the GENERAL RADIO Experimenter

INDEX

TO

GENERAL RADIO EXPERIMENTER

VOLUMES XVI AND XVII

June, 1941 to May, 1943

GENERAL RADIO COMPANY

CAMBRIDGE MASSACHUSETTS

U. S. A.

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THE General Radio EXPERIMENTER

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

AN A-C-OPERATED POWER SUPPLY FOR THE SOUND-LEVEL METER

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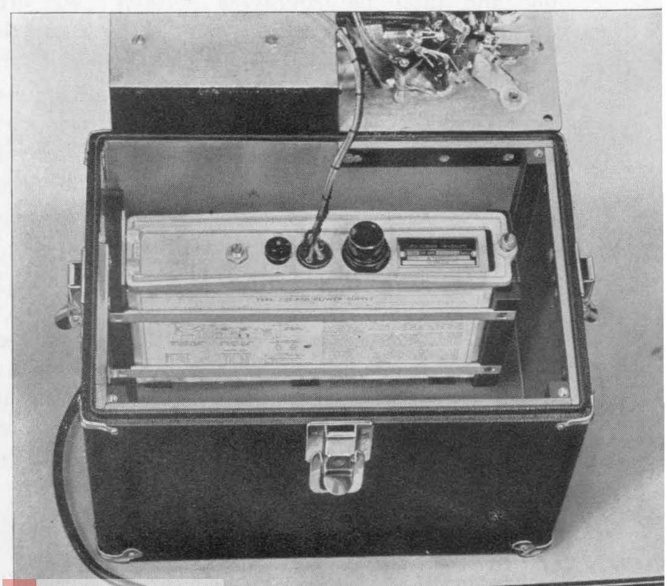
● MANUFACTURERS OF SOUND-LEVEL METERS are generally divided on the subject of a-c operation versus battery operation. For portable use, battery operation is almost essential. For stationary operation or use only indoors, where power lines are available, a-c operation has some advantages, while for production work, where continuous operation is required, the a-c power supply has a distinct advantage in that it eliminates the need for frequent replacement of batteries.

It is apparent that, for really universal application, a sound-level meter should be capable of operation either from the a-c power line or from batteries. However, since conventional tube types do not operate equally satisfactorily on both alternating current and small, portable dry cells, and because high-gain amplifiers represent a serious problem in regard to hum elimination when a-c operated, most sound-level meters have been designed for ei-

FIGURE 1. View of the power-supply unit installed in the battery compartment of the sound-level meter cabinet.



ther one or the other type of power supply, but not for universal operation. This seriously restricts the use of any such instrument. Many engineers and laboratories have need for only one sound-level meter and are somewhat at a loss, therefore, as to which type will best suit their requirements.

A logical answer to this problem is provided by the new TYPE 759-P50 Power Supply developed by the General Radio Company for use with its TYPES 759-A and 759-B Sound-Level Meters. In keeping with the company's policy of retarding obsolescence so far as possible, the power supply can be used with the earliest meters in the 759 series as well as the latest. It is small, light, and compact and fits directly in the battery compartment of the sound-level meter in place of the batteries.

The power supply includes an oxide rectifier and suitable filter circuits which supply either 3 or 1.5 volts for operation of the two filaments (depending, respectively, upon whether an A-type or B-type sound-level meter is used). The unit includes also a vacuum-tube rectifier and filter for supplying suitable plate voltage.

Hum level in the power supply has been kept so low that the TYPE 759-B Sound-Level Meter can be used over its entire sensitivity range with this new power supply. The TYPE 759-A Meter can be used down to 34 decibels, which is entirely adequate for most machinery problems such as are encountered in production testing.

In designing the new power supply, an interesting problem was encountered in the development of a suitable filter for the filament circuit. Because of the high gain of the sound-level meter amplifier and the fact that filament-type tubes, originally intended only for battery operation, are used, it was found that small line-voltage fluctuations would momentarily shift the gain, causing fluctuation in the reading of the meter. While high-capacity electrolytic condensers satisfactorily eliminated all ripple frequencies from the filament voltage, some low-frequency variations were present, particularly with a poorly regulated or noisy power-supply line. The final solution was to use two flashlight cells in the last stage of the filter in place of a condenser. These function satisfactorily as a condenser, but also have the additional advantage that they maintain substantially constant voltage. When used with the TYPE 759-B Sound-Level Meter the cells are connected in parallel, and when used with the A-type they are connected in series. This transformation is accomplished by a simple plug which is inserted into a socket on the top of the power-supply unit.

When the instrument is operating, the cells are charging slightly, so that their life is practically equal to their normal shelf life. When the instrument is turned off, a small relay, built as part of one of the filter chokes, opens the circuit so that the cells will not run down. The cells are of the standard flashlight variety, readily replaceable, and cost only ten cents each. However, under normal line-voltage conditions, their life is six months, a year, or even longer.

The convenience of the power supply is its outstanding feature. At any time it is possible to interchange power supply and batteries immediately without any rewiring or circuit changes. This makes the same sound-level meter readily adaptable for production testing or field work and is a real source of economy in laboratories requiring only one meter.

No alterations to the sound-level

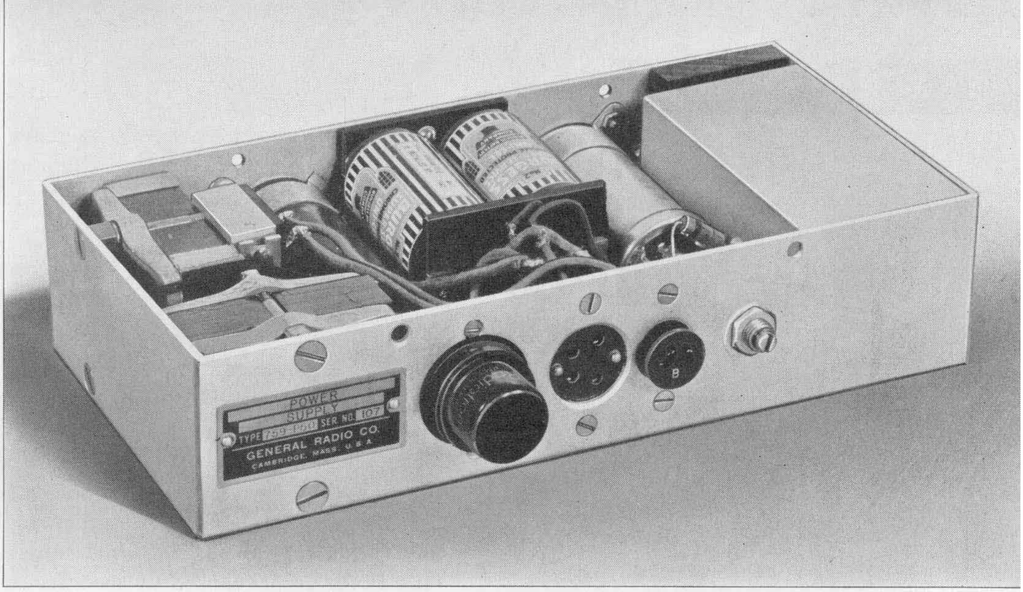


FIGURE 2. View of the power-supply unit showing the compact arrangement of parts.

meter are necessary when the power supply is installed in a TYPE 759-B Sound-Level Meter, and only minor changes are required for the TYPE 759-A.

Complete directions and a kit of parts are supplied with each power unit so that these alterations can be easily made by the user. It is not necessary to return the instrument to the factory.

— H. H. SCOTT

SPECIFICATIONS

Output: 1.5-volt and 3-volt filament supplies; 90-volt plate supply.

Hum and Noise Level: Sufficiently low to assure satisfactory operation over the entire range of the TYPE 759-B Sound-Level Meter when the supply-line frequency is 60 cycles. On the TYPE 759-A Sound-Level Meter, satisfactory operation is obtained on all ranges except at the 60 db attenuator setting, provided the a-c line frequency is 60 cycles. Operation from line frequencies below 60 cycles is possible, but is not recommended.

Input: 105 to 125 volts, 40 to 60 cycles. The power input is less than 8 watts at 115 volts, 60 cycles.

Tube: One type 6H6 tube is supplied.

Terminals: An output socket fits the plug on the battery cable of the TYPE 759-B Sound-Level Meter.

Dimensions: (Length) 10 x (width) 2¼ x (depth) 5 inches, over-all.

Net Weight: 7 pounds, 6 ounces.

Type		*Code Word*	*Price*
759-P50	A-C Power Supply......	NUTTY	$55.00

PRIORITIES

Because practically all of our manufacturing facilities are devoted to National Defense projects, a preference rating certificate or other approved priority rating will be necessary to secure delivery. At the present time a rating of A-10 or higher is required for delivery of all instruments and parts, but for certain items in especially heavy demand a rating of A-2 or higher may be necessary to insure reasonable delivery.

IMPEDANCE BRIDGES ASSEMBLED FROM LABORATORY PARTS

PART VII—MEASUREMENT OF DIRECT CAPACITANCE*

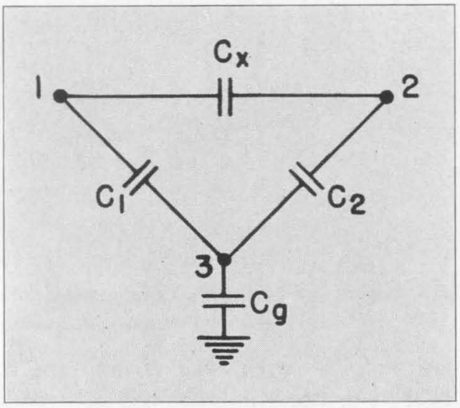


FIGURE 1. Representation of a capacitor which has, in addition to its direct capacitance (C_X), capacitance from each terminal to a third terminal. The third terminal can represent a shield, which itself has a capacitance to ground.

● **ANY CAPACITOR** which does not have one of its terminals grounded is effectively a three or four-terminal impedance. In addition to the desired capacitance between terminals, there exist stray capacitances from each terminal either to ground or to a shield, which has a capacitance to ground. The situation is illustrated in Figure 1. Since measuring circuits of the type under discussion are grounded at some point, consideration must be given to these capacitances when it is desired to measure the direct capacitance between terminals 1 and 2. In many practical cases, of course, the extraneous capacitances are negligible compared to the desired direct capacitance, and accurate measurements can be made simply by connecting terminals 1 and 2 to the measuring circuit. When C_1, C_2, and C_g are of the same order of magnitude as C_X, however, a direct measurement can obviously not be made. If, however, the terminal impedances are large compared to the bridge arms, it is possible to connect the third terminal to the bridge in such a manner that the terminal impedances and the direct impedance are separated, and the latter can be measured subject only to the errors caused by the effect of the terminal impedances placed across the bridge arms. A few typical measurements of this sort are illustrated in Figure 2. Figure 2 (a) shows a three-terminal condenser connected to a capacitance bridge, the junction of whose capacitance arms is grounded. As shown, the third terminal is connected to the junction of the ratio arms, placing C_2 across the arm B, and placing C_1 in parallel with C_g across opposite corners of the bridge, where they do not influence the balance. The presence of C_2 across the arm B causes the dissipation factor reading of the bridge to be high by an amount $R_B\omega C_2$. The bridge reads correctly for C_X unless C_2 is sufficiently large to bring in the terms which are normally neglected in the simplified balance equations, or unless the losses and leakage in C_2 are sufficient to reduce appreciably the effective parallel resistance of the B arm.

