, was formerly with Landers, Frary and Clark. He was also New York City manager of motor appliances for GE and mid-western division manager for Ceco Manufacturing Company.

* * *

FORMICA PLASTICS FILM SHOWN IN N. Y.

A sound film in color prepared by The Formica Insulation Company, Cincinnati, Ohio, depicting the history of the laminated plastics industry, had its premiere showing in New York recently.

The film, entitled *The Formica Story* reviewed the development, design and application of laminated plastics, during the past fifty years.

The sound film on 16mm is available for meetings of engineers, executives, employee organizations, and for plastic classes in colleges.

* * *

BALTIN APPOINTED TO TBA

Will Baltin, former program manager of the DuMont television station W2XWV, has been appointed secretary treasurer of Television Broadcasters Association. Mr. Baltin will handle the business affairs of the association, and coordinate the activities of member television companies for the advancement o elevision during wartime and in the post-var period.

* * *

CONTINENTAL ELECTRIC ELECTRONIC TUBE CATALOG

A catalog, RS, with data on phototubes, ectifier tubes, and grid control tubes thyratrons) has been published by Continental Electric Company, Geneva, Illiois.

* * *

"OCCUPATIONS IN ELECTRONICS" OLDER

A 6-page folder on *Occupations in Electronics* by Forrest H. Kirkpatrick of Bethany College, and John E. Crawford f RCA, has been published by Occuational Index, Inc., New York Univrsity, New York 3, New York. The older describes the job opportunities in ostwar electronics. Single copies are 25c.

* * *

SPRAGUE ISSUES "FOR IMMEDIATE DELIVERY" PAMPHLET

A listing of *immediate delivery* items is ncluded in a 4-page pamphlet being disributed by Sprague Products Company f North Adams, Massachusetts. Included in the listings are bathtub type metal rectangular units in single and ual capacities and in voltages from 50 o 1750 volts d-c (tolerance is said to be minus 20%, plus 30%); oil-filled, oilmpregnated can type units range in capaities from 1.0 to 17 mfd and in a variety f a-c/d-c voltages. Mica condensers vailable for immediate delivery include nits in d-c test voltage range of 1000, 500, and 5000.

* * *

ABRASION TEST FOLDER

A repeated-scrape abrasion tester for omparing the toughness of insulation on lm-coated magnet wires is described in G. E. folder, GEA-4166.

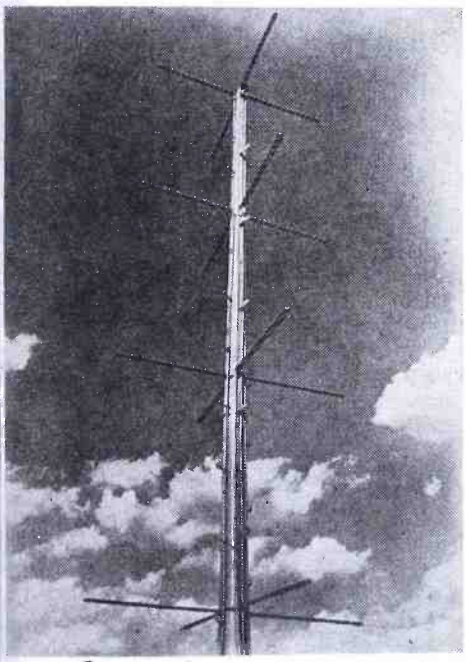
* * *

SOLAR EMPLOYEE JOURNAL

A journal covering activities of emloyees is now being published monthly y the Solar Manufacturing Corporation. Sylvan A. Wolin is editor.

•

F-M TURNSTILE ANTENNA



(*Courtesy Andrew Company*)

recent installation in the Zenith F-M staon WWZR atop the Field Building in hicago. Feeding the four bays of the turnile antenna are eight 1 3/8" diameter coax-l cables. These lines as well as the 4 1/8" iameter cables feeding power from the ansmitter, are used in a "back-to-back" onnection to provide a balanced 140-ohm transmission line.

WHEATSTONE BRIDGE

(*Continued from page* 52)

kinds, but come under only a few general classifications. These are:

(1)—Grounds of one or more wires, which may be of high, low or intermediate value.

(2)—Shorts or crosses of one or more wires of a pair, crossed or shorted to one or more wires of a line or cable.

(3)—Opens, which may affect one or more wires of a pair, quad or an entire line or cable.

(4)—Combinations of grounds, crosses or opens, which may affect one or more wires of a line or cable.

For a solid short condition, a loop measurement is indicated. This cannot be relied upon until it has been verified by measurement from both ends into the trouble, and by comparison with a spare or another pair of like gauge, that the trouble is solid and free from variation caused by contact resistance. The grounded Varley or the metallic Varley may also be used where possible to check the resistance in the short, if any exists. If none exists, the Varley measurement will indicate zero.

For a variable short circuit condition, the best method to use is the metallic Varley using a split pair. This places the variable resistance in the battery branch of the circuit, thus removing it from the balancing arms. The accuracy of location will suffer to a greater and greater degree as the resistance of the fault increases since under this condition, the sensitivity of the bridge circuit goes down. This would necessitate using a higher voltage battery which is not desirable for two reasons. One is that the hazards of ruining either the galvanometer or the resistances or both are too great in the event that the high resistance suddenly changes to a low resistance. The other is that whenever the high resistance is caused by poor contact at the fault, the excess current may cause the trouble to suddenly disappear only to reappear with the removal of the battery. In all cases of a high resistance contact such as may be caused by very fine wire or very rusty or corroded wire, etc., a low battery potential should be used to avoid the burning out of the contact.

Open measurements should be made using the bridge set up for Murray loop measurements and preferably using a very low a-c frequency of 4 to 20 cycles per second as previously outlined. It is important that the insulation of the faulty wire be good, but if an appreciable leak is found, less ccurate results are only to be expected.

In all of the cases that have been nalyzed and discussed, it has been citly assumed throughout that the ffect of intermediate connecting cable, pen wire, and equipment with their ccompanying variations in resistance ith changes in weather and temperaure have been taken into account. he theoretical resistances and disnces apply to the conductors only. , for example, a circuit is composed 100 feet of 22-gauge cable from the cation of the measuring bridge, 800 et of 19-gauge cable from the office ible terminal, 27 miles of 104 copper, 5 miles of 16-gauge cable at the far nd of the open wire to the other terinating office, 200 feet of 20-gauge able and finally terminating into a ansformer or repeating coil, account ust be taken of each length of copper aired cable and equipment. The type measurement may in this case be a op followed by a Varley if one wire the open wire pair is grounded. rom the total loop measurement of he bridge, the excess amounts of quipment and cable resistance must e deducted. This gives the actual esistance of the open wire pair which ould be approximately 270 ohms. A arley measurement made from the ome office or near end would include he loop resistance of the open wire us the total loop resistance of cable, onnecting wire, and repeating coil indings at the far end. The differnce of the loop and Varley measurenents would cancel out the excess able and equipment resistance at the ir end, leaving only the near end to e accounted for. This difference obained would have deducted from it he excess cable and bridge lead restance giving the actual resistance to he trouble in ohms. The quotient of he loop minus the Varley, and the oop corrected as outlined will give he fractional part of the total disance to the trouble from the measurng or near end. This when multiplied the total length of the open wire ill give the actual miles to the troue from the junction of the open wire ith the near end cable. If the total ngth of the open wire is not known, he temperature coefficient in ohms er loop mile may be used. In this nstance, the quotient of the corrected oop minus Varley and the ohms per oop mile will give the distance to the ouble.

nsulation Resistance by the oltmeter Method

Although this type of testing is not rectly allied with the Wheatstone ridge circuit, it is so closely tied up

(Continued on page 72)

WHEATSTONE BRIDGE

(Continued from page 71)

with the applications of that circuit in practically all of the measurements described, that it is included here for the sake of continuity and completeness.

Let us consider Figure 12, in which a battery and a high resistance voltmeter are connected in series to a wire *n* miles long, having assumed uniform leakage to ground throughout its length. If we let internal resistance of the battery be assumed as negligible, in comparison with the magnitude of the resistances being measured, and

R = resistance of the voltmeter, in ohms,

D = deflection of the meter indicator, in volts, and

X = total line leakage, in ohms;

$x = nX$ = leakage per mile, in ohms, or $n\,X/10^6$ megohms per mile.

