the GENERAL RADIO Experimenter

INDEX

TO

GENERAL RADIO EXPERIMENTER

VOLUMES XX AND XXI

June, 1945 to May, 1947

GENERAL RADIO COMPANY

CAMBRIDGE MASSACHUSETTS

U. S. A.

I N D E X
TO GENERAL RADIO EXPERIMENTER
Volumes XX and XXI, June 1945 through May 1947

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THE General Radio EXPERIMENTER

VOLUME XX No. 8 JANUARY, 1946

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

THE SERIES AND PARALLEL COMPONENTS OF IMPEDANCE

Also
IN THIS ISSUE

● WE HAVE RECEIVED lately a number of inquiries about the meanings of such terms as "series capacitance," "parallel capacitance," "series resistance," "parallel resistance," etc., as they are used in instruction books for General Radio bridges. Although most engineers think in terms of the series components of impedance, many types of problems, particularly those involving vacuum tubes, are more simply handled in terms of the parallel components. Certain bridge circuits give directly the series components of an impedance, while others can be arranged to give the parallel components, the choice depending on the intended application. Discussion of the relationship between the series and parallel components, however, seldom appears in the elementary textbooks.

That any impedance can be represented both ways is clear from the fact that measurements on it at a single frequency can determine only the relationship between the voltage across the impedance and the in-phase and quadrature components of the current flowing through it. Stated in terms of power engineering, a circuit element draws a certain amount of power at a particular value of power factor, and these two quantities completely define the effective impedance of the element for the conditions applying. It is sometimes convenient to represent the impedance as a pure resistance in series with a pure reactance, but it is very often more convenient to consider it as made up of a different value of resistance in parallel with a reactance.

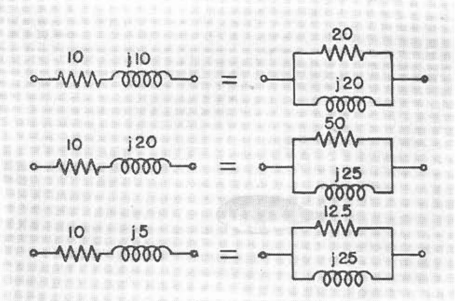


Figure 1. Examples of equivalent series and parallel circuits.

The two representations, however, are completely equivalent and either pair of components can be simply determined in terms of the other pair.

For example it will be seen that, in three cases shown in Figure 1, the first configuration of series elements would draw the same current, both in phase and magnitude as the second configuration, consisting of resistive and reactive elements in parallel. The two arrangements of each case are indistinguishable from each other by measurements made at their terminals at a fixed frequency.

The general relationship between the elements of the series and parallel arrangements can be simply found by equating the current drawn in the two cases.

$$i = \frac{e}{R_s + jX_s} = \frac{e}{R_p} + \frac{e}{jX_p} \qquad (1)$$

where R_s and X_s are the series components and R_p and X_p are the parallel components. Rationalizing and equating the real and imaginary terms,

$$\frac{R_s}{R_s^2 + X_s^2} = \frac{1}{R_p}$$

or

$$R_p = R_s\left(1 + \frac{X_s^2}{R_s^2}\right) \qquad (2)$$

$$\frac{X_s}{R_s^2 + X_s^2} = \frac{1}{X_p}$$

or

$$X_p = X_s\left(1 + \frac{R_s^2}{X_s^2}\right) \qquad (3)$$

The quantity X_s/R_s is the familiar Q or storage factor of an inductor or capacitor, and its reciprocal is the dissipation factor D, more frequency employed in describing the losses in capacitors. Substituting these quantities in Equations (2) and (3),

$$R_p = R_s\,(1 + Q^2) = R_s\left(1 + \frac{1}{D^2}\right) \qquad (4)$$

$$X_p = X_s\left(1 + \frac{1}{Q^2}\right) = X_s(1 + D^2) \qquad (5)$$

These equations give the parallel components of impedance directly in terms of the series components. The relationships, however, serve equally well when the series components are required and the parallel components are given, because the quantity Q or D can be determined directly from either the series or parallel components. Dividing (4) by (5),

$$\frac{R_p}{X_p} = \frac{R_s}{X_s}Q^2 = \frac{R_s}{X_s}\frac{1}{D^2}$$

or

$$Q = \frac{1}{D} = \frac{X_s}{R_s} = \frac{R_p}{X_p} \qquad (6)$$

so that Q can be determined immediately, whichever components are given, and used in Equations (4) and (5) to obtain the other components. A further simplification is that only one of the two Equations (4) and (5) need be employed with (6) to make the complete transformation. The three steps in each case are as follows:

Given R_s and X_s

(1) $Q = \dfrac{X_s}{R_s}$

(2) $R_p = R_s\,(1 + Q^2)$

(3) $X_p = \dfrac{R_p}{Q}$

Given R_p and X_p

(1) $Q = \dfrac{R_p}{X_p}$

(2) $R_s = \dfrac{R_p}{1 + Q^2}$

(3) $X_s = R_sQ$

If it is preferred to work in terms of dissipation factor the corresponding steps are:

Given R_s and X_s

$$(1)\quad D = \frac{R_s}{X_s}$$

$$(2)\quad R_p = R_s\left(1 + \frac{1}{D^2}\right)$$

$$(3)\quad X_p = R_p D$$

Given R_p and X_p

$$(1)\quad D = \frac{X_p}{R_p}$$

$$(2)\quad R_s = \frac{R_p}{1 + \frac{1}{D^2}}$$

$$(3)\quad X_s = \frac{R_s}{D}$$

It is seen that use of Q or D, which are associated with equal simplicity with either the series or parallel components, greatly facilitates the transformation. Since (6) is readily borne in mind, the only relation that need be remembered is that, as seen from (4), the ratio between the parallel and series resistances is the quantity $1 + Q^2$. It should be noted that the parallel resistance and parallel reactance are always greater than the corresponding series components. It is obvious that for large Q the series resistance must be small compared with the series reactance, but the parallel resistance must be large compared with the parallel reactance.

One of the simplest examples of the utility of the parallel impedance components is in parallel resonant circuits where the coil losses are high. It will be seen in Figure 2 that parallel resonance

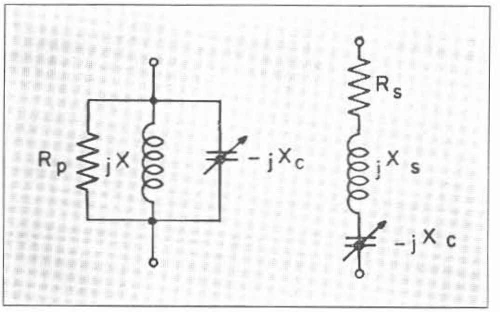


FIGURE 2. Series and parallel resonant circuits. The capacitance necessary to resonate with a given inductance will depend upon whether the elements are connected in series or in parallel.

occurs when the condenser reactance is exactly equal to the parallel reactance of the inductor, regardless of the coil losses. If the tuning capacitance for parallel resonance is determined from the series components of the coil impedance, on the other hand, the required value depends both on the resistance and on the reactance. In the series circuit the opposite applies and resonance occurs when the condenser reactance is exactly equal to the series reactance of the inductor. Where the Q of the coil is high, the difference between its series and parallel reactance is negligible in ordinary applications. Even with a Q of 10 the difference is only one per cent. But for lower values of Q the difference rapidly increases. The parallel reactance of an inductor with a Q of 1 is twice the series reactance, so that only half the capacitance is required to tune it to resonance in a parallel circuit as in a series circuit.

— W. N. TUTTLE

HEAT DISSIPATION FROM CABINETS FOR ELECTRICAL INSTRUMENTS

● **IN SELECTING THE TYPE** of cabinet construction to be used in a particular instrument design, performance considerations are generally thought to dictate such matters as shielding requirements and anti-vibration treatment, whereas the importance of aesthetic appeal, portability, compactness and cost is evaluated in terms of customer demand. The fact that practically all instruments, at least those using vacuum tubes, dissipate a certain amount of heat is not only a design problem that is often neglected in the first draft of an instrument, but may be the limitation on both the quality of shielding, if ventilation is necessary, and the minimum practical size to which a unit can be built. Taking account of the heating immediately raises the problems of what is an allowable maximum operating temperature, and how to predict the dissipation capacity of a cabinet of given construction and size.

Although the highest ambient temperature expected in the field of application generally determines the criterion for temperature rise, the hot spot temperature is the limiting factor where the deterioration of components is concerned, and the average temperatuie rise is restricted by precise circuit elements when temperature coefficients obtain. Taking 50° C (122° F) as the maximum ambient at which a laboratory instrument is likely to be used, we find that for most circuits a 40° C rise for the hot spot and a 20° C rise for the maximum air temperature give conservative operating conditions.

Not only are the thermal conditions existing in even an idealized cabinet so complicated as to defy analysis, but they are widely different from those met in an actual instrument, so that the exact physical analysis, if it were available, would be impractical to apply to a specific design. A general understanding of the physical phenomena involved is, however, helpful in suggesting methods of improving the heat dissipation of a cabinet.

If we examine the three possible methods for heat to escape from an unventilated box containing the usual complement of vacuum tubes and transformers, it becomes evident that conduction accounts for most of the heat loss and that convection and radiation are small factors, because of the low temperatures of the walls. Heat coming from the air inside must have a temperature drop to push it through the cabinet wall as well as through the air films immediately inside and outside of the wall. The resistance to heat flow of a metal wall is negligible in comparison to that of these air films, although a thick wood or fabric covered wood wall has a resistance comparable to that of one of the air films. An appreciable proportion of the heat is usually generated at high temperatures, and its transfer to the inside surface of the case is therefore by radiation, so that the more closely the inside of the case resembles a black body the better will it be. In spite of the fact that the absorption of radiant energy raises the temperature of the inside surface of the case and thus reduces the temperature difference that drives the heat through the inside air film, the net effect is to increase the heat flow by raising the outside surface temperature.

Even at the low temperatures that are satisfactory for the outside surface of a cabinet, a small gain in heat dissipation results from improving the radiation efficiency of the outside surface. In short, a metal case with dull black finish inside and outside is the most efficient for this class of electrical equipment.

Experimental evidence indicates that in cabinets of 100 square inches or larger the heat dissipation capacity at a given temperature is proportional to the area. The following table shows clearly the effect mentioned above of the various finishes and materials used in conventional cabinet construction, and these data give a means of making sufficiently accurate calculations to be useful as a design tool provided wide departures from the average conditions stated are not encountered.

Panels metal and vertical.
Heat sources distributed.
Maximum air temperature rise = 20° C
Maximum hot spot temperature rise = 40° C

$$P = kA$$

P = Allowable power dissipation, watts.
k = Empirical constant, watts/square inch.
A = Area, square inches.

Type of Construction	k
Aluminum, panel and box unfinished.	0.04
Aluminum, outside of panel and inside and outside of box painted black.	0.08
Airplane Luggage, ⅜″ thick plywood, black fabric lined and covered.	0.05
Airplane Luggage, ⅜″ thick plywood, 0.005″ copper lined, black fabric covered.	0.04
Airplane Luggage, ⅜″ thick plywood, 0.005″ copper lined, lining painted black, black fabric covered.	0.05
Walnut, ½″ thick.	0.05
Walnut, ½″ thick, 0.005″ copper lined.	0.03
Walnut, ½″ thick, 0.005″ copper lined, lining painted black.	0.06
Relay Rack, outside of panel painted black, box nickel plated.	0.04
Relay Rack, outside of panel and inside of box painted black, outside of box nickel plated.	0.05
Relay Rack, outside of panel, and inside and outside of box painted black.	0.07

The design engineer is generally confronted with one of two problems, either the size of a unit is fixed, and the question is whether ventilation will be required or not, or the size is not yet determined, and a decision must be made as to whether the physical size of the circuit components or the heat dissipation capacity of the cabinet will limit the minimum size. The empirical method will invariably be used to determine the final solution to these problems, but much effort can be saved by a preliminary calculation.

— H. C. Littlejohn

AN ENGINEERING APPROACH TO TROUT FISHING

● **UNDER THE TITLE** "Busman's Holiday" we mentioned in our September, 1945, issue the measurements made by Mr. Robert F. Field of the General Radio engineering staff on water depth and temperature in Lake Winnepesaukee. Apparently there are a few engineers who like to spend their leisure hours in the highly optimistic pursuit known as angling, and some interest has been expressed by them in a more detailed statement of Mr. Field's investigation.

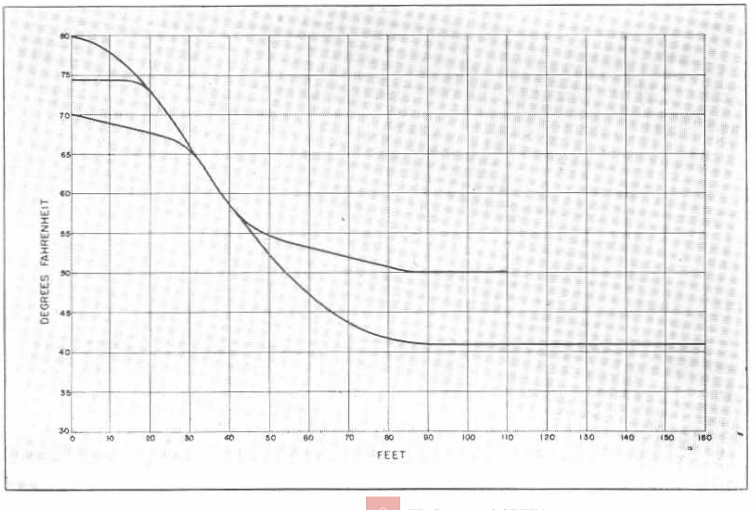
The project had two underlying motives, one economic, the other scientific. From the economic viewpoint it was desirable to find where the trout were in summer in order to increase the protein content of the family food supply; the scientific objective was to establish a relationship between water depth and temperature, and to find the deepest spot in the lake. Both objectives, we are happy to report, were attained.

That excellent fish, the lake trout, abhors warm water. Consequently, when summer comes and the water at the lake surface and in shallow areas reaches temperatures of 70 to 80 degrees Fahrenheit, he beats a hasty retreat to the deep water, where more comfortable temperatures can be found. At any lake, the old timers eagerly point out where the deep spots are located. Unfortunately, their stories seldom agree closely enough to permit the fisherman to troll through the exact spot and pull in the trout. At times, the stories follow a pattern and are obviously folklore. At others, a comparison of data from different sources produces more than a suspicion of organized conspiracy to suppress the true facts. An objective investigation, however, will invariably settle the matter.

The necessary equipment consisted of a boat, a minimum-reading thermometer and a line marked at 10-foot intervals.

In the accompanying plot are shown the results of the investigation. For the first 30 feet of depth, the temperature depends upon the time of day and upon such factors as the weather for the previous several days. The two upper curves show the difference between late afternoon and the following morning, and the lowest curve is the result of several days of cool, rainy weather. Below these surface differences, the ultimate depth does not affect the temperature-depth relationship until a 40-foot depth is reached. At greater depths, the data follow two well-defined curves. The upper curve is for ultimate depths of about 100 feet or less and a minimum temperature of about 50 degrees. When the ultimate depth is around 150 feet, the temperature below 90 feet drops to 41 degrees and stays constant thereafter. The 41-degree water is where the lake trout are found. This temperature is close to the theoretical value of 39.2 degrees where water has its maximum density.

From the plot, it is evident that, if the temperature readings fall along the upper curve, no 41-degree water (and no trout) will be found at that spot. If the readings follow the lower curve, how-



Temperature of water as a function of depth. Near the surface, the temperature depends upon the time of day and the weather for the preceding several hours, and the three branches of the curve illustrate the differences encountered. At depths below about 40 feet the curve has two branches, one for ultimate depth of about 100 feet, the other for ultimate depths of 160 feet or more.

ever, a deep area at depths up to 150 feet or more is indicated. It should be noted that 41-degree water is possible at ultimate depths less than 150 feet if the spot is spring fed. From the fisherman's viewpoint, this is quite satisfactory, since temperature, rather than depth, is his primary concern.

In summary, it can be said that the results were eminently satisfactory. In the deep spots located, lake trout were plentiful, and a 159-foot depth was determined to be the maximum in that part of the lake covered by the survey.

ACKNOWLEDGMENT

● **DURING THE WAR** the load on the American industrial machine was enormous — but it didn't break down; it delivered about what was expected of it. In our own small corner of industry we were no exception. The requirements of our Armed Services and of our Allies for our regular catalog products grew beyond anything that had ever been estimated, and kept right on growing. On top of this we were called upon to produce to government specifications large numbers of complicated and still classified measuring equipment which was as precise as the best laboratory types.

We regularly manufacture a variety of test equipment and precision components with almost two hundred different type listings in our present catalog. Most of these products are made in relatively small quantities even under wartime conditions, but they generally are important production tools or components of other electronic equipment; thus the demand was most urgent. We were uniquely in a position to build some of them, having the facilities, the machines, the skilled manpower (although badly reduced by the draft), and the requisite engineering experience.

To permit concentration of all our effort upon the making of these products we released, on our own initiative, complete drawings and all technical information in our possession for a great many of our other products to other manufacturers who took on the task of making them under their own separate contracts with the Government to help fill the urgent war needs of the moment.

In most cases the new manufacturers took hold and did an outstanding production job. In some instances they had little previous experience with this class of manufacture which makes their achievement even more impressive.

There was and is no connection between these organizations and the General Radio Company, not even to the extent of royalty payments. We delivered the manufacturing and engineering details at no charge and collected no royalties — the only considerations were that the production be for direct war purposes and that the manufacture would cease at the end of the war emergency.

In many ways these contacts were a pleasant and useful experience for us. We got to know better many members of our own and allied industries and we discovered, by attempting to explain them, some of the weaknesses in our own designs which has helped us with new designs.

Among the companies with whom we worked and our products with which they were concerned are:

Automatic Signal Corp., So. Norwalk, Conn.	TYPE 572-B Microphone Hummer
W. W. Boes Company, Dayton, Ohio	TYPE 583-A Output Power Meter TYPE 578-B Shielded Transformer
Bond-Halstead, Inc., Wilmington, Delaware	TYPE 380-P3 Switch
Eastern Company, Cambridge, Massachusetts	TYPE 487-A Megohmmeter
Hugh H. Eby, Inc., Philadelphia, Pennsylvania	Terminals
Federal Manufacturing & Engineering Company, Brooklyn, New York	TYPE 804-C Standard-Signal Generator TYPE 605-B Standard-Signal Generator
Florida Aircraft Company, Fort Lauderdale, Florida	TYPE 774 Plugs and Patch Cords
Freed Radio Corporation, New York City	TYPE 568 Condenser
General Communications Company, Boston, Mass.	TYPE 834-B Electronic Frequency Meter TYPE P-496 Electronic Frequency Meter
M and Z Industrial Development Co., Bayonne, N. J.	TYPE 434-B Audio-Frequency Meter
The Muter Company, Chicago, Illinois	TYPES 214, 314, 371, 471 and 433 Rheostat-Potentiometers
National Electrical Machine Shops, Washington, D. C.	TYPE 774-YB Terminal Units
Frank Reiber, Inc., Los Angeles, California	TYPE 804-C U-H-F Signal Generator
Stoddart Aircraft Radio Company, Los Angeles, Cal.	TYPE 714-A Amplifier TYPE 672-A Power Supply TYPE 673-A Power Supply
Ucinite Company, Newtonville, Massachusetts	Plugs and Dials

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THE General Radio EXPERIMENTER

VOLUME XX No. 9 FEBRUARY, 1946

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

Also IN THIS ISSUE

THE CONSTANT WAVEFORM FREQUENCY METER

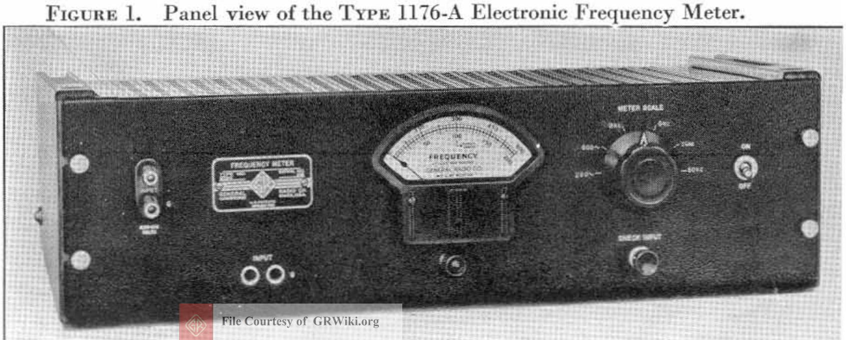
● **IN THE COMMUNICATIONS INDUSTRY** there has been a steadily growing demand for direct-reading frequency meters. Various counting type circuits utilizing thyratrons, multivibrators, etc. have been designed to provide direct frequency indications on a meter scale. General Radio Type 834-A Frequency Meter, for instance, which has been a very popular instrument, is of the thyratron type. The range of this instrument, however, extends only up to 5 kilocycles which, for many applications, is inadequate.

In many laboratory and production measurements, as well as in the monitoring of high-frequency radio transmitters, much higher frequencies must be measured, and, for a general-purpose instrument, a range of about 50 kilocycles is desirable. To provide this greater range the General Radio Company has developed a new circuit* which is considerably simpler than those commonly used.

One of the simplest circuits that can be used for measuring frequency is a resistance-reactance combination, which has a frequency-dependent transmission characteristic. Figure 2 shows the elementary form of such a circuit with a rectifying diode included. When the resistance, *R* (including the diode resistance), is made small compared to the reactance of the capacitor, *C*, the current, and hence the voltage drop across *R*, will be directly proportional to the capacitance and to

*U. S. Patent No. 2,362,503

Figure 1. Panel view of the Type 1176-A Electronic Frequency Meter.



the frequency. Such a device can be made to provide a substantially linear variation in transmission as a function of frequency, which gives a desirable linear meter scale. It has also the important advantage that the calibration will depend mainly upon a single capacitance-resistance combination, and consequently will have a high degree of stability. If, therefore, provision is made to impress upon such a circuit a waveform which is constant in amplitude and wave shape, regardless of frequency, a simple vacuum-tube voltmeter connected across the resistor will give an accurate indication of the transmission and consequently of frequency.

The simplest waveform to generate is a square wave. The problem becomes, therefore, mainly one of designing suitable limiting and wave shape circuits so as to reduce to a minimum any error resulting from changes in amplitude or waveform of the signal applied to the RC circuit.

The new Type 1176-A Frequency Meter operates on this principle. As shown in the elementary schematic diagram of Figure 3, the circuit consists of a series of limiting amplifiers and diodes, with automatic biasing circuits which provide satisfactory operation over a wide range of input voltages ranging from ¼ volt to 150 volts, and for practically any type of waveform which will ordinarily be encountered. A push button is provided on the panel to check that the input voltage is sufficient to insure accurate readings.

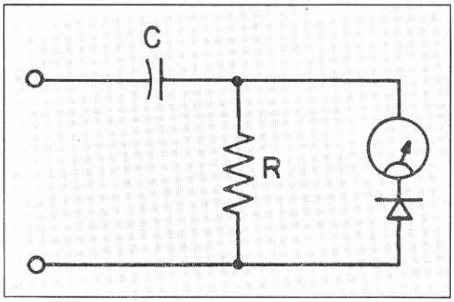


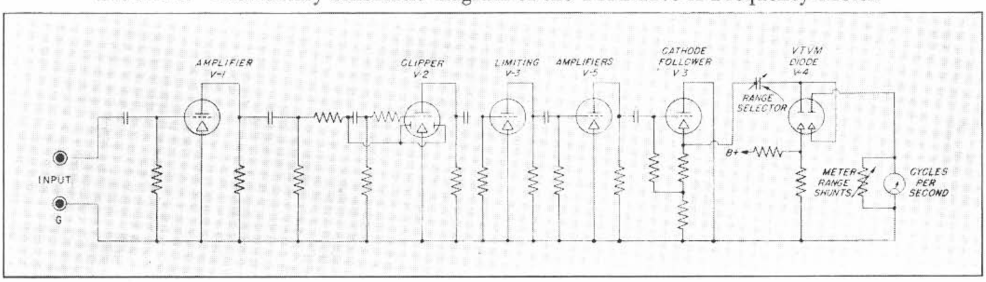
Figure 2. Simple frequency-dependent circuit, consisting of a capacitor and a resistor in series.

The RC circuit is switched to provide six ranges giving full-scale meter deflection of 200, 600, 2000, 6000, 20,000, and 60,000 cycles. The indicating meter is actuated by a full wave vacuum-tube meter circuit, which provides additional compensation for any dissymmetry in the waveform after limiting. The circuit is provided with voltage regulation, and the heater current for the vacuum-tube voltmeter is also regulated. Thus a high degree of stability and accuracy is assured, and the permanence of the calibration is such that no trimmer adjustments are required on the front panel.

The meter scale is linear, and provision is made for the addition of an external extension meter or recorder through a multipoint connector at the rear. Two sets of input terminals are provided on the panel and another set on the multipoint connector at the rear. Plugging into the W. E. panel jacks automatically disconnects the rear terminals. The full-scale current is 0.2 milliamperes.

In a time of rising prices, the sim-

Figure 3. Elementary schematic diagram of the Type 1176-A Frequency Meter.



plicity of the new frequency meter allows it to be sold at a price actually lower than the older narrower range type. While designed originally as a part of a new police and aviation radio monitor, the TYPE 1176-A Frequency Meter will also be catalogued separately to replace the TYPE 834 Electronic Frequency Meter.

— H. H. SCOTT

SPECIFICATIONS

Range: 25–60,000 cycles per second in six ranges. Full-scale values are 200, 600, 2000, 6000, 20,000, 60,000 cycles.

Accuracy: ± 2 cycles $\pm 2\%$ of full scale, for all ranges. When operating on the 60,000-cycle range, with less than 0.5 volt input, the accuracy becomes $\pm 3\%$ of full scale.

Input Voltages: 0.25–150 volts.

Input Resistance: 500,000 ohms, for all ranges. One side grounded.

Input Waveform: The readings are substantially independent of waveform, so long as the dissymmetry of the positive and negative portions of the wave is less than 8:1.

Power Supply: 105–125 (or 210 to 250) volts, 50-60 cycles.

Power Input: Approximately 50 watts.

Vacuum Tubes:

1—type 6H6	1—type 6SN7-GT
1—type 6SQ7	1—type 6J5
1—type 6X5	2—type 6SJ7
1—type 6V6	1—type OA3/VR75

1—Amperite 3-4

Mounting: Relay-rack panel; walnut end frames are available to convert to table mounting.

Accessories Supplied: Spare fuses; spare pilot lamp, multi-point connector, all vacuum tubes.

Dimensions: Panel, 19 x 5¼ inches, depth behind panel, 11¼ inches.

Net Weight: 19½ pounds.

Type		*Code Word*	*Price*
1176-A	Frequency Meter................	TIMID	$185.00

A HANDY PAIR OF BRIDGES

● **IN THE LABORATORY** or on the production line the need frequently arises for the rapid measurement of capacitors or inductors with a moderate degree of precision. For this purpose, a pair of twin units, the TYPE 1614-A Capacitance Bridge and the TYPE 1631-A Inductance Bridge, is now offered, each designed to accomplish a specific purpose at a moderate price consistent with reliable performance and simplicity of operation. To a first approximation these new bridges supply, as separate units and in more portable form, the reactance bridge circuits available in the popular General Radio Impedance Bridge, TYPE 650-A. They do not supersede, but rather supplement the latter.

Each of these self-contained bridges is housed in a covered walnut cabinet measuring 13½″ x 8½″ x 7″ and provided with a handle on one end for easy portability (13 pounds). Mounted on slip hinges, these covers can be opened for exposing the operating panel or removed entirely if desired. Explicit operating instructions are attached to the inside of the cover. While the bridges must be operated with their control panels approximately horizontal, they can be transported or stored in any position.

Either bridge is energized by a one-kilocycle microphone-driven reed "hummer," operated by dry cells which are housed in the cabinet. The hummer unit is provided with a flexible mounting which minimizes injury from shock and reduces the hum tone transmitted to the instrument panel and, hence, into the operating room.

Sufficient bridge sensitivity permits a pair of headphones, without amplification, to be used as the null-balance detector. A pocket compartment is provided in the cabinet for storing these phones. In order that these bridges shall be available for instant use at all times, the phones are internally attached to the instrument so that they cannot conveniently be "borrowed" for another job and thus, perhaps, not be available when wanted.

The main six-inch control dial of each bridge is provided with a slow-motion drive to facilitate accurate adjustment. This dial drives a six-inch rheostat logarithmically wound so that, over the major range of the dial covering two decades of capacitance (1614-A) or of inductance (1631-A), these values can be read with nearly the same fractional accuracy at all scale points. Sufficient overlap is provided at both the upper and lower extremities of this double-decade dial. These large rheostats are equipped with an adjustment cam which permits a differential displacement between the rheostat arm and the control dial, a feature which affords an accurate calibration of each individual dial in the manufacture of these instruments.

The 1614-A Capacitance Bridge

While the Schering bridge provides the most precise measurement of capacitance, it requires expensive adjustable capacitors. Nominal precision can readily be obtained by what is known as the series-resistance bridge circuit depicted in Figure 2. One bridge arm consists of the large logarithmic rheostat controlled by the main dial, which is calibrated directly in microfarads. The multiplier arm consists of one of three fixed resistors selected by a triple position panel switch and providing the following ranges:

Multiplier	*Capacitance Range*
1.0	1 to 100 μf
0.01	.01 to 1 μf
0.0001	.0001 to .01 μf
0.0001	10 to 100 $\mu\mu$f*

*Reduced precision on this lowest decade.

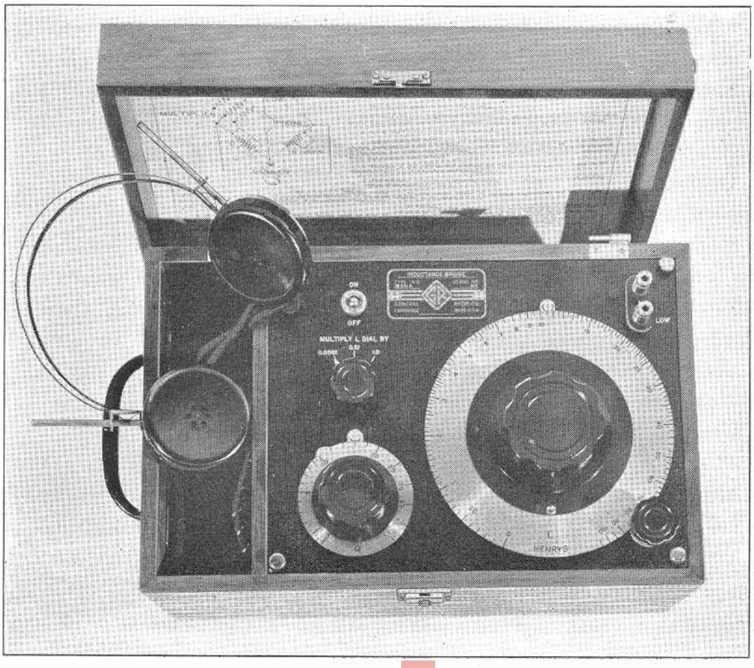


FIGURE 1. View of the TYPE 1631-A Inductance Bridge. Telephones are permanently attached to the instrument, and all necessary operating instructions are mounted in the cover. The TYPE 1614-A Capacitance Bridge is similar in appearance.

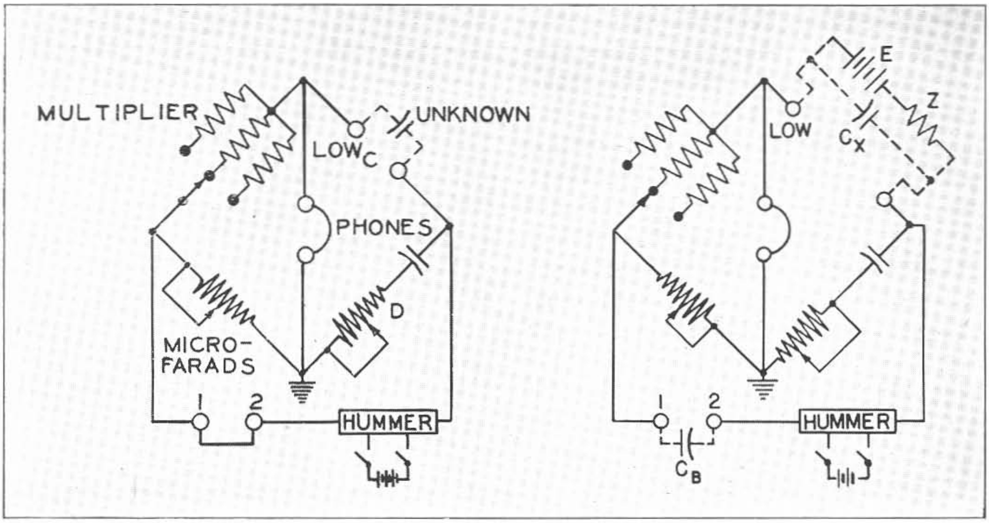


FIGURE 2. Schematic wiring diagram of the TYPE 1611-A Capacitance Bridge.

FIGURE 3. Bridge connections for capacitance measurement with polarizing voltage.