In Figure 2(b) the third terminal is connected to the junction of the A and N arms. Here C_2 is effectively removed from circuit, while C_1 and C_g parallel the standard condenser, causing a direct

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*Much of the material discussed in this installment has appeared in previous *Experimenter* articles and elsewhere, but is included here for the sake of completeness of the current series.

error in the capacitance balance, the magnitude of which depends on the ratio $\frac{C_1 + C_g}{C_N}$. The dielectric losses associated with C_1 and C_g increase the effective dissipation factor of the standard arm. Consequently the dissipation factor reading of the bridge is low, and may easily become negative.

If the third terminal is ground (C_g infinite in Figure 1), a direct measurement of C cannot be made with a bridge that is grounded at the junction of the capacitance arms, since the direct capacitance is paralleled with one of the terminal capacitances when it is connected to the bridge. By making three sets of measurements, however, with successive pairs of the three capacitances connected in parallel, data can be obtained from which the constants of the terminal capacitances as well as the direct capacitances can be computed.[1]

With a bridge grounded at any point other than one side of the unknown, the direct measurement can be made even when the third terminal is itself grounded. This is illustrated by Figure 2(c), wherein the junction of the ratio arms is grounded. With this connection the capacitance C_2 is placed across arm B, introducing an error of $R_B\omega C_2$ in the dissipation factor balance. Figure 2(d) shows a fourth arrangement, wherein the bridge is grounded at the junction of dissimilar arms. With the third terminal connected between the ratio arms, C_2 parallels the B arm and the ground capacitance C_g parallels the A arm, while C_1 does not affect the measurement. Under these conditions the error in the dissipation factor caused by the presence of the extraneous capacitances is $(R_B\omega C_2 - R_A\omega C_g)$, while the capacitance balance is substantially correct.

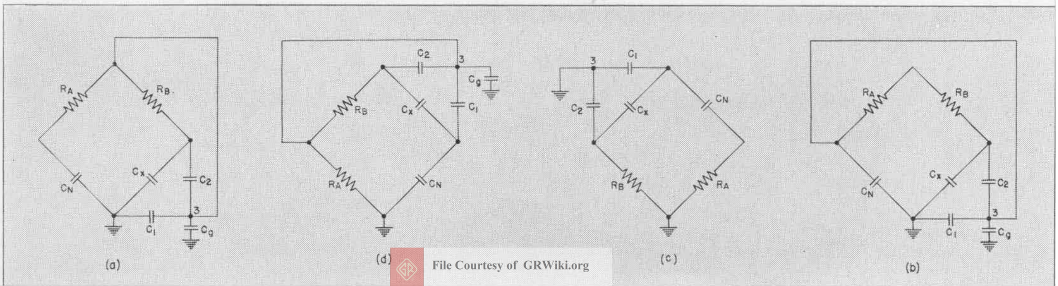
As is the case with most impedance measurements, somewhat better accuracy can generally be obtained by the use of a substitution method, as indicated in Figure 3. If, as is customary, the disconnection is made at the high side of the N arm, an error equal to $A\omega C_1$ is introduced in the dissipation factor reading by the capacitance C_1 being placed across the A arm when the connection is made. If, on the other hand, the disconnection is made at the ground side, the capacitance C_1 is across the A arm for both balances, and the change in the capacitance shunting R_A is $\frac{C_X C_2}{C_X + C_2}$. Thus the error encountered in the dissipation factor measurement is different for the two methods of disconnection. The best method can be determined only by a consideration of the relative magnitudes of C_X, C_1, and C_2.

GUARD CIRCUITS

If direct measurements, independent of the terminal capacitance, are desired, the measuring circuits and procedure must necessarily be made more complex. This is commonly done by means of Wagner grounds, guard circuits, and

[1] R. F. Field, "Direct Capacitance and Its Measurement," General Radio *Experimenter*, Vol. VIII, No. 6, November 1933.

FIGURE 2. Showing various methods of connecting the capacitance network of Figure 1 into a bridge, to obtain a measurement of C_X.



similar arrangements which provide an auxiliary circuit to which the third terminal can be connected. The terminal impedances then become a part of the auxiliary circuit and are balanced out. Although a number of different arrangements are possible, they all serve essentially the same purpose — namely, to bring the third terminal of the unknown impedance to the same potential as one corner of the bridge to which it is not normally connected.[2]

Figure 4 illustrates an impedance bridge with all four corners coupled to a common fifth point. In spite of the presence of the four additional impedances, a true balance of the main bridge circuit can be obtained, provided that a certain relationship is maintained between the impedances of the auxiliary network and those of the bridge. These relationships are:

$$\frac{Z_A}{Z_B} = \frac{Z_N}{Z_P} = \frac{Z_S}{Z_T} \qquad (1)$$

or

$$\frac{Z_A}{Z_N} = \frac{Z_B}{Z_P} = \frac{Z_F}{Z_H} \qquad (2)$$

For either type of balance, four equations of balance must be satisfied, since all impedances involved are complex.

By connecting the third terminal of the unknown to the fifth, or guard, terminal the two unwanted impedances are made part of the auxiliary circuit and can be balanced out.

An excellent method of making the auxiliary balance[2] is to connect the generator and detector to the bridge in the usual way, and to provide a switch to connect the guard terminal to one of the corners of the bridge. This places either Z_F and Z_H or Z_S and Z_T across a pair of bridge arms. Now, the relations of Equations 1 or 2 can be satisfied by first balancing the bridge alone, and then balancing the bridge and auxiliary circuit in parallel by adjusting the auxiliary circuit. Using this method, either of the two auxiliary circuits can be balanced without changing the generator and detector connections, and with relatively simple switching arrangements.

Let us consider the arrangement of Figure 2 (d). As redrawn in Figure 5 the

[2]R. F. Field, "A Guard Circuit for Capacitance Bridge Measurements," General Radio *Experimenter*, Vol. XIV, No. 10, March, 1940.

FIGURE 3. Connections for a substitution measurement of C_X.

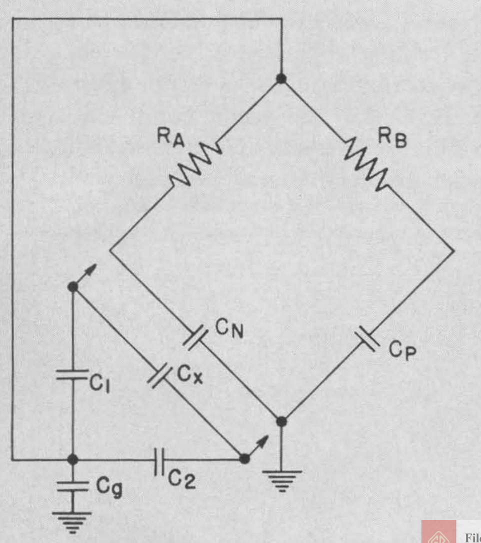
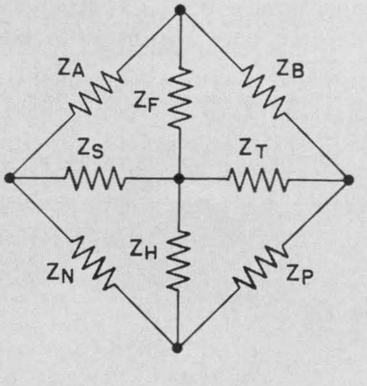


FIGURE 4. A five-terminal bridge network, showing an impedance between the fifth terminal and each corner of the bridge.



similarity to the generalized circuit of Figure 4 is apparent at once. Simply by providing an additional variable condenser across C_2 or C_g, it will be possible to balance partially the auxiliary circuit (R_A—R_B—C_2—C_g). As pointed out previously, both reactive and resistive balance of the auxiliary circuit must be provided for complete balance, to satisfy the expression $D_N + Q_A = D_P + Q_B = Q_F + Q_H$. It will frequently be adequate, however, to provide only for balancing the principal component of the auxiliary circuit[3]. Thus, in the simple case illustrated above, if C_2 and C_g are properly balanced, any unbalance of their resistive components will have only a negligible effect on the bridge balance, provided C_1 is small compared to C_N.

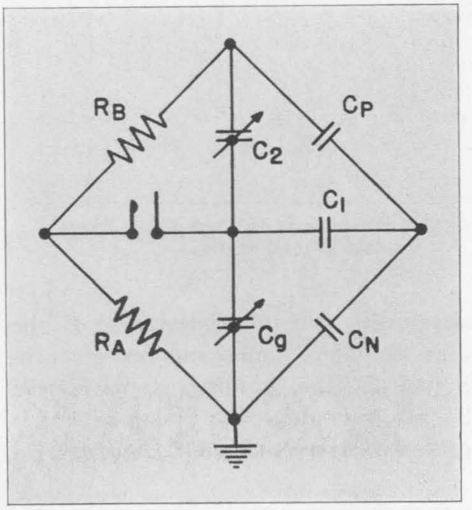


FIGURE 5. The arrangement of Figure 2 (d), redrawn to show its similarity to the generalized circuit of Figure 4.

The guard circuit may also be used to eliminate stray impedances associated with the bridge terminals and arms, as well as those associated with the unknown, by making ground the fifth or guard terminal. The earliest and perhaps the best known of this type of circuit is the Wagner Ground, used to remove from circuit the stray capacitances to ground of the generator terminals. As commonly used, balance is achieved by alternate adjustments of the bridge and the guard circuit, the detector being switched from one to the other. The detector may be left connected, however, and the balance made as suggested above, by adjusting the guard circuit with the bridge grounded at the junction of the ratio arms, and adjusting the bridge with this point ungrounded.

[3]Balsbaugh, Howell, and Dotson, "Generalized Bridge Network for Dielectric Measurement," Trans. AIEE, 1940, pps. 950-956, has shown that, if the guard and coupling circuits are not in true balance, the error introduced in the equation of balance of the bridge is proportional to the *product* of the unbalances of these auxiliary circuits.

A 500,000-OHM RHEOSTAT POTENTIOMETER

● **THE LATEST ADDITION TO OUR LINE** of rheostat potentiometers is a large-size, high-resistance unit, TYPE 433-A, having a total resistance of 500,000 ohms.

This wire-wound variable resistor is similar in construction to other General Radio rheostat-potentiometers. Figure 1

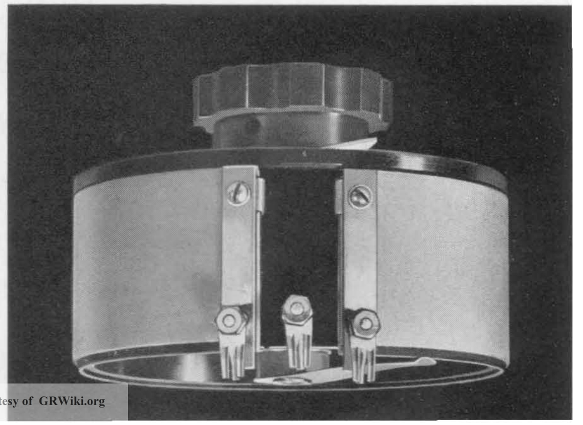


FIGURE 1. View of TYPE 433-A Rheostat Potentiometer.

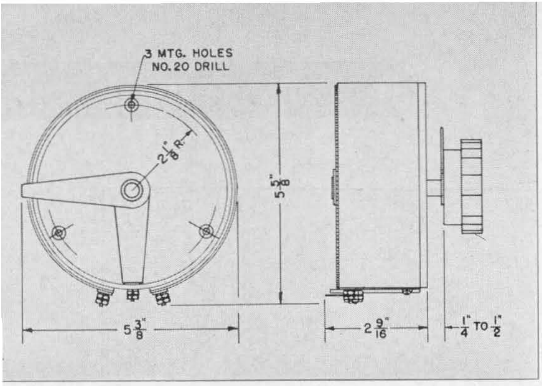


FIGURE 2. Dimensions of TYPE 433-A Rheostat Potentiometer.

shows the general appearance of the unit; complete dimensions are given in Figure 2. The winding is distributed linearly over an arc of 315 degrees and is protected from mechanical damage by a linen bakelite cover strip. Continuous contact with the winding is maintained by a phosphor-bronze blade mounted on a 3/8-inch bakelite shaft. The control knob is a TYPE 637-Q. Connections to the ends of the winding and to the contact arm are brought out to screw terminals with 3-fingered tinned soldering lugs.

