the GENERAL® RADIO Experimenter

INDEX

TO

GENERAL RADIO EXPERIMENTER

VOLUMES XIV AND XV

June, 1939 to May, 1941

GENERAL RADIO COMPANY
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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

Also

IN THIS ISSUE

A NEW NULL INSTRUMENT FOR MEASURING HIGH-FREQUENCY IMPEDANCE

● **THE DESIGN** of impedance-measuring equipment, in general, involves two fundamental choices, namely, the selection of an impedance standard, or standards, for comparison with the unknown impedance, and the selection of a method for indicating when a known relation between them is established.

It has been common experience that null methods of comparison yield the highest precision of setting. At commercial and audio frequencies, where there is relatively little difficulty in obtaining adequate impedance standards, bridge methods have therefore found almost

Figure 1. View of the Type 821-A Twin-T Impedance-Measuring Circuit with cover removed. The airplane-luggage type of case is provided with a carrying handle and the instrument is easily portable. Connecting cables and instruction book are mounted in the cover.

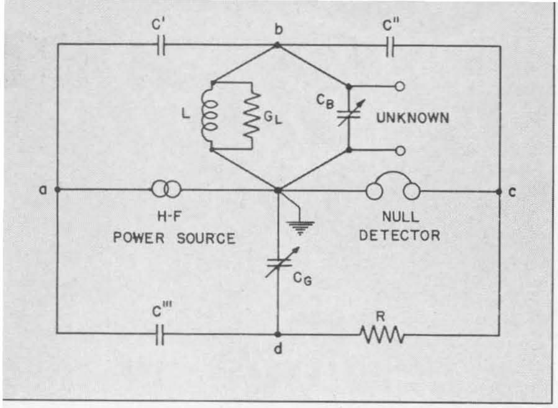

FIGURE 2. Basic circuit diagram of Twin-T impedance-measuring circuit. Losses in the tuning coil, L, are represented by the conductance, G_L, to simplify the balance equations.

universal acceptance. As the frequency is raised, however, residual parameters in the impedance standards and in the wiring cause more and more serious departures from idealized behavior, and, at radio frequencies, it is generally found that bridges designed for low-frequency operation become so inaccurate as to be useless.

Through proper choice of impedance standards and improvement in circuit configurations, the upper frequency limit for accurate bridge measurements has, in recent years, been progressively increased. This process of refinement, however, restricts more and more severely the choice of bridge circuits that can be used and thereby limits the convenience and adaptability that can be obtained.

Another approach to the problem of obtaining suitable null methods at high frequencies can be made by devising entirely different types of circuits rather than by refining existing bridge circuits. The new parallel-T circuits,[1] for instance, are generally adaptable for this service, and the TYPE 821-A Twin-T Impedance-Measuring Circuit, described in this article, uses one that has been found particularly satisfactory.[2]

THEORY OF OPERATION

The basic circuit used is illustrated in Figure 2.

Balance is obtained when the transfer impedances[3] of the two parallel T circuits *a–b–c* and *a–d–c* are made equal and opposite. For this condition the balance equations become:

$$G_L - R\omega^2 C'C''\left(1 + \frac{C_G}{C'''}\right) = 0 \quad (1)$$

$$C_B + C'C''\left(\frac{1}{C'} + \frac{1}{C''} + \frac{1}{C'''}\right) - \frac{1}{\omega^2 L} = 0 \quad (2)$$

If the circuit is initially balanced to a null and then rebalanced by means of the condensers, C_G and C_B, when an unknown admittance, $Y_x = G_x + jB_x$, is connected to the terminals marked UNKNOWN in Figure 2, the unknown conductive and susceptive components can be found from

$$G_x = \frac{R\omega^2 C'C''}{C'''}(C_{G2} - C_{G1}) \quad (1a)$$

$$B_x = \omega(C_{B1} - C_{B2}) \quad (2a)$$

in which C_{G1} and C_{B1} represent capacitance values for initial balance, and C_{G2} and C_{B2} capacitance values for final balance.

ADVANTAGES OF CIRCUIT

Used in this way, the circuit is seen to provide a parallel-substitution measurement of the unknown admittance, with the conductive component proportional to the incremental value of one variable air condenser and the susceptive component proportional to the incremental value of another air condenser. Since each balance is independent of the other, the circuit is well fitted for use in

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[1] W. N. Tuttle, "Bridged-T and Parallel-T Null Circuits for Measurements at Radio Frequencies," Proc. I.R.E., Vol. 28, pp. 23-29, January, 1940.

[2] D. B. Sinclair, "The Twin-T. A New Type of Null Instrument for Measuring Impedance at Frequencies up to 30 Megacycles," Proc. I.R.E., Vol. 28, pp. 310-318, July, 1940.

[3] Defined as the ratio of the input voltage to the output current when the output terminals are short-circuited.

a direct-reading instrument for measuring admittance.

Two features of the circuit that make it particularly useful for measurements at radio frequencies are:

1. There is a common ground point for one side of the generator, one side of the detector, one side of the conductive-balance condenser, C_G, one side of the susceptive-balance condenser, C_B, and one side of the unknown admittance, Y_x. Not only does the common ground eliminate the need for the shielded transformer required in bridge circuits, but it renders innocuous many of the residual circuit capacitances, as can be seen from Figure 2. Capacitances from points *a* and *c* to ground, for instance, fall across the generator and detector where they cause no error. Capacitances from points *b* and *d* to ground fall across the susceptance-balance condenser, C_B, and the conductance-balance condenser, C_G. When substitution measurements are made in terms of capacitance increments they cancel out.

2. The conductive component is measured in terms of a fixed resistor and a variable condenser. This combination, providing the equivalent of a continuously variable resistance standard, has been found much freer from residual parameters than any variable resistor yet devised.

These two features, in themselves, either minimize or eliminate certain unwanted residual parameters. The general circuit arrangement, in addition, disposes of others. Capacitance between points *a* and *b* of Figure 2, for instance, falls across condenser C', capacitance between points *b* and *c* falls across condenser C'', and capacitance between points *a* and *d* falls across condenser C'''. These residual capacitances can all be included as parts of the various capacitances listed and taken into account in the instrument calibration.

DESCRIPTION OF INSTRUMENT

Figure 4 is a panel view of the TYPE 821-A Twin-T Impedance-Measuring Circuit. The controls, shown in the photograph, are:

1. A precision-type variable air condenser used to measure susceptive components and having a dial directly calibrated from 100 to 1100 $\mu\mu$f.

2. An auxiliary condenser, consisting of a bank of fixed condensers controlled

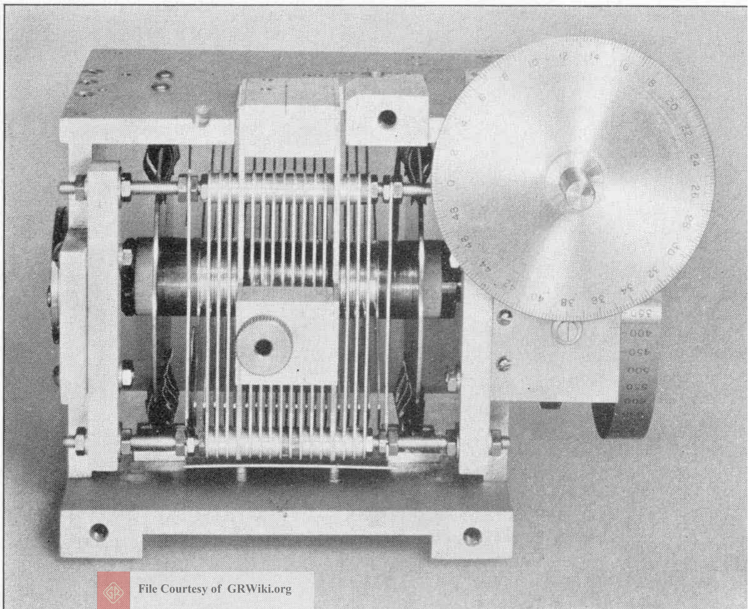

FIGURE 3. View of susceptance condenser C_B showing the two aluminum blocks used to feed from the stator to the internal circuit and to the panel terminal, and brass discs grounding the rotor to the frame through low-inductance brushes.

by push buttons and a small variable condenser, in parallel with the susceptance condenser, used to establish the initial susceptance balance at any chosen setting of the susceptance condenser.

3. A coil switch, marked with the frequency range covered by each tuning coil.

4. A variable air condenser used to measure conductive components and having two scales, one reading from 0 to 100 μmhos and one reading from 0 to 300 μmhos.

5. A 4-position switch used to establish a scale on the conductance dial from 0 to 100 μmhos at 1 Mc, from 0 to 300 μmhos at 3 Mc, from 0 to 1000 μmhos at 10 Mc, and from 0 to 3000 μmhos at 30 Mc. At other than these discrete frequencies, the dial reading must be multiplied by the square of the ratio of the frequency used to the nominal frequency indicated by the 4-position switch.

6. Two small variable condensers, in parallel with the conductance condenser, used as coarse and fine controls to establish the initial conductance balance at zero setting of the conductance condenser.

APPLICATION OF INSTRUMENT

Greatest convenience is obtained with the Twin-T in the measurement of admittances having relatively small conductive components since, for these measurements, the instrument is direct reading. By the use of series fixed condensers, however, admittances having relatively large conductance components can also be measured.

In the first class, namely, admittances having small conductive components, fall condensers, coils, dielectric samples, parallel-tuned circuits, high-resistance units, and antennas and unterminated transmission lines near half-wave resonance. In the second class, namely, admittances having large conductive components, fall series-tuned circuits, terminated transmission lines and matching sections, and antennas and unterminated transmission lines near quarter-wave resonance. Some typical measurements on a few of these devices will serve to indicate the general technique of measurement.

1. Measurement of a 500 μμf condenser at 10 Mc.

Set the 4-position switch at 10 Mc and the coil switch on the 10.0–20.0 Mc

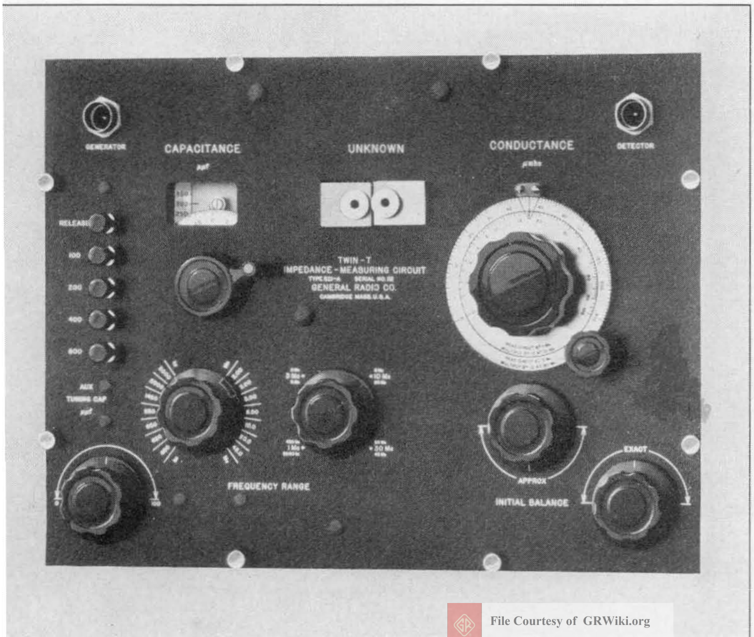

FIGURE 4. Panel view of experimental model of Twin-T impedance-measuring circuit. At the left of the panel are the susceptance condenser (CAPACITANCE μμf) and the auxiliary tuning condenser (AUX. TUNING CAP.). At the right are the conductance condenser (CONDUCTANCE μmho), and the parallel trimmer condensers (INITIAL BALANCE). The remaining controls (FREQ. RANGE) are the coil switch, at the left, and the conductance range switch, at the right.

range. Set the susceptance condenser at some high value, say 1000.0 $\mu\mu$f, and the conductance condenser at zero. By varying the auxiliary condensers in parallel with the susceptance and conductance condensers, adjust to an initial balance.

Connect the condenser to be measured across the UNKNOWN terminals and, with the susceptance and conductance condensers, adjust to a final balance. Let the susceptance condenser setting be 442.4 $\mu\mu$f and the conductance condenser setting be 80 μmho.

Then the unknown parallel capacitance, C_x, and conductance, G_x, are:

$$C_x = 1000.0 - 442.4 = 557.6\ \mu\mu\text{f}$$
$$G_x = 80\ \mu\text{mho}$$

If it is desired to express the condenser loss in terms of dissipation factor, D_x:

$$D_x = \frac{G_x}{\omega C_x} = \frac{80 \times 10^{-6} \times 100}{2\pi \times 10^{7} \times 557.6 \times 10^{-12}} = 0.23\%$$

2. Measurement of 1 μh coil at 25 Mc.

Set the 4-position switch at 30 Mc and the coil switch on the 20.0–45.0 Mc range. Set the susceptance condenser at some low value, say 100.0 $\mu\mu$f, and the conductance dial at zero. Establish the initial balance as described in the previous example.

Connect the coil to be measured across the UNKNOWN terminals and establish the final balance as before. Let the susceptance condenser setting be 139.8 $\mu\mu$f and the conductance condenser setting be 90 μmho.

Then the unknown susceptance, B_x, and conductance, G_x, are:

$$B_x = 2\pi \times 25 \times 10^{6}(100.0 - 139.8) \times 10^{-12} \times 10^{6} = -6250\ \mu\text{mho}$$

$$G_x = 90 \times \left(\frac{25}{30}\right)^2 = 62.5\ \mu\text{mho}$$

The unknown parallel inductance, L_x, and storage factor, Q_x, can easily be found to be:

$$L_x = \frac{10^{6}}{2\pi \times 25 \times 10^{6} \times 6250 \times 10^{-6}} = 1.02\ \mu\text{h}$$

$$Q_x = \frac{6250}{62.5} = 100$$

3. Measurement of matched 72-ohm coaxial line at 830 kc.

Set the 4-position switch at 1 Mc and the coil switch on the 620-850 kc range. Establish an initial balance with the conductance condenser set at zero and the susceptance condenser at some value near mid-scale. Connect the impedance to be measured to the UNKNOWN terminals with a small "postage-stamp" type fixed condenser in series with the ungrounded lead. Change this series condenser to find the largest value for which a balance on the conductance dial can be obtained. Say this is 150 $\mu\mu$f, nominal value.

Leave the ground terminal of the unknown impedance connected to the grounded UNKNOWN terminal. With the fixed condenser connected to the ungrounded UNKNOWN terminal, but free at the far end, establish an initial balance with the conductance condenser set at zero and the susceptance condenser at some relatively high value, say 500 $\mu\mu$f.

Connect the free end of the series condenser, C_a, to the grounded UNKNOWN terminal and rebalance. If there is any appreciable change in conductance balance, rebalance with the zero-adjustment trimmers across the conductance condenser, leaving the conductance dial set at zero. Let the susceptance condenser reading be 352.5 $\mu\mu$f. Then:

$$C_a = 500 - 352.5 = 147.5\ \mu\mu\text{f}$$

$$X_a = \frac{-1}{2\pi \times 830 \times 10^3 \times 147.5 \times 10^{-12}} = -\ 1300 \text{ ohms}$$

Disconnect the far end of the series condenser from the grounded UNKNOWN terminal and connect it to the ungrounded terminal of the unknown impedance. Rebalance with the susceptance and conductance condensers. Let their readings be 353.6 $\mu\mu$f and 60.8 μmho. Then the conductance and susceptance components of the series circuit are:

$$G = \left(\frac{0.83}{1}\right)^2 \times 60.8 = 41.9\ \mu\text{mho}$$

$$B = 2\pi \times 830 \times 10^3 \times (500 - 353.6) \times 10^{-12} \times 10^6 = 764\ \mu\text{mho}$$

From these figures, the resistance and reactance are:

$$R = \frac{41.9 \times 10^{-6}}{(764^2 + 41.9^2)10^{-12}} = 71.6 \text{ ohms}$$

$$X = \frac{-764 \times 10^{-6}}{(764^2 + 41.9^2)10^{-12}} = -\ 1306 \text{ ohms}$$

The reactance of the line itself is found by subtracting the reactance of the series condenser:[4]

$$R_x = R = 71.6 \text{ ohms}$$

$$X_x = -\ 1306 - (-\ 1300) = -\ 6 \text{ ohms}$$

— D. B. SINCLAIR

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[4] The possibility of making substantial errors in reactance through taking the difference between two large numbers can be avoided by assuming that the conductance, G, is negligible compared with the susceptance, B, and taking the difference between the reactances corresponding to 147.5 $\mu\mu$f and 146.4 $\mu\mu$f. This gives a rough check figure of

$$X = \frac{10^{12}}{2\pi \times 830 \times 10^3}\left(\frac{146.4 - 147.5}{146.4 \times 147.5}\right) = -10 \text{ ohms}$$

SPECIFICATIONS

Frequency Range: 420 kc to 30 Mc.

Capacitance Range: 100 to 1100 $\mu\mu$f on standard condenser, direct reading.

Conductance Range:

$$\left.\begin{array}{l} 0 - 100\ \mu\text{mho at } 1 \text{ Mc} \\ 0 - 300\ \mu\text{mho at } 3 \text{ Mc} \\ 0 - 1000\ \mu\text{mho at } 10 \text{ Mc} \\ 0 - 3000\ \mu\text{mho at } 30 \text{ Mc} \end{array}\right\} \text{Direct Reading}$$

Between these direct-reading ranges the range of the conductance dial varies as the square of the frequency.

Accuracy: ± 1 $\mu\mu$f ± 0.1% for capacitance. For conductance, ± 0.1% of full scale ± 2% of actual dial reading.

Accessories Supplied: Coaxial cables for connections to generator and detector.

Accessories Required: A suitable radio-frequency generator and detector are required. Either TYPE 684-A Modulated Oscillator (with the addition of a coaxial output jack) or TYPE 605-A Standard-Signal Generator is a satisfactory generator. A well shielded radio receiver covering the desired frequency range is recommended for the detector.

Mounting: The instrument is mounted in an airplane-luggage type of case with carrying handle and removable cover.

Dimensions: 17¾ x 12 x 9½ inches, over-all.

Net Weight: 26 pounds.

Type		*Code Word*	*Price*
821-A		LAGER	$340.00

NOTES ON THE CARE AND MAINTENANCE OF VARIACS

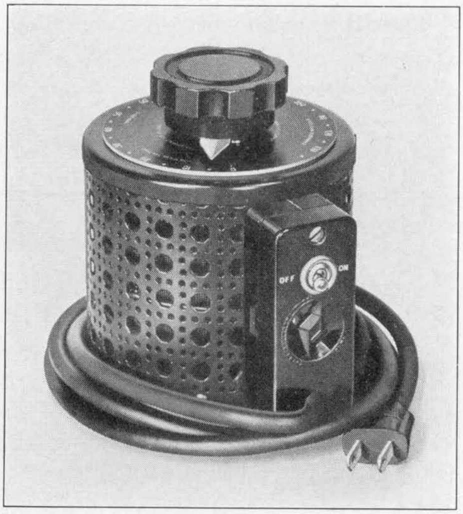
● **MUCH INTEREST** has been shown in the recent article in the General Radio *Experimenter* which outlined a general maintenance and service program for General Radio instruments. A number of requests for maintenance notes on particular instruments have now been received, and, because of the fact that over 35,000 Variacs are in use, it is believed that many customers would welcome specific instructions for the maintenance of these controls.

Inspection of the Variacs returned for repairs shows such conditions as worn brushes and damaged windings to be most prevalent. Careful maintenance would have prevented most of this damage, with the exception of that caused by overloading.

The brushes should be inspected regularly to make certain that excessive wear has not taken place. The interval between inspections may be determined by the frequency at which the Variac voltage settings are changed. When the brush is worn the brass holder may come in contact with the winding surface and cause immediate fusing of the turns short-circuited by the holder. A worn brush may also cause arcing to the winding, and the resultant roughened areas on the contact surfaces will cause further wearing and damage to both the brush and the winding. It is recommended that a small stock of replacement brushes be ordered as part of the maintenance procedure.

Many Variacs are operated in locations where there is considerable dirt, dust, and grit, and even corrosive fumes. Such Variacs require frequent cleaning of the winding in order to insure positive contact between the brush and the winding. If this is not done, erratic operation may result, due to arcing and lack of proper contact, so that eventually the winding must be replaced.

When the surface on which the brush bears becomes blackened or corroded, it should be cleaned with crocus cloth or a very fine sandpaper, making certain that all rough places are smoothed. Remove the loose particles with a fine brush and then clean with carbon tetrachloride or some similar highly volatile cleaning agent.



Excessive heating usually will be caused from too much current flowing in the load circuit. The portion of the winding affected depends in most instances upon the position of the brush. While the winding may not be damaged if the overload is removed quickly, the carbon of the brush may disintegrate. A new brush should be installed before the Variac is again placed in service.

Overheating the winding will cause the turns to loosen and, in cooling, they may not return to their original positions. A raised turn may cause a brush to wear excessively or even break.

The instructions that are included in every Variac shipment state that, when a Variac is used to control the voltage in the primary of a high-voltage transformer or other highly inductive load, it is necessary that either the voltage setting of the Variac be reduced to zero or the output circuit opened before the line circuit is broken. If neither of these procedures is followed, it is possible that a surge will cause serious damage to the winding, although each Variac is tested with 2000 volts d-c between the winding and the frame.

Since the Variac is an auto-transformer, it should never be connected to a load circuit containing a ground. The only exception is when one side of the line and one side of the load are both grounded; these may be connected to the common input-output terminal of the Variac.

Adequate fusing in both line and load circuits is recommended. Replacement fuses are considerably cheaper than replacement Variacs. — H. H. Dawes

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MISCELLANY

● **AT THE ANNUAL BANQUET** of the Emporium Section, I.R.E., H.B. Richmond, Treasurer of the General Radio Company, was the guest speaker. His subject was "Observations of an Engineer on the Continent and in the Near East on the Threshold of War."

● **THE FOLLOWING ERRORS** occurred in the December issue of the *Experimenter*.

In the caption to Figure 3, R_o at 10 Mc should be 0.13 Ω.

In the caption to Figure 9, the values for R_o should be

0.004 Ω at 2 Mc
0.007 Ω at 5 Mc
0.01 Ω at 10 Mc

In the description of Type 318-C Dial Plate, it was stated that the scale progresses in a counterclockwise direction. As the photograph clearly shows, the progression is clockwise. In addition, the photograph should show a nickel silver border around the dial, identical with that on Type 318-B.

● **EFFECTIVE JANUARY 1,** the price of the replacement Type 631-P1 Strobotron for use in Type 631-B Strobotac is reduced to $4.50, net, f.o.b. Cambridge.

● **RECENT VISITORS** to the General Radio laboratories include Mr. W. J. Kroeger of the Frankford Arsenal, Mr. G. Forrest Drake, Chief Engineer of Woodward Governor Company, Dr. Frederick E. Termon, President of the Institute of Radio Engineers, and Messrs. W. R. Knotts, R. Howell, and F. R. Flansburg of RCA.

The General Radio EXPERIMENTER is mailed without charge each month to engineers, scientists, technicians, and others interested in communication-frequency measurement and control problems. When sending requests for subscriptions and address-change notices, please supply the following information: name, company name, company address, type of business company is engaged in, and title or position of individual.

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BRANCH ENGINEERING OFFICES
90 WEST STREET, NEW YORK CITY
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THE General Radio EXPERIMENTER

VOLUME XV No. 8 FEBRUARY, 1941

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

Also

IN THIS ISSUE

AN IMPROVED ULTRA-HIGH-FREQUENCY SIGNAL GENERATOR

● **THE NEED CONTINUES TO INCREASE** in communication and allied research for an accurately known source of radio-frequency voltages in the frequency range extending upward from 10 megacycles. In the General Radio *Experimenter* for November, 1939, the Type 804-A U-H-F Signal Generator was announced. This instrument furnished a test signal at frequencies up to 330 Mc. The new Type 804-B U-H-F Signal Generator employs the same basic design as the older instrument, but incorporates a number of refinements and improvements.

To meet the requirements of most users, a signal generator must

Figure 1. Panel view of the Type 804-B U-H-F Signal Generator. The knob in the center controls the band-change switch, and the dial below it, the tuning condenser. The output control is at the lower right.

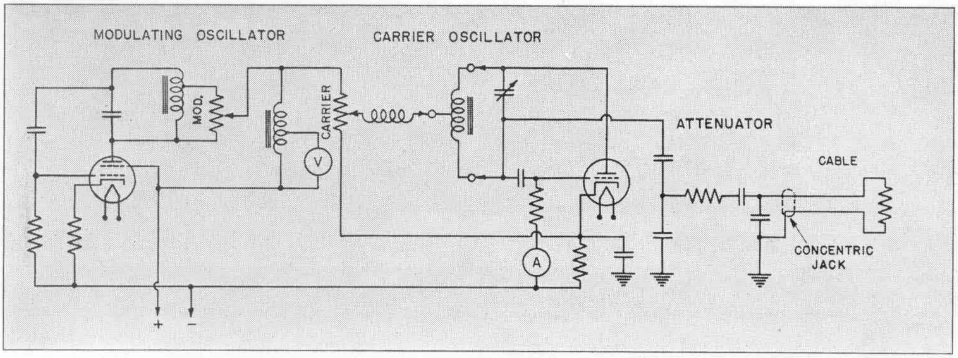


Figure 2. Schematic circuit diagram of Type 804-B U-H-F Signal Generator. The a-c power supply is not shown.

combine a wide frequency range with the ability to be set easily and accurately. The frequency range of the Type 804-B U-H-F Signal Generator extends from 7.5 to 330 megacycles in five overlapping ranges. The five corresponding scales on the main frequency dial are direct reading. The desired range is selected at will by an ingenious coil-switching mechanism, which brings the proper coil into position immediately above the variable tuning condenser and associated 955-type oscillator tube. The range selector is designed to bring the proper direct-reading frequency scale into view through the window in the protective dial housing. The scales not being used are masked, thus doing away with the confusion of multiple scales simultaneously in view. Each frequency range is carefully aligned with the direct-reading frequency scale so that the frequency can be set to well within 2% of the desired value.

The five coils for the five overlapping frequency ranges are fastened to a mycalex disc, which is rotated by the range selector. Mycalex is used because it has very low loss at radio frequencies. At each coil position, silver contacts on the disc engage silver brushes on the condenser frame, as the particular coil is moved into operating position. Lead length between the coil and the condenser is a minimum. A sixth position is provided on the range switch with a blank plug-in form, which can be wound by the user to cover frequency bands in or below the normal range of the signal generator, or to provide band spread over a limited frequency range.

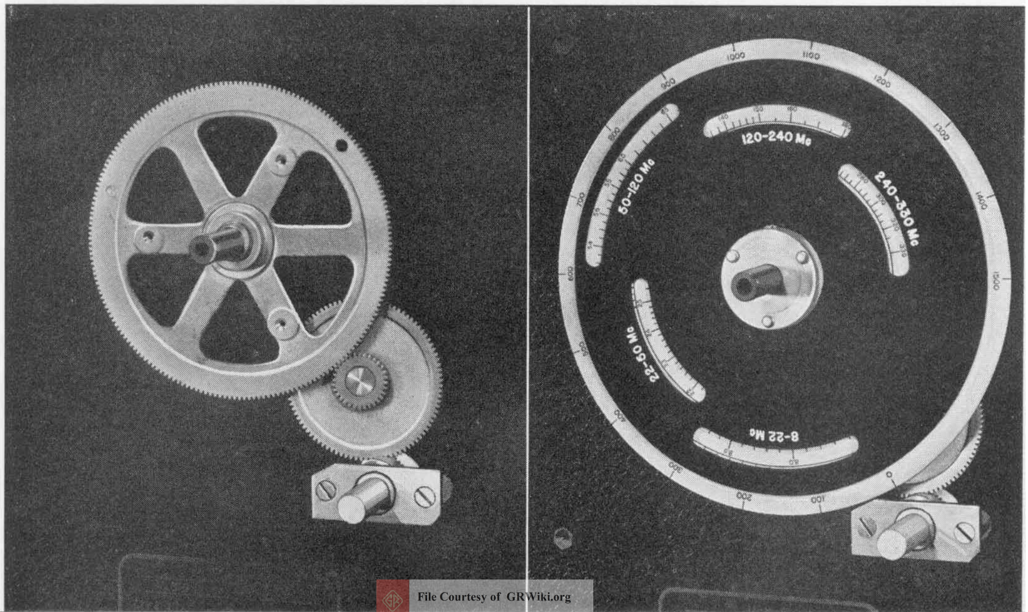
Over each range, the frequency is controlled by a worm-driven variable air condenser. The vernier dial has 100 divisions and requires 15 turns to cover each frequency range. The tuning condenser therefore has 1500 scale divisions for each range, making possible a precision of setting to better than 0.1%. The carrier oscillator uses a Hartley circuit in conventional arrangement.

Radio-frequency leakage in the Type 804-B U-H-F Signal Generator has been reduced to a minimum and is not noticeable with receivers that are now available. The design of the new instrument is such that there are no openings or windows in the panel or cabinet to allow a small amount of radio frequency field to "spray" out. Complete shielding of the two cases of the panel voltmeters has effectively prevented leakage through the meters. Direct-reading, protected frequency scales have been obtained without allowing appreciable leakage to occur.

The output voltage is continuously variable from 1 microvolt to 20 millivolts up to 100 megacycles. Above this frequency, the maximum output is somewhat less than 20 millivolts. The attenuator dial is calibrated in microvolts and millivolts, and is direct reading when the oscillator amplitude is adjusted to a standard value as indicated by the carrier-level grid-current meter. The attenuator dial controls a capacitive attenuator. A step-down gear reduction is provided for ease in adjusting to the desired output voltage. Adjusting screws, set at the factory, allow the output voltage to conform accurately to the dial calibration over the entire scale of the attenuator dial. The capacitance of the attenuator is variable between 0 and 15 $\mu\mu$f in each of two sections, forming a capacitive voltage divider. Regardless of setting, the attenuator presents a constant capacitance to the oscillator circuit so that the carrier frequency does not change with adjustment of the attenuator control.

The voltage from the attenuator is impressed across a 100 $\mu\mu$f condenser. The output voltage is obtained from this condenser through a series resistor of 75 ohms, and is made available at the concentric shielded jack at the right-hand side of the panel. A three-foot concentric shielded output lead, having a characteristic impedance of 75 ohms and fitting the shielded jack, is furnished with the signal generator. The calibration of the attenuator dial is the actual open-circuit voltage at the output jack. Above 30 megacycles the voltage at the far end of the output cable is very nearly equal to that at the output jack. Below 30 megacycles a correction factor for the voltage at the far end of the cable must be applied. A plot of this correction is included in the instruction book. Above 30 megacycles a 10:1 shielded attenuator can be used with the output cable to

FIGURE 3. Two views of the frequency-control drive mechanism partially disassembled. In the left-hand view are shown the gears through which the direct-reading frequency dial is driven from the worm shaft of the condenser. The dial attaches directly to the large gear. At the right, the dial and mask are shown assembled. The bakelite shaft concentric with the large gear drives both the coil switch and the mask for the frequency scales.



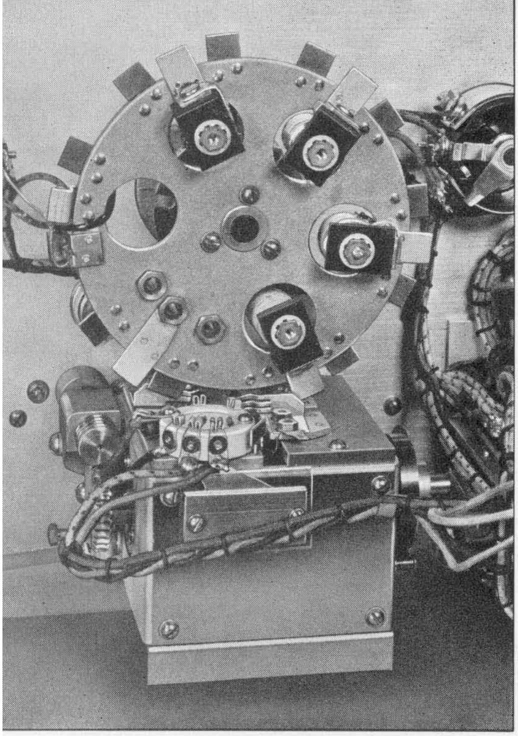


FIGURE 4. Rear view of the coil switching mechanism. A positive wiping contact between silver-plated surfaces is obtained. Leads are made short and direct in order to keep residual inductance and capacitance low.

reduce the indicated output voltage by a factor of ten.

The TYPE 804-B U-H-F Signal Generator will furnish either an amplitude-modulated or an unmodulated output signal. The type of modulation desired is controlled by a three-position panel switch; one position for an unmodulated signal, the second for a 400-cycle modulated signal from an internal oscillator, and the third for modulation from an external source. The modulation is adjustable up to 60%. The percentage-modulation indicator on the panel measures the a-c modulating voltage which is inserted in series with the d-c plate supply of the oscillator tube. As the latter voltage is maintained constant by voltage regulation, the indicating meter can be calibrated directly in percentage modulation, since the modulating voltage is proportional to the percentage of modulation.

When it is desired to employ external modulation, the internal modulation oscillator tube is used as an amplifier. The input impedance for external modulation is 0.25 megohm. Approximately seven volts across the input terminals are required for 50% modulation. The modulation characteristic is flat within ±2 db from 100 to 20,000 cycles per second.

The TYPE 804-B U-H-F Signal Generator operates from the a-c line, 105 to 125 volts, 42 to 60 cycles. A voltage regulator circuit eliminates difficulties due to fluctuating line voltage. Provision is made for altering the connections to the primary of the power transformer so that the instrument can be used on 210 to 250 volt lines. Adequate radio-frequency filtering is provided in the power supply to prevent leakage back into the power line.

— MYRON T. SMITH

SPECIFICATIONS

Carrier Frequency Range: 7.5–330 Mc in five ranges — 7.5–22, 22–50, 50–120, 120–240, 240–330 Mc.

Frequency Calibration: Direct reading within 2%.

Output Voltage Range: 1 microvolt to 20 millivolts for 7.5–100 Mc; 1 microvolt to 10 millivolts for 100–330 Mc.

Output System: Output is obtained across a 75Ω cable.

Modulation: The generator is amplitude modulated. Continuously adjustable 0–60%. Internal: 400 cycles ±5%. External: Flat within 2 db from 100 to 20,000 cycles. Seven volts are required for 50% modulation. The input impedance is 0.25 megohm. Frequency modulation is present, particularly at the higher frequencies. For testing selective receivers, therefore, it is recommended that the generator be used unmodulated.

Stray Fields: Stray fields will not be noticeable with receivers of poorer sensitivity than 1 microvolt.

Power Supply: 105–125 (or 210–250) volts, 40–60 cycles, 25 watts.

Tubes: 955, 6G6G, 6X5G, VR150.

Accessories Supplied: Three-foot output cable, 75-ohm impedance. Six-foot cable for a-c line connection. One blank coil form for additional frequency range. One terminal unit 774-YA-1. One external attenuator 774-X-1.