The series resistance of the wire itself is also considered to be negligible in comparison with the leakage resistance. By Kirchoff's second law, the equation of voltage equilibrium is

$$E = R\,i + n\,X\,i = i\,(R + n\,X) \qquad (97)$$

Since the deflection of the meter is proportional to the potential across the meter, or $D = R\,i$, each side of *97* may be multiplied by *R*, from which we get,

$$E\,R = R\,i\,(R + n\,X) = D\,(R + n\,X) \qquad (98)$$

and since, the insulation resistance per mile, $x = n\,X$, we may solve for x in *98* using this substitution, or

$$x = n\,R\left(\frac{E}{D} - 1\right) \text{ ohms per mile} \qquad (99)$$

which may be converted to the more common form of

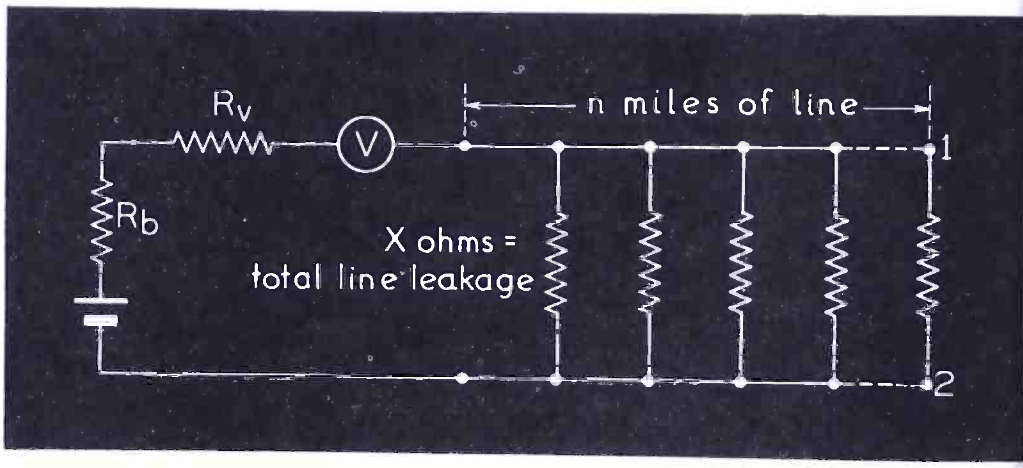
$$x = \frac{n\,R}{10^6}\left(\frac{E}{D} - 1\right) \text{ megohms per mile} \qquad (100)$$

If a limit of M-megohms per mile is set as the allowable limit below which the insulation shall not be permitted to fall, then with $x = M$, and *100*, the deflection maximum is

$$D = \frac{10^{-6}\,n\,E\,R}{M + 10^{-6}\,n\,R} \text{ volts} \qquad (101)$$

If the minimum insulation resistance is set at 25-megohms per mile and a meter having 100,000 ohms is used with a battery of 100 volts, the deflec-

Figure 12
Insulation resistance measurement using a high resistance voltmeter.



tion obtained will be

$$D = \frac{10\,n}{25 + 0.1\,n} \text{ volts} \qquad (102)$$

For example if a wire 50 miles long were tested, the highest reading that could be allowed without exceeding the limits required for a satisfactory commercial line would be 16.7 volts, while a 100-mile line should not give a reading in excess of 28.6 volts.

Since the sensitivity of these high resistance voltmeters is fairly good, they may be used to check for crosses, shorts and grounds on cables and lines. The Wheatstone bridge circuit would then be used for the actual location as outlined in the preceding theory. The exact application which would have to be used in any particular case would be determined largely by the nature of the trouble as revealed by the voltmeter. Of course, if the voltmeter is absent, a listening test may reveal that trouble exists, but it is most frequently a poor criterion. The best procedure in this instance would be to go ahead with loop and Varley tests. The loop may then be either longer or shorter than expected, or the Varley may read either zero or some definite value. If the loop is longer than expected, a series unbalance or an open condition in which both wires have made contact with a ground path is indicated. If the Varley reads zero, then either a ground at the far end with a short, or a short alone, or contact with a foreign wire which causes a capacity unbalance without a direct ground path is indicated. If a loop indicates shorter than expected, a shorted pair is most certainly indicated, the nature of which has to be determined by one of the methods outlined. The short may be solid, intermittent solid, greater than zero and solid, or greater than zero and intermittent or variable, or both. Thus from a few considerations of the simplest of the conditions which might be met with in practice, it can be appreciated that the better knowledge of theory of the bridge circuit that one possesses, the more accurate may be the location of a fault consistent with speed and expediency.

There is no substitute for experience in dealing with the multitude of possible arrangements and faults that may develop when many circuits are in close proximity. The general headings of the causes of trouble run into the hundreds, while the specific causes multiply this figure by several times. The author has had practical field experience in locating line faults and thus has a deep appreciation of the difficulties which beset those upon whom the responsibility for clearance and maintenance of the lines of communications rest.

RADIO RANGE BEACONS

(*Continued from page* 58)

of the loop (indicated as antennas *1* and *2*) and the course bending antenna *R*. In this example it is assumed that the electrical spacing between the vertical sections of the loop is 45°, and that the reflector (course bending antenna) is spaced 10° to the left of the center of the loop.

The vertical antennas *1* and *2* have a resultant field V_{1-2} to the left (see *Appendix B*). In addition each antenna energizes the reflector 180° out of phase with its own field.

As the energy due to antenna *2* passes toward the left it energizes the reflector which, in turn, produces a field 180° out of phase with the field which energized it. Therefore, V_{2R}, the reflector field due to antenna *2*, is drawn in 180° out of phase with V_2, the vector of antenna *2*.

The energy due to antenna *1*, however, must travel to the reflector, is reversed 180°, and travels back to antenna *1* before it can regain the field which energizes it. In this time the net phase difference becomes 180° — 25°. The 25° represents the time taken for the energy to travel to the reflector and back. V_{1-2} is therefore drawn in, reversed 180° from V_1 and retarded an additional 25°.

By completing the parallelogram, V_R, the net reflector voltage is secured. The sum of V_R and V_{1-2} produces V_W, the resulting field to the left (or west). This is shown in *C* so as not to confuse *B*.

The radiation to the right is analyzed in a similar manner. In this case the reflector field (V_{1R}) due to antenna *1* is 180° out of phase with the field that produces it (V_1). The field due to antenna *2* is 180° out of phase, with an additional retardation of 65°, the time the energy takes to travel from antenna *2* to the reflector and back again. The resulting field to the right (or east) is V_E.

The resulting pattern for the case in point indicates that the smaller lobe of the non-symmetrical figure-of-eight is in the direction of the reflector.

Figure 15 illustrates various field combinations for various locations of the reflector.

AF COMMUNICATIONS

(*Continued from page* 26)

s, both giving push-button selection four channels. One works on the band and the other is a v-h-f job ed to all fighters and close-support craft.

This fighter set, model TR1143, has compact box-like chassis consisting four sections which plug together. ese sections include the receiver, nsmitter, audio amplifier and annel-selector mechanism. It weighs pounds and is both easy to install d service.

Tubes are the familiar 6 volt, 3 npere octal-based types. Their se- ence in the receiver is . . . radio equency amplifier, crystal-controlled cillator working in the 6-mc band, times multiplier, cathode-injection ixer, three intermediate frequency ages operating at about 9 mc, and uble-diode-triode for demodulation, tomatic volume control, and first low equency amplification.

The transmitter uses a 6-mc crystal- ntrolled oscillator followed by two ebler stages and a doubler stage. A ish-pull power amplifier stage gives watts r-f output.

The amplifier-modulator unit con- ins a low frequency output pentode d a push-pull modulator circuit for e transmitter. An impulse-type motor employed in the push-button channel lector mechanism.

Power is drawn from a rotary con- rter run from the *floating* aircraft ttery. Heater current as well as plate d bias supplies are taken from sepa- te generators. Ample filtering sys- ms are provided.

With this transmitter the range is 20 miles from 10,000 feet, but greater stances are covered, particularly om higher altitudes. The range is *ptical* and rests very largely on the cation of the aerial and the height of e aircraft.

Royal Air Force sets are built to ithstand temperatures between -30° d + 60° C and are tested in low essure chambers and in heavy at- ospheric humidity. They are proof gainst dust and sand.