TYPE 433-A Rheostat Potentiometer is available from stock in the 500,000-ohm size only. The power dissipation for the whole winding is 25 watts, for a temperature rise of 50° to 60° Centigrade. This corresponds to a maximum current of 7 ma. The net weight of the unit is 1 pound, 2 ounces.

Type	*Code Word*	*Price*
433-A	IMBUE	**$13.50**

MISCELLANY

● **THE CURRENT INSTALLMENT** of "Impedance Bridges Assembled from Laboratory Parts" concludes the series. Owing to the interest expressed in these articles, particularly by readers in engineering schools, we plan to reprint the entire series in a single booklet. Copies will be furnished free of charge to readers who request them and, if the demand warrants, they will be made available in reasonable quantities to teachers for student use.

● **THE SERVICE AND MAINTENANCE NOTES**, originally scheduled for mailing in September, 1941, have been unavoidably delayed. Publication is nearly completed, however, and we hope to mail them during January to those who have requested them.

GENERAL RADIO COMPANY

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THE General Radio EXPERIMENTER

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

THE MEASUREMENT AND ANALYSIS OF LINEAR AND TORSIONAL VIBRATIONS WITH ELECTRONIC INSTRUMENTS

Also

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A 100-Watt Output Power Meter 6

Rubber-Covered Cables 8

● **THE CURRENT DIFFICULTY** in obtaining instruments and equipment, irrespective of priority rating, has focused attention on the problem of extending the usefulness and application of instruments already on hand. One of the most fruitful possibilities in this direction is that of utilizing sound-measuring and analyzing equipment for the study of vibrational phenomena.

The study of the theoretical as well as the practical aspects of vibrations in mechanical systems has been greatly stimulated by the current emphasis on the mechanical side of warfare. The development of fast fighting planes, long-range bombers, high-speed torpedo boats, and armored vehicles depends, in no small measure, on the ability to measure and to control vibrations in the machinery and struc-

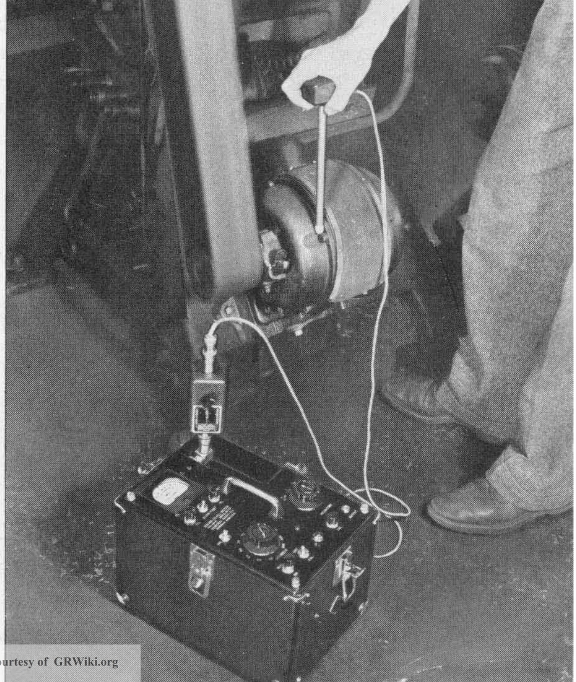


Figure 1. Measuring vibration in a punch-press motor with the vibration pickup and sound-level meter.

ture. The present article will show that, if suitable pickup mechanisms are available, the TYPE 759-B Sound-Level Meter and the TYPE 760-A Sound Analyzer (or the TYPE 736-A Wave Analyzer) can be used for satisfactory measurement and analysis of linear and torsional vibrations. One or more of these instruments will be found in most industrial research laboratories.

LINEAR VIBRATIONS

The TYPE 759-B Sound-Level Meter can be readily adapted for vibration measurements simply by replacing the microphone with a suitable vibration pickup. Since this instrument has a high input impedance, for use with a crystal microphone, a crystal pickup can be substituted directly.

The reading of the sound-level meter will be a measure of the acceleration[1] of the vibrating member to which the pickup is applied, because the crystal pickup, as commonly manufactured, is inertia-operated. More frequently, however, an indication of the *amplitude* of the vibration, independent of frequency, is desired, and occasionally a measure of vibration *velocity*. Fortunately it is a simple matter to convert the output voltage of the pickup to a voltage that is proportional to velocity or displacement. The voltage across the condenser in a series resistance-capacitance circuit is proportional to the integral of the impressed voltage at all frequencies for which the reactance of the condenser is small compared to the series resistance. Acceleration integrated once gives velocity, and the velocity integrated once gives displacement, so that a simple combination of resistors and condensers, together with a selector switch, can be used to provide any one of the three types of response.

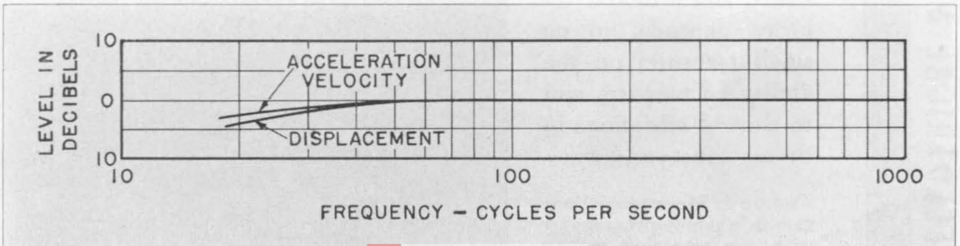
The TYPE 759-P35 Vibration Pickup and the TYPE 759-P36 Control Box, designed particularly for use with the TYPE 759-B Sound-Level Meter, have been available for some time.[2] The design of the control box is such that flat response is maintained to a frequency very close to the resonant frequency of the crystal pickup. Calibration figures are supplied, so that the decibel readings of the Sound-Level Meter can be converted to absolute values of vibration displacement, velocity, or acceleration. Figure 2 shows the over-all frequency response of a pickup, control box, sound-level meter combination. The low-frequency response is limited by the response of the amplifier in the sound-level meter, which, in its normal application, is not required to amplify frequencies below 20 or 30 cycles per second.[3]

[1]This is equivalent to saying that, for a constant amplitude, the response is proportional to the square of the frequency.

[2]Literature describing these units will be sent on request.

[3]The TYPE 761-A Vibration Meter, an instrument that is functionally identical to the combination described above, was announced in the June, 1941, *Experimenter*. Being specifically designed for vibration work, however, this instrument is direct reading in vibration displacement, velocity, and acceleration, and can be used at frequencies as low as 2 cycles.

FIGURE 2. Over-all frequency response of the vibration pickup, control box, and sound-level meter.



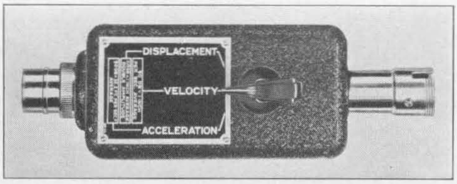


FIGURE 3. View of the TYPE 759-P36 Control Box. Nominal calibration data are given on the nameplate.

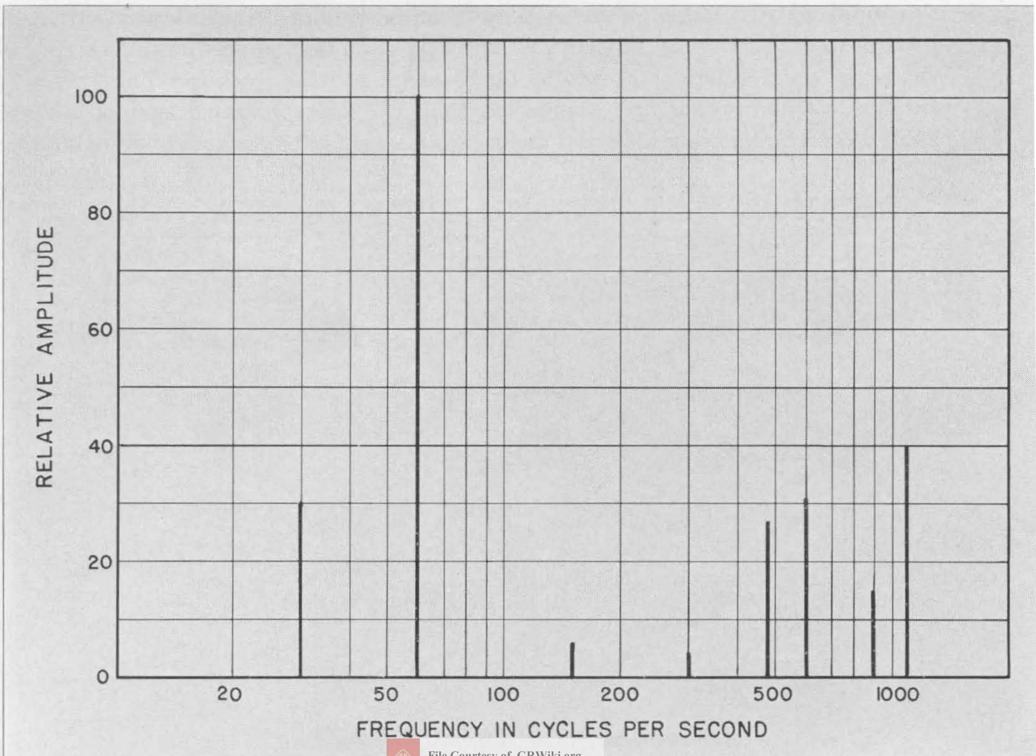
ANALYSIS

In many applications a knowledge of the frequency distribution and the relative amplitudes of the various components of the vibrational disturbance is important. If the TYPE 760-A Sound Analyzer is used at the output of the sound-level meter, the entire vibration spectrum can be scanned in a few minutes, and the frequency and magnitude of the significant components noted. In Figure 4 is shown an analysis, made with the equipment described above, of the vibrations occurring in a large surface grinder.

The TYPE 736-A Wave Analyzer may also be used to evaluate the frequency components of the output voltage of the sound-level meter. The selectivity of this analyzer,[4] however, is so high that unless the vibration is very stable, it may be difficult to obtain satisfactory readings at the higher frequencies. Furthermore, at the lower frequencies, the band width of 4 cycles may be too broad to obtain a satisfactory separation of non-harmonic components that may fall very close together in frequency. In addition, the instrument is somewhat bulky as compared to the TYPE

[4]The band width is approximately 4 cycles, and is independent of frequency. In contrast to this, the TYPE 760 has a constant fractional band width, giving the same effective separation and tuning stability at all frequencies.

FIGURE 4. Frequency analysis of the vibrations occurring at the exhaust horn of a vertical grinding machine. The large component at 60 cycles corresponds to the second harmonic of the rotational speed of the driving motor. The 1100-cycle component is caused by a resonance in the motor housing.



760 and requires preliminary adjustments that are relatively critical compared to the simplicity of operation of the Type 760.

TORSIONAL VIBRATION

Of utmost importance in the design and testing of cam shafts, drive shafts, propellers, and other rotating mechanisms that transmit or deliver power is a knowledge of the frequency and amplitude of the torsional vibrations that exist. Only when a means for measuring such vibrations is available can an intelligent program be directed toward their reduction or elimination.