This available range is thus ten millionfold. The multiplier switch also changes (not shown) the operating impedance of the hummer in the interests of sensitivity. The arm opposite the microfarads rheostat contains the unknown capacitor under test, while the arm opposite the multiplier contains a high grade standard capacitor in series with a smaller logarithmic rheostat D. This rheostat is adjusted by the smaller control dial which is calibrated directly in the dissipation factor of the unknown over a working range from 0 to 0.45. Bridge balance is easily attained by the joint manipulation of the MICROFARADS and D dials, with the multiplier initially set for the appropriate range.

The necessary parameter relationships to achieve a balance of this capacitance bridge are expressed by the two simultaneous equations:

$$C_x = \left(\frac{C_1}{R_3}\right) R_2 \qquad (1)$$

and

$$D_x = (2\pi f C_1) R_1 \qquad (2)$$

where C_1 is the standard capacitor, R_1 the resistance (small rheostat) in series with C_1, R_2 is the large rheostat and R_3 is the multiplier value used. For a fixed value of C_1 it will be seen that the existence of R_2, in Equation (1) only, permits the large dial to be calibrated in terms of C_x for a specific C_1/R_3 ratio, while changing the multiplier R_3, also uniquely in Equation (1), by double decade steps changes one hundredfold the magnitude of C_x for any specific R_2 value. Likewise the existence of R_1, in Equation (2) only, permits the small dial to be calibrated directly in terms of D_x for a specific value of frequency f, — in this case one kilocycle.

By a slight modification indicated in Figure 3, the 1614-A Bridge can be used to measure electrolytic capacitors having an applied polarizing voltage. A suitable d-c voltage source E in series with a resistor Z is applied, with the correct polarity, across the terminals of C_x. To eliminate errors, the value of Z should exceed 100 times the reactance of C_x at 1000 cycles. If C_x passes any leakage current, the actual voltage on C_x will be the value of E diminished by the drop in Z. For this purpose, a jumper connecting the internal points 1 and 2 must be removed and a capacitor C_B (not provided) of 2 μf (or larger) inserted in this position. Space is available for storing this capacitor within the cabinet. For ordinary uses of the bridge, C_B (Figure 2) should be removed and the jumper replaced to achieve maximum sensitivity and precision of balance.

The 1631-A Inductance Bridge

There are several bridge circuits, such as the Owen, Hay, etc., available for the measurement of inductance, but for measuring inductors whose Q value does not exceed about 50, the most convenient circuit is that of the Maxwell bridge depicted in Figure 4. One bridge arm consists of the large logarithmic rheostat controlled by the main dial which is calibrated directly in henrys. Opposite this is the multiplier arm comprising one of three fixed resistors selected by a triple-position panel switch and providing the following ranges:

Multiplier	*Inductance Range*
1.0	1 to 100 henrys
0.01	.01 to 1 henrys
0.0001	.0001 to .01 henrys
0.0001	10 to 100 microhenrys*

The available range is thus ten millionfold. The multiplier switch also changes (not shown) the operating impedance of the hummer in the interests of sensitivity. A third arm of this bridge consists of the unknown inductor under test. The opposite arm contains a high-grade standard capacitor which is shunted by a smaller logarithmic rheostat Q. This rheostat is adjusted by the smaller control dial which is calibrated directly in terms of the storage factor of the unknown over a working range from 0 to 45. With the multiplier initially set for the appropriate range, a bridge balance is easily attained by a joint manipulation of the henrys and Q dials.

*Reduced precision on this lowest decade

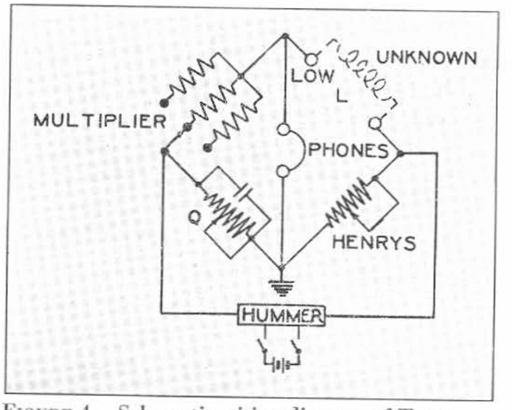


FIGURE 4. Schematic wiring diagram of TYPE 1631-A Inductance Bridge.

A balance of this Maxwell bridge is indicated by the two simultaneous equations:

$$L_x = (R_3C_1)R_2 \qquad (3)$$
$$Q_x = (2\pi fC_1)R_1 \qquad (4)$$

wherein C_1 is the standard capacitor, R_1 the resistance (small rheostat) in parallel with C_1, R_2 is the large rheostat and R_3 is the multiplier value used. Starting with a fixed C_1 value, since R_2 is uniquely in Equation (3) the large dial can be calibrated in terms of Lx for a specific R_3C_1 product, while changing the multiplier R_3 by a double decade step modifies one hundredfold the magnitude of C_x for any specific R_2 value. Likewise, the existence of R_1, in Equation (4) only, permits the small dial to be calibrated in terms of Q_x for a specific value of frequency f, — in this case one kilocycle.

Parenthetically, Equation (4) sets a maximum limit to the Q_x value attainable, with a specific fC product, as determined by the maximum resistance R_1 for which it is practical to wind a calibrated rheostat. Higher values of Q_x may be measured by the Hay bridge in which the standard capacitor and the small-valued Q rheostat are in series. The Hay bridge, however, requires a troublesome correction factor to be applied to the inductance scale values of low Q inductors. Hence the Maxwell bridge was chosen for this purpose, on the assumption that rarely do the 1-kc Q values of inductors exceed 45.

It should be noted that when iron-cored inductors are measured on such a bridge as this, having no control over the applied generator voltage, the L and Q values obtained are the 1 kc values corresponding to an arbitrary degree of magnetization in the core which is indeterminate unless a vacuum-tube voltmeter is applied across the terminals of the inductor in the balanced bridge. If the ferromagnetic core does not contain an appreciable air gap, this indeterminate magnetization will, in general, considerably exceed that corresponding to initial permeability.

— Horatio W. Lamson

SPECIFICATIONS FOR TYPE 1614-A CAPACITANCE BRIDGE

Capacitance: Range, 10 $\mu\mu$f to 100 μf in three steps: 10 $\mu\mu$f to 10,000 $\mu\mu$f; 0.01 μf to 1.0 μf; and 1.0 $\mu\mu$f to 100 μf.

Accuracy: $\pm 2\%$, except on the lowest range, where, after the zero capacitance of 9 $\mu\mu$f is subtracted, the accuracy is $\pm(2\mu\mu\text{f} + 2\%$ of the dial reading).

Dial Calibration: Approximately logarithmic (uniform fractional accuracy) over two main decades, with a compressed lower decade which is used only for measurements below 100 $\mu\mu$f.

Dissipation Factor: Range, 0 to 45%. Accuracy: On the lowest range, the error, expressed in per cent dissipation factor, is $\pm(2\% + 0.1 \times$ dial reading); on the other two ranges, $\pm(0.2\% + 0.1 \times$ dial reading).

Frequency: The internal oscillator furnishes the necessary bridge power at a frequency of 1000 cycles $\pm 5\%$.

Power Supply: 6-volt dry battery. Two Burgess F2BP units connected in series are recommended, and are supplied with the instrument. Space for these is provided in the cabinet.

Accessories Supplied: Head telephones and batteries.

Accessories Required: When a d-c polarizing voltage is used, a 2 μf blocking capacitor is required. This condenser is not supplied with the instrument, but space for a General Electric Type 55X-629 is provided in the cabinet.

Mounting: Walnut cabinet with removable hinged cover.

Dimensions: 13½ x 8½ x 7 inches, overall.

Net Weight: 13¼ pounds.

Type		*Code Word*	*Price*
1614-A	Capacitance Bridge..............	LAPEL	$90.00

SPECIFICATIONS FOR TYPE 1631-A INDUCTANCE BRIDGE

Inductance: Range, 10 μh to 100 h in 3 steps, 10 μh to 10,000 μh; 0.01 h to 1 h; and 1 h to 100 h.

Accuracy: $\pm 2.5\%$ of dial reading between 100 μh and 10 h. Below 100 μh the error varies inversely as the magnitude of the unknown. Above 10 h the error increases to $\pm 10\%$ dial reading at 100 h.

Dial Calibration: Approximately logarithmic (uniform fractional accuracy) over two main decades, with a compressed lower decade which is used only for measurements below 100 μh.

Q: Range, 1 to 45. Accuracy, $\pm 10\%$ of dial reading for values of Q between 2 and 10. For higher values the error increases progressively to $\pm 15\%$ at a Q of 45. For lower values, the error increases to $\pm 20\%$ at a Q of 1.

Other specifications are identical with those for Type 1614-A Capacitance Bridge.

Type		*Code Word*	*Price*
1631-A	Inductance Bridge...............	LARVA	$98.00

PERSONNEL

Kipling Adams, who takes charge of General Radio's Chicago office.

● **MESSRS.** Hermon H. Scott, Lucius E. Packard and Raymond W. Searle have severed their connections with General Radio to form a new company for the present purpose of manufacturing technical equipment of a type not made by General Radio. They have located in Waltham, Massachusetts.

Mr. Scott, who came with us in 1931, was a development engineer, and Mr. Searle was one of our production foremen. Mr. Packard, who joined our engineering staff in 1936, has been in charge of our District Office at 920 South Michigan Avenue in Chicago since its organization over two years ago.

That very active office is now managed by Mr. Kipling Adams, who for the past several years has been Assistant Manager of the Service Department. Mr. Adams received his technical education at the Massachusetts Institute of Technology and came to the General Radio Company in 1934. During his first few years here he was with the Calibration Laboratory where final tests and calibrations are made on our instruments, and since then he has been in the Service Department. These activities have given him an intimate and well-rounded knowledge of General Radio products, which qualifies him to be of the maximum assistance to our many friends in the Chicago and Middle Western area.

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THE General Radio EXPERIMENTER

VOLUME XX No. 10 MARCH, 1946

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A CONVENIENT AMPLIFIER AND NULL DETECTOR

Also IN THIS ISSUE

● **MOST BRIDGE MEASUREMENTS** and various other laboratory procedures require the use of an amplifier to obtain sufficient sensitivity. The TYPE 814-A Amplifier has for years been a popular item in the General Radio line and has been widely used with audio-frequency bridges.

Recent developments in tubes and circuits have prompted the design of a new model, retaining all the advantages of the old, but with several distinct improvements.

The TYPE 1231-A Amplifier and Null Detector consists of a high-gain resistance-coupled amplifier, using the new miniature-type tubes, mounted on a shock-absorbing suspension. The maximum gain in the

FIGURE 1. Panel view of the TYPE 1231-A Amplifier and Null Detector.

middle of the audio-frequency range is approximately 90 db, and the instrument is usable over the range from 20 cycles to 100 kc.

A most important feature of the new instrument is the built-in null-indicator circuit, which is essentially a semi-logarithmic vacuum-tube voltmeter, utilizing a single multi-section tube as an a-c amplifier, diode voltmeter, and d-c amplifier, thus allowing a high degree of sensitivity. The voltage developed by the diode is applied as a gain-controlling grid bias to the preceding a-c amplifier section of the tube producing the semi-logarithmic response. No additional indicating devices are necessary, therefore, to use the new instrument as a bridge null indicator, although a pair of phones can be plugged into the output if desired.

The amplifier and null detector may also be made selective with regard to frequency by plugging in the Type 814-P Tuned Circuits. The Type 814-P2 operates at either 400 or 1000 cycles, and the Type 814-P3 at 60 cycles. Other types of tuned circuits or filters can also be used with the instrument.

The new amplifier and null detector is unusually compact and convenient to operate. Normally, the instrument is enclosed in a small walnut case, matching other General Radio equipment, and is operated from internal batteries which will have a long life because of the low current drain. It is also possible where desirable to operate the instrument from the 60-cycle lines by use of the Type 1261-A Power Supply unit. This is the same power supply that is used for operating the General Radio Sound-Level Meters and other battery instruments.

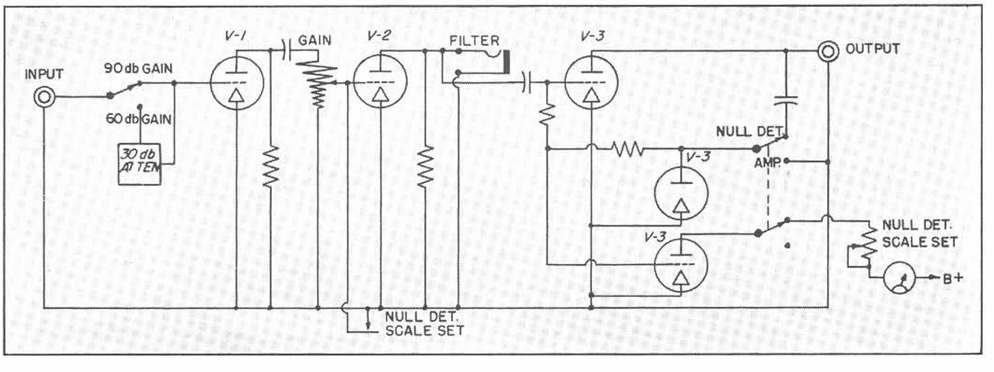
For relay-rack mounting, a panel extension can be provided which mounts the two standard Type 814-P Filters also, thus providing a complete unit assembly.

The main gain control of the amplifier is a high-grade, wire-wound unit, which will last indefinitely with normal use. Push buttons are provided to reduce the input voltage and gain for high-level signals and for selecting operation as a straight amplifier or as a null detector with the semi-logarithmic features. Other push buttons allow checking of the battery voltage and exact setting of the null point for critical bridge measurements.

The input and output connections will take either General Radio Type 774-E Coaxial Connectors or the usual Type 274-M Plugs.

Figure 3 shows the frequency characteristics of the amplifier and complete specifications are appended to this article. The instrument was purposely designed for maximum gain in the

Figure 2. Schematic circuit diagram of the Type 1231-A Amplifier and Null Detector.



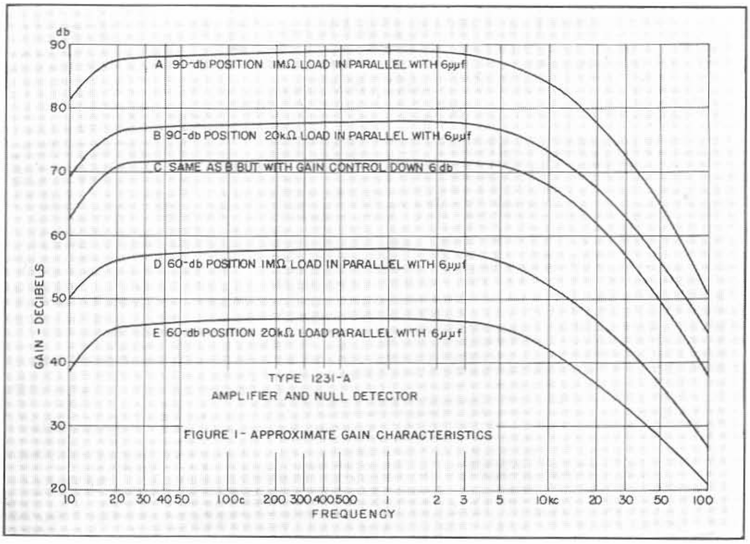


FIGURE 3. Approximate gain characteristics as a function of frequency.

important audio-frequency range, but, for operation as a bridge null detector, it will provide substantial gain well beyond this range.

It is expected that the TYPE 1231-A Amplifier and Null Detector, because of its small size, high sensitivity, and general convenience, will meet a real need in most communications laboratories.

— H. H. SCOTT, W. F. BYERS

SPECIFICATIONS

Input Impedance: 1 megohm in parallel with 50 micromicrofarads.
Maximum Gain: 90 db with 1 megohm load — down 6 db at 12 and 11,000 cycles — 24 db down at 50 kc.
Null Detector Sensitivity: Less than 100 microvolts input is required to give 10% indication on the meter at 1 kc.
Output Impedance: Approximately 50,000 ohms.
Output Voltage: 5 volts into 20,000 ohms; 20 volts into 1 megohm.
Battery Life: Between 200 and 250 hours at 8 hours per day.
Frequency Response: See curves.
Noise Level: Less than 0.5 volt at full gain with battery operation; less than 1.0 volt at full gain with a-c power supply.
Tubes: The instrument requires two TYPE 1L4 and one TYPE 1D8GT Tubes which are supplied in the instrument.
Power Supply: Burgess TYPE 6TA60 (Signal Corps BA48) Battery Pack is supplied in place in the instrument. When a-c supply is desired, TYPE 1261-A Power Supply can be used.
Accessories Available: TYPE 814-P2 (400 and 1000 cycles) and TYPE 814-P3 (60 cycles) Tuned Circuits are available for providing selectivity. These were described in Catalog K, first (page 112) and second (page 90) editions. For facilitating connections to the input and output, two TYPE 274-M Plugs are supplied. TYPE 274-NC or TYPE 274-NE Shielded Connectors may be used. Where complete shielding is required, TYPE 774 Coaxial Connectors are recommended.
Dimensions: 12¼ x 8 x 10¾ inches, overall.
Net Weight: 23¼ pounds, including batteries.

Type		*Code Word*	*Price*
1231-A	Amplifier and Null Detector......	VALID	$160.00

DECIBELS

We have received through the courtesy of the author, Mr. V. V. Rao of Madras, India, a copy of his recent book, entitled "The Decibel Notation," a comprehensive treatise covering the history, details, and application of the decibel concept. This book is a welcome addition to the communications engineer's library, since it embraces, in convenient form, a considerable fund of information on a subject that is usually treated only briefly in engineering texts. Detailed reviews have appeared in many of the electronic periodicals.

TYPE 1261-A POWER SUPPLY

An a-c operated power supply for instruments which use the Signal Corps BA48 Battery Block (Burgess Type 6TA60)

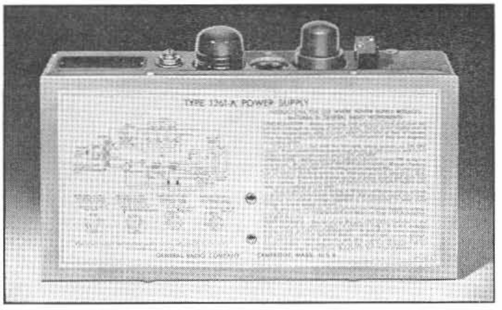


FIGURE 1. View of the TYPE 1261-A Power Supply.

● **BATTERY OPERATION** of a measuring instrument is necessary if the instrument is to be truly portable and easily used in any location. Many instruments designed primarily for portable use can be and are used in production test set-ups where they are in continuous use. A-c operation for such instruments would eliminate the need for frequent battery replacement, but would limit portability.

The TYPE 759-B Sound-Level Meter is a good example of a portable battery-operated instrument which is often used in production testing. Early in 1942* the TYPE 759-P50 power supply was brought out as an a-c operated power source which was interchangeable both electrically and mechanically with the battery block in the sound-level meter. Thus, the sound-level meter became readily adaptable for either production work or for field use. This power supply could also be used in the older TYPE 759-A Sound-Level Meter.

Other instruments intended for portable use, and hence battery operated, have been developed and manufactured since the TYPE 759-P50 Power Supply became available. A number of these instruments use the same battery block as the TYPE 759-B Sound-Level Meter, but differences in plate or filament supply loads and other small details prevented the use of the TYPE 759-P50 Power Supply in these later instruments. For all these instruments, a new power supply, the TYPE 1261-A, has been developed which incorporates the changes required to make a more complete replacement for the BA-48 battery block.

This new supply, by means of suitable selector plugs, can be used in the following General Radio instruments: TYPE 759-A and TYPE 759-B Sound-Level Meters, TYPE 720-A Heterodyne-Frequency Meter, and TYPE 1231-A Audio Amplifier and Null Detector. Octal selector plugs inserted into a socket on the top of the power supply are used to select either a 1.5- or 3.0-volt filament supply and, in cases where the current drain of the instrument to be supplied is less than required for normal operation of the power supply, the plug is used to add load resistors which insure normal operation.

The TYPE 1261-A Power Supply is not limited to use in General Radio instruments. An unwired selector plug, which can be wired by the customer, is available so that the power supply can be used in any battery operated instrument which has plate and filament requirements within its scope. Intended as a general purpose replacement for the BA48 battery block, the power supply is a light and compact unit that fits in the battery compartment, and has a four-

*See "An A-C Operated Power Supply for the Sound-Level Meter," by H. H. Scott, General Radio *Experimenter*, January 1942, Vol. XVI, No. 8.

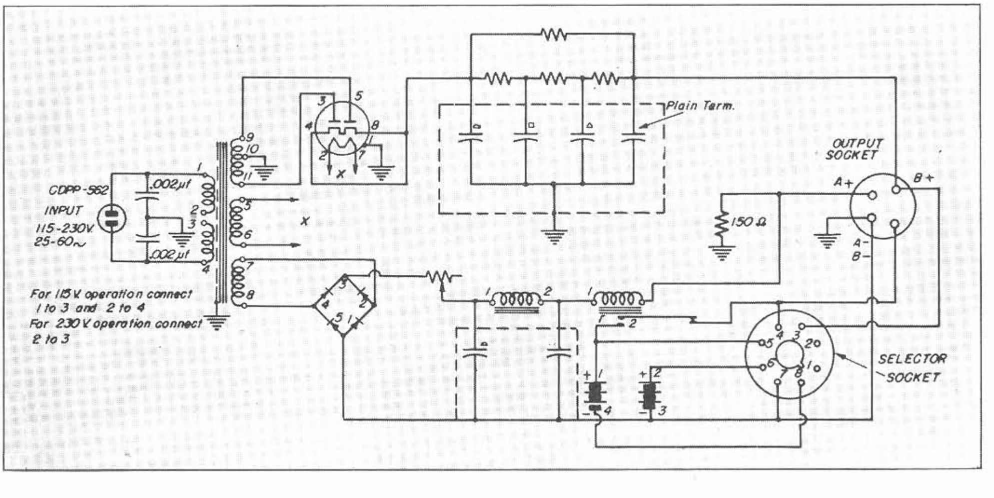
terminal output socket which fits the plug on the battery cable of instruments using the BA48 battery.

The filament supply, which consists of a selenium rectifier with an L-C filter, has two small flashlight cells connected across the output. These cells are connected in series for 3.0-volt output, and in parallel for 1.5-volt output by means of the selector plug. The cells act as a regulator for the filament supply and also provide an extremely low a-c output impedance. Hum and low-frequency changes in output caused by a-c power line fluctuations are reduced to the point where satisfactory operation of the high-gain amplifiers used in the sound-level meters and in the TYPE 1231-A Audio Amplifier and Null Detector is obtained.

A small push button switch, located on top of the power supply, makes it possible to disconnect the cells at any time and to set the output voltage equal to the cell voltage. Under operating conditions, normal line voltage variations cause the cells either to charge or to discharge slightly, and when the instrument is turned off a small relay opens the circuit so that the cells will not be damaged. These flashlight batteries cost only ten cents apiece and, as used, their life is extremely long. A conventional vacuum-tube rectifier is used to obtain the d-c output voltage for the plate supply. A four section R-C filter attenuates the hum or a-c ripple voltage to an exceptionally low value. The plate supply is not regulated.

The TYPE 1261-A Power Supply replaces the older TYPE 759-P50 for use in the TYPE 759-A and TYPE 759-B Sound-Level Meters, but is not directly interchangeable with it. A modification incorporated in the new power supply to protect the small flashlight cells, used in the filament circuit, requires a simple wiring change in the battery cable of older instruments. This change does not affect battery operation of the instrument and has been made standard wiring for current and future lots of instruments. Complete instructions are furnished with the power supply so that the customer can make the required change in older instruments, and once the change is made the TYPE 1261-A Power Supply is directly interchangeable

FIGURE 2. Schematic circuit diagram of the power supply.



with the BA48 battery block. Similar changes must be made in existing models of the TYPE 720-A Heterodyne Frequency Meter. For instruments shipped in the future, no changes are necessary.

— E. E. GROSS

SPECIFICATIONS

OUTPUT:

Filament Supply: 1.5 volts or 3.0 volts up to 350 ma. Normal current through filter choke to operate relay = 300 ma. Bleeder resistor in selector plug needed for lower current requirements.

Plate Supply:

133 volts open circuit	For 115-volt 60-cycle power line with normal filament current of 300 ma.
107 volts at 3 ma	
89 volts at 5 ma	
72 volts at 7 ma	

Maximum output current = 8 ma

Selector Plugs: One of the following is furnished. Please specify type wanted.

Selector Plug 1261-P1 — Provides proper voltages for TYPE 759-A Sound-Level Meter. Battery Plate of Sound-Level Meter must be replaced by four-terminal plug to fit output socket of TYPE 1261-A Power Supply. Full sensitivity of instrument cannot be used. Attenuator settings below 50 db on B and C weighting networks and below 40 db on the A weighting network are not recommended.

Selector Plug 1261-P2 — Provides proper voltages for TYPE 759-B Sound-Level Meter.

Selector Plug 1261-P3 — Provides proper voltages for TYPE 720-A Heterodyne-Frequency Meter.

Selector Plug 1261-P4 — Provides proper voltages for TYPE 1231-A Audio Amplifier and Null Detector. On Null Detector use, the plate supply regulation causes the meter to overshoot somewhat upon rapidly approaching a null.

Selector Plug 1261-P5 — To be wired by customer to meet his own requirements.

Hum and Noise Level — Sufficiently low, when operated from 60-cycle supply line, to assure satisfactory operation of instruments listed under conditions specified.

Input Voltage: 105–125, or 210–250, volts, 40 to 60 cycles.

Input Power: Less than 10 watts at 115 volts, 60 cycles.

Tube: One TYPE 6H6 is supplied.

Batteries: Two Burgess No. 2 uni-cells which are floated across the output of the Filament Supply are furnished.

Terminals: A four-terminal output socket fits the plug on the battery cable of the TYPE 759-B, TYPE 720-A, and TYPE 1231-A.

Dimensions: (Length) 10 x (width) 2¼ x (depth) 5 inches.

Net Weight: 7¼ pounds.

Type		*Code Word*	*Price*
1261-A	Power Supply	NUTTY	$55.00

When ordering, specify type of selector plug desired. See list above in specifications. Also mention type number and serial number of instrument with which it is to be used.

TYPE 1260-A VARIAC RECTIFIER

● **DURING THE PAST FEW YEARS** selenium rectifiers have come into wide use as a means for obtaining d-c output voltage from a-c supply lines. They are extremely simple, rugged, compact, and efficient, and they have practically unlimited life. Being conservatively rated, they will carry heavy overloads without damage. In most of its applications the selenium rectifier is chosen to do a particular job, wired permanently into a circuit, and then forgotten. However, there are many cases, especially in standardizing laboratories and in development work, when a continuously variable source of d-c voltage is very useful. Storage batteries are not too convenient and usually require charging just when they are needed. Furthermore, the output voltage of a storage battery is not easily adjustable. The development engineer often needs varied d-c voltage

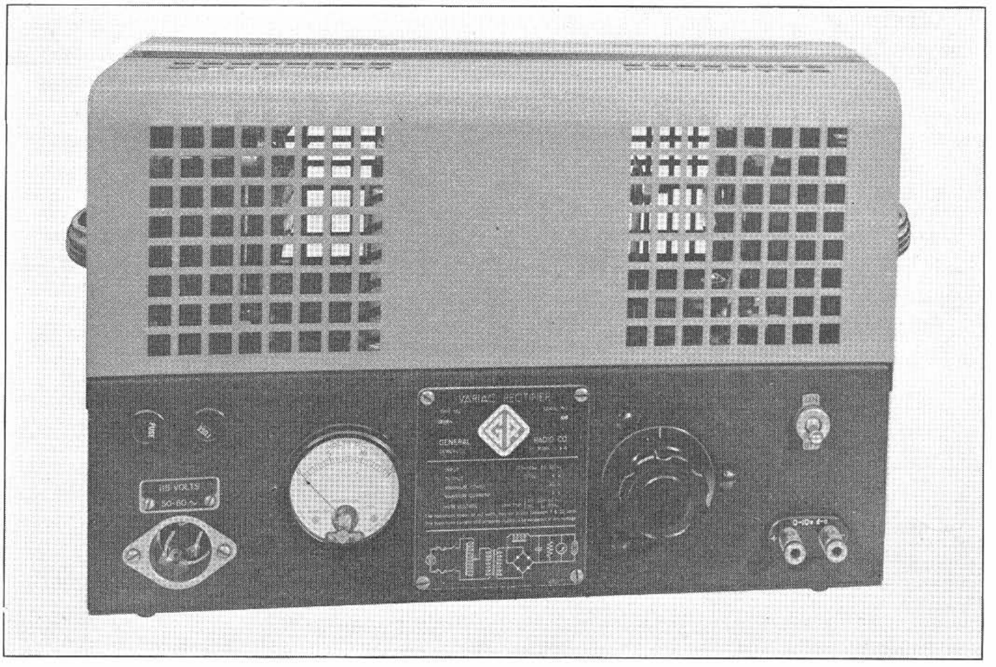


FIGURE 1. View of the TYPE 1260-A Variac Rectifier.

supplies. His requirements may be 1.0 volt at 4.5 amperes filament supply for a power oscillator today, 10.0 volts at 20 ma for a bridge supply tomorrow, 6.0 volts at 5 amperes to test a small d-c constant-speed motor the next day, and so on for a large number of relatively short-time applications, no two of which have the same requirements.

It was to meet these needs that the TYPE 1260-A Variac-Rectifier was developed. This power supply is a combination of a TYPE 200-B Variac, a step-down transformer, a selenium rectifier, a filter, and an output voltmeter arranged electrically as shown in the name-plate wiring diagram (see Figure 1).

By supplying continuously adjustable primary voltage between 0 and 115 volts to the step-down transformer, the TYPE 200-B Variac provides extremely smooth control of the d-c output voltage from 0–15 volts. A maximum power of 40 watts may be drawn from the power supply between 7 and 10 volts, but the maximum current of 6 amperes should not be exceeded at any voltage. There are many applications where the unfiltered d-c output from the rectifier would be adequate; however, the possible uses of the power supply are increased many fold if the d-c output is filtered so that the a-c ripple is a small part of the output. The output filter of the TYPE 1260-A Variac-Rectifier reduces the ripple or hum voltage to a value low enough for most applications. The d-c internal output impedance of the power supply is approximately 0.6 ohm. This means that there will be an increase in output voltage of about 2.4 volts if a load drawing a current of 4 amperes is removed.

The TYPE 1260-A Variac-Rectifier is a compact and convenient power unit. It is mounted in a sturdily constructed steel case; it is light in weight and easily portable. The various components have been carefully selected to obtain maximum output ratings and yet insure long life and trouble-free operation. Several of these units, built a few years ago, have been used in many ways in our standardizing laboratory and in our engineering laboratories as replacements for storage batteries. They have been operated for long periods at maximum ratings with no appreciable drift in output voltage. These power supplies were designed to be convenient, useful sources of low-voltage d-c power, and our experience with them over the past few years has proved that the design objectives were well met.

— E. E. GROSS

SPECIFICATIONS

Output Range: 4 amperes at 0–10 volts, dc; maximum power, 40 watts; maximum current, 6 amperes; no load voltage, over 15 volts.

Meters: The output voltage is indicated by a voltmeter mounted on the cabinet. A panel knob controls the output voltage.

Power Supply: The unit will operate from a 105- to 125-volt, 50- to 60-cycle line. A suitable power cord is supplied with the instrument.

Power Input: At the full 40-watt load, the power input from the a-c line is about 75 watts.

Hum Voltage: At 10 volts, 4 amperes, the hum voltage is less than 100 millivolts or 1% of the output voltage when the instrument is operated on a 60-cycle line. At 2 volts, 6 amperes, the hum voltage is less than 60 millivolts or 3% of the output voltage when the instrument is operated on a 50-cycle line. For lower load currents, the hum decreases.

Accessories Supplied: Spare fuses.

Mounting: The instrument is mounted in a metal cabinet suitable for table use.

Dimensions: (Length) 16 x (depth) 7 x (height) 9½ inches, overall.

Net Weight: 26¾ lbs.

Type		*Code Word*	*Price*
1260-A	Variac-Rectifier.................	VALET	$90.00

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THE General Radio EXPERIMENTER

VOLUME XX No. 11 APRIL, 1946

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

HAVE YOU A TYPE 650 IMPEDANCE BRIDGE?

Also
IN THIS ISSUE

● **ONE OF THE MOST POPULAR** General Radio laboratory instruments, since it was first introduced in 1933, has been the Type 650 Impedance Bridge — witness a sale of over 3700 of these instruments to date. By means of this bridge resistors, inductors and capacitors can be measured rapidly and conveniently over a wide range of values.

On many occasions, the need arises for using an auxiliary audio amplifier to increase the sensitivity of bridge balance when employing either headphones or an a-c galvanometer as the null detector. For this purpose General Radio has successively offered the Type 514 and Type 814 Amplifiers, and at the present time the Type 1231 Amplifier and Null Detector.