Mounting: Black crackle aluminum panel, walnut cabinet, hinged cover.

Dimensions: (Length) 19½ x (depth) 9 x (height) 11⅝ inches.

Net Weight: 32 pounds.

Type		*Code Word*	*Price*
804-B	....	DENSE	$350.00

This instrument is licensed under patents of the American Telephone and Telegraph Company solely for utilization in research, investigation, measurement, testing, instruction, and development work in pure and applied science.

TYPE 804-B U-H-F Signal Generator is not in stock at present, and nearly all units now in production have already been reserved for customers. After the few remaining units are sold, deliveries cannot be made before next June.

BROADCAST FREQUENCY REALLOCATION

● **BROADCAST ENGINEERS** are urged to place their orders *now* for new frequency monitor crystals required by the frequency reallocation plan effective March 29.

The TYPE 376-L Quartz Plate is recommended for use in all approved GR monitors. The special price effective until March 29 is $65.00. Send $35.00 with your order, which will be shipped COD for the balance unless credit has been established. If cash for the full amount is sent in advance, we prepay postage.

SERVICE DEPARTMENT NOTES

RETURNING INSTRUMENTS FOR REPAIR

● **WHEN GENERAL RADIO** instruments are returned to the Service Department for repair or reconditioning, the time consumed in handling the job can be held to a minimum if the procedure outlined here is followed.

Before returning an instrument to our factory for any reason whatever, a letter giving *complete* information about its operation should be sent to the Service Department. It is essential that both the type and serial numbers be given so that our records may be checked and the history noted. In many instances, if the instrument is not operating properly, it is possible to diagnose the trouble from the information supplied, and to correct it by furnishing a new part such as a tube, resistor, inductor, or capacitor.

If it is necessary that the instrument be returned to the factory, the Service Department will so advise the owner and will furnish complete shipping instructions.

When an instrument is to be returned for reconditioning and recalibration, which may require a week to ten days to complete, it is sometimes possible to speed the work if a letter is written requesting us to proceed at once upon receipt of the instrument and a confirming purchase order is then sent within a few days. This procedure should be followed only in an emergency, as it is easier for all concerned when packing and shipping instructions are supplied by the Service Department. Because of possible damage from excessive handling in transit, we do not recommend shipment via freight or overland trucking, but prefer railway express or parcel post, depending upon the size and weight of the instrument.

The letter or purchase order authorizing necessary work to an instrument should always be mailed so as to arrive before the shipment. A packing slip referring to the letter or purchase order should be enclosed in the shipment. If this is not done, serious delays will result, as in some cases we would have no way of knowing by whom the instrument was shipped.

In accordance with the procedure of purchasing divisions, some of our customers request a quotation to cover the cost of reconditioning equipment. It is our practice to submit an estimate based on records of previous charges for equipment of the same type and age. This estimate is not a definite quotation, but is the form of minimum and maximum prices.

We have found it necessary to follow this plan rather than to test a returned instrument completely and to quote an exact charge, which would inevitably be in excess of what the customer expected because of the excess laboratory time required. As an example, let us consider the return of a Type 736-A Wave Analyzer about two years old. A charge between $45.00 and $60.00 would be quoted to cover complete reconditioning and recalibration, guaranteed for one year. However, if this instrument were sent to our laboratory for test to determine the extent of necessary reconditioning and recalibration, at least eight hours would be spent before an exact quotation could be made. Following acceptance of the quotation by the customer, repairs would be made in the shop, after which the eight hours' laboratory test time would have to be duplicated. In such instances, costs may be almost 50% greater than that compared with an amount quoted between minimum and maximum prices.

The following table can be used to estimate the approximate charge for an instrument, not obsolete, that requires normal reconditioning and recalibration.

List Price	*Maximum Reconditioning Charge % of List Price*
Up to $100	25
$100 to 200	20
200 to 400	15
400 up	10

The use of this table will often avoid delays by giving the customer's purchasing division an idea of the approximate cost. If, upon inspection of a returned instrument, it is found that the cost of reconditioning will be in excess of normal charges, the customer is advised the maximum cost, and no work is done until a reply is received.

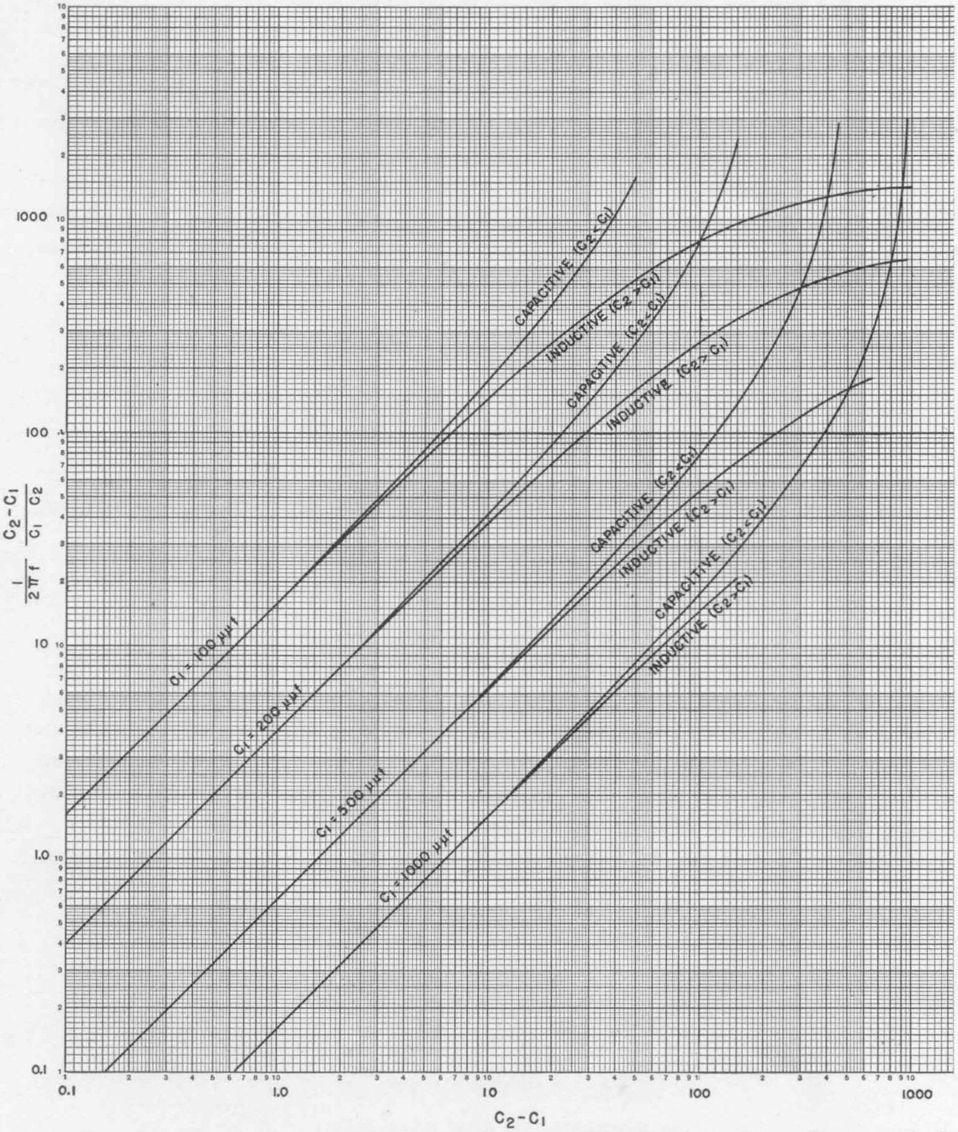
The procedure that is followed in reconditioning a returned instrument is to clean it thoroughly; check and resolder any connections that may have weakened; replace or repair any component part that has become worn, deteriorated, or damaged; tighten all assembly and mounting screws; clean the panel and polish the cabinet.

It is then sent to the laboratory for final test and recalibration. The instrument must pass the same test as a new instrument. If an obsolete type, it is tested under the specifications that were used when it passed through the laboratory originally. Because of the careful and complete reconditioning in our shop and accurate testing in our laboratory, we are able to guarantee an instrument for one year, which guarantee is identical to that which applied to it when it was originally sold.

— H. H. Dawes

A CHART FOR USE WITH THE TYPE 516-C RADIO-FREQUENCY BRIDGE

● **FOR ACCURACY** and flexibility, measurements with the TYPE 516-C Radio-Frequency Bridge are ordinarily made with a series-condenser substitution. A fixed condenser of the appropriate capacitance is first placed across



Plot of change of reactance of bridge condenser at 1 Mc. For the series-condenser method this is equal to the unknown reactance. For use at other frequencies, divide by the frequency in megacycles. If the unknown reactance is inductive, C_2 is greater than C_1; if the unknown reactance is capacitive, C_2 is less than C_1.

the UNKNOWN terminals and the bridge balanced, giving a capacitance reading C_1. The unknown impedance to be measured is then connected in series with the auxiliary fixed condenser and the bridge rebalanced, yielding a capacitance setting C_2.

The reactance of the unknown impedance is calculated from the expression

$$X_X = \frac{1}{2\pi f}\left(\frac{C_2 - C_1}{C_2 C_1}\right)$$

Where many measurements are to be made, these calculations are tedious, so that the use of a chart for determining X_X saves considerable time. The reactance X_X can be plotted against the capacitance difference $C_2 - C_1$, using C_1 as a parameter.

Since only a few fixed condensers are necessary for covering all ordinary conditions of measurement, only a few curves need be plotted. The condensers ordinarily used with the bridge are Type 505 Mica Units having capacitances of 100, 200, 500, and 1000 $\mu\mu$f. Using these values, the curves on the preceding page have been plotted.

We shall be glad to send an enlarged copy of this chart to any user of the Type 516-C Radio-Frequency Bridge who requests it.

USING THE VARIAC AT LOW FREQUENCIES

● **IN THE ARTICLE** bearing this title, which appeared in the December, 1940, issue, the current ratings for Type 200-CUH and Type 200-CMH Variacs are incorrectly listed. Rated current for these units is 2 amperes, and maximum current 2.5 amperes.

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THE General Radio EXPERIMENTER

VOLUME XV No. 9 MARCH, 1941

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

DELIVERIES

● **OUR MANUFACTURING AND ENGINEERING FACILITIES** are heavily overloaded at present with urgent work, most of it directly connected with the National defense program. Because of this, we are unable to accept most orders for equipment of special design.

We are making every effort to maintain adequate stocks and to make prompt deliveries of standard apparatus, much of which is also necessary for the defense program, but owing to slow deliveries of raw materials and parts, as well as to our own overtaxed facilities, deliveries on some items will necessarily be slow. We ask your indulgence and, on our part, we will do our best to get equipment to you as promptly as possible.

TRANSMITTER MAINTENANCE IN THE MODERN BROADCASTING STATION

By Charles Singer

(*Technical Supervisor, Station WOR**)

● **AS IN MANY OTHER PUBLIC SERVICES,** reliability in broadcasting is achieved only through careful and unceasing plant maintenance operations, which require not only competent personnel but accurate measuring instruments as well.

(*Continued on page 2*)

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*Bamberger Broadcasting Service, Inc., Carteret, N. J.

ALSO IN THIS ISSUE:

Commercial measuring equipment plays a vital role in the maintenance of transmitters. It makes the testing of transmitting equipment simple and accurate, and it saves many dollars in costly failures. Tubes, condensers, inductances, and all radio equipment can, by proper maintenance, be made to operate more efficiently, thereby achieving many more hours of operating life. In this way the instruments more than pay for themselves.

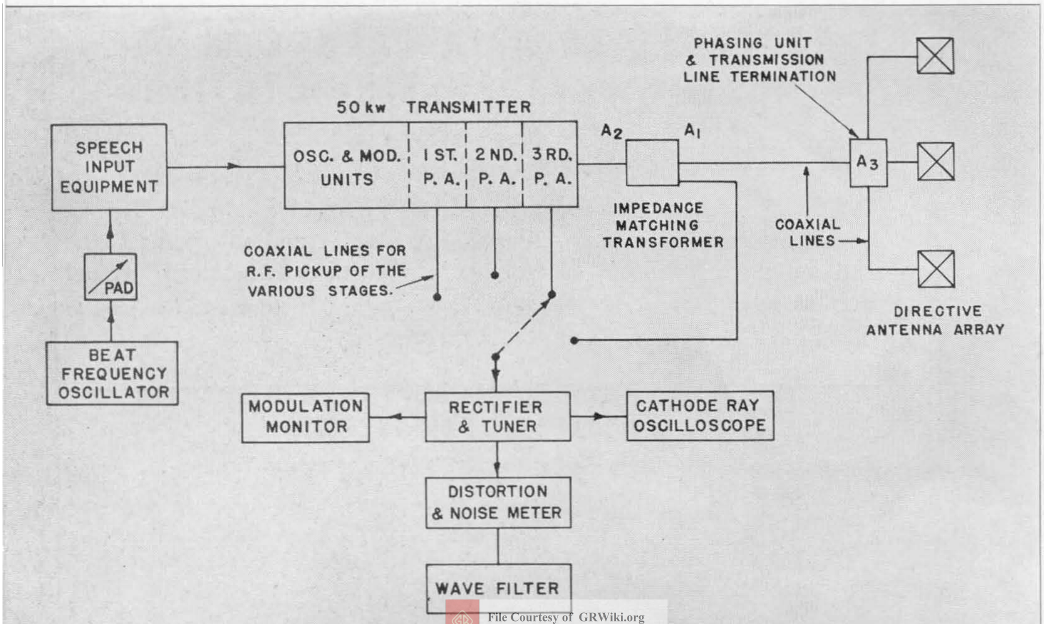
At WOR, the entire transmitting plant is operated and maintained from the instructions given in four operation routine manuals, namely:

1. Operating Routine
2. Night Maintenance
3. Day Maintenance
4. Special Maintenance

In these books, which were developed from WOR's operating experience, there is set forth, each day, a number of items which outline the maintenance to be done. For each individual item there is a detailed maintenance schedule describing the work to be done. Detailed maintenance of this type may seem unnecessarily complicated. It is, however, a vital operation if transmitter failures are to be avoided.

With the routine maintenance system of transmitter operations each component part is soon known to have certain measured values. At the completion of a given test, the operator enters his results on a log sheet beside the previous measurement of that component. Should any difference be observed, it is immediately corrected, and, in the majority of cases, a failure can be avoided. For example, each night after sign-off of WOR, the first maintenance operation is to feel all condensers. Two men are on watch; consequently, it takes a relatively short time to do this. Should a condenser have a hot spot on it, it is removed and checked on the radio-frequency bridge. Usually, these hot-spot condensers will show a change in power factor. Failure to remove such a condenser may result in "dead-air." All observations, whether

FIGURE 1. The arrangement of equipment for routine modulation, distortion and noise tests at WOR. Coaxial lines from each power amplifier stage are brought to the measurements room and permit measurements to be made on individual stages. Antenna impedance is measured at A_1, A_2, and A_3.



any change was or was not noted, must be logged.

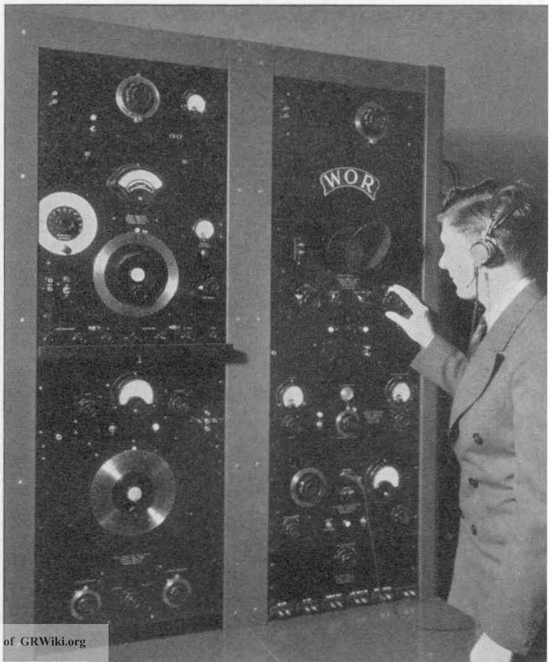
Another important use of the r-f bridge is in antenna impedance measurements. Such measurements are essential to broadcasting stations using directive arrays, as the replacement of almost any condenser in the phasing units will necessitate a minor adjustment in the impedance matching transformers to obtain the most efficient operation. Replacement may affect critically the impedance match of the individual lines causing a considerable loss of energy, as well as modifying the current balance among the towers, and this cannot be detected unless impedance measurements are made. It is not safe to permit maintenance adjustments in an antenna coupling unit unless an r-f bridge is available, so that, during the inspection, if anything is moved which results in a change in current balance, it may be readily corrected. At WOR the transmission line leading to the phasing unit is measured weekly, and it is assumed that if no change is noted all is normal. However, if a change of 2% in impedance is noted, which cannot be accounted for, a more thorough inspection of the phasing and termination units is undertaken.

The impedance of the individual towers may be measured at A_3 (see Figure 1). The impedance at A_1 in Figure 1 is checked weekly, and is normally of the order of 75 ohms. A_2 is checked monthly, since this transmission line does not undergo radical temperature changes. All measurements are made with the General Radio TYPE 516-C Radio-Frequency Bridge, and in conjunction with this bridge a TYPE 605-B Standard-Signal Generator is used as the r-f power source. The frequency calibration of the generator is checked by obtaining a zero beat against the transmitter crystals. The impedance measurement is then carried out with almost any number of observations over a range of plus or minus 20 kc from the carrier frequency.

Distortion measurements also play an important part in the maintenance routine of a broadcast station. Too often stations are run on the theory "It sounds all right to me!" This hit and miss analysis is not adequate today, because instruments are available for measurements to prove the efficiency of the transmitter adjustment. These instruments are easily used and provide an accurate analysis of transmitter performance. The General Radio TYPE 732-B Distortion and Noise Meter and the General Radio TYPE 732-P1 Range-Extension Filters are used for both audio and modulated r-f carrier measurements.

The transmitter is measured every

FIGURE 2. Equipment for measuring modulation and distortion consists of wave analyzer, beat-frequency oscillator, modulation meter, distortion and noise meter, range-extension filter, and cathode-ray oscillograph. These are conveniently mounted on a pair of relay racks as shown here.



Sunday night for distortion and noise. This test is expedited by the explanation in the routine book, and it requires only about fifteen minutes to make a complete set of measurements. The operations are as follows: First, the General Radio TYPE 713-B Beat-Frequency Oscillator is fed into the speech input equipment, and several levels are selected to provide 15%, 37.5%, 50%, 75%, 85%, and 100% modulation. At each level the distortion meter is used to determine the over-all distortion at a given audio frequency. Figure 1 illustrates this arrangement in block form. The equipment used for the test is shown in Figure 2. The test is repeated for various frequencies, namely, 50, 100, 400, 1000, 5000, and 7500 cycles. A noise measurement is also made, and positive and negative modulation peaks are checked on the TYPE 731-A Modulation Monitor. At the same time the modulated r-f carrier is observed on a cathode-ray oscilloscope. As was pointed out, this check is made weekly, but whenever a tube is removed in any of the modulated r-f stages, an additional check is made on that stage. This is made possible by coaxial lines which are brought to the measuring room from the 1st, 2nd, and 3rd power amplifiers as well as from a coil coupled to the impedance matching transformer, as illustrated in Figure 1. From previous measurements, each stage is known to perform in a given way, and any changes in its characteristics are immediately detected. Thus, distortion measurements have, in many instances, helped to prevent costly breaks during air shows.

The routine may seem a bit elaborate for a 15-minute test, but it is easily completed within the allotted time after the operations have been repeated a few times. Data sheets are supplied for the measurements, and, since all equipment is housed in one group of racks, all measurements and transmitter adjustments are made from the measuring room. Once each month an over-all check is made from the master control studio in New York, that is, a distortion check is made through the studio equipment to insure low line noise as well as low distortion in the studio equipment itself. Once each year a more comprehensive study is made with the General Radio Wave Analyzer. These checks are important and give the operator a feeling of security because everything is normal.

The routine for the 50 kw transmitter does not tell the entire maintenance story. Other important maintenance operations are carried out on short-wave broadcasting equipment. The many re-

FIGURE 3. Equipment used on the test bench includes impedance bridges, vacuum-tube voltmeters, standard-signal generators, and other general-purpose test instruments.

ceivers and portable transmitters require careful testing with adequate testing equipment to insure the continuity of short-wave broadcasts. To take care of the necessary maintenance of the short-wave equipment, as well as many other needs which have currently arisen from FM, a measuring bench was made to house these instruments.

Many instruments are capable of being utilized for a wide variety of tests. For example, the TYPE 605-B Standard-Signal Generator is used in almost every activity from alignment of short-wave receivers to field-strength measurements. In field-strength measurements the signal generator is used as a calibration oscillator in the 40-50 Mc band. For short-wave broadcasting it is used as an alignment oscillator for receivers in the high-frequency and ultra-high-frequency bands. It is used for the power source in r-f bridge measurements and for the measurement of all receiver characteristics. It is also used frequency modulated for oscilloscope studies.

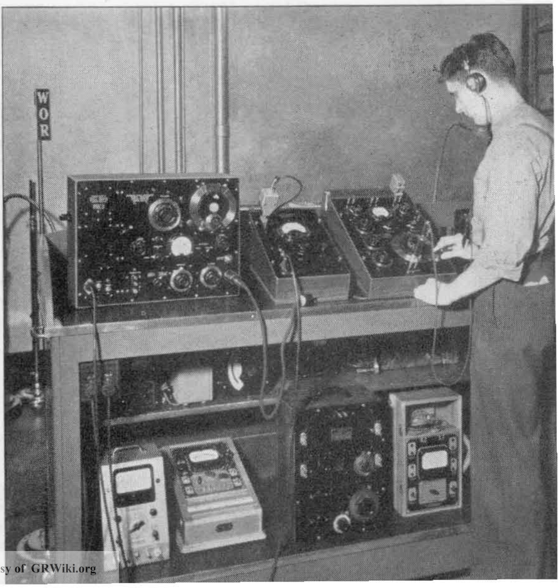
The TYPE 650-A Impedance Bridge also has many applications. It is particularly useful in measuring condensers and resistors in the regular nightly maintenance routine on the 50 kw transmitter. For example, overload resistors have very low values, and ohmmeter measurements are unreliable if accuracy is desired. The impedance bridge does an excellent job of checking these component parts at WOR.

The General Radio TYPE 726-A Vacuum-Tube Voltmeter has been used to check r-f voltages on transmission lines; it assists in receiver alignment and is used to measure transmission line voltages in portable high-frequency transmitters.

The high-frequency and ultra-high-frequency transmitters are subject to the same scrutiny as is the standard broadcast transmitter. Distortion measurements, frequency runs, etc., are also made on these transmitters, even when the power radiated is as low as 0.2 watts. A maintenance schedule is necessary for short-wave broadcast equipment, as each transmitter is designed to be used under a number of different conditions. These transmitters are designed and constructed in the laboratory of WOR's Technical Facilities Division, each with different characteristics to meet different conditions. The frequency-response characteristics used with a given transmitter depend upon the background noise conditions. For example, the low frequencies need more attenuation in airplanes than in motor boats and submarines, and this requires that each type of broadcast be given separate consideration.

Many other instruments are used at WOR and, since it would require entirely too much space to explain their specific uses in detail, they are merely listed here. These instruments have an

FIGURE 4. Another view of the test bench showing the TYPE 650-A Impedance Bridge in use.



ever-increasing utility, and, after continual usage, they become almost indispensable.

1. Wave Analyzer
2. Audio Oscillator
3. Q Meter
4. Heterodyne Frequency Meter
5. Frequency-Limit Monitor
6. Standard-Signal Generator
7. A-C Vacuum-Tube Voltmeter
8. Impedance Bridge
9. R-F Bridge
10. Cathode-Ray Oscilloscope
11. Distortion and Noise Meter
12. Range-Extension Filter
13. Modulation Monitor
14. Beat-Frequency Oscillator
15. Capacity Bridge
16. D-C Vacuum-Tube Voltmeter
17. Field-Strength Measuring Set
18. D-C Amplifier and Recorder
19. Square-Wave Generator
20. U-H-F Standard-Signal Generator
21. Portable Cathode-Ray Oscilloscope

A modern broadcast station needs these instruments to attain the best possible conditions for efficient operation as it is felt that only through such a system can broadcasting provide a better public service.

CALIBRATING THREE-PHASE POWER-FACTOR METERS WITH THE VARIAC*

● **IT CAN BE SHOWN** that a single-phase source of controllable voltage plus a voltmeter to read the voltage is all that is necessary to calibrate three-phase electrodynamic power-factor meters, but, because of the usual difficulty of obtaining the variable voltage, it has been customary in the past to use ammeters, voltmeters, and a three-phase wattmeter on a three-phase line with a load to make this calibration. Since the Variac is a convenient device for varying the voltage easily and with negligible phase shift, it becomes an important aid in more easily calibrating these meters.

A description of the power-factor meter itself and its principle of operation will help show the method of using the Variac for calibration. Both single and polyphase power-factor meters have

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*We are indebted to Mr. Paul McGahan of the Westinghouse Electric and Manufacturing Company for the original suggestion and details of this application.

FIGURE 1 (*left*). Schematic diagram of the single-phase power-factor meter.

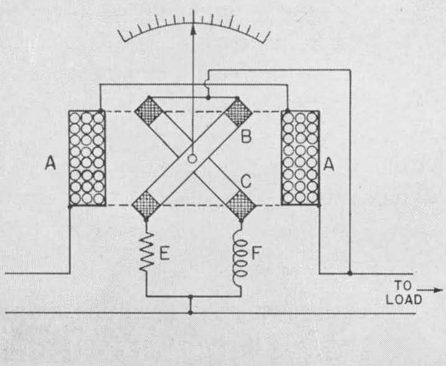
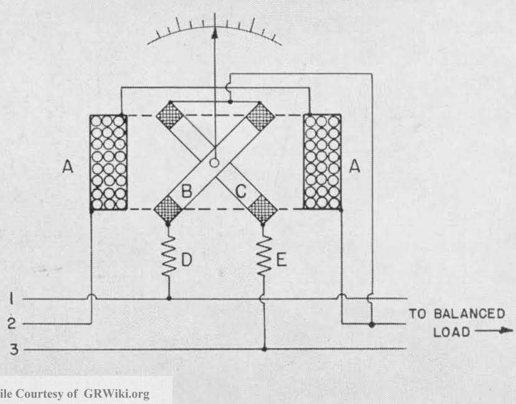


FIGURE 2 (*right*). Schematic diagram of the three-phase power-factor meter.



a stationary series coil, *A*, and two movable voltage coils, *B* and *C*. The two moving coils are fixed with respect to each other, at an angle of 90°, but rotate freely with respect to the fixed coil. No springs or other restoring devices are used, and the only torque produced on the coils is that caused by the action of the electric fields produced when the instrument is connected to the power line. In the single-phase models the voltage is fed to one of the movable coils through a resistance and to the other one through a reactance, as shown in Figure 1. Thus the current through one coil is in phase with the line voltage, while the current is out of phase with the voltage by 90°. With this arrangement there is no torque between the latter coil, *C*, and the fixed coil at unity power factor, and so the movable coil, *B*, orients itself parallel to the fixed coil. As the power factor decreases, the torque between the fixed and in-phase movable coil, *B*, decreases, while the torque between coil *C* and the fixed one increases. Thus, as the power factor decreases, the coils rotate farther and farther from the unity-power-factor position before equilibrium is reached, until finally, at zero power factor, there is no torque between the inphase coil, *B*, and the fixed coil, and coil *C* aligns itself parallel to the fixed one.

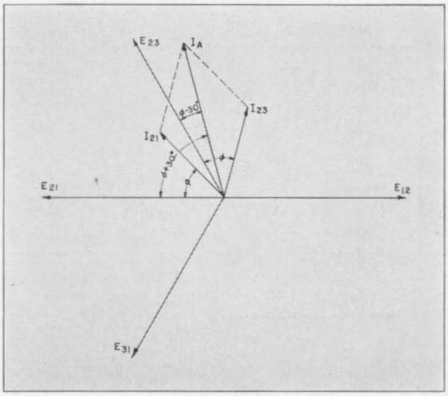


FIGURE 3. Vector diagram of the voltage and current relations in the three-phase power-factor meter. The current through the fixed coil is I_A, and the voltages across coils B and C are E_{21} and E_{23} respectively.

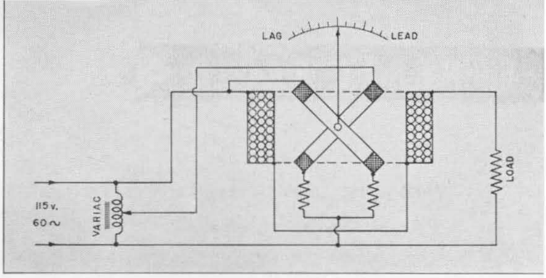


FIGURE 4. Diagram showing how the Variac is connected to the power-factor meter. With connections as shown, one-half the scale can be calibrated. By shifting the Variac to the other movable coil, the other half of the scale can be covered.

In the three-phase power-factor meter, use is made of the different phase voltages instead of using a reactance to shift the phase between the two movable coils. Because of this arrangement, and the fact that only one line current is applied to the instrument, the three-phase instrument operates satisfactorily on the assumption of balanced load conditions only. Figure 2 shows the method of connection to the line and load. The equations for the torque produced between the various coils are:

Torque between *A* and *B* =

$$I_A \times E_{21} \times \cos(\theta + 30°) \times \sin AB$$

and torque between *A* and *C* =

$$I_A \times E_{23} \times \cos(\theta - 30°) \times \sin AC$$

where I_A is current through the fixed coil and E_{21} and E_{23} are the respective phase voltages. The phase angle of the load is θ, and the cosine terms represent the electrical angle between I_A and E_{21} or E_{23}, as shown in the vector diagram of Figure 3. The sine terms represent the mechanical angles between the moving and fixed coils. When the two torques are equal but opposite, the moving coils will come to rest, and their angles with respect to *A* can be determined from the two equations by equating them thus:

$$I_A \times E_{21} \times \cos(\theta + 30^\circ) \times \sin AB = I_A \times E_{23} \times \cos(\theta - 30^\circ) \times \sin AC$$

$$\text{or } \frac{E_{21}}{E_{23}} \times \frac{\cos(\theta + 30^\circ)}{\cos(\theta - 30^\circ)} = \frac{\sin AC}{\sin AB}$$

Thus it will be seen that only when the load is balanced and E_{21} equals E_{23} can the mechanical angle of the movable coils be calibrated in terms of the phase angle. It will also be seen from the equation that the mechanical angle can be changed, for a given phase angle, by varying E_{21} and E_{23} with respect to each other. This fact points to the new method of calibration — keeping θ fixed and at zero while varying the ratio E_{21}/E_{23}. The diagram of Figure 4 shows how the Variac and power-factor meter are connected for this method.

The ratio E_{21}/E_{23} need only be set equal to the ratio $\frac{\cos(\theta+30^\circ)}{\cos(\theta-30^\circ)}$ to make the movable coils rotate to the proper angle for a phase angle θ. Hence, we can make up the table shown below.

The negative values indicate a change in phase of 180°, but, since most power-factor meters are calibrated from 50% lag through 100% to 50% lead, the low power factors are of lesser importance. Below is a connection diagram showing how half the scale would be calibrated. By changing the Variac to the other movable coil the other half of the scale would be covered.

% PF *cos θ*	θ	$(\theta + 30^\circ)$	*Cos* $(\theta + 30^\circ)$	$(\theta - 30^\circ)$	*Cos* $(\theta - 30^\circ)$	$\frac{\cos(\theta + 30^\circ)}{\cos(\theta - 30^\circ)}$	E_{21} *% of Normal*	E_{23} *% of Normal*
0 lead	−90°	−60°	.500	−120°	−.500	−1.	100	−100
30 lead	−72°32′	−42°32′	.740	−102°32′	−.213	−3.47	100	−28.8
50 lead	−60°	−30°	.866	−90°	0	0	100	0
70 lead	−45°34′	−15°34′	.964	−75°34′	.253	3.80	100	26.3
100	0°	30°	.866	−30°	.866	1.	100	100
70 lag	45°34′	75°34′	.253	15°34′	.964	.263	26.3	100
50 lag	60°	90°	0	30°	.866	0	0	100
30 lag	72°32′	102°32′	−.213	42°32′	.740	−.288	−28.8	100
0 lag	90°	120°	−.500	60°	.500	−1.00	−100	100

— Martin A. Gilman

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THE General Radio EXPERIMENTER

VOLUME XV No. 10 APRIL, 1941

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

INSTRUMENT SPECIFICATIONS

● **OWING TO GOVERNMENTAL RESTRICTIONS** on the use of certain materials it has become necessary to use substitutes wherever possible. While the performance specifications of our instruments are not affected, substitute materials are being used. In accepting orders during the present situation brought about by the national defense program, we reserve the right to make any necessary substitutions in the constructional features.

NEW TERMINALS FOR USE WITH COAXIAL TRANSMISSION LINES

● **AT HIGH AND ULTRA-HIGH FREQUENCIES** the desirability of constraining electro-magnetic fields within very definite confines has led largely to the adoption of coaxial transmission lines for the transference of power from one point to another. Ideal transmission lines of this type, having inner and outer conductors of zero resistance and an intervening medium of zero power factor, are theoretically capable of transferring power with zero energy loss and with zero external field, and the properties of actual lines approach those of the ideal very closely.

(*Continued on page 2*)

ALSO IN THIS ISSUE:

For measuring circuits, in particular, coaxial lines have been found almost indispensable. Because of the uniform distribution of inductance and capacitance along their lengths and the absence of appreciable losses from radiation, they generally follow the conventional "engineering solution" of the long line with great accuracy at frequencies extending into the hundreds of megacycles. To the properties of low losses and of low external field must therefore be added the further virtue of accurate predictability.