Perhaps the most widely accepted method of analyzing torsional vibrations is to make an oscillogram of the vibration, using a torsional pickup, integrating amplifier, and recording oscillograph.[5] A mechanical harmonic analyzer is then used to obtain the magnitudes of the different components of the complex wave. Although this method is admittedly the most thorough (it yields the phase relations as well as the amplitudes), its use has been somewhat limited by the relatively high cost of the equipment involved and by the fact that considerable time is consumed in developing and analyzing the records. For production or maintenance testing, in particular, the latter objection can become sufficiently serious to rule out the method.

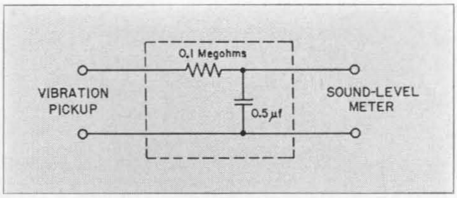


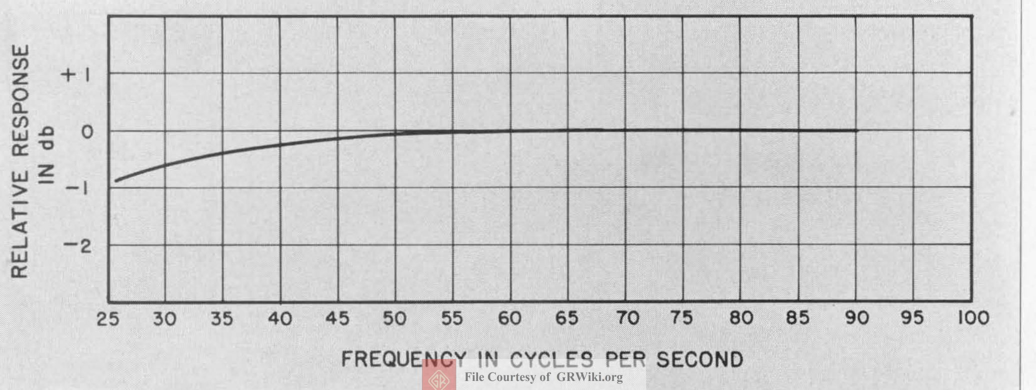
Figure 5. A simple integrating network that can be used to convert the output of the No. 89246 torsional pickup to a voltage proportional to angular displacement.

Owing to the speed and convenience with which measurements can be made, the combination of sound-level meter and analyzer offers an attractive possibility for use in conjunction with a torsional pickup. Tests made with a Sperry[6] pickup (No. 89246) indicate that entirely satisfactory performance can be obtained from a lower limit of 25 cycles to an upper limit determined solely by the pickup. This type of pickup, when mounted against a shaft, is insensitive to the steady-state angular

[5] The sound analyzer has been found extremely useful with this equipment, since it is often desirable to know, while the tests are being run, what the frequencies and relative amplitudes are.

[6] Vibration pickups, amplifiers, and recorders formerly manufactured by the Sperry Gyroscope Company are now made and sold by Consolidated Engineering Corp., 1255 East Green Street, Pasadena, California.

Figure 6. Over-all response of the torsional pickup, the integrating network of Figure 5, and the Type 759-B Sound-Level Meter. Data were not taken above 90 cycles because of the limitations of the mechanical calibrator used, but an essentially flat response should be maintained to about 1000 cycles, the upper frequency limit of the pickup.



motion, but produces a voltage proportional to the angular velocity of any torsional vibration that is superimposed. To convert this voltage to a voltage proportional to amplitude of angular displacement, it is only necessary to integrate once, using a series resistance-capacitance circuit. A satisfactory integrating network can be made with a 0.1-megohm resistor and a 0.5 μf condenser, as shown in Figure 5. The response of the sound-level meter with the network of Figure 5 interposed between pickup and amplifier is shown in Figure 6. A constant calibration amplitude of 2.02 degrees was maintained as the frequency of vibration was varied from 25 cycles to 90 cycles.[7] The minimum measurable amplitude is limited by the sensitivity of the pickup to about 0.1° at a frequency of 60 cycles, while the maximum amplitude that can be measured is essentially limited by the maximum angular displacement that the pickup can safely accommodate. The response of the pickup is linear up to 10° double amplitude.

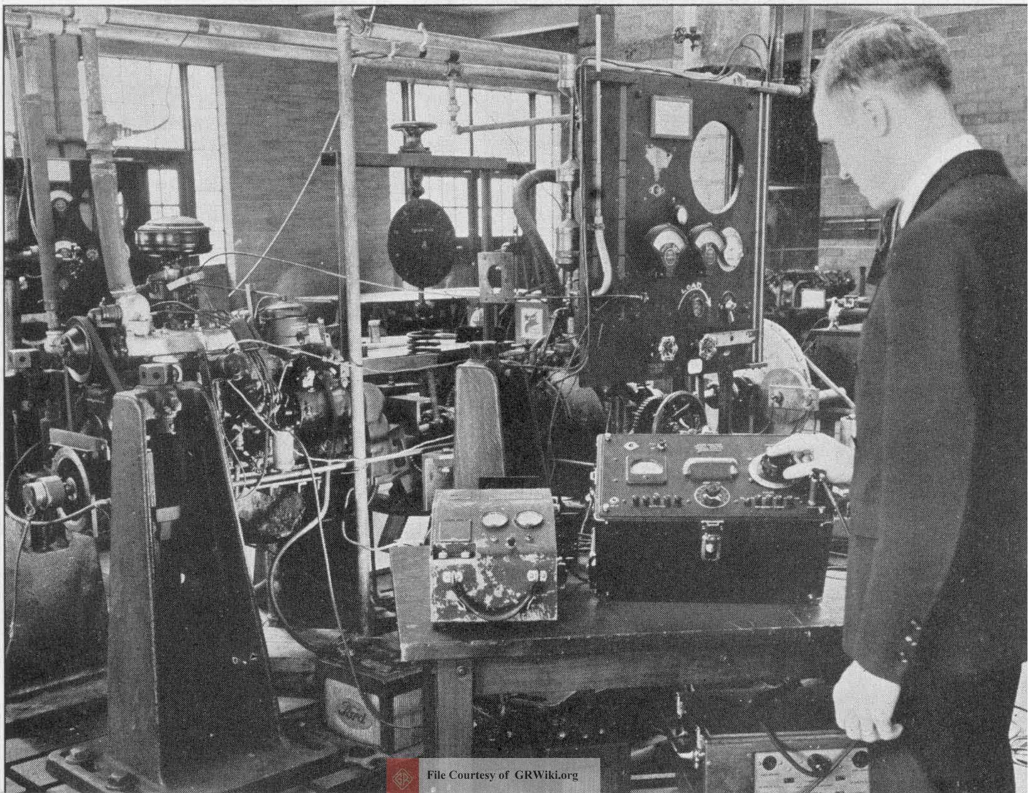
[7]This test was made in the laboratories of the Ranger Engineering Corporation, Farmingdale, Long Island, New York.

Here again, the use of the TYPE 736-A Wave Analyzer (directly across the output of the integrating network) is feasible but, as previously explained, is not generally as satisfactory as the scheme outlined above.

While of course it cannot be pretended that the various arrangements discussed above can completely replace equipment specifically designed for vibration measurement and analysis, experience has shown that entirely satisfactory results can be had in the study of vibrations at frequencies above 20 or 25 cycles.

—IVAN G. EASTON

FIGURE 7. The TYPE 760-A Sound Analyzer set up for analyzing torsional vibration. The pickup is shown on the end of the crankshaft at the extreme left. The sound-level meter is not shown in this photograph.

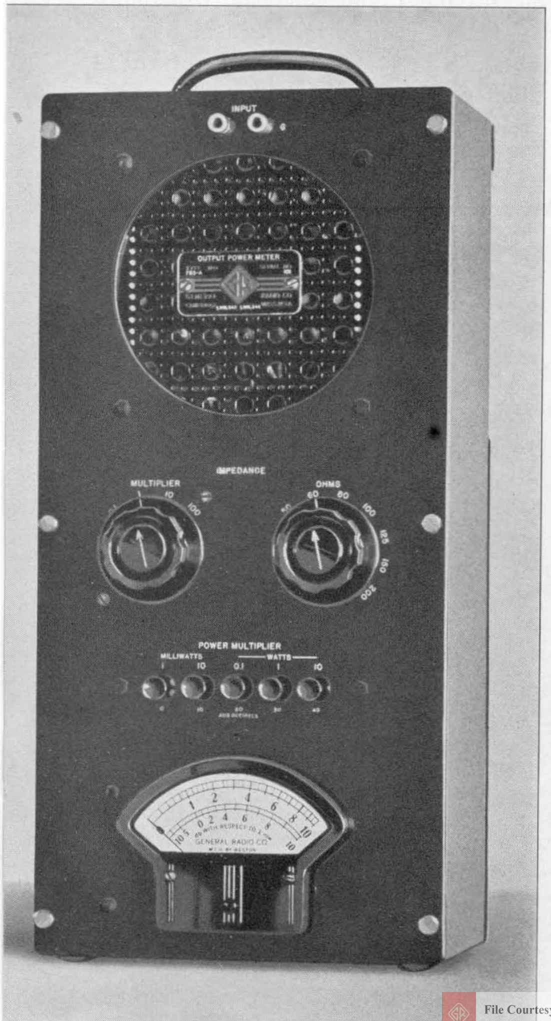


A 100-WATT OUTPUT POWER METER

● **THE OUTPUT POWER METER** for power-output and internal-impedance measurements on radio receivers, amplifiers, and oscillators was first introduced by General Radio nearly ten years ago.* Over a thousand of these instruments have been sold and, with the development of the art, their general utility around the communications laboratory is constantly increasing.

*"A Power Meter with a Wide Frequency Range," *Experimenter*, May, 1932. "A Direct-Reading Meter for Power and Impedance Measurements," *Experimenter*, November, 1932.

FIGURE 1. Panel view of the TYPE 783-A Output Power Meter.



It has been evident recently that there exists a field for an instrument of the same type but capable of dissipating greater amounts of power, and the new TYPE 783-A Output Power Meter has been designed to meet this need.

Nearly as sensitive at low power levels as the older TYPE 583, this new instrument has a much wider power range extending to a maximum of 100 watts. The power scale on the indicating meter extends from 0 to 10, and is used in conjunction with a set of five push-button-operated decade multipliers. An auxiliary decibel scale is provided on the meter, extending from −10 db to +10 db, referred to a level of 1 milliwatt.

The impedance range is 2.5 ohms to 20,000 ohms, covered by means of two switches, one direct reading in ohms, the other a multiplier.

The accuracies of both power and impedance indications are maintained over a considerably wider frequency range than in the TYPE 583.

A functional schematic diagram of the TYPE 783-A Output Power Meter is given in Figure 2. As can be seen from this diagram, the instrument is equivalent to an adjustable load impedance, across which is connected a voltmeter calibrated directly in watts dissipated in the load. It consists essentially of a voltage divider and an autotransformer for adjusting the impedance level, and a set of resistive pads for adjusting attenuation.

The operation of the output power meter is extremely simple. For measuring the power that a circuit is capable of delivering into a given impedance, the impedance switch and multiplier are set to the desired value, and the power is then indicated by the meter and its multiplier. The internal impedance of the source under test can also be determined since it is equal to the impedance into which maximum power is delivered.