With light weight they combine rength and reliability. In one month 1943, code transmitting unit failures Britain's Bomber Command were ss than .1 per cent of sorties flown.

•

PAPER IS VICTORY POWER! CON- ERVE IT. DON'T WASTE IT!

•

BOOK TALK . . .

RADIO MATERIEL GUIDE

By *Lieut. Francis E. Almstead*, USNR, U.S. Naval Training School, Noroton Heights, Connecticut, and Comdr. *F. R. L. Tuthill*, USNR, Eastern Sea Frontier Command . . . 242 pp. . . . New York: McGraw-Hill Book Company, Inc. . . . $2.00.

A condensation of fundamentals so effective in rapid training techniques serves as the basis of this compact volume. The treatment and data presented closely parallels that offered in the courses of study of Army and Navy radio schools.

The text opens with a review of mathematics needed in elementary radio, such as frequency determination, impedance, power factor, etc. Other chapters cover the nature and control of electrical energy, measuring instruments, sonic wave behavior, circuits and their calculation, radio parts and their care. More than one-hundred and fifty illustrations accompany the text.

The book is recommended as a reference manual for practicing radio operators, as well as prospective operators and men preparing for the armed services.—*J. M. L.*

• • •

MAINTENANCE AND SERVICING OF ELECTRICAL INSTRUMENTS

By *James Spencer*, Director, Instrument and Relay Department, Meter Division, Westinghouse Electric & Manufacturing Company, Newark, New Jersey: *Edited by Major M. F. Behar*, Editor, Instruments . . . 256 pp. . . . Pittsburgh, Pennsylvania: Instruments Publishing Company . . . $2.00.

The material contained in this study is a reprint of a serial published in *Instruments* in 1941, and was originally developed from class notes which the author used in teaching the subject to trade schools and to new personnel of an instrument manufacturer.

Chapters list general type of instruments in common use, with simple explanations of their principle of operation and detailed instructions on maintenance and repair. The repair of parts and the making of new parts is not dealt with, except for a chapter on pivot and bearing troubles.

Covered are direct-current instruments, a-c ammeters and voltmeters, a-c wattmeters, frequency meters, instrument transformers, synchroscopes, power factor and rva meters, reactive factor meters, and others, with 274 charts, diagrams and photographs.—*O. R.*

• • •

THE RADIO AMATEUR'S HANDBOOK

By The Headquarters Staff of the American Radio Relay League . . . 664 pages . . . West Hartford, Connecticut: The American Radio Relay League, Inc. . . . $1.00 (paper cover).

In this 1944 edition appears an expansion and revision of the chapters on fundamental principles and design.

The *construction and data* section of the book closely follows the pattern of recent editions, except for the addition of a chapter on *Carrier-Current Communication.* The War Emergency Radio Service chapter has been completely rewritten and considerably expanded.

Also revised are the classified vacuum-tube tables, which this year include some fifty new tubes. A supplementary cross-index of tubes by type numbers has been added.—*J. M. L.*

• • •

MATHEMATICS OF RADIO COMMUNICATIONS

By T. J. Wang, Ph.D., Instructor in Electrical Engineering, Ohio State University . . . 371 pp. . . . New York: D. Van Nostrand Company, Inc. . . . $3.00.

For the student or engineer who wishes to study communications, but cannot devote the time necessary to the study of all the associated mathematics, this book will prove quite useful. Only those parts of algebra, trigonometry, calculus, etc., necessary for understanding fundamental electric and radio theory are stressed.

There are 15 chapters covering laboratory work with graphical representations and practical computations, elementary direct current theory and circuits, and elementary alternating current theory and circuits.

The mathematical coverage includes simple equations, algebra, quadratic and simultaneous equations, exponents, square root, trigonometry, radian measure of angles, vectors and vector forms, logarithms, differential and integral calculus and Fourier series.

While intended primarily for students, this book is an excellent review for the radio engineer who wants to brush up on his mathematics.

Answers to the exercises are not included, but the chapters contain numerous examples of mathematical operations for the branches covered.—*A. M.*

THE INDUSTRY OFFERS . . . — . . . — . . . —

G. E. FLUXMETER

A light-beam instrument, the fluxmeter, for measuring either flux density or the total magnetic flux in magnetic circuits has been announced by G. E.

The instrument, applicable wherever permanent magnets or d-c electromagnets are used, is similar to a light-beam galvanometer and has a suspension-type element. Different sensitivities are available by utilizing galvanometers of various characteristics. These galvanometers are interchangeable. The instrument has a scale of 50-0-250 millimeters.

The control box, which contains a dry cell, resistors, rheostat, push button, and a switch, provides a means of introducing a voltage into the electric circuit to compensate for small but undesirable spurious voltages that may otherwise cause errors in high-sensitivity instruments of this type. The control box provides a means of returning the spotlight index to the zero position after a measurement has been made.

A direct current, moving-coil galvanometer of the compensated type is used. In this design, the restoring torque of the suspension is, for all practical purposes, eliminated. After a flux change has been indicated, the light beam index remains at the point, giving ample time for accurate readings.

* * *

IDEAL COMBINATION ETCHER AND DEMAGNETIZER

A new tool that is ready for instant use as either an etcher or demagnetizer has been announced by The Ideal Commutator Dresser Company, 4025 Park Avenue, Sycamore, Illinois.

Fourteen heats are provided: lo-90, 150, 200, 250, 350, 450, 600; hi-300, 400, 500, 650, 850, 1100, 1350 watts. This wide range is useful for marking all iron, steel and their alloys from small delicate parts up to large smooth castings.

Overall dimensions are 8¼" x 11½" x 8⅝". Size of work-plate, 8¾" x 7¼". Includes 5' primary cord and plug, heat-resisting etching tool with 5' lead, extra point, ground clamp and 4' lead. Weight, 39 lbs.

* * *

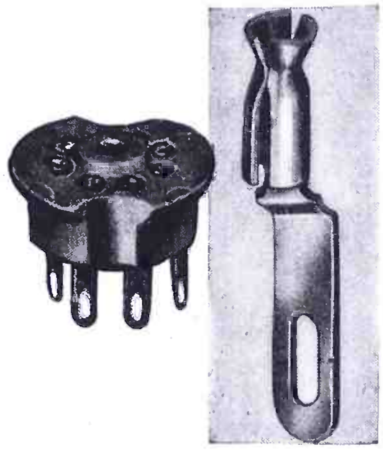
OXFORD-TARTAK CABLE CLAMP

A compression type cable clamp, O-T, has been announced by Oxford-Tartak Radio Corporation, 3911 South Michigan Avenue, Chicago 15, Illinois.

EBY MINIATURE TUBE SOCKETS

A miniature tube socket with micro-processed beryllium copper silver-plated contacts has been developed by Hugh H. Eby, Inc., 18 West Chelten Avenue, Philadelphia 44, Pa.

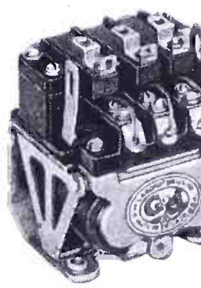
The socket is available in two types. One is a low loss type with Navy grade G steatite casting, having loss factor of .016 or less when tested in accordance with ASTM D 150-42T. Its capacity is said to be 1.5 mmfd or less at 10 mc. The second type is a general purpose unit with mica filled plastic casting having a loss factor of .05 or less when tested in accordance with ASTM D 150-42T. Its capacity is said to be 5 mmfd or less at 10 mc.



* * *

G-M LAB SENSITIVE POWER RELAYS

Relays for circuits requiring small multi-pole relays have been developed by G-M Laboratories. Known as the type 27 sensitive relays, they are available in 1, 2, and 3-pole sizes, approximately 1⅝" x 1 11/16" x 2" high. Five-pole relays are also available. These are approximately 2" x 2½" x 2⅛" high. Weight, 4½ to 6½ ounces; vibration and acceleration: 10 g.; contact capacity, up to 10 amperes, depending on available power and on coil and circuit characteristics; contact pressure, 30 to 50 grams; temperature, humidity, elevation and salt spray test to usual aircraft specifications.