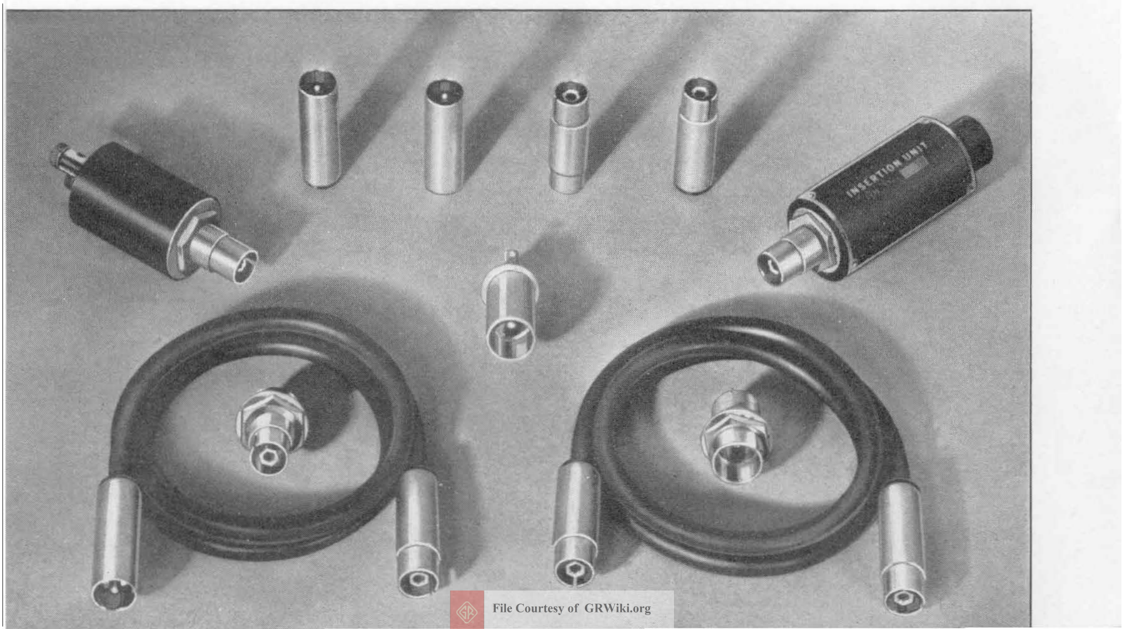
To obtain maximum benefit from coaxial lines, however, it is of great importance to use proper terminal equipment. For convenience, a plug and jack system is often highly desirable for use in setting up measuring systems. The plugs and jacks used should have two very definite properties, namely, (1) they should be as short as possible and have a characteristic impedance differing from that of the line as little as possible, in order to minimize reflections resulting from impedance mismatch, and (2) they should have as continuous an external shield as possible, in order to minimize external fields.

The TYPE 774 Coaxial Terminals, shown in Figure 1, have been designed with these two properties in mind. In order to reduce impedance mismatch in lines having different characteristic impedances they have been made with as short internal conductors as possible and with as low capacitance as possible. In order to provide as continuous an external shield as possible, lugs have been provided for four connections to the outer shell from the cable sheath at points uniformly distributed around the circumference.

The solid dielectric is polystyrene, which has both a low dielectric constant and a low power factor. These properties make possible the low capacitance and low losses of TYPE 774 Coaxial Terminals.

A plug unit and a jack unit are available for mounting on panels, and a similar pair of units for terminating coaxial cables. The plug connector and the jack connector make it possible to join two cables having identical terminations, that is, two plugs or two jacks.

FIGURE 1. A group of TYPE 774 Coaxial Terminals.



For many applications the capacitance of these units is the factor to be considered in determining their suitability. The capacitance for each Type 774 Unit is listed in Table I. In addition to the total capacitance listed in the first column, there is given, for many units, a figure called "insertion capacitance," which is the capacitance added to a circuit when that particular unit is plugged in. This is lower than the total capacitance because of the overlapping when a plug unit is plugged into a jack.

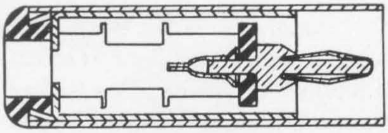
In addition to the connectors listed in Table I, another unit, for use with Type 684-A Modulated Oscillator, is available. This adapter can be installed in place of the output binding posts provided on the oscillator. This type of output terminal is necessary when the oscillator is used as a power source for impedance measurements at frequencies above a few megacycles, as for instance with the Type 821-A Twin-T.

The importance of maintaining the continuity of the external conductor in

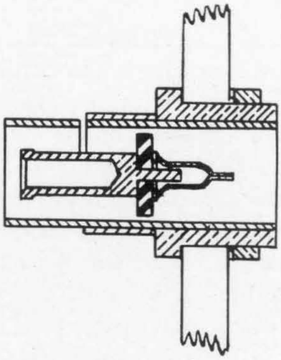
TABLE I

Type		*Total Capacitance*	*Insertion Capacitance*
774-P	Panel Jack	2.8 $\mu\mu$f	1.7 $\mu\mu$f
774-G	Panel Plug	2.4	1.3
774-M	Cable Jack	2.8	1.7
774-E	Cable Plug	2.5	1.4
774-F	Plug Connector	3.6	1.3
774-N	Jack Connector	4.2	2.0
774-X	Terminal Unit	6.0	4.9
774-M 774-G	Cable Jack and Panel Plug	4.1	....
774-P 774-E	Cable Plug and Panel Jack	4.1	....

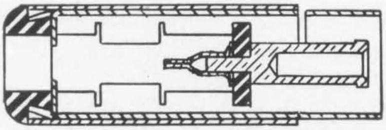
Figure 2. Panel view of a Type 684-A Modulated Oscillator with Type 774-V Coaxial Adapter installed. The adapter fits one of the mounting holes for the standard binding post terminals supplied with the instrument. A metal button to cover the other hole is furnished.



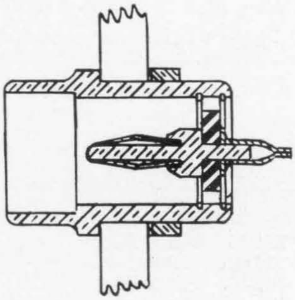
Type 774-E Cable Plug.



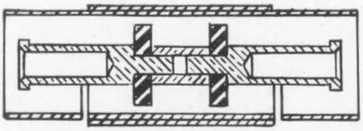
Type 774-P Panel Jack.



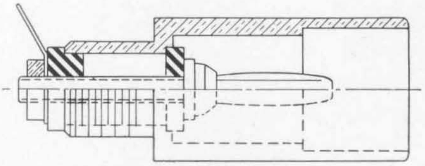
Type 774-M Cable Jack.



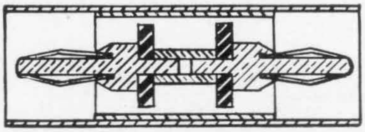
Type 774-G Panel Plug.



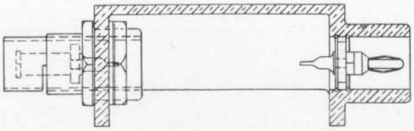
Type 774-N Jack Connector.



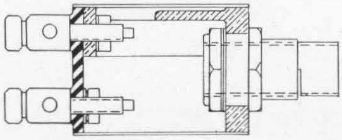
Type 774-V Adapter.



Type 774-F Plug Connector.



Type 774-X Insertion Unit.
(One-half actual size.)



Type 774-YB Terminal Unit.
(One-half actual size.)

Figure 3. Sectional drawings showing construction of Type 774 Coaxial Terminals. Drawings are full size except where indicated.

measurements with this instrument can be seen from the following example:

A measuring system comprising a Type 684-A Modulated Oscillator, a Type 821-A Twin-T Impedance-Measuring Circuit, and a radio receiver are connected as shown in Figure 4.

A small amount of series inductance in the ground side of the generator cable is designated as L_G, a similar inductance in the receiver cable as L_R, and the common ground lead as L_M. The voltage

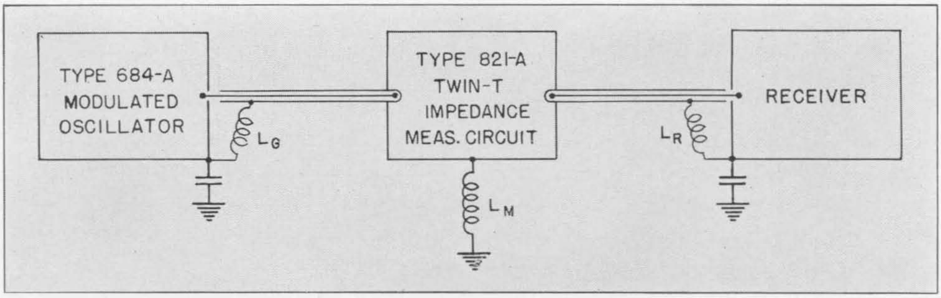


FIGURE 4. Block diagram illustrating the effect of series inductance in the generator and detector leads.

drop in L_G produces a flow of current around the loop consisting of the cable sheath, the ground lead, L_M, and the ground capacitance of the oscillator. Similarly, current flows in the right-hand loop that includes L_R.

The voltage applied to the receiver has, therefore, two components, one from the Twin-T, the other from the drop across L_R. When a null point is reached, therefore, the Twin-T is out of balance by an amount necessary to cancel the effect of the extraneous voltage from L_R, that is, to make the vector sum of the Twin-T output voltage and the extraneous voltage equal to zero.

The error in measurement caused by this series inductance is one of the most serious encountered in null measurements at radio frequencies, and, in order to avoid it, coaxial terminals should be used on both generator and receiver.

TYPE 774 Coaxial Terminals can also be supplied already assembled as patch cords with 3-foot lengths of concentric cable, for use in measuring circuits. Two assemblies are available: TYPE 774-R1, which has a plug at each end; and TYPE 774-R2, which has one plug and one jack.

The cable is made to our specifications by Simplex Wire and Cable Company and consists of a stranded beryllium-copper conductor separated from a braided tinned-copper shield by Anhydrex A insulation, with an over-all covering of abrasion-resistant rubber. The nominal characteristic impedance of the cable is 72* ohms; the nominal capacitance is 26 $\mu\mu$f per foot; and the power factor is 2% or less at 1000 cycles.

— D. B. SINCLAIR

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*This is subject to a variation of ±10%.

Type	*Description*	*Code Word*	*Price*
774-E	Cable Plug	ACCESSOEYE	$1.50
774-M	Cable Jack	ACCESSOMUD	1.50
774-G	Panel Plug	ACCESSOGOD	1.00
774-P	Panel Jack	ACCESSOPOP	1.00
774-F	Plug Connector	ACCESSOFIG	1.00
774-N	Jack Connector	ACCESSONUT	1.00
774-R1	Patch Cord	ACCESSORIM	4.00
774-R2	Patch Cord	ACCESSORAT	4.00
774-X	Insertion Unit	ACCESSOXEB	4.50
774-YB	Terminal Unit	ACCESSOYAM	3.50
774-V	Adapter for TYPE 684-A Modulated Oscillator	ACCESSOVAN	1.75
774-A	Concentric Shielded Cable (3 feet)	ACCESSOAPE	1.00

A NEW AUDIO-FREQUENCY MICROVOLTER*

● **IN MANY TYPES OF WORK,** such as the measurement of amplifier gain, hum level, overload points, and transformer characteristics, the TYPE 546-B Audio-Frequency Microvolter is an extremely useful accessory for obtaining accurate answers rapidly, for by using the microvolter with an audio oscillator an accurately known source of continuously variable voltage is made available over the range from 1.0 microvolt to 1.0 volt.

The microvolter consists of an adjustable attenuator, the input to which is standardized by means of the voltmeter on the panel of the instrument. The attenuator settings are made by means of two panel controls, one a six-step decade multiplier and the other an individually calibrated dial which gives continuous variation over each decade. This dial has a scale which is essentially logarithmic and is calibrated in decibels as well as voltage. The voltmeter is of the copper-oxide rectifier type and is so designed that it has negligible frequency error over the range from 50 to 40,000 cycles. The meters are standardized at 77° F., and a slight correction is necessary if ambients differing widely from this value are encountered. A curve of this correction is given in Figure 2.

Several features of the new microvolter are outstanding. By the elimination of the input transformer and by careful design of the attenuator, the leakage and extraneous pickup are reduced to less than 0.1 microvolt. Therefore excellent voltage accuracy is obtained even at levels as low as 1.0 microvolt, and so gain and overload characteristics of high-gain amplifiers can be measured accurately with only a few microvolts on the input. Furthermore, the decibel calibration makes it possible to obtain gain or loss values, in decibels, for amplifiers, transformers, lines, and other networks, without the necessity of manipulating voltage ratios and then converting them.

The absolute accuracy of the output voltage has been made 3% ±0.1 micro-

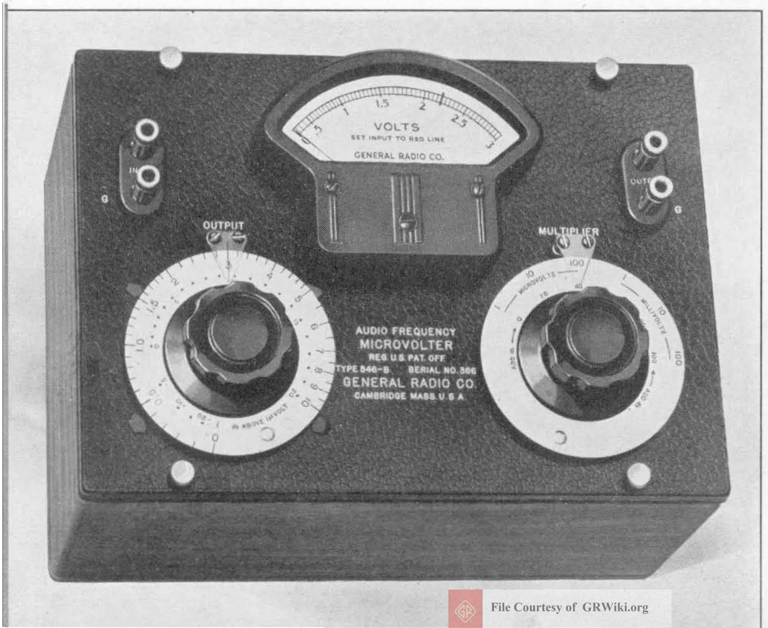

FIGURE 1. Panel view of TYPE 546-B Microvolter. The output scale is spread out for precise reading and is approximately logarithmic. The auxiliary decibel scale carries a dot for each decibel between 1 and 20.

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*Reg. U. S. Pat. Off.

volt at all voltages above 1.0 microvolt over the entire frequency range from 50 to 40,000 cycles. Because the output impedance is only 200 ohms no corrections for the load impedance are necessary in most measurements. Where voltage ratios are all that are important, or where any error in the input voltmeter can be eliminated, such as by use of an external meter, the accuracy is 2%. The frequency range may be extended to 100,000 cycles with this same accuracy if the output level is greater than 100 microvolts. — Martin A. Gilman

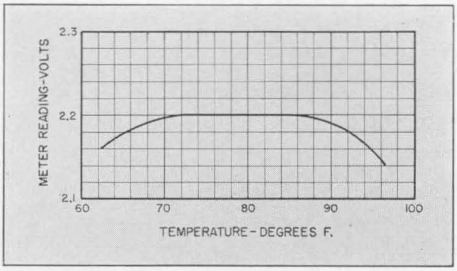


Figure 2. Temperature characteristic of meter used in Type 546-B Microvolter.

SPECIFICATIONS

Output Voltage Range: From 0.1 microvolt to 1.0 volt, open circuit, when the input voltage is set to the standardized reference value.

Accuracy: For open-circuit output voltages the calibration is accurate within ±3% ± 0.1 microvolt for output settings above 1 microvolt and for all frequencies between 50 and 40,000 cycles. This accuracy applies only where waveform and temperature errors are negligible (see below). Below 1 microvolt the error increases owing to crowding of the scale.

For ratios or increments of voltage, at a given frequency, the accuracy of any reading is within ±2% ± 0.1 microvolt, at frequencies up to 100,000 cycles. At the higher frequencies this accuracy applies only at levels above 100 microvolts.

Output Impedance: The output impedance is approximately 200 ohms and is constant with setting within ±5%. This impedance is sufficiently low so that no correction on the output voltage is necessary for load impedances of the order of 50,000 ohms and greater.

Input Impedance: Approximately 430 ohms, substantially independent of output setting on all but the highest multiplier position.

Waveform Error: The accuracy of the microvolter as a calibrated attenuator or voltage divider is independent of waveform. The absolute accuracy of the output voltage calibration depends on the characteristics of the input voltmeter, which has a small waveform error that depends in turn on both the phase and the magnitude of harmonics present in the input. This error in the voltmeter can, in general, be neglected when the microvolter is used with ordinary laboratory oscillators.

Temperature Error: The accuracy of the calibration is independent of temperature when the microvolter is used as an attenuator or voltage divider. The absolute accuracy is affected slightly by temperature because of change in the voltmeter characteristics. The necessary correction for temperatures from 65° to 95° F. are furnished with the instrument.

Power Source: The driving oscillator must be capable of furnishing about 2.2 volts across 430 ohms, or about 11 milliwatts.

Accessories: Two Type 274-M Plugs are supplied.

Terminals: Jack-top binding posts are mounted on standard ¾-inch spacing.

Mounting: The instrument is mounted on an aluminum panel in a shielded walnut cabinet.

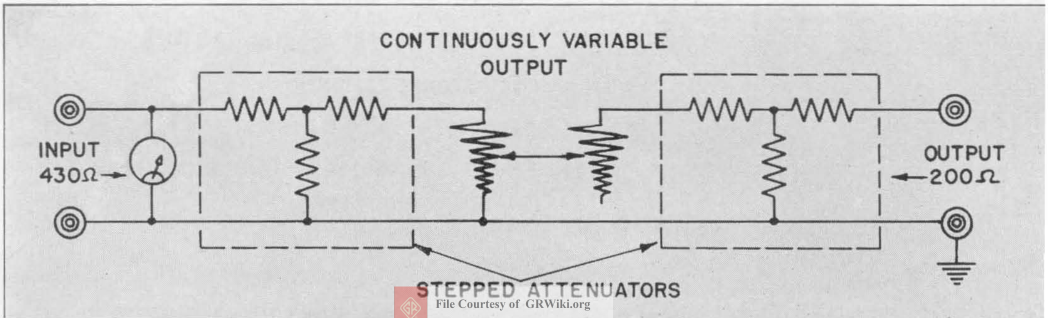
Dimensions: (Length) 10 x (width) 7 x (height) 6⅜ inches, over-all.

Net Weight: 6½ pounds.

Type		*Code Word*	*Price*
546-B	Microvolter	CROWN	$80.00

Type 546-B Microvolter is temporarily out of stock. Deliveries will be resumed in August, 1941.

Figure 3. Schematic circuit diagram of Type 546-B Microvolter.



SOUND-LEVEL METER JOINS POLICE FORCE

● **IN THE CURRENT PHASE** of New York City's anti-noise campaign, renewed on January 4 by Mayor LaGuardia, 10,000 taxicabs are having their horns rated by the General Radio Sound-Level Meter under the supervision of Thomas W. Rochester, chief engineer for the Police Department. The accompanying photograph shows the test arrangement used in preliminary measurements made for the purpose of collecting basic data from which a standard horn rating can be determined. Measurements were made at distances of 100 feet and 20 feet from the sound-level meter. These preliminary measurements indicated that the range of sound levels encountered extended from a maximum of 90 db at 20 feet to a minimum of 75 db at 100 feet. For most horns, a difference of about 5 db in level between the two distances was noted.

MISCELLANY

● **COVER-TO-COVER READERS** must have been somewhat baffled by the following paradoxical statement appearing in Mr. Gilman's article in the March *Experimenter:* "Thus the current through one coil is in phase with the line voltage, while the current is out of phase with the voltage by 90°." This condition is admittedly hard to achieve in practice, and the best way out of the difficulty is to insert the words "in the other coil" after the word "while." In order to fit the last piece into the puzzle, the editor admits that ϕ, in Figure 3, and θ, in the balance of the article, are one and the same angle.

GENERAL RADIO COMPANY
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THE General Radio EXPERIMENTER

VOLUME XV No. 11 MAY, 1941

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

Also IN THIS ISSUE

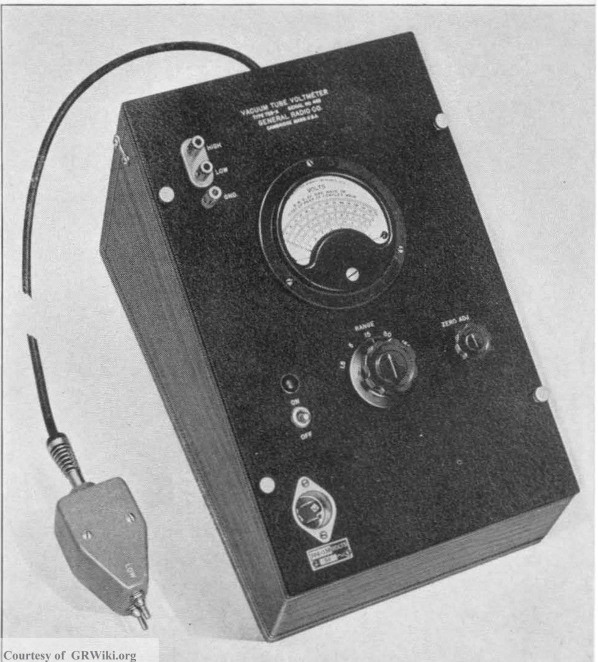
RADIO-FREQUENCY CHARACTERISTICS OF THE TYPE 726-A VACUUM-TUBE VOLTMETER

● **BECAUSE OF THE WIDE ACCEPTANCE** of the TYPE 726-A Vacuum-Tube Voltmeter[1] for the measurement of voltage and current[2] at radio frequencies, a discussion of its behavior at frequencies up to 150 megacycles may be of some general interest.

A knowledge of the effective input impedance at the terminals of a voltmeter is desirable so that the instrument can be used with confidence in a given application. With an exact knowledge of the input impedance and its variation with frequency, it will, in general, be possible to estimate or compute the effect of the voltmeter on the circuit under measurement.

In most applications, however, one component only of the impedance is of real significance. For example, when measuring the voltage across tuned circuits it is usually possible to retune the circuit

FIGURE 1. Panel view of the TYPE 726-A Vacuum-Tube Voltmeter.

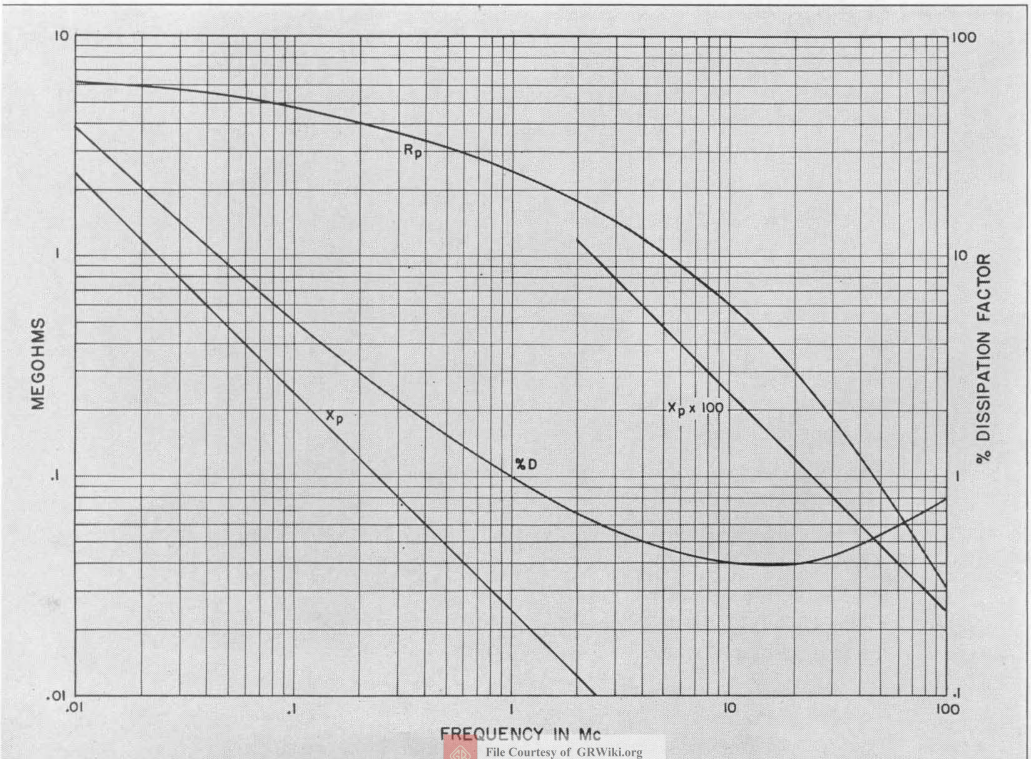


to compensate for the capacitance, so that the parallel resistance R_p becomes the effective input impedance. When making measurements on untuned circuits, on the other hand, the capacitance is the significant component, its impedance being much smaller than R_p at high frequencies.

When the voltmeter is used as an indicator in the parallel-resonance method of impedance measurement, both components of impedance are absorbed by virtue of the substitution method employed. In this case, however, the finite impedance of the vacuum-tube voltmeter acts to reduce the "resolving power" of the method of measurement. Thus, even here, a knowledge of the impedance is useful, as it permits us to estimate the resulting limitations on range and accuracy.

Figure 2 shows the measured variation of the effective parallel resistance R_p, as well as the parallel capacitive reactance X_p of the Type 726-A Vacuum-Tube Voltmeter. The new Type 821-A Twin-T Impedance-Measuring Circuit[8] was used in making these measurements over the frequency range from 1 megacycle to 30 megacycles. At the lower frequencies the Type 516-C Radio-Frequency Bridge was used, while in the region from 30 to 100 megacycles a susceptance-variation circuit[9] was employed. The measurements were made on several voltmeter probes, with the plug tips removed. At low frequencies the input impedance is equivalent to a resistance of approximately 6 megohms, shunted by a capacitance of 6.6 $\mu\mu$f. The reduction in R_p at the higher frequencies is caused by dielectric losses. These losses occur in the yellow bakelite housing between the input terminals, in the ceramic tube socket, in the blocking condenser and the resistors,

Figure 2. Plots of input reactance, resistance, and dissipation factor of Type 726-A Vacuum-Tube Voltmeter as a function of frequency.



and in the glass envelope of the diode. (The diode conductance loss is unimportant at high frequencies.) The total input impedance may also be considered as that of a capacitance whose dissipation factor varies with frequency. The effective dissipation factor $\frac{1}{R_p \omega C}$ of the input capacitance is also plotted in Figure 2.

When the Type 726-P1 Multiplier[3] is used with the vacuum-tube voltmeter the effective input impedance is even higher than that indicated by Figure 2, being approximately equivalent to that of a 4.5 $\mu\mu$f condenser of less than 0.5% power factor. The multiplier thus is an excellent means of obtaining extremely high input impedances, if the 10:1 reduction of sensitivity is permissible.

It is well known that certain frequency effects are present in any vacuum-tube voltmeter circuit, which cause the voltage indications at very low and at very high frequencies to differ from the true value of the applied voltage. At some low frequency the reactance of the condenser in the diode rectifier circuit will become sufficiently high so that a significant fraction of the applied voltage appears across it rather than across the diode. In addition, the dynamic characteristics of the indicating meter and the characteristics of the amplifier can also become important in determining the response to the impressed voltage. In the Type 726-A Vacuum-Tube Voltmeter, however, these other effects become significant at frequencies much lower than those at which the reactance error first becomes appreciable. The performance is substantially independent of frequency, even at the lowest audio frequencies. At 20 cycles the error is less than 1%.

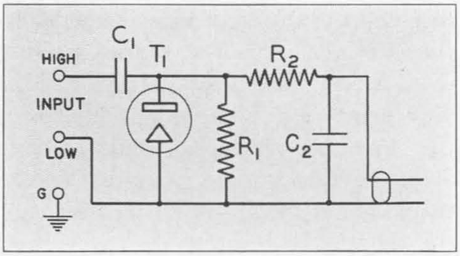


Figure 3. Schematic diagram of the input circuit of Type 726-A Vacuum-Tube Voltmeter.

At high radio frequencies, on the other hand, two important phenomena come into play that have a progressively more important effect on the performance of the voltmeter as the frequency is raised. The most commonly known of these is resonance in the input circuit of the voltmeter. To a first approximation, it can be considered that the inductance of the input leads resonates simply with the anode-to-cathode capacitance of the diode, causing the voltage acting on the diode to increase. As a consequence, the voltmeter reads higher than the true value of the impressed voltage. The resulting increase in voltmeter indication can be calculated with a fair degree of accuracy for frequencies up to one-half or one-third of the resonant frequency,* and the result is independent of the voltage level.

The second important phenomenon which affects the behavior of a voltmeter at high frequencies is the finite time of transit of electrons from the cathode to anode of the diode rectifier. This effect has been widely discussed in the literature [4,5,6] under the various names of "electron-inertia error," "transit-time effect," and "premature cut-off."

Referring to Figure 3, the condenser C_1 tends to charge to the voltage required to turn back electrons at the anode of the diode.[7] If the time of elec-

---

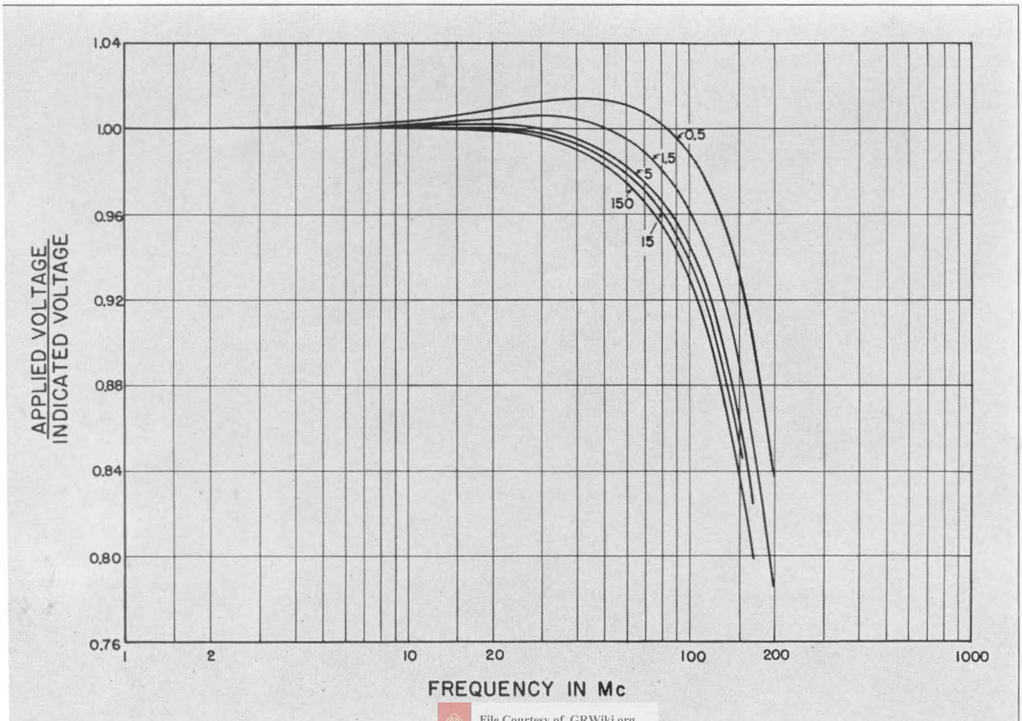
*The resonant frequency is approximately 400 megacycles, with the probe tips removed.

tron flight in the diode is negligible compared to the period of the alternating voltage, the condenser voltage will be very nearly equal to the peak value of the impressed voltage. Actually, however, the transit time is comparable to the period at the higher frequencies, and the electric field acting on an electron changes while it is in flight. The retarding field acting during the inverse portion of the cycle effectively reduces the voltage required to turn the electron back. Consequently, the voltage on *C* will not reach the value that would occur for a negligible transit time. It can be seen that the effect of this phenomenon is a voltage indication that is lower than the true value. Furthermore, since the time of transit is a function of the plate-to-cathode potential, the deviation is a function of the applied voltage as well as of the frequency.

Figure 4 shows the ratio of applied voltage to indicated voltage for the TYPE 726-A Vacuum-Tube Voltmeter as a function of frequency, for several different voltage levels. These data were taken with the plug tips of the voltmeter probe removed. If the tips are in place and the voltage is applied near the ends, the deviation is larger than that indicated, since the resonant frequency is lowered by the added inductance and capacitance of the tips.

It will be observed that, at the high voltage levels, the percentage deviation is essentially independent of voltage. That is, the electron transit time is small by virtue of the relatively high accelerating field acting, and resonance is the controlling factor in the behavior. At lower levels, however, the transit time phenomenon introduces a compensating effect which reduces the deviation between the true value and the indicated value of voltage. At levels in the vicinity

FIGURE 4. Plot of the ratio of applied voltage to indicated voltage as a function of frequency for several different levels of indicated voltage. The voltage levels are indicated by the figures associated with each curve.



of ½ volt the two effects very nearly cancel each other, and the frequency error is less than 2% up to 100 megacycles. At frequencies up to about 50 megacycles the frequency error is less than ±2% for all levels.

The curves of Figure 4 also apply when the Type 726-P1 Multiplier is used, since it introduces no appreciable frequency error in the range from 1 to 100 megacycles.

When non-sinusoidal voltages are being measured at high frequencies, the peak value of the voltage acting on the diode may be markedly different from the peak value at the probe terminals, because of the different value of resonant rise experienced by the various components of the voltage. Consequently, considerable caution must be exercised in interpreting readings at the higher frequencies, if the voltage under measurement is not very nearly sinusoidal. For sinusoidal voltages, however, it is possible to make measurements at frequencies up to 150 megacycles with substantially the same accuracy that obtains at lower frequencies, by using the correction factors indicated in Figure 4.

— Ivan G. Easton

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

[2] D. B. Sinclair, "The Type 726-A Vacuum-Tube Voltmeter as a Radio-Frequency Ammeter" — General Radio *Experimenter*, Vol. XIII, Nos. 3 and 4, August-September, 1938.

[3] D. B. Sinclair, "A Voltage Multiplier for Use With the Vacuum-Tube Voltmeter at Radio Frequencies" — General Radio *Experimenter*, Vol. XIV, No. 12, May, 1940.

[4] L. S. Nergaard, "Electrical Measurements at Wave Lengths Less Than Two Meters" — Proc. I.R.E., Vol. 24, No. 9, p. 1207, September, 1936.

[5] E. C. S. Megaw, "Voltage Measurements at Very High Frequencies" — *Wireless Engineer*, Vol. XIII, No. 149, p. 65, Vol. XIII, No. 150, p. 135, and Vol. XIII, No. 151, p. 201.

[6] C. L. Fortescue, "Thermionic Peak Voltmeters for Use at Very High Frequencies" — Proceedings of Wireless Section, I.E.E., Vol. X, No. 262, 1935.

[7] C. B. Aiken, "Theory of the Diode Voltmeter" — Proc. I.R.E., Vol. 26, No. 7, p. 859, July, 1938.

[8] D. B. Sinclair, "A New Null Instrument for Measuring High-Frequency Impedance" — General Radio *Experimenter*, Vol. XV, No. 7, January, 1941.

[9] R. F. Field, "An Improved Measuring Circuit for the Susceptance Variation Method" — General Radio *Experimenter*, Vol. XV, No. 4, September-October, 1940.

● **LAST MONTH** we pointed out that, in order to conserve essential materials for National defense, substitutes are being used in General Radio instruments. An example of this is the use of plastics instead of aluminum for panels where the substitution does not impair the performance of the instrument. Important mechanical characteristics such as strength and stability, as well as appearance, have been considered in selecting the most acceptable substitute, so that the high standard of quality in General Radio instruments will be maintained.

Maintenance of quality is particularly important at this time, because the bulk of our products is going either to manufacturers and laboratories engaged in National defense work or directly to the various government activities.

THE GENERAL RADIO STANDARDIZING LABORATORY

● **QUALITY CONTROL IN INSTRUMENT MANUFACTURE** is an extremely important function. However well-designed an instrument may be, accurate calibration and reliability in service are the qualities that determine its ultimate usefulness. Careful inspection, adjustment, and standardization are necessary to achieve the reliability and accuracy that are promised in the manufacturer's published specifications.