The Type 650 Impedance Bridge is self-energized by four No. 6 dry cells housed in the bridge cabinet. These dry cells supply d-c power for the Wheatstone bridge measurement of resistance and drive a one kilo-

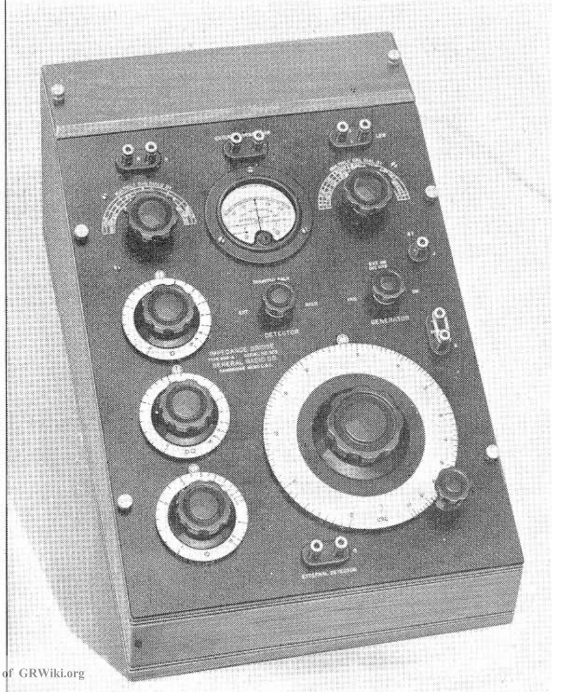


Figure 1. View of the Type 650-A Impedance Bridge with battery compartment at the rear.

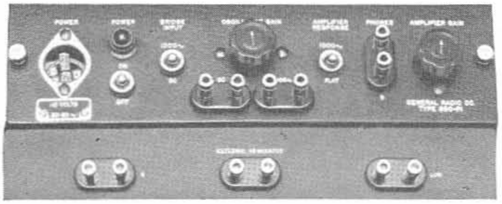


FIGURE 2. Panel view of the TYPE 650-P1 Oscillator-Amplifier, showing top of bridge panel. Connections are easily made from either the d-c or 1000-cycle terminals to the EXTERNAL GENERATOR terminals of the bridge.

cycle microphone hummer for bridge measurements of either inductors or capacitors. While such an arrangement is essential for field use where power lines are not available, most of these bridges are used exclusively in laboratories where a-c mains are always available, in which case it would be desirable to dispense with the dry cells and operate the bridge directly from the power line.

To meet these two desirable requirements, the General Radio Company now offers the TYPE 650-P1 Oscillator-Amplifier, a useful combination unit which is designed to fit into the cabinet compartment formerly housing the dry cells. This 650-P1 comprises:

(1) A source of d-c voltage for resistor measurements.

(2) A one-kilocycle vacuum-tube oscillator to replace the microphone hummer.

(3) An amplifier of sufficient gain for all uses of the bridge.

This auxiliary unit is energized from a single-phase a-c power line at a frequency of 50 to 60 cps, and having a nominal voltage of either 115 or 230. The power consumed is about 10 watts. The unit is assembled in a metallic cabinet with a top control panel which replaces the wooden cover of the battery compartment.

When installed, a short plug-terminated jumper connects the bridge input terminals to adjacent d-c or one-kilocycle terminals on the panel of the auxiliary unit. By means of an internal shielded cable a connection is made between the bridge detector terminals and the input of the amplifier. The phones or a-c galvanometer (neither of which are supplied) that are to be used as the null detector are then attached to terminals on the auxiliary unit. The TYPE 483-F Output Meter is a useful a-c galvanometer for this purpose. A complete set of instructions for installing and operating the TYPE 650-P1 is supplied.

The power-supply rectifier is a 6H6 duplex diode functioning as a voltage doubler. The maximum permissible value of the filtered d-c output is applied to the bridge. This is considerably in excess of the 6 volts originally available from batteries, so that the sensitivity of balance when measuring the larger values of resistance (one megohm maximum) is definitely enhanced using the galvanometer already incorporated in the TYPE 650 Bridge.

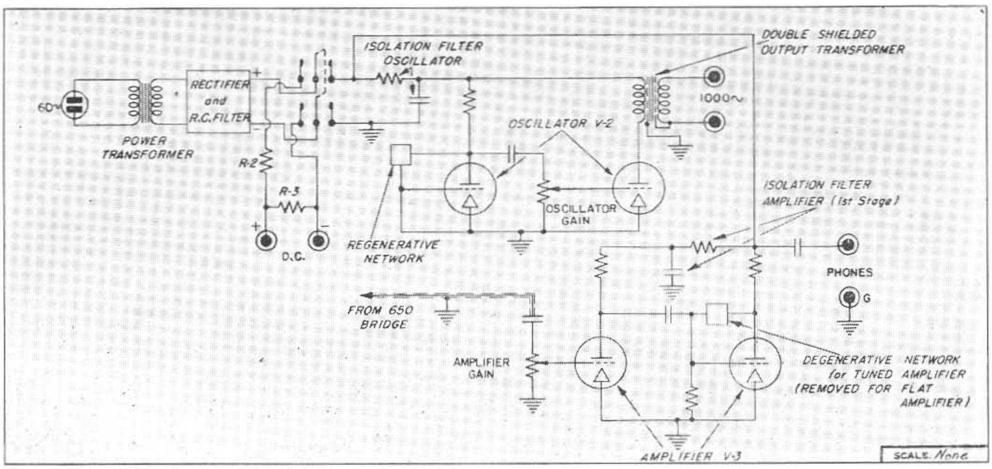
The one kilocycle oscillator employs a 6SL7GT duplex triode, one triode functioning as an oscillator and the other as a buffer amplifier. The oscillator is of the R-C feedback type and can be more accurately tuned to one kilocycle than is practical with the microphone hummer, thus minimizing inherent frequency errors in the reading of the Q or D dial of the bridge, since the calibration of these dials is a function of the operating frequency. The oscillator frequency is subject to a 20-cycle warm-up rise over a one-hour interval and is adjusted at thermal equilibrium to within ± 5 cycles of one kilocycle. Between the oscillator and the buffer amplifier is inserted an oscillator gain control permitting a full range adjustment of the a-c voltage applied to the bridge. This feature, which is not available in the original TYPE 650 Bridge unless a separate oscillator is used, is desirable in measuring iron-cored inductors at low flux densities approach initial permeability.

Between the output of the buffer amplifier and the bridge input terminals is interposed a 5/1 step-down transformer which is doubly shielded and designed to give maximum over-all efficiency with the TYPE 650 Bridge and to minimize certain residual capacitance errors. This transformer contains separate electrostatic shields around the primary and secondary windings with a substantial separation between them. The primary shield is grounded, while the secondary shield is connected to that secondary terminal which leads to the junction of the *A* and *N* arms of the bridge. This arrangement places a negligible capacitance of 9$\mu\mu$f across the standard capacitor (10,000 $\mu\mu$f) bridge arm, and introduces a capacitance of less than 36 $\mu\mu$f (4.4 megohms reactance) across the CRL rheostat arm (10 kilohms maximum), so that the error introduced is also negligible.

The full output of the oscillator somewhat exceeds that of the microphone hummer and has a distinctly purer wave form. Twenty milliwatts into a matching load of 2 kilohms may be obtained with a harmonic content of less than 2% at full output. The open circuit voltage is in excess of 12 volts, and the hum level is less than 15 millivolts.

The amplifier comprises two stages utilizing a second 6SL7GT tube. The first stage is tuned by means of a degenerative R-C network to afford peak gain and maximum selectivity at one kilocycle, thus facilitating bridge balances by minimizing harmonics, especially when measuring nonlinear circuit elements such as iron-cored inductors. An open circuit voltage-ratio gain in excess of 50 db is available with a discrimination in excess of 10 db at 0.5 and 2 kilocycles. With average phones a gain of about 45 db is realized. This amplifier is preceded by a gain control affording full range adjustment of detector sensitivity. The hum level is less than 10 millivolts. A blocking capacitor removes any d-c component from the output to the phones or a-c galvanometer and par-

FIGURE 3. Elementary schematic circuit diagram of the TYPE 650-P1 Oscillator-Amplifier.



tially resonates the average headphones to one kilocycle. Crosstalk between the oscillator and amplifier components is reduced to an imperceptible minimum by the use of isolation filters and compartment shielding.

In order that this same amplifier may conveniently be used when making a bridge measurement at some frequency other than one kilocycle (using, of course, some other oscillator) a switch is provided whereby the tuning network may be removed, thus giving the amplifier a flat gain characteristic. This flat amplifier has a gain of about 60 db at one kilocycle, dropping by 0.6 db at 100 cycles and 3.5 db at 5 kilocycles. The hum level is less than 400 millivolts.

The hum levels for the two amplifiers quoted above are obtained when employing a resistive load such as the TYPE 483 Output Meter. Due to resonance selectivity, when using head phones these hum levels are reduced about 20 db and become audibly imperceptible.

On the control panel of the TYPE 650-P1 are the following items: an input socket for attachment to an a-c power line together with a line switch and pilot light; d-c, l-kc and phone terminals; a switch for selecting either d-c or 1-kc excitation of the bridge; a switch for rendering the amplifier flat or tuned to 1 kc; and separate control knobs (uncalibrated) for adjusting the oscillator output level and the amplifier gain.

While the TYPE 650 Impedance Bridge will continue to be sold equipped with the microphone hummer and dry batteries, the addition of this new auxiliary unit should prove beneficial to many old and new users of this bridge.

The TYPE 650-P1 Oscillator-Amplifier is specifically designed to be used with the TYPE 650 Bridge. However, as an individual item, it should prove useful with many other bridge systems, providing, in compact form, both a one kilocycle oscillator for exciting such a bridge and an amplifier for the bridge detector. The unit can also serve as a convenient source of well-filtered d-c power (but cannot be used simultaneously as an oscillator-amplifier). The open-circuit d-c voltage is 190 volts with a hum level less than 0.1 volt, and the regulation is closely linear and represented by the equation:

$$\text{Terminal Voltage} = 190 - 23I$$

wherein I is the load current in milliamperes. No harm is done by short-circuiting the d-c terminals which affords a maximum current of 8 milliamperes.

— HORATIO W. LAMSON

SPECIFICATIONS

Oscillator: Frequency — 1000 cps ±1%; Harmonics — less than 2% at full output; Open-circuit Voltage — continuously adjustable up to maximum of 12 to 15 volts; Internal Impedance — 2000 ohms; Hum Level — 15 millivolts.

Amplifier: Voltage Gain (with average phones) — continuously adjustable up to about 45 db; Attenuation to Second Harmonic (when tuned to 1000 cycles) — approximately 15 db. Hum Level — inaudible.

D-C Output: Maximum Current — 8 milliamperes, no adjustment provided; can be short-circuited without damage; Hum Level (no load)—less than 100 mv; Open-Circuit Voltage—190 volts, approximately; Internal Resistance — 23,000 ohms.

Power Supply: 105 to 125 (or 210 to 250) volts, 50 to 60 cycles.

Power Input: 10 watts.

Vacuum Tubes (supplied):
1 — 6H6 type 2 — 6SL7GT type

Accessories Supplied: Connector for use between bridge and oscillator-amplifier; line cord-and-plug assembly; spare pilot lamp; one TYPE 274-M Plug.

Dimensions: Cabinet — $10\frac{1}{2}$ x $2\frac{1}{2}$ x $6\frac{3}{4}$ inches; Panel — 12 x $3\frac{3}{8}$ inches.

Net Weight: 9 pounds.

Type		*Code Word*	
650-P1	Amplifier-Oscillator	BOGUS	$140.00

MULTIPLE PHOTOS WITH THE MICROFLASH

AN ELECTRONIC SEQUENCE TIMER, developed by the Air Technical Service Command at Wright Field, permits multiple exposures on a single film to be made with the Microflash. Developed by Captain C. H. Coles of the ATSC engineering division, the timer utilizes the constant rate of voltage increase across a linearly charged condenser to excite six amplifiers which are set to progressively decreased sensitivity. Each amplifier trips one Microflash lamp.

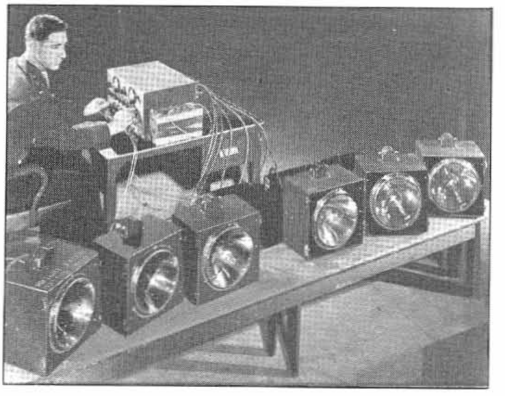


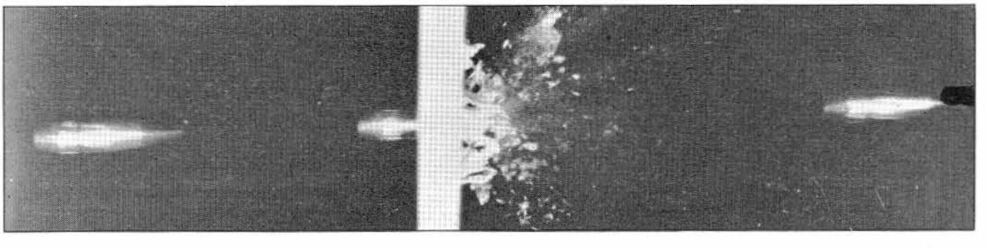
FIGURE 2. The electronic sequence timer and a battery of six Microflash units. Captain C. H. Coles of the Air Technical Service Command, who developed the timer, is shown at the controls.

The interval between flashes is varied by simply turning dials on the sequencer panel, and the total time required for all six pictures can be adjusted between 0.6 second and 0.0003 second.

The photograph of Figure 2, which shows three exposures of a 0.50 caliber machine-gun bullet was timed by the electronic sequencer.

ASTC engineers found many uses for the Microflash during the war, in addition to bullet photography. Among these was the study of rupturing propeller blades. By using X-ray film, an f/2.5 night aerial camera lens, and a special developer, they have taken Microflash pictures successfully at distances of 40 to 50 feet from the lamp.

FIGURE 1. Three-exposure photograph of a 0.50 caliber machine-gun bullet.



MEASURING LATERAL MOTIONS IN A ROTATING SYSTEM WITH THE STROBOLUX

● **A NOVEL APPLICATION** for the TYPE 648-A Strobolux has recently turned up in connection with the measurement of small lateral motions in a rotating shaft.

An investigation of these minute irregularities in the moving parts of their line of tachometers was being carried out by the Barbour Stockwell Company of Cambridge, Massachusetts. These tachometers are of the conventional fly-ball type, having three weights spaced at 120° around the spindle, each connected by one link to a sliding collar near the other end of the spindle. A helical compression spring surrounding the spindle between the fixed and sliding collars tends to separate these collars. Centrifugal force moves the weights outward from the shaft, thus causing the links to swing outward from their stationary position parallel to the shaft. This moves the sliding collar along the shaft toward the fixed collar in opposition to the restoring force furnished by the helical spring. The position of the sliding collar is translated into dial reading by means of a spring-loaded follower, bearing against the face of the sliding collar and connected to the dial mechanism.

In the course of the investigation, a contour comparator of the familiar shadow-projection type was used. This consists of a light source with a suitable lens and mirror combination so arranged that a sharply defined and greatly enlarged shadow of the sample is projected upon a ground glass screen. For measurement purposes, the screen had vertical and horizontal reference lines. Graduated micrometer screws were provided to move the stage in a plane normal to the light beam so any point on the image could be set to the fixed reference lines.

It then occurred to Mr. Frank P. Wilkins, one of the engineers working on the project, that the dynamic conditions might be rather different from the static conditions. These could be readily observed if the light source of the comparator were replaced with a stroboscopic light. The arrangement shown in Figure 1 was set up with the lamp housing of a TYPE 648-A Strobolux replacing the incandescent lamp. The convex glass was removed from the Strobolux lamp housing and a piece of heavy paper substituted. This had a 2½" hole cut in the center to admit light

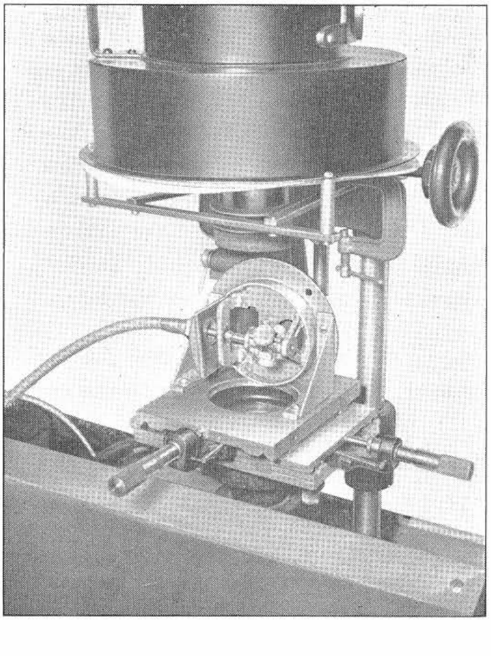


FIGURE 1. View of the tachometer mounted in the contour comparator with Strobolux lamp above.

to the condensing lens, while shielding the ground glass screen from direct light. It was recognized that this arrangement would pass only about 10% of the available light into the system.

The tachometer sample was then rotated at a speed of 2500 rpm, and the Type 631-B Strobotac, which controlled the Strobolux, was set to flash at a speed only a few rpm less than this. The picture shown in Figure 2 illustrates the image which appeared, rotating slowly, on the ground glass screen. A cyclic motion of the spring-loaded follower bearing against the face of the sliding collar was immediately seen. This appeared to be caused by a very slight wobble on the face of the collar, although the condition had not been noticed on a static check when the spring surrounding the shaft was extended. Only a little less obvious was a small erratic lateral motion of the shaft which appeared to take place slowly under the effect of the light from the Strobolux. It was also possible to observe and to measure the very small variations in the radial displacement of the three weights.

Two separate improvements in the tachometers were made possible or greatly facilitated by the use of the Strobolux with this arrangement.

The first was the elimination of a small amount of bouncing of the spring-loaded follower against the face of the rotating sliding collar. Before the investigation this had been manifested only as a very small and erratic motion of the pointer on the dial at medium high speeds.

The second improvement was effected by experimenting with the ball bearings on the spindle and observing the results. This enabled the engineers to fix tolerance standards for these bearings that will eliminate the small lateral motion of the spindle.

The engineers on this project felt that the results were quite satisfactory but decided to reduce the loss in the stroboscopic illumination by constructing a small holder and reflector for the Strobolux lamp which could concentrate a greater part of the available light into the small condensing lens of the contour comparator. This will produce an even sharper and brighter image of the sample permitting greater accuracy in the measurements.

— Kipling Adams

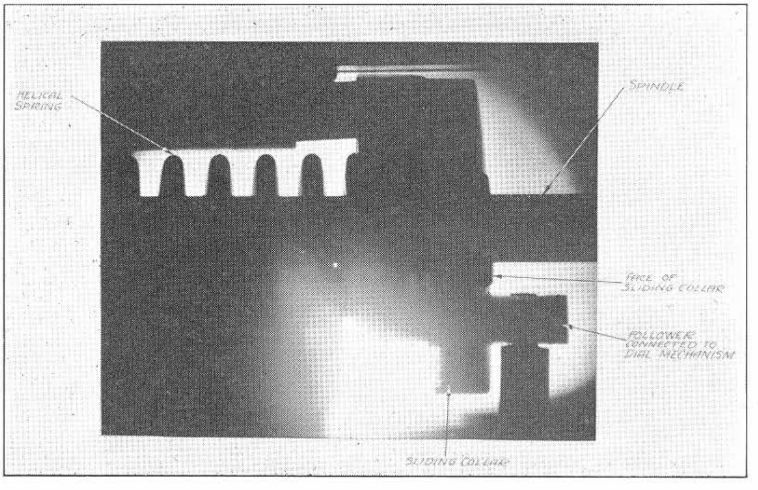
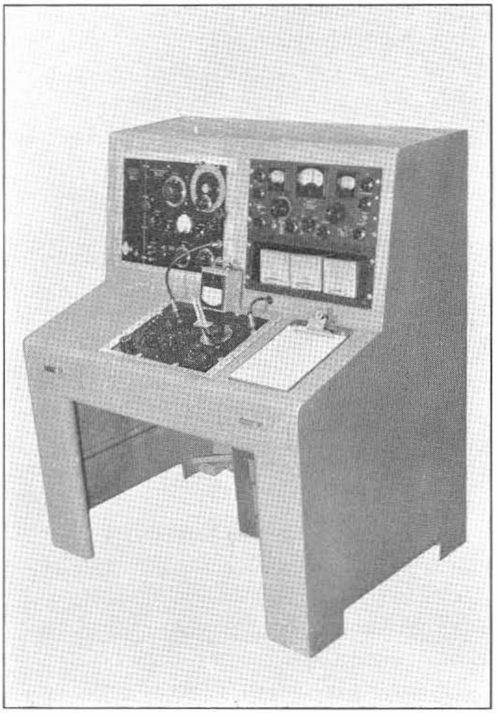


Figure 2. View of the image appearing in the ground glass screen of the comparator.



MISCELLANY

● **FOR RADIO-FREQUENCY IMPEDANCE** measurements Globe Union, Inc., of Milwaukee, manufacturers of Centralab radio products, have designed and built the convenient and attractive console shown here. The sloping desk top houses a TYPE 821-A Twin-T Impedance Measuring Network, while in the vertical panel are the r-f power source, a TYPE 605-B Standard-Signal Generator, and a radio receiver for use as a null detector. Drawers in the front provide a convenient place for storing accessories.

Manufacturers of capacitors and inductors find the Twin-T the most precise instrument available for the measurement of reactance and loss at radio frequencies.

IMPORTANT

Enclosed with this issue of the *Experimenter* is a new net price schedule for all General Radio products. This new schedule becomes effective on and after April 15, 1946.

Since a number of price changes have been made, all are urged to consult the new revised list before ordering.

GENERAL RADIO COMPANY
275 MASSACHUSETTS AVENUE
CAMBRIDGE 39 MASSACHUSETTS
TELEPHONE: TROWBRIDGE 4400

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90 WEST STREET
TEL.—WORTH 2-5837

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TEL.—WABASH 3820

LOS ANGELES 38, CALIFORNIA
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TEL.—HOLLYWOOD 6321

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THE General Radio EXPERIMENTER

VOLUME XX No. 12 MAY, 1946

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

Also
IN THIS ISSUE

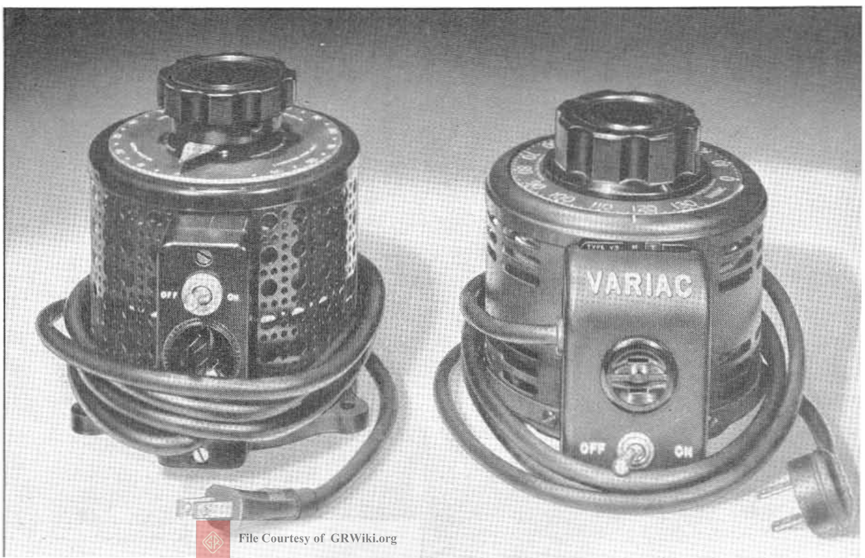
V-5 SERIES VARIACS — NEW, IMPROVED MODELS REPLACE 200-C SERIES

● **FOR MANY YEARS** the 200-C Variac has been our most popular single instrument. We might, perhaps, have chosen to let well enough alone and continued 200-C sales indefinitely. Yet, inevitably, such a course collides with the hard fact that failure to progress is to retrogress.

Mere change, however, is not progress. In supplanting the 200-C with the V-5 we have made every effort to incorporate features that are truly improvements. Careful consideration has been given to each factor, including greater convenience and utility, reliability, improved appearance, and increased value.

For your convenience the V-5 has the new General Radio Unit Brush,

Figure 1. The old and the new. At the right is shown the new Type V-5MT, at the left the old Type 200-CM.



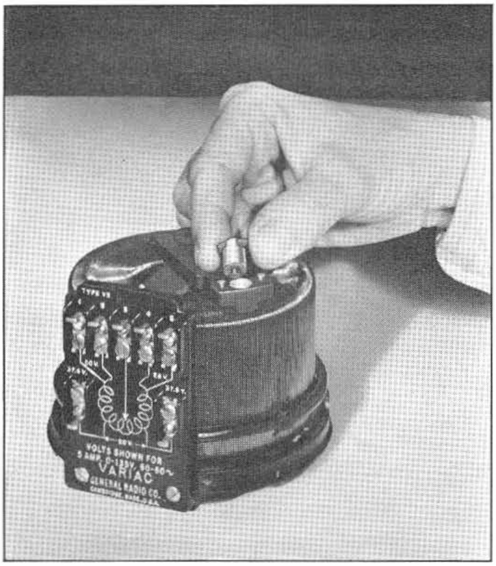


FIGURE 2. "The new General Radio unit brush . . . can be changed without tools."

which can be changed (Figure 2) without tools. The V-5 Brush has a further advantage in that its low sprung weight reduces hammering and arcing under vibration conditions, injurious to both brush and winding, and it is designed to prevent contact of the brush holder with windings and consequent short-circuit damage. Proper brush pressure is assured by the use of an accurate coil spring, which in turn is electrically shunted so that the spring is not subjected to load current. Brush travel is limited by a resilient stop.

No longer is it necessary to take practically the whole moving assembly apart to reverse a shaft in changing from panel to table mounting, or vice versa. A single screw (Figure 3) loosens the shaft for removal or tightens it in assembly. Brush adjustment and the resilient stop setting are unaffected by shaft reversal.

Variacs have to be carried, and knobs are poor and unreliable handles, relying, as they do, on set screws. Figure 4 shows the V-5 cord with molded-on plug which, wound about the Variac and plugged into the Variac outlet, serves as an excellent carrying strap.

Rubber feet, that squeeze back into their sockets for panel mounting, make V-5 Variacs considerate of your bench, table, or desk, and provide sufficient friction to prevent slipping even on smooth surfaces. Rounded contours, devoid of sharp or irregular projections, still further protect against damage to adjacent objects (including hands and feet) in the inevitable bumps and drops that occur in the life of an instrument.

Dials with BIG calibration figures and additional scale points make V-5 Variacs easy to read and to reset. Clockwise rotation to increase output is the rule.

Terminals have both solder and screw connection facilities, are easy to get at, and logically arranged. Barriers reduce the hazard of short-circuits from loose strands. A circuit diagram, integral with the terminal strip, identifies each lead and indicates normal voltages (not turns, as heretofore) between leads as a wiring aid. Extra

FIGURE 3. "A single screw loosens the shaft for removal or tightens it in assembly."

terminals (6 and 7) greatly extend the usefulness of the V-5 Series Variacs when used with supplementary transformers. The terminal strip itself is a high-impact-strength molding, protected by a metal, fiber-lined cover to reduce further the possibility of breakage. If, however, you are ingenious enough to break it, the only tool required for replacement is a screw driver.

Next to a zipper, the V-5 case and terminal cover assembly method is the fastest. Two screws only are required for the assembly of the two pieces to the base. A cover band, carrying integral rivets, cooperates with the screws to form a secure yet rapid assembly. Type numbers distinguishing cased and uncased models are automatically formed by registering tabs on the parts.

A new, heavy-duty switch breaks both sides of the line in mounted V-5 models. Provision is made for polarity indication for use with grounded loads, which, while seldom recommended, are sometimes required.

General Radio Company has never featured appearance for appearance's sake. Yet V-5 Variacs have a distinctly modern look, naturally and functionally derived. Curves instead of angles result from the maintenance of fixed clearances with a minimum of enclosing material. The top "band" is used to position the perforated cover side when it is welded to the top ring. The bottom "band" which forms the other termination of the cylindrical portion of the enclosure carries fastening rivets as previously explained. Both are functional. The flush dial is a perfectly logical method of conserving space and material, as it is eminently suitable to form a portion of the enclosure. The perforations were deliberately chosen to accent the cylindrical nature of the design. Perforations

FIGURE 4. ". . . the V-5 cord with molded-on plug which, wound about the Variac and plugged into the Variac outlet, serves as an excellent carrying strap."

were required in any event; better attractive than not.

The finish selected for V-5 Variacs is an exceptionally durable matte baking lacquer. Its lack of high lights prevents distraction of attention and preserves the unity of the basic design. The use of aluminum for structural parts with attendant corrosion resistance is further enhanced by this wear-resistant and attractive covering.

Last but not least, don't overlook the fact that V-5 Variacs deliver 25 per cent more KVA per pound than equivalent 200-C models. Grain-oriented strip cores permit reductions in both iron and copper for comparable ratings. Aluminum instead of zinc and steel in structural parts furthers this weight reduction.

— GILBERT SMILEY

SPECIFICATIONS

Note: Models are designated by type number. The basic VARIACS, V-5 (for 115-volt input) and V-5H (for 230-volt input), are supplied with terminal strip, but without case, terminal box, switch, convenience outlet, and cord. Models V-5M and V-5HM include the case. Models V-5MT and V-5HMT are complete mounted models with case, terminal box, switch, cord, and outlet.

Dials: Dials are engraved for overvoltage connection (135 or 270 volts maximum). Special dials are available for 115- and 230-volt maximum output. Dial is reversible, one side for table mounting, the other for panel.

Type	*V-5*	*V-5M*	*V-5MT*	*V-5H*	*V-5HM*	*V-5HMT*
Load Rating (KVA)	.862	.862	.862	.575	.575	.575
Input Voltage	115	115	115	230 or 115	230 or 115	230 or 115
Output Voltage (Zero to)	135 115	135 115	135 115	270 230	270 230	270 230
Rated Current (Amperes)	5	5	5	2 1*	2 1*	2 1*
Maximum Current (Amperes)	7.5	7.5	7.5	2.5	2.5	2.5
No-Load Loss — 60 ∽ (Watts)	9	9	9	9	9	9
Driving Torque (Inch — Ounces)	30–50	30–50	30–50	30–50	30–50	30–50
Overall Height for Table Mounting (Inches)	5	5	5	5	5	5
Maximum Panel Thickness (Inches)	3/8	3/8	3/8	3/8	3/8	3/8
Depth behind Panel (Inches)	3 21/32	3 21/32	3 21/32	3 21/32	3 21/32	3 21/32
Diameter of Variac Cylinder (Inches)	4 13/16	4 15/16	4 15/16	4 13/16	4 15/16	4 15/16
Add for Terminals (Inches)	9/16	9/16	1 15/16	9/16	9/16	1 15/16
Net Weight (Pounds)	6 3/4	7	7 5/8	6 1/2	6 3/4	7 3/8
Code Word	COBRA	COPAL	CORAL	CULPA	CUMIN	CUPID
Price	$16.50	$17.50	$20.00	$21.50	$22.50	$25.00

*With 115-volt input applied across half the winding.

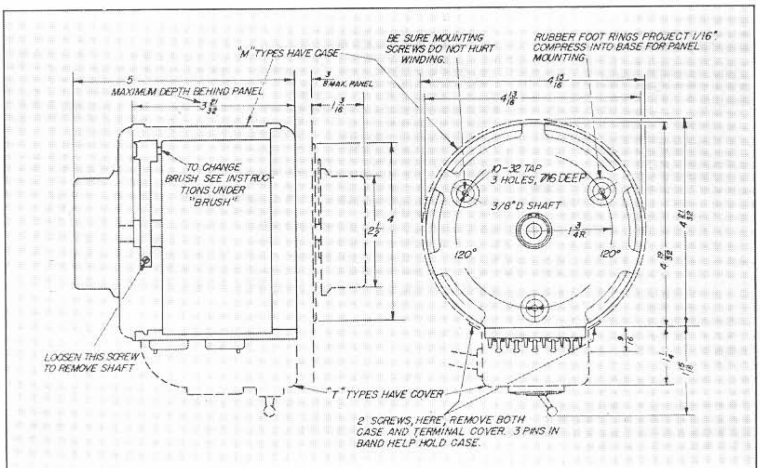


FIGURE 5. Universal dimension drawing for the V-5 series of Variacs. The basic V-5 Variac is shown in full lines. The case (M) and the terminal-box cover (T) are shown dotted. The knob and dial in panel mounting position are also shown dotted.

D AND Q

● **THE POWER LOSSES** in both capacitors and inductors can be expressed in several different ways. In a perfect reactor the current either leads or lags the voltage by exactly 90° as shown in Figure 1. When power losses occur, the phase angle θ becomes less than 90° by an angle δ called the loss angle or phase defect angle, as shown in Figure 2. The power loss W is a direct function of this loss angle:

$$W = EI\cos\theta = EI\sin\delta \qquad (1)$$

where the trigonometric functions are defined as the power factor.

The reactor can be represented either as a parallel or series circuit,[1] as shown in Figures 3 and 4. For these circuits, the power loss is

$$W = \frac{E^2}{R_p} = E^2G = I^2R_s \qquad (2)$$

The ratio of parallel reactance and parallel resistance is called the dissipation factor D, while its reciprocal is called the storage factor Q.