* * *

KIRKLAND RECESSED-LAMP INDICATING LIGHT

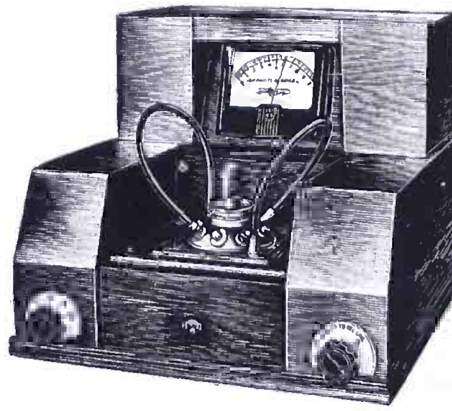
A pilot light for either panel or switchplate mounting with a recessed bulb has been developed by the H. R. Kirkland Company, Morristown, N. J.

According to the manufacturers, this unit, S11, provides such exterior ventilation to the lamp bulb that a much greater lamp-life is to be expected. It is available in all colors.

INTERELECTRODE CAPACITY INSTRUMENT

A meter that provides measurement of vacuum tube interelectrode capacities on a production basis has been announced by the Technical Apparatus Company, 1171 Tremont Street, Boston 20, Massachusetts. The instrument identified as model 37 interelectrode capacity meter has a range from .001 to 100.0 mmfd. Accuracy is said to be 5% or better on all the five steps in which this range is provided. On the lowest range, increments as low as .00001 mmfd may be used. Measurement is made at radio frequency in a crystal-controlled circuit having both primary and secondary voltages automatically regulated for maximum operating stability.

The connector base accepts adapters for tubes up to 8-pin and provides for connection of coaxial cables to any pair of elements whose capacity is to be measured. A universal shield is furnished with the instrument to accommodate tubes up to 2⅝" in diameter and 4¾" high; standard RMA shields, as specified for various tube types, may also be used. Special shields for various tube types can be supplied when required.



* * *

FAIRCHILD AMPLIFIERS AND RECORDERS

A portable recorder and a portable amplifier-equalizer are now being produced by the Fairchild Camera and Instrument Corporation, 88-06 Van Wyck Boulevard, Jamaica 1, N. Y. The unit now in production has a new magnetic cutter head replacing the old crystal cutter head; a low-pressure pickup with a choice of crystal or dynamic type; and the weight of the machine has been lightened to 70 pounds.

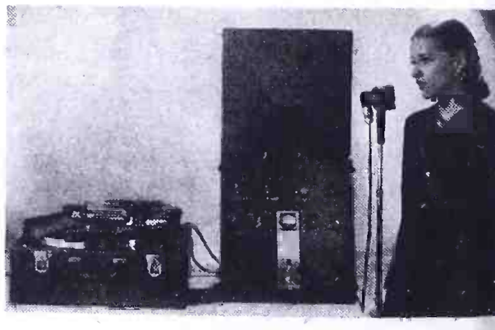
According to George Rattray, assistant vice president in charge of engineering, and Theodore Lindenberg, sound equipment project engineer, the general vibration level of the recorder has been reduced by 5 to 8 decibels.

The recorder is designed for portable field service, for lateral recording and use of 78 rpm or 33.3 rpm discs up to 16".

The recorder mechanism has a selection of four cutting pitches; either from the inside of the record to the outside or from the outside in, without the necessity of changing feed screws. The turntable speed is changed from 78 rpm to 33.3 rpm by shifting a push rod on the center pivot of the turntable. For operation on a 115 volt/60 cycle power line. Power consumption, 70 watts.

On the recording head carriage mechanism, an octagonal scale reading in minutes is mounted directly back of the carriage mechanism. This scale is calibrated to meet all sixteen conditions of pitch, direction and speed by simply rotating the scale until the calibration that matches the conditions of the recording appears on top. In addition, this scale shows the minutes of recording completed

(*Continued on page* 80)



THE INDUSTRY OFFERS . . . —

(*Continued from page 78*)

and the minutes of blank disc remaining to be recorded.

Pitches of 98, 141, and 161 lines per inch are selected by snapping into place any one of four standard spur gears supplied.

The 33.3 rpm speed is obtained by a 54 to 1 gear and worm reduction of the 1800 rpm synchronous motor speed.

The 78 rpm speed is secured from a friction step-up device from 33.3 to 78 rpm. This consists of a group of balls riding between two conical shaped races held together under spring pressure. The speed shift pin, when in the up position, interposes this step-up mechanism between the 33.3 drive and the turntable, driving the turntable at 78 rpm.

The portable recorder is equipped with 110-125 volt, 60 cycle synchronous motor which is dynamically balanced and selected for quiet operation.

The equalization element of the amplifier-equalizer is provided to permit a boost of the extremes to compensate for falling off of high or low frequencies. At any frequency from 20 to 100 cycles, a boost of from 0 to 20 db is said to be available; and at any frequency from 4,000 to 10,000 cycles, the manufacturer states that 0 to 20 db may be added. There is a negligible interaction between the controls and they can be pre-set at any time to provide known amounts of equalization for any particular purpose.

This unit is provided with receptacles for instantaneous hookup on the rear of the chassis for the accommodation of amphenol input plugs for a microphone of 50 or 250 ohms impedance, two crystal pickups of 10,000 ohms impedance and both a **T** unbalanced and an **H** balanced 500-ohm impedance for radio tuner or line from an external mixer. Outputs are connections provided for two portable recorders as well as for a **T** unbalanced and an **H** balanced 500-ohm output line.

The amplifier, speaker and control panel are mounted in a metal trunk with an overall width of 17″ and height of 18¼″. The greatest depth is 11″.

There are 11 tubes in the circuit: 2—1620's, 1—6J7, 4—6J5's, 2—6L6G's, 1-6H6, and 1—5U4G.

Maximum undistorted output is said to be 12.51 watts from the 500-ohm winding of the **T** or **H** position on the output. The noise level below zero level of .006 watts at full gain of amplifier with bass and high controls off is said to be minus 22 decibels, and the hum level is said to be minus 40 decibels. The microphone input positions are said to have a gain of 107 decibels. Both pickup positions are said to have a gain of 61 decibels, and both line input positions a gain of 67 db.

If the dynamic pickups are used, the pickup inputs will be wired for 50-ohm impedance at 170 db gain.

The master gain control is provided with a logarithmic taper.

A phone monitor jack is provided on the panel so signals may be monitored with headphones instead of the loud speaker. This jack is a separate 8-ohm output winding.

* * *

BERYLLIUM COPPER SPRING GRADE WIRE

A spring grade beryllium copper wire with a new type of silver coating is now being produced by Little Falls Alloys, Inc., 189 Caldwell Ave., Paterson, New Jersey. Wire is available in sizes .064″ to .007″. This wire, known as Silvercote, is said to have a guaranteed minimum tensile strength of 185,000 pounds after hardening.

* * *

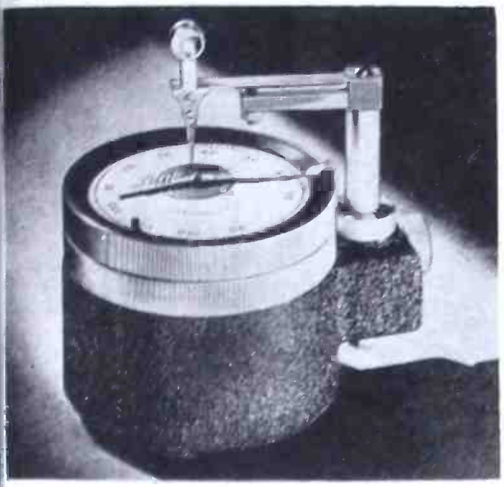
SHEFFIELD TORSIOMETER

An instrument that measures the torsion of small spiral springs has been produced by the Sheffield Corporation, Dayton, Ohio, as a result of collaboration with the Gruen Watch Company and the Manross Division of the Associated Spring Corporation. A direct reading of the torsion measurement of the spring being checked can be made in millimeter grams on a graduated Manross scale. This permits the identifying and classifying of springs as to torsion before assembling in instruments.

The torsiometer will accommodate springs up to 2⅞″ in diameter with a maximum torsion measurement of 49.5 millimeter grams.

In operation, the inner end of the spiral spring to be checked is attached to the center post. The outer end is grasped by the suspended tweezers. The outer marker is set at 120 on the scale which represents the zero point. The dial is then turned one complete revolution. The point at which the indicator rests on the scale represents the torsion classification of the spring.

The indicator spindle rides on a jewel bearing and has a resistance of nearly zero. The pointer is made of non-magnetic material and the pointer assembly is statically balanced.