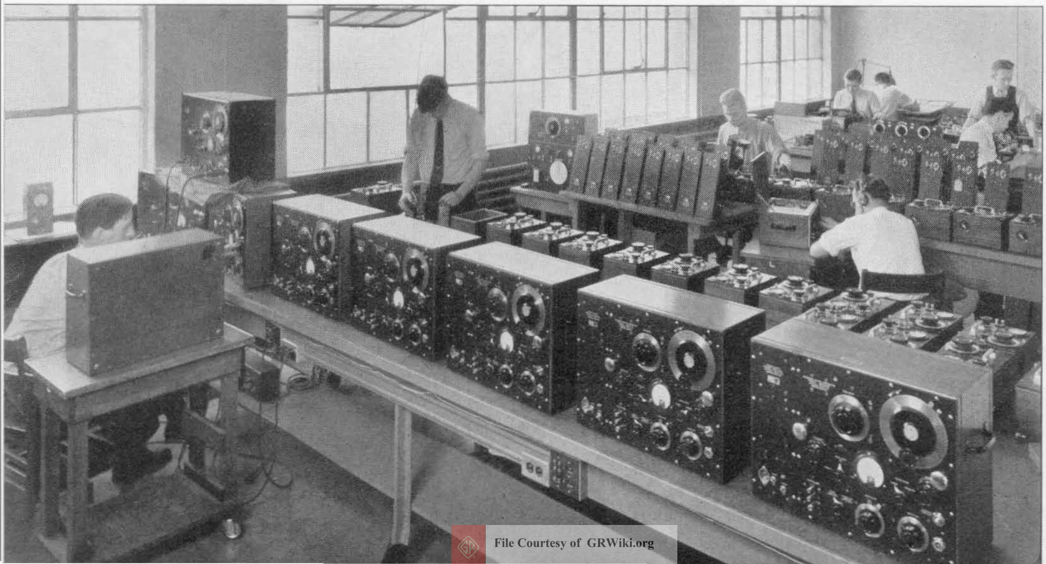
At the General Radio Company, the Standardizing Laboratory is the final link between the Company and the customer. Any errors or defects that may have occurred in manufacture must be caught and corrected in this laboratory. Instruments are then adjusted for optimum performance and calibrations are made. Obviously, the laboratory's job must be well done or customer dissatisfaction is bound to result.

The principal function of the Standardizing Laboratory is the performance of standard engineering-test and calibration operations as a routine production procedure. Closely allied to this are the inspection and test of component parts before their assembly into complete instruments. The complete production of piezo-electric quartz crystals is also carried on as an activity of the laboratory, since the preparation of these is primarily a precision calibration operation involving adjustments in terms of a known standard.

Capable personnel, adequate test equipment, and standardized test procedure are essential if a uniformly accurate and reliable product is to be turned out. Of the twenty men who test General Radio instruments, six have engineering degrees and ten others are graduates of engineering institutions. In addition to permanent laboratory personnel, the staff usually includes several students pursuing co-operative courses in electrical engineering at nearby engineering schools and colleges.

Some $25,000 worth of standard General Radio instruments are permanent equipment in the laboratory, including 3 Wave Analyzers, 6 Beat-Frequency

FIGURE 1. View of a portion of the Standardizing Laboratory, showing a group of TYPE 605 Standard-Signal Generators undergoing test and calibration.



Oscillators, 6 Vacuum-Tube Voltmeters, 7 Bridges, 8 Heterodyne-Frequency Meters, and at least one each of most other instruments in the General Radio catalog; and there are, of course, many special instruments and assemblies designed to meet specific test requirements.

Definitely prescribed test schedules are carried out on each instrument. The tests to be made are specified by the engineer responsible for the development of the instrument, and the detailed testing specifications are worked out jointly by the engineer in charge of testing and the development engineer.

The thoroughness of the test procedure followed on most instruments can be illustrated by listing the tests performed on a TYPE 736-A Wave Analyzer. In brief, these consist of

1. Over-all Inspection of Mechanical Assembly.
2. Input Power Measurement.
3. Adjustment of Plate Supply Voltage.
4. Adjustment of Oscillator Range.
5. Balance Adjustments on Detector.
6. Preliminary Adjustment of Crystal Filter to Obtain Desired Band-Pass Characteristics.
7. Detector Tuning Adjustment.
8. Final Crystal Filter Adjustments.
9. Over-all Gain Measurement.
10. Detector Distortion Measurement.
11. Attenuator Check.
12. Meter Check.
13. Input Multiplier Check.
14. Phase Inverter Linearity Test.
15. Frequency Response Measurement.
16. Absolute Voltage Calibration.
17. Hum Measurement.
18. Frequency Calibration.

These instruments are tested in groups of five, that is, each test is performed on each instrument of the group before the test man proceeds to the next test. In this way, more efficient use of time and equipment is possible than when the complete test schedule is performed on each instrument in turn. Yet the minimum test time per instrument is about six man hours, when no difficulties that require trouble shooting are encountered.

FIGURE 2. Calibrating a TYPE 722 Precision Condenser in the Standardizing Laboratory.

The activities of the Standardizing Laboratory, however, are not confined to production alone, but also touch engineering, sales, and service.

The engineering functions of the laboratory are threefold, embracing the determination of performance data on new instruments, the maintenance of standards, and the training of engineering assistants for the development engineering group.

Trial production lots of new instruments (usually numbering either five or ten units) are given particularly comprehensive tests. Over-all performance tests are made to determine catalog specifications; errors and omissions in manufac-

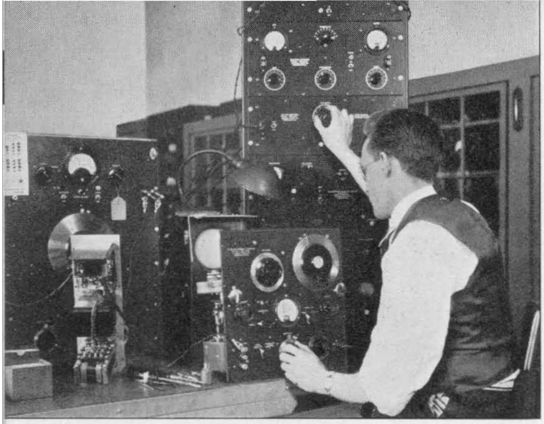


FIGURE 3. Modulation tests on a TYPE 605 Standard-Signal Generator.

turing specifications are corrected; and, finally, operating tests under severe temperature and humidity conditions are made.

In order to calibrate instruments for voltage, resistance, capacitance, inductance, and frequency, accurate standards are necessary. These are maintained in the laboratory with the exception of the Primary Standard of Frequency, which is located in the Engineering Department. All working standards of resistance, inductance, and capacitance are intercompared periodically, and several are sent yearly to the U. S. Bureau of Standards for recalibration.

Testing specifications allow in general about three-quarters of the tolerance given in the catalog specifications. That is, a resistor with a published accuracy of 0.1% is rejected by the laboratory if its error is over 0.075%.

The Standardizing Laboratory provides excellent training for engineering assistants, and, from time to time, laboratory personnel are assigned to the development engineering group for work under the supervision of development engineers.

Not so obvious, but extremely important is the laboratory's connection with sales. It is the responsibility of the laboratory administration to control inventories so that no more than four months' supply of major instruments is kept in the stockroom. This policy has two beneficial results. It permits laboratory time to be allotted most efficiently in terms of salable instruments, and it assures the customer that the calibration of any instrument that he purchases is no more than four months old. Some particularly critical calibrations, however, are made only upon receipt of a customer's order.

Each repaired instrument is completely tested and calibrated to the same specifications as a new instrument of the same type. Since the volume of repairs cannot be easily controlled, close cooperation between the laboratory and the Service Department is essential in order to avoid delays in returning repaired instruments to their owners.

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VOLUME XVI No. 1 JUNE, 1941

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

THE VIBRATION METER — A NEW ELECTRONIC TOOL FOR INDUSTRY

VIBRATION AND ITS MEASUREMENT

● **ALTHOUGH THE LOW-FREQUENCY LIMIT** of hearing for average individuals is about 16 cycles per second, frequencies lower than this can still cause considerable annoyance.

For instance, a man riding in an automobile may experience a sensation that he recognizes as either "noise" or "vibration," but which is really neither of these alone but a combination of the two. In the low-frequency region, vibrations may be felt at frequencies so low that they cannot be heard. This ability to feel the vibration through the sense of touch extends well up into the audible range and may be detected even

(*Continued on page 2*)

● **THE OFFICE OF PRODUCTION MANAGEMENT** requires that we make shipments on defense orders in accordance with their preference rating, and before shipments are made on any non-defense requirements for the same item. The great majority of orders now are for defense. Therefore, be sure to indicate the preference rating on all purchase orders. We will frequently have to request an official extension of the preference rating on the Priorities Division Form PD-3, but this will be done only when necessary for extension to subcontractors.

When an instrument is needed quickly we urge you to avoid the formality of asking for a quotation first before placing the order. The net prices, f.o.b. the factory, are given in the catalog or the *Experimenter*, and not infrequently we find that, although we can make delivery from stock when a quotation is made, the stock is exhausted when the order is received. That is why we are compelled under the present circumstances to make all delivery promises on quotations subject to prior sale.

at very high frequencies when the amplitude of vibration is sufficient.

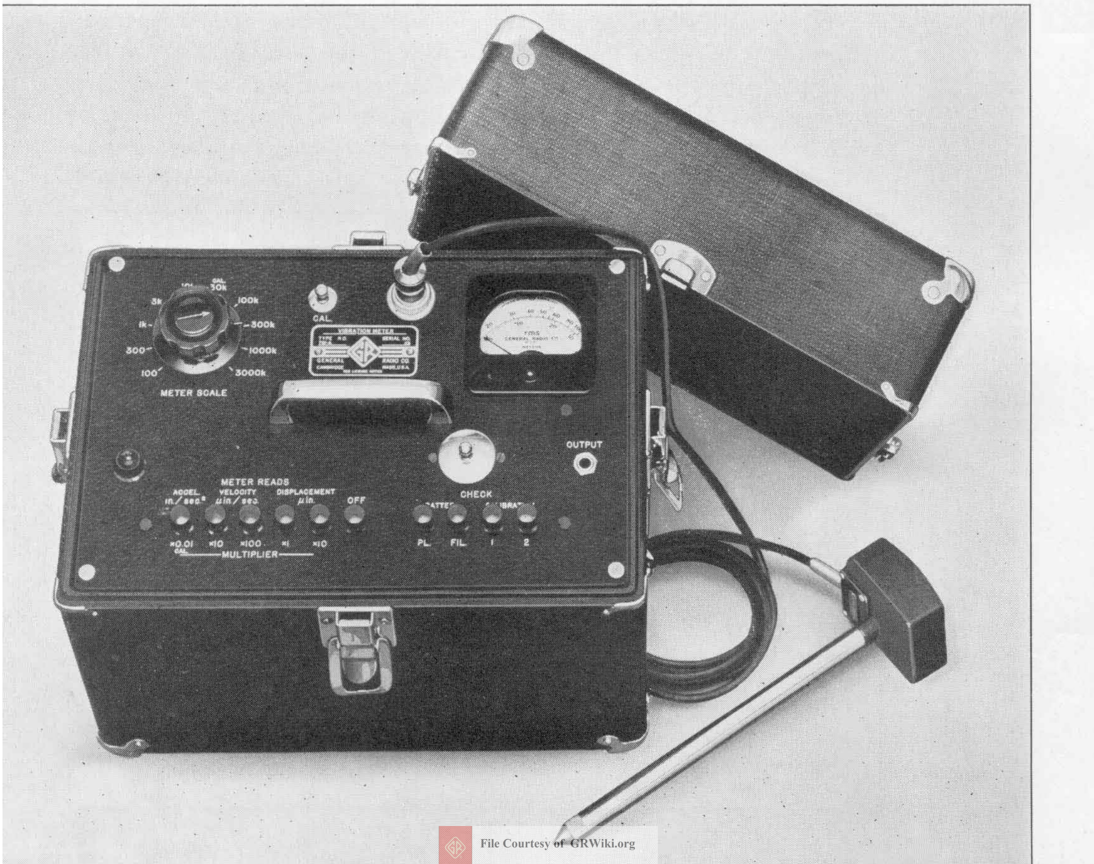
Sound, as generally encountered, is a vibration propagated through the air, but it is also transmitted through various other media and generally originates as an actual mechanical vibration in or of some solid structure. To trace noise to its source, therefore, through the many materials or mechanisms that may transmit it, some means of measuring vibration is desirable.

For some years sound and noise measurements have been increasing in importance to all branches of industry. A wide variety of mechanical devices, from clocks to automobiles, are checked with sound-level meters. The same type of meter is used for measurements on sound-absorbing and sound-insulating materials. Analyzers and various accessories have widened the scope of usefulness of this instrument, until today there are few engineering jobs which can be undertaken without some sort of sound-level measurements.

To adapt the sound-level meter to vibration measurement, a vibration pickup, replacing the microphone, can be used. Essentially, a vibration pickup is similar to a microphone, excepting that the pickup is designed to respond to solid-borne rather than air-borne vibrations. Hence, if the pickup is held in a sound field, the output, if any, will be relatively small compared to that of a microphone, but if it is held against some vibrating mechanical part, the output will be relatively high. This is essentially a matter of mechanical impedance and coupling.

The sound-level meter as a vibration-

FIGURE 1. View of the TYPE 761-A Vibration Meter with cover removed.



measuring instrument is limited in its frequency-response characteristics by the sound-level meter itself, and such instruments are seldom satisfactory below 25 cycles per second, although, for vibrations within the audible range, the vibration pickup has proved extremely useful. The need to extend the range of vibration measurements into the lower-frequency region, so that sub-audible vibrations can be measured, has led to the development of a new instrument, the Type 761-A Vibration Meter, which operates on the same principle as the sound-level meter.

The vibration meter as such does not replace the sound-level meter, but rather complements it. The two instruments cover overlapping frequency ranges, and throughout these ranges both types of measurement are useful. The vibration meter, like the sound-level meter, provides definite and reproducible readings, unaffected by human judgment, which are consequently invaluable for record and comparison purposes.

CHARACTERISTICS OF VIBRATION

The design of sound-level meters has been standardized and simplified through the general adoption by all manufacturers of the American Standards Association's tentative standards. These include frequency-response curves that approximate the characteristics of the human ear at various sound intensities, and, while these curves are not absolutely accurate for any particular individual, they represent an average of a large number of people. Hence the sound-level meter can be used not only to measure the physical characteristics of the sound, but also to approximate to a considerable degree the physiological and psychological effects.

The effects of vibration on human beings are less well known than those of sound, and, consequently, the main purpose of a vibration meter is to measure the actual physical magnitude of the vibration, while the problem of correlating such readings with the psychological and physiological effects is left to the user. It has not been found, however, that this presents any serious limitation to the use of vibration measurements, since for any particular problem the correlation between vibration and its undesirable effects can be determined relatively easily on an empirical basis.

The troubles arising from vibration can be generally divided into four classes, as follows: (1) noise, (2) annoyance, (3) deflection, (4) stress.

Under condition (1) the vibration is undesirable because of its resulting noise. This applies mainly to vibrations in the audible range and is the condition most generally associated in the public mind with the word "vibration."

Under the second condition the vibration is annoying in itself and is actually felt rather than heard, since it is mainly in the sub-audible range. This again is rather frequently encountered. A good illustration of this is found when standing on an upper deck at the stern of a steamship, where little noise is noticeable but where considerable vibration may generally be felt.

The third condition occurs when the vibration causes a deflection of certain members in the vibrating structure or mechanism, resulting in actual contact or striking, with consequent noise or breakage. This is a very severe condition, which obviously should never be encountered in finished mechanical designs, but it is frequently met in the laboratory where design work is in progress.

Under the fourth condition the actual

stresses set up in mechanical members of the vibrating machine cause a strain so severe as to exceed the elastic limits of certain mechanical parts, thus causing failure. This again is generally a design problem rather than one encountered in the field. Such conditions are usually encountered in equipment which must be built to the very closest limits of safety in order to save weight or size, as in airplane design.

Obviously, the characteristics of most simple vibrations can be satisfactorily described through measurements of amplitude and frequency, and in advance design work such measurements are generally required. In more simple cases, however, it is desirable to obtain a single figure which expresses reasonably well the characteristics of the vibration, and, with this idea in mind, various types of vibration-measuring devices have been used at one time or another. Such devices have generally read one of three characteristics of the vibration — namely, displacement, velocity, or acceleration — depending mainly upon the inherent characteristics of the particular vibration-measuring device.

In the design of a new and universal type of vibration meter, selection of any one of these three characteristics as standard did not seem desirable, any more than in the design of a sound-level meter only one weighting characteristic would be desirable. Under a particular condition any of the three types of vibration measurements might be the best.

The displacement of a vibration is, of course, merely a measure of the actual amplitude, without respect to frequency, and is of importance mainly under conditions (2) and (3), as outlined above. Under other circumstances some definite relationship between the frequency and the magnitude of the vibration is desirable in the reading, and the choice between velocity and acceleration characteristics is based upon both practical and mathematical considerations. The velocity, being the derivative of the displacement, differs from it by a factor proportional to the frequency, while the acceleration, being the second derivative of the displacement, differs from it by a factor propor-

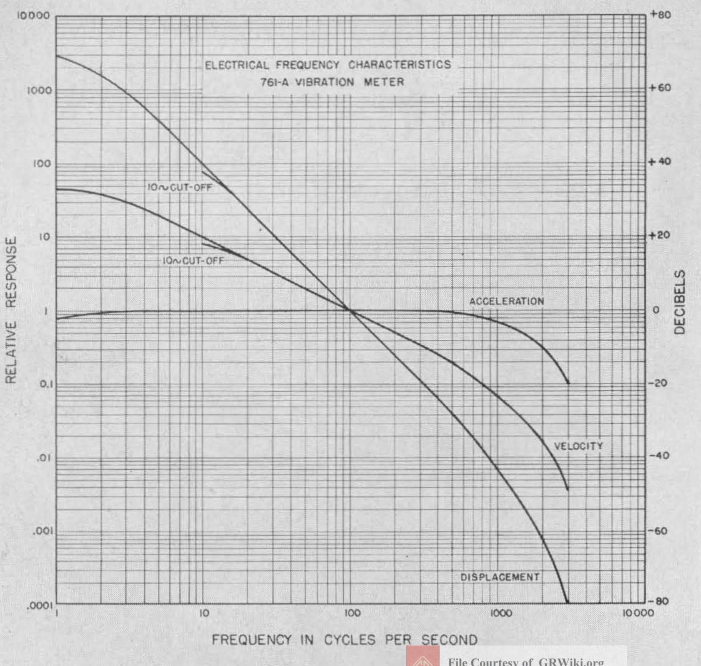


FIGURE 2. Frequency response of the electrical circuits in the TYPE 761-A Vibration Meter.

tional to the square of the frequency.

In particular, velocity measurements have been found most useful under conditions (1) and (2) and give a good indication of the actual amount of noise caused by a vibration when the radiating surfaces are large compared to the wavelength of the vibration. Acceleration measurements are useful under conditions (1) and (4). So far as noise is concerned [condition (1)], acceleration measurements are most valuable when the radiating surfaces are small compared to the wavelength of the sound. Since the stress in any particular part will vary directly with the acceleration, it is obvious that this is the type of measurement to be used under condition (4).

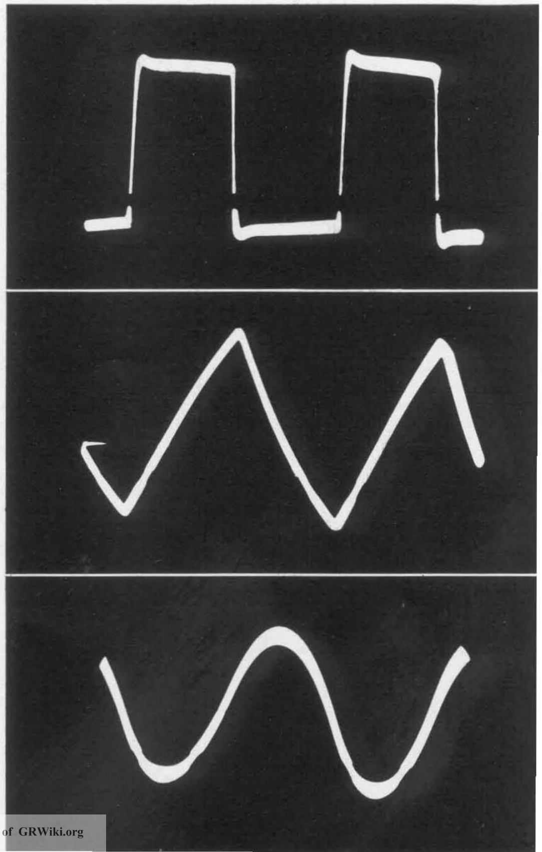
DESIGN FEATURES OF THE VIBRATION METER

Heretofore vibration meters have generally been designed with only one or two particular applications in mind. In many cases the design of the meter has been influenced more by what was readily achievable in the way of response characteristic, rather than by what was needed for actual practical use. Hence there have been vibration accelerometers, vibration velocity meters, and vibration displacement meters, but few if any instruments have combined all three of these characteristics.

The need for inclusion of all three characteristics was first met by a special control box and vibration pickup designed for use with the General Radio Sound-Level Meter. This equipment, which has been available for several years,[1] consists essentially of a two-stage integrating circuit, which, when operated in conjunction with an acceleration-type pickup, provides a choice of acceleration, velocity, or displacement response over the frequency range of the sound-level meter.

Of primary importance in the design of a new vibration meter was the inclusion of all three response characteristics and extension of the frequency range well below the audible limit, in order to cover the range in which most of the important machinery vibrations are found. This involved considerably more than a mere redesign of the sound-level meter, although all of the essential and

FIGURE 3. The action of the integrating circuits is shown by these oscillograms of the output voltage of the vibration meter for the three types of response when a square wave is applied to the input. At the top is shown the displacement response with wave shape unaltered; the velocity is shown at the center; and the lower oscillogram shows the acceleration.



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[1] The TYPE 759-P15 Vibration Pickup and TYPE 759-P16 Control Box are for use with the TYPE 759-A Sound-Level Meter, and the TYPE 759-P35 Vibration Pickup and TYPE 759-P36 Control Box for use with the TYPE 759-B Sound-Level Meter.

desirable features of its mechanical and electrical design have been retained. The new instrument has been designed for useful response down to a lower limit of 2 cycles per second, and the integrating circuits are built into the amplifier itself rather than added at the input, thus minimizing the effects of tube noise, microphonic vibration, etc., and making possible the use of the instrument over a wide range of vibration amplitudes.

Figure 2 shows the actual frequency response of the electrical circuits of the vibration meter and demonstrates more clearly than words the unusual features of the design. Since the vibration pickup is of the acceleration type, the response must vary inversely as the square of the frequency in order to provide an over-all flat displacement characteristic, and this type of response must be maintained from the lower limit of 2 cycles up above the range of the vibration pickup. The amplifier circuit used is similar to the self-stabilizing circuit already in use in the Types 759-A and 759-B Sound-Level Meters, but with such changes as are necessary to produce the extended low-frequency response and to provide stability and freedom from unwanted regeneration or degeneration in the extreme low-frequency ranges.

An interesting example of the function of the integrating circuits is shown in Figure 3, which shows the output voltage for the instrument when a square wave is applied to the input terminals for the three types of response. The actual over-all frequency response of the instrument, including the vibration-pickup unit, is shown in Figure 4.

Figure 1 shows the appearance of the complete vibration meter and is ample proof of the high degree of simplicity and convenience which have been attained. The vibration values are read directly from the scale of the indicating instrument, and the sensitivity is adjustable over a wide range by means of the knob marked METER SCALE. The actual range covered between the smallest vibration which can be detected and the greatest which the instrument will indicate is 300,000 : 1[2].

The row of push buttons at the lower

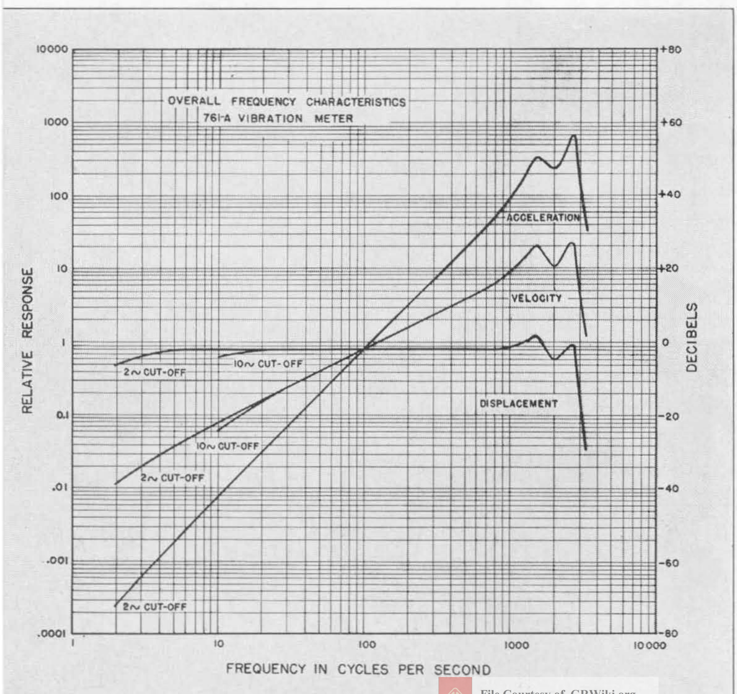

Figure 4. Over-all frequency response of the vibration meter, including the vibration pickup.

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[2] 3,000,000 : 1 for frequencies above 10 cycles.

edge of the panel provides all the additional controls necessary in the operation of the vibration meter. The meter is automatically turned on when any of the ACCELERATION, VELOCITY, or DISPLACEMENT buttons is depressed. These buttons control the response characteristic of the meter, as previously outlined. It will be noted that two each are provided for the velocity and displacement characteristics. In both of these cases the additional button provides an extra degree of sensitivity over what could otherwise be attained, through the expedient of limiting the low-frequency response below 10 cycles. The extra buttons, therefore, do not affect the normal operation of the instrument in any way, but provide an additional tenfold increase in sensitivity for use at vibration frequencies above 10 cycles. It will be noted that each of the buttons is marked with a multiplying factor to be applied to the meter readings when that particular button is used. The circuit values have been so chosen that the multiplying factors are all equal decimal units, so that multiplying becomes merely a matter of shifting a decimal point.

The four buttons at the right-hand end of the panel provide means for checking the accuracy of the instrument at any time. The PL and FIL buttons provide battery checks. For satisfactory operation it is merely necessary that the batteries provide a meter deflection above the B mark. The other two buttons, marked CALIBRATION, control an internal calibrating circuit which allows the amplifying circuit to be set accurately at any time to the original factory adjustment by comparing the gain with the loss in a fixed attenuator. In use, this is accomplished by connecting the instrument to a power line (60 cycles or any other power-line frequency), setting the other panel controls as indicated by the engraved words CAL, and pressing first CALIBRATION button No. 1 and then No. 2. If the gain in the amplifier is correct, the meter will show the same deflection when either button is depressed. If the two deflections differ, the sensitivity may be readily readjusted by means of a screw driver inserted in the opening just above the name plate. This calibration depends only upon the internal attenuator and is independent of the voltage or frequency of the power line.

The jack marked OUTPUT connects directly to a separate output amplifying stage entirely distinct from that which drives the indicating instrument. When the vibrations are in the audible range, a pair of phones plugged into this jack will provide a means of listening. The jack may also be used for connecting to an analyzer, recorder, or other auxiliary equipment. The external circuit connected to the output jack will have no effect on the operation of the meter.

Users of the Type 759 Sound-Level Meters will readily recognize similarities between those instruments and the new vibration meter. In the design of the new instrument full advantage has been taken of the experience gained in building hundreds of sound-level meters. The rubber suspension for the tube shelf, the compensated circuit to minimize the effects of battery voltages, the design of the attenuators, the shaped-pole-piece indicating instrument, etc., follow the same practice as in the widely used General Radio Sound-Level Meters. The same type of battery is used, and a similar type of airplane-luggage case. Provision is made in the cover of the case for holding the vibration pickup with spare tips and probe.

USE OF THE VIBRATION METER

In use the vibration meter has exceeded all expectations. The ease with which readings may be taken, the provision of the three characteristics — namely, acceleration, velocity, and displacement — and the extension of the range down to 2 cycles per second with substantially flat response characteristics provide a degree of flexibility hitherto unapproached in commercially available vibration-measuring apparatus. It has already been applied to a wide range of applications, in the manufacture and design of airplanes, automobiles, and various mechanical devices, both large and small, and in the study of vibrations in structures.

— H. H. Scott

The price of Type 761-A Vibration Meter is $260.00, net, f. o. b. Cambridge. Complete specifications will be sent upon request.

SERVICE AND MAINTENANCE NOTES TO BE AVAILABLE

● **CONSIDERABLE INTEREST** has been shown in the article in the September-October, 1940, *General Radio Experimenter* which outlined a maintenance and service program for General Radio instruments. The number of requests that have been received for maintenance notes on particular instruments suggests that such notes would be welcomed by most General Radio customers. As a consequence, we are preparing comprehensive maintenance and service notes for a number of our most commonly used instruments. We are planning to send these notes, as they become available, to those customers who indicate an interest in them. If service and maintenance notes are desired, please write to the Service Department, specifying the type and serial numbers of your instruments. To obtain the maximum benefit from this program, it would be helpful if you could suggest to us the individual or group in your organization to whom such notes will be most useful.

We believe that this program will eventually result in longer life, improved reliability, and lower service costs for your General Radio instruments and parts.

— H. H. Dawes

ERRATA—TYPE 774 COAXIAL CONNECTORS

● **TYPE 774-R1** and Type 774-R2 Patch Cords were incorrectly described in the April issue of the *Experimenter*. Type 774-R1 has a plug unit at one end and a jack at the other, while Type 774-R2 has a jack at each end.

GENERAL RADIO COMPANY

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BRANCH ENGINEERING OFFICES

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THE General Radio EXPERIMENTER

VOLUME XVI No. 2 JULY, 1941

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

IMPEDANCE BRIDGES ASSEMBLED FROM LABORATORY PARTS

● **ON MANY OCCASIONS** when the need for accurate impedance measurements arises, a commercial self-contained bridge is not available, and its purchase may not be economically justified. It is generally possible, however, using standard components that are readily available in the average laboratory, to set up bridge circuits on the bench that, at audio frequencies, will realize accuracies approaching those of commercial bridges over a wide range of impedances. It must be recognized, however, that some sacrifice of convenience and speed of measurement will necessarily be involved.

The unavoidable stray capacitances that are associated with such circuits lead many laboratory workers to view them with considerable suspicion, particularly for the measurement of high impedances or of small quadrature components. But, with shielded components, and with

(*Continued on page 2*)

PRIORITIES

● **SINCE A LARGE PROPORTION** of our products is now being shipped on national defense orders, deliveries are more and more frequently controlled by the priority rating of the order. When the equipment that you are purchasing from us is to be used directly or indirectly in the execution of any defense contract or subcontract, *be sure to indicate this fact on your purchase order, giving the prime contract or order number and the preference rating.*

This will greatly assist us in making deliveries in accordance with the need of the material for defense, and will assist you in getting better deliveries on urgently required material. This information is necessary to us in obtaining many of the raw materials from which the products are manufactured.

a good bridge transformer, such as the Type 578 Shielded Transformer, the disposition of the stray capacitances can be definitely controlled. The magnitudes of these capacitances can then be determined with considerable accuracy and allowance made for their effects so that the accuracy of measurement will depend mainly upon the accuracy of the standards and of the bridge arms.

On the following pages are described bench setups of a number of conventional bridge circuits using standard components, and the disposition and measurement of the various stray and residual circuit impedances are discussed.

CAPACITANCE BRIDGES

Figure 1 shows a generalized capacitance bridge circuit and the complete equations of balance. This circuit can represent any one of a number of well-known capacitance bridge circuits, depending upon the relative magnitudes of the various impedances shown. For example, if capacitances C_A and C_B are zero, we have the familiar series-resistance type of capacitance bridge, wherein the losses in the P arm are balanced by the resistance N in the standard arm. This circuit is used in the General Radio Type 740 Capacitance Test Bridge, with the resistance R_N calibrated directly in dissipation factor* at 60 cycles. It is also used in the Type 650-A Impedance Bridge. Here again the N arm resistance is calibrated to read directly in dissipation factor, in this case at 1000 cycles.

As another example, if C_N and C_P are made infinite, the network reduces to the parallel-resistance type capacitance bridge in which the standard and unknown arms are A and B. In this circuit the effective parallel resistance of the unknown capacitance is balanced by the resistance, R_A, in parallel with the standard capacitor, C_A.

If D_N and Q_B are small compared to D_P and Q_A the network assumes the form of a Schering bridge, which is characterized by the fact that the dissipation factor of the P arm is balanced by parallel capacitance in the opposite arm. This is the circuit used in the Type 716

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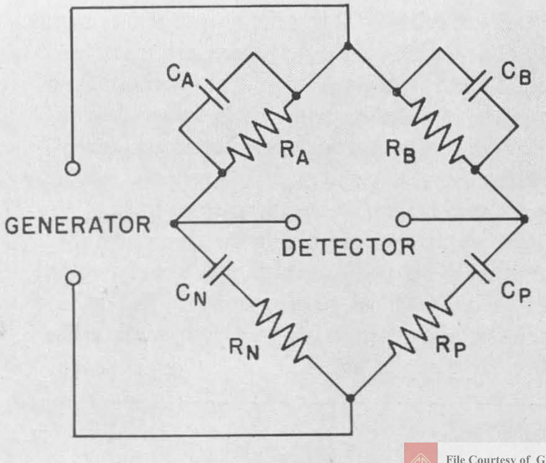
*The ratio of series resistance to series reactance, equal to $R\omega C$ for a condenser in series with a resistance. The symbol D, with appropriate subscripts, will be used throughout for this quantity.

The ratio of series reactance to series resistance, frequently called storage factor, will be designated as Q. For a capacitance in parallel with a resistance, we have $Q = R\omega C$.

Figure 1. Generalized capacitance bridge circuit, with equations of balance. Note that the expressions are written in terms of the dissipation and storage factors of the arms, so that frequency does not enter explicitly.



$Q_A = R_A\omega C_A$ $D_P = R_P\omega C_P$

$Q_B = R_B\omega C_B$ $D_N = R_N\omega C_N$

$$C_P = C_N\left(\frac{A}{B}\right)\frac{1+Q_B^2}{1+D_N(Q_B-Q_A)+Q_AQ_B}$$

$$D_P = \frac{D_N+Q_A-Q_B(1-Q_AD_N)}{1+D_N(Q_B-Q_A)+Q_AQ_B}$$

Capacitance Bridge, a highly precise direct-reading bridge for the measurement of capacitance and dissipation factor.

The fundamental equations of balance shown hold for all the reduced circuits mentioned, and, if accurate results are to be obtained, the complete expressions must be retained and examined for the effect of stray capacitances and other residual impedances in the bridge arms.