The output power meter is extremely useful in experimental work where a number of power and impedance measurements must be made as the characteristics of the circuit under measurement are varied. It is a valuable aid in the design and testing of amplifiers, oscillators, filters, transformers, and other networks, in making standard tests on radio receivers, and in measuring the power output of vacuum tubes. Its impedance range is wide enough to simulate all types of loudspeakers, and its sensitivity is sufficient to measure directly the output and internal impedance of a magnetic phonograph pickup.

Another use is in the measurement of the loss in a transformer working out of a given source impedance. The maximum output of the source is determined, after which the transformer is interposed between the source and the meter, and the maximum output of the transformer is found. The difference between the two readings on the decibel scale gives the transformer loss directly.

SPECIFICATIONS

Power Range: 0.2 milliwatt to 100 watts in five ranges (10 and 100 milliwatts, 1, 10, and 100 watts, full scale). An auxiliary decibel scale reads from −10 to +50 db referred to a level of 1 milliwatt.

Impedance Range: 2.5 to 20,000 ohms. Forty discrete impedances, distributed approximately logarithmically, are obtained by means of a ten-step OHMS dial and a four-step MULTIPLIER.

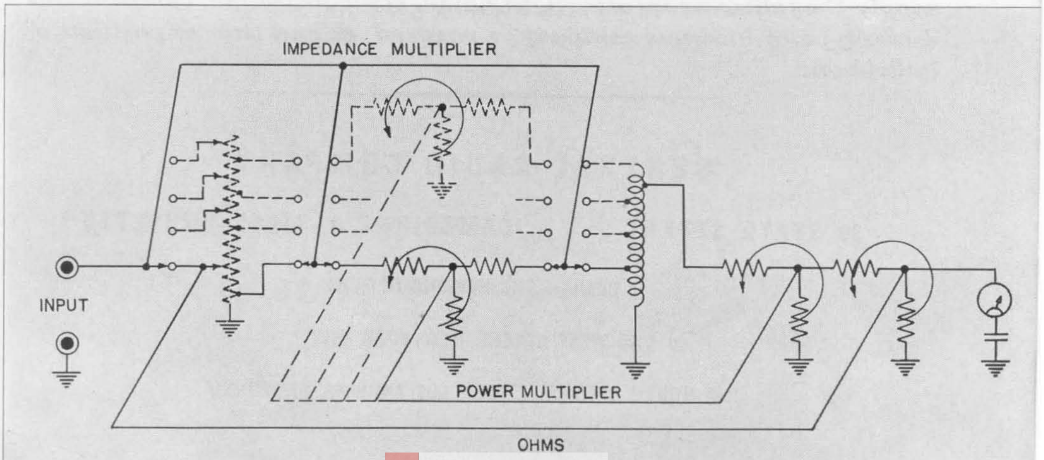
Impedance Accuracy: The input impedance is within ±2% of the indicated value, except at the higher audio frequencies, where the error for the higher impedance settings may exceed this value. At 15,000 cycles the input impedance error is about 5% for impedances from 10,000 to 20,000 ohms.

Power Accuracy: The indicated power is accurate to ±0.25 db at full-scale reading. At the lowest impedance multiplier setting (2.5 to 20 ohms) there may be an additional error of 0.2 db due to switch contact resistance when the power multiplier is set at 10 (10 to 100 watt range).

The over-all frequency characteristic of the power indication is flat within ±0.5 db from 20 cycles to 10,000 cycles; within ±0.75 db to 15,000 cycles.

Waveform Error: The indicating instrument used is a copper-oxide rectifier meter, calibrated in r-m-s values for a sinusoidal applied voltage. When non-sinusoidal voltages are applied an error in indication may occur, since the meter is not a true r-m-s indicating device. The error

FIGURE 2. Schematic circuit diagram of the TYPE 783-A Output Power Meter.



will depend on the magnitude and phase of the harmonics present, but, with waveforms normally encountered in measurement circuits at communications frequencies, will not be serious.

Temperature and Humidity Effects: Humidity conditions have a negligible effect on the accuracy of the instrument.

The instrument is calibrated at 77° Fahrenheit, and if the ambient temperature departs widely from this value, additional errors of indication may be expected. At high temperatures (95° Fahrenheit) this additional error may approach the nominal calibration error, particularly at the higher frequencies.

The heat dissipated by the instrument itself has no effect on the accuracy.

Accessories Supplied: One Type 274-M Plug.

Mounting: The instrument is mounted on a bakelite panel in a walnut cabinet.

Dimensions: 8 x 18 x 7 inches, over-all.

Net Weight: 17 pounds.

Type		*Code Word*	*Price*
783-A	Output-Power Meter	ABBEY	$185.00

This instrument is manufactured and sold under United States Patents Nos. 1,901,343 and 1,901,344.

RUBBER-COVERED CABLES

● **FOR THE PRESENT**, at least, we have sufficient rubber-covered power-supply cables and concentric-shielded cables on hand to supply with new equipment. We cannot yet estimate how long our supply will last, but as long as it does we will continue to furnish them. In the meantime we are searching for adequate substitutes.

Because of the now very limited supply, we are sorry that we will not be able to furnish rubber-covered cables either as spares with new equipment or for replacements. Although we believe that the cables we now supply as standard accessories are the best available, there are adequate substitutes for the power cables. Users are urged to conserve and repair broken concentric conductors, however, because substitutes for these, not employing rubber, are much more difficult to find.

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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VOLUME XVI No. 10 MARCH, 1942

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

HOW GOOD IS AN IRON-CORED COIL?

● **WHEN MR. ARGUIMBAU WROTE HIS ARTICLE** for the November, 1936, issue of the General Radio *Experimenter* on "Losses in Audio-Frequency Coils," he approached the subject from a refreshingly new point of view, and gave a number of very useful new concepts to help understand the behavior of such coils. An example is the single template which can be used to draw the curve, on log-log paper, of storage factor Q against applied frequency for *any* coil, provided only that the maximum Q of the coil and the frequency at which it occurs are known, and that resonance is remote. Alternatively, the template can be used to draw a smooth curve through a number of experimentally determined points. One such template is shown on page 2.

Employment of a useful tool like this soon makes it part of one's mental equipment. Then comes a desire to extend its usefulness by making it help answer a wider field of questions. Predictions would be very useful indicating the maximum Q of a coil and the frequency at which it occurs as changes are made in the characteristics of the iron core. For instance, how much would doubling the stack height of the iron increase Q_{max}, or what would decreasing the thickness of the laminations do to Q_{max} and f_{max}? Many other similar questions will occur.

If a theory can be provided to answer such questions, the behavior of almost any projected construction can be extrapolated from empirically obtained information. This need be less extensive than might be expected. Experimental values of Q_{max} and f_{max} for several different air gaps in an average-sized lamination would suffice, although more data would be preferable in that they would permit intercomparisons.

The extrapolated results, which must be based on certain assumptions, are necessarily approximate but still are accurate enough for most design calculations. Knowledge of the exact value of Q is rarely necessary, order of magnitude generally being sufficient.

CONDITIONS AND ASSUMPTIONS

These conditions and assumptions are as follows:

(1) Measurements are to be made at such a level that the iron has its initial permeability, that is, that the flux density, B, is vanishingly small.

(2) Under condition (1) the hysteresis loss in the iron core-material vanishes. This point is taken up in more detail on pages 11 and 12.

(3) Skin effect of the copper wire of which the coil is wound is negligible. This assumption is justified since skin effect at audio frequencies is encountered only in rather large copper wires, larger than one would be likely to employ in winding coils for use at those frequencies, such as coils for wave filters.

(4) There is negligible leakage flux traversing the copper winding. This means that eddy-current losses in the copper can be neglected. It also postulates uniform B throughout the whole magnetic path. When air gap becomes large, leakage flux is no longer negligible. Then eddy-current losses increase and the effect of the air gap on μ and on f_m cannot be calculated from simple theory.

(5) There are negligible eddy currents between adjacent laminations.

(6) Resonance is remote.

What, then, is the good of results applicable only when B is almost zero? Of course, many iron-cored coils are power transformers, operating at 10 to 12 kilogausses, some regulating types even working purposely in the saturation region at still higher flux densities. But in the communications field there are many applications for transformers and reactors operating at exceedingly low levels — for example, microphone or interstage transformers, low-level wave filters. In fact, low level is often a hindrance in the design of a transformer, because the iron permeability is so low that more turns are needed to provide the requisite minimum inductance. Even audio transformers in higher-level stages must be designed to have adequate inductance at initial permeability to prevent distortion when the audio signal drops to a very low value, such as during pianissimo orchestral passages. Furthermore, even though this analysis can be used directly only at very small B, once it is thoroughly understood it is not difficult to make estimates of the modifications required at higher levels, where the effect of hysteresis is no longer negligible. Increasing hysteresis losses decrease Q_m and mask out somewhat the contributions to Q_m, as frequency varies, of ohmic and eddy-current losses. Frequency f_m is unaffected. [See discussion accompanying Equation (20) and Figure 3 in Appendix.] Although the Q–f curve has a flatter top at higher B's, the remote wings are the same and the maximum occurs at the same frequency.

Template
for theoretical
relation
Q vs. frequency
(correct for log-log paper, $2\frac{1}{2}''$ per decade,
K & E, No. 359-120).
A cardboard template will be sent
upon request.

EXPRESSIONS FOR Q_m AND f_m

If the listed conditions are met, the following expressions, taken from a detailed derivation in the Appendix, give the maximum Q of the coil and the frequency at which it occurs:

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

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

(For meanings of the symbols, consult Glossary near beginning of Appendix.) These properties (Q_m and f_m) are given in terms of dimensions of the lamination, resistivities of the copper and iron, and the permeability of the core material. Rewriting Equations (22) and (23) as follows will show more clearly the nature of the separate contributing factors:

$$\begin{array}{ccccccccccc}
Q_m = & \sqrt{3} & \cdot & \sqrt{\dfrac{A\alpha}{\delta^2 l}} & \cdot & \sqrt{\rho_i} & \cdot & \sqrt{\dfrac{S}{t}} & \cdot & \sqrt{\dfrac{1}{\rho_c}} & (22a) \\
 & \uparrow & & \uparrow & & \uparrow & & \uparrow & & \uparrow & \\
 & \text{numeric} & & \text{geometry of core} & & \text{material of core} & & \text{geometry of coil} & & \text{material of coil} & \\
 & \downarrow & & \downarrow & & \downarrow & & \downarrow & & \downarrow & \\
f_m = & \dfrac{10^9\sqrt{3}}{4\pi^2} & \cdot & \sqrt{\dfrac{l}{\delta^2 A\alpha}} & \cdot & \sqrt{\dfrac{\rho_i}{\mu^2}} & \cdot & \sqrt{\dfrac{t}{S}} & \cdot & \sqrt{\rho_c} & (23a)
\end{array}$$

All of the factors determining Q_m and f_m are purely physical properties of the core and coil structure, with the single exception of the factor S, which is the effective copper winding area, and which in a sense is a derived property of the core structure.

It must be clearly understood that the permeability appearing in the formulae is the effective permeability of the path in the structure employed, which in general must be less than that of the iron obtained with ring samples. Equation (3) or (3a) of the Appendix gives an expression relating effective and true incremental permeabilities. However, there are many uncertainties in its employment. The effective length of the gap is almost never the same as the measured gap, for a variety of reasons. It is usually greater, but in the case of very large gaps may be less because of the effects of fringing. It is not completely satisfactory to regard, as some have suggested, every gap as being effectively longer than it really is by a fixed length equal to the equivalent length of a butt joint. A further complication arises from the fact that a gap in the iron leg inside the coil has more effect than one of the same length in a leg (or legs) outside the coil.