* * *

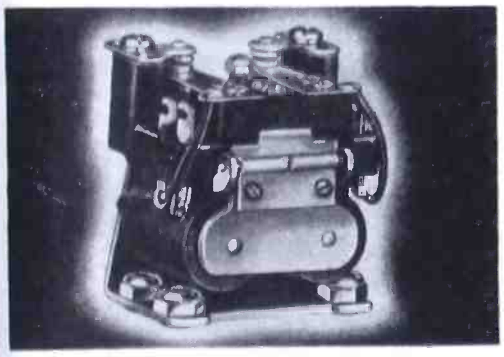
ALLIED CONTROL DUAL-COIL RELAY

A dual-coil power relay with a semibalanced armature has been developed by the Allied Control Company, Inc., 2 East End Avenue, New York City 21, N. Y.

The double coil construction is said to make this type of unit suitable for use in plate circuit applications where the limited amount of current is insufficient to operate a single coil power relay directly.

The two coils of the relay, known as BOY, may be connected in series for operation at one voltage and in parallel for a second voltage; or one coil can be used for operation while the other is used for holding. The coils are cellulose acetate sealed.

Dimensions are 1⅝" x 1⅞" x 1½"; weight, five ounces.



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REINER 5" OSCILLOSCOPE

A 5" oscilloscope, model 550-A, for square wave analysis has been produced by Reiner Electronics Company, 152 West 25th St., New York 1, N. Y. The instrument is said to afford reproduction of square-wave signals from 40 cycles per second to 50 kc per second. A vertical amplifier has a sine wave frequency response of 500 kc per second, ± 1 db. The voltage gain is said to be 625.

The oscilloscope is equipped with a detachable coaxial cable with input capacity of 8 mmfd. A compensated 4-step attenuator permits observation of voltages up to 175 volts, with a variable gain control on the second stage providing continuous variation of gain.

* * *

CML VOLTAGE REGULATED SUPPLIES

Two power supplies, model 1100 table model and model 1110 rack mounting, using the series regulator circuit, have been announced by the Communication Measurements Laboratory, 116 Greenwich Street, N. Y. City, N. Y.

Two stage control circuit is used. The high voltage output can be shifted through a range of 225 to 325 volts by means of a potentiometer control on the front panel. The maximum current drain is 200 milliamperes from 225 to 300 volts and 180 milliamperes from 300 to 325 volts. Under these conditions the sum of all a-c components present in the output is said to be less than 5 millivolts. The change in voltage output from no load to full load is said to be less than one volt.

The primary of the power transformer is tapped for use at 105, 115 and 125 volts on a 50-60 cycle source. An unregulated heater supply winding of 6.3 volts at 5 amperes is furnished.

* * *

ALLIANCE AIRCRAFT D-C MOTOR

An aircraft type d-c series motor is now being produced by the Alliance Mfg. Company, Lake Park Blvd., Alliance, Ohio.

Primarily designed to operate blowers for cooling purposes in aircraft equipment, the unit operates on a 28 volt d-c source at 0.75 ampere delivering a 1/80 h-p at 8000 rpm. Totally enclosed with ball bearing construction. Measures overall, less the ¼" diameter shaft extension, 3" in length by 1⅞" diameter. Weighs 17 ounces. Said to have low temperature rise permitting operation under high ambient temperatures.

Basic design can readily be modified to meet other volume applications with either shunt or series winding for desired voltage, current drain and horsepower output up to 1/50 consistent with speed and duty cycle.

* * *

BRADLEY LAB C-O RECTIFIERS

A group of copper oxide rectifiers, coprox, has

(Continued on page 82)

THE INDUSTRY OFFERS . . . —

(*Continued from page* 81)

been announced by Bradley Laboratories, Inc., 82 Meadow Street, New Haven 10, Conn. Features of the rectifiers are said to be gold contacts on the copper oxide pellets, and pre-soldered lead wires.

Among the types available is the BX-100, a center tap full-wave rectifier completely enclosed in bakelite. It has been designed for use in special circuits up to 8 megacycles. Another type produced, BX-22.5, is a single half-wave rectifier. A full wave, BX-22.2, and a double half-wave type, BX-22.4, are also produced.

BX-22.3

BX-100

BX-22.4

* * *

KELLOGG FIELD PHONE GENERATOR

A portable generator, GN-38-B, designed by the Kellogg Switchboard and Supply Company, Chicago, is now being used in the EE-8 portable field telephone of the Signal Corps.

* * *

TAB-TYPE WIRE MARKERS

A tab or apron type of electrical conductor identification marker for very fine or small diameter conductors, such as sizes smaller than No. 10, is now available from William Brand & Co., 276 Fourth Avenue, New York, N. Y.

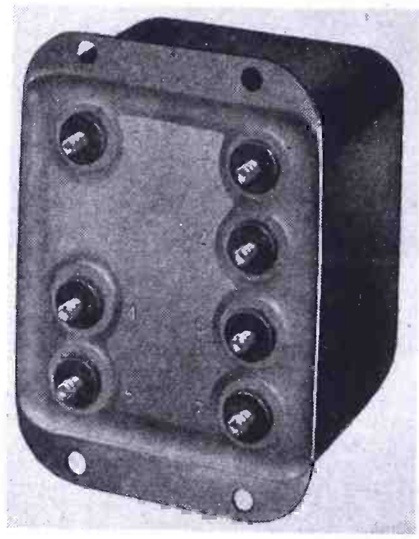
* * *

PEERLESS TRANSFORMER

A hermetically sealed transformer, type S, has been developed by Peerless Electrical Products Co., 6920 McKinley Ave., Los Angeles 1, Calif.

The new transformer is enclosed in a cold drawn copper cadmium plated steel case.

Terminal is molded into a plastic block which has a metal flange molded into its periphery. This flange is then soldered into the case.



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G. E. THYRATRON WELDING CONTROL

A thyratron welding control for low-capacity spot welders has been announced by the industrial control division of G. E. Coupled with a suitable welding transformer, this control can be used with either welding tongs or a small bench welder. It is said to be particularly suitable for the spot-welding of vacuum tube parts.

Other applications for which the control, in combination with the proper welding transformer, is desirable, include the welding of solid or stranded wires to terminals of copper, brass, bronze, steel, or ferrous alloys; joining two tinned-copper, steel, or alloy wires; and spot-welding thin pieces of various alloys.

* * *

RADIOTONE PLANT SOUND UNIT

A portable sound system with recording,

receiving, phonograph and p-a features has been produced by the Radiotone division of The Robinson Houchin Optical Company, 79 Thurman Avenue, Columbus, Ohio.

• • •

INDUSTRIAL SPECIALTY ELECTRON MICROSCOPE CAPACITOR

A .01-mfd, 40,000 volt, d-c capacitor has been built for use in a special application of the electron microscope by the Industrial Specialty Company, 1725 West North Avenue, Chicago 22, Illinois. This capacitor is said to be capable of continuous operation at 80°C.

The case is welded steel 4 11/16" x 5¾" x 7", with a stand-off insulator 8½" high. This insulator is the Westinghouse solder seal type, soldered directly to the case.

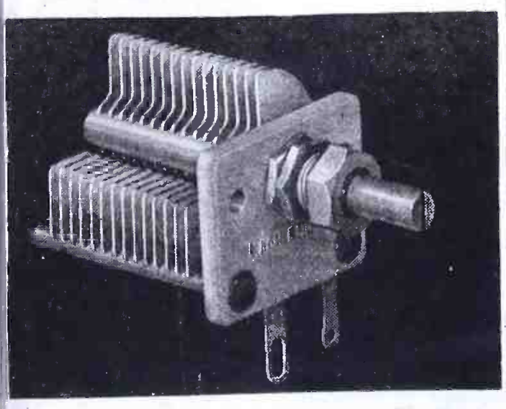
* * *

KAAR VARIABLE CONDENSERS

A new line of variable condensers with capacities of from 12 to 140 mmfd are now being offered by Kaar Engineering Company, Palo Alto, California.

Condensers feature shafts slotted for screwdriver adjustment. Tapered lock nuts and split bushings are also used.

Special types with double rotors and stators, high maximum capacities, or special mounting brackets are also available.



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AERCO MULTI-GRIP CHUCK

A hydraulic multi-grip milling machine chuck has been announced by Aerco Corporation, Hollydale, California. The chuck has two rows of five collets. Both rows of parts may be milled simultaneously, according to the manu-

(Continued on page 84)

THE INDUSTRY OFFERS ...