GENERAL CONSIDERATIONS

In the circuit shown in Figure 1 the generator and detector terminals are shown merely as two pairs of terminals brought out from opposite corners of the bridge. So far as the bridge balance equations are concerned, it is immaterial whether generator and detector are connected as shown or are interchanged. Their location is usually governed by considerations of sensitivity, the connections being made in the manner which yields the maximum output voltage for a given input voltage and a given unbalance of the bridge.

Inasmuch as high-input-impedance amplifiers preceding the null detector are almost universally used in a-c bridge measurements, the sensitivity problem is best discussed on the basis of the use of a detector of infinite impedance. It can be shown[1] that for a given unbalance of the bridge, the ratio of open-circuit output voltage to input voltage (with the generator across a pair of resistive arms) is

$$(1)\qquad \frac{E_o}{E_i} = \frac{\dfrac{A}{B}}{\left(1+\dfrac{A}{B}\right)^2}\, d$$

where A and B are the resistance of the arms across which the generator is connected, and d is the fractional change in the unknown from the condition of true balance. If the generator is connected across unlike bridge arms (one resistive, the other reactive) Equation (1) becomes

$$(2)\qquad \frac{E_o}{E_i} = \frac{\dfrac{A}{B}}{1+\left(\dfrac{A}{B}\right)^2}\, d$$

where either A or B is a reactance.

Either the detector may be grounded and the generator connected to the bridge through a shielded transformer or vice versa. When the detector is grounded, its capacitance to ground becomes part of its own terminal impedance, and does not affect the bridge balance. The terminal-to-ground capacitances of the generator are replaced by those of the shielded transformer. Although these capacitances appear across the bridge arms,* they are small, localized, and measurable; and allowance can be made for their effects on the measurement. When the generator is grounded, the transformer is used between the detector and the bridge.

Two of the points at which the bridge may be grounded leave both sides of the unknown above ground potential, while grounding either of the other two points grounds one side of the unknown capacitance. The choice of the point of grounding will depend somewhat upon the type of measurement that it is desired to make. For capacitors that are physically large, and hence subject to electrostatic pickup and stray capacitance effects, or for unknowns that have one side normally grounded, it is preferable to ground the bridge at one of the unknown terminals. On the other hand, if both unknown terminals are above ground it is possible to measure direct

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*By the use of Wagner grounds or guard circuits the effect of these impedances may be removed from the measuring circuit. This subject will be treated in some detail in a later issue.

capacitance between any two terminals of a three-terminal capacitance.

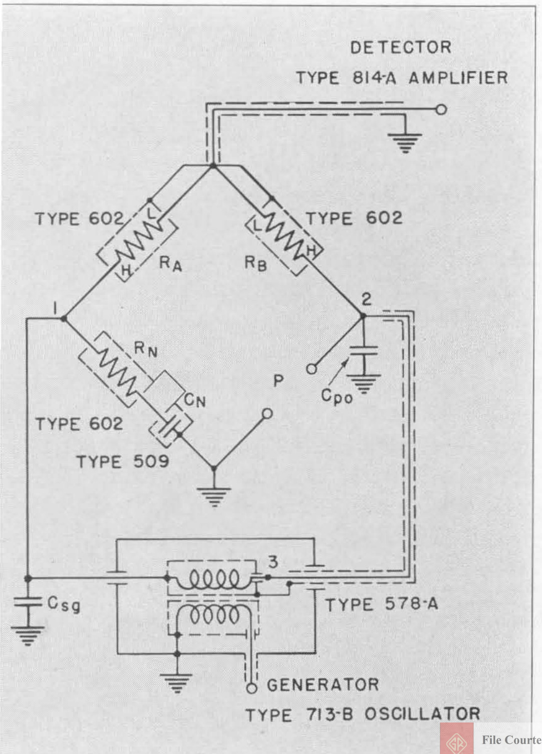
Extraneous coupling from the voltage source* to the bridge arms, to the unknown, or to the ungrounded detector terminals can cause serious errors in direct-reading measurements, and second-order errors in substitution measurements. Difficulties from this cause can be largely overcome by using well shielded oscillators and amplifiers and connecting them to the bridge with shielded leads.

In order to obtain maximum sensitivity and ease of balance it is desirable to use a tuned detector to eliminate harmonics, hum, and residual noise, which may otherwise obscure the fundamental balance and make precise balances difficult or impossible.

FIGURE 2. Connections for a bench layout of a series-resistance type of capacitance bridge. The approximate equations of balance are

$$C_P = C_N \frac{A}{B}$$

$$D_P = D_N + Q_A - Q_B$$



THE SERIES-RESISTANCE BRIDGE

Perhaps the simplest and most straightforward of the bridges mentioned is the series-resistance bridge, using variable ratio arms to balance the unknown capacitance against a fixed standard and a variable resistance in series with the standard condenser to balance the losses in the unknown arm. The arrangement illustrated in Figure 2, using TYPE 602 Decade-Resistance Boxes, a TYPE 578-A Transformer, and a TYPE 509 Standard Condenser has been found very satisfactory for measurements over a wide range of capacitance and power factor.

With this bridge, accurate results can be obtained if the various circuit and circuit-element residuals are measured and their effects on the capacitive and resistive balances computed.

CAPACITIVE BALANCE

Circuit Residuals

The capacitance C_{sg}, consisting of inter-shield capacitances of the transformer winding and the capacitance or point (1) to ground, appears across the *N* arm and causes an error depending directly on the ratio of its magnitude to that of the standard. Across the unknown arm *P* is placed C_{Po}, which is the capacitance of point (2) to ground.

The magnitudes of the two capacitances thus placed across the lower arms of the bridge can be determined quite accurately by balancing the bridge with C_{sg} connected alternately across the *N* and *P* arms. With the connections shown, C_{Po} is measured directly (if the standard

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*The effect of the extraneous voltage introduced into the bridge circuit by a capacitance or resistance coupling between the windings of the transformer is particularly interesting and will be discussed in some detail in a later issue.

condenser is large compared to C_{sg}); with the leads from the transformer reversed at the bridge, a value for C_{sg} is obtained that will be in error by the amount of capacitance contributed by the point (1) to ground. With careful wiring and arrangement of components, this capacitance will be small. The error in the measurement of these quantities probably does not exceed a micromicrofarad for C_{Po} and five micromicrofarads for C_{sg}. Typical values are 100 $\mu\mu$f for C_{sg} and 10 $\mu\mu$f for C_{Po}. In general, it is probably more desirable to use the circuit with C_{sg} across the N arm. If the capacitance of the standard is 0.01 μf, the resulting correction for C_{sg} will be of the order of 1%, subject to an error of only a few hundredths per cent.

INTERNAL RESIDUALS IN STANDARDS

Next in order of importance are the terminal-to-shield capacitances of the decade-resistance boxes. The TYPE 602 Decade-Resistance Boxes are completely shielded, with a separate shield terminal provided on the panel. Figure 3 represents, to a first approximation, the shielded decade box and the associated capacitances. The terminal connected to the highest resistance decade is designated as *H*, while the terminal connected to the lowest resistance decade, and located nearest to the shield terminal, is designated as *L*.

The terminal capacitances C_L and C_H can be measured directly with the bridge, of which the decade box is a part. Consider the ratio arm *B* of Figure 2. It is shown with the *L* terminal and the shield connected to the junction of the ratio arms. Hence C_L is short-circuited, and C_H parallels the resistance R_B and does not affect the capacitance balance. If the bridge is balanced with the *P* arm open circuited a value for C_{Po} will be obtained, as previously. If the shield is connected to ground, C_H is thrown across the *P* arm, in parallel with C_{Po}, and the bridge at balance will indicate the value of $C_H + C_{Po}$. The difference of the two readings is clearly C_H. By interchanging the *L* and *H* terminals of the box and following the same procedure, the value of C_L is obtained.

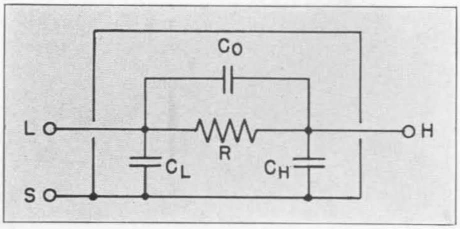

FIGURE 3. Simple equivalent circuit of a TYPE 602 Decade-Resistance Box with its associated capacitances. The panel terminal located nearest to the shield terminal is designated as *L*.

A series of measurements on a TYPE 602-L Decade-Resistance Box (a four-dial box with a maximum resistance of 111,100 ohms) show that, while C_L and C_H both vary with the resistance setting of the decade, their sum is constant. Comparison of these data with previous studies[2,3] of the TYPE 602 indicates that the capacitance C_o is of the order of 5 $\mu\mu$f, and at audio frequencies is practically constant, so that the major portion of the capacitance shunting R_B is C_L or C_H, depending upon which terminal of *R* is tied to the shield.

Let us examine the standard arm *N* of Figure 2. With the connections as shown the shields of both the decade box and the standard condenser are grounded, the capacitance C_L of the decade is thrown across the standard, and C_H parallels the series combination of R_N and $(C_N + C_L)$. It can be shown, however, that if C_N is large compared to $C_H +$

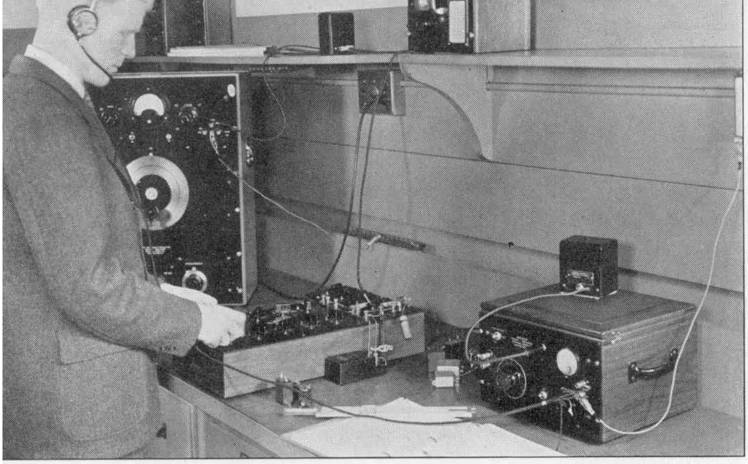

FIGURE 4. View of the series-resistance capacitance bridge assembled on a laboratory bench. The TYPE 578-A Transformer is on the shelf above the bridge.

C_L, and if the resistance of the arm is not large compared to the reactance of C_N, then the arm can be considered as a capacitance equal to $C_N + C_H + C_L$ in series with R_N. Now, knowing C_{sg} from the measurements previously outlined, and measuring $C_L + C_H$ as described above (or assuming a nominal value of 15 $\mu\mu$f per decade) we have the capacitance of the standard arm to an accuracy that will be limited solely by the accuracy to which we know the value of the standard condenser itself.

There remains the effect of the terminal-to-shield capacitances placed across the ratio arms. With the shields of the ratio-arm decade boxes connected to the junction of A and B the capacitance C_H of each arm is thrown in parallel with that arm, and the capacitance of the shield to ground is placed across the detector terminals, where it is harmless. As may be seen from an inspection of the complete balance equations of Figure 1, capacitance in parallel with a ratio arm has only a second-order effect on the capacitance balance, provided that the resultant Q's of the ratio arms are small. At a frequency of 1000 cycles the Q of a decade box set at 100,000 ohms is less than 0.01 and can safely be ignored so far as the capacitance balance is concerned.

DISSIPATION-FACTOR BALANCE

The capacitances across the ratio arms have an important effect on the dissipation-factor balance. To a first approximation the equation governing this component of the balance may be written as

$$D_P = D_N + Q_A - Q_B. \qquad (3)$$

It is, of course, desired to compute D_P from the equation $D_P = D_N = R_N \omega C_N$. To be able to do this with reasonable accuracy it is necessary to keep the difference $Q_A - Q_B$ small with respect to D_P.

At this point it may be well to discuss the effect of small residual inductances in the ratio arms. Figure 5 is a representation of a resistance with inductance in series, shunted by a capacitance. The Q of this circuit can be expressed as

$$Q = \frac{\omega L}{R} - R\omega C \qquad (4)$$

provided that $\omega^2 LC \ll 1$. Hence the series inductance introduces a Q term opposite in sign to that introduced by the shunt capacitance,* and the equations previously written may still be used by substituting for Q_A and Q_B their values modified by the presence of the series

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*Grover has suggested a bridge utilizing series inductance in the ratio arms to balance for dissipation factor. Since the inductive Q is negative in terms of the equations we have written, resistance in the P arm is balanced by series inductance in the B arm.

inductance. For high resistance settings of a decade box[2,3] the inductive Q is generally completely negligible compared with the Q contributed by the shunt capacitance. At low settings, however, the inductance will predominate, and the error in the dissipation factor due to that particular arm will change its sign.

From Equation (3) it is clear that the accuracy that can be obtained by computing the dissipation factor directly from the value indicated by the setting of R_N will be limited by the sum of the effective residual Q's and D's of the network, so that $Q_A - Q_B + D_{No}$ (where D_{No} is the residual dissipation factor of the standard arm), will be the error encountered. We must allow for the possibility that all three terms are additive in the worst case. What then is the order of magnitude of the error that may be expected?

A good mica standard in the N arm will of itself have a power factor less than 0.0005, but this may be somewhat increased by the losses in C_{sg} shunting it. At one kilocycle the Q of a TYPE 602-J Decade Box (for instance) will vary roughly from -0.0008 at 10,000 ohms to $+0.0007$ at 10 ohms, so that the difference $Q_A - Q_B$ may range from zero to 0.0015. The maximum error to be expected, then, in measuring the total dissipation factor of the P arm, will be of the order of 0.0025. If the ratio arms are nearly equal, however, the difference between their Q's will be small, and the maximum value of $Q_A - Q_B + D_{No}$ may be as low as 0.001. If $Q_A - Q_B$ be made equal to D_{No} the error will approach zero. This condition may be approached by connecting a capacitor of known dissipation factor in the P arm, setting the resistance box in the N arm to the proper value, and balancing the bridge by means of a variable condenser across the A or B arm, as may be required. A known resistance in series with a known capacitance, or an air condenser of extremely small dissipation factor, may be used in the P arm.

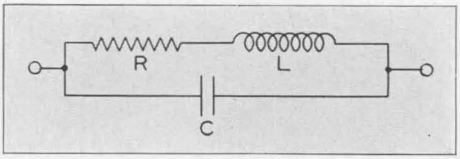


FIGURE 5. Simple representation of a resistor having residual inductance and capacitance.

This latter procedure must be followed through for each setting of R_B that is to be used and is a relatively long and tedious procedure that is generally not justified. If it is followed with care, however, the only remaining error is that caused by the variation of the Q of the A arm (something of the order of ± 0.0005 at 1000 cycles).

In all cases, if best accuracy is desired, two balances must be made, one with the unknown disconnected. Allowance for the initial capacitance and dissipation factor of the bridge can easily be made, as follows:

$$C_X = C_2 - C_1$$

$$D_X = \frac{D_2C_2 - D_1C_1}{C_X}$$

If accurate measurements of small power factors are desired, however, it is better to use a substitution method with a good calibrated air condenser in the N arm, either in the circuit described above or in the Schering bridge arrangement to be discussed in a later issue.

— IVAN G. EASTON

REFERENCES

[1] Radio Engineering Handbook, 3rd Edition, page 179, section 7.

[2] D. B. Sinclair, "Radio-Frequency Characteristics of Decade Resistors"— General Radio *Experimenter*, Vol. XV, No. 6, December, 1940.

[3] R. F. Field, "Frequency Characteristics"— General Radio *Experimenter*, Vol. VI, No. 10, February, 1932.

A MULTIPLIER FOR THE VACUUM-TUBE VOLTMETER

● **THE TYPE 726-P1 MULTIPLIER** was developed after the publication of our current catalog and, consequently, many recent purchasers of the TYPE 726-A Vacuum-Tube Voltmeter may not be aware that a multiplier is available.

This accessory to the vacuum-tube voltmeter, which was described in the May, 1940, issue of the *Experimenter*, extends the upper limit of voltage measurement to 1500 volts.

The multiplier is a capacitive voltage divider with a ratio of 10 : 1, and is intended primarily for use at frequencies above one megacycle. The frequency error is shown in Figure 1; the mechanical features in Figure 2.

Type		*Code Word*	*Price*
726-P1	Multiplier	ALOUD	**$15.00**

Delivery can be made from stock.

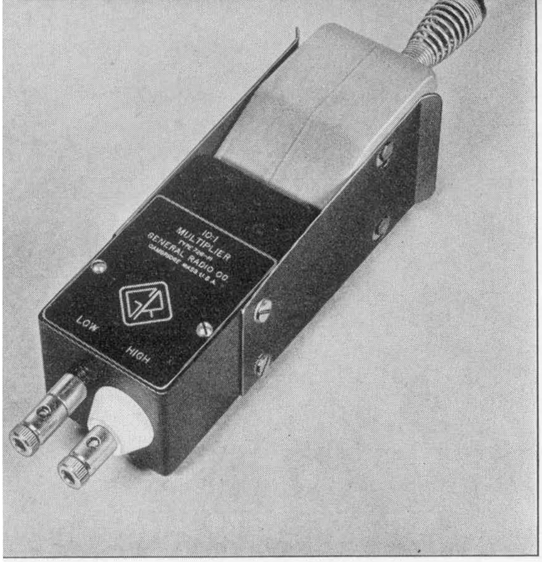
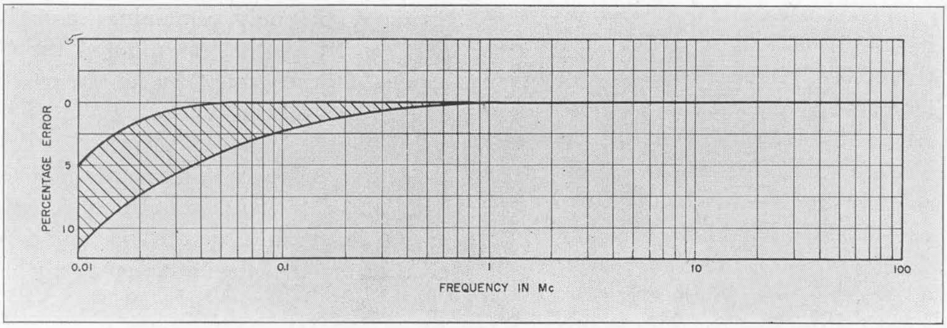


FIGURE 1 (*above*). View of the TYPE 726-P1 Multiplier with the voltmeter probe plugged in.

FIGURE 2 (*below*). Plot of the error in multiplier ratio as a function of frequency. The input admittance of the multiplier is equivalent (between 100 kc and 100 Mc) to that of a 4.5 $\mu\mu$f condenser of less than 0.5% power factor. The shaded area shows the variation in low-frequency error with different voltmeters.



SERVICE AND MAINTENANCE NOTES

● **WE HAVE RECEIVED** so many requests for service and maintenance notes that it is impossible for us to acknowledge them individually. The notes are now in preparation and will be mailed early in the fall.

GENERAL RADIO COMPANY
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THE General Radio EXPERIMENTER

VOLUME XVI No. 3 AUGUST, 1941

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

Also

IN THIS ISSUE

● **THE GREATLY INCREASED DEMAND** for equipment for the defense program made it necessary to call to the attention of our customers through these columns the fact that there would be delays in the supplying of some equipment. Fortunately, due to fairly large stocks that had been built up in anticipation of these requirements, and also due to good stocks of basic materials, these delays have been kept to a minimum.

A new situation has now arisen, and it has become quite acute. Many of the materials going into the manufacture of our products are on the mandatory priority list. This means that in order to obtain such materials we must furnish a preference rating to our suppliers. This is obviously a matter over which we do not have control. It means essentially that we are in a position to fill orders only when the items are in some manner associated with the defense program. Therefore, BE SURE TO SHOW DEFENSE CONTRACT NUMBER AND PREFERENCE RATING ON ALL PURCHASE ORDERS AND INQUIRIES.

The Defense Supplies Rating Plan of the Office of Production Management allows us to purchase certain scarce materials under an A-10 priority rating when the material is to be used in the manufacture of General Radio equipment entering into the National Defense Program.

Under this plan, documentary evidence is required to show that the products are almost entirely used for defense. Consequently, we are asking our customers to supply us with affidavits stating the percentage of purchases made by them for defense uses month by month. You may already have received the first of these affidavit forms. They are very simple in form, and we hope that you will return them promptly, as it is only by this means that we can continue to maintain a smooth flow of production to meet your requirements.

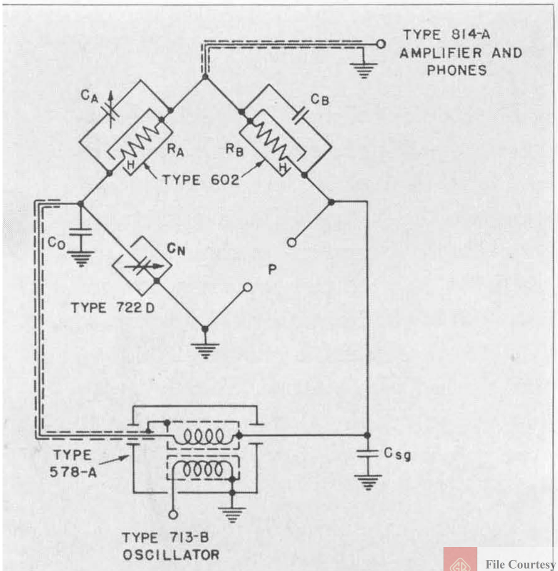
IMPEDANCE BRIDGES ASSEMBLED FROM LABORATORY PARTS

PART II THE SCHERING CIRCUIT

● **IN LAST MONTH'S EXPERIMENTER** a bench layout of a series-resistance type of capacitance bridge using standard laboratory components was described. An alternative method of balancing the dissipation factor of the unknown arm is the Schering method, which utilizes a capacitance across the opposite ratio arm. This arrangement was first used by Thomas and, following the later work of Schering, came into wide use for the measurement of dielectric constant and power factor of insulating materials and dielectrics.

The advantage of this circuit over the series-resistance and parallel resistance types lies largely in the fact that an initial dissipation factor balance can be easily established, so that the dissipation factor of the unknown can be read as an *increment* of a calibrated condenser. A variable air condenser is commonly used in the *N* arm, one ratio arm being fixed and the other variable in decade steps. The maximum error that will occur if the initial balances are properly established, and if the condenser C_A is correctly calibrated in capacitance differences, will be due to the variation in dissipation factor of the standard capacitance as its setting is varied. This variation, for the Type 722 Precision Condenser, is less than 0.0004.

Figure 1. Connections for a bench layout of a Schering bridge.



In the design of commercial, shielded, self-contained Schering bridges it is customary to place the terminal capacitances of the shielded transformer across the ratio arms, where they are relatively innocuous, becoming a part of the initial dissipation factor balance. In a bench layout, however, this is generally not possible. Figure 1 shows an arrangement wherein the terminal capacitances are placed across the capacitance arms, in the same manner as in the series-resistance bridge.

A Type 722-D Precision Condenser (direct reading in capacitance from 100 to 1100 $\mu\mu$f, and from 25 to 110 $\mu\mu$f) is ideal for use as the capacitance standard. Any shielded air condenser, such as the Type 539-A, may be used for dissipation factor balance. An accurate capacitance calibration of this condenser can be obtained by a direct substitution method, with the Type 539 Condenser placed across the standard condenser, and with an appropriate condenser in the *P* arm for balance. Any variable air condenser of suitable size can be used for the dissipation factor balance while C_A is being calibrated.

As shown, the larger terminal capacitance is placed across the unknown arm while the smaller (5–10 $\mu\mu$f) is placed across the standard arm. Both of these terminal capacitances can be measured with good accuracy, as follows:

To measure C_o, the capacitance across the standard arm (C_N), balance the bridge with the ratio arms set equal and with a balancing condenser of from 100–1000 $\mu\mu$f in the P arm. Change the setting of the A arm and obtain a second balance. Denoting the readings of the standard condenser by C_{N1} and C_{N2}, respectively, the value of C_o is given by

$$C_o = \frac{C_{N_1} - \frac{R_A}{R_B} C_{N_2}}{\frac{R_A}{R_B} - 1} \quad \ldots\ldots (1)$$

The initial capacitance (C_{SG}) across the unknown arm is obtained by a direct measurement, it merely being necessary to add to C_N the value of C_o just obtained. That is

$$C_{SG} = \frac{R_A}{R_B}(C_N + C_o) \quad \ldots\ldots (2)$$

After these quantities have been determined, a known constant capacitance can be added to the standard arm reading at all times. Similarly, the zero capacitance (approximately 100 $\mu\mu$f) must be subtracted from the measured value of the unknown.

The dissipation factor of the zero capacitance across the P arm can be determined as follows: Establish an initial balance (C_{SG}, D_{SG}) with the P arm open, setting C_A at some large value (an additional condenser across the B arm will be necessary). Place a 1000 $\mu\mu$f condenser in the P arm, and rebalance the bridge by means of C_N and C_A. The dissipation factor of the zero capacitance will then be given by

$$D_{SG} = \frac{\Delta Q_A - \frac{D_1 C_1}{C_2}}{1 - \frac{C_{SG}}{C_2}} \ldots\ldots (3)$$

In this equation, C_{SG} is the zero capacitance, C_2 is the total P arm capacitance, and D_1 is the dissipation factor of the condenser C_1 placed across the P arm. If a good condenser is used for C_1, the ratio D_1C_1/C_2 will be very small and can be neglected.

To establish the initial balance for any setting of the ratio arms it is merely necessary to connect a capacitance of known dissipation factor (best obtained by a known resistance in series with a known capacitance) in the P arm, to set C_A to the proper value, and to balance by means of C_N and an additional condenser in parallel with B. After a reference point on C_A has been thus determined, it will be possible to compute D_P directly from any setting of C_A, or, at any given frequency, C_A can be calibrated directly in dissipation factor. The accuracy of measurement will be limited by (1) the accuracy of calibration of C_A, by (2) the accuracy with which the reference dissipation factor was known, and by (3) the variation in the dissipation factor of the TYPE 722 as its setting is varied. A direct-reading accuracy better than 0.001 (at 1000 cycles) can be attained if care is taken.

SUBSTITUTION METHOD

For the measurement of capacitance below 1000 $\mu\mu$f, excellent accuracy for dissipation factor as well as capacitance can be attained by the use of a substitution method in the standard arm. Although the procedure for this type of measurement is fairly obvious, it is outlined here because of its importance.

An initial balance is made with a capacitance of about 1000 $\mu\mu$f in the P arm, with the unknown in place and connected to the standard on the grounded side. The high-potential connection should be made with a bare self-supporting lead. For the initial balance this lead should be left ¼″ or ½″ from the high terminal of the unknown. The condenser C_A must be set at some fairly large value, since its setting must be reduced when the unknown is connected. When the initial balance is made and the settings of C_A and C_N noted, the unknown is connected and the bridge rebalanced. Denoting the new settings by C_N' and C_A', and the increments by ΔC_N and ΔC_A we have (provided D_X is less than about 10%)*

$$C_X = C_N - C_N' = \Delta C_N \ldots . (4)$$

$$D_X = R_A \omega \Delta C_A \left(\frac{C_N}{C_X} \right) \ldots\ldots . (5)$$

The capacitance so determined will be accurate within ±2 $\mu\mu$f, a possible maximum error of ±1 $\mu\mu$f for each capacitance setting. The possible error in D will range from 0.00005 for a 1000 $\mu\mu$f condenser to 0.0005 for a 100 $\mu\mu$f condenser.

If the LOW section (25–110 $\mu\mu$f) of the TYPE 722-D is used, capacitances up to 85 $\mu\mu$f can be measured to ±0.4 $\mu\mu$f, and the errors in D become 0.00005 for an 85 $\mu\mu$f condenser, 0.0005 for 10 $\mu\mu$f.

—IVAN G. EASTON

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*The complete expressions are:

$$C_X = \Delta C_N \frac{1 + \Delta^2 D \left(\frac{C_N}{\Delta C_N} \right)^2}{1 - \Delta^2 D \left(\frac{C_N}{\Delta C_N} \right)}$$

$$D_X = A \omega \Delta C_A \left(\frac{C_N}{C_X} \right) \frac{1}{1 - \Delta^2 D \left(\frac{C_N}{\Delta C_N} \right)}$$

Next month's installment of Mr. Easton's article will include numerical examples illustrating the order of magnitude of the stray impedances, and their calculation, as well as typical measurements with both the series resistance and the Schering bridges. — **EDITOR**

THE TYPE 757-A ULTRA-HIGH-FREQUENCY OSCILLATOR

(150 TO 600 MEGACYCLES)

● **THE RECENT EXTENSIVE APPLICATION** of ultra-high frequencies to communications and allied fields has made desirable the development for those frequencies of commercially available measuring instruments that approach in convenience and reliability the more common instruments operating at lower frequencies. Power sources have logically been first in order for this development, since no physical experiment can be done at a given frequency until energy at that frequency is available. The early work on power sources was naturally directed toward simply obtaining an oscillator that supplied some power at the desired frequency without much regard for its other characteristics. More recently the technique at ultra-high frequencies has reached the point where power sources of a specialized nature can be developed, and primary emphasis can be placed on some highly desired characteristic, such as large power output, a high degree of frequency stability, or small size. In addition, general-purpose oscillators that satisfactorily effect a compromise among a wide variety of desirable characteristics have resulted from the application of this advanced technique, and descriptions of these oscillators have been given in the literature.[1] In this type of oscil-

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[1] L. S. Nergaard, "Electrical Measurements at Wave Lengths Less Than Two Meters," Proc. I.R.E., September, 1936, Vol. 24, No. 9, pp. 1209–1211.

W. L. Barrow, "An Oscillator for Ultra-High Frequencies," Review of Scientific Instruments, June, 1938, Vol. 9, No. 6, pp. 170–174.

Ronold King, "A Variable Oscillator for Ultra-High Frequency Measurement," Review of Scientific Instruments, November, 1939, Vol. 10, No. 11, pp. 325–331.

lator, emphasis is given to the ability to deliver energy over a wide frequency range, a characteristic that is usually difficult to obtain for ultra-high-frequency oscillators unless a series of adjustments is made for each operating frequency.

The General Radio Type 757-A Ultra-High-Frequency Oscillator, shown in Figure 1, is an oscillator of this general-purpose type with a convenience for use hitherto unattained in commercial oscillators of its frequency range. The oscillator can be set by means of a single crank control to any frequency between 150 and 600 megacycles (a range of wavelengths in air of 200 cm to 50 cm). In spite of this wide range of frequencies, the lead-screw drive allows small frequency increments to be easily obtained, a feature that is necessary in using a power source for measurements of phenomena highly dependent on frequency.

The means by which the convenience and the excellent general characteristics are obtained can be seen from Figures 2 and 3. Figure 2 is a cut-away view which illustrates the simplicity of the oscillator and shows certain mechanical details; for instance, the long, horizontal, lead-screw drive; the piston with its spring contacts; and the outer containing cylinder, which serves as the

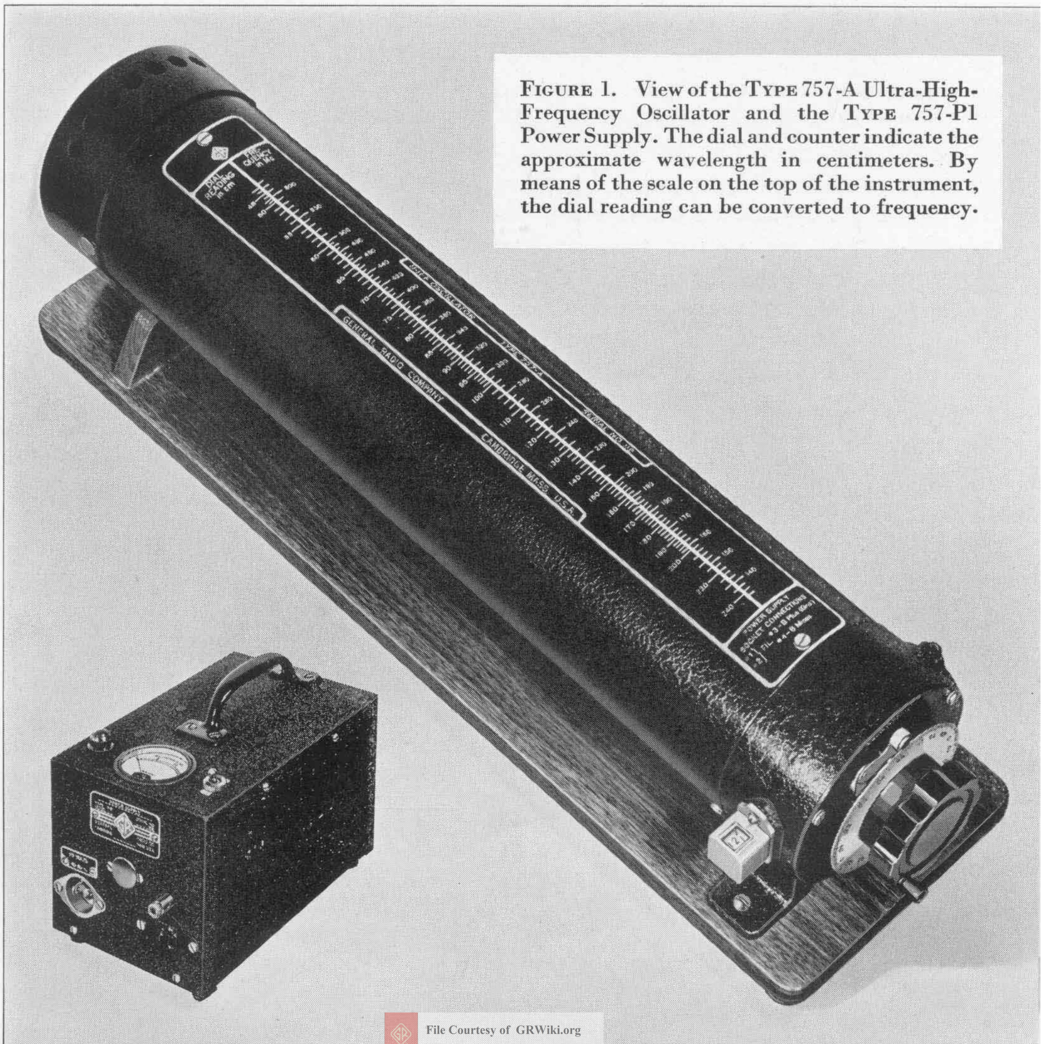

FIGURE 1. View of the Type 757-A Ultra-High-Frequency Oscillator and the Type 757-P1 Power Supply. The dial and counter indicate the approximate wavelength in centimeters. By means of the scale on the top of the instrument, the dial reading can be converted to frequency.