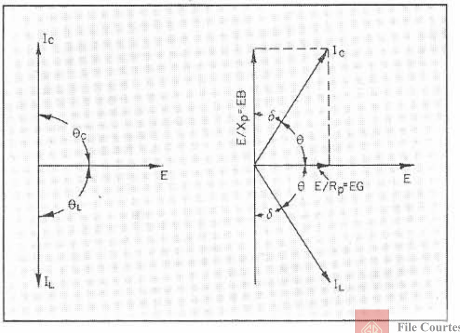
$$D = \frac{1}{Q} = \frac{X_p}{R_p} = \frac{G}{B} = \frac{R_s}{X_s} \qquad (3)$$

Combining Equations (2) and (3)

$$W = \frac{E^2D}{X_p} = \frac{E^2}{QX_p} = I^2DX_s = \frac{I^2X_s}{Q} \qquad (4)$$

[1] Tuttle, W. N., "The Series and Parallel Components of Impedance," *General Radio Experimenter*, XX, 8, January, 1946.

FIGURE 1 (*left*). Relation between current and voltage for a perfect reactor. Phase angle θ is 90°.
FIGURE 2 (*right*). Relation between current and voltage for a reactor having losses. Phase angle θ differs from 90° by loss angle.



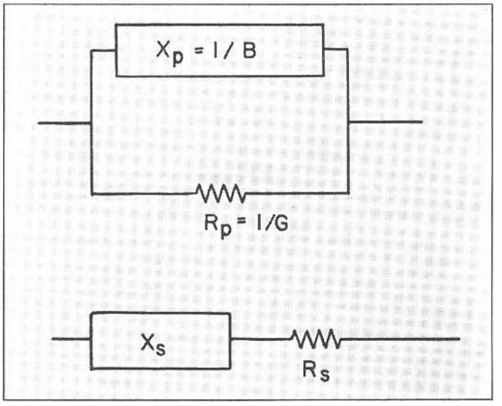


FIGURE 3 (*above*). Parallel components of a reactor.
FIGURE 4 (*below*). Series components of a reactor.

The names of these factors are appropriate, for dissipation factor D is proportional to the power dissipated in the resistive elements, and storage factor Q is proportional to the power stored in the reactive elements. The relation of these factors to the phase and loss angles as shown in Figure 2 is

$$D = \frac{1}{Q} = \cot\theta = \tan\delta \qquad (5)$$

Dissipation factor and power factor differ by less than 1% when their values are less than 0.15.

Dissipation factor and storage factor are used as figures of merit, or quality factors, for capacitors and inductors. It has become customary to use dissipation factor D for capacitors and storage factor Q for inductors. There are, however, many reasons for using dissipation factor exclusively. In the first place, any power loss is a defect, a departure from perfection, and the factor that measures it should vary directly with it and go to zero for zero loss. Dissipation factor meets this requirement and storage factor does not. Values for several types of capacitors are given in Table I.

TABLE I

Capacitors	*D*	*Q*
Paper, Ordinary	0.02	50
Paper, Quality	0.005	200
Mica, Ordinary	0.002	500
Mica, Quality	0.0005	2000
Polystyrene	0.0002	5000

It is sometimes urged that dissipation factor is awkward because it is a decimal, with several ciphers before the significant figure. To remedy this, dissipation factor has been expressed in per cent, as is customary for power factor when used for power transmission. This has led to endless confusion when it becomes necessary to state the accuracy of its measurement also in per cent. For this reason, it is expressed as a ratio in all ASTM specifications and in all General Radio publications. There is only a slight gain in the number of figures required if storage factor Q is used instead, as is shown in Table I, and the ability to show accuracy by the number of significant figures is impaired.

In the second place, whenever there is more than one source of loss in a reactor, the dissipation factors representing such losses add directly to give the total dissipation factor. Multiple storage factors can be added only by taking the reciprocal of the sum of their reciprocals, in other words by first calculating dissipation factor. A good example of multiple dissipation factors is offered by a capacitor with a solid dielectric,[2] such as mica. Over a wide frequency range there are three kinds of losses, as shown in Figure 5. At low frequencies interfacial polarization produces a dissipation factor D_i, which decreases with increasing frequency in such manner that it results in a straight line with a slope less than 45° on a log-log plot of dissipation factor against frequency. Such a law would produce a very low dissipation factor at high frequencies. There is in addition a residual polarization, whose origin is not understood, which produces a minimum dissipation factor D_r, that is constant with frequency. At high frequencies ohmic resistance in the leads to the capacitor becomes an important source of loss, both because the reactance decreases with increasing frequency and because skin effect in the leads increases their resistance as the square root of the frequency. Hence, dissipation factor D_s from this source increases as the 3/2 power of frequency. Total dissipation factor is the sum of the three separate dissipation factors. Obviously, storage factor is useless in such a summation.

Storage factor Q has been used for inductors for a long time in connection with tuned circuits. In a tuned circuit in which a voltage e is introduced either through a series resistance as

[2] Field, R. F., "Frequency Characteristics of Decade Condensers," *General Radio Experimenter*, XVII, 5, October, 1942.

FIGURE 5. Component dissipation factors of a mica capacitor; D_i from interfacial polarization, D_r from residual polarization, and D_s from series resistance.

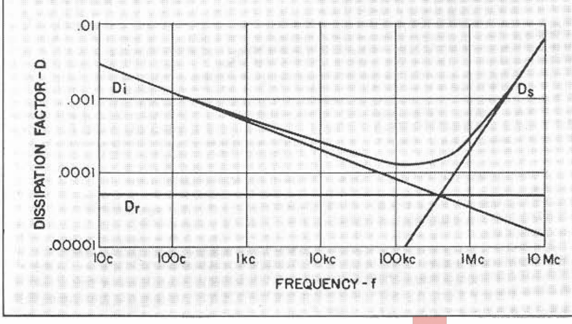
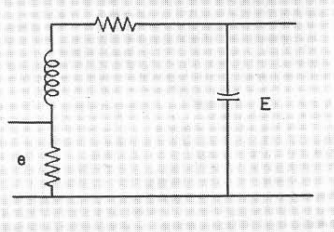


FIGURE 6. Circuit for measuring step-up voltage ratio.



shown in Figure 6, or inductively, the ratio of the voltage developed across the tuning capacitor to the input voltage is the storage factor Q of the circuit.

$$\frac{E}{e} = Q = \frac{\omega L}{R} = \frac{1}{R\omega C} \qquad (6)$$

If an air capacitor or a mica capacitor is used, the losses of the inductor are usually so much greater than those of the capacitor that it has become common practice to consider that this voltage step-up is the storage factor Q of the inductor. But even here there may be multiple losses. With care it is possible to design an air core inductor using insulated stranded wire (litzendraht) having a dissipation factor of 0.002 (Q = 500). A poor mica capacitor could easily have a dissipation factor of 0.001 (Q = 1000). Here is a 50% error if the voltage step-up is used. If both losses are considered, it is much faster to note that the total dissipation factor is 0.003 than to add the reciprocals of 500 and 1000 which, when it is done, amounts to first calculating dissipation factor.

Inductors furnish just as good an example of multiple dissipation factors as capacitors. In an iron-core coil there are three sources of loss[3], as shown in Figure 7. At low frequencies, the dissipation factor D_c produced by the ohmic resistance of the inductor varies inversely with frequency and is therefore a straight line sloping down at 45° on a log-log plot of dissipation factor against frequency. Hysteresis losses in the iron core provide a dissipation factor D_h which is constant with frequency and whose magnitude increases with flux density. Eddy current losses in the iron laminations increase with frequency and produce a dissipation factor D_e which increases directly with frequency. The total dissipation factor is the sum of the three separate dissipation factors. As noted under capacitors, it would only complicate matters to try to use storage factor Q.

Air-core inductors used near their resonant frequencies offer a different set of three losses. Ohmic resistance furnishes a dissipation factor D_c varying inversely with frequency as shown in Figure 8. There are eddy current losses in the copper winding which behave in the same manner as eddy current losses in iron and give a dissipation factor D_e which increases directly with frequency. Finally the distributed capacitance of the inductor has a dissipation factor D_0 of its own and determines, with the inductance of the coil, a natural frequency f_0. The dissipation factor D_f

[3] McElroy, P. K., and Field, R. F., "How Good is an Iron-Cored Coil?", *General Radio Experimenter*, XVI, 10, March, 1942.

FIGURE 7. Component dissipation factors of an iron-cored coil; D_c from ohmic resistance, D_h from hysteresis, and D_e from eddy currents.

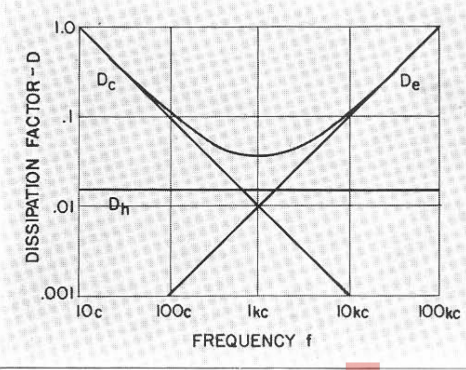
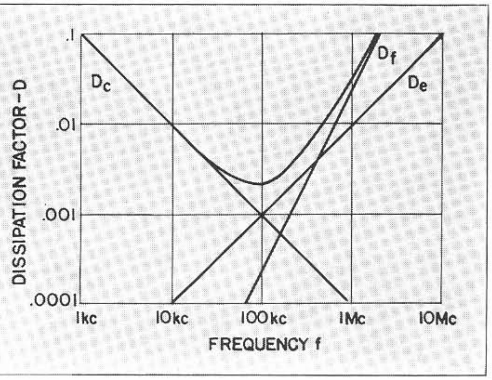


FIGURE 8. Component dissipation factors of an air-cored coil; D_c from ohmic resistance, D_e from eddy currents, and D_f from distributed capacitance.



which this capacitance produces in the coil varies with the square of the frequency, having, of course, the value D_0 at f_0. Again this total dissipation factor is the sum of the three separate dissipation factors.

An even more complicated case is a multiple layer toroid with a low permeability dust core for use at high frequencies. There are five sources of loss, ohmic resistance, eddy currents in both iron and copper, hysteresis losses in the iron, and dielectric losses in the distributed capacitance.

It thus appears that in both the analysis and the synthesis of losses in capacitors and inductors, dissipation factor best expresses their losses. Since both capacitors and inductors are now used over very wide ranges of frequency, it is logical to use for all other calculations the factor in terms of which the losses are best expressed, namely dissipation factor D.

— Robert F. Field

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

OUTPUT SYSTEMS OF SIGNAL GENERATORS

● **INTRODUCTION.** Recently, the question of standardizing the output of different models of signal generators has become one of increasing importance. Measurements made using one type of signal generator frequently do not agree with those made with another type. While this lack of agreement does, on occasion, occur at frequencies in the standard broadcast band, the differences are generally much more serious at higher frequencies. The recent extensive use of signal generators at frequencies above the standard broadcast band has brought about this demand for standardization.

The differences in these measurements are generally caused by the differing output impedances of the generators. When a load (usually the input of a receiver) is connected to the signal generator, the voltage appearing at the generator terminals is not the open circuit voltage but is a value determined by the impedances of the generator and load as well as by the indicated open-circuit voltage. Furthermore, the leads connecting the generator and load also affect the voltage actually applied to the device under test. Consequently, the load voltage can be determined from the indicated open-circuit voltage of the generator only if the generator impedance, the load impedance, and the characteristics of the leads are known.

Zero Source Impedance

One approach to this problem is to make these complicating effects negligible, and to a certain extent this situation is achieved at low frequencies. At standard broadcast frequencies and lower, no difficulty is experienced in making the cable con-

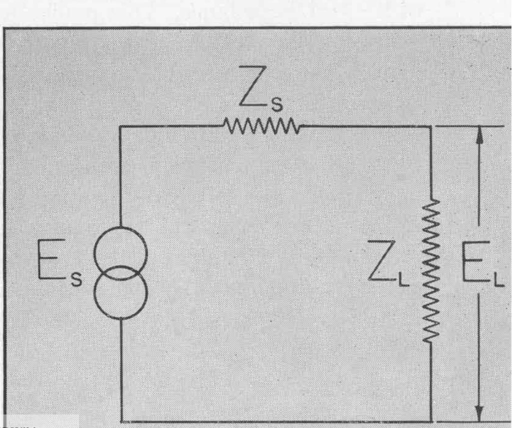


FIGURE 1. A generator of internal voltage E_s and internal impedance Z_s connected to a load Z_L. The load voltage E_L is easily determined by equation (1).

necting the generator and the load so short that there is no appreciable difference between the voltages at the generator end and the load end of the cable. Then, if the voltmeter in the generator is connected to measure the voltage at the generator terminals, this measured voltage is also the voltage applied to the load. If the load impedance is changed, the voltmeter will still read the voltage at the load. Since this arrangement needs no correction of voltage for the effect of load, it can be considered to be a generator with a voltage source of magnitude equal to that indicated on the voltmeter acting in series with a zero source impedance.

Generally, the direct measurement in the signal generator of voltages less than about one-tenth volt is neither convenient nor economically feasible. In order to work down to output voltages of one microvolt and even less, it is customary in low- and medium-frequency generators to insert a low-impedance attenuator between the voltmeter and the output terminals.

Equivalent Source

In order to calculate the effects of the load on the voltage at the terminals of a signal generator using this arrangement of voltage control, it is convenient to use the Thévenin-theorem approach and to represent the generator as a voltage source "behind" — i.e., in series with — an impedance. This equivalent voltage is the open-circuit voltage of the signal generator, and the equivalent impedance is that seen looking into the system with the terminals at which the voltage is measured short-circuited. The signal generator is usually arranged to indicate the value of this equivalent voltage by a combination of the voltmeter reading and the attenuator setting. The equivalent source impedance is generally made as low as is practical, being frequently 10 ohms, or even less. At low and medium frequencies with these low impedances, it is frequently possible to neglect the effect of the load on the output voltage and to assume that the terminal voltage is the open circuit voltage. In any case, it is relatively easy to determine the correction for the load, as indicated in Figure 1, by the formula:

$$E_L = \frac{Z_L}{Z_L + Z_s} E_s$$

$$= \left(1 - \frac{Z_s}{Z_s + Z_L}\right) E_s \qquad (1)$$

where E_L = Load Voltage
E_s = Source Voltage
Z_L = Load Impedance
Z_s = Source Impedance

When the generator source impedance is small compared to the load impedance, only an approximate value for the load impedance need be used for the calculation of load voltage.

Dummy Antenna

A signal generator with low source impedance, however, is not the complete answer even at low frequencies. For example, the adjustment of a receiver input system to give maximum voltage sensitivity with a low-impedance signal generator will lead to the use of a high ratio transformer at the input with correspondingly high apparent voltage sensitivity. However, the actual sensitivity with a normal antenna will be much less than the measured value. There are several ways of avoiding this discrepancy between measured and actual sensitivities. Fundamentally, it is desirable that the effective source impedance of the generator be essentially the same as that of the energy source with which the device under test is to be used.

One procedure for achieving this condition is to add at the generator output terminals a series impedance of the value necessary to produce the desired source impedance. The standard dummy antenna adopted by the Institute of Radio Engineers for the medium and high frequencies is such a series impedance. Its value has been selected to make the source impedance of the generator approximate that of a representative antenna. An impedance network or a section of special attenuating cable is sometimes used when the simple series arrangement is not practical.

The relatively high impedance of the dummy antenna helps to mask the variations in generator output impedance that are caused by residual reactances. It also aids in isolating the generator from the load to the extent that variations in load impedance do not appreciably affect the voltage indicated by the signal-generator voltmeter. This isolation is particularly helpful when the input impedance of the device under test is to be adjusted during the course of the measurement, since the specified source impedance of the generator is based on maintaining the voltmeter reading constant for varying conditions of load (or at least correcting for the variations of the voltmeter reading).

Loop Antennas

Receivers using loop antennas are usually treated in a manner that makes the loop antenna the source impedance or dummy antenna. In some cases the voltage from the signal generator may be inserted in series with the loop. Then it is important that the generator source impedance be small as compared with the loop antenna resistance. The preferred arrangement uses a small loop antenna connected to the signal generator. This generator loop is so placed as to couple to the receiving loop antenna. In this case the generator output source impedance is usually modified by series resistance to yield a desired field intensity relationship at the receiving loop.[1]

High-Frequency Effects

At frequencies of 5 Mc and higher it is often impractical to make the connection between the signal generator and the device under test so short that its effect on the measurement can be neglected. Then the problem of determining the terminal voltage becomes more involved. This determination can be made in a number of different ways, but it is frequently convenient to associate the connecting cable with the signal generator and then evaluate the characteristics at the load end of the cable in terms of the open-circuit voltage and output impedance. When these characteristics are available, the voltage at the terminals of the load can be determined by the simple relationship of Equation 1, where the end of the cable is taken as the generator terminals.

In order to analyze the different typical conditions that occur as a result of the connecting cable, they can be separated into five specific arrangements, as shown in Figure 2. In each case, the connecting cable is assumed to have negligible loss and be a uniform transmission line of characteristic impedance Z_0 which is a pure real, that is, resistive. Furthermore, the connecting fittings are assumed to be a uniform extension of the line. (For accurate measurements a

[1]W. O. Swinyard, "Measurement of Loop-Antenna Receivers," *Proc. I.R.E.*, 29, 7, July, 1941.

cable with clip leads should not be used at frequencies above about 5 megacycles. The coaxial fitting at the end of the cable must be connected directly to the chassis of the device under test.) With a properly arranged set-up, these conditions are generally fulfilled for all practical purposes.

Line Matched At Both Ends

The first arrangement, labeled as I in Figure 2, is readily analyzed. The cable is terminated at both ends in a resistance equal to the characteristic impedance of the line. In the first of the two alternative situations, the cable and generator source impedance have been chosen to be alike. In the second alternative, the source impedance of the generator has been increased by the use of a series resistor to a value equal to the characteristic impedance of the line. The output system of the Type 805-C Signal Generator is of this type when used with the termination unit. The situation of a generator source impedance larger than that of the line is not often encountered, but the impedance can be adjusted with a shunt resistor to give a matched condition, in which case correction must be made for the voltage drop of the combination of resistances.

The termination at the load end of the cable in this first arrangement matches the line, with the result that the voltage along the line is constant in magnitude but varying in phase. The impedance seen looking into the cable at the generator end is the characteristic impedance of the cable, and the voltage appearing at this end of the cable is one-half the generator voltage. This voltage is then transferred to the other end of the cable without change in magnitude. Therefore, the equivalent source voltage is one-half the generator voltage, or $E_s = E_G/2$. The equivalent source impedance is readily seen to be a parallel combination of a properly terminated line and a resistor equal in value to the characteristic impedance. The effective source impedance is equal to one-half the characteristic impedance, or $Z_s = Z_0/2$. This arrangement gives a source voltage and a source impedance that is independent of frequency, certainly a desirable condition. In actual practice, however, residual reactances in the generator and the termination introduce a dependence on frequency. This behavior is illustrated by the frequency characteristic of the source impedance of the Type 805-C Signal Generator, shown in Figure 3. In this generator the generator impedance is 75 ohms, and the terminating resistor is 75 ohms. The corresponding source impedance is then 37.5 ohms at low

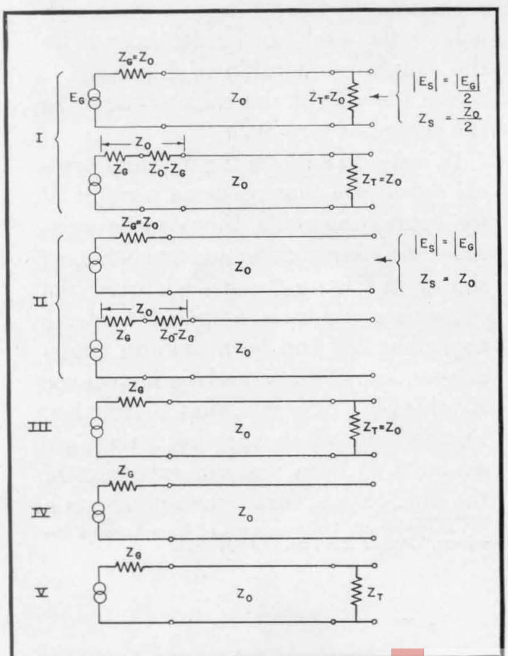


Figure 2. Schematic diagrams of four types of generator output systems. Case V is the general case.

frequencies. The residual inductance in the attenuator and shunt capacitance in the termination unit switch cause the deviations at high frequencies from the low-frequency value. For most applications the deviations shown are sufficiently small that no corrections for them need be made.

While there is no standing wave on the connecting cable with no load on the output for arrangement I, any load will alter the condition of proper termination, and a standing wave will exist on the line. This standing wave, however, does not upset the validity of the Thévenin-theorem approach (the use of equivalent source voltage and source impedance).

Line Matched at Generator End

The second arrangement, designated in Figure 2 as II, is probably the most generally useful system. The generator impedance is made equal to the characteristic impedance of the line with the result that the source at the end of the cable is the same as the generator except for a shift in phase of the voltage, that is, $|E_s| = |E_G|$, and $Z_s = Z_0$. That the source impedance is Z_0 can be readily seen from the fact that the line is terminated at the generator end in its characteristic impedance. To show that the magnitude of the open-circuit voltage at the end of the cable is equal to $|E_G|$, compare the power available at this point to the power available at the generator terminals. If there are no losses in the line, the two available powers must be the same. It follows that the open-circuit voltage at either point must have the same magnitude, since the output impedance is the same, and the same power is available. This voltage can, of course, also be derived from

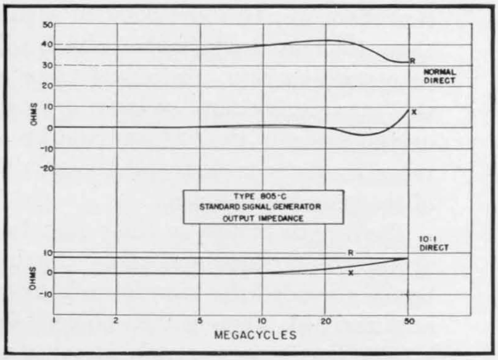


FIGURE 3. Output impedance of TYPE 805-C Standard-Signal Generator as a function of frequency.

the standard transmission-line equations.

This arrangement has a source impedance and output voltage independent of frequency, and it has the advantage over the previous arrangement that, for a given power supplied at the generator terminals, appreciably more power is available at the output terminals than in the previous case. A disadvantage is that it is somewhat more difficult to avoid a noticeable frequency dependence of the equivalent source at high frequencies.

It is interesting, although not essential to the foregoing discussion, to notice that unless the load is matched to the line a marked standing wave will exist on the line. The open-circuit condition is particularly interesting. Then the voltage at the end of the line remains in magnitude always equal to the generator voltage E_G, but the voltage at the input to the line varies from that value to practically zero as the frequency increases from very low frequencies. The zero value of voltage occurs for frequencies at which the transmission line is an odd multiple of a quarter wavelength. At frequencies higher than that for which the line is a quarter wave-length long, in the open circuit condition there will be some point or points along the line at which the voltage will be practically zero. The actual value of this "practically-zero" voltage depends on the losses in the line and will be of the order of one-tenth to one-thousandth of the generator voltage.

The length of line in these first two arrangements does not affect the behavior provided the impedance is properly matched. There is then no need for a cable that is permanently attached to the signal generator if the connectors and cables used form a uniform transmission line of the correct value of characteristic impedance. At ultra-high frequencies and higher, uniformity is particularly important. At these frequencies a succession of even small discontinuities can cause a very serious degradation of the characteristics of a transmission system, and, therefore, connectors and cables specially designed for this frequency region should be used.

Line Matched at Load End

Arrangement III is used in some signal generators. The transmission line is terminated in its characteristic impedance when there is no load on the line, but its characteristic impedance is not equal to the generator impedance. The no-load terminal voltage is given by the simple formula:

$$|E_s| = \left|\frac{Z_0}{Z_0 + Z_G}\right| \; |E_G|$$

If the generator impedance is a pure resistance, the output voltage is independent of frequency. The frequency characteristic of the source impedance is more complicated, but can be readily derived from the transmission line equations. It can be expressed as

$$Z_s = \frac{Z_0\left(Z_G + jZ_0 \tan\frac{2\pi l}{\lambda}\right)}{(Z_0 + Z_G)\left(1 + j\tan\frac{2\pi l}{\lambda}\right)}$$

where l = length of transmission line
λ = wave-length of signal in transmission medium

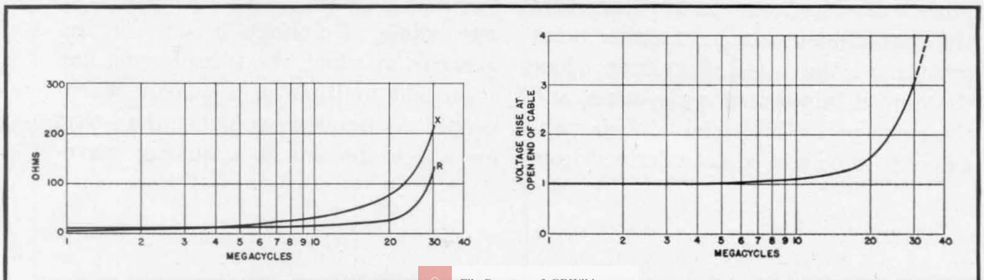
The source impedance varies with frequency as can be seen from this equation. If the generator impedance is a pure resistance, the source impedance varies in magnitude as a function of frequency between the extremes $Z_G Z_0/(Z_G + Z_0)$ (the parallel combination of Z_0 and Z_G) when the cable is an integral multiple of a half wave-length long and $Z_0^2/(Z_G + Z_0)$ when the cable is an odd multiple of a quarter wave-length long. At these extremes the source impedance is a pure resistance, while at other frequencies there is a reactive component. The maximum value of reactance occurs when the cable is an odd multiple of an eighth wave-length long. For this condition the source impedance is

$$Z_s = \frac{Z_0}{2} \pm j\,\frac{Z_0(Z_0 - Z_G)}{2(Z_0 + Z_G)}:$$

Line Unmatched

The next arrangement, IV, yields a source whose impedance and open-circuit voltage both vary with fre-

FIGURE 4. Output impedance and voltage of TYPE 605-B Standard-Signal Generator and TYPE 774-R Cable as a function of frequency.



quency. The transmission-line equations show that the equivalent source for this line terminated only at the generator end in a resistance different from the characteristic impedance is

$$Z_s = \frac{Z_0\left(Z_G + jZ_0 \tan \dfrac{2\pi l}{\lambda}\right)}{Z_0 + jZ_G \tan \dfrac{2\pi l}{\lambda}}$$

$$E_s = \frac{Z_0 E_G}{Z_0\left(\cos \dfrac{2\pi l}{\lambda}\right) + jZ_G\left(\sin \dfrac{2\pi l}{\lambda}\right)}$$

If the generator impedance is real, the source varies between the extremes

$$Z_s = Z_G, \,|\, E_s \,|\, = \,|\, E_G \,|$$

when the transmission line is a multiple of a half wave-length long, and

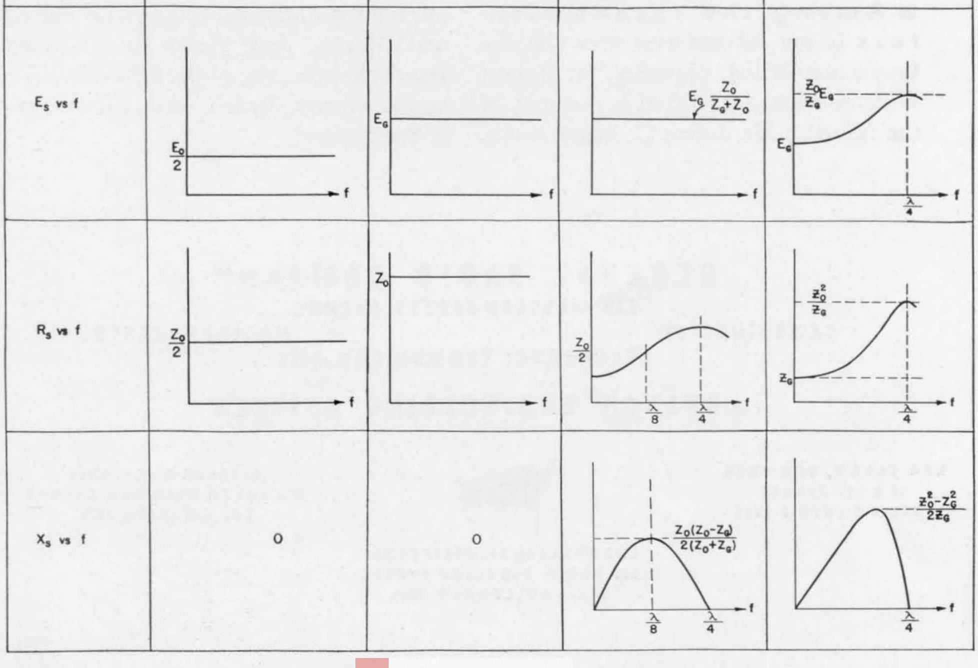
$$Z_s = \frac{Z_0^2}{Z_G}, \,|\, E_s \,|\, = \frac{Z_0}{Z_G} \,|\, E_G \,|$$

when the transmission line is an odd multiple of a quarter wave-length long. At these extremes the source impedance is a pure resistance, while at other frequencies there is a reactance component.

This arrangement is the one that occurs when the Type 605-B Signal Generator is used with a Type 774-R coaxial cable. The Type 605-B has a 10-ohm attenuator, and the cable impedance is 72 ohms. The measured results for the equivalent source of this combination are shown in Figure 4. The impedance was measured by the Type 916-A Radio-Frequency Bridge and the voltage was measured by comparing receiver response with and without the cable. As a

Figure 5. Summary of impedance and voltage characteristics of the four types of terminations shown in Figure 2. The relative phase relationships of the voltages are not included, and it is assumed that Z_G is a pure resistance.

	I	II	III	IV
E_s	$\frac{E_G}{2}$	E_G	$E_G \frac{Z_0}{Z_G + Z_0}$	$\frac{Z_0 E_G}{Z_0 \cos \frac{2\pi l}{\lambda} + jZ_G \sin \frac{2\pi l}{\lambda}}$
Z_s	$\frac{Z_0}{2}$	Z_0	$\frac{Z_0(Z_G + jZ_0 \tan \frac{2\pi l}{\lambda})}{(Z_0 + Z_G)(1 + j \tan \frac{2\pi l}{\lambda})}$	$\frac{Z_0(Z_G + jZ_0 \tan \frac{2\pi l}{\lambda})}{Z_0 + jZ_G \tan \frac{2\pi l}{\lambda}}$



result of departures from the idealized conditions of termination, the equivalent source impedance cannot in general be easily calculated. The actual value of source impedance can be most readily determined by direct measurement with a bridge, for example, the TYPE 916-A Radio-Frequency Bridge.

General Case

The fifth arrangement, labeled in Figure 2 as V, is the general case with terminations at both ends. For this case the equivalent source is

$$E_s = E_G \frac{Z_0 Z_T}{\left(Z_0(Z_G + Z_T)\cos\frac{2\pi l}{\lambda} + j(Z_G Z_T + Z_0^2)\sin\frac{2\pi l}{\lambda}\right)}$$

and

$$\frac{1}{Z_s} = \frac{1}{Z_0}\,\frac{Z_0 + jZ_G\tan\frac{2\pi l}{\lambda}}{Z_G + jZ_0\tan\frac{2\pi l}{\lambda}} + \frac{1}{Z_T}$$

This case is included only to complete the record, since the previous ones are special cases of this general one.

Summary

In order to determine the voltage applied to a device from a signal generator it is necessary to know the load impedance, the characteristics of the connecting cable, and the generator voltage and source impedance. The second and third can be combined to give an equivalent source for the combination. Certain specific characteristics of generator and cable are to be preferred for this equivalent source, notably that in which the generator impedance is equal to the characteristic impedance of the cable. Measurements made with a radio-frequency bridge are the most reliable method of obtaining the necessary impedance values.

— ARNOLD PETERSON

MISCELLANY

● **Among the recent visitors** to our laboratories were the distinguished British physicist, Sir Robert Watson-Watt; Dr. Alfred L. Loomis, of the NDRC; Mr. Johan C. Lagercrantz, our representative in Stockholm, Sweden; and Messrs. Jurg Keller and Jacques Baerlocher-Sarasin, of the firm of Seyffer and Company, Zurich, who represent us in Switzerland.