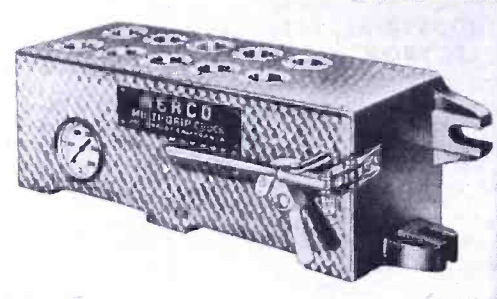
(*Continued from page* 83)

facturer. A hydraulic hand pump supplies the necessary pressure for closing all collets simultaneously.

Self-centering aligning collets are said to insure perfect on-center milling. An automatic safety stop prevents damage to collets should all stations not be filled. The collets themselves operate on a double taper, one at each end. An adjustable depth stop is provided for each collet.

The new chuck comes in two models. The pump, valves, pressure gauge, depth gauge and collet assemblies are all incorporated in the main body of the chuck. Model 10-B measures 14¼" x 4½" x 4½" with a collet capacity of from ⅛" to ½". Model 10-E measures 21" x 6¾" x 6¾". Its collet capacity is from ½" to 1¼".

Round, square and hexagon collets are available for both chuck models.



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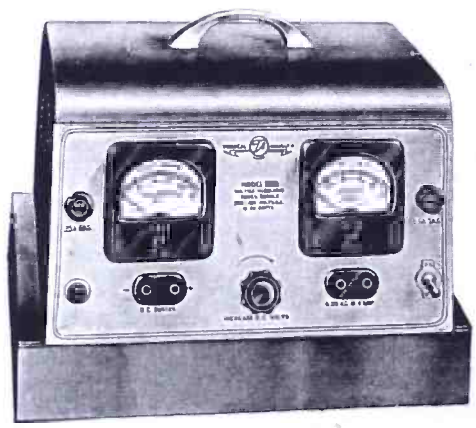
VOLTAGE REGULATED POWER SUPPLY

A voltage regulated power unit, model 1218, supplying d-c loads up to 40 watts at 200 to 400 volts, with voltage variation of less than 1% from zero to full load, has been developed by the Technical Apparatus Company, 1171 Tremont Street, Boston 20, Massachusetts.

Current may be drawn up to 100 milliamperes at 400 volts and increasing to 200 milliamperes at 200 volts.

A built-in voltmeter and milliammeter permit direct reading of output delivered at a safety jack located on the front panel. A second output jack affording 4 amperes a-c at 6.3 volts (unregulated) is also provided. Both the d-c output and the a-c input are fused for protection against overload.

The power supply is housed in a ventilated steel cabinet 13" wide, 9" high, and 9" deep.



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BARBER TUBE VOLTMETER

A wide range vacuum tube voltmeter with a h-f probe is being made by Alfred W. Barber Laboratories, 34-02 Francis Lewis Blvd., Flushing, N. Y. This probe is cone shaped with the high terminal in its nose. The probe is molded from low-loss material.

The voltmeter, model VM-27E, is said to measure voltages from 0.1 to 100 volts at d-c, a-c and r-f frequencies to over 100 megacycles.

•

PAPER PACKS A WAR PUNCH
FIGHT WASTE!

•

HIGHLIGHTS OF DR. ROTERS' AIEE PAPER ON D-C ELECTROMAGNET DESIGN CONSIDERATIONS

FOR an intelligent application of magnetic circuits to instruments, motors and control apparatus, a working knowledge of the fundamental properties of electromagnets is essential. Graphical solutions to magnetic design problems as a substitute for cut-and-try programs, or just plain guesswork were offered by Dr. Herbert C. Roters, Director of Research, Fairchild Aviation Corporation, in his paper.

The magnetization curve was presented to provide an analysis (useful flux in the operating air gap versus NI, the ampere-turns) for all the iron parts of the magnet and the air gap which is used to determine the optimum operating point.

In tractive magnets, the useful work done equals the *force* at the start of the stroke times the *stroke*. In designing a magnet it was shown that it is first necessary to calculate its index number which is $= \sqrt{F}/S$ where $F =$ force and $S =$ stroke. It is then necessary to pick the general type of solenoid which operates most efficiently for that index number. A long, thin magnet called the *solenoid* and *plunger* type (or *leakage flux* type) has the lowest index number; hence is most suitable for long stroke and small force. A tapered plunger design comes next. A cone-faced plunger has an intermediate value. A flat-faced plunger magnet has a high index number, being suitable for developing high forces at short strokes, while a flat-faced armature type of lifting magnet has the highest force and smallest stroke. The common horseshoe magnet was shown to be impractical for any single job but, when large pole pieces are attached, it can provide satisfactory results for a wide range of operating points.

The following expression was presented in the discussion of force:

$$\text{Average force} = \frac{B^2 S}{2\mu} = \frac{B^2 r^2}{11.45} \text{ lbs. for}$$

lifting magnet, and $\dfrac{B^2 r^2}{22.9}$ for the flat-faced plunger magnet, where

r = radius of core in inches
B = kilomaxwells per square inch

Particular attention was given to relays which are usually limited in power consumption. For high speed action, according to Dr. Roters, inductance must be held to a minimum, making a high permeance and current necessary. It is also desirable to laminate or slot the core to decrease the lag due to eddy currents. The advantages of Permalloy were stressed, including the low residual which permitted the omission of the usual non-magnetic spacer to prevent sticking. This in turn, permitted a shorter stroke, less leakage, less NI, a smaller coil . . . etc., offering a cumulative advantage.

Coil Design Problems

Dr. Roters also brought out many interesting facts about coil design. Where power can be neglected (as in many aircraft applications) a considerable reduction in size is possible when the coil is wound with Formex or Formvar vinyl-acetal-insulated magnet wire. These coils seem to stand 150° C indefinitely and, in some cases, even 160°. This allows a temperature rise of at least 80° C over an ambient of 70° C, a favorite present high limit. Because this type of insulation does not become brittle as does standard enamelled wire, the paper used between layers may often be eliminated, further reducing the coil size.

It is obvious that a smaller coil permits a smaller magnetic circuit. But it does more, according to Dr. Roters. When the length is decreased, the NI and leakage coefficient are decreased. This permits a smaller core, which further reduces the coil size, etc., again offering a cumulative effect. As the unit decreases in bulk, the inductance also decreases, making faster operation possible. Similarly, increasing the flux density by the use of special alloys (where price will permit) results in a series of cumulative benefits which reduces the magnet size considerably. The Ferrocobalts are particularly good. Dr. Roters gave as an example a solenoid of about 1 cubic inch wound with Formex which withstood 250 watts for intermittent periods of a second or so in regular operation.

In the discussion which followed the paper, coil construction, special alloys, comparison of core materials and the advantages of annealing the core in low NI applications were all given consideration. The speaker also referred to his book, *Electromagnetic Devices*, published by John Wiley, which, incidentally is an outstanding work.—(*E. G.*)

IMPEDANCE TRANSFORMATION

(*Continued from page* 44)

appeared from time to time, but are not generally available.

Knowing the *admittance* $Y = G + jB$ of the receiver, the two numbers

$$g = G R_o \qquad b = B R_o$$

may be obtained from R_o, the characteristic impedance of the line.

The two corresponding circles of the *z-plane* map will be found to intersect at some point. This intersection should be marked on the map. Then the *rotating dial* (Figure 12) which we imagine to have been enlarged and placed in position, should be set to read *zero* on the dial, (the dial is read at the point $g = 0$, $b = 0$ of the chart).

In the next operation we make a point mark on the dial sheet opposite the intersection previously determined; after which the dial is rotated, until the dial reading is equal to L/λ, the length of the line in wavelengths. The point mark which we have made on the dial sheet will have moved on an arc of circle, and its new position will determine a new intersection, hence two new values of *b* and *g*, from which the input admittance of the line may easily be determined (dividing by R_o).

The chart of Figure 12 may be redrawn to any desired accuracy by using the construction of Figure 12a.