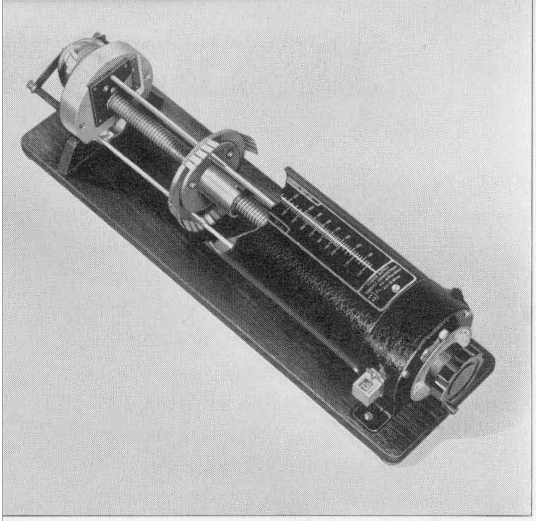

FIGURE 2. Cut-away view of TYPE 757-A Ultra-High-Frequency Oscillator, showing the internal construction.

radio-frequency shield, carries the lengthwise, square key for the keyway of the piston plunger, and forms with the lead screw a coaxial line that is a principal part of the oscillatory circuit.

In order to illustrate the electrical circuit of the oscillator, Figure 3 has been drawn as a simplified schematic diagram. The coaxial line between the oscillator tube and the short-circuiting piston is energized by the tube, a WE-316-A, and variation of the length of this section of the line produces the desired change in wavelength and frequency. This variation is obtained by rotation of the lead screw, which in turn drives the short-circuiting piston along the length of the coaxial line. Backlash in this drive mechanism has been reduced by the use of a double-nut spring take-up in the piston. The pitch of the lead screw and the ratio of the gearing to a counter at the drive end have been adjusted to make the counter indicate the approximate value of the wavelength in air of the generated oscillations. A conversion chart attached to the top of the oscillator makes simple the conversion of this dial reading to frequency.

The output is obtained by coupling inductively to the main oscillatory circuit. At the coaxial outlet jack, which is provided at the tube housing, the source impedance is about 75 ohms in order to give approximately full output into a properly terminated coaxial line. The coaxial type outlet is in keeping with the shielded oscillator system, and, when coaxial lines are used for coupling the oscillator to the instrument using the high-frequency energy, the combination aids in confining the energy within a shielded system.

When the maximum allowable plate voltage for the tube is used on the oscillator, at frequencies up to 400 Mc (wavelengths down to 75 cm), a power output of about 4 watts can be obtained. The output at higher frequencies decreases gradually with frequency, but considerable power is still available at 600 Mc. This available power is much greater than normally required for measurements, but it permits the use of high voltage levels at the measuring instruments with the concomitant simplification of detection problems.

In order to have the full output of the oscillator available at the outlet terminals, no isolation means have been used. The result is that a reactive load coupled directly to the oscillator shifts the frequency of oscillation,[2] but for the usual conditions of loading the frequency shift is generally less than 2%.

The power supply required for this oscillator is 2 volts at 3.65 amperes for the filament, which may be obtained from the a-c line, and up to 450 volts plate supply with a maximum plate current of 80 milliamperes. The plate supply must be one that permits grounding of the

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[2] D. C. Prince, "Vacuum Tubes as Power Oscillators," Chapter V, Proc. I.R.E., August, 1923, Vol. 11, No. 4, pp. 409–418.
"Vacuum Tubes as Oscillation Generators," Part V, General Electric Review, July, 1928, Vol. 31, No. 7, pp. 388–394.

positive side, since the positive side of the high voltage circuit is connected to the outer brass container.

A power supply for the oscillator can be purchased separately. This power supply, the Type 757-P1, shown in Figure 1, provides the necessary fixed voltages for the operation of the oscillator. At a slight sacrifice in maximum oscillator output, the plate voltage delivered by the Type 757-P1 is arranged to be somewhat less than the maximum allowable for the oscillator tube in order that the tube will not be damaged even though oscillations should cease. This power supply includes a meter that is connected to indicate the oscillator grid current, a measure of the intensity of oscillation; and, of course, the power cable necessary for connection with the oscillator unit is also supplied.

— Arnold Peterson

SPECIFICATIONS FOR TYPE 757-A ULTRA-HIGH-FREQUENCY OSCILLATOR

Frequency Range: 150 to 600 Mc (200 to 50 cm wavelength).

Calibration: The frequency determined from the chart converting dial reading to frequency is accurate to ± 2½% with no load connected to the oscillator.

Output Power: 3 to 4 watts up to 400 Mc (75 cm) decreasing with frequency to about 1 watt at 600 Mc.

Output Impedance: Effectively of the order of 75 Ω.

Frequency and Wavelength Control: A slow-motion drive, which carries a dial calibrated in divisions representing approximately 0.01 cm change in wavelength, is used for changing the wavelength and frequency. A counter to the left of this dial, together with the reading of this dial, constitutes the dial reading used for the chart to convert dial reading into frequency.

Tube: A WE-316-A is required and is furnished with the instrument.

Power Supply: Filament: 3.65 amperes at 2 volts. Plate: 450 volts (max.), 80 ma. (max.). The Type 757-P1 Power Supply (see below) may be used in place of batteries.

Accessories Supplied: One power cable and one Type 774-E Cable Plug.

Mounting: The oscillator is mounted on a walnut base.

Dimensions: (Length) 21 × (width) 6 × (height) 6 inches, over-all.

Net Weight: 13¼ pounds.

Type		*Code Word*	*Price*
757-A	U-H-F Oscillator	SIREN	$195.00

PATENT NOTICE. This instrument is manufactured and sold under patents of the American Telephone and Telegraph Company solely for utilization in research, investigation, measurement, testing, instruction, and development work in pure and applied science.

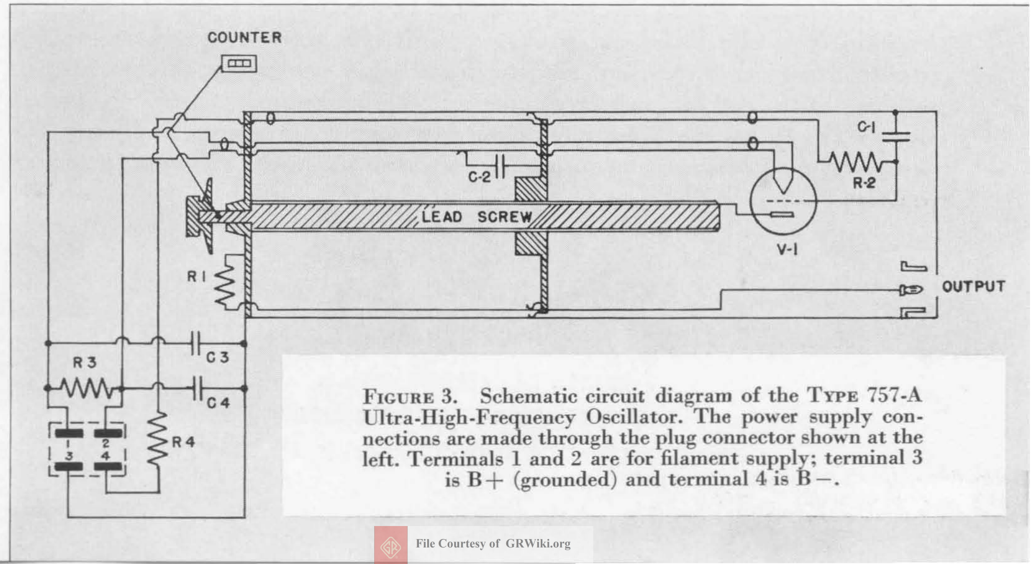

Figure 3. Schematic circuit diagram of the Type 757-A Ultra-High-Frequency Oscillator. The power supply connections are made through the plug connector shown at the left. Terminals 1 and 2 are for filament supply; terminal 3 is B+ (grounded) and terminal 4 is B−.

SPECIFICATIONS FOR
TYPE 757-P1 POWER SUPPLY

Power Input: 105 to 125 volts or 210 to 250 volts, 40 to 60 cycles. At 115 volts, 60 cycles, the power drawn is less than 70 watts.

Output: 2 volts a-c, and 300 volts d-c, for the Type 757-A U-H-F Oscillator.

Tube: 1 Type 5Y3-G is supplied.

Accessories Supplied: A seven-foot line connector cord, an interconnecting cord and plug, and spare pilot lamps.

Dimensions: (Width) 5¾ inches × (height) 6¾ inches × (length) 9¾ inches.

Net Weight: 11¼ pounds.

Type		*Code Word*	*Price*
757-P1	Power Supply	SIRENAPACK	$45.00

NEW PRICES FOR TYPE 274 PLUGS AND JACKS

● **RISING COSTS** for screw-machine parts have made necessary a revision of the prices of Type 274 Plugs and Jacks. New prices are as follows:

Type	*Unit Price*	*Package of 10*	*Package of 100*
274-P Plug	$0.12	$0.90	$6.25
274-J Jack	.10	.55	3.50
274-M Double Plug*	.50	3.50	——

*The Type 274-M Double Plug is a new model molded from polystyrene to give low capacitance and low losses.

ERRATUM—TYPE 761-A VIBRATION METER

● **IT HAS BEEN CALLED TO OUR ATTENTION** that the caption for Figure 3 of the article describing Type 761-A Vibration Meter, appearing in the June *Experimenter*, is incorrect. Since the response is flat for acceleration, the top oscillogram is acceleration, the middle one velocity, and the lowest, displacement.

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THE General Radio EXPERIMENTER

VOLUME XVI No. 4 SEPTEMBER, 1941

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

THE MICROFLASH—A LIGHT SOURCE FOR ULTRA-HIGH-SPEED PHOTOGRAPHY

Also

IN THIS ISSUE

● **THE PHOTOGRAPHY OF MOVING OBJECTS** by means of extremely short light flashes is rapidly becoming a familiar technique. Although associated in the popular mind mainly with the striking pictures of athletes, dancers, and others appearing in the press, this type of photography was first used for scientific and industrial purposes. A previous article* discussed the use of the General Radio Strobolux for single-flash photographs in industry. Like the available commercial equipment, however, the Strobolux is limited to those applications where a flash duration of 30 microseconds ($\frac{1}{30,000}$ second) will not blur the image.

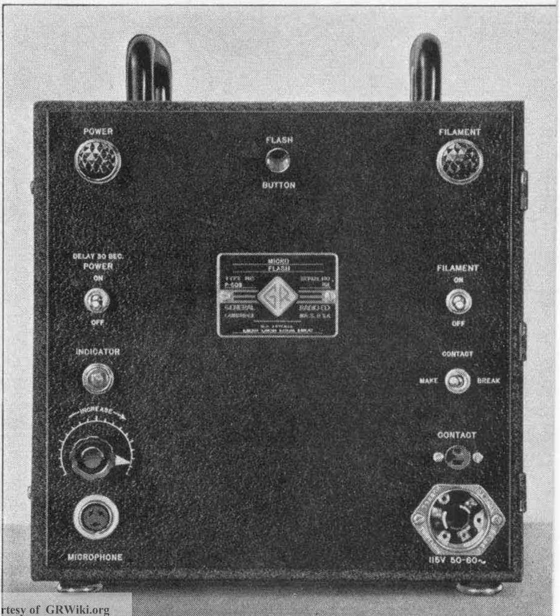

Figure 1. Panel view of the Microflash. The lamp is mounted in the rear of the case.

There are many scientific and industrial problems where flashes of much shorter duration are required. An example occurring in the field of ballistics is the photography of projectiles in flight. A rifle bullet traveling at a speed of 2400 feet per second would travel nearly an inch in 30 microseconds. To record the bullet on photographic film without appreciable

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*J. M. Clayton, "Single-Flash Photography with the Strobotac and Strobolux," *Experimenter*, November, 1940.

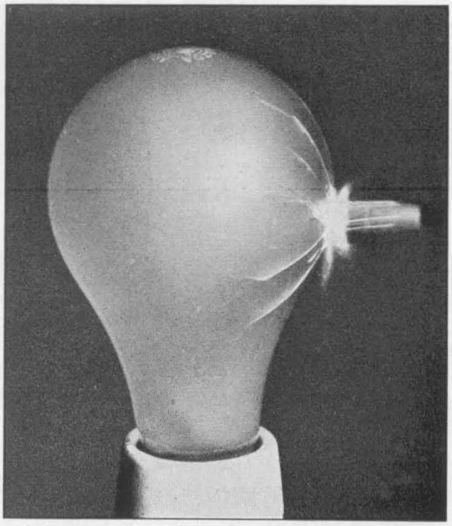

FIGURE 2. This photograph of a bullet entering an electric lamp bulb is an excellent example of the motion-arresting properties of the Microflash.

blur would require a flash duration no greater than about one microsecond.

To meet the specialized requirements of this and similar problems, the TYPE P-509 Microflash, shown in Figure 1, has been developed.† The functional diagram of Figure 4 shows the principle of operation. A high-voltage transformer and rectifier, operating from the a-c line, charge a condenser to a potential of about 7000 volts. Across the condenser is connected the flash lamp, filled with rare gases. When a pulse is impressed on the flashing electrode by the trigger circuit the gas in the lamp ionizes, producing a conducting path through which the condenser discharges. The discharge produces a brilliant flash of light lasting for about one microsecond.

The flashing circuit consists essentially of a gas-discharge tube working into an induction coil. A vacuum-tube amplifier is included to provide sufficient gain to permit operation from a conventional crystal microphone.

A microphone provides the most convenient method of triggering when the phenomenon to be photographed is accompanied by sound. For photographing a rifle bullet, for instance, the microphone is placed in such a position that

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†Only a few of these instruments have been built, and they are not available for general sale.

FIGURE 3. A Microflash photograph of the explosion of a shotgun shell as it leaves the muzzle of the gun, which is out of the picture at the left. Note that the wads can be seen as well as the cluster of shot.

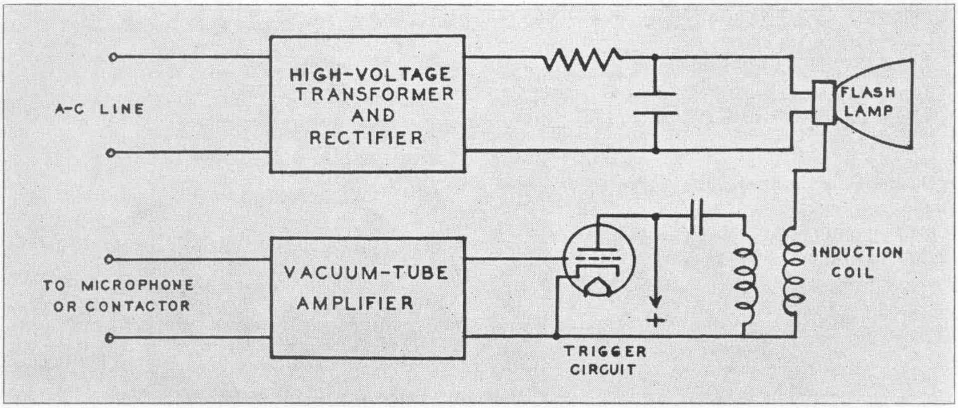

FIGURE 4. Schematic circuit diagram of the Microflash.

the sound of the report reaches the microphone at the same instant that the bullet passes the camera. Other methods of tripping the flash can, of course, be used, such as mechanical contactors and photo-electric cells. Conditions peculiar to each problem will usually determine the best method to use.

Figures 2 and 3 are actual photographs taken with the Microflash.

The Microflash is an example of the work being done by General Radio for National Defense. It will eventually lead to better stroboscopes for industry.

IMPEDANCE BRIDGES ASSEMBLED FROM LABORATORY PARTS

PART III — EXAMPLES OF MEASUREMENTS

● **THE DETERMINATION OF THE MAGNITUDES** of stray capacitance errors and the method of calculating the results of measurements with the series-resistance and Schering bridges may be made somewhat clearer through numerical examples.

Examples of actual measurements with both types of bridges are given below.

Series-Resistance Bridge

(a) Determination of C_{SG} and C_{PO} (see Figure 2, July *Experimenter*):

$$B = 10{,}000 \text{ ohms} \qquad C_N = 0.01\ \mu\text{f}$$
$$P = \text{Open-circuit} \qquad (1)$$
$$A = 13.6 \text{ ohms}$$

$$C_{PO} = \frac{13.6}{10{,}000}(0.01) \times 10^{-6} = 13.6\ \mu\mu\text{f}.$$

Reverse leads from transformer, at the bridge.

$$B,\ C_N,\ P, \text{ as above}$$
$$A = 104.3 \text{ ohms} \qquad (2)$$
$$C_{SG} = \frac{104.3}{10{,}000}(0.01) \times 10^{-6} = 104.3\ \mu\mu\text{f}.$$

Change leads back to original connection, placing C_{SG} across C_N. In addition, 60 $\mu\mu$f from the TYPE 602 is across C_N. The total value of the N arm capacitance is now 0.01016 μf (assuming the value of C_N to be exactly 0.01 μf).

(b) A measurement on a 500 $\mu\mu$f air condenser with 10,000 ohms in series gave the following data:

$$B = 10{,}000 \text{ ohms} \qquad C_N = 0.01016\ \mu\text{f}$$
$$A = 510.1 \text{ ohms} \qquad R_N = 530 \text{ ohms}$$

$$C_P = \left(\frac{510.1}{10{,}000}\right)(0.01016) = 519\ \mu\mu\text{f} \qquad (3)$$

$$C_X = 519 - 13.6 = 505\ \mu\mu\text{f}$$

$$D_P = (530)\ (0.01016)\ (6280) \times 10^{-6} = 3.38\%$$

For more accurate results a zero reading was made:

$B = 10{,}000$ ohms $\qquad A = 13.6$ ohms

$R_N = 747$ ohms $\qquad D_{PO} = (747)\ (6280)\ (0.0106) = 4.68\% \qquad (4)$

$$D_X = \frac{(3.38)\ (519) - (4.68)\ (13.6)}{505} = \frac{1755 - 63}{505} = 3.33\%$$

The true values were known to be

$C_X = 503.7\ \mu\mu\text{f} \qquad D_X = 3.15\%$

Schering Circuit

(a) Determination of C_o

$A = B = 1000$ ohms

(approximately) 1000 $\mu\mu$f in P arm (1)

$C_{N1} = 1092.3\ \mu\mu\text{f}$

$A = 10{,}000$ ohms $\qquad B = 1000$ ohms

$C_{N2} = 102.1\ \mu\mu\text{f}$ (2)

From Equation (1),

$$C_o = \frac{1092.3 - (10)\ (102.1)}{10 - 1} = 7.9\ \mu\mu\text{f} \qquad (3)$$

(b) Determination of C_{SG}

$A = 100$ ohms $\qquad B = 1000$ ohms

P arm open

$C_N = 1004.1\ \mu\mu\text{f}$ (4)

From Equation (2),

$$C_{SG} = \tfrac{1}{10}\ (1004.1 + 7.9) = 101.2\ \mu\mu\text{f} \qquad (5)$$

(c) Calibration of C_A

A calibration curve for capacitance increments of a Type 539-A Condenser is obtained by the substitution method described above.

(d) Measurement of dissipation factor of the zero capacitance in the P arm:

$A = 10{,}000$ ohms $\qquad B = 10{,}000$ ohms

P arm open, 539-A across A arm

$$\begin{aligned} C_N &= 93.6\ \mu\mu\text{f} \\ C_A &= 498\ \mu\mu\text{f} \qquad (6) \\ C_{SG} &= 93.6 + 7.9 \\ &= 101.5\ \mu\mu\text{f} \end{aligned}$$

0.001 μf (Type 505-F) across P arm

$$\begin{aligned} C_N &= 1092.2\ \mu\mu\text{f} \\ C_A &= 380\ \mu\mu\text{f} \qquad (7) \\ C_{P2} &= 1092.2 + 7.9 = 1099.9\ \mu\mu\text{f} \\ \Delta Q_A &= \Delta C_A \,.\, R_A \,.\, \omega \\ &= (500 - 380) \times 10^{-12} \times 10{,}000 \\ &\quad \times 6280 \qquad (8) \\ &= 0.00754 \end{aligned}$$

From Equation (3),

$$D_{SG} = \frac{\Delta Q_A}{\frac{C_{SG}}{C_{P2}}} = \frac{.00754}{1 - \frac{101.5}{1099.9}} \qquad (9)$$

$$= 0.0083 \qquad = 0.83\%$$

(e) Establishing an initial dissipation factor balance:

$A = B = 1000$ ohms

1000 $\mu\mu$f air condenser connected in P arm. The D of the P arm is then

$$D_P = \frac{D_{SG}C_{SG} + D_1C_1}{C_P} \qquad (10)$$

$$= \frac{(0.83)(102.1) + (.004)(1000)}{1102.1}$$

$$= 0.081\%$$

The condenser C_A is now set to a value corresponding to

$$\frac{D_P}{R_A\omega} = C_A = \frac{.00081}{1000 \times 6280} = (129)\ \mu\mu\text{f}$$

and balance established by means of a small air condenser across the B arm. The bridge is now ready for use.

As a check on over-all accuracy a measurement on a 0.001 μf mica condenser in series with 1000 ohms yielded the following data:

$$\begin{aligned} A &= B = 10{,}000 \text{ ohms} \\ C_A &= 102\ \mu\mu\text{f} \\ C_N &= 1094.4\ \mu\mu\text{f} \\ C_X &= (1094.4 + 7.9) - 101.2 = \mathit{1001.1}\ \mu\mu\text{f} \\ D_P &= (10{,}000)(6280)(102) = 0.64\% \\ D_X &= \frac{(0.64)(1102.3) - (0.83)(101.2)}{1001.1} \\ &= \mathit{0.62\%} \end{aligned}$$

These results compare very well with the (known) true values of 1002.3 and 0.65% for capacitance and dissipation factor, respectively.

As another example, the following data represent a measurement of a Type 505, 0.001 μf condenser, by a substitution method:

$C_A = 153\ \mu\mu\text{f} \qquad C_N = 1102.5\ \mu\mu\text{f}$

$C'_A = 102\ \mu\mu\text{f} \qquad C_N' = 104.2\ \mu\mu\text{f}$

$\Delta C_A = 51 \qquad \Delta C_N = 998.3\ \mu\mu\text{f}$

$$D_X = (1000)(6280)(51)(\times 10^{-12})\ \frac{1022.3}{998.3}$$

$$= 0.035\%$$

$$C_X = \Delta C_N = 998.3$$

Accurate measurements on this condenser by our Standardizing Laboratory yielded:

$C = 998.6\ \mu\mu\text{f} \qquad D = 0.037\%$

— Ivan G. Easton

BRINGING THE CATALOG UP TO DATE

● **THERE HAVE BEEN A NUMBER OF ADDITIONS TO** and deletions from our line of instruments since the publication of Catalog K. In addition, several instruments have been temporarily discontinued in order that our facilities may be more efficiently used for the production of items needed for National Defense.

NEW INSTRUMENTS ADDED TO LINE

Type		*Described in EXPERIMENTER*
50	Variac	July 1939
318-C	Dial Plate	Dec. 1940
700-P1	Voltage Divider	Aug. 1939
701-A, -K	Dials	Aug. 1940
703-K, -L	Dials	Aug. 1940
717-K, -L	Dials	Aug. 1940
726-P1	Multiplier	May 1940
729-A	Megohmmeter (Battery Operated)	July 1940
740-BG	Capacitance Test Bridge	Feb. 1939
755-A	Variable Air Condenser	Aug. 1939†
757-A	U-H-F Oscillator	Aug. 1941
757-P1	A-C Power Supply	Aug. 1941
758-A	U-H-F Wavemeter	Aug. 1940
759-P35	Vibration Pickup	*
759-P36	Control Box	*
761-A	Vibration Meter	June 1941
769-A	Square-Wave Generator	Dec. 1939
774	Coaxial Terminals	April 1941
804-B	U-H-F Signal Generator	Feb. 1941
821-A	Twin-T Impedance-Measuring Circuit	Jan. 1941
25-A	Frequency Monitor (Consists of TYPE 475-C Frequency Monitor and TYPE 681-B Frequency Deviation Meter)	Jan. 1940†

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*Descriptive folder available on request.
†Reprint available.

INSTRUMENTS REPLACED BY NEW MODELS

Type		*New Model*	*Described in EXPERIMENTER*
100-K	Variac	100-Q	July 1939
100-L	Variac	100-R	July 1939
475-B	Frequency Monitor	475-C	Jan. 1940
546-A	Microvolter	546-B	April 1941
561-C	Vacuum-Tube Bridge	561-D	
614-B	Selective Amplifier	614-C	
616-C	Heterodyne Frequency Meter	616-D	
675-M	Piezo-Electric Oscillator	675-N	

Type		New Model	Described in EXPERIMENTER
681-A	Frequency-Deviation Meter	681-B	Jan. 1940
695-B	Charging Equipment	695-C	
696-B	Power Supply	696-C	
716-A	Capacitance Bridge	716-B	
759-A	Sound-Level Meter	759-B	April 1940*
759-P1	Tripod and Cable	759-P21	
834-A	Electronic Frequency Meter	834-B	

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*Descriptive folder available on request.

INSTRUMENTS DISCONTINUED

Type 70-A, -B Variac Transformers
Type 80-A, -B Variac Transformers
Type 90-B Variac
Type 138-A Binding Post
Type 138-D Switch Contact
Type 154-A, -B Voltage Dividers
Type 246-L, -M, -P Variable Air Condensers
Type 247-G, -F Variable Air Condensers
Type 274-K Binding Post Assembly
Type 274-L Binding Post Assembly
Type 293-A Universal Bridge
Type 293-P1 Bridge Transformer
Type 293-P2 Bridge Transformer
Type 293-P3 Slide Wire
Type 329-J Attenuation Box
Type 410-A Rheostat Potentiometer
Type 480-A, -B Relay Racks
*Type 516-C Radio-Frequency Bridge
Type 525-C, -D, -F, -H, -L Resistors
Type 544-P2 90-Volt Power Supply
Type 574 Wavemeter
Type 578-AR, BR, CR, AT, BT, CT Transformers
Type 586-DM, -DR, -EN, -ER, -P5, -P6, -Q1 Power-Level Indicators
Type 613-B Beat-Frequency Oscillator
Type 625-A Bridge
Type 641-A, -B, -C, -D, -E, -F, -G, -H, -J, -K, -L Transformers
Type 664-A Thermocouple
Type 666-A Variable Transformer
Type 671-A Schering Bridge
Type 677-U, -Y Inductor Forms
Type 677-P1 Spacer
Type 678-P Plug Base
Type 678-J Jack Base
Type 682-B Frequency-Deviation Meter
Type 684-A Modulated Oscillator
Type 716-P2 Guard Circuit
Type 721-A Coil Comparator
Type 722-FU Precision Condenser (Unmounted)
Type 733-A Oscillator
Type 741-G, -J, -P Transformers

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*To be replaced by Type 916-A.

INSTRUMENTS TEMPORARILY DISCONTINUED

Type 202-A, -B Switches
Type 202-Y, -Z Knobs
Type 219-L, -N Decade Condensers
Type 334-F, -K, -R, -T, -Z Variable Air Condensers
Type 335-Z Variable Air Condenser
Type 505-R, -T, -U, -X Condensers
Type 526-A, -B, -C, -D Mounted Rheostat-Potentiometers
Type 539-X Variable Air Condenser
Type 611-C Syncro Clock
Type 642-D Volume Control
Type 646-A Logarithmic Resistor
Type 669-R Slide-Wire Resistor
Type 670-BW, -FW Compensated Decade Resistors
Type 739-A, -B Logarithmic Air Condensers

NEW RESISTANCE VALUES FOR RHEOSTAT-POTENTIOMETERS

● **ALTHOUGH THERE ARE NO ACCEPTED STANDARDS** or preferred values for variable resistors, it is desirable, in order that a given resistance range be covered by as few models as possible, to standardize on a logarithmic distribution of resistance values. The values listed in our current Catalog K for General Radio rheostat-potentiometers are the outgrowth of customer demands, over a period of several years, for resistors to be used in particular applications. Most of the resistance values so determined no longer have any significance, as, for instance, the low-resistance models which originally met the NEMA standards for filament rheostats in vacuum-tube circuits.

In order to cover the available resistance range for a given type of unit more effectively, the resistance values have been revised to give an approximately logarithmic distribution for each model. A 1-2-5-10 system has been adopted for all models except Types 333 and 533, which already use a 1-3-10 system.

The new listings are as follows:

Type	*Maximum Resistance*	*Maximum Current*	*Code Word*	*Price*
301-A	5	0.9 a	PALSY	$1.00
301-A	10	0.65 a	REMIT	1.00
301-A	20	450 ma	RENEW	1.00
301-A	50	280 ma	RIFLE	1.00
301-A	100	200 ma	RIGID	1.00
301-A	200	140 ma	REBUS	1.00
301-A	500	90 ma	RIVAL	1.00
301-A	1000	65 ma	RAVEL	1.00
301-A	2000	45 ma	READY	1.00
301-A	5000	28 ma	ROMAN	1.00
*301-A	10,000	17 ma	CURRY	1.50
*301-A	20,000	12 ma	CRUMB	1.50
214-A	10	1.0 a	RURAL	2.00
214-A	20	0.7 a	RAZOR	2.00
214-A	50	450 ma	RAPID	2.00
214-A	100	320 ma	RIVET	2.00
214-A	200	220 ma	EMPTY	2.00
214-A	500	140 ma	ROSIN	2.00
214-A	1000	100 ma	ENACT	2.00
214-A	2000	70 ma	SYRUP	2.00
214-A	5000	45 ma	ROWEL	2.00
214-A	10,000	32 ma	RUMOR	2.00
371-A	1000	120 ma	REDAN	4.00
371-A	2000	90 ma	REFIT	4.00
371-A	5000	55 ma	ROTOR	4.00
371-A	10,000	38 ma	ROWDY	4.00
371-A	20,000	28 ma	RULER	4.00
*371-A	50,000	16 ma	SATYR	4.00
*371-A	100,000	11 ma	SEPOY	4.00
371-T	10,000	28 ma	SULLY	4.00
*314-A	1000	90 ma	DIVAN	4.00
*314-A	2000	65 ma	ENEMY	4.00
*314-A	5000	40 ma	ENJOY	4.00
*314-A	10,000	28 ma	DIVER	4.00
*314-A	20,000	20 ma	ENROL	4.00
*314-A	50,000	13 ma	DONAX	4.00
*314-A	100,000	9 ma	DONGA	4.00

(*Continued on page 8*)

Type	*Maximum Resistance*	*Maximum Current*	*Code Word*	*Price*
*471-A	10,000	35 ma	ERECT	$6.00
*471-A	20,000	25 ma	HUMAN	6.00
*471-A	50,000	15 ma	ERODE	6.00
*471-A	100,000	10 ma	ERUPT	6.00
*471-A	200,000	8 ma	ESKER	6.00
333-A	1	8.5 a	VALOR	4.00
333-A	3	5.8 a	VAPID	4.00
333-A	10	2.5 a	VENUS	4.00
333-A	30	1.8 a	VIGIL	4.00
333-A	100	1.0 a	VIGOR	4.00
333-A	300	0.6 a	VILLA	4.00
333-A	1000	0.25 a	HUMOR	4.00
533-A	1	15.8 a	MOLAR	6.00
533-A	3	9.1 a	MONAD	6.00
533-A	10	5.0 a	MORAL	6.00
533-A	30	2.9 a	MOTTO	6.00
533-A	100	1.6 a	MUGGY	6.00
533-A	300	0.9 a	MUMMY	6.00
533-A	1000	0.5 a	HUSSY	6.00

*Supplied with linen-bakelite protecting strips.

Power ratings have been revised to give consistent values based on a temperature rise of approximately 50° to 60° C. above the ambient, except those for TYPE 333 and TYPE 533, which are based on a temperature rise of approximately 250° C.

Type	*Power Rating*	*Type*	*Power Rating*
301-A	4 watts	371-A with protecting strip	12 watts
301-A with protecting strip	3 watts	371-T	8 watts
214-A	10 watts	471-A	12 watts
314-A	8 watts	333-A	100 watts
371-A	15 watts	533-A	250 watts

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A VACUUM-TUBE-DRIVEN TUNING-FORK OSCILLATOR

Also IN THIS ISSUE

● **THE ELECTRO-MECHANICAL OSCILLATOR,** whose frequency is controlled by a tuning fork, has many uses in electrical communications. The Type 813 Oscillator and its predecessor the Type 213 have been used, for instance, as power sources for bridge measurements and for transmission measurements on lines and cables, as modulating sources for test oscillators and radio-beacon transmitters, and as test-tone sources for communication systems. For applications where a constant-frequency audio oscillator is needed for permanent installation in other equipment or circuits, the tuning-fork type is usually the most economical.

The Type 813 Oscillator, like its predecessor the Type 213, is a microphone-button-driven type. For some applications, in particular those requiring very low distortion, stability of output voltage, or a-c operation, vacuum-tube drive is more satisfactory. To meet these requirements, the new Type 723 Vacuum-Tube Fork, shown in Figure 1, has been designed. This instrument has extremely low distortion so that it can be used for distortion measurements without additional filters, and its output is much more stable

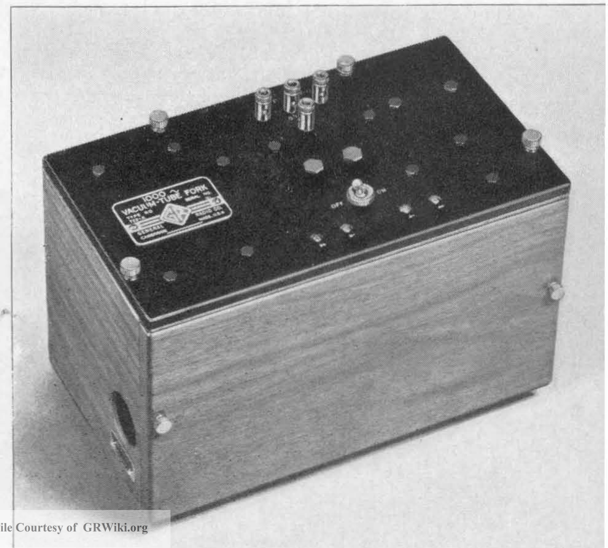


Figure 1. View of the Type 723 Vacuum-Tube Oscillator, showing the panel.

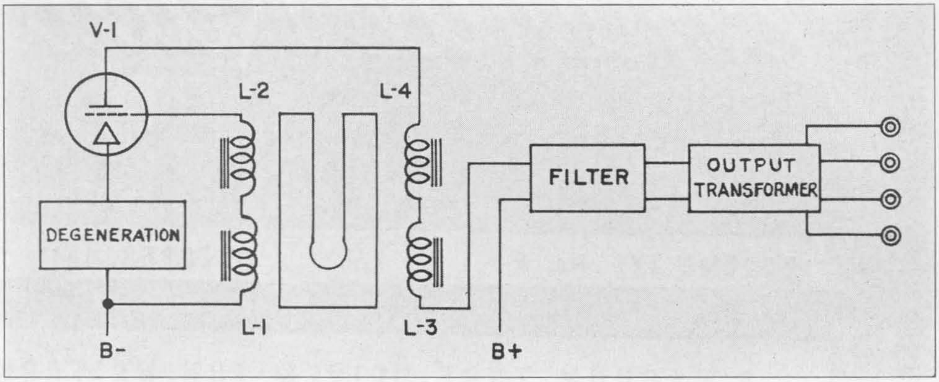


FIGURE 2. Functional circuit diagram of the TYPE 723 Vacuum-Tube Fork.