It is better, therefore, to use the empirical approach in getting the original data, the springboard from which to jump. The Q_m and f_m should be obtained for at least one core structure at a number of air gaps covering the range from complete interleaving (no gaps) to the largest practical gap. Interpolation between experimentally derived points can be done directly on the log-log plot, like Figure 4 of Mr. Arguimbau's paper, or, better still, by using an auxiliary curve (on the same sheet, if desired) of length of air gap, g, against f_m. Three such curves are the inclined, dashed ones on Figure 1. The Q_m for any one structure will be sensibly independent of the air gap. It will be found safe to predict μ on the basis of gap-length ratio (g/l); that is, if one structure has twice the l of another and twice the g, the μ will be the same for purposes of Equation (23).

FIRST DEDUCTIONS

What are the most obvious facts to be gleaned from these two expressions, Equations (22) and (23)?

1. If cores and coils are considered having similar proportions but different sizes (every dimension altered by the same factor), f_m is inversely proportional and Q_m is directly proportional to any homologous dimension. This means, for instance, that for a 1½″-tongue lamination f_m would be ½ as great and Q_m twice as great as for a ¾″-tongue lamination (lamination thickness being unchanged).

2. f_m and Q_m are inverse with δ, the thickness of laminations.

3. f_m is inverse with μ, which is an effective μ that takes into account the effect of any air gaps in the magnetic circuit.

4. Q_m is independent of μ and hence also of air gap.

5. f_m is independent of hysteresis loss, hence of B, the flux density.

SPECIFIC EFFECTS

Now, suppose these general observations be applied to specific problems which might be encountered in practice. What will happen, for instance, if:

1. The whole structure is changed in size but not in shape, each homologous dimension being multiplied by a factor r? Q_m increases and f_m decreases by this factor r.

2. Lamination thickness is diminished? First, δ is smaller. Also, α becomes smaller because a smaller effective amount of iron can be assembled into the coil. This is because: (a) the scale makes up a bigger proportion of the core; and (b) it is impossible to pack in the iron so tightly, since it gets too flimsy to withstand such heavy pushing forces. Since α decreases, Q_m does not increase quite inversely with δ and f_m increases slightly faster than inversely with δ, the disparity being a factor $\sqrt{\alpha}$.

3. The window is not filled with copper (S below normal)? Q_m decreases and f_m increases as the square root of the copper factor decreases, unless t also changes. (Example: What are Q_m and f_m of a transformer primary, or secondary, only?)

4. Similarly, the core is not filled with iron (α of iron less than normal)? As in 3 just above, Q_m decreases and f_m increases as the square root of the stacking factor of the magnetic material decreases.

5. The coil is wound for a higher stack of iron, the laminations having the same contour? A would increase, making Q_m increase and f_m decrease proportionally to the square root of the stack height of the iron were it not that t is increased simultaneously. This partially reduces the effect of the higher stack so that the changes in Q_m and f_m are less than proportional to the square root of the stack height. For example, let us consider, for lamination proportions usually encountered, that the stack height is changed from once to twice the width of the center leg of the lamination. In this case it has been found that Q_m and f_m change by a factor of approximately 1.25, instead of 1.41 (the square root of 2, the stack-height factor).

6. ρ_i is decreased, say by substituting A-metal for silicon-steel laminations? Q_m and f_m would decrease with the square root of ρ_i. However, in this particular case f_m would decrease still further, because the initial μ of A-metal is so much larger than that of silicon steel.

7. ρ_c is increased, say by winding the coil with resistance wire? Q_m would decrease and f_m would increase with the square root of ρ_c.

8. One or more air gaps are inserted in the magnetic circuit? Q_m would be unchanged, f_m would vary inversely with

the effective μ of the magnetic circuit (or, expressed differently, with the inductance L of any particular coil).

9. Combinations of the above changes are made? The net result will be an alteration which is measured by the product of the alterations produced by each of the individual changes.

EXPERIMENTAL CONFIRMATION

How well do the experimental facts bear out this theoretical analysis? This will be shown by three examples of varying complexity selected from the information on a chart herewith, Figure 1, similar to the one on page 4 of Mr. Arguimbau's article but containing a great deal more information subsequently obtained. In each example two different cases will be compared. The data will be presented in columnar form for greater ease in comparison. Where a ratio is used, it is expressed as the ratio of the second case to the first.

1. This is an example where coils are compared, wound on square cores using two standard General Radio laminations, of the same thickness (0.0188″), of approximately the same proportions, but of different sizes. Each magnetic circuit contains an air gap in the center leg only, proportional to the length of the magnetic circuit in each instance, which should keep the effective μ of the circuit the same.

Case	*I*	*II*
GR Type	345	485
Q_m	39	48
f_m	310 c	250 c
Air Gap	0.010″	0.0133″
Width of Center Leg	¾″	15/16″

The ratio of the Q_m's would be by theory the ratio of two homologous dimensions, or $\frac{15/16''}{3/4''} =$ 1.25. Compare the measured ratio: $\frac{48}{39} = 1.23$.

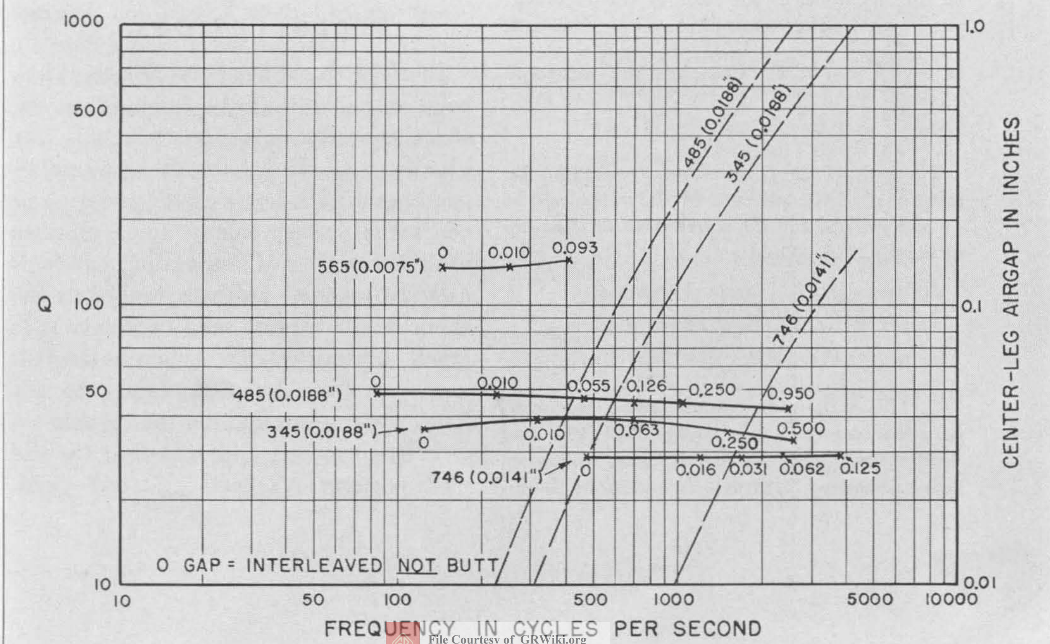
Similarly, the ratio of the f_m's by theory would be inverse with the homologous dimen-

FIGURE 1. Plots of Q and g vs. f.

Horizontal solid curves show locus of Q_{max} as a function of f_{max}. Curves are labeled by GR Type number of lamination, and by thickness of each lamination in parenthesis.

Inclined, dashed curves show locus of f_{max} (frequency at which Q is maximum) as a function of center-leg air gap.

Laminations progress from small to large in this order: 746, 345, 485, 565. The 565 curve is so far separated from the others not so much because of larger size as because of thinner laminations.



sions, which would be $\frac{3/4''}{15/16''} = 0.80$. The measured ratio is $\frac{250}{310} = 0.81$.

2. This example (taken from data *not* shown in Figure 1) compares Q_m and f_m of two identical coils on our TYPE 485 Core having different thicknesses of completely interleaved laminations (closest possible approach to zero air gap).

Case	*III*	*IV*
Q_m	39	93
f_m	110 c	410 c
δ	0.0192″	0.0075″
L	6.35 h	3.7 h
Weight of Iron	23.5 oz.	18.5 oz.

δ is changed by a ratio of 0.39.

α is changed by the ratio of the weights of iron, 0.787. This neglects the effect of scale on the iron, which at a thickness of 0.0075″ has not yet become an appreciable fraction of the total thickness.

However, a still further factor must be considered; μ is lower for the thin iron. It will be noted that L is changed by a ratio of 0.583, more than can be accounted for by the change in α. The difference, or a ratio of $\frac{0.583}{0.787} = 0.741$, can be ascribed to the decrease in μ. It is known that permeability does decrease for the thin, heavily-worked gauges of silicon steel.

The ratio of the Q_m's by theory would then be the reciprocal of the ratio of the δ's multiplied by the square root of the ratio of the α's, or $\frac{\sqrt{0.787}}{0.39} = 2.27$. Compare the measured ratio: $\frac{93}{39} = 2.38$.

The ratio of the f_m's by theory would be the reciprocal of the product of the ratios of the δ's and the μ's, and the square root of the ratio of the α's. This would be

$$\frac{1}{0.392 \times 0.741 \sqrt{0.787}} = 3.87.$$

The measured ratio is $\frac{410}{110} = 3.73$.

3. This is the most complicated comparison, between two coils, one using GR TYPE 345 Laminations and the other using the very small Allegheny TYPE F12 Laminations. Lamination thicknesses and stack heights vary as well as the dimensions of the laminations themselves.

Case	*V*	*VI*
Type	GR-345	F12
Q_m	38	29
f_m	980 c	2350 c
δ	0.0188″	0.0141″
Air Gap	0.111″	0.062″
Width of Center Leg	3/4″	11/32″
Stack Height	3/4″	23/32″

The ratio of the δ's is $\frac{0.0141}{0.0188} = 0.75$.

The ratio of homologous sides equals

$$\frac{11/32}{3/4} = 0.46.$$

In addition to these factors, others are necessitated by the extra stack height of the F12 lamination; ratio of A to that of a square-center-leg stack equals $\frac{23/32}{11/32} = 2.09$. The ratio of t's $= 1.34$.

The ratio of the Q_m's by theory equals $\frac{1}{0.75}(0.46)\sqrt{\frac{2.09}{1.34}} = 0.765$. Compare the measured ratio: $\frac{29}{38} = 0.764$.

The ratio of f_m's by theory equals

$$\frac{1}{0.75} \times \frac{1}{0.46} \sqrt{\frac{1.34}{2.09}} = 2.32.$$

The measured ratio is $\frac{2350}{980} = 2.40$.

CONCLUSIONS

In all of the above comparisons it has been assumed that the laminations are strictly similar in shape, which is not exactly true. However, the reasonably good agreement between the theory and the actual measurements for a number of different sizes of laminations can logically be taken to indicate, first, that the theory is adequate and, second, that small departures of the lamination dimensions from strict similarity do not have any major effect on the results.

It is, therefore, apparent that the use of Equations (22) and (23) will yield, with satisfactory approximation, a good picture of the behavior of a particular proposed coil structure, provided there is a small amount of reliable information on which to base the predictions.

HOW TO USE

To discover and put to use any extrapolated information such as has been described, proceed as follows:

1. List all of the properties and dimensions which differ for the two cases to be compared. Values for l will be needed if gap lengths g are to be altered to keep μ unchanged.

2. Calculate Q_m and f_m for the new structure from the known corresponding values of the old structure and the information in 1.