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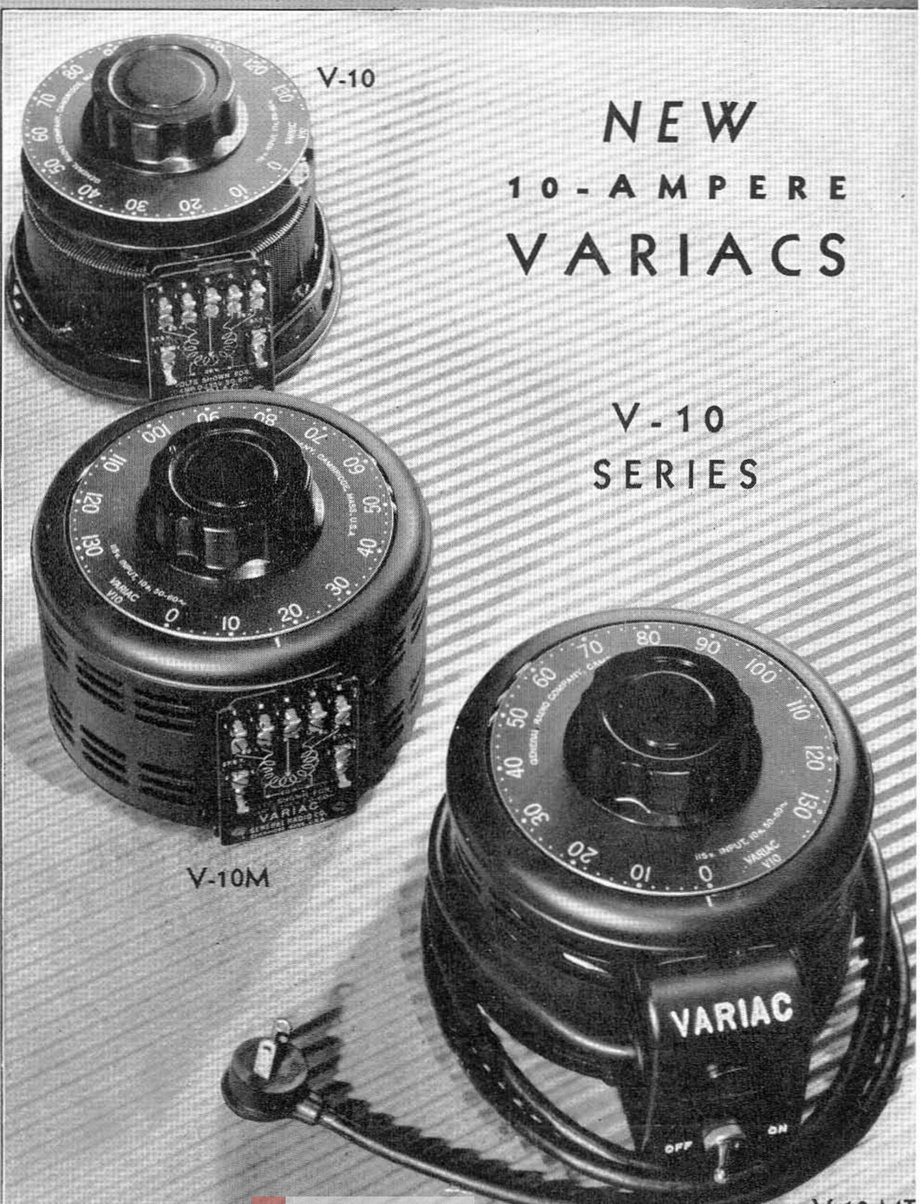
ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS



V-10 SERIES VARIACS—NEW, STANDARD MODELS INTERMEDIATE BETWEEN 200-C AND 100 SERIES

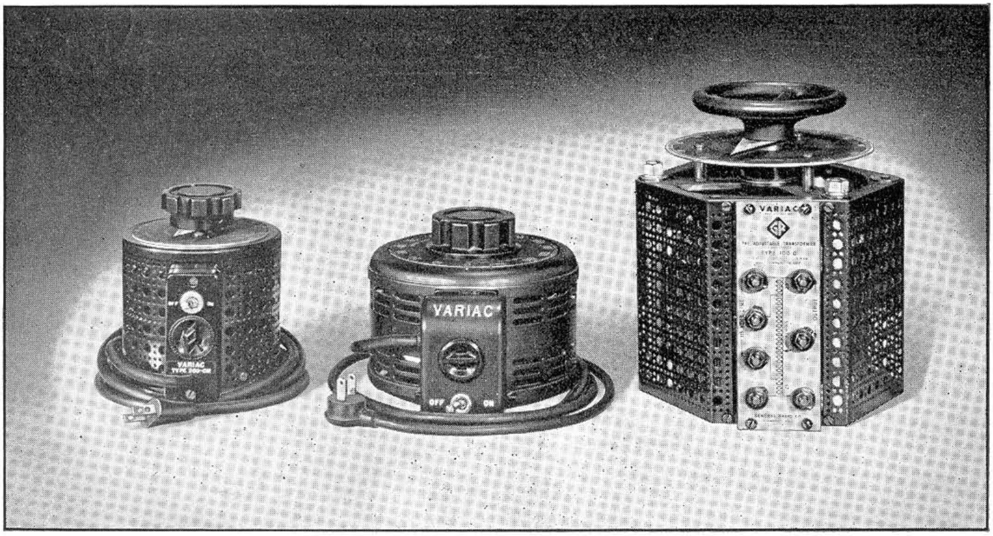


FIGURE 1. Intermediate between the old TYPE 200-CM (left) and the TYPE 100-A (right), the new TYPE V-10 Variac is rated at 10 amperes for 115-volt input.

● **THERE HAS BEEN** an awkward gap between TYPE 200-C and TYPE 100 Variacs ever since the latter's introduction. Loads but slightly exceeding 200-C capacities called for the much larger and more costly TYPE 100 size.

V-10 Series Variacs neatly plug this gap, with ratings just double those of the new V-5 or old 200-C Variacs. The 115-volt V-10's are rated at 10 amperes, with a 15-ampere maximum, coinciding with the capacity of commonly used outlets, plugs, cords, and No. 14 wire circuits.

As contrasted with older models, V-10MT (115-volt mounted model) delivers 112 per cent *more* KVA per pound than 100-Q and 59 per cent *more* than 200-CM. This startling gain in output to weight ratio stems from three sources, recently derived design formulae for the most favorable distribution of copper and iron, the use of a low-loss scroll core, and light-weight structural parts.

The article on V-5 Series Variacs[1] dealt mainly with "user" features. V-10 Series Variacs have these same improvements, unit brush, simplified shaft reversal, better and extended terminal facilities, modern appearance, and so forth.

The present article will be chiefly devoted to certain highlights of V-10 manufacture in the hope that an exposition of these unseen processes may not only be of interest but may further an appreciation of built-in Variac quality.

V-10's, like all Variacs, start with an "iron" core. Actually cold-finished silicon steel strip with a guaranteed maxi-

[1]Gilbert Smiley, "V-5 Series Variacs," *EXPERIMENTER*, XX, 12, May 1946.

ALSO IN THIS ISSUE

mum core loss[2] is used. The strip is wound on precision mandrels exactly as ribbon would be wound. Side guides prevent wandering; limit switches control outside diameter. Ends are cut and "tack" welded to prevent unwinding, and cores are removed from mandrels.

These cores, however, are far from finished. The strains and stresses of winding have increased losses to some ten times the guaranteed maximum. Like balls of dough in which lie incipient loaves of bread, these coils of steel have, dormant within their fibers, the excellent magnetic properties inherent in the strip material. To make bread, bake dough. To make cores, bake coils of strip for three hours at 1450° F in a protective, non-oxidizing atmosphere, followed by a slow cool. Each core is now eight pounds of top magnetic performance.

A finished core is next sandwiched between top and bottom grooved phenolic winding plates which, with inner and outer wrappers, mutually insulate core and winding. The grooves, cooperating with the winding machine, accurately position each turn, and make possible a precision double-banked winding on the inner edge.

The gaping jaws of the V-10 winding machine receive the sandwich (Fig. 2), which is firmly gripped by the rotating vise. The jaws close; the magazine ring is locked to the power-driven winding ring and filled with wire as required to wind a V-10 (Fig. 3). After the magazine ring is freed from the winding ring, the wire is threaded over the take-off guide, through the winding guide, and anchored to the rotating vise, which is positioned for the first turn and latched to the gear train that advances it one wire per rev-

[2]Core loss is guaranteed not to exceed a specified maximum.

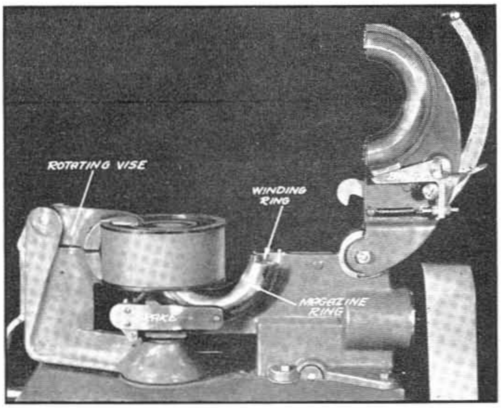
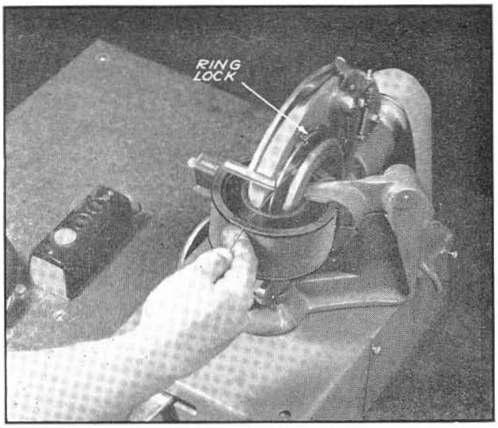
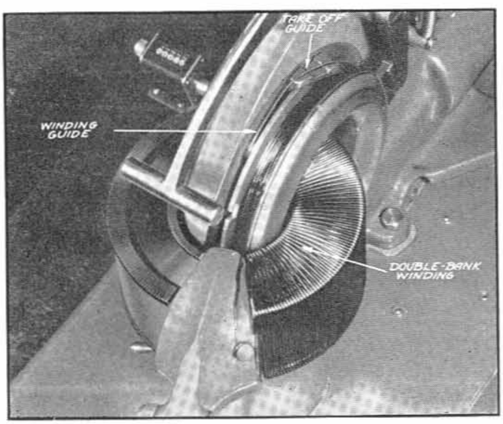


FIGURE 2. Core and winding plate assembly are inserted with winding head assembly opened. Note particularly the concentricity of the winding ring about the core section.



(*Above*) FIGURE 3. Magazine ring being loaded with wire prior to actual winding.
(*Below*) FIGURE 4. A partially wound V-10 coil.



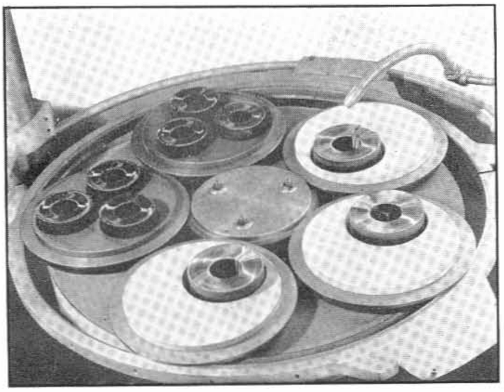
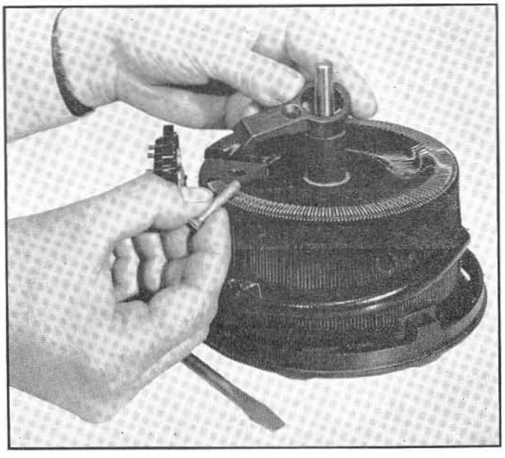
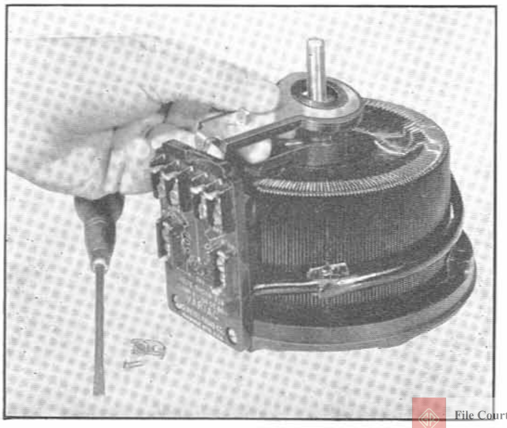


FIGURE 5. Variac coils are here shown in place in the oscillating work holders of the "Hyprolap".



(*Above*) FIGURE 6. The coil is firmly held to the base by the coil clamp here shown during installation.

(*Below*) FIGURE 7. Installing the brush take-off sub-assembly which carries brush current from radiator to terminal, supports the upper end of the terminal strip and serves as a resilient brush stop.

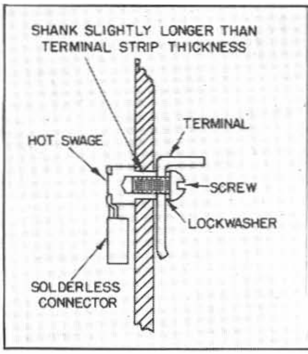


olution of the winding ring. The drive is reversed to winding position, brake tension is applied to the magazine ring, and winding starts, the winding ring pulling wire from the magazine ring as it wraps turn after turn around the core, which moves smoothly into position for each wire by the action of the rotating vise. Figure 4 shows a partially completed winding. Note particularly the uniform banking formed on the inside of the coil.

The V-10 winding machine represents (to the best of our knowledge) a radically new approach to the toroidal winding problem. Unlike earlier machines, the winding ring is concentric with the core center, resulting in a more uniform rate of wire removal from the magazine ring and permitting a five-fold increase in winding speed by eliminating excessive acceleration and deceleration of the magazine and its load. A further advantage is the close proximity of the winding guide to the winding, yielding more accurate wire positioning. This winding machine is a product of our own tool room, developed and designed by General Radio.

After the operations of anchoring end turns, coining to leave the brush track raised, and attachment of taps, the brush track is "Hyprolapped" to a plane, highly polished surface (Fig. 5). Coil and base then meet, insulation is installed, and the coil clamped down (Fig. 6) to avoid passing bolts through the core structure.

The sub-assembly of brush take-off lead and insulator, next to be added (Fig. 7), has three separate functions. It replaces a pigtail, serves as a brush stop, and supports the top of the terminal strip. Figure 8 shows the metal-to-metal pressure construction of V-10 terminals. No dependence is placed upon bakelite under pressure.



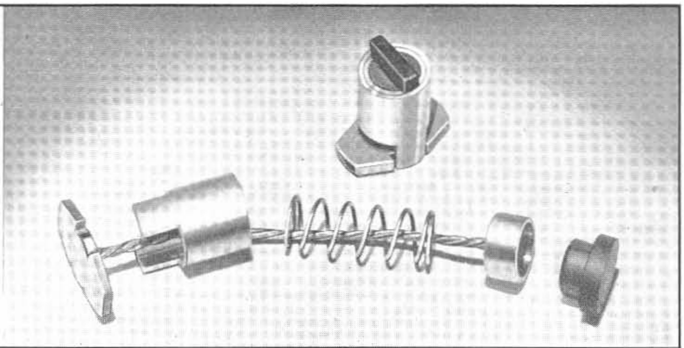


Figure 8. Sectional drawing of terminal assembly.
Figure 9. Assembled and exploded views of the V-10 Brush.

The addition of the radiator, insulation, and shaft parts completes assembly of uncased models except for brushes. Figure 9 gives assembled and exploded views of the V-10 unit brush. Extended research on carbon-to-metal contacts have convinced us that pressure, properly applied, yields the lowest resistance. Solder or electro-deposited metals do not "wet" carbon, they simply cling to a relatively few surface irregularities. Such contacts improve with pressure but never equal direct pressure contacts between carbon and brush holder. So we press the carbon into the brush holder, and the holder into the brush shell. Then we spot weld a flexible copper shunt (ever try a copper to brass spot weld?) between the movable brush parts to prevent load current from traversing the spring and spoiling its tension. The diamond-shaped retainer is snapped into place, and the unit brush is assembled.

The V-10 case parts are spot welded notched for terminal strip registration, finished, and then assembled with cover band, terminal cover, knob, and dial to the V-10 to make a V-10MT cased model.

The scope of this article is too limited for a more complete exposition of the tools and tricks, gadgets and "know-how" that go into Variac production and that make the new Variacs more useful, more reliable, more efficient than ever before. We hope we have been able to show you something of the care and precision that make today's Variacs better.

—Gilbert Smiley

SPECIFICATIONS

Note: Models are designated by type number. The basic VARIACS, V-10 (for 115-volt input) and V-10H (for 230-volt input), are supplied with terminal strip, but without case, terminal box, switch, convenience outlet, and cord. Models V-10M and V-10HM include the case. Models V-10MT and V-10HMT are complete mounted models with case, terminal box, switch, cord, and outlet.

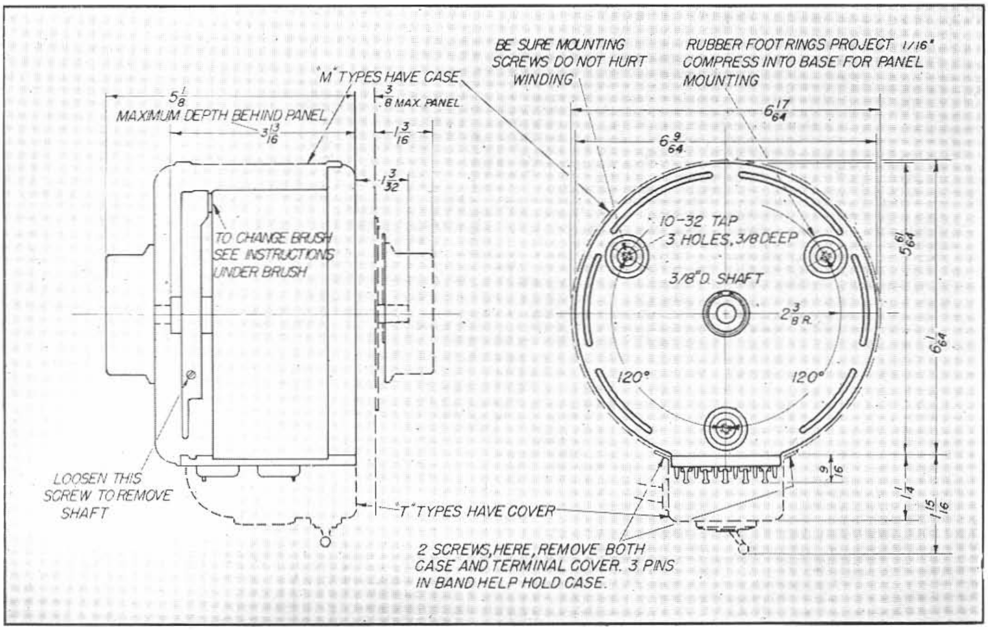
Dials: Dials are engraved for overvoltage connection (135 or 270 volts maximum). Special dials are available for 115- and 230-volt maximum output. Dial is reversible, one side for table mounting, the other for panel.

Type	*V-10*	*V-10M*	*V-10MT*	*V-10H*	*V-10HM*	*V-10HMT*
Load Rating (KVA)	1.725	1.725	1.725	1.15	1.15	1.15
Input Voltage	115	115	115	230 or 115	230 or 115	230 or 115
Output Voltage (Zero to)	135 115	135 115	135 115	270 230	270 230	270 230
Rated Current (Amperes)	10	10	10	4 2*	4 2*	4 2*
Maximum Current (Amperes)	15	15	15	5	5	5
No-Load Loss — 60 ∽ (Watts)	17	17	17	17	17	17
Overall Height for Table Mounting (Inches)	5⅛	5⅛	5⅛	5⅛	5⅛	5⅛
Maximum Panel Thickness (Inches)	⅜	⅜	⅜	⅜	⅜	⅜
Depth behind Panel (Inches)	3 13/16	4 5/32	4 5/32	3 13/16	4 5/32	4 5/32
Diameter of Variac Cylinder (Inches)	6 9/64	6 17/64	6 17/64	6 9/64	6 17/64	6 17/64
Add for Terminals (Inches)	⅜	5/16	1 11/16	⅜	5/16	1 11/16
Net Weight (Pounds)	11¼	11⅝	12⅛	10⅝	11	11½
Code Word	HAZEL	HEAVY	HELOT	HINNY	HOARY	HOBBY
Price	$27.50	$29.00	$31.50	$31.50	$33.00	$35.50

*With 115-Volt input applied across half the winding. Load rating is reduced to one-half the value shown.

DELIVERIES TO START IN NOVEMBER

FIGURE 10. Universal dimension sketch for V-10 series Variacs.



AIRLINE RADIO TESTING WITH GR INSTRUMENTS

• **THE FINAL STEP** in the testing of radio direction finders used in American Airlines Flagships is checking them against General Radio Signal Generators.

American, which flies the greatest fleet of commercial aircraft in the world, has two of these signal generators in constant use in their Radio Overhaul shop at La Guardia Field, New York. In the course of a regular 90-day radio overhaul, the direction finders are put on the revolving test stand whose aerial is connected with a Type 605-B Standard Signal Generator. When the overhaul is finished, and immediately before the unit is sent down to the hangars for installation in one of the giant Flagships, it is sent to the "screen room" to be checked out against a Type 805-B Generator. American owns 200 radio direction finders, which are completely overhauled after every 90 days of service.

American Airlines has found these generators to be among the best obtainable. For American's purposes a signal generator must transmit a signal that is not only measurable but extremely accurate. The generators in use in their

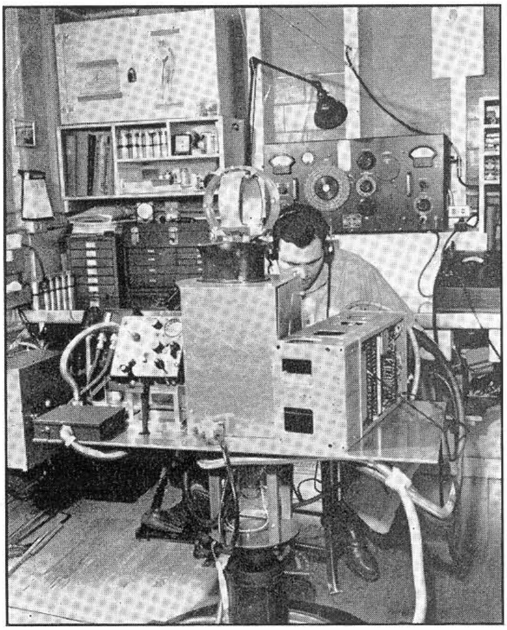


Figure 1. In the screen room of American Airlines radio-overhaul shop, automatic direction finders are mounted on a revolving test stand and checked against a General Radio Type 805-B Standard-Signal Generator. Here lead mechanic Gerard Miller tests the set with a Type 726-A Vacuum-Tube Voltmeter.

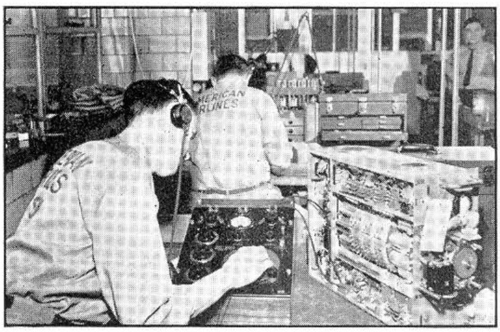


Figure 2. The Type 631-B Strobotac is used to check vibrator reeds by James Hargreaves, lead mechanic in radio-overhaul shop.

radio shops cover all frequencies in the high frequency band.

The Signal Generators are but two of the pieces of General Radio Co. equipment in use by this great airline. Also in constant use in their shops at La Guardia are the TYPE 650-A Impedance Bridge, the TYPE 631-B Strobotac, the TYPE 726-A Vacuum Tube Voltmeter, and the TYPE 813-A Audio Oscillator.

The Voltmeter and the Bridge have come to be highly important pieces of equipment in the routine operation of the shop. While the Impedance Bridge is actually a laboratory instrument, American radio engineers have found it to have an important daily use in the testing of various parts. With it the capacity of condensers, resistance of resistors, and inductance of coils can be immediately ascertained.

A stroboscope is not only a great time and trouble saver, but has come to be a true necessity in the modern radio shop. American Airlines has in its Radio Overhaul a General Radio Strobotac, TYPE 631-B. This piece of equipment is small enough and light enough to be handily moved to wherever it might be needed. It is in constant use by American radio repairmen, who bring it into play in analyzing anything that moves. With it vibrator reeds are checked for proper contact, armatures are searched for correct brush contact, and rotating gyroscopes are inspected.

The Strobotac is frequently used in the American shops to check items which do not in themselves move, but which are affected by movement. This is done on a "shaker table". The table is mounted on shock mounts, and a high speed air motor with an attached eccentric weight is connected to the table itself. An item which might be affected by vibration is put on the table. Then the table is put into motion and the item in question scanned with the Strobotac, which will show up any cracks or bad connections caused by the stress of the motion.

Possibly the best record set for American Airlines by a General Radio product is that of the TYPE 813-A Audio Oscillator. This Oscillator, which transmits a constant 1000 cycle tone, was chosen by American for use in their range stations in Mexico. There are five such stations in the company's Mexican leg. These stations were put into operation on September 5, 1942. On March 11, 1946, the first one of the Oscillators to need repair was received in the American Airlines shops in New York. This makes a total of 41 months of constant service by five General Radio Audio Oscillators before one of them needed adjustment.

Text and photographs for this article were furnished through the courtesy of American Airlines, Inc.

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

Also

IN THIS ISSUE

A NEW VACUUM-TUBE VOLTMETER

● **THE TYPE** of peak-reading vacuum-tube voltmeter now considered standard in the communications industry was first introduced by the General Radio Company some ten years ago when the Type 726-A Vacuum-Tube Voltmeter was announced.[1] The novel design and excellent performance of this voltmeter has maintained its popularity over the intervening years considerably beyond the time when advances in the art would normally necessitate a new design. The development, during this period, of new tubes, circuits, and construction techniques has, however, progressed continuously, and these now make it possible to design a completely new instrument having even better performance.

In the design of the new Type 1800-A Vacuum-Tube Voltmeter, which replaces the Type 726-A, effort has been directed to correcting the minor disadvantages that have been found in the older voltmeter as well as to incorporating desirable new features that make the instrument more flexible and convenient.

[1] W. N. Tuttle, "Type 726-A Vacuum-Tube Voltmeter," General Radio *Experimenter*, Vol. XI, No. 12, May 1937, pp. 1–6.

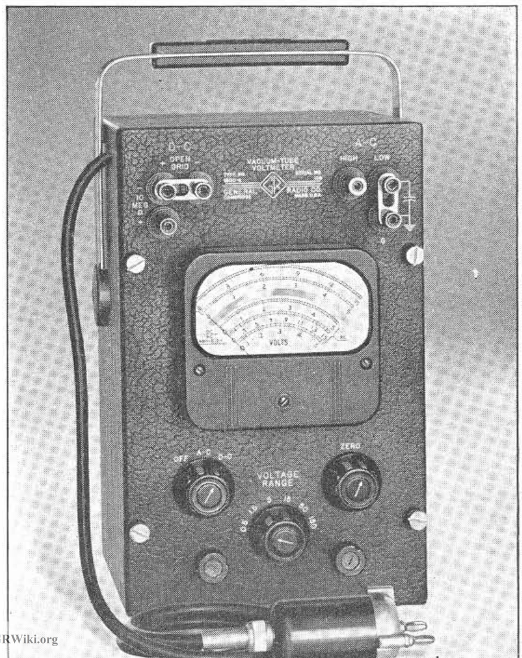


Figure 1. Panel view of the Type 1800-A Vacuum-Tube Voltmeter.

Some of these disadvantages were:

1. The shielding of the probe of the Type 726-A was not complete, and trouble with pickup from strong fields at frequencies over about 50 Mc was sometimes encountered.

2. Individual zero adjustments for each range were used in the Type 726-A, and these sometimes drifted from their correct setting, so that readjustment of the zero when switching from range to range was necessary.

3. The voltage-regulating transformer in the Type 726-A was explicitly designed for one line frequency. In addition, the transformer radiated an appreciable 60-cycle field.

4. The Type 726-A probe was relatively bulky and difficult to get into confined spaces.

5. The Type 726-A meter was somewhat difficult to read at a distance because of the knife-edge pointer and was confusing because of the arrangement of the scales. Reflections from the glass were often bothersome, and the light from the pilot light distracting.

6. The cabinet of the Type 726-A, while convenient for ordinary bench use, was rather large and could not readily be used in other positions than the normal one in which the panel was at a 15° angle to the horizontal.

These disadvantages, with others of a more minor nature, have all been eliminated in the Type 1800-A, and many new features have been added. These include:

1. The natural frequency of the probe has been increased by nearly 3:1, and the gain in high-frequency performance because of the complete shielding is even greater.

2. The input capacitance has been reduced by a factor of 2, and the parallel resistance component increased fourfold at low frequencies.

3. An additional 0-0.5 volt scale has been added that extends the sensitivity by a factor of 3.

4. A complete set of d-c scales has been added, covering the same ranges as the a-c scales.

5. A set of terminal parts has been provided that makes it possible to assemble, on the end of the probe, connectors that will fit all standard General Radio Type 138, 274, and 774 terminals.

6. The new instrument is smaller, lighter, and easier to use.

7. It can be operated with the panel horizontal, vertical, or inclined.

METER

Figure 2 shows the arrangement of the five scales on the face of the meter. The two outer scales, 0 to 15 and 0 to 5, are linear. On these two scales are read all d-c voltages and all a-c voltages above 5 volts. The non-linear inner scales, 0 to 0.5, 0 to 1.5, and 0 to 5, are used only for a-c voltage measurements of 5 volts or less. The a-c scales are calibrated to read the r-m-s value of a sine-wave voltage or 0.707 times the peak value of a voltage of complex wave shape. The accuracy of all d-c ranges and all a-c ranges at low frequencies is $\pm$ 2% of full scale.

A mirror has been located between the inner and outer groups of scales. The mirror, of course, does not increase the accuracy of the voltmeter, but it does make it possible to read a voltage more precisely than when no mirror is present. The extra precision of reading or setting is most important when small differences must be observed. Without a mirror, parallax makes it very difficult to observe small movements of the pointer, especially when the attention of the observer must be removed momentarily from the meter to some part of the circuit under test. With a mirror, the eye can be placed precisely over the pointer for every reading. The upper portion of the pointer has been made knife-edged to facilitate further the precise observation of small differences. The lower portion has been made broad so that the approximate position of the pointer can be seen from some distance away.

Quite often small differences must be observed in an a-c voltage that is not properly read on one of the linear scales under the knife-edged portion of the pointer. If the voltage is above approximately 0.5 volt, the difference can be observed on the corresponding linear scale with only a small error. If the voltage is less than approximately 0.5 volt, the difference read on the linear scale must be corrected for the nonlinearity of the a-c scale. Of course, the actual voltage level, if it must be known, should be read on the correct a-c scale.

The meter face is illuminated by two lamps inside the meter case. The illumination makes a pilot light unnecessary and eliminates bothersome reflections from the glass over the meter face.

PANEL

On the panel, in addition to the meter, are input terminals for a-c and d-c voltage measurements, and three operating controls. The panel is shown in Figure 1. The d-c terminals, at the upper left, provide two different input impedances. The input resistance at the vertical pair is 10 megohms, while the horizontal pair apply the voltage directly to the grid of the d-c amplifier tube. A link is provided to short-circuit the pair of terminals not being used. The a-c terminals, which can be used only when the probe is plugged into its position behind the panel, are at the upper right. The LOW terminal is not grounded directly to panel but is isolated by a blocking condenser. A short-circuiting link is provided to ground the LOW terminal if it is desired to do so. Internal connections for a-c or d-c voltage measurement are made by the combined power and selector switch located at the left under the meter.

The center knob operates the VOLTAGE RANGE selector switch, which is used to select the appropriate voltage range for either a-c or d-c measurements. The zero control is operated by the knob at the right. The zero setting is not changed by operation of the VOLTAGE RANGE switch, although slight readjustment may be required when the A-C — D-C switch is operated. The line fuses, located at the bottom of the panel, can be replaced with the aid of a screwdriver without removing the instrument from its cabinet.

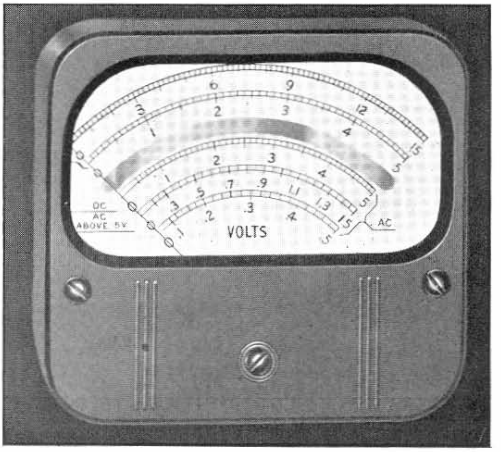


Figure 2. Close-up of the meter scales. The legend at the left is carried to the top two scales by a line that, in this view, is hidden by the meter case.

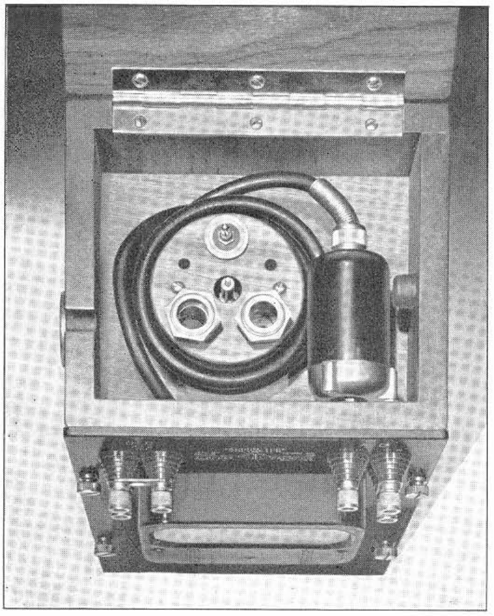


Figure 3. View of the storage compartment at the top of the cabinet. The probe fittings are stored here. The preliminary model used for this photograph has only one slot for the probe cable. Final models have a slot on each side of the cabinet.