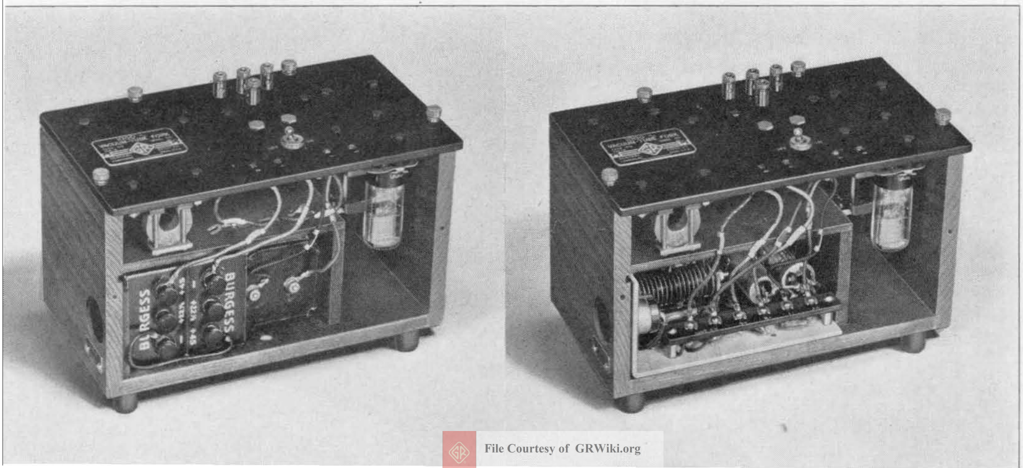
than that of the microphone-button type.

The tuning fork itself is identical with that used in the TYPE 813 Oscillator. Figure 2 shows the electrical circuit. Driving coils L_3 and L_4, wound on U-shaped permanent magnets to give a constant polarization, are arranged along one tine of the fork, and similar coils L_1 and L_2, facing the other tine, serve as pickup coils. A 1A5G-type vacuum tube provides the necessary amplification between pickup and driving circuits. The output filter is in series with the driving coils and feeds an output transformer. A tapped secondary winding provides output impedances of 50, 500, and 5000 ohms.

The TYPE 723 Vacuum-Tube Fork is available in two models, 1000 cycles and 400 cycles, and with either of two types of power supply, batteries, or an a-c power pack. The a-c power unit uses selenium rectifiers and includes a voltage regulator tube to minimize the effect of line voltage fluctuations. The two power supplies are interchangeable mechanically and mount in the oscillator cabinet as shown in Figure 3.

Because magnetic drive and pickup impose comparatively little restraint upon the vibration of the fork, the frequency stability and waveform of this oscillator are considerably better than those of microphone-button-driven types, and the effective Q of the fork is higher. Another factor contributing to low distortion is the greater degree of linearity obtainable in the operation of a vacuum tube than in a microphone but-

FIGURE 3. Views showing the power-supply compartment.
Left, battery supply and, right, a-c power supply.



ton. The output filter further reduces the harmonic amplitudes, so that the total harmonic content is less than 0.5% when the oscillator is operated into a load equal to or greater than the nominal output impedance.

The rated output is 50 milliwatts into a matched resistive load. The fork, the tube, and the associated circuit elements are mounted on the under side of a bakelite panel. The walnut cabinet includes space for the power supply.

SPECIFICATIONS

Frequency: The Type 723 Vacuum-Tube Fork is supplied for two operating frequencies, 1000 cycles and 400 cycles. (See price list below.)

Frequency Stability: The temperature coefficient of frequency is approximately −0.008% per degree Fahrenheit. The frequency is entirely independent of load impedance. When the a-c power supply is used, an initial downward drift of frequency occurs as the temperature of the fork is affected by heat generated in the power-supply unit. The total frequency drift is of the order of .15% to .2%. Most of this drift, however, occurs in the first 30 minutes of operation.

Accuracy: The frequency is adjusted to within ±0.01% of its specified value, at 77° Fahrenheit, using battery power supply.

Output: The output to a matched load is approximately 50 milliwatts.

Internal Output Impedance: Output impedances of 50, 500, and 5000 ohms are provided.

Waveform and Hum Level: The total harmonic content is less than 0.5%. The hum is negligible.

Terminals: Binding posts for the output circuit are mounted on the panel. Battery terminals are brought out to sunken screw heads on the panel to permit measurements of the battery voltages.

Power Supply: The instrument is available for either battery operation or for operation from 105 to 125-volt, 50 to 60-cycle line. For battery operation one Burgess type 4FA (1½ volt) and two Burgess type Z30-N (45-volt) are required. The batteries and a-c power supply are interchangeable. The power supply, Type 723-P1, is available separately. (See price list.) The ON-OFF switch is arranged to control the a-c line or the battery current.

Vacuum Tubes:

For battery supply: 1 type 1A5G
For a-c supply: 1 type 1A5G
1 type VR-105-30

The necessary tubes are supplied.

Accessories Supplied: A seven-foot line connector cord is supplied with the a-c operated model.

Mounting: The oscillator assembly is mounted on a bakelite panel and is enclosed in a walnut cabinet.

Dimensions: (Length) 10⅝ × (width) 6¼ × (height) 7¾ inches, over-all.

Net Weight: 11¼ pounds, including batteries; 8 pounds 14 ounces, with a-c supply; a-c power supply alone, 1¼ pounds.

Type	*Frequency*	*Power Supply*	*Code Word*	*Price*
723-A	1000 cycles	Batteries	SNAKE	$70.00
723-C	1000 cycles	105 to 125 volts, 50 to 60 cycles	SOLID	90.00
723-B	400 cycles	Batteries	STORY	70.00
723-D	400 cycles	105 to 125 volts, 50 to 60 cycles	SULKY	90.00
723-P1	A-C Operated Power Supply Only		SNAKEYBATT	22.00
723-P2	Set of Replacement Batteries		SNAKEYPACK	2.00

NOTE TO EXPEDITERS

To insure an early reply to follow-up inquiries on undelivered orders, please state in your inquiry the order number, the date of the order, and, if space permits, the material ordered. This information will assist us materially in locating and tracing your order.

IMPEDANCE BRIDGES ASSEMBLED FROM LABORATORY PARTS

PART IV—TRANSFORMER ERRORS

● **THE IDEAL SHIELDED TRANSFORMER** interposed between a generator and bridge would provide complete isolation between the two windings, except for the desired inductive coupling. Even in a carefully designed and constructed transformer, however, there will be some small residual electrostatic and leakage coupling between the windings, which will introduce extraneous voltages into the bridge circuit. This type of error may best be analyzed by considering the effect of introducing (across any arm of a bridge) an external voltage E' in series with an impedance Z_T, as illustrated in Figure 1. The equation of balance for this network is

$$Z_P = \frac{Z_B}{Z_A} \, \frac{Z_N\left(\dfrac{Z_T}{\beta}\right)}{Z_N + \left(\dfrac{Z_T}{\beta}\right)} \; * \qquad (1)$$

The parameter β is given by

$$\beta = 1 \pm \alpha\left(1 + \frac{Z_B}{Z_A}\right) \; *$$

where $\alpha = \dfrac{E'}{E_O}$.

It will be observed that the balance Equation (2) is identical to that which would be obtained if a passive impedance equal to $\dfrac{Z_T}{\beta}$ were connected in parallel with Z_N. This fact makes it possible to estimate readily the effect of stray coupling from the voltage source to the bridge.

By properly choosing the magnitude and polarity of E' the expression for β can be reduced to zero (for a given value of the ratio $\dfrac{Z_B}{Z_A}$). Under this condition no error will be caused by the presence of Z_T, as it will effectively appear as an infinite impedance across Z_N. In Figure 3 is presented an experimental verification of this fact. The observed change in power-factor reading of a Type 716-A Capacitance Bridge is plotted against the computed value of the parameter β for various values of E'. Theoretically, this plot should be a straight line, passing through the origin, of slope equal to the coupling resistance R_T. The agreement obtained is well within the limit of experimental error.

In the practical case where E' is the voltage source supplying the bridge, the ratio α is the effective voltage ratio of the bridge transformer. Thus from the point of view of reducing the effect of residual interwinding coupling, there is an optimum turns ratio for any given bridge. In general, of course, other considera-

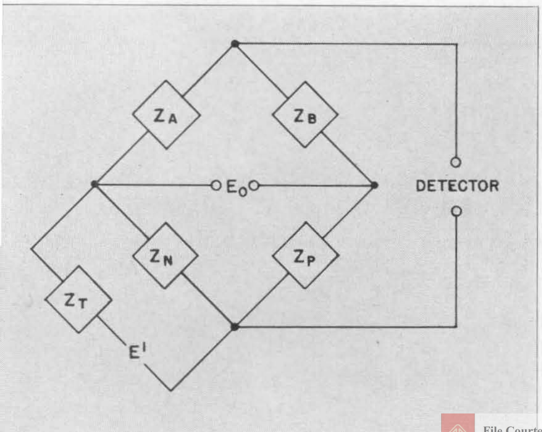


FIGURE 1. Showing an external voltage connected across a bridge arm through an impedance Z_T.

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*These equations are strictly true only for the case of a zero-impedance generator. For a generator of finite impedance, Equation (1) may still be used, but the expression for β becomes more complicated.

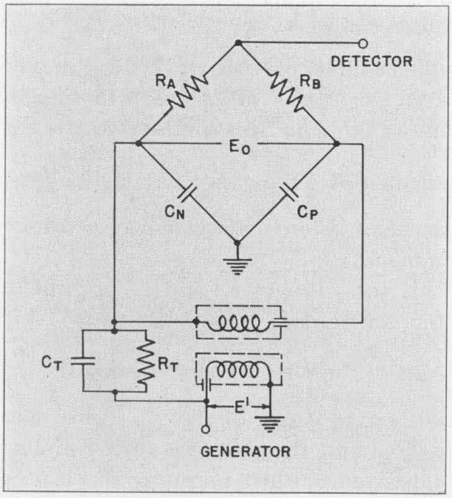

FIGURE 2. Showing how an impedance between windings of a TYPE 578-A Shielded Transformer gives rise to the situation shown in Figure 1.

tions dictate the choice of turns ratio, but it is well to keep in mind that there is a best choice of *polarity*.

Using the TYPE 578-A Shielded Transformer in the bridge arrangements previously described, the residual coupling between windings effectively places a small capacitance (βC_T) in parallel with large resistance $\frac{R_T}{\beta}$ across one of the *capacitance* arms of the bridge.

This can be seen by comparing Figures 1 and 2. With the bridge grounded at the junction of the capacitance arms, the oscillator voltage E', in series with C_T, R_T, is placed across the N arm. Although there is some coupling from the high side of the primary to either secondary terminal, the only path shown is to that terminal which is connected to the secondary shield. The other is negligible in comparison.

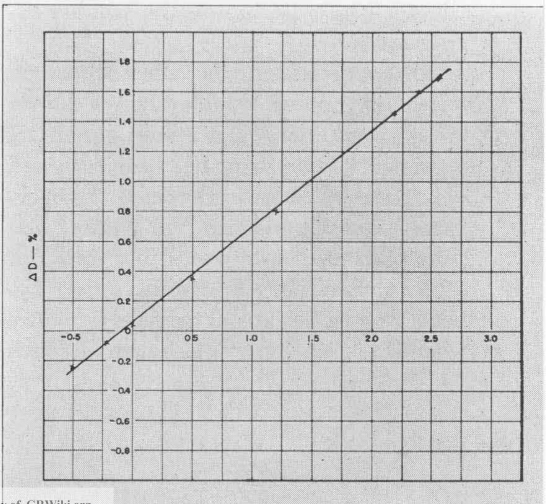

FIGURE 3. Plot of observed change in dissipation factor as a function of β. The slope corresponds to a resistance of 226 megohms. The measured d-c value of the coupling resistance used was 221 megohms.

The capacitance C_T is of the order of a few tenths of a micromicrofarad, while the d-c leakage resistance is of the order of 10,000 megohms or greater. In addition to the d-c leakage, there is a component of resistance contributed by the dielectric losses associated with C_T. At 1000 cycles this component is roughly of the same magnitude as the leakage resistance. At frequencies much below this, the latter component is more important, while at higher frequencies the dielectric losses in C_T become the significant component.

In the previously suggested arrangement of the Schering circuit the impedance $\frac{Z_T}{\beta}$ is placed across P, or unknown arm of the bridge, and becomes part of the zero capacitance and dissipation factor of that arm. Since β varies with the setting of the ratio arms, however, this component is not constant, and the value of C_{SG}[†] and D_{SG}[†] will not be the same for all settings of the ratio arms. For large values of $\frac{A}{B}$ a fairly large capacitance is added, but in this case the unknown (C_P) is large and the error remains negligible.

In the series-resistance bridge as de-

[†] See August and September, 1941, *Experimenters*.

scribed, $\frac{Z_T}{\beta}$ is placed across the N, or standard arm of the bridge, where it is completely negligible if β remains reasonably small, since $C_N = 10{,}000\ \mu\mu\text{f}$. In the extreme case, however, of $\frac{B}{A}$ equal to 1000, a capacitance of the order of 25 $\mu\mu$f will be placed across C_N, i. e., $\beta = 1 \pm \frac{1}{4}(1 + 1000)$. Although small compared to 10,000 $\mu\mu$f, this value is not negligible. But a value of $\frac{B}{A}$ equal to 1000 corresponds to a capacitance in the P arm of only 10 $\mu\mu$f. Thus the error of 25 parts in 10,000 introduced corresponds to only a few hundredths of a micromicrofarad in the determination of the unknown capacitance.

Similar arguments will show that the effect of $\frac{R_T}{\beta}$ is generally negligible to other sources of error, at 1000 cycles. For extreme values of bridge arm impedances, however, it is well to estimate the effect of βC_T and $\frac{R_T}{\beta}$ for any given measurement before a statement of accuracy is made.

If the voltages E' and E_O are not in (or 180° out of) phase the ratio α becomes complex. The impedance $\frac{Z_T}{\beta}$ is then no longer of the same character as Z_T, and the presence of a coupling capacitance, which normally affects only the capacitance balance, will produce a shift in dissipation factor balance. Similarly, a resistance coupling will affect the capacitance balance. This can occur at higher frequencies, where the leakage reactance of the transformer becomes appreciable, producing a phase shift between input and output voltages.

— Ivan G. Easton

MORE INFORMATION, PLEASE!

● **TIME**--a particularly valuable commodity to defense industries — can frequently be saved when equipment is returned for repairs if an effort is made to supply our Service Department with detailed information on the trouble experienced.

Often instruments have been returned with no greater trouble than a blown fuse or deteriorated vacuum tube. Occasionally equipment has been returned in first-class condition because the operating instructions supplied with it had not been followed, or possibly were not fully understood. Frequently what appears to be unsatisfactory performance can be traced to the external circuit with which the instrument is being used.

The unfortunate aspect of such cases is that our Standardizing Laboratory must spend considerable time in checking these returned instruments to make sure that there are no other sources of difficulty and no intermittent troubles which may be lying dormant only to reappear at the customer's laboratory. If these checks reveal no evidence of faulty operation we must then write the customer for information which might better have been furnished earlier and might even have eliminated the necessity for returning the equipment, thereby saving time, inconvenience, and expense.

Even when equipment is actually in

need of repairs, a detailed statement of the trouble may be very helpful. Analysis of this information sometimes shows that the trouble is caused by a single defective component which can be easily replaced. Where time is important, this part can be supplied to the customer with instructions for its installation in the instrument. If, on the other hand, the equipment must be returned, our laboratory will at least know what condition needs correction, and no time will be lost in making extra preliminary checks. Since repair charges depend upon the cost of the labor and material involved, a considerable saving can thus be made in the repair charge.

A further advantage of the detailed report is that it eliminates the possibility of our overlooking an intermittent fault while some other defect is discovered and corrected.

The following procedure is recommended when trouble apparently develops in General Radio equipment:

(1) Study the instruction book carefully.

(2) Check all fuses.

(3) Check all batteries under normal load. Replace if they show less than rated voltages. If batteries are old, it is advisable to try a new set, because the internal resistance may be high enough to cause trouble without showing a serious drop in voltage.

(4) Check all tubes.

(5) Check carefully the external circuit with which it is being used, particularly if the circuit is a new one.

If the source of trouble cannot be located, write our Service Department, giving the type and serial number of the instrument, a description of the exact nature of the defect, telling whether it developed suddenly or over a period of time, and a diagram of the circuit connected to it showing component values and locations of grounds. When measuring instruments, such as bridges or meters, give inaccurate readings, a sheet of sample data and a description of the method of measurement should be given. When the instruction book gives several paragraphs under the titles of INSTALLATION and OPERATION, it is especially helpful to describe the observed effects of carrying out the consecutive steps in each paragraph.

Upon the receipt of this information the Service Department will be in a position to render prompt assistance and can advise you at once how your service problem can best be handled.

When repairs are to be made on instruments needed in the execution of a National Defense contract, the work can be greatly facilitated if the preference rating is mentioned in the repair order.

— Kipling Adams

FREQUENCY MODULATION

● **BY NOW** almost everyone has felt the impact of the National Defense Program in one way or another. For many manufacturers, including the General Radio Company, it has meant a complete rearrangement and expansion of manufacturing facilities. Most of our plans for new equipment, especially for non-defense material, have been completely changed, and we are putting our maximum effort into the production of urgently required defense supplies.

One of the non-defense activities that we have been especially loath to postpone is the production of instruments for broadcast frequency modulation.

Designs had been completed and production was well under way for a modulation monitor for FM transmitters and for a standard-signal generator for FM broadcast receiver testing. We believe that both of these instruments are designed to cover adequately the requirements of their special fields, and under ordinary circumstances they would have been on the market some time ago. The manufacture of them has had to be put aside, at least temporarily, but we are glad to say that we are able still to continue engineering development of improved circuits and methods; in the end, of course, we intend that the instruments shall be up to the minute when they are finally released.

Another instrument that is being developed in the laboratory is a frequency monitor for FM, and this work will be continued up to the point of production so that we can start it immediately when production facilities are again available.

These are examples of the work that is continuously going on in our laboratories. Many of our engineers are working on instruments directly connected with the defense program, but a part of the staff is continuing the development of more generally useful equipment. As has happened before in a situation of this kind, the great concentrated effort on military equipment will result in a big step forward for the communications art in general, and all lines of radio development will eventually receive the benefits of the things that are being learned in doing the defense job.

— A. E. Thiessen

ABOUT THAT SHOTGUN SHELL

We have received a number of comments from readers about Figure 3 of the article on the Microflash appearing in last month's *Experimenter*. The caption is undoubtedly incorrect, and the shell was fired from the *right* of the picture.

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THE General Radio EXPERIMENTER

VOLUME XVI No. 6 NOVEMBER, 1941

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A REDESIGN OF THE VACUUM TUBE BRIDGE

Also

IN THIS ISSUE

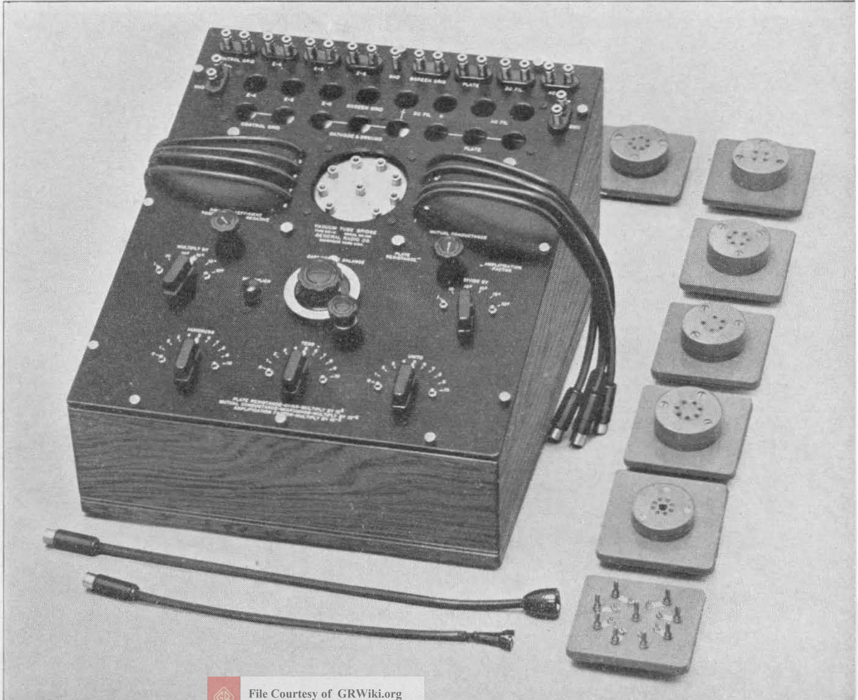
Page

● **THE TYPE 561-A VACUUM TUBE BRIDGE*** was introduced over nine years ago in order to provide a means of measuring the three dynamic tube coefficients — transconductance, amplification factor, and electrode resistance — over the wide ranges of values encountered under various operating conditions in the many new types of tubes then coming into use. The bridge embodied new measuring circuits to provide flexibility of operation and to permit the balancing of currents through the tube capacitances without affecting the measurement.

---

*General Radio *Experimenter*, May, 1932.

FIGURE 1. Panel view of the bridge with its plug-in socket adapters.

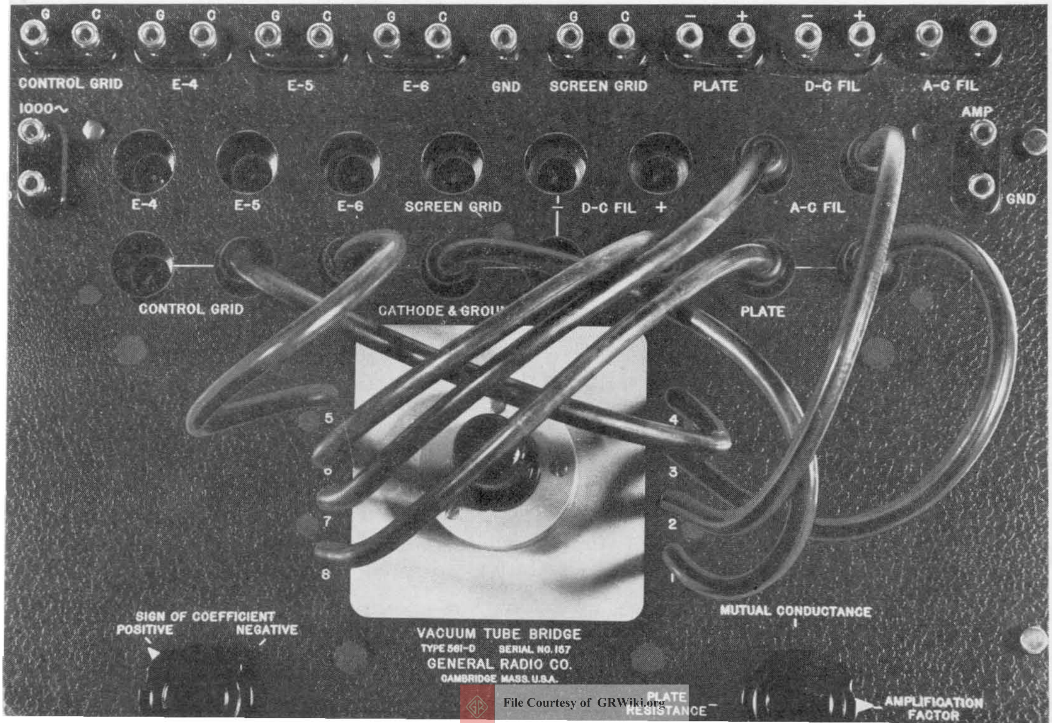


Although the original instrument has undergone only minor design modifications, it has been in steadily increasing demand ever since its introduction. The measuring circuits themselves have not been changed and have proved fully equal to the problems which have been encountered. The socket arrangement, however, has not proved adequate, although new adapters have been supplied as new requirements arose. Tube manufacturers and other users have usually found it desirable in setting up the bridge to employ their own socket arrangement and power-supply circuits instead of those supplied. A redesign of this portion of the bridge has been needed to provide a truly universal arrangement.

There is no evidence that the introduction of new base types is approaching an end, and the logical solution of the problem seems to be to eliminate the sockets entirely from the bridge and to provide a jack plate, into which any desired socket can be inserted. Moreover, the tube industry has not found it possible to maintain standard base connections, and several different arrangements are now found even of cathode, heater, and shield connections. The decision was made, therefore, to provide a correspondingly numbered cord and plug for each socket terminal. By this means the desired electrodes can quickly be connected to the measuring circuits and power supplies, and several electrodes can readily be connected in parallel, as when testing a pentode connected as a triode.

Figure 1 is a photograph of the redesigned bridge, called Type 561-D, showing the jack plate with the Type 274 Jacks to receive the plugs on the socket plate. Figure 2 is an enlarged view of the tube-control portion with an octal socket inserted in position and the cords plugged in for measurements on a 6SK7 tube operated as a triode. It will be noted that three jacks, parallel connected, are

Figure 2. View of the tube control portion of the panel with cords plugged in for a typical measurement.



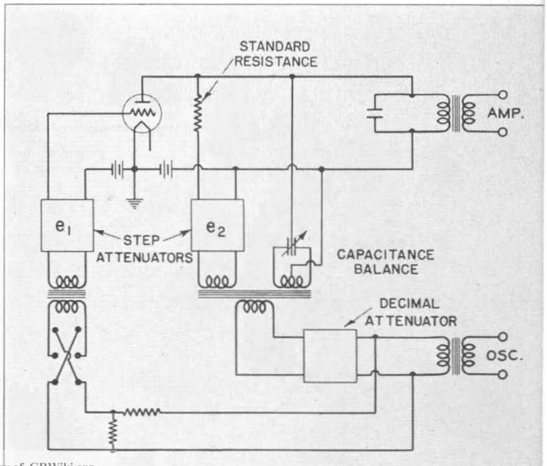
marked PLATE. These run to the plate portion of the measuring circuit and thence to the plate supply battery. In the example shown, the tube is operating as a triode, so cords from the suppressor, screen, and plate, bearing the numbers of the socket terminals 3, 6, and 8, respectively, are plugged into the three plate jacks. The cord from the control grid, number 4, is plugged into one of the two CONTROL GRID jacks, which lead through the grid portion of the measuring circuits to the grid bias source. The shell and cathode cords, numbered 1 and 5, are plugged into two of the CATHODE AND GROUND jacks and cords 2 and 7 into the A–C FIL jacks, thus completing the socket terminal connections. Leads to the required plate and control-grid voltage sources are not shown, but would be attached to the correspondingly marked pairs of binding posts at the top of the panel.

The general arrangement is very flexible and any imaginable electrode connections can be made quickly and easily. It will be noted that terminals for three extra voltage sources are provided, in addition to filament or heater, control grid, screen grid, and plate. This is so that measurements can be made with the suppressor and other auxiliary electrodes maintained at specified voltages other than cathode potential.

The cord and jack arrangement has the additional advantage of permitting full shielding of the various terminal connections, so that the stray capacitances which are balanced out in the measurement are only those of the tube and socket. The cords have a concentric braided shield, over which is molded rubber composition of high durability. Concentric plugs and jacks are used so that the shields run all the way to the jack plate, no matter what cross-connections are required. In the measuring circuits a guard connection is employed, which is at plate potential, to shield the plate electrode and the wiring connected to it. This is brought to the outside terminals of the three plate jacks and thence through the shield braid on the cord to the socket terminals of the plate and other electrodes connected in parallel with it. When the shields of the other cords are plugged into any of the other jacks they are connected to ground. This double system of shielding, although involving complications in the design of the instrument, permits a very simple and straightforward technique on the part of the user, as no precautions need be taken in the measurement even when extreme values of the tube coefficients are encountered.

It is evident that any electrode or combination of electrodes can be connected to function as the control grid and any other electrode or combination as the plate or operating electrode. It is sometimes important, for example, to determine the dynamic resistance in the control grid circuit when the grid is taking current, and the parameters corresponding to amplification factor and transconductance can be measured showing the influence of the other elec-

FIGURE 3. Schematic diagram of the circuit for measuring transconductance.



trodes on the grid current. Another quantity frequently of interest is the amplification factor of the screen electrode with respect to the plate current. The new cord and jack arrangement permits special measurements of this kind to be made as simply and directly as those of the usual quantities.

The three-tube coefficients are defined in terms of small incremental voltages or currents in the grid and plate circuits, and the bridge provides two independent 1000-cycle sources which can be varied over wide limits and inserted in any required circuit of the tube. Figure 3 shows the arrangement for transconductance measurements. The two separately-variable voltages for the grid and plate circuits are derived from the same 1000-cycle source. In order that the power supplies can be at ground potential and connected at one side directly to the cathode, the two voltages are obtained from the windings of separate transformers. On the secondary side of each transformer is a step attenuator for varying the voltage by factors of 10. In the primary of one of the transformers is a decimal attenuator giving to three significant figures the relative value of the plate-circuit voltage e_2. The grid-circuit voltage e_1 causes an alternating current to flow through the output transformer in the plate circuit. The voltage e_2, acting through the standard resistance R_s sends a current in opposite phase, and balance is obtained when the two currents through the output transformer cancel. The condition of balance is $e_1 S_m = e_2/R_s$ or $S_m = \frac{e_2{}^1}{e_1 R_s}$ and the transconductance is determined by the ratio of the two test voltages and the standard resistance. Any quadrature component through the output transformer, resulting from the tube interelectrode capacities, can be balanced out by the voltage of the extra split secondary, acting through the double-stator condenser. This adjustment does not affect the balance conditions for the in-phase components and consequently has no effect on the measurement.

A simple rearrangement of the two standard voltage sources permits the measurement of each of the other two coefficients, amplification factor, and plate resistance. In each case, the required quantity is expressed simply in terms of the ratio of the two test voltages, and can be read to three significant figures by the setting of the decimal attenuator. This is adjusted by the three controls at the bottom of the panel. A detailed analysis of the measuring circuits can be found in a paper† describing the original TYPE 561-A instrument.

— W. N. TUTTLE

General performance specifications for the TYPE 561-D Vacuum-Tube Bridge are identical with those for the older TYPE 561-C. The price remains unchanged at $375.00.

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†W. N. Tuttle, "Dynamic Measurement of Electron Tube Coefficients," Proc. I.R.E., No. 21, pp. 884–857, June, 1933.

CORRECTION

In the price list for TYPE 723 Vacuum-Tube Fork, appearing in the October issue of the *Experimenter*, there occurred a transposition of code words for the a-c power supply and the set of batteries. The correct code word for the TYPE 723-P1 A–C Power Supply is SNAKEYPACK, and that for the set of replacement batteries is SNAKEYBATT.

IMPEDANCE BRIDGES ASSEMBLED FROM LABORATORY PARTS

PART V—INDUCTANCE MEASUREMENTS

● **THE MEASUREMENT** of capacitance and dissipation factor of condensers by means of "bread board" bridges assembled from standard laboratory components has been discussed in previous *Experimenter* articles. Using similar techniques and equipment it is possible to make excellent measurements of inductance and reasonably good measurements of Q.[1]

Inductance can be measured by comparison with a known inductance using a ratio arm,[2] or "simple impedance" bridge, or in terms of capacitance using a product arm[2] bridge. In the latter class are included the bridges of Hay, Owen, and Maxwell. These all measure inductance in terms of the product of two resistances and a capacitance, but differ among themselves in the method of obtaining the resistance balance. In Figure 1 is shown a generalized bridge circuit suitable for the measurement of inductance, together with the complete equations of balance.

COMPONENTS REQUIRED

In setting up circuits to measure inductance, it would be desirable to utilize the same elements that were used for the capacitance bridges previously discussed. For the series-resistance bridge, the availability of three decade-resistance boxes, and a good mica standard condenser (0.01 μf) as well as a Type 578-A Transformer and the necessary oscillator and detector was assumed. For the Schering circuit a Type 722 Precision Condenser and an inexpensive air condenser were required in addition. Circuits very similar to these two capacitance bridges, differing only in that one resistance arm and the capacitance arm are interchanged in po-

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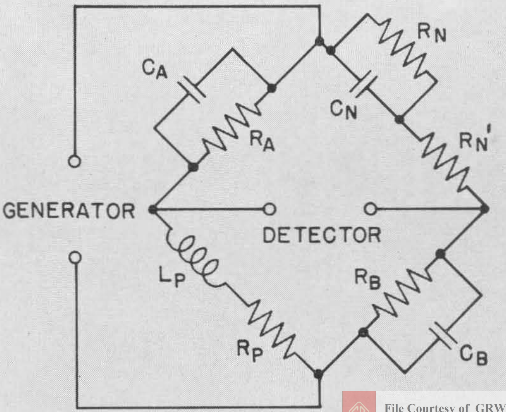
[1] The symbol Q is used to denote the ratio of series resistance to series reactance. For an inductor, this is $\frac{\omega L}{R}$. The term "storage factor" is also used for this quantity.

[2] The designations "ratio arm bridge" and "product arm bridge" mean simply that, for the former, the unknown reactance is determined by the *ratio* of two resistance arms, while for the latter the two resistance arms appear as a product in the balance equation. This terminology, although not generally accepted, has found limited use in the literature. See, for example, "Classification of Bridge Methods," J. G. Ferguson, B.S.T.J., Vol. XII, 1933, pp. 452–468; also, "A Brief Summary of Bridge Networks," W. T. Seeley, *Electrical Engineering*, March, 1940, pp. 108–113.

Figure 1. A generalized product arm bridge, showing complex impedances in all arms. The inductance in the P arm is measured in terms of the resistive product arms, A and B, and the capacitance standard C_N.



$$D_N = \frac{1}{Q_N} \cong R'_N\omega C_N + \frac{1}{R_N\omega C_N}$$

$$D'_N = R'_N\omega C_N$$
$$Q_A = R_A\omega C_A$$
$$Q_B = R_B\omega C_B$$
$$Q_P = \frac{\omega\, L_P}{R_P}$$

$$L_P = R_A R_B C_N \frac{1 - D_N Q_A - D_N Q_B - Q_A Q_B}{(1 + {D'_N}^2)\,(1 + Q_A^2)\,(1 + Q_B^2)}$$

$$Q_P = \frac{Q_N - Q_A - Q_B - Q_N Q_A Q_B}{1 - Q_A Q_B + Q_N Q_A + Q_N Q_B}$$

$$= \frac{1 - Q_A Q_B - D_N Q_A - D_N Q_B}{D_N + Q_A + Q_B - D_N Q_A Q_B}$$

sition, can be set up for inductance measurements, utilizing the same components.

THE HAY AND MAXWELL BRIDGES

A close analogue of the series-resistance type of capacitance bridge is the Hay bridge, shown in Figure 2a with simplified equations of balance.

The expression for the storage factor may be written in reciprocal form as

$$\frac{1}{Q_P} = D_P = \frac{R_P}{\omega L_P} = D_N + Q_A + Q_B + D_{N_O}$$

where D_{N_O} is the dissipation factor of the standard condenser itself, and D_N that added by the series resistor N.