3. Plot the point corresponding to the new Q_m and f_m on the log-log paper. Lay the template on the paper with the long straight side parallel to the f-axis and with the (marked) center of the hump of the curve at the point just plotted. The behavior of *any* coil wound on this structure over a wide range of frequencies will be shown by the template [barring, of course, skin effect (very large wires), resonance (very high inductance), or other anomalous circumstance]. The curve may be actually drawn using the template, or, if it would cause confusion, on a sheet bearing a great deal of information, values could be read directly from the edge of the template.

APPENDIX

Losses in an iron-cored coil come about from four sources: namely, I^2R loss and eddy-current loss in the copper, hysteresis and eddy-current loss in the iron. Eddy-current losses in the copper will be ignored in this analysis for two reasons. The first is that the audio frequencies considered will be too low and/or the wire sizes too small to have appreciable eddy-current loss. The second, and more important reason, is that the iron core effectually prevents most of the flux from traversing the window in which the copper of the coil is located.

GLOSSARY

The symbols used are tabulated next, with their definitions and dimensions.

E = r-m-s alternating emf across coil; volts $(= I\omega L)$.

I = r-m-s alternating current through coil; amperes.

L = inductance of coil; henrys.

f = frequency of alternating voltage and current.

$\omega = 2\pi f$.

$\mathfrak{F}$ = r-m-s magnetomotive force; gilberts.

H = r-m-s magnetic force produced by current I; oersteds.

B = r-m-s flux density within the iron; gausses.

B_m = max. instantaneous value of alternating flux density $= B\sqrt{2}$; gausses.

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

$\mathcal{R}$ = reluctance of magnetic path.

P_h, P_e = power dissipated in the iron by hysteresis and eddy currents, respectively; watts.

R_h, R_e = equivalent resistances corresponding to P_h and P_e; ohms.

R_c = ohmic resistance of the copper in the coil; ohms. (Assumed the same as the d-c value; that is, no skin effect.)

D_c, D_h and D_e = dissipation factors corresponding to R_c, R_h, and R_e, obtained by relating each equivalent resistance to the coil reactance ωL; dimensionless.

D = total dissipation factor = sum of D_c, D_h, and D_e.

ρ_c = resistivity of copper; ohm-cm.

d = wire diameter (exclusive of insulation); cm.

T = total length of copper wire; cm.

t = length of average turn; cm.

N = number of turns of wire.

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

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

δ = lamination thickness; cm.

A = total geometric cross section of magnetic path; cm^2.

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

l = mean length of flux path; cm.

V = volume of magnetic material = $lA\alpha$; cm^3.

g = total length of air gaps; cm.

μ = incremental permeability (effective) of magnetic circuit.

μ_t = incremental permeability (true, ring-sample) of magnetic material.

η = hysteresis constant.

ϵ = Steinmetz exponent.

DERIVATION

Most of the basic equations given below can be found in any textbook or handbook of electricity. The first six define $\mathfrak{F}$, $\mathfrak{R}$, μ, Φ and L in terms of coil parameters and current through the coil:

$$\mathfrak{F} = \frac{4\pi NI}{10} \tag{1}$$

$$\mathfrak{R} = \frac{l-g}{\mu_t A\alpha} + \frac{g}{A} = \frac{1}{A\alpha}\left(\frac{l-g}{\mu_t} + g\alpha\right) = \frac{l}{\mu A\alpha} \tag{2}$$

where

$$\mu = \frac{\mu_t}{1 + \dfrac{g}{l}(\mu_t\alpha - 1)} \tag{3}$$

and, since usually $\mu_t\alpha \gg 1$, approximately

$$\mu = \frac{\mu_t}{1 + \dfrac{g}{l}\mu_t\alpha} \tag{3a}$$

$$\Phi = \frac{\mathfrak{F}}{\mathfrak{R}} = \frac{4\pi NI\mu A\alpha}{10l} \tag{4}$$

$$\Phi = BA\alpha \tag{5}$$

Also, using Equation (4):

$$L = \frac{\Phi N}{10^8 I} = \frac{4\pi N^2\mu A\alpha}{10^9 l} \tag{6}$$

COPPER LOSS

The series ohmic resistance of the coil is given, from resistivity, by:

$$R_{c(ser)} = \rho_c \frac{T}{\dfrac{\pi}{4}d^2} = \frac{4\rho_c Nt}{\pi d^2} \tag{7}$$

The dissipation factor corresponding to this can be reduced by the use of Equation (6):

$$D_c = \frac{R_{c(ser)}}{\omega L} = \frac{4\rho_c Nt}{\pi d^2} \cdot \frac{10^9 l}{2\pi f 4\pi N^2 \mu A\alpha}$$
$$= \frac{10^9 l\rho_c t}{8\pi^2 f N d^2 \mu A\alpha} = \frac{10^9 \rho_c tl}{8\pi^2 f\mu S A\alpha} = \frac{c}{f} \quad (8)$$

This dissipation factor is found to be inversely proportional to frequency, the factor of proportionality being:

$$c = \frac{10^9 \rho_c tl}{8\pi^2 \mu S A\alpha} \quad (9)$$

HYSTERESIS LOSS

The power expended in hysteresis loss is given by (since $B_m = B\sqrt{2}$):

$$P_h = \eta V f B_m^{\epsilon} 10^{-7} = \eta V f 2^{\epsilon/2} B^{\epsilon} 10^{-7} \quad (10)$$

Since power equals $\frac{E^2}{R}$, the equivalent parallel resistance of the hysteresis loss is, using also Equations (5) and (6):

$$R_{h(par)} = \frac{E^2}{P_h} = \frac{I^2\omega^2 L^2}{P_h}$$
$$= \frac{4\pi^2 N^2 B^{(2-\epsilon)} f A\alpha}{2^{\epsilon/2} 10^9 \eta l} \quad (11)$$

Similarly, the equivalent series resistance is:

$$R_{h(ser)} = \frac{2^{\epsilon/2} 16\pi^2 \eta \mu^2 N^2 A\alpha f}{10^9 B^{(2-\epsilon)} l} \quad (11a)$$

The corresponding dissipation factor, reduced by Equation (6), is:

$$D_h = \frac{\omega L}{R_{h(par)}}$$
$$= \frac{2\pi f 4\pi N^2 \mu A\alpha}{10^9 l} \cdot \frac{2^{\epsilon/2} 10^9 \eta l}{4\pi^2 N^2 B^{(2-\epsilon)} f A\alpha}$$
$$= \frac{2^{\epsilon/2} 2\eta\mu}{B^{(2-\epsilon)}} = h \quad (12)$$

This factor is independent of frequency and has the value:

$$h = 2^{(1+\epsilon/2)} \eta\mu B^{(\epsilon-2)} \quad (12a)$$

EDDY-CURRENT LOSS (IRON)

The power expended in eddy-current loss in the iron is given by:

$$P_e = \frac{\pi^2 f^2 B_m{}^2 \delta^2 V}{6 \times 10^{16} \rho_i} = \frac{\pi^2 f^2 2 B^2 \delta^2 V}{6 \times 10^{16} \rho_i} \quad (13)$$

Since power $= \frac{E^2}{R}$, the equivalent parallel resistance, reduced by Equations (5) and (6), is:

$$R_{e(par)} = \frac{E^2}{P_e} = \frac{I^2\omega^2 L^2}{P_e} = \frac{12\rho_i A\alpha N^2}{\delta^2 l} \quad (14)$$

The corresponding dissipation factor, reduced by Equation (6), is:

$$D_e = \frac{\omega L}{R_{e(par)}} = \frac{2\pi f 4\pi N^2 \mu A\alpha}{10^9 l} \cdot \frac{\delta^2 l}{12\rho_i A\alpha N^2}$$
$$= \frac{2\pi^2 \delta^2 \mu f}{3\rho_i 10^9} = ef \quad (15)$$

This dissipation factor is directly proportional to frequency, and the factor of proportionality is:

$$e = \frac{2\pi^2 \delta^2 \mu}{3\rho_i 10^9} \quad (16)$$

EQUIVALENT CIRCUIT REPRESENTING LOSSES

It is interesting to note how this analysis demonstrates the correctness of the usual method of showing the equivalent circuit of an iron-cored coil or transformer, as in Figure 2. Here R_c and R_e are resistances independent of frequency, representing respectively ohmic loss in the copper and eddy-current loss in the iron. Equations (7) and (14) show the invariance of R_c and R_e with frequency. R_h, on the other hand, whether calculated as a series or as a parallel resistance [Equations (11) and (11a)], varies with the first power of frequency. Hysteresis loss, therefore, cannot be represented as a resistance independent of frequency and hence is not shown in the equivalent circuit of Figure 2.

TOTAL LOSS

The total dissipation factor, D, is the sum of the three separate dissipation factors:

$$D = \frac{c}{f} + h + ef \qquad (17)$$

OPTIMUM CONDITIONS

When plotted on log-log paper each component is a straight line as shown in Figure 3, that for hysteresis being horizontal and those for copper and eddy current being slanted down and up at 45°, respectively. Minimum D occurs where the c and e lines cross, at a frequency given by

$$f_m = \sqrt{c/e} \qquad (18)$$

At this frequency the minimum D is

$$D_m = h + 2\sqrt{ce} \qquad (19)$$

When a curve of D for any coil has been found experimentally, numerical values for the three coefficients c, h, and e, can be found by drawing 45° asymptotes to the curve. The intercepts of these lines with the 1-cycle axis are the values of c and e. The value of h is the difference between the observed minimum and twice the value of the two asymptotes at their crossing point.

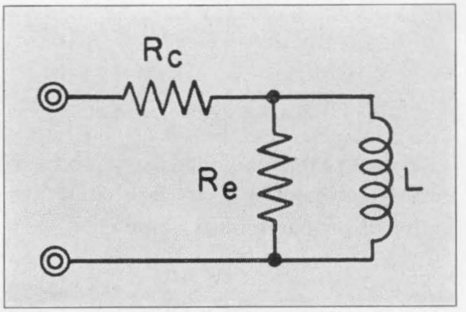


FIGURE 2. Equivalent circuit of a low-frequency coil.

Q—STORAGE FACTOR

Engineers in the radio and audio fields are more accustomed to think in terms of Q, the reciprocal of D, than in terms of D. Unfortunately, the expression just developed gives:

$$Q_m = \frac{1}{h + 2\sqrt{ce}} \qquad (20)$$

This is easily enough calculated in a given case, but it does not lend itself readily to quick mental calculations because of the presence of h. However, there is a fortunate circumstance which makes neglect of h in this expression allowable.

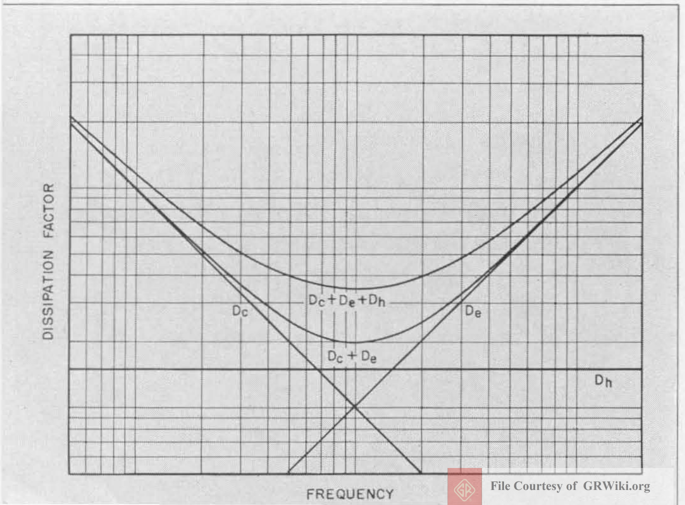


FIGURE 3. Dissipation factor-frequency relationships.