CABINET

The shielded walnut cabinet is provided with a storage compartment at the top for the probe and the various terminals supplied. Figure 3 is a view of the compartment with the probe plugged into position to allow use of the panel a-c input terminals. There is a slot in each side of the cabinet for passage of the probe cable when the probe is used externally and the compartment cover is closed. The two slots have proved to be very convenient when the probe must be used at some distance from either side of the instrument.

The voltmeter can be used without separate support in any one of three positions: vertical, horizontal, and inclined. The three positions of use, together with the small size and light weight, make the instrument easy to use in practically any set-up.

In an inclined position, the instrument is supported by the metal carrying handle at an angle of approximately 30 degrees from the horizontal. The handle locks automatically either in the vertical position, i.e., parallel to the panel, or in the horizontal position. It is released merely by pressing in on the handle near each hub.

PROBE CIRCUIT

The a-c input voltage is rectified by a diode, which, at input voltages over 5 volts or so, gives a linear d-c voltage output very nearly equal to the peak value of the a-c input voltage. At lower voltages, the d-c voltage output tends to vary as the square of the a-c input voltage. Therefore, non-linear scales are needed for measurement of a-c voltages up to 5 volts.

The tube chosen for use as the diode rectifier is the Type 9005 acorn tube. The 9005 was chosen because its natural frequency of 1500 megacycles is the highest of any diode commercially avail-

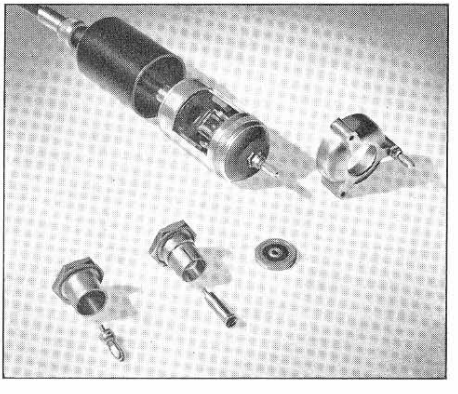
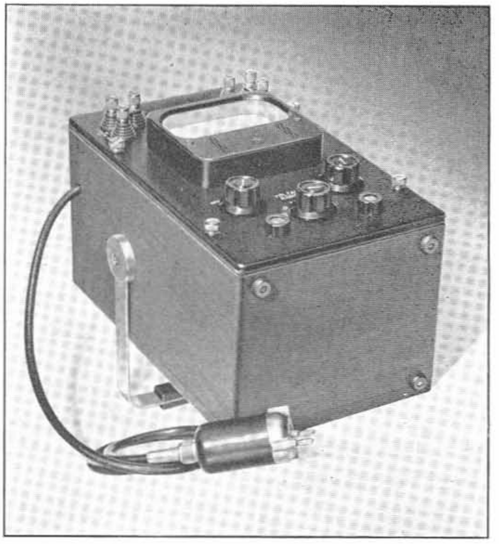


Figure 4. Exploded view of the probe and the various terminal fittings.

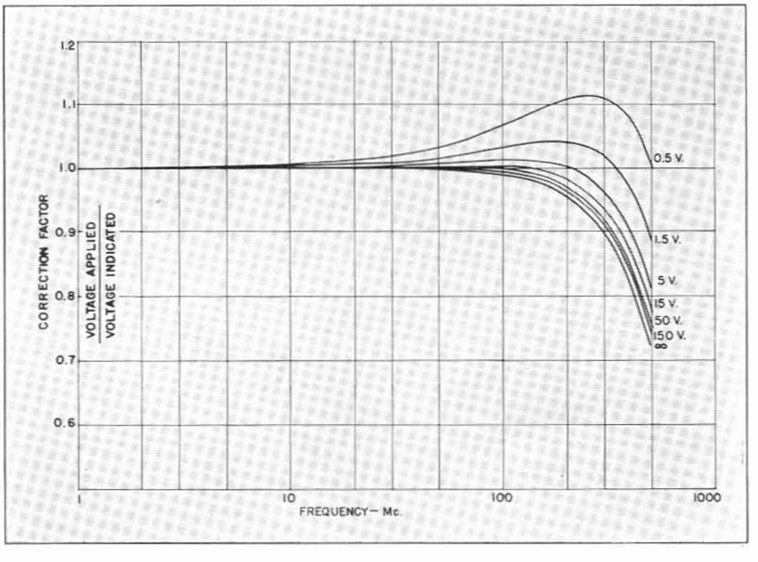
Figure 6. As shown here, the voltmeter can be used with the panel inclined, support being provided by the handle, which locks into position.



able. Because of great care taken in the design of the probe containing the diode and the associated circuit elements (Figure 4), the resonant frequency of the probe input circuit has been kept at 1050 megacycles — two-thirds as high as the resonant frequency of the tube itself. The high resonant frequency was obtained by making the probe small and compact and by minimizing the length of all input connections. The plate of the diode is coupled directly to the high input terminal by a button-type condenser, and the cathode of the diode is connected to the probe shell by a very short connecting strip.

The resonant frequency of 1050 megacycles makes it possible to measure voltages over a wide frequency range without applying corrections. Up to 300 Mc, for instance, the maximum error is $\pm 12\%$ when the indicated voltage is 0.5 volt or more. If the indicated voltage is less than 0.5 volt, the error will be larger because of the increase of the transit-time error of the diode. In Figure 5 are shown curves from which corrections for frequency error can be obtained. The curves show that the frequency error caused by input circuit resonance is in the opposite direction from the transit-time frequency error. For instance, at 0.5 volt indicated, the correction for transit-time balances the correction for resonance at 500 Mc and the net error is zero. When accurate measurements are not necessary, the instrument can be used as a voltage indicator up to 2500 Mc.

Figure 5. Frequency correction for a number of different indicated voltages. The curve marked ∞ is included to show the complete resonance effect.



The transit-time (or premature-cutoff) error varies directly with the interelectrode spacing and would, therefore, be less with a tube having more closely spaced elements. The spacing, however, is only a few thousandths of an inch, and it seems impractical to decrease this spacing without lowering the voltage rating to a point where the high-voltage limit would be too low for an instrument designed to respond to voltages up to 150 volts.

An inspection of the curves makes it clear that the premature cutoff, in general, contributes substantial errors at all voltages at frequencies for which the resonance-rise corrections can be made with fair accuracy. The performance of the instrument is therefore rather considerably determined by the limitations of the tube itself, and further improvement in resonance-rise conditions would lead to only a minor increase in the useful frequency range. For this type of voltmeter, then, it would seem that the practical high-frequency limit with existing tubes had been approached closely.

The user must remember that in order to obtain the frequency characteristics given in Figure 5, he must be very careful when making connection to the probe. The relationship between the voltage at the probe terminals and the voltage that it is desired to measure at some point in a circuit depends upon the connections used. The voltage applied to the probe is that at its terminals, which is not necessarily the same as that at the far end of the connecting leads. At high frequencies the connection must be as short and direct as possible.

The three types of terminals supplied for use with the probe are shown in Figure 4. They include a Type 274 plug terminal and Type 774 male and female coaxial terminals. A 50-ohm disc-type resistor is supplied for use with the coaxial terminals. The metal cap shown directly in front of the probe is used to attach all terminals to the probe. There are three threaded holes around the edge of the cap instead of only the one which is necessary for attachment of the Type 274 plug low terminal. The three holes are provided so that the cap can be fastened to a metal sheet or other flat metal surface when a minimum ground-connection inductance is desired. A hole in the metal sheet or surface makes the high probe input terminal accessible.

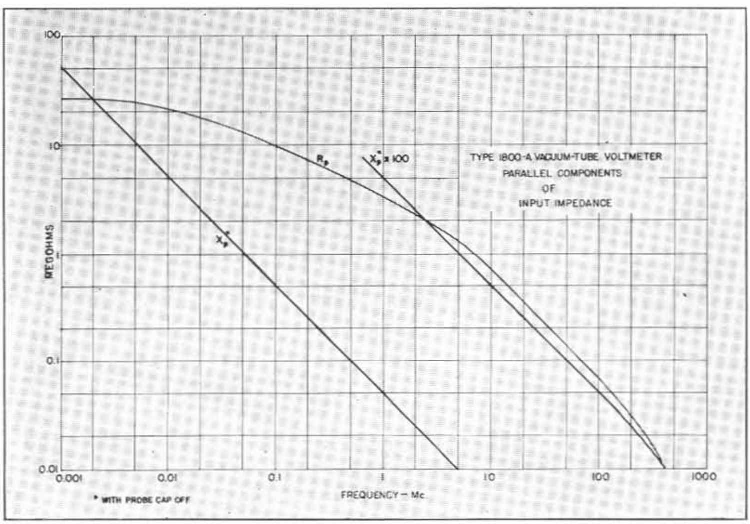


Figure 7. Plot of impedance vs. frequency for the parallel components of input impedance. The equivalent parallel capacitance is 3.1 $\mu\mu$f.

At frequencies low enough so that higher input capacitance and lower input natural frequency are not important, the probe can be plugged into position in the storage compartment and the a-c voltage applied to the panel a-c input terminals.

The probe shell is equipped with a cylindrical shield that has an aperture through which the diode is accessible. When the shield is rotated until the aperture into the probe is closed, the probe is completely shielded. A molded bakelite shell slips over the probe and is held in place by an insulated nut. The exposed metal at the front end of the probe is the only exposed part of the probe that can be at a d-c potential to ground. Connection to the d-c amplifier is made with a three-wire, shielded cable about three feet long.

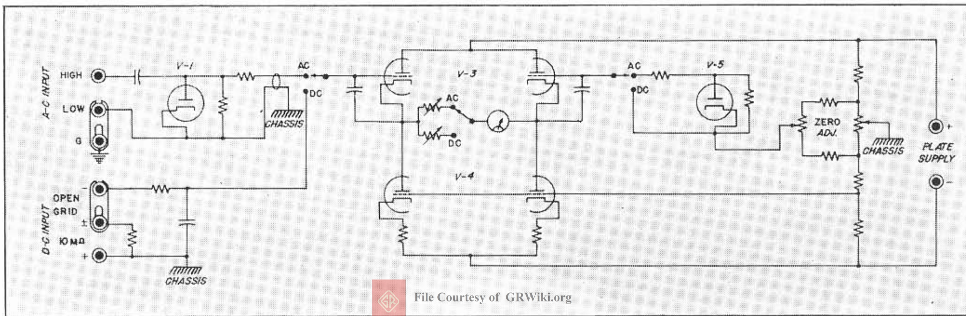
When voltage measurements are made on electronic circuits, or, for that matter, on circuits of any kind, the measuring device should disturb the circuit as little as possible. That is, the input impedance of the measuring device should be as high as possible. A 100-megohm resistor is shunted across the diode rectifier and another 100-megohm resistor is connected in series with the lead from the diode plate to the grid of the d-c amplifier tube. These resistance values give an effective parallel input resistance of 25 megohms at low frequencies. At high frequencies the input resistance is reduced by the Boella effect[2] in the resistors and by increased losses in the dielectrics. The input capacitance to the probe proper is about 3.1 micromicrofarads of which 0.8 micromicrofarad is the plate-to-cathode capacitance of the diode. When the cap used for attaching the terminals is screwed on the probe and the Type 274 plugs are used, the input capacitance is increased about 1 micromicrofarad. Figure 6 shows curves of probe input resistance and input reactance versus frequency. The input reactance curve is calculated for an input capacitance of 3.1 micromicrofarads.

Amplifier

The bridge-type amplifier is shown in simplified form in Figure 8. The indicating meter is connected in series with the appropriate precision wire-wound resistor between the cathodes of the twin-triode voltmeter tube, V-3. The cathodes of V-3 are connected directly to the plates of V-4, another twin triode. The latter tube is used to provide high degeneration without using high values of resistance from the cathodes of V-3 to ground and, consequently, a high plate-supply voltage. Each triode section of V-4, together with its cathode resistor, has the effect of a 7-megohm incremental resistance to voltage changes from the cathode of V-3 to B-; yet the plate-supply voltage need be only 450 volts because of the low d-c drop in the

[2] Boella, M., "Sul comportamento alle alta frequenze di alcuni tipi di resistenze elevate usate nei radio-circuiti," *Alta Frequenza*, Vol. 3, p. 132, April 1934.

Figure 8. Elementary schematic circuit diagram of the Type 1800-A Vacuum-Tube Voltmeter.



tube. Since the incremental resistance from the V-3 cathodes to B- is high compared to the resistance of the meter circuit, the amount of degeneration is determined almost completely by the meter-circuit resistance. At voltage ranges of 15 volts full scale and higher, the degeneration is sufficient to prevent tube changes from affecting the calibration of the instrument. Controls are provided for adjusting the calibration of the 0.5, 1.5, and 5-volt a-c and d-c ranges in case it is ever necessary to change V-3. Change of any other tube except V-1, the diode rectifier, has no effect on the calibration. Changing V-1 may require readjustment on the low a-c ranges.

Input to the voltmeter tube, V-3, for a-c or d-c measurements is selected by the combined power and selector switch mentioned previously. The d-c input system provides an R-C filter to prevent any ripple voltage from reaching V-3. The desired input resistance is obtained by proper connection of the short-circuiting link at the panel D-C input terminals. The input resistance of the OPEN GRID circuit is determined by the insulation resistance and the grid current which flows from V-3 through the voltage source. When the a-c input is selected, a diode, V-5, and circuit identical to that in the probe is connected to the grid of the inactive triode of V-3. The purpose of this diode is solely to balance the voltage developed at the other grid of V-3 from contact potentials and initial velocity of electrons in the probe diode. V-1 and V-5 must be selected so that each tube gives approximately the same value of initial developed voltage. The heater current for the two diodes, V-1 and V-5, is regulated with an Amperite 3-4 ballast lamp in order to maintain the zero of the low voltage a-c ranges as stable as possible.

Zero adjustment of the indicating meter is accomplished by adjusting the grid biases of V-3 very slightly. Bias adjustment is made on one grid to take care of large zero shifts such as may be encountered when V-3 is changed. Bias adjustment on the other grid, which should be sufficient for all normal adjustment necessary during actual use of the instrument, is made with the panel ZERO control. A third zero control is provided to adjust the a-c zero into coincidence with the d-c zero. The control varies the division of heater current between the two diodes, V-1 and V-5, and so changes the initial voltage developed by each diode. Coincidence of the two zeros is of interest only for the sake of convenience. Lack of coincidence has no effect on the accuracy of measurement. A change of resistance in series with the indicating meter, which happens whenever the VOLTAGE RANGE switch is rotated, has no effect on the zero reading.

Power Supply

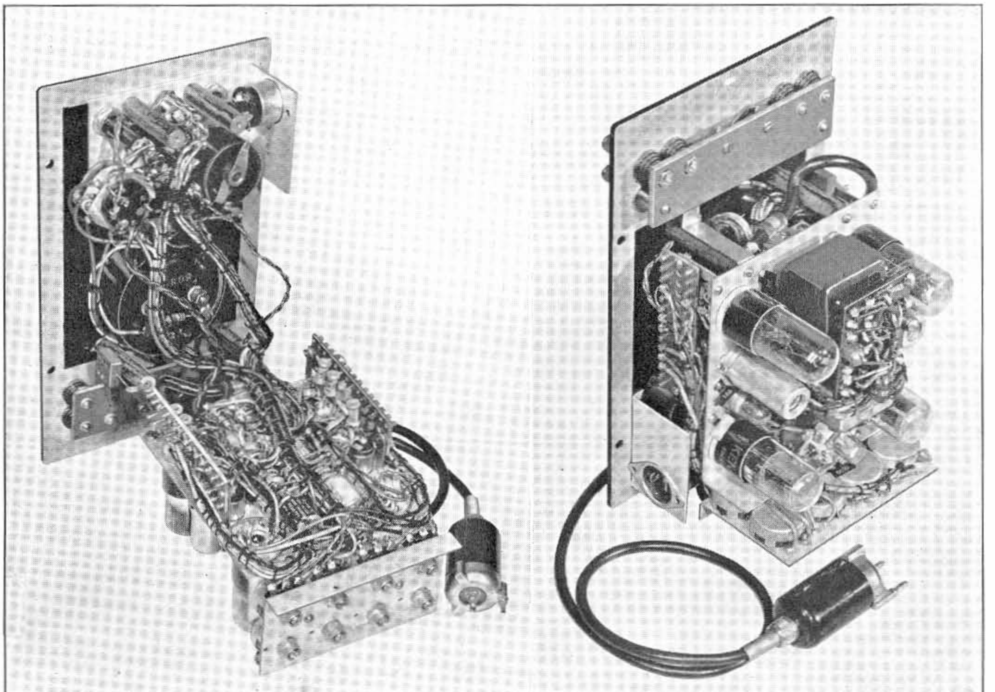
In order to maintain an accurate, stable meter reading on the low-voltage ranges, the plate supply voltage must be held constant regardless of power-line voltage fluctuations. Stabilization of the plate-supply voltage is obtained by use of an electronic voltage-stabilizing circuit, consisting of two vacuum tubes and two neon lamps. Both the shunt amplifier tube of the stabilizing circuit and the variable-resistance series tube are miniature types. The voltage across two neon lamps in series is used as the stable reference voltage of the system. When adjusted properly, the circuit maintains the plate-supply voltage within a volt or two of 450 volts as the power-line voltage is varied from 105 to 125 volts. Heater voltage changes in all tubes except the two diodes, V-1 and V-5, have no appreciable effect on the meter reading. The heater current of V-1 and V-5 is, as mentioned above, regulated by a ballast lamp.

The controls used for calibration and circuit adjustment are mounted on a shelf at the bottom of the chassis where they are easily accessible when the instrument is removed from its cabinet. All of the circuit elements of the voltmeter except those in the probe are mounted on the panel and a small, compact chassis. The chassis circuit elements and the panel circuit elements are connected together with a flexible cable so arranged that the panel and chassis may be swung apart to allow easy access to all portions of the circuit. Figure 9 shows a view of the complete assembly and a view of the chassis and panel separated to expose all parts of the circuit for servicing. The two sections are separated by removing four screws and unsoldering two direct connections to the grids of the voltmeter tube, V-3.

The Type 1800-A Vacuum-Tube Voltmeter has many advantages over the older Type 726-A. It is smaller, lighter, and easier to use; it can measure d-c voltages; it is more stable against line-voltage changes; its input impedance is higher and its frequency range is substantially greater. Mechanically and electrically, full advantage is taken of modern techniques and new components to produce an outstanding new voltmeter.

C. A. WOODWARD, JR.

Figure 9. Two views of the chassis. At the right is shown the complete assembly, while the view at the left shows how chassis and panel separate for servicing.



SPECIFICATIONS

Voltage Range: 0.1 to 150 volts, a-c, in six ranges (0.5, 1.5, 5, 15, 50, and 150 volts, full scale); 0.01 to 150 volts, d-c, in six ranges (0.5, 1.5, 5, 15, 50, and 150 volts, full scale).

Accuracy: D-C, $\pm 2\%$ of full scale on all six ranges. A-C, $\pm 2\%$ of full scale on all six ranges for sinusoidal voltages.

Waveform Error: On the a-c voltage ranges, the instrument operates as a peak voltmeter calibrated to read r-m-s values of a sine wave, or 0.707 of the peak value of a complex wave. On distorted waveforms the percentage deviation of the reading from the r-m-s value may be as large as the percentage of harmonics present.

Frequency Error: At high frequencies resonance in the input circuit and transit-time effects in the diode rectifier introduce errors in the meter reading. The resonance effect causes the meter to read high and is independent of the applied voltage. The transit-time error is a function of the applied voltage and tends to cause the meter to read low. The curves of Figure 5 give the frequency correction for several different voltage levels. It will be noted that at low voltages the transit-time and resonance effects tend to cancel, while at higher voltages the error is almost entirely due to resonance.

This voltmeter may be used at frequencies as low as 20 cycles with a frequency error of less than 2%.

Input Impedance: At low frequencies the equivalent parallel resistance of the a-c input circuit is 25 megohms. At higher frequencies this resistance is reduced by losses in the shunt capacitance. The equivalent parallel capacitance at radio frequencies is $3.1 \mu\mu f$ with the probe cap and plug removed. At audio frequencies this capacitance increases slightly. The probe cap and plug add approximately $1 \mu\mu f$. The accompanying plot gives the variation of R_p and X_p with frequency.

On the d-c ranges two values of input resistance are provided, 10 megohms and open grid.

Power Supply: 105 to 125 volts or 210 to 250 volts, a-c, 50 to 60 cycles. The instrument incorporates a voltage regulator to compensate for supply variations over this voltage range. The power input is less than 25 watts.

Tubes: Two Type 9005, two Type 6SL7-GT, one Type 6AT6, one Type 6C4, one Type 6X5-GT, one Type 3-4, and two Type 991 are used; all are supplied.

Accessories Supplied: A seven-foot line connector cord, spare meter lamps and fuses; Type 274 and Type 774 terminations and 50-ohm terminating resistor for probe.

Mounting: Black crackle finish aluminum panel mounted in a shielded walnut cabinet. The carrying handle can be set as a convenient support for the instrument when placed on a bench with the panel tilted back.

Dimensions: (Width) 7⅜" x (depth) 7½" x (height) 11⅛", over-all.

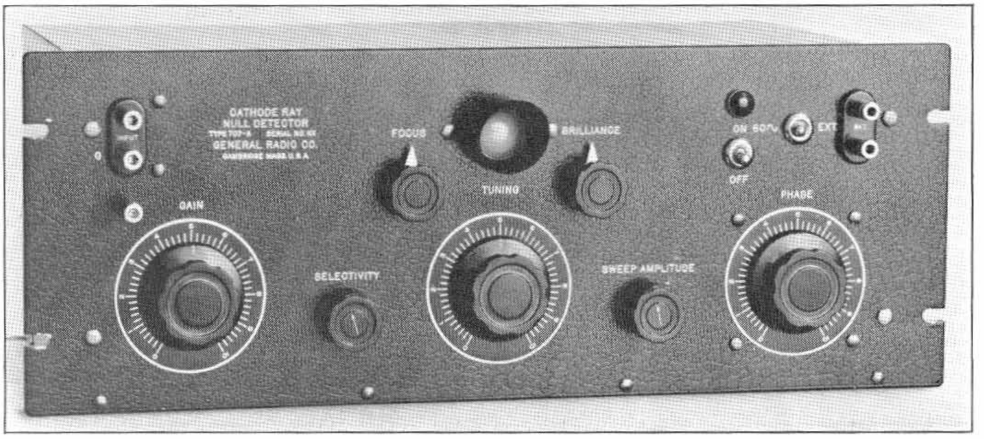
Net Weight: 13 pounds.

Type	*Description*	*Code Word*	*Price*
1800-A	**Vacuum-Tube Voltmeter**	DUCAT	**$305.00**

CATHODE-RAY NULL DETECTOR
Now Available

During the war, the Type 707-A Cathode-Ray Null Detector was temporarily discontinued in order that production facilities might be concentrated on more urgent items.

Repeated inquiries from our customers have made it evident that a demand exists for this instrument, and we are glad to announce that it is again available.



The Type 707-A Cathode-Ray Null Detector was described in the first and second editions of Catalog K. The catalog description is reprinted below.

TYPE 707-A CATHODE-RAY NULL DETECTOR

USES: This visual null indicator is intended for use as a balance detector in bridge and other null-method measurements at power-line and audio frequencies. When calibrated for a given frequency, it can be operated as a limit indicator. It can also be used for comparing frequencies by means of Lissajous figures or, when calibrated, used as an a-c millivoltmeter.

DESCRIPTION: The output of the bridge is applied through an 80-db highly selective amplifier,* operating on the degenerative principle, to the vertical deflecting plates of a one-inch cathode-ray tube. The bridge generator voltage is applied through an adjustable phase-shifting network to the horizontal plates. The tilted ellipse so formed is reduced to a horizontal straight line at balance.

FEATURES: Independent indications are given of the effect of balancing either the reactive or the resistive bridge control separately. This adds considerably to the speed and convenience of routine bridge measurements, and permits either bridge control to be balanced accurately without necessitating an accurate balance of the other. Indication is also given of the direction off balance of either one of the bridge controls, chosen at will.

This null indicator cannot be injured by overloading and is instantaneous in response and recovery. External fields do not affect its operation, and it radiates no appreciable field.

For bridge balancing, it is less fatiguing than headphones and can be used in noisy locations.

*U. S. Patent No. 2,173,426

SPECIFICATIONS

Input Impedance: One megohm.

Sensitivity: 150 μv at 60 cycles; 200 to 300 μv at 1000 cycles.

Selectivity: 40 decibels against second harmonic.

Frequency Range: Plug-in units tune the amplifier for any desired operating frequency between 20 and 2000 cycles. Continuous tuning range ±5% for each unit.

Temperature and Humidity Effects: When this instrument is operated under severe conditions of temperature and humidity some decrease in sensitivity and selectivity may be expected. For the low frequency tuning units the sensitivity may be reduced by as much as 6 db, while for all units the selectivity to the second harmonic may be reduced by 5 db. The above figures are for a relative humidity of 80% at 95° Fahrenheit.

Controls: Panel controls are provided for adjusting the focus and brilliance of the cathode-ray pattern, the phase and amplitude of the horizontal sweeping voltage, and the gain, selectivity, and tuning of the amplifier.

Accessories Supplied: A 7-foot line connector cord, spare pilot lamps and fuses, one Type 274-M Plug, and one Type 274-NC Shielded Conductor.

Accessories Required: One plug-in phasing circuit is used at any frequency below 400 cycles; one plug-in tuning unit for each operating frequency used. These are not included in the price of the instrument. (See price list on page 12.)

Power Supply: 105 to 125 volts, 40 to 60 cycles.

Power Input: 20 watts at 60 cycles.

Vacuum Tubes: One 6K7-G pentode, one 6F8-G twin triode, one 6J5-G triode, one 913 cathode-ray tube, and one 6X5 rectifier; all are supplied with the instrument.

Mounting: Standard 19-inch relay-rack panel. Walnut end brackets are supplied for table mounting.

Dimensions: Panel, 19 x 7 inches; depth behind panel, 9 inches.

Net Weight: 29 pounds.

Type		*Code Word*	*Price*
707-A	Cathode-Ray Null Detector†	NULTY	$250.00

†Less Phasing Unit and Tuning Unit (see next page).

PLUG-IN UNITS FOR TYPE 707-A CATHODE-RAY NULL DETECTOR

These units are required for use with TYPE 707-A Cathode-Ray Null Detector and are not included in the price of that instrument.

A phasing unit is necessary for operation at any frequency *below* 400 cycles. At 400 cycles and above, none is required. A tuning unit is required for each operating frequency. The tuning range is ±5%.

All units plug into mounting jacks provided inside the null detector.

PHASING UNITS

Type	*Description*	*Code Word*	*Price*
707-P1	**For Frequencies Below 100 cycles**	NULLTECANT	**$8.00**
707-P2	**For Frequencies Between 100 and 400 cycles**	NULLTECBOY	**8.00**

AMPLIFIER TUNING UNITS

Type	*Frequency*	*Code Word*	*Price*
707-P42	**42 cycles**	NULLTECCAT	**$30.00**
707-P50	**50 cycles**	NULLTECDOG	**30.00**
707-P60	**60 cycles**	NULLTECEYE	**30.00**
707-P100	**100 cycles**	NULLTECTAP	**30.00**
707-P400	**400 cycles**	NULLTECFIG	**30.00**
707-P1000	**1000 cycles**	NULLTECGUM	**30.00**
707-P2000	**2000 cycles**	NULLTECHIM	**30.00**

MISCELLANY

We were pleased to welcome recently, as a postwar visitor, Mr. Lewis M. Lyons, a member of the firm of Claude Lyons, Ltd., who have for many years represented the General Radio Company in Great Britain.

Other recent visitors to our plant and laboratories include our representative in Finland, Mr. K. L. Nyman of Helsinki, who was accompanied by Mr. K. S. Sainio, Chief Engineer of the Finnish Broadcasting Company; Dr. Alfred P. De Quervain, Assistant Chief Engineer, Electronics Department, Brown-Boveri Company, Ltd., of Baden, Switzerland, and Mr. Robert C. Habich of Berne, Switzerland, and Port Washington, New York; Dr. Stig Ekelöf, Professor of Theoretical Electricity and Electrical Measurements, Chalmers Tekniska Högskola, Gothenburg, Sweden, and Mr. P. Poppe of Oslo Lysverker, Oslo, Norway; Dr. R. W. Guelke of the South African National Physical Laboratory, Pretoria, South Africa; and Professor F. Dacos, Institute Montefiore, Liege, Belgium.

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A PEAK-READING VOLTMETER FOR THE U-H-F RANGES

Also
IN THIS ISSUE

• **PROGRESS IN AN** engineering field depends to a considerable degree upon the development of measuring instruments. As the field is extended through laboratory research, instruments are "tailor-made" for each project, but later commercial development usually brings with it simple, general-purpose direct-reading instruments. Initially of only moderate accuracy, they eventually are developed to an accuracy comparable with that achieved in older fields.

In the electrical-communication industry, the voltmeter affords an excellent example of this type of development. The frequency range of accurate measurements with voltmeters has been progressively extended from the audio-frequency limits of 20 years ago to the present figure of several hundred megacycles[1] approaching the limit to which currently available diodes can be pushed.

[1]C. A. Woodward, Jr., "A New Vacuum-Tube Voltmeter," *General Radio Experimenter*, September, 1946, pp. 1-10.

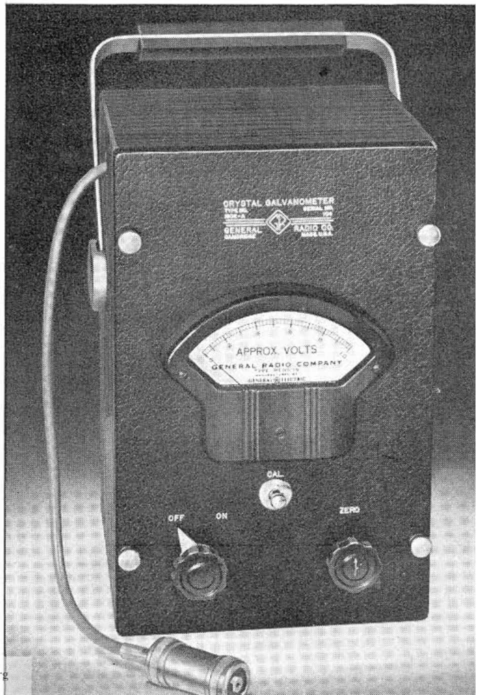


Figure 1. Panel view of the Type 1802-A Crystal Galvanometer.

For still higher frequencies, there is an obvious need for a direct-reading voltmeter. Such an instrument must necessarily be of the pioneering type, in which accuracy is sacrificed for convenience and extended frequency range. The new Type 1802-A Crystal Galvanometer is exactly this type of instrument. Its range of direct measurement is 0.1 to 1 volt with an accuracy of ±5%, and two multipliers are furnished to extend the range to 10 volts and 100 volts. It can be used as a direct-reading instrument at frequencies up to 1000 megacycles and for the measurement of voltage ratios, the frequency limit is well above 1000 megacycles.

Functionally, this new voltmeter is a peak-reading instrument, consisting of a rectifier and a d-c amplifier. The extended frequency range is obtained through the use of a crystal rectifier in place of the thermionic diode used in vacuum-tube voltmeters. Crystal rectifiers, however, cannot yet be produced to a degree of uniformity comparable with that of the thermionic diode, and this fact explains the difference in accuracy between the crystal instrument and the vacuum-tube voltmeter. The scale is calibrated directly in volts, but the instrument has been named a crystal galvanometer to indicate that its accuracy is not as good as the 2% we have come to expect from the vacuum-tube voltmeter.

The crystal rectifier used in this voltmeter is one of the new units developed during the war[2,3], which were widely used as mixers and as uncalibrated voltage indicators. In a previous article[3] the characteristics of these crystals as voltmeter rectifiers were discussed, and it was pointed out that their excellent high-frequency characteristics considerably outweighed their lack of uniformity.

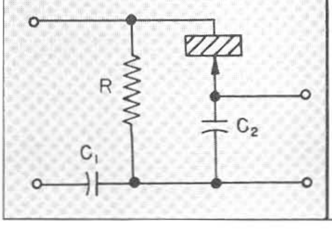
These crystals are commonly used as simple rectifiers in series with a meter. The use of a peak-reading circuit, however, has two important advantages: the input resistance is higher; and the variation in response between different crystals is less. The resistance of the 1N21B-type crystal is of the order of a few hundred ohms in the "forward" direction and from 15,000 to 100,000 ohms in the "reverse" direction. In the simple crystal-meter circuit the input resistance is approximately twice the forward resistance, while in the peak-reading circuit it is approximately one-third the back resistance, or from 5,000 to 30,000 ohms. Variations in crystal forward resistance affect the calibration directly in the series circuit, but in the peak-reading circuit, only the small difference between the peak applied voltage and the developed d-c voltage depends upon the crystal characteristics.