AUDIO REACTORS

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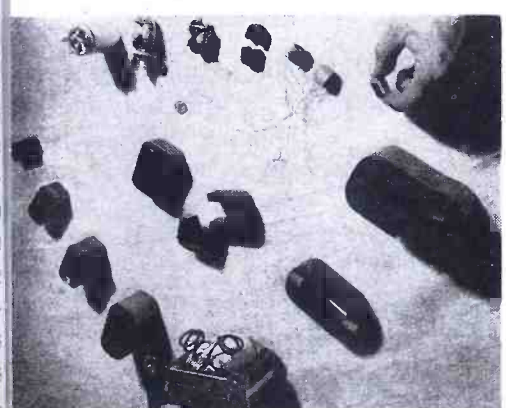

(*Courtesy Westinghouse*)

Hipersil ribbon-wound cores.

the change in inductance and also the change in Q.

If the inductance is properly designed, the inductance (in henrys) will not change perceptibly even with this large range in voltage, but of course, the Q will change.

From this it follows that if the Q or inductance of a coil is to be expressed, the applied voltage and frequency must also be stated.

This non-linear response will produce waveform distortion and inter-

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AUDIO REACTORS

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modulation, especially if the operating range is through a high value of flux density. If the device is operated much above its maximum permeability, the distortion is horrible, showing itself chiefly as third harmonic. Swing chokes and voltage regulating transformers operate on the negative slope of this permeability curve. To make an inductance stable and suitable for equalizers or filters, all of the characteristics mentioned above must be suppressed. This is done chiefly by introducing reluctance into the magnetic circuit. This reduces the effective permeability of the magnetic circuit, but more important, makes the apparent permeability constant for different values of applied voltage. The loss in permeability is generally a very large factor, in most cases will probably be greater than 10.

Now in terms of transformer design this would be a very great loss, but in terms of inductances, the permeability, even though reduced by this large amount, is still very much greater than with no core. A design might call for a permeability of 50, even though a permeability of 1,000 was possible with the same laminations interleaved. The loss of inductance would be 20 to 1. However, the inductance in henries would be 50 times as great as if no core was used. Obviously, if you have a permeability of 50, the amount of winding for a given inductance can be materially reduced, resulting in smaller size or increased Q or both.

The effect of the introduction of an air gap (called magnetic resistance or reluctance) in reducing the non-linear distortion is comparable to the introduction of resistance into the plate circuit of a vacuum tube amplifier. The tube is basically not free from distortion since the plate current varies as the 3/2 power of the applied plate voltage, and can only approach freedom from distortion when resistance is introduced.

The apparent loss of permeability is expressed as

$$\mu = \frac{\mu_{AC}}{1 + [(a/e)\ \mu_{AC}]}$$

where μ = permeability of finished inductance, and μ_{AC} = permeability of core material. a/1 is the ratio of air gap to the length of the magnetic path. A graph showing this curve appears in Figure 8. The precise control of inductance, accomplished by varying the air gap, may be used to maintain production (of a typical 1,000-cycle Q_{max}

inductance) to within 1% of specifications.

The practical application of iron core inductances requires an air gap in the magnetic circuit for the purpose of adjusting the inductance as well as stabilizing. A design for Q_{max} to occur at low frequencies such as 50 cycles will probably have a very small air gap, a few thousandths of an inch at most. It is quite obvious that if this air gap changes even slightly, the inductance will change greatly, so the mechanical problem of maintaining an accurate air gap under these circumstances is acute. There is a solution which is quite unique; an F-shaped lamination was chosen which had the center leg cut at 45°. The two half sections of *F* laminations were held butted up with a clamp, as shown in Figure 9, and the inductance was regulated by sliding the *F* pieces, which open up the air gap, at the 45° cut of the lamination. The particular virtue of this design lies in the location of the air gap which falls in the center of the winding, and also in that there is a minimum of leakage flux to intercept the clamp, there being an intimate contact where the two F-sections butt.

At medium audio frequencies, air gaps, by necessity, assume a/l ratios of .001 to .005, so the matter of precise air gap spacing is very much less critical. A design might be illustrated by an *E-I* core structure using a bakelite spacer of .050 with the *E* and *I* pieces held by a hard fiber frame. The screws are brass in order to reduce the hysteresis loss caused by steel screws.

Typical power filter reactors are examples of how not to build high-*Q* inductances. To begin with, the core material is usually several grades poorer than *audio* grades, and is thick. This makes eddy current loss and hysteresis loss high. The shape will invariably be *E* and *I* which will result in leakage flux. In order to clamp the *I* pieces together, keepers are used along with some kind of a shell or bracket made of cold-rolled steel.

The presence of a keeper or iron clamp at a point of leakage flux will distort the magnetic field due to the permeability of the keeper being many times that of air, (Figure 9). The lines of force which are supposed to travel from the *E* to *I* crowd into the keeper, which produces a very high flux density in the keeper, resulting in the generation of spurious frequencies which may be difficult to filter. To demonstrate this point, one has only

(*Continued on page* 90)

(*Continued from page* 89)

to tune the filter choke to parallel resonance at 120 cycles and then connect a high gain amplifier to the power supply. The noise generated by the choke will be quite noticeable. However, if the choke is constructed along the idea of a high-Q audio reactor, the tuned filter system will be effective in reducing hum.

Reactors which must exhibit high Q at high frequencies require magnetic structures which have very low eddy current losses. This in turn calls for thinner and thinner laminations, and, at the same time, a large air gap. The search for thinner and thinner laminations ends with powdered core. The air gap in this case is the binder and insulating material, and of course is uniformly distributed throughout the length of the magnetic circuit. The usual design employs the classic ring core; however, the powdered core is not as ideal as might be expected since the permeability is so low that the core is not altogether effective in confining the flux. Powdered cores for audio frequencies have much more active material; they are more dense than the familiar radio frequency cores. Permeabilities range around 25 and 50; the permeability for radio frequency cores is usually from 2 to 5.

Most inductances for equalizers and wave filters, etc., use conventional transformer cores and transformer type windings. As the requirements for Q become more severe, special high efficiency shapes are used along with certain peculiar devices.

A recent transformer core development of the Westinghouse Company, hipersil, has been used effectively in high-Q inductance, (Figure 10). The material is made in ribbon form and so heat treated and processed that it has very low losses. This is accomplished chiefly by the pronounced development of grain structure. It is layer wound on a mandrel, annealed at high temperature, baked with a binding material, then cut into two halves, U-shaped. The layers are so well bound together that they appear to be solid. The cut surfaces are

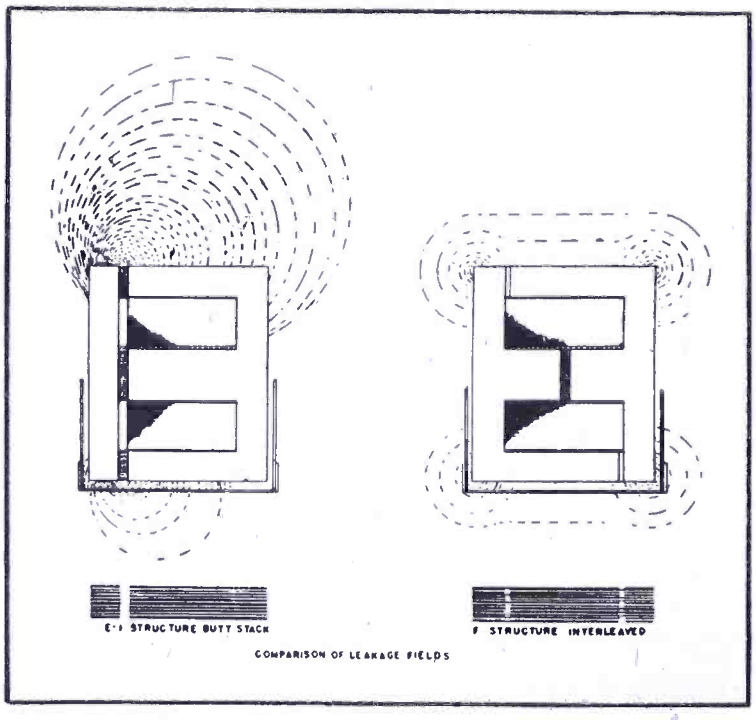

Figure 9

Here we see how the presence of a keeper or iron clamp at a point of leakage flux will distort the magnetic field due to the permeability of the keeper being many times that of air.

ground flat so that when the core is eassembled into a transformer the air ap at the butt will be a minimum. There are three outstanding adantages in this type of core. First, it as unusually low losses; second, it is eadily adaptable to an astatically balnced winding; and third, it may be ecurely clamped to maintain the air ap without loss in Q.