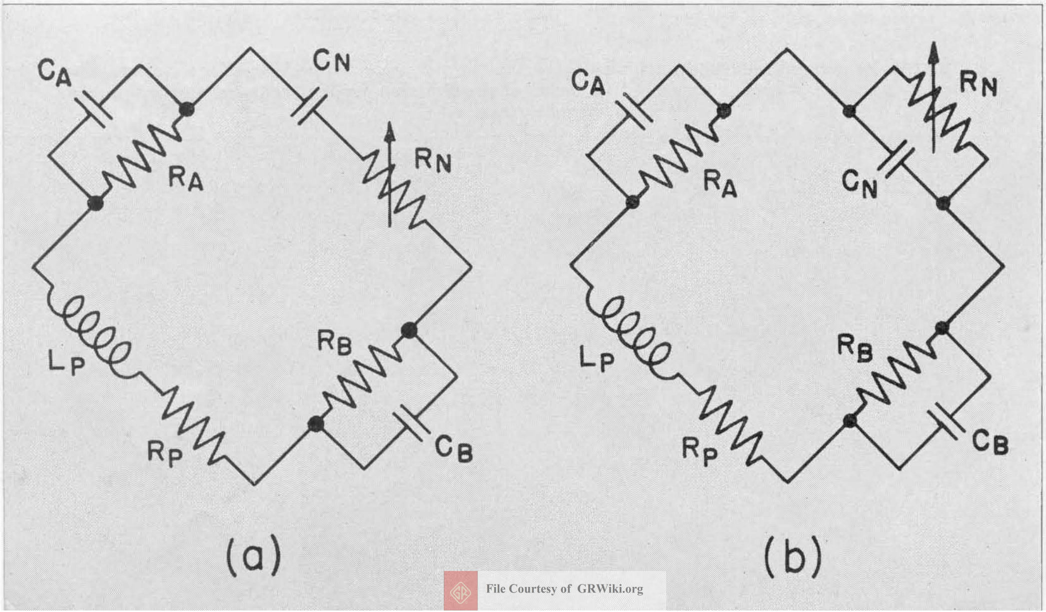
In this form it is identical with the corresponding equation for the analogous capacitance bridge, *except* that all the residual factors *add*, whereas the Q's of the ratio arms in a capacitance bridge are *subtractive.* This fact warns us immediately that, in general, less accuracy can be expected in measuring Q than for the corresponding quantity for a condenser. It is also evident that the residuals cannot be neutralized by adjusting capacitance across resistive arms.

MAGNITUDE OF RESIDUALS

From our knowledge of the magnitude of the residuals we can readily estimate the order of magnitude of the error to be expected. Suppose we connect the bridge transformer in such a manner as to place the secondary shield-to-ground capacitance across the standard condenser (0.01 μf), and set R_A and R_B to (say) 10,000 ohms. We may then use the values given in a previous article[3] discussing the series-resistance capacitance bridge. These values were $Q_A = Q_B = 0.0008$, $D_{N_O} = 0.001$, which give approximately 0.0025 for the maximum value of $Q_A + Q_B + D_{N_O}$. This corresponds to an error of 25% in measuring a Q of 100, and to an error of 2% for $Q = 10$,

FIGURE 2. Showing the Hay (*a*) and Maxwell (*b*) bridges. The two circuits differ only in the method of balancing the resistive component of the unknown. The Hay circuit utilizes resistance in *series* with the standard condenser, while in Maxwell's arrangement a parallel resistance is used. The approximate equations of balance are:

Hay: $L_P = \frac{R_A R_B C_N}{1 + D_N^2}$ $\quad Q_P = \frac{1}{D_N + Q_A + Q_B}$ $\quad$ Maxwell: $L_P = R_A R_B C_N$ $\quad Q_P = \frac{Q_N}{1 + Q_N(Q_A + Q_B)}$



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[3] General Radio *Experimenter*, July, 1941, p. 7.

if the simple relation $Q = \frac{1}{D_N}$ is used.

This accuracy, at high values of Q, is rather poor but in many practical cases a knowledge of the approximate value of Q is sufficient, even when it is desired to know the inductance quite accurately. The error in the indicated value of Q can always be reduced somewhat, of course, by inserting the estimated values of the residual terms into the equations of balance.

The accuracy of inductance measurement with the Hay circuit will be limited largely by the accuracy with which the capacitance standard is known. With the standard known to ±0.1% an accuracy of approximately ±0.3% can be achieved. One inconvenient feature of the Hay bridge, however, is the factor $\frac{1}{1 + D_N{}^2}$ which appears as a multiplier in the equation for series inductance. For a coil whose Q is 30, this correction amounts to 0.1%, while at $Q = 10$ the correction is 1%. This correction cannot be ignored except at high values of Q, and for this reason Maxwell's arrangement is frequently preferred for coils of low Q.[4] In this arrangement (Figure 2b) the coil resistance is balanced by a resistance in *parallel* with C_N, and no first order correction factor appears in the inductance equation. The difference in the inductance equation between these two bridges can be explained by saying that the product arm bridge measures series inductance in terms of the parallel capacitance of the opposite arm.[5]

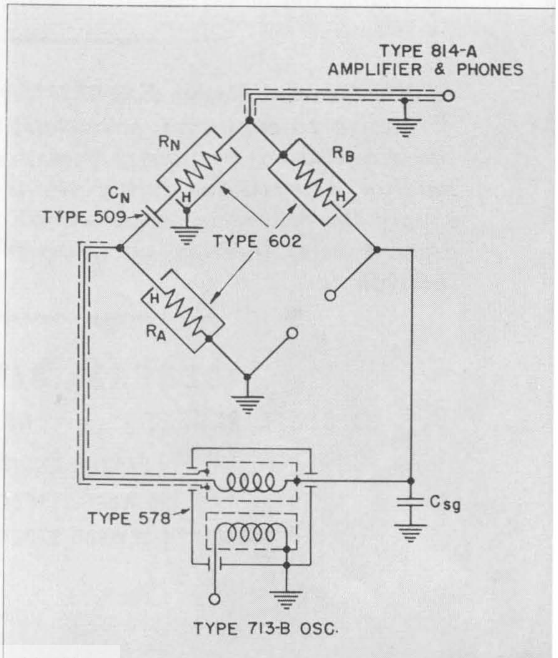
A bench layout utilizing the Hay and Maxwell circuits is shown in Figure 3. The components required are the same as those used for the series-resistance type of capacitance bridge described in a previous article. The disposition of the transformer terminal capacitances is different, however. The small (5 − 10 μμf) capacitance is placed across a resistive arm, while the larger (approximately 100 μμf) is placed across the unknown arm. The effect of this capacitance on inductors of 0.1 henry or less is negligible. For higher inductances, a satisfactory correction may be made, assuming a value of 100 μμf for this capacitance.[6]

For this particular configuration the sum of the residual Q's and D's is approximately 0.002.

A 100,000-ohm resistance box across C_N (= 0.01 μf) will balance Q's up to 6.3, while the same box may be placed in series with C_N for higher values of Q.

—IVAN G. EASTON

FIGURE 3. Connections for a bench layout of an inductance bridge. As shown, it is a Hay bridge, but can be converted to a Maxwell bridge by placing R_N in parallel with C_N.



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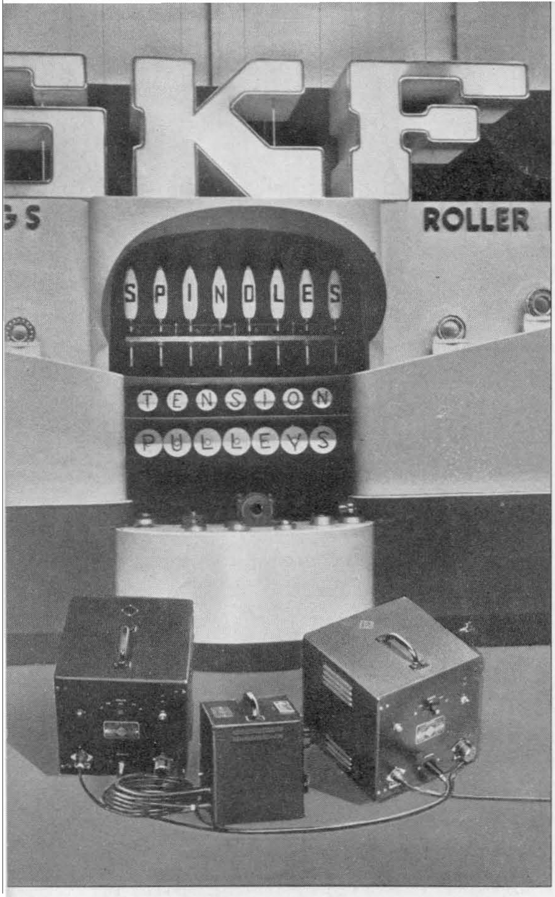
[4] In the TYPE 650-A Impedance Bridge, for instance, the Hay circuit is used for Q's above 10, and the Maxwell circuit for Q's below 10.

[5] Or conversely, measures *parallel* inductance in terms of *series* capacitance. This point will be discussed more fully in a subsequent installment.

[6] The correction is 1% for $L = 2$ henries. An error of 10% in the assumed value of C_{SG} will, therefore, introduce an error of 0.1% in the computed value of L.

GETTING DISPLAY INTEREST WITH THE STROBOLUX



● **AN INTERESTING APPLICATION** of the Strobotac and Strobolux to advertising is shown in the accompanying photograph of the display used by SKF Industries, Inc., at the Southern Textile Exhibit held last spring.

The eight spindles, each bearing a letter, were driven at a constant speed by a chain drive. The two rows of tension pulleys were driven similarly, but at a different speed from that of the spindles. Contactors on one spindle and one tension pulley provided a means of flashing the Strobotac. Controlled by the Strobotac were two Strobolux units, illuminating the spindles and tension pulleys.

A motor-driven switching system provided a repeating cycle of operation. First an ordinary incandescent lamp showed all parts rotating at high speed. Next the Stroboluxes, synchronized to the spindle speed, showed the spindles apparently stationary and spelling out the word SPINDLES. Finally the Stroboluxes were controlled by the speed of the tension pulleys, making the letters on them readable and those on the spindles illegible.

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THE General Radio EXPERIMENTER

VOLUME XVI No. 7 DECEMBER, 1941

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

IMPEDANCE BRIDGES ASSEMBLED FROM LABORATORY PARTS

PART VI—INDUCTANCE MEASUREMENTS

Also
IN THIS ISSUE

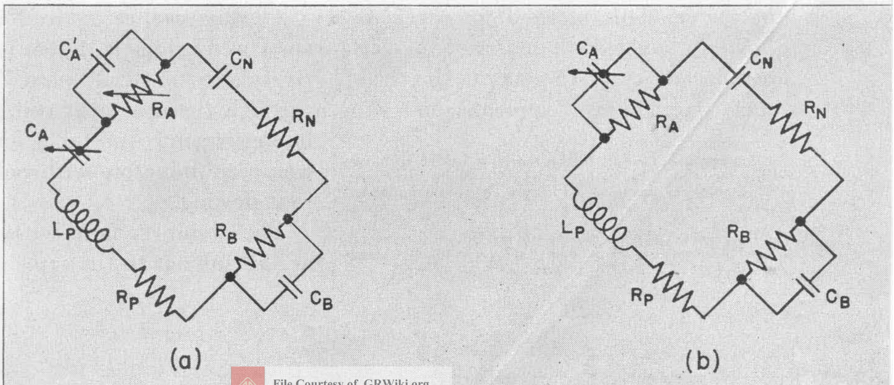
INDEPENDENCE OF BALANCE

● **IN ANY ALTERNATING-CURRENT BRIDGE** there are two conditions that must be simultaneously satisfied to obtain a true null balance. For maximum convenience in the use of the bridge it is desirable that the two adjustments for balance be independent of each other, so that the element that is varied to secure one balance shall not affect the other balance. Otherwise, the condition commonly known as a "sliding zero" occurs. It is characterized by the fact that balance must be approached by comparing a number of successive adjustments for minimum. The degree of dependency of the two components of balance

FIGURE 1. Schematic diagrams of (a) the Owen bridge, and (b) the "Schering," or parallel, form of the Owen bridge. The conventional form of the Owen bridge balances the losses in the P-arm by a condenser in series with R_A, while the circuit shown in (b) utilizes a parallel condenser across the A-arm. Circuit residuals are represented by R_N, C'_A and C_B in both diagrams. The approximate equations of balance are

$$\text{Owen:}\qquad L_P = R_A R_B C_N;\ Q_P = \frac{D_A}{1 + D_A(Q_B + D_N + Q'_A)}$$

$$\text{"Schering":}\ L_P = \frac{R_A R_B C_N}{1 + Q_A{}^2};\ Q_P = \frac{1}{Q_A + Q_B + D_N}$$



on each other (i.e., the amount of "sliding") is dependent only on the storage factor, Q, of the unknown impedance.[1] The higher the storage factor of the unknown impedance, the less pronounced is the sliding effect.

It can be shown that truly independent balances are obtained only when the two adjustments for balance *are made in the same arm*, or when one adjustment is made in each complex arm. An example of the first method is the Owen bridge, while a bridge devised by Sinclair[2] illustrates the second.

The two elements that provide independent balances can be made direct reading for the resistive and reactive (or conductive and susceptive) components of the unknown impedance, independent of each other.

As an example, consider the Hay and Maxwell bridges, already discussed. Both bridges are subject to a sliding balance, since the elements that are varied to secure balance do not satisfy the conditions stated above. The Hay circuit is commonly used for inductors of high Q, and the effect is not pronounced. The Maxwell bridge, however, is frequently used for measurements of low storage factors, and in this case the sliding zero becomes very noticeable. If C_N and R_N (where N is the arm opposite the unknown inductance) are chosen as variables, the two components of balance are completely independent, as pointed out above. This method, however, requires a standard condenser of the variable, decade type. In general, the use of such a condenser will mean some sacrifice in accuracy of inductance measurements, as compared to the accuracy attainable with a fixed standard condenser. Furthermore, the bridge no longer can be made direct reading for Q. For these reasons, this arrangement is not commonly used, and the method suggested earlier (varying R_A and R_N) is generally preferred in spite of the concomitant sliding zero.

THE OWEN BRIDGE

Another well-known circuit for the measurement of inductance is the Owen bridge shown in Figure 1(a). In this circuit the resistive component of the unknown impedance is balanced by a capacitance in series with one of the resistive arms. If the reactive balance is obtained by varying this resistive arm, the two balances are independent, as shown by the equations of Figure 1. We have here a situation analogous to that pointed out above for the Maxwell bridge, wherein independent balances are secured if the resistance and capacitance of the standard arm are both varied.

For the Owen circuit the variable condenser determines the resistance balance, whereas for the Maxwell bridge the variable condenser determines the inductance balance. A decade condenser, with its relatively poor accuracy (typically 1%), is generally satisfactory for the resistance measurement, however, as larger errors from other sources generally determine the accuracy of measurement of this component.

Another well recognized advantage of the Owen bridge is that it is a comparatively easy matter to pass direct current through the unknown coil. This bridge is consequently suitable for measuring iron-core inductors with polarizing current flowing.

The circuits so far discussed, however, are all subject to the same limitation in

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[1] This statement refers to the four-arm bridge with two complex arms. If three or more arms are complex, the degree of dependency is expressed in a somewhat more complicated fashion.

[2] D. B. Sinclair, "A Radio-Frequency Bridge for Impedance Measurements from 400 Kilocycles to 60 Megacycles," *Proc. I.R.E.*, November, 1940, pp. 497–503.

measuring coils of high Q, namely that the error in the determination of Q (or of resistance) is directly proportional to Q, and an error of 25%, 50%, or even greater is not uncommon when Q is of the order of 100 or greater.

THE RESONANCE BRIDGE

The resonance bridge, shown in series form in Figure 2, is undoubtedly the most accurate method available for measuring coil resistance at audio frequencies. Since the reactance of the P arm is reduced to approximately zero, the resistance balance becomes the dominant one, and the circuit residuals have only a second-order effect on it, while having a comparatively large effect on the reactance balance. This method, then, although quite accurate for resistance measurements, is not very accurate for inductance measurements.

To obtain the coil resistance, when the total P arm resistance is known, obviously requires a knowledge of the effective resistance of the tuning condenser. Additional measurements are thus required, unless condensers of known dissipation factor are used, or unless R_C is negligible with respect to R_L.

THE "SCHERING" CIRCUIT

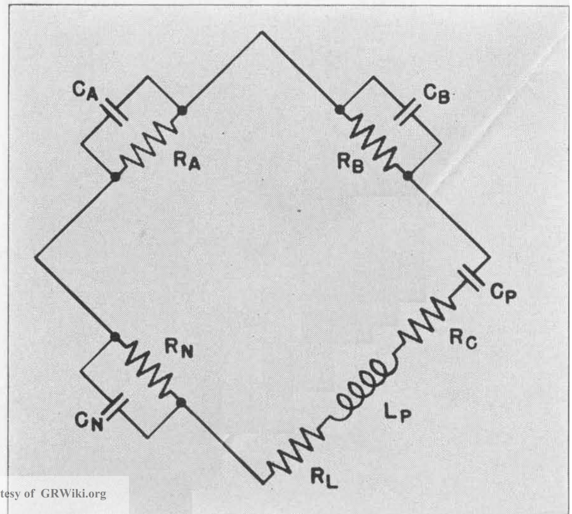
The analogue of the Schering (capacitance) bridge discussed in a previous article is the circuit shown in Figure 1(b).[3] Here the inductance is balanced by a precision air condenser in the opposite arm, while the resistance of the unknown is balanced by a capacitance across one of the fixed resistance arms. As is the case for the Hay bridge, the two components of balance are dependent on each other, but, for storage factors greater than 10, the annoyance from the sliding balance is not serious.

For capacitance measurements, one of the advantages of the Schering circuit lies in the fact that it is a relatively simple matter to establish an initial balance by adjusting the difference $Q_A - Q_B$, using trimmer condensers across the ratio arms. For inductance measurements, however, the Q's of the resistance arms and the condenser are directly additive, as pointed out in a previous article. Thus, in this case, it is not possible to use parallel capacitance across the resistance arms. The possibility immediately suggests itself, however, of using *series inductance* in one of the resistance arms[4] to compensate for the residual Q's of the circuit. The establishment of the initial balance requires an inductance of high Q, the value of which is known to a moderate accuracy, or, alternatively, a moderate value of Q, accurately known. This, in turn, re-

FIGURE 2. The series-resonance bridge. The capacitances C_A, C_B, and C_N are circuit residuals. The equations of balance are

$$R_P = \frac{R_B R_N}{R_A} \frac{1 - Q_A{}^2}{1 + Q_N(Q_A + Q_B) + Q_A Q_B}$$

$$L_P = \frac{1}{\omega^2 C_P} \text{(approximately)}$$



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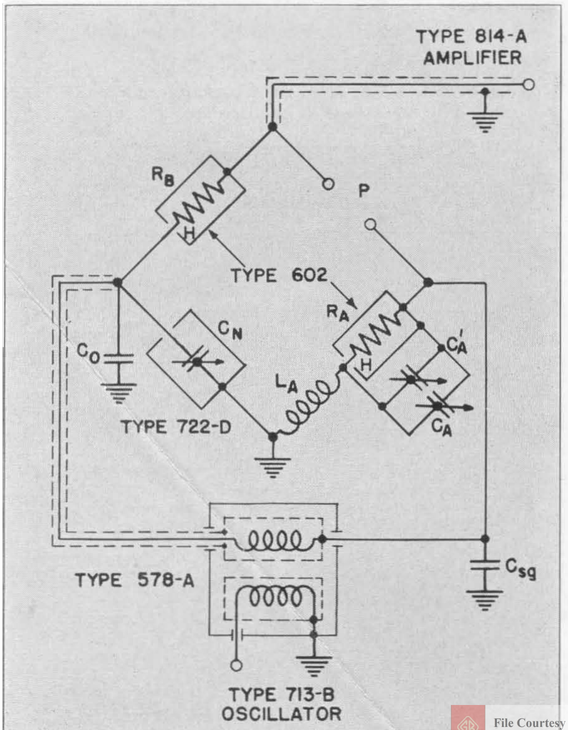
[3] This circuit may equally well be considered as a parallel form of the Owen bridge.

[4] This is comparable to Grover's arrangement for using series inductance in a ratio arm to measure the D of a condenser.

quires an independent method of measuring resistance accurately, but fortunately such a method is available in the series-resonance bridge.

A compensated circuit of the type outlined, using the same components as the Schering capacitance bridge already described,[5] is shown schematically in Figure 3. The secondary shield-to-ground capacitance of the transformer is placed across the resistance arm A, with the smaller (10 μμf) terminal capacitance across the capacitive arm N (Type 722-D Precision Condenser). A fixed inductance, L_A, is used, with an additional trimmer capacitor C'_A, to make the final adjustment in establishing the initial balance. For an A arm resistance of 20,000 ohms, a 50 or 100 mh choke[6] may conveniently be used for L_A, together with a 100 μμf condenser used for C_A.

Figure 3. Connections for a bench set-up of a parallel form of Owen bridge.



The stray capacitance placed across the standard condenser can be determined by a method similar to that used with the capacitance bridges. If two balances are made for a given L_P, one with the standard condenser disconnected at its high terminal, the other in the usual manner, with C_N set at about 100 μμf, the stray capacitance C_O will be given by

$$C_O = C_{N1} \frac{1}{\dfrac{B_2}{B_1} - 1}.$$

Here B_1 is the reading of the B arm resistance box with C_N disconnected, B_2 its reading with the standard set at a value of C_N.

With this arrangement an initial balance was established against a Q of 20 (known to about ±1%, from resonance bridge measurements). The Q's of several other coils were then measured, the results checking the known values within essentially the accuracy of reading of the condenser C_A. Additional correction terms are introduced into the inductance equation by the transformer capacitance shunting the series inductance L_A. For the case cited, this correction is negligible, but it can become significant if large inductances are used in the A arm.

— Ivan G. Easton

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[5] General Radio *Experimenter*, August, 1941, p. 2.

[6] The resistance of the choke must, of course, be added to R_A. The choke resistance is small compared to R_A, however, and the d-c value may be used without introducing any appreciable error.

THERMOCOUPLES

We regret that, owing to circumstances beyond our control, we can no longer supply vacuum thermocouples. Consequently, all models of Type 493 Thermocouples are discontinued, effective December 1.

USING THE CATHODE-RAY OSCILLOGRAPH IN FREQUENCY COMPARISONS

● **IN FREQUENCY MEASUREMENT AND CALIBRATION,** where an unknown frequency is to be compared with, or adjusted to, a standard frequency, the cathode-ray oscillograph offers a convenient and precise means of making the necessary comparison. The following summary of the various ways in which a cathode-ray oscillograph can be used advantageously in frequency measurement presents no new methods. The methods discussed are presented particularly for those who may have available a cathode-ray oscillograph, but who may not appreciate its potentialities in this field.

I LISSAJOUS FIGURES

Starting with the simplest comparison, if a voltage from a frequency standard is applied to one pair of the deflecting plates (say, the horizontal) of the cathode-ray oscillograph and a voltage from a source whose frequency is to be adjusted in terms of the standard to the other pair, patterns of the type illustrated in Figure 1 will be obtained.

These are the well-known Lissajous figures. For simple frequency ratios, expressible by small whole numbers, the patterns are not too complicated, and identification of the frequency ratio is possible, even when the pattern is rotating slowly.

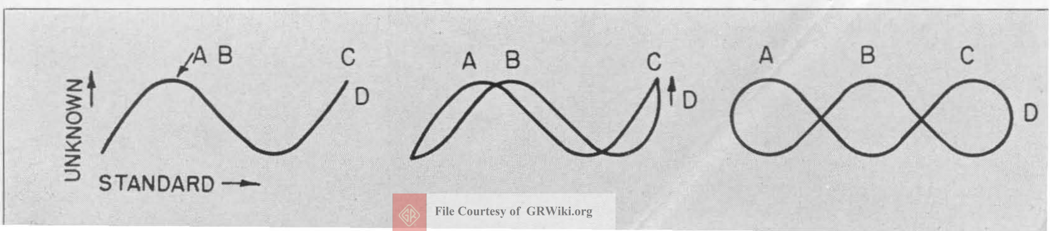
If the pattern can be made to be nearly stationary, by adjustment of the frequency to be checked, then the frequency ratio is found as follows: Count the horizontal tangent points (such as *A*, *B*, *C*, Figure 1); count the vertical tangent points (such as *D*). The frequency ratio is the ratio of the number of horizontal points to the number of vertical points, which, for the example of Figure 1, is 3 : 1. If the unknown frequency is on the vertical plates, then the unknown is three times (as illustrated in Figure 1) the standard frequency.

As indicated by the successive parts of Figure 1, the appearance of the pattern changes progressively if there is a slight difference in frequency between the unknown and standard frequencies. Under such conditions the tangent points can be counted only for simple frequency ratios. If a frequency ratio of 7 : 5 is obtained, for example, the pattern appears almost as a network covering the area and, unless the pattern is steady, it is very difficult to count the tangent points.

II MODULATED WAVE PATTERNS

A variation of this procedure occurs in the problem of matching a low beat frequency, obtained as a result of comparing an unknown radio frequency with a standard frequency. If this beat is obtained in an oscillating receiver, the oscillating frequency being offset appreciably (by a kilocycle or so from both of the radio frequencies), then the output of the receiver is an audible tone, the

FIGURE 1. Lissajous figures for a frequency ratio of 3 : 1. At the left, the two frequencies are shown in phase; at the center, slightly out of phase; and at the right, in quadrature.



amplitude of which waxes and wanes at a rate equal to the low beat frequency difference of the two radio frequencies.

If one of the two radio frequencies is of somewhat greater amplitude than the other, and if the receiver output is connected to the vertical plates while a calibrated audio oscillator is connected to the horizontal plates, then patterns of the type shown in Figure 2 are obtained.

The receiver output is equivalent to an audio-frequency carrier modulated by the beat-frequency difference. The pattern of Figure 2 is familiar as a means of checking the percentage modulation and indicating roughly the quality of modulation. The advantage of this method in frequency comparisons is that the audio-frequency system of the receiver is not called upon to transmit a frequency of only a few cycles; it transmits the audible carrier, which may be placed anywhere in the audible range. Also, the matching oscillator can be operated at a multiple of the beat frequency, in a range where the accuracy of its calibration is usually much improved.

In Figure 2, the illustration is made for a case in which one radio frequency is roughly twice the amplitude of the other, resulting in a modulation of the audible carrier of $A/B = 0.5$ approximately, or 50%. If the matching oscillator frequency is not exactly equal to the beat frequency, the pattern will slowly progress through the sequence indicated. The illusion of a three-dimensional figure is strong; the pattern appears like a tube, with the ends cut at an angle to the axis.

If the matching oscillator frequency is adjusted to a multiple of, or in simple ratios to, the beat frequency, the pattern developed at the ends of the figure corresponds to the Lissajous figures for the same ratios. If the matching oscillator is adjusted to twice the beat frequency, a pattern of the type shown in Figure 3 results. It will be seen that a "two-to-one" pattern is developed at each end of the figure or that, in illusion, two tubes are developed, side by side. If the matching frequency is not exactly twice the beat frequency, the pattern changes progressively through the sequence indicated.

III CIRCULAR SWEEP PATTERN

Very useful patterns are obtained if a circular sweep is used. These patterns are of a form in which the frequency ratio is easily identified, even when the ratio is expressed by numbers which are not small integers.

To produce a circular sweep, it is necessary to obtain two equal voltages having a phase difference of 90 degrees from the standard frequency source. One method is illustrated in Figure 4.

The standard frequency is supplied through transformer *T*-1, to match the total load. Resistances, *R*, and reactances, *C*, are made equal in magnitude,

FIGURE 2. Modulated wave pattern. Left, in phase; center, slightly out of phase; right, in quadrature.

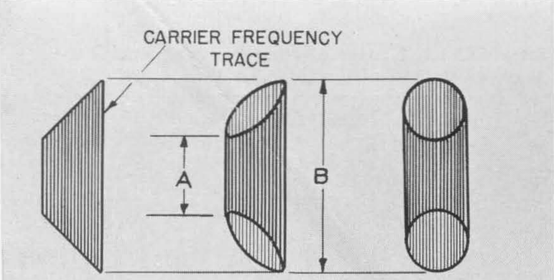
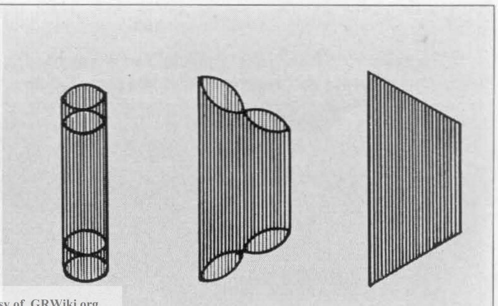


FIGURE 3. Successive phases of a modulated wave pattern when the matching frequency is twice the modulation frequency.



for convenience. A value of 10,000 ohms at the standard frequency is handy. The phases of the voltages taken to the primaries of *T*-2 and *T*-3 are 45 degrees ahead and behind the voltage supplied from *T*-1, or have a phase difference of 90 degrees. If the impedances of *T*-2 and *T*-3 are high (interstage coupling transformers with a step-up ratio of 6 : 1 are suitable), then connecting them across the elements *R* and *C* will not materially affect the phase of the voltages. The combined effects of the transformer loading and of the impedances across the secondaries generally make it necessary to readjust the elements of the phase shifter somewhat.

If the standard frequency supply is distorted in waveform, a circular sweep cannot be obtained. A low-pass filter is then necessary, as indicated in the diagram.

If a cathode-ray oscillograph having a radial deflector is used, the unknown frequency can be placed on that electrode. In the more general case, it is convenient to introduce the unknown on the vertical plates as shown at *T*-4.

The type of patterns obtained with a circular sweep are illustrated in Figure 5. With no unknown frequency introduced, the pattern is a circle, with the spot traveling once around the circle for each cycle of the standard frequency. If a frequency equal to five times the standard frequency is introduced on the vertical plates, a pattern as illustrated in the second part of the figure will result. The frequency ratio can be determined by counting the tops of the waves, as at *A*, *B*, *C*, *D*, *E*. (If radial deflection is used, the pattern is not distorted, and the frequency ratio is found by counting the outer tips of the waves.)

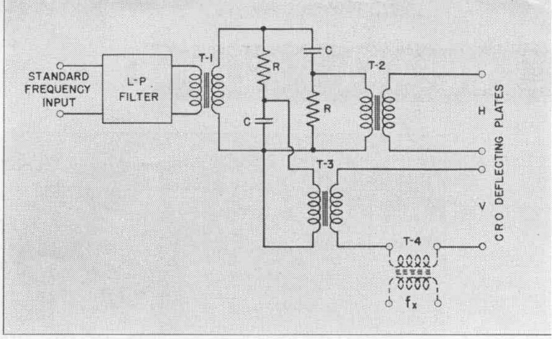


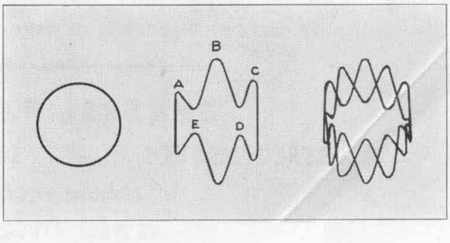
FIGURE 4. Connections for obtaining a circular sweep.

Note that, even when the unknown frequency is not set exactly to five times the standard frequency, the pattern does not change form, but appears to *rotate* slowly. The successive transitions from one line to two lines (as in Figure 1) do not occur. The pattern remains a "one-line" pattern as long as the frequency ratio is not far from any integral multiple of the standard frequency.

If the unknown were adjusted to one-fifth the standard frequency, a set of five nearly concentric circles would be seen. (Not illustrated.)

The "one-line" pattern is extremely convenient when using equipment whose calibration is known to be nearly correct. For example, an audio-frequency oscillator may be known to have no error greater than a very few cycles at any point in its range. If it is desired to set this oscillator to, say, 1700 cycles in terms of the standard (say 100 cycles), then simply set the oscillator to 1700 cycles by its calibration. The pattern will then be a "one-line" pattern (having 17 "tops," but it is not necessary to

FIGURE 5. Circular sweep patterns. Left, standard-frequency circular sweep; center, circular pattern with superimposed frequency equal to five times the standard; right, with superimposed frequency equal to 9/2 the standard.



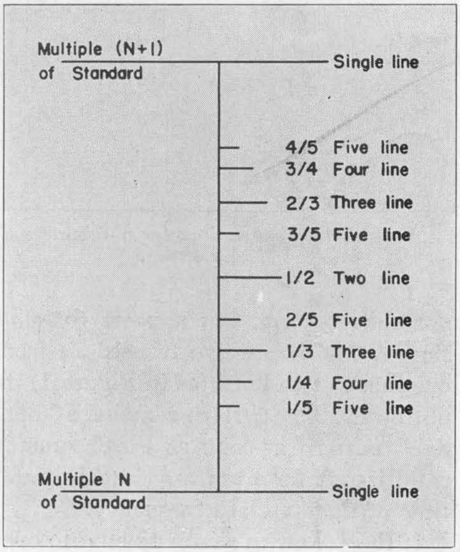


Figure 6. Table showing the sequence of patterns in each standard-frequency interval.

count them), rotating at a rate depending on the error of the oscillator at the setting of 1700 cycles. Readjust the oscillator slightly until the pattern stands still, when the frequency will be 1700 cycles in terms of the standard.

It will be seen from the above that an audio-frequency oscillator can be calibrated using "single-line" patterns only, at every 100 cycles (from a 100-cycle standard) up to the highest frequency at which the successive waves on the pattern can still be distinguished.

If the oscillator is set to an odd multiple of one-half the standard frequency, a "two-line" pattern, illustrated in the third part of Figure 5, is obtained. Using "one-" and "two-line" patterns, the oscillator can be calibrated at every 50 cycles up to a limit determined by the ability of the observer to distinguish the pattern.

In a similar manner, "three-line" patterns are obtained when the oscillator is set at one-third and two-thirds of the way between successive 100-cycle points; "four-line" patterns when set at one-quarter and three-quarters, etc.

The important feature of this method is that the sequence of patterns repeats in *each* 100-cycle interval, as illustrated in Figure 6. Since the 100-cycle multiples are readily identified, it becomes a simple matter to calibrate an oscillator within 20 cycles of any desired audio frequency.

For purposes of illustration, reference has been made throughout the above to a standard frequency of 100 cycles, which is a convenient value for use in measurements up to a few thousand cycles. If the standard frequency be multiplied by 10, the patterns obtained will be identical at frequencies ten times higher than before, giving a useful range up to the low radio frequencies.

The General Radio Type 699 Comparison Oscilloscope has been designed particularly for use with the Class C-21-HLD Primary Frequency Standard and Frequency Measuring Equipment. Provision is made for permanent shielded wiring to all necessary components of the standard and measuring equipment. By means of key switches, the desired sources and method of comparison can be selected. One hundred-cycle and one thousand-cycle filters and phase shifters are provided for obtaining circular sweeps at either frequency.

— J. K. Clapp

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