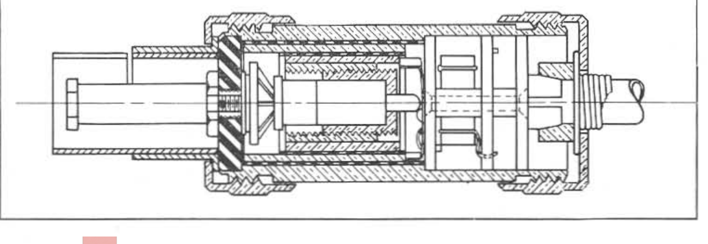
PROBE

The design of the probe in which the crystal rectifier is mounted is one of the most important factors in determining

[2]W. E. Stephens, "Crystal Rectifiers," *Electronics*, 19, 7, July, 1946, pp. 112-119.
[3]Arnold Peterson, "Vacuum-Tube and Crystal Rectifiers as Galvanometers and Voltmeters at Ultra-High Frequencies," *General Radio Experimenter*, 19, 12, May, 1945, pp. 1-7.

Figure 2. Schematic (*left*) and cross section (*right*) of the probe.





the high-frequency performance. A cross-section of the probe and the schematic of the crystal and its associated circuit elements are shown in Figure 2.

The r-f voltage is applied to the crystal through the series combination of C_1 and C_2. The impedance of this combination at frequencies above about 10 Mc is so low compared with that of the crystal that the a-c voltage is effectively applied across the crystal. The rectified current charges up the capacitance C_2 with the discharge path through the high resistance, R, and the crystal back resistance. The resulting d-c voltage developed across this capacitance is then applied to the amplifier through the probe cable.

To obtain the highest possible natural frequency, the blocking condenser is in the ground lead, thus avoiding the inductance and capacitance to ground of a blocking condenser in the high potential lead. The use of a battery-operated amplifier avoids the hum problem that occurs with this connection in an a-c operated diode voltmeter.

The probe assembly has been designed to minimize inductance throughout by using cylindrical blocking and by-pass condensers, as illustrated in the sectional view, Figure 2. The outer shell forms the ground connection, and, with the first inner cylinder, forms the capacitance C_1. The first inner cylinder, in turn, forms the capacitance C_2 with the center assembly. The resistor R is formed by a resistive coating on the surface of the mycalex insulating disc that carries the center terminal of the probe.

HIGH-FREQUENCY ERROR

This design together with the high resonant frequency of the crystal cartridge itself yields a probe with a resonant frequency of about 1800 Mc. The actual value for any unit depends on the particular crystal cartridge used. The resonant frequency varies with different units from 1650 to 2000 Mc. This value is sufficiently high that the resonance correction is only about 30 per cent at 1000 Mc. No transit-time or other frequency corrections need be applied. The correction factor for resonance is shown in Figure 3.

Frequently, the voltmeter is used only as an indicator of the existence of voltage. Then the resonant frequency is not a limiting factor on the useful frequency range of the instrument, and it can be used at frequencies up to and beyond

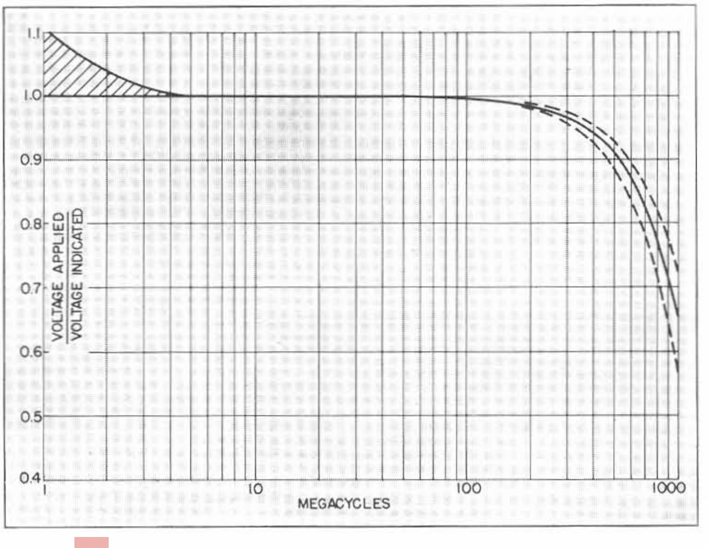


Figure 3. Frequency correction for resonance in the probe. The dash lines show the range of variation among normal crystals. Additional correction curves for use with the multipliers are supplied in the instruction book.

3000 Mc. Since there is no appreciable frequency correction that depends upon voltage level, measurements of voltage ratios can also be made at frequencies well above 1000 megacycles.

LOW-FREQUENCY ERROR

At frequencies below a nominal value between 10 and 30 Mc, the indication falls off as a result of the small series capacitance in the probe. The frequency at which this effect becomes noticeable depends on the characteristics of the individual crystal, particularly the back resistance. The voltmeter is not intended for use at frequencies below 10 Mc.

CALIBRATION

The crystal develops across the blocking condenser about 75 per cent of the peak voltage applied. Variations between crystals, therefore, will affect the calibration and show up as differences in absolute level and changes in the shape of the calibration curve.

In order to set the absolute level, provision is made for calibrating the crystal rectifier at the 0.7-volt level. The power cord in the upper compartment is used to connect to an a-c power line of 115 volts, and a voltage divider in the cabinet derives from this voltage the necessary 0.7 volt to be applied to the crystal rectifier. This calibration check can be used whenever the crystal is replaced.

The meter scale is calibrated for the average crystal and, when the level is standardized at 0.7 volt, will match the individual Type 1N21B Crystal Rectifier to ±5%.

AMPLIFIER

The d-c amplifier uses a degenerative cathode-follower circuit with a 1R5-type tube as a triode. The circuit is arranged as a bridge system, as shown in Figure 4. The reading of the indicating meter is set initially to zero by adjusting the bias resistor until balance is obtained. The resistor is connected so that, as the bias is varied, the degenerative resistance for voltages applied by the crystal is maintained nearly constant. This arrangement makes the sensitivity practically independent of the zero setting. The sensitivity of the amplifier is changed by the adjustable series resistance in the meter circuit, and this adjustment does not affect the zero setting.

PROBE FITTINGS

Since the inverse peak voltage rating of the high-frequency crystal rectifiers is only a few volts, the range of the voltmeter can be extended only slightly by reducing the sensitivity of the d-c amplifier. Voltages above one volt are measured by the use of capacitance voltage dividers placed ahead of the probe. Two of these multipliers are furnished to give a complete coverage to 100 volts in decade steps.

In addition to the multipliers, fittings are provided for two coaxial connectors of the General Radio 774 type. A 50-ohm disc-type resistor is also provided. These are illustrated in Figure 5. The 50-ohm resistor can be used in conjunction with the voltmeter for making approximate measurements of power up to 1000 Mc.

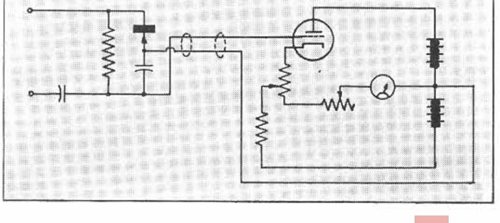


Figure 4. Elementary schematic circuit diagram of the complete voltmeter.

PANEL AND CABINET

As shown in Figure 1, the meter, mounted in the center of the panel, carries only the basic range of 0 to 1.0 volt. The two main controls, which are the on-off switch and the zero set, are operated by the knobs below and on either side of the meter.

The shielded walnut cabinet is of the same construction as that for the TYPE 1800-A Vacuum-Tube Voltmeter with the storage compartment at the top for the probe and the various fittings supplied. The handle can be locked in either the position parallel to the panel or perpendicular to it, thus permitting the instrument to be set in a vertical, a horizontal, or a 30-degree position.

The TYPE 1802-A Crystal Galvanometer brings to the u-h-f range of frequencies the convenience and flexibility of the vacuum-tube voltmeter. The variety of probe terminations including coaxial connectors facilitates its use with other laboratory instruments. Experimental models have already been used by a few laboratories engaged in microwave research. Their experience indicates that it is a most useful and convenient voltmeter for the u-h-f ranges.

— ARNOLD PETERSON

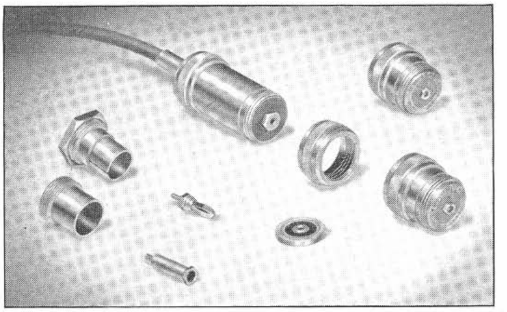


Figure 5. View of the probe and fittings. The capacitance-type voltage multipliers are shown at the right.

SPECIFICATIONS

Voltage Range: 0.1 to 1 volt, direct-reading; 1 to 10 volts and 10 to 100 volts direct-reading with multipliers supplied.

Accuracy: $\pm 5\%$ of full scale on sinusoidal voltages, subject to frequency correction.

Waveform Error: The meter response approaches that of a peak voltmeter, but the scale is calibrated in r-m-s values for a sine-wave input. On distorted waveforms the percentage deviation of the reading from the r-m-s value may be as large as the percentage of harmonics present.

Frequency Error: High-frequency error is caused by resonance effects. The resonant frequency depends on the particular crystal used. With a 1N21B crystal in the probe, the resonant frequency is between 1650 and 2000 Mc; with the 10:1 multiplier attached, between 1700 and 2200 Mc; and with the 100:1 multiplier attached, between 1350 and 1650 Mc. For frequencies below resonance the applied voltage at the terminals is approximately equal to the indicated voltage multiplied by $1 - \left(\frac{f}{f_o}\right)^2$, where f is the operating frequency and f_o is the resonant frequency. At frequencies below a nominal value between 10 and 30 Mc the output falls off as a result of the small capacitance in the probe. The frequency at which this effect becomes noticeable depends upon the characteristics of the individual crystal.

Input Impedance: The input capacitance is nearly independent of the standard crystal used, but the input conductance depends on the frequency, voltage level, and the crystal characteristics. For 1N21B crystals representative values are:

Probe:	Capacitance	— 5 $\mu\mu f$
	Conductance	— 100 μmhos
10:1 Multiplier:	Capacitance	— 2.5 $\mu\mu f$
	Conductance	— Less than 25 μmhos
100:1 Multiplier:	Capacitance	— 1.6 $\mu\mu f$
	Conductance	— Less than 10 μmhos

Power Supply: One Burgess Z3ON and one Burgess 2F batteries are furnished.

Vacuum Tube: One 1R5-type vacuum tube is supplied.

Crystal: One 1N21B-type crystal is supplied.

Mounting: Walnut cabinet; cable and probe are stored in cabinet.

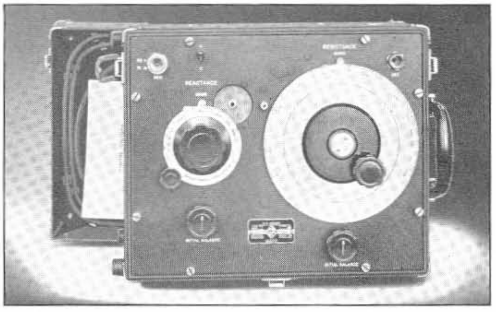
Accessories Supplied: One TYPE 1802-P1 10:1 Multiplier, one TYPE 1802-P2 100:1 Multiplier, one TYPE 1802-P3 Cable Termination, fittings for plugging into TYPE 774-G or 774-M coaxial connectors.

Dimensions: 7 x 7 x 10½ inches, overall.

Net Weight: 10¾ pounds.

Type		*Code Word*	*Price*
1802-A	**Crystal Galvanometer**	CONIC	**$175.00**

LEAD CORRECTIONS FOR THE RADIO-FREQUENCY BRIDGE



In the measurement of impedance with the TYPE 916-A Radio-Frequency Bridge, the bridge settings at balance indicate the magnitude of the unknown impedance as seen at the bridge terminals, which includes the impedance of the connecting leads. Standard leads of known reactance are supplied with the bridge to facilitate making the necessary connections for lead reactance to ground.

In the instruction book for this bridge, approximate expressions for the correction are given, i.e.

$$R_x = R_e \left[1 + 2\frac{X_e}{X_a} - \left(\frac{R_e}{X_a}\right)^2\right]$$

$$X_x = X_e + \frac{X_e^2 - R_e^2}{X_a}$$

where R_x = Resistance of unknown
X_x = Reactance of unknown
R_e = Resistance indicated by bridge setting.
X_e = Reactance indicated by bridge setting.
X_a = Lead reactance (obtained from chart in instruction book).

There are a number of cases where these approximate expressions are not adequate and where their use will lead to appreciable errors in the result. Because of this, the following more complete expressions have been derived:

$$R_x = R_e\left[1 + 2\frac{X_e}{X_a} + 3\left(\frac{X_e}{X_a}\right)^2 - \left(\frac{R_e}{X_a}\right)^2\right]$$

$$X_x = X_e + \left[\frac{X_e^2 - R_e^2}{X_a} + \frac{X_e}{X_a}\left(\frac{X_e^2 - 3R_e^2}{X_a}\right)\right]$$

These corrections and a corrected set of illustrative examples of their use have been printed in the form of an errata sheet to be inserted in the instruction book for TYPE 916-A Radio-Frequency Bridge. A copy of this sheet will be sent to any user of the bridge upon request.

When asking for the errata sheet, please give the serial number of the bridge with which it is to be used.

DELIVERIES

Every now and then in these pages we try to keep our readers posted on the operations of the General Radio Company. We have in recent years published several articles about our organization, how we try to operate, and what we are trying to do.

Just now the thing that probably concerns all of us most is the question of deliveries. As was to be expected, the war brought about an enormous demand for precision test equipment. Looking back, we are now rather proud of our record and the record of our industry in this all-out race for production. It is not easy to produce precise and frequently intricate technical equipment. It requires highly trained manpower, complicated production procedures, and frequently very scarce components. In spite of these factors, we were able to increase production to more than six times its prewar peak. We had thought that when the war was over we could expect a large decrease in demand which could easily be met by the now existing facilities. What we did not fully realize was the extent of the demands of the starved civilian industry nor the size of the requirements for the rehabilitation of the war-ruined industries and laboratories abroad.

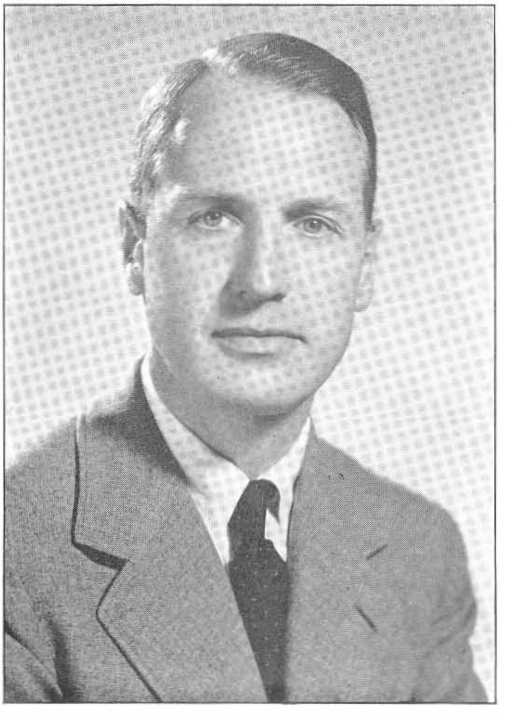
These demands, coupled with a scarcity of component parts and materials, many of which are in acutely short supply, have resulted in delays in delivery of many of our products which must often seem to be unreasonable.

Those of our readers who are in the manufacturing business will know what we mean by the extreme delays involved in obtaining key components. Materials like copper wire and some classes of transformer steel are almost unobtainable anywhere on any delivery schedule or for any price.

We are glad that the great majority of our customers are well acquainted with these problems and are being patient when we often must quote delivery delays of many months. We wish to assure all of our customers who have been patiently waiting for delivery that we are doing everything we can to get the instruments produced; and that, in spite of acute spot shortages, our entire enlarged plant is operating at full capacity, and, finally, that we will not sacrifice quality and care in production just to speed deliveries.

MISCELLANY

The Medal of Freedom was presented to Dr. W. N. Tuttle of the General Radio Company on August 27 at an informal ceremony in the engineering library of the Company, witnessed by his colleagues on the engineering staff.



The presentation was made by Major General Harold M. McClelland, formerly Air Communications Officer of the Army Air Forces and now commanding Army Air Forces Air Communications Service. The award was made "for exceptionally meritorious achievement which aided the United States in the prosecution of the war against the enemy in Continental Europe as head, radar subsection, Operational Analysis Section, Headquarters, Eighth Air Force, from 16 October 1942 to 3 October 1944. During this period Dr. Tuttle distinguished himself by adapting blind bombing equipment and technique for introduction into the Eighth Air Force, making a very valuable contribution to the success of numerous close cooperation attacks. His vision and untiring efforts reflect high credit upon him and were of great importance to the operations of the armed forces of the United States."

Dr. Tuttle is a graduate of Harvard, Class of 1924, and received his S.M. in 1926 and his Ph.D. in 1929 from that University. He is a senior member of the Institute of Radio Engineers, a member of the Acoustical Society of America, of the American Physical Society, and of Sigma Xi.

VISITORS

A recent visitor to General Radio, returning after several weeks in Paris, was Mr. Paul Fabricant, of the Paris firm Radiophon, our representatives in France. Other visitors include Mr. Frederick S. Barton, of the British Air Commission; Dr. Lewis M. Hull, of Aircraft Radio Corporation; Mr. O. B. Ottesen, of Jan Wessels Radiofabrikk, Oslo, Norway; and Mr. Louis E. LeBel, of Marianno Soares & Cia., Ltda., Rio de Janeiro and São Paulo.

NOTICE OF PRICE CHANGE

Because of rapidly rising costs, all prices given in our current price lists are increased by 10% effective October 1, 1946, for shipments to be made after December 31, 1946.

This will, of course, not affect any firm orders that you may now have with us, but is applicable only to new orders.

We regret the necessity of this price change, which we have resisted as long as possible, but we know that our customers will understand the conditions that have made it necessary.

We also regret that, contrary to our long-established practice of quoting firm delivered prices, it is now necessary to limit to six months the period for which we will guarantee a price. For shipments which cannot be made within six months from our receipt of the order, customers will be advised prior to shipment if any price changes have been made, at which time the order may be canceled without obligation.

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THE General Radio EXPERIMENTER

VOLUME XXI No. 6 NOVEMBER, 1946

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

THE NEW BROADCAST MONITORING EQUIPMENT

Also IN THIS ISSUE

• EVER SINCE commercial broadcasting began in 1922, General Radio monitoring equipment has been used to measure station performance. The first General Radio frequency monitors were general purpose wavemeters, later displaced by narrow-range, more precise types as frequency tolerances were narrowed. Simple crystal oscillators were used in the late 1920's and, by 1929, temperature control was required. The direct-reading frequency-deviation meter[1] was added to the improved crystal oscillator[2] in 1932 to provide a continuous indication of magnitude and direction of transmitter frequency drift. Successive models incorporated a number of improvements, but the basic design remained unchanged until the new, postwar Type 1181-A Frequency Deviation Monitor appeared.

[1]U.S.Pat. No. 1,944,315.
[2]U.S.Pat. No. 2,012,497.

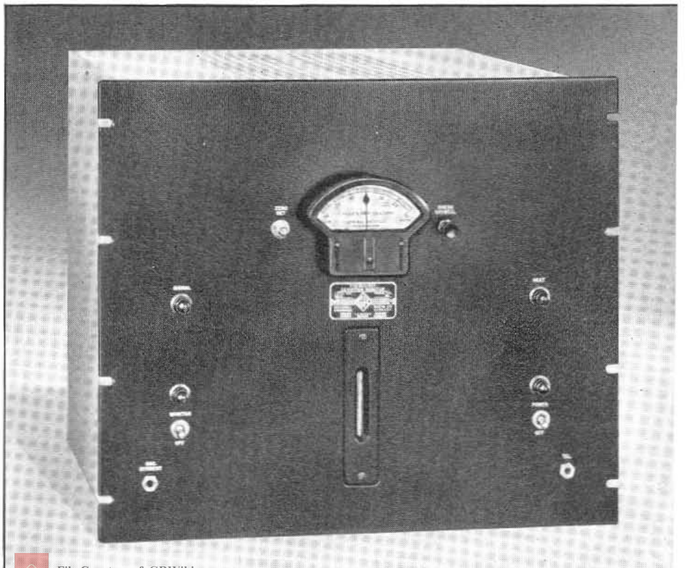


Figure 1. Panel view of the Type 1181-A Frequency - Deviation Monitor.

Development of the modulation monitor has followed a similar course. An instrument for the direct measurement of modulation percentage, the Type 457-A Modulation Meter[3], was introduced by General Radio in 1930. This was followed in a few years by one of the first continuously indicating modulation monitors[4], the Type 731. The latest model, Type 1931-A, embodies a number of improvements not found in older models.

The new frequency and modulation monitors are the result of many years' experience in the design and construction of this class of equipment. Designs were partially complete in 1941, but priority regulations did not permit the manufacture of broadcast monitors during the war. These new instruments are now in production, although deliveries are still hampered by material shortages. No announcement of the new monitors has hitherto been made in the *Experimenter*, although some hundreds of monitors have already been sold. The following brief descriptions may be of interest to many of our readers who are not directly connected with the broadcasting industry.

TYPE 1181-A FREQUENCY DEVIATION MONITOR

The outstanding operating feature of this monitor is its ability to operate on a modulated signal. Hitherto, it has been necessary to couple the monitor to a point in the transmitter where an unmodulated signal was available. This often posed a problem, particularly in transmitters with low-level modulation. Operation directly from the modulated transmitter output is a considerable improvement, simplifying both installation and operation.

The system[5] for making the deviation indication independent of waveform is indicated in the block diagram of Figure 2. By means of clipping and limiting circuits, the beat frequency waveform is modified to a constant, square waveform, thus eliminating the amplitude variations caused by modulation. After suitable amplification, these square waves are applied to a discriminator circuit (or frequency meter), similar to that used on previous models, but modified somewhat for operation on square waves.

As a further simplification of the coupling problem, the sensitivity has

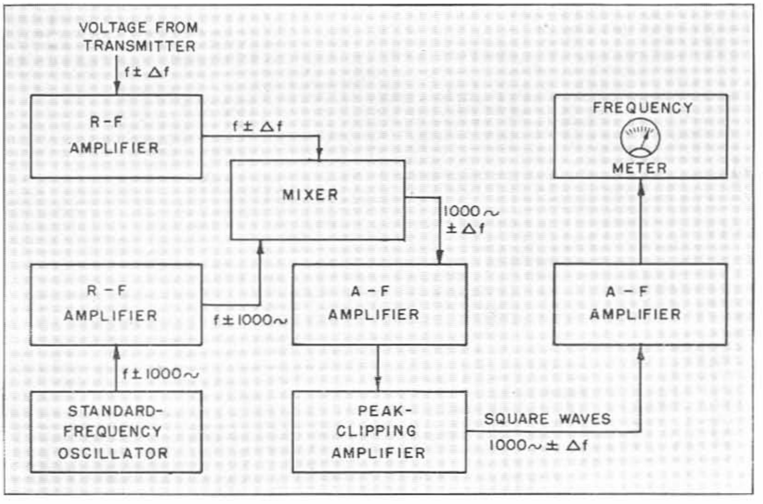


Figure 2. Functional block diagram showing the operation of the frequency-deviation monitor.

[3]U.S. Pat. No. 2,012,291.
[4]U.S. Pat. No. 2,069,934.
[5]U.S. Pat. No. 2,362,503.

been increased so that sufficient pick-up is obtained on a few feet of wire, acting as an antenna.

An aperiodic amplifier is used for the r-f signal, thus assuring sufficient voltage to saturate the square-wave generating circuits under all normal conditions of operation and for high levels of modulation. Sufficient sensitivity is provided to permit remote monitoring if a tuned antenna is used.

The constant-temperature oven in the new monitor is smaller than that used in previous models, since modern, low-temperature-coefficient crystals do not require as precise a temperature control as the older types. This leads to an appreciable saving in space, which, with other improvements in mechanical design, makes the new monitor considerably less bulky than the two-panel arrangement previously used.

Mechanically, the assembly is designed for easy access to all parts and for efficient ventilation. Chassis assemblies run vertically at each end of the panel, so that the heat generated in tubes and transformers can be carried out at the top by the natural air stream through the center of the instrument. Submounted parts are accessible when cover plates are removed at the sides.

The illuminated meter is easily read at a considerable distance. Both the meter lamp and a pilot lamp are extinguished when the r-f input signal falls below the required minimum. A push button is provided for checking the crystal signal amplitude. As in previous models, a heat fuse protects the temperature-control system from burnout in case of thermostat failure.

Accuracy, convenience, and reliability were the design objectives in the Type 1181-A Frequency Deviation Monitor, and it meets or exceeds all F.C.C. specifications given in the Standards of Good Engineering Practice. It is now undergoing F.C.C. approval tests.

TYPE 1931-A AMPLITUDE-MODULATION MONITOR

While operating on the same principle as previous models, the new Type 1931-A Modulation Monitor embodies a number of improvements.

Various circuit changes have been made to achieve greater dependability and accuracy, particularly in the flashing lamp circuit. The meter scales are illuminated and easily readable from across the transmitter room.

One new feature of considerable importance to users of high-fidelity transmitters is an additional detector circuit which provides a demodulator output with less than 0.1% distortion. Measure-

Figure 3. Panel view of the Type 1931-A Modulation Monitor.

ments of audio distortion in the transmitter can be made by feeding this output to a TYPE 1932-A Distortion and Noise Meter or a TYPE 736-A Wave Analyzer. A low-impedance, 600-ohm output circuit is also provided for audible monitoring.

The TYPE 1931-A Amplitude-Modulation Monitor is approved by the F.C.C. and has been assigned Approval No. 1555.

A WIDE-RANGE U-H-F TEST OSCILLATOR

One of the instruments developed for the armed services by General Radio during the war was a low-power, portable, u-h-f oscillator which was used for making rough field checks of the alignment, calibration, and performance of several types of wide-range radar-intercept receivers. These receivers, covering the frequency range from 40 to above 1000 Mc, were used in the radar-countermeasure program for monitoring and locating enemy radar stations, for setting a jamming transmitter on the frequency of an enemy radar, and for many other purposes. Consequently, large numbers of them were installed on ships, in aircraft, and at ground positions. In order to attain maximum performance from the receivers, it was found desirable to test them frequently under actual operating conditions, as installed in a ship or aircraft. Since it was usually impractical to transport bulky test equipment to the operating site, the test instrument had to be a single small unit, simple to use and capable of operating from a wide variety of power sources. The unit developed by General Radio to fulfill this requirement was later manufactured by the Fairchild Camera and Instrument Company* to General Radio drawings, and bore the service designation, TS-47/APR Test Oscillator.

As can be seen from the panel view in

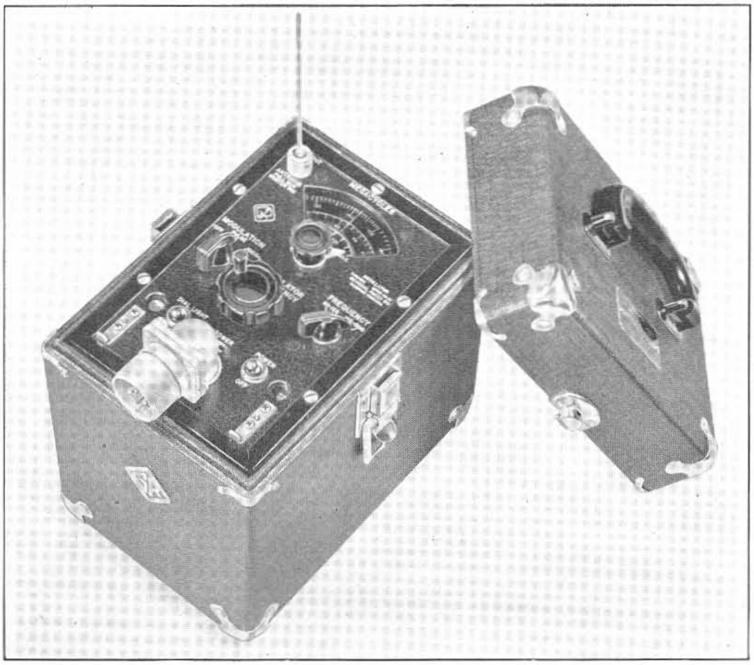


Figure 1. View of the instrument in its cabinet with cover removed.

*David W. Moore, Jr., "Test Oscillator TS-47/APR," *Radio News*, May, 1946.

Figure 1, the oscillator was unusually small and compact. The unit shown, containing a tunable oscillator, a modulator, an a-c power supply, and a built-in antenna, was housed in a moderately well-shielded case. The dimensions were only 10″ x 8″ x 11″ overall and the weight about 15 pounds. Oscillator tuning was controlled by a single knob, and the dial indicated the operating frequency directly within ±1%. The frequency range between 40 and 500 megacycles was covered on the fundamental of the oscillator with a maximum output of at least 5 mw at frequencies below 400 Mc and a somewhat lower output at higher frequencies. The output signal was rich in harmonics, and relatively strong usable signals were present up to frequencies of about 1500 Mc. Since only rough performance checks were required, no provision was made for monitoring the power output, but a simple output control was included. In order to make the instrument as generally useful as possible, internal circuits were provided for amplitude modulating the carrier with a 1000-cycle sine wave or a fairly long pulse.

R-F CIRCUIT

Since simplicity was a prime factor in the design of the instrument, a very elementary r-f circuit was chosen. Basically it consisted of a plate-modulated oscillator whose output was fed to an output connector or an antenna through a coaxial cable. The coupling between cable and oscillator was accomplished by means of a rotatable loop which served as the output control.

In the frequency range covered by the instrument, a butterfly circuit[1,2,3] has several characteristics which make it a logical choice for use as the oscillator resonator. Some of these characteristics are:

1. A butterfly is capable of tuning over very wide frequency ranges at relatively high frequencies, thus minimizing the complexity of band switching problems.

2. With a butterfly of the proper design, wide-range oscillators usually can be made to operate up to about 80% of the natural resonant frequency of the oscillator tube.

3. The frequency of oscillation is varied simply by a single control.

4. A butterfly circuit has no sliding contacts carrying large r-f currents.

5. The unit is small and compact.

In order to keep the size of the oscillator at a minimum and to simplify the installation of a band changing switch, a semi-butterfly of the type shown in Figure 2 was selected.

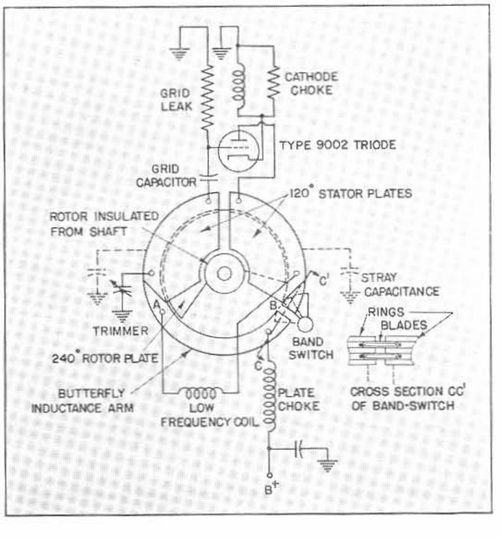
A butterfly resonator, which is normally a two-terminal network, is well suited for use with a triode tube in a

[1] E. Karplus, "The Butterfly Circuit," *General Radio Experimenter*, Vol. XIX, No. 5, October, 1944.

[2] E. Karplus, "Wide-Range Tuned Circuits and Oscillators for High Frequencies," *Proceedings of the I.R.E.*, Vol. 33, No. 7, July, 1945, pp. 426-441.

[3] U. S. Patent No. 2,367,681.

Figure 2. Diagram of the r-f oscillator showing general construction of the butterfly and the band switch.



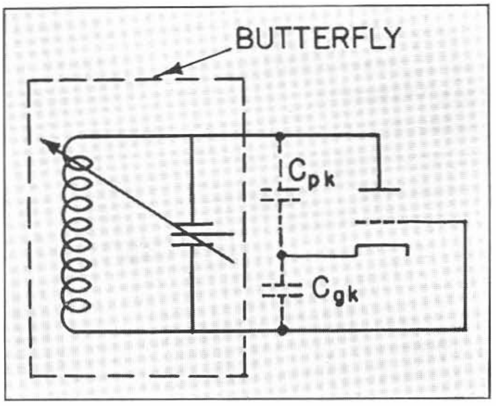


Figure 3. Basic r-f oscillator circuit.

modified Colpitts oscillator circuit in which the interelectrode tube capacitances form the feedback circuit. A basic schematic of the circuit is shown in Figure 3.

The Type 9002 Miniature Triode was selected as the oscillator tube because it was a preferred type and because of its relatively high natural resonant frequency, small size, and low power requirements. Since continuous coverage is not possible with the butterfly resonator at frequencies higher than about 80% of the natural resonant frequency of the tube, the upper frequency limit was fixed by this choice of tube.

A semi-butterfly resonator is actually a parallel resonant circuit which has a capacitive branch and an inductive branch. From Figure 2 it can be seen that the capacitive branch is made up mainly of the capacitance from one stator section to the rotor in series with the capacitance from the rotor to the other stator section, and that the inductance of the arm between points A and B constitutes the main portion of the inductive branch. As the rotor is turned in the counterclockwise direction from the position shown in the figure, the capacitance from the right-hand stator section to the rotor decreases, thus decreasing the total effective capacitance. The rotor also advances along the inductance arm, partially shielding it magnetically and hence reducing its inductance. As the result of the decrease in inductance with decreasing capacitance, a much wider tuning range is obtained than can be produced by varying the capacitance alone. In this butterfly a 2.25 to 1 change in inductance over the tuning range was obtained.