Summation

(1) All iron core inductances hould have an air gap for stability nd distortion-free operation unless hey operate at extremely low flux ensity.

(2) The formulas for $f(Q_{max})$ etc., old only for low values of flux ensity, where the hysteresis loss is mall compared with the eddy current oss. Furthermore, if the operating ux density is high, the $f(Q_{max})$ will e somewhat lower than the formulas ndicate, since the increased hysteresis oss is effectively a shunt resistance cross the inductance in parallel with he eddy current loss.

(3) In order that Q be a maximum: *a) The core area should be large; b) the copper space should be large nd the length of the turn short; (c) he length of the magnetic circuit hould be short; (d) the electrical resistance of the core material should be high; (e) the lamination of the core material should be thin; (f) the flux density should be low; and (g) there should be a minimum of leakage flux.*

(4) For high Q at low frequency, all of the above, but most important, the effective permeability of the magnetic circuit should be large, and the air gap should be small.

Conclusion

Iron core inductances can account for the following improvements over air core inductances at audio frequencies.

(1) Smaller physical space.

(2) Higher Q.

(3) A means of adjusting or trimming the inductance to exactly the right value.

(4) A reduction in the external field.

(5) An inductance which can be conveniently shielded both electrostatically and electromagnetically.

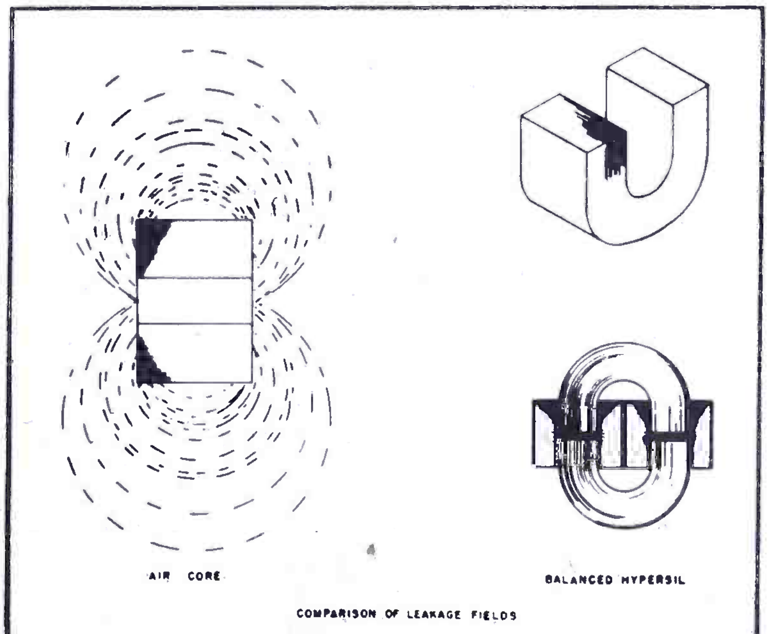

Figure 10
Characteristics of hipersil, with a comparison of the leakage fields.

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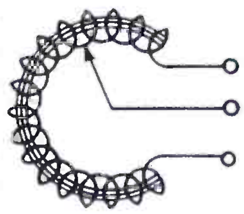

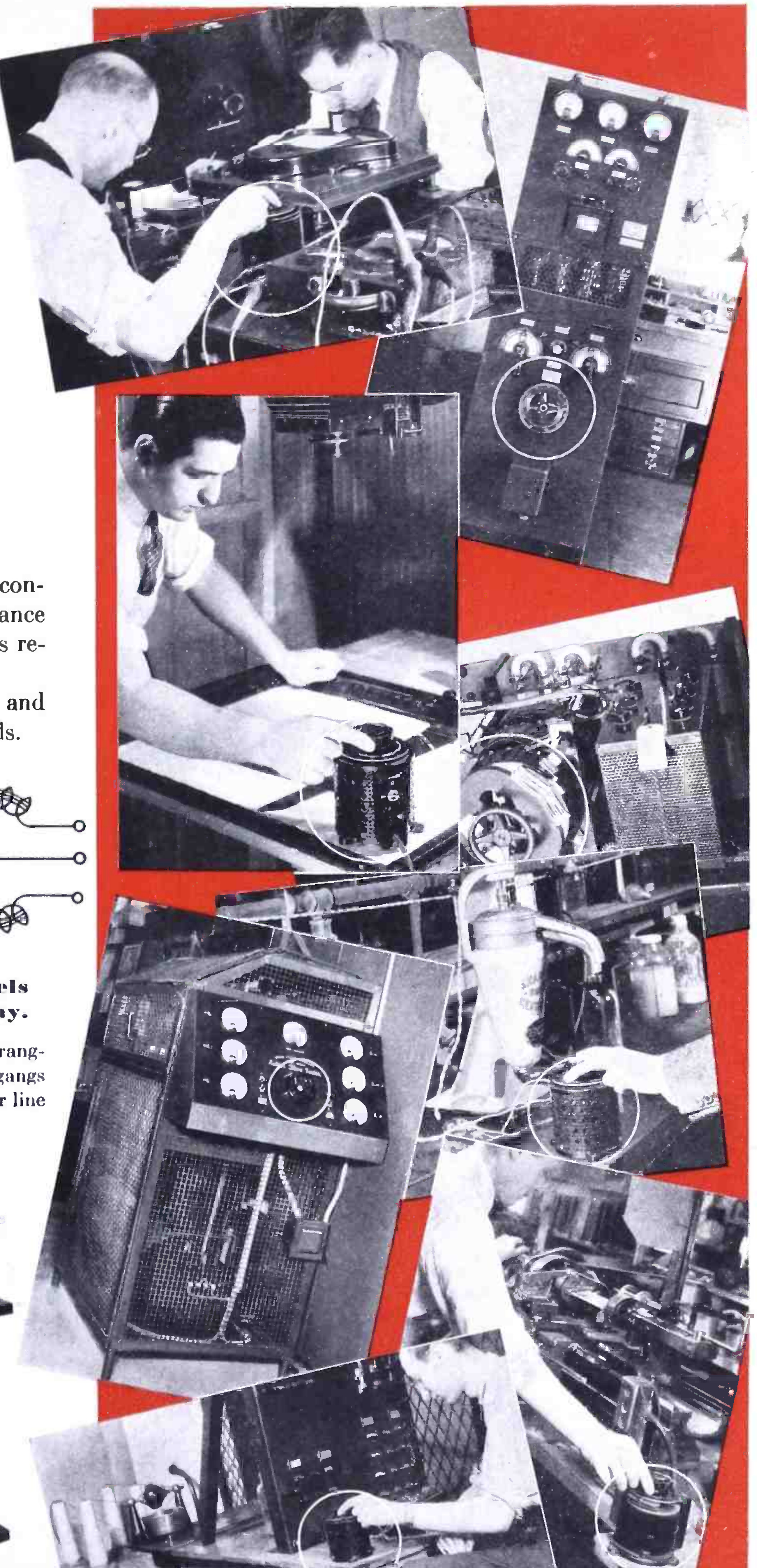

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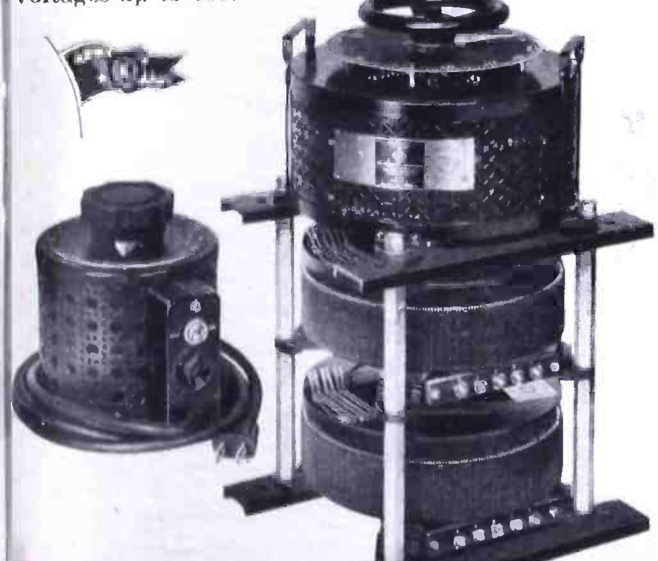

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