In spite of the advantage gained from the variation of both L and C in a butterfly, it was found to be impractical to cover the entire band from 40 to 500 Mc in one step, and it was necessary to break it into two bands. The high-frequency band from 115 to 500 Mc was covered with the resonator acting as a butterfly as described above. However, for the 40 to 115 Mc band, the butterfly inductance arm was open-circuited by means of the band switch which placed a low-frequency coil across the capacitance sections of the butterfly as shown in Figure 2. In the low-frequency band, the inductance remained fixed and the tuning was accomplished by means of the variation in capacitance alone.

The design of the band switch was important, as the inductance it introduced in its closed position had to be small compared to the minimum inductance of the inductive arm of the butterfly, or the tuning range would have been appreciably reduced. The losses it introduced in the circuit also had to be small to avoid reducing Q below the value required to sustain oscillation. Therefore, it was constructed of a multiple set of blades which interleaved the rings forming the inductance arm as shown in Figures 2 and 4. Each blade consisted of two spring leaves which were compressed when the blade was between the rings and made a rigid, low-loss, low-inductance connection.

In the type of Colpitts circuit used, the whole oscillator circuit, including the cathode of the tube, is floating with respect to ground and connections were made to the cathode, heaters, and plate through chokes. However, a choke is not an infinite impedance, and it was found that a "hole" in the oscillations occurred over a narrow frequency band, which appeared to be caused by a resonance in the circuit consisting of the cathode lead inductance, the choke impedance, and the stray capacitances of the stator sections to the shield. Although the frequency at which the hole appeared could be shifted by changing the reactance of the choke, a practical choke design was not found which would shift it out of the desired range. The hole was eliminated by connecting a resistor across one of the chokes as shown in Figure 2, to reduce the Q of the undesired resonance.

At the upper end of the frequency range, the operating frequency is largely dependent on the tube capacitances. Therefore, in order to compensate for changes in tube capacitance when the oscillator tube is replaced, a small adjustable trimmer capacitor was connected from one stator section to ground.

Mechanically, the butterfly stator was mounted on three insulating posts as shown in Figure 4, and was covered by an insulating plate that performed the dual function of supporting the oscillator tube socket and of clamping and aligning the two stator sections of the butterfly. The insulated rotor shaft turned in a set of ball bearings mounted in a casting below the butterfly. However, it was found that quarter-wave resonance occurred in the shaft circuit near the upper end of the tuning range, and it was necessary to ground the upper end of the metal shaft with a grounding strap as shown in Figure 4 in order to shift the spurious resonances out of the operating range.

The orientation of the output coupling loop was adjusted by a knob on the panel and was arranged to have its minimum and maximum coupling to both the butterfly inductance arm and the low frequency coil at approximately the same position. The power from the coupling loop was fed through a coaxial

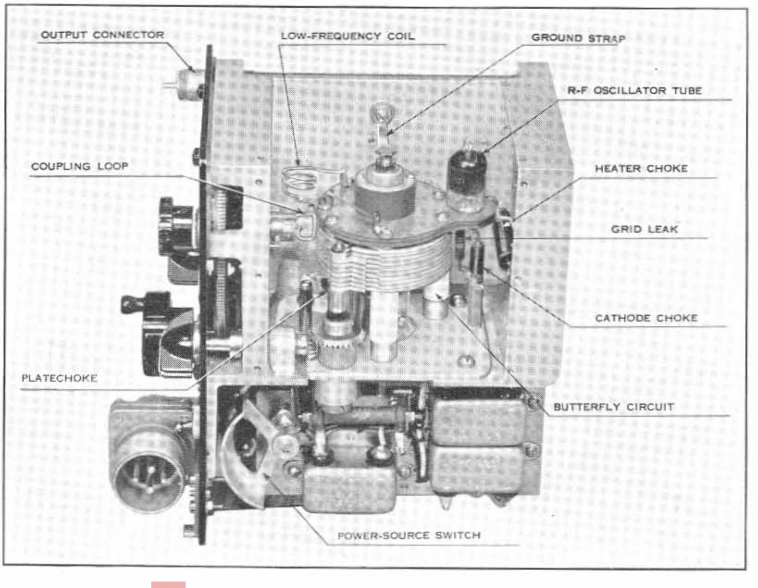


Figure 4. Interior view of the instrument with shields removed.

transmission line to the modified type N connector located on the panel. This connector was constructed in a novel manner as it could be used as a standard connector or an antenna. For use as an antenna, the center conductor was pulled out to any desired length up to 6 inches, as illustrated in Figure 1. The center conductor was formed of a tightly coiled spring and hence was not damaged by bending.

MODULATOR AND POWER SUPPLY

The modulator was somewhat unconventional, consisting of a single tube which could be made to produce sine wave or pulse modulation. For sine wave modulation, a conventional Hartley circuit was used with the primary of the modulation transformer acting as the tuning inductor in the resonant circuit. For pulse modulation, the transformer was connected to form an oscillating circuit with an abnormally large amount of feedback and a natural frequency of about 7 kc, determined by stray capacitances. The grid resistance was so chosen that only one cycle of 7 kc oscillation occurred before the grid blocked. The first half cycle, which was positive, plate-pulsed the r-f oscillator. Although a long pulse was produced by this modulator, it was found to be adequate for many field testing purposes.

For maximum utility in its intended application, the instrument was designed to operate at all of the supply voltages and frequencies then in use by the Allied military services: 80, 115, and 230 volts, at frequencies between 50 and 2600 cycles. Provision was also made for operation from external batteries. A selector switch made the proper connections for the various types of power sources, and the type of power required was indicated on a drum attached to the switch which was visible through an opening in the panel.

MECHANICAL FEATURES

In order to facilitate quantity production and to simplify maintenance, unit construction was used throughout. The whole oscillator assembly was made one removable unit and the power supply and modulator section was another unit. Connections were made between units by means of jumper wires connected between adjacent terminal strips.

Other mechanical features contributed materially to the utility and acceptability of the instrument. Among these was an edge-illuminated frequency dial which could be read regardless of external lighting conditions.

The unit was adapted for mounting in a standard ATR rack, and the whole assembly was shock mounted within its carrying case, a feature that made it possible for the instrument to withstand without damage the severe conditions of military use. —R. A. Soderman

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THE General Radio EXPERIMENTER

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

VOLUME XXI No. 7 DECEMBER, 1946

DECADE VOLTAGE DIVIDER NOW AVAILABLE

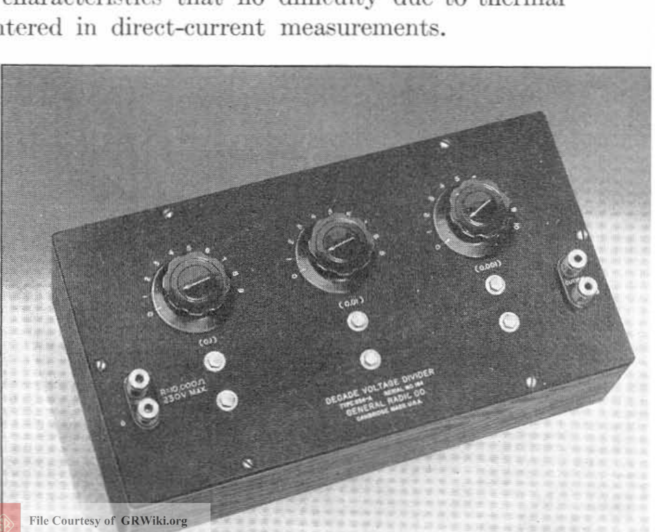
Also

IN THIS ISSUE

● **THE TYPE** 654-A Voltage Divider, always a useful laboratory device, was discontinued during the war in order to concentrate production facilities on more urgently needed items. We have had a number of requests for this item and are glad to announce that it is again available from stock.

As shown in the accompanying photograph, this voltage divider has three dials, giving division factors of 0.1, 0.01, and 0.001, respectively. Voltage ratios between 0.001 and 1.000 can be obtained in steps of 0.001 with an accuracy of $\pm 0.2\%$. The division is accomplished by the equivalent of two three-dial decade resistance boxes, so connected that when resistance is taken from one box it is added to the other to maintain the total resistance constant at 10,000 ohms. This action is accomplished through the use of two Type 510 Decade-Resistance Units operated from each control knob by means of a chain drive. All resistors are wound with an alloy wire of such characteristics that no difficulty due to thermal emf will be encountered in direct-current measurements.

The Type 654-A Voltage Divider is currently priced at $100.00 plus 10%. Complete specifications will be sent on request.



Panel view of the decade voltage divider.

THOSE IRON-CORED COILS AGAIN

Since publication of the article on iron-cored coils in the *Experimenter* a few years ago[1], considerable use has been made during the intervening years of the theory developed therein. It is not surprising that there have emerged, during that period, a number of ways in which the use and application of the theory may be simplified.

PART I

INTERCHANGEABILITY OF PERMEABILITY AND FREQUENCY

One of the more significant improvements has to do with the method of securing empirically the information needed for a particular lamination structure. This needed information consists of the maximum storage factor Q of a full coil wound on the lamination structure, the frequency at which it occurs, and the law of variation of inductance with center-leg air gap.

Theory

If reference be made to page 9 in the Appendix portion of the March, 1942, article, it will be seen that for a given structure D_c varies inversely while D_e varies directly as the product $f\mu$. Hence, if frequency is held constant, dissipation factors D_c (copper loss) and D_e (eddy-current) are functions solely of effective permeability μ. Thus,

$$D_c = \frac{c'}{\mu} \quad \text{and} \quad D_e = e'\mu \qquad (24)\ \&\ (25)$$

where,

$$c' = \frac{10^9 \rho_c tl}{8\pi^2 f S A\alpha} \quad \text{and} \quad e' = \frac{2\pi^2\delta^2 f}{3\rho_i 10^9} \qquad (26)\ \&\ (27)$$

(Measurements are postulated at very small B, where hysteresis loss is negligible.) Consequently, the same shape of Q curve will be obtained at constant frequency and varying permeability as with constant permeability and varying frequency.

The Q of the coil will vary from a low value with the core interleaved up through a maximum value at some intermediate air gap and down again to a low value at a large air gap.

[1]P. K. McElroy and R. F. Field, "How Good is an Iron-Cored Coil?" *General Radio Experimenter*, March, 1942.

Technique

It will be necessary to choose an appropriate frequency at which to make the measurements, in order to be sure that the peak occurs somewhere near the geometric center of the inductance range covered. While experience will usually provide a guide to the selection of the measuring frequency, a more reliable one is provided by the expression for f_m and a few measurements on various types of cores.

The frequency at which the Q of a coil is a maximum is given by the expression

$$f_m = \frac{10^9}{4\pi^2\mu\delta}\sqrt{\frac{3\rho_c\rho_i tl}{SA\alpha}} \qquad (23)$$

These frequencies are made lower as the size of the core is increased, as the laminations are made thicker, as the permeability of the core is increased, or as the resistivity of the core is decreased. Both of the last two reasons, for instance, operate to make an A-metal core require a lower measuring frequency than a silicon-steel one.

If the frequency chosen were that corresponding to Q_m for an interleaved core, the measured points would all fall at and on one side (the left) of the maximum, while for the frequency corresponding to no iron (air core) the distribution of points would be on the other side of the maximum. A convenient frequency is the geometric mean of the two values of f_m corresponding to Q_m for the interleaved core and Q_m for the air core. This frequency corresponds to a μ which is the geometric mean between that of the interleaved material and the effective value for the air core.[2]

[2]This value is greater than unity for at least two reasons: (1) The turns are not concentrated in a single layer, but fill the whole window, thus yielding a larger inductance; and (2) the flux path external to the center leg is a much greater fraction of the total.

For example, measurements of a number of coils with cores of 4% silicon steel indicate that the ratio of inductance, and hence equivalent permeability, between air core and full-interleaved core is about 65. For an initial permeability of the core material of 470, the equivalent air core permeability is then about 7. Using the geometric mean of 60 in Equation 23 gives the optimum frequency for taking data for the humped Q curve. For A-metal, the permeability values are 2500 and 7, giving a geometric mean of about 130.

The following table indicates the optimum measuring frequencies for gathering data for humped curves as functions of the lamination size and material (in the order of size). The dimensions of the four standard GR lamination shapes here discussed appear in Figure 1.

Lamination	4% Silicon Steel	A-Metal
746	1805 cps	782 cps
345	823	439
485	612	327
365	462	—

These figures can be compared with the notes following shortly below and with the distribution of plotted points on the curves in Figures 2 to 11.

In the examples illustrated by those figures, various combinations of lamination shape and material appear. It should be noted, in connection with all the curves mentioned here, that a point representing measurements on an air-core coil (ferro-magnetic material completely removed) generally lies right on the curve which goes through the points representing conditions where there is some ferromagnetic material in the circuit. These points are identified on the plots by an adjacent letter "A." A letter "B" on each plot will be placed adjacent to the point representing a butt joint in the center leg (interleaved outside legs), and a letter "I" beside the point representing the completely interleaved condition.

Some notes correlating the symmetry of the humped curves with the measuring frequency are given in the captions. Unless otherwise stated, fair symmetry was achieved.

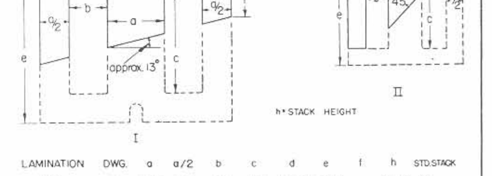
Plotting the Curves

The plots are all made with values of Q as ordinates, and values of inductance as abscissae. Since the only information to be extracted from the humped curves is the maximum Q reached, there is no necessity to reduce the data so that plotting can be done against the effective permeability to which the inductance is, by definition, strictly proportional.

Each of Figures 2 to 8 will be seen to carry two additional curves. The second curve, which slopes downward from left to right, is obtained (from the original data) by plotting measured center-leg air gap in mils against measured inductance (at the initial permeability level)[3]. These curves would have a constant downward 45° slope (inductance inverse with air gap), were it not for the effects of fringing, of the reluctance of the magnetic material, and of the equivalent reluctance of a butt joint.

The third curves, sloping upward to the right, are derived from the second ones. For them, the ordinates are the same as for the second ones, namely, measured air gap. The abscissae, however, are frequencies in cycles according to the scale at the top of the figure. These frequencies are the frequencies at which the maximum Q occurs for various air gaps. This curve can be used to decide what air gap will give the best compromise between stability of inductance against voltage changes and high Q at the desired frequency. When the air gap has been decided upon, the frequency at which the maximum Q will occur can be read from the curve.

[3]No points for air-core, interleaved, or butt joints are included, since it is impossible to assign specific values of air gap to these conditions on the log-log chart.



LAMINATION	DWG.	a	a/2	b	c	d	e	f	h	STD. STACK
746	II	.344	.172	.250	.750	1-3/16	1-3/32	—	23/32	2:1
345	I	.750	.375	.531	1-1/8	2.5625	1-7/8	5/8	3/4	SQUARE
485	I	15/16	15/32	21/32	1-19/32	3-3/16	2-17/32	29/32	15/16	SQUARE
365	I	1-3/16	19/32	25/32	2-1/8	3-15/16	3-5/16	1-3/16	1-3/16	SQUARE

Figure 1. Dimensions of core laminations used in measurements.

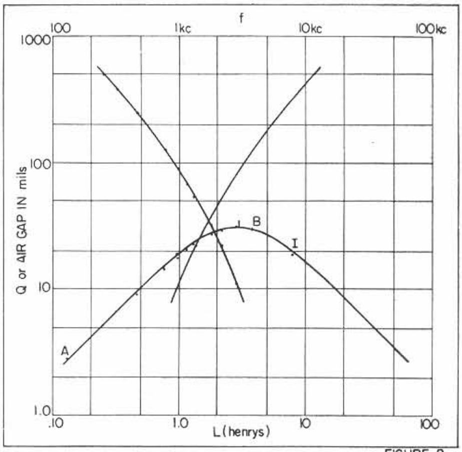


FIGURE 2

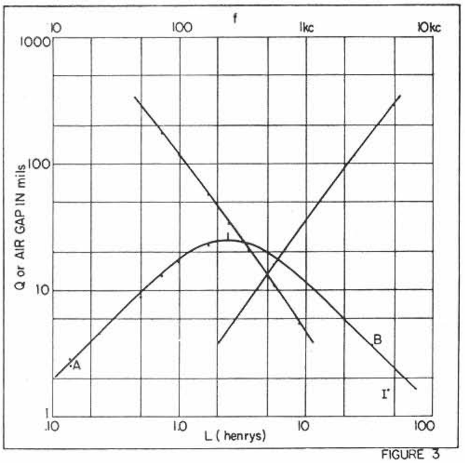


FIGURE 3

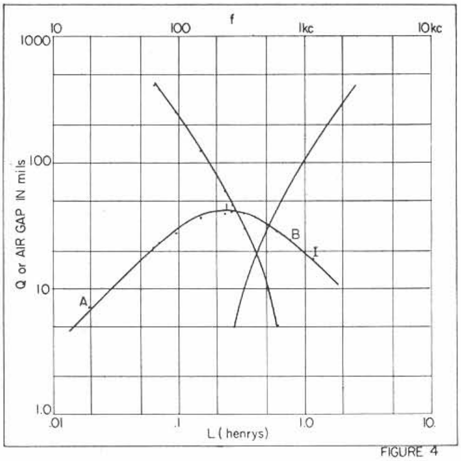


FIGURE 4

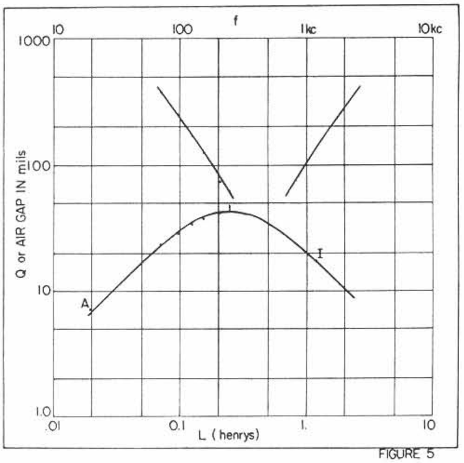


FIGURE 5

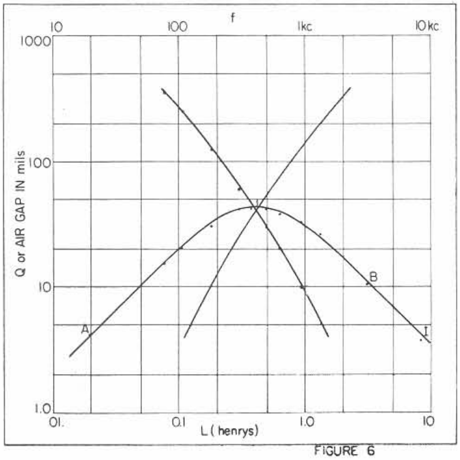


FIGURE 6

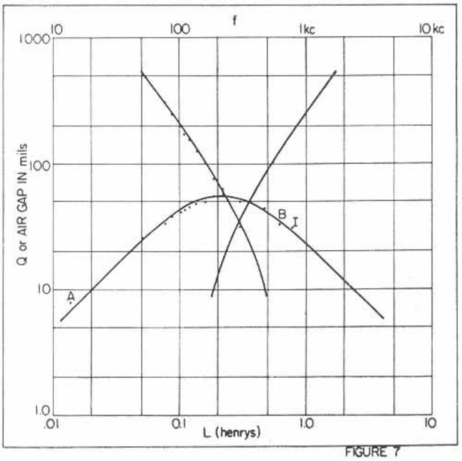


FIGURE 7

If, now, the center point of a humped cardboard template be laid on the chart at the point where a vertical line corresponding to the frequency indicated as above intersects a horizontal line representing the Q_m characteristic of this structure independent of air gap (passing through the peak of the original humped curve), one can read the Q of coils on this structure (having the chosen air gap) at any other frequency.

To derive the third curve from the second, use is made of the relationship that Q_m occurs at a frequency which is inverse with effective permeability, i.e., with inductance. The product, then, of inductance and frequency for Q_m at all points is the same. The value of this product can be obtained by multiplying the frequency of measurement by the inductance at which maximum Q of the humped curve occurs. If Figure 8 is taken as an example, the maximum of the humped curve occurs at 0.245 henries, and the measuring frequency is 400 cycles. The product of the two is 98. The air gap corresponding to 0.245 henries is 68 mils at 400 cycles. To obtain a point on the third curve corresponding to any point on the second curve, divide the abscissa for the second curve into 98, from which will be obtained the abscissa for the third curve. The ordinate is the same for both. The end result is two curves, mirror images of one another.

Figure 2. A coil wound on a small, or postage-stamp, core (GR 746) of 29-gauge 4% silicon-steel laminations with an $\frac{11}{32}''$ wide center leg stacked $\frac{23}{32}''$ high was measured at 1 kc, but the plot was quite lopsided, with most of the points on the low-inductance side of the peak. The frequency should have been higher.

Figure 3. A coil on the same core as that of Figure 2, employing 29-gauge A-metal laminations, was measured at 1 kc, which was nearly the correct frequency.

Figure 4. A coil on a 345 core of 26-gauge 4% silicon-steel laminations, having a $\frac{3}{4}''$ square center leg, was measured at 700 cycles. The laminations had a blunt-angled nose. (See Figure 12.)

Figure 5. The same coil as in Figure 4, but with square-nosed laminations. Unfortunately, they were available only with a center-leg air gap no smaller than $\frac{1}{16}''$. Therefore, there are no points between approximately 0.25 henries (corresponding to $\frac{1}{16}''$ air gap) and 1.3 henries (corresponding to completely interleaved laminations). The two sloping curves are short for the same reason. Within the range where both have points, the curves of Figures 4 and 5 should be and actually are sensibly in the same locations.

Figure 6. A coil on the same size core, but made from 29-gauge A-metal laminations, was measured at 400 cycles.

Figure 7. A coil on a 485 core of 26-gauge 4% silicon-steel laminations, having a $\frac{15}{16}''$ square center leg, was measured at 400 cycles.

Figure 8. A coil on the same size core, but made from 29-gauge A-metal laminations, was measured at 400 cycles.

Figure 9. A coil of 8800 turns of No. 30 enamel on 365 core of 26-gauge 4% silicon-steel laminations, having a $1\frac{3}{16}''$ square center leg, was measured at 400 cycles.

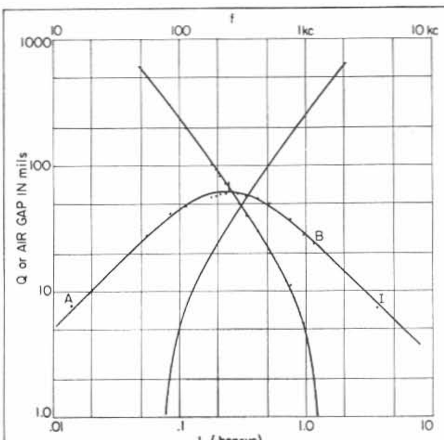


FIGURE 8

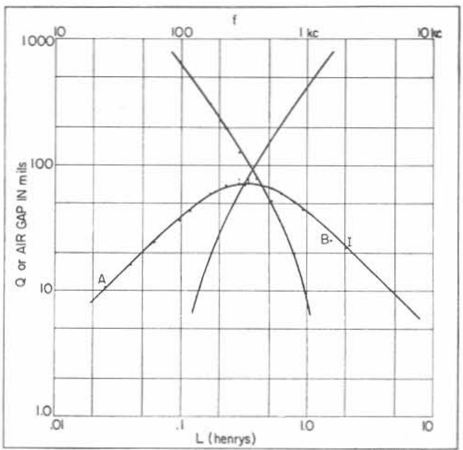


FIGURE 9

Comparison of Graphic Results With Theory

The following equations are repeated from the March, 1942, article:

$$L = \frac{4\pi N^2 \mu A \alpha}{10^9 l} \qquad (6)$$

$$f_m = \frac{10^9}{4\pi^2 \mu \delta}\sqrt{\frac{3\rho_c \rho_i t l}{S A \alpha}} \qquad (23)$$

If they are combined by eliminating μ, the following equation results:

$$L = \frac{N^2}{\pi \delta f_m}\sqrt{3\rho_c \rho_i \frac{t}{S} \cdot \frac{A\alpha}{l}} \qquad (28)$$

This equation yields a figure for the value of inductance at which the top of the humped curve should occur for the particular measuring frequency employed. This is compared in the following table with the value obtained graphically from each plot.

		L		Q	
lam.	mat.	calc.	act.	calc.	act.
746	SS	2.76	2.95	28.0	30
746	A	2.59	2.43	26.3	25
345	SS	0.219	0.24	40.6	42
345	A	0.442	0.42	46.9	43
485	SS	0.198	0.216	54.9	54
485	A	0.228	0.244	63.4	63
365	SS	28.7	33.5	72.1	71.5

In the same way the following equation

$$Q_m = \frac{1}{\delta}\sqrt{\frac{3\rho_i S A \alpha}{\rho_c t l}} \qquad (22)$$

gives the maximum height which should be reached by the humped Q curve. This likewise is compared with the graphically obtained value in the table.

It will be noted that the agreement between theory and measurement is fairly good when the vagaries of ferromagnetic materials are borne in mind.

Air-core Point

The fact that the point representing measurements on an air-core coil falls on the curve as closely as other points may at first thought seem strange. However, this point represents the extreme limit of low permeability where copper loss is the sole factor in determining Q. For this condition the flux cutting the copper is identical with that for an interleaved core.

In order to demonstrate that this point belongs on the curve, extra points to define the curve were taken for Figure 9. The two points to the right of point A represent air gaps of 1¾ inches and 1⅛ inches, respectively. The third point corresponds to a ⅝-inch gap, the longest normally measured.

Eddy Currents in Copper

It will be noted that a number of points in the middle of the Q curves (that is, for moderate-length air gaps) of Figures 10 and 11 do not fall on the curve. This is attributable to increased eddy-current losses in the copper.

Figure 10. A coil wound with 972 turns of No. 20 enamel, measured at 400 cycles.

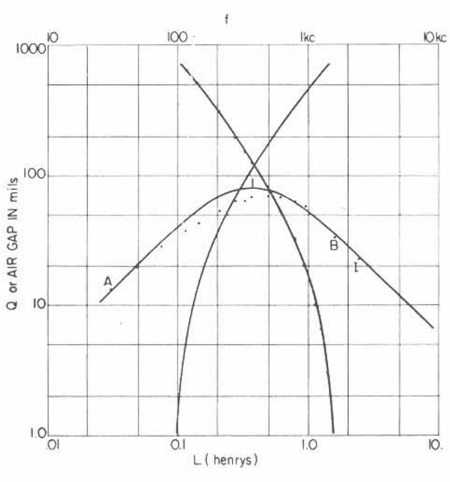
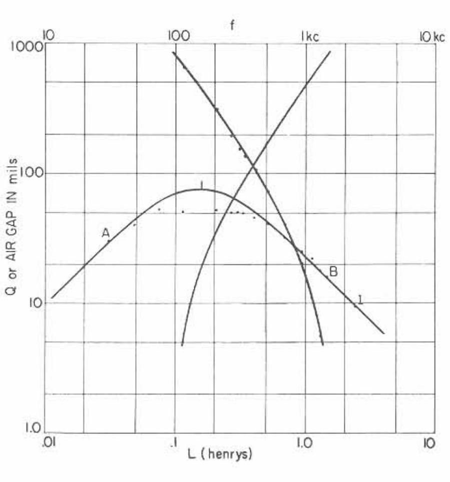


Figure 11. The same coil measured at 1 kilocycle.



The original theory[1] included the assumption that eddy-current losses in the copper were negligible. For moderate air gaps, however, this assumption no longer is valid. Unless the air gap is quite small, nearly all the magnetomotive force is concentrated across the gap. This results in excessive fringing at the gap, and the lines of fringing flux cut the copper, inducing eddy currents. The resultant increased loss in the copper reduces the Q below the ideal value to be expected from the shape of the curves. With larger gaps, approaching the length of the center leg, the leakage flux field has expanded considerably, and much of it goes out around the copper. Meanwhile, the ohmic losses in the coil have gone up considerably, owing to the increased current, and are far larger than the remaining eddy-current losses.

From the winding data given in the captions to Figures 9, 10, and 11, it will be seen that the loss increases with increasing wire size and with increasing frequency, as eddy-current loss would be expected to do.

As was done in the case of Figure 9, two extra points were taken, those at the right of point A, to see whether this analysis was correct and the points would gradually approach the humped curve. They did so very nicely on Figure 10 and pretty well on Figure 11, although the right-hand one of the two was somewhat of a sport.

Eddy Currents in Iron

There is one further phenomenon indicated by the plotted points. In Figure 3 the point marked "I" is way off the humped curve. A ready explanation comes to hand. This figure represents data taken on a core of A-metal laminations. These laminations had no insulating coating to prevent interlamination eddy currents in the way that the scale on silicon steel acts. It is believed, therefore, that there were at the time of measurement excessive interlamination eddy currents. These acted to push the point away from the curve by two effects. First, the extra losses reduced the Q, which pushed the point down. Second, the extra eddy currents reduced the inductance of the coil, thus pushing the point to the left.

Over-all Results

It can now be seen what a considerable amount of information has been made available as a result of a single series of measurements, with varying air gaps in the center leg at a single frequency. The Q_m of the magnetic structure has been determined, from which the Q_m for other alloys and thicknesses of laminations can be easily deduced by methods given in the March, 1942, paper. The second or downward-sloping curve, when taken in conjunction with the known number of turns of the coil used in making the measurements, gives all the information needed to choose the number of turns for a coil of any inductance (at initial permeability) with a given air gap. The third or upward-sloping curve tells what air gap to choose in order to get the maximum Q at any particular frequency. Once a specific design has been determined by use of the second and third curves, its Q behavior at all frequencies can be forecast by using the cardboard template, as suggested in a prior paragraph and more fully covered in the March, 1942, article.

The method of securing information suggested in this article is much simpler than that used in the prior one, since data need be taken for only one instead of several humped Q curves. Lest it be thought that the more complicated method used to secure the information diagrammed in Figure 1 of the March, 1942, article was unwarranted, it should be noted that the simplicity of this present method is dependent on facts adduced therein. If it had not already been demonstrated by actual measurement that the Q_m of a coil occurred at a frequency f_m which varied inversely with the inductance, that is, was higher with larger air gaps, the construction of the third curves from the second ones, purely mathematically, could not have been justified.

[1] Loc. cit.

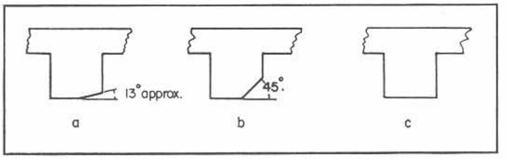


Figure 12. Types of laminations: (a) blunt-angled; (b) sharp-angled; (c) flat.

Example of Use of Complete Theory

An example might be given of a place where the information just described would be needed and used. Suppose it was desired to build a series of high-pass filters having the same impedance and attenuation characteristics but with a series of cut-off frequencies. Figure 13 shows the simple "T" configuration of such a filter. The requirement of similar attenuation characteristics must be interpreted as requiring the same Q for all the coils. By the rules of filter design, the inductance is inverse with the cut-off frequency of the filter. By the rules which have just been developed, the permeability (or inductance) of a coil should be inverse with the frequency at which its Q_m is to occur. Since the Q_m should occur in all filters with the same relation to the cut-off frequency, that is, approximately at cut-off frequency, the conclusion is that the design of this group of filters is quite simple.

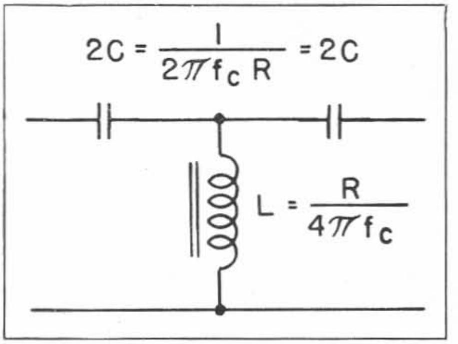


Figure 13.

The same coil winding can be used for all of them (barring unforeseen difficulty). The only differences between coils for filters of different frequencies will be the center-leg air gaps in those coils. The capacitance values will vary inversely with the frequency. The number of different coils required is thus kept at a minimum, although the laminations must be assembled into them differently.

— P. K. McElroy

GLOSSARY

f = frequency of alternating voltage and current.

L = inductance of coil; henrys.

Φ = total r-m-s flux in the iron; maxwells.

μ = incremental permeability (effective) of magnetic circuit.

t = length of average turn; cm.

N = number of turns of wire.

$S = N \cdot \frac{\pi d^2}{4}$ = effective window area (total copper cross section); cm².

δ = lamination thickness; cm.

A = total geometric cross section of magnetic path; cm².

α = stacking factor of iron; dimensionless (ratio of effective area of core material to inside area of coil tube; deficiencies are occasioned by scale, burrs, bent laminations, core-plating, etc.).

l = mean length of flux path; cm.

ρ_c = resistivity of copper; ohm-cm.

ρ_i = resistivity of the lamination material; ohm-cm.

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