Separate dissipation factor curves are the three labeled straight lines.

At initial permeability ($D_h = 0$), the lower curve "$D_c + D_e$" represents coil behavior. Invert it to secure a Q-curve.

Above initial permeability ($D_h \neq 0$), the upper curve "$D_c + D_e + D_h$" is representative of the blunting action of a finite D_h. f_{max}, although less easily determined, is unchanged

SIMPLIFICATION OF EXPRESSION FOR Q

The factor, h, Equation (12a), contains a term $B^{(\epsilon-2)}$. All of the measurements which we have described in Mr. Arguimbau's article and in this one have been made with such a low flux density in the iron that the initial permeability plateau has been reached. This, for high-silicon steels, is in the region of B below one gauss, or, really, a place where B is approaching zero. It is very helpful to make inductance measurements on this plateau, since unavoidable small changes in the supply voltage make imperceptible changes in the permeability and, hence, the inductance. Contrariwise, once the B has become large enough so that permeability has begun to increase, it is no longer possible to have μ independent of the effects of voltage applied to a coil having a core of ferro-magnetic material (with the notable exception of some dust cores). The initial permeability plateau is the most easily reproducible measuring condition and its only drawback is the high gain required ahead of the detector in a measuring circuit. One can always be sure that he is on this plateau when making measurements by continuously reducing the voltage applied to the coil until successive reductions make no further change in the measured value.

A curve is shown in Figure 4 of the relation, at very low magnetizing forces, between H and μ for a high-silicon steel, reproduced by permission from a paper by H. W. Lamson.* This shows the plateau just referred to, a phenomenon not widely known. Note that the curve starts to ascend from the plateau ($\mu_t = 470$) at about $H = 0.001$ oersted, which corresponds to a $B(=\mu_t H)$ of about 0.5 gauss. This will indicate the order of smallness of excitation to reach the plateau for a 4% silicon steel.

If the customarily used value of ϵ, namely, the 1.6 figure of Steinmetz, holds good for these very low flux densities, then h becomes infinite, since B (approaching zero) goes into the denominator. This, we know physically, is not true, and there is implicit corroboration in the three examples of this article. However, there is a better authority than this. It is well, but not generally, known that the Steinmetz exponent is not a constant at all, but varies with flux density B and happens to have a value very close to 1.6 in the middle region, say between 2 and 10 kilogausses. On the other hand, the value of this exponent increases for both very high and very low flux densities. Values as high as three or more can readily be found in the literature for

*Proc. I.R.E., Vol. 28, No. 12, p. 546 (Dec., 1940).

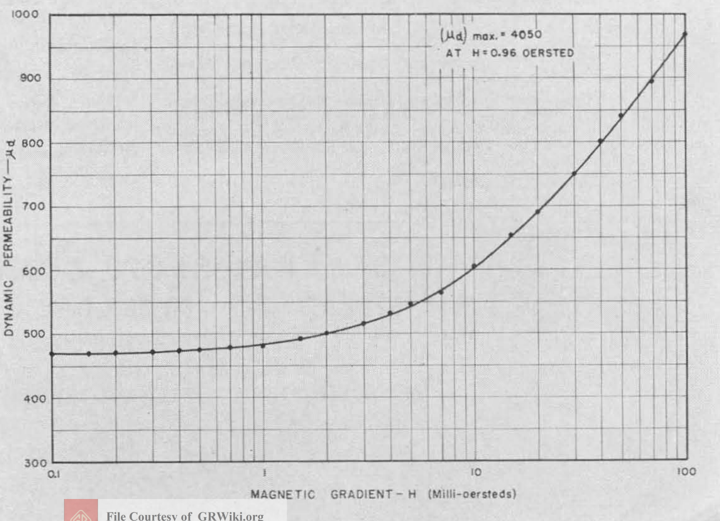


FIGURE 4. Permeability of silicon steel at low inductions, showing the initial-permeability plateau.

very high values of B,† and an exponent of 2.4 for very low B values is given.†† Further, Rayleigh, in 1887, showed that $\epsilon = 3$ at low flux densities, and this has been confirmed recently by Elwood.‖

If the Steinmetz exponent in the initial permeability region is greater than 2, then h approaches zero as B approaches zero. This means that, in the initial permeability region, h can be neglected and

$$Q_m = \frac{1}{2\sqrt{ce}}. \tag{21}$$

FINAL EXPRESSIONS FOR Q_m AND f_m

Substituting the values for c and e from Equations (9) and (16):

$$Q_m = \frac{1}{\delta}\sqrt{\frac{3\rho_i SA\alpha}{\rho_c tl}} \tag{22}$$

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

EFFECTS OF TEMPERATURE

The variation of Q_m and f_m with temperature can be calculated using the temperature coefficients of copper and iron resistivities. ρ_c has a temperature coefficient of $+0.4\%$ per degree C, while ρ_i has one of $+0.5\%$. Temperature coefficient of c is then $+0.4\%$, of e -0.5%. D_m (or Q_m) is sensibly constant with temperature, but f_m increases about 0.5% per degree C.

†Spooner, "Properties and Testing of Magnetic Materials" (1927), p. 24.

††Page 325 of Vol. 2 of the Dictionary of Applied Physics, 1922 Edition, quoting an article by A. Campbell, "Magnetic Properties of Stalloy in Weak Alternating Fields," Phys. Soc. Proc. 1920, XXXII, 232. (Stalloy is an English high-silicon steel.)

‖Physics 6, 215, 1935.

NOTES ON EXPRESSIONS (22) AND (23)

Although this article concerns itself only with using the equations for comparative, not absolute, purposes, the Expressions (22) and (23) will determine Q_m and f_m from constants of the core structure. This has actually been done and fairly good agreement with measured values obtained.

For example, Q_m for GR-345 core (¾″-tongue) calculates 37.7 and measures from 33 up to 40. Also, for GR-485 core (1 5⁄16″-tongue) Q_m calculates 49.5, measures 43 to 48. In each case the very low, disagreeing, measured values occur at high frequencies (large air gaps) where the uncertainties of Q measurements increase and where there is more probability of eddy-current losses in the copper because of fringing.

Calculations of f_m in cycles per second using as a basis the plateau μ_t of 470 (from Figure 4) in Equation (3a) are tabulated below:

Center-Leg Gap	*f_m-345 Coil Calc.*	*f_m-345 Coil Meas.*	*f_m-485 Coil Calc.*	*f_m-485 Coil Meas.*
Interleaved	97	122	74	84
1⁄16″	631	700	379	480
⅛″	1164	1050	685	700
¼″	——	——	1294	1040
½″	4360	2600	——	——
0.95	——	——	4700	2500

It will be noted that this table bears out the statement made under heading EXPRESSIONS FOR Q_m AND f_m early in the paper that the equivalent air gap is almost never the same as the measured gap, being larger than the measured value for small gaps and smaller for large gaps (larger gap means smaller μ, which means larger f_m). The μ for interleaved laminations is indicated to be slightly less than that for a ring-sample (truly gapless) material. Contrariwise, for the largest illustrative gaps in these examples the μ is larger than one would be led to expect by theory.

—P. K. McElroy and R. F. Field

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

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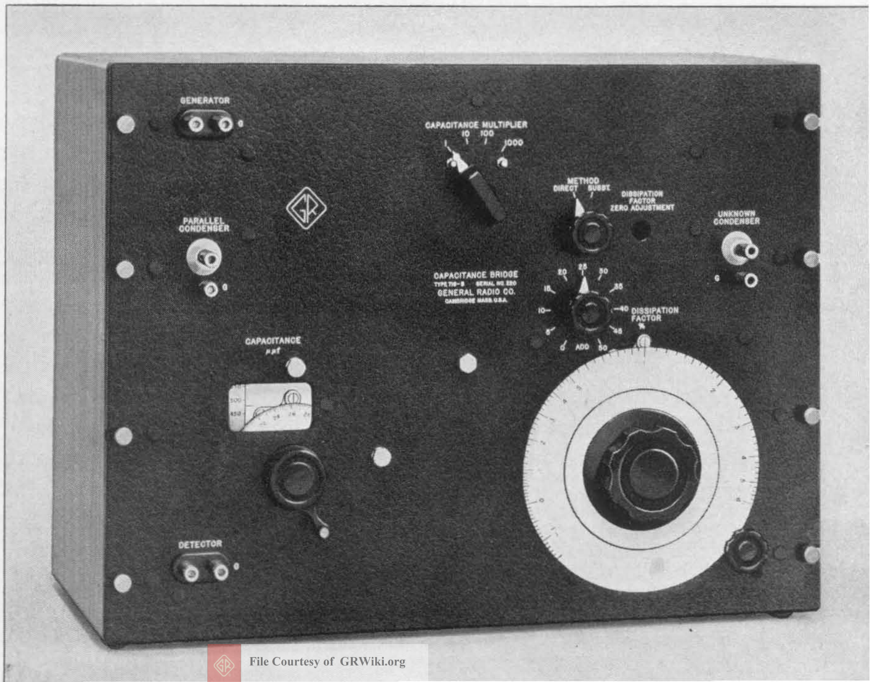
INCREASED POWER-FACTOR RANGE FOR THE CAPACITANCE BRIDGE

● **IN THE FIVE YEARS SINCE ITS ANNOUNCEMENT** the TYPE 716-A Capacitance Bridge has fulfilled its design specifications as an accurate instrument for the measurement of capacitance and as a worthy successor of the old TYPE 216 Capacity Bridge. Its direct-reading precision condenser, in conjunction with four decade-spaced ratio arms, makes it possible to measure capacitances from 100 $\mu\mu$f to 1 μf. Its use of the Schering bridge circuit simplifies the

FIGURE 1. Panel view of the TYPE 716-B Capacitance Bridge. The new features are the DISSIPATION FACTOR switch which controls a step condenser and the METHOD switch which allows either DIRECT or SUBSTitution measurements to be made.



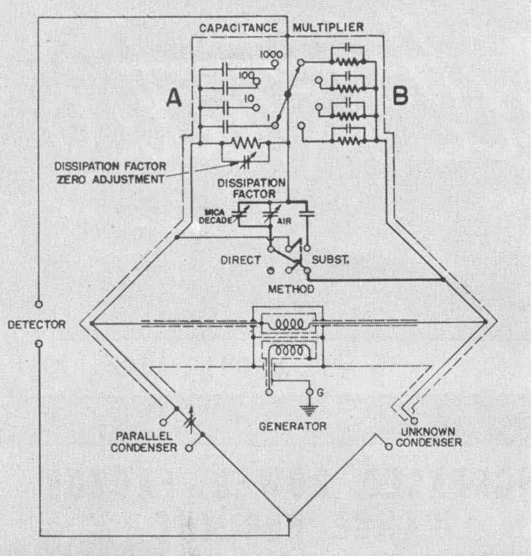


FIGURE 2. Circuit diagram of the new bridge. Changes and additions are shown by heavy lines. The mica decade condenser increases the dissipation factor range to 56%. The reversing switch makes possible a positive dissipation factor reading in substitution measurements.

measurement of dielectric losses by allowing the dial of the air condenser connected across the fixed ratio arm to be calibrated directly in dissipation factor.* The range of this dial, 0.06 (or 6%), has limited somewhat the usefulness of the bridge, particularly at frequencies below 1 kc, where the dissipation factor of commercial dielectrics tends to increase. It is both for this reason and to simplify the use of the bridge in substitution measurements that a new model, the TYPE 716-B Capacitance Bridge, is now introduced.

The circuit diagram of the new model is shown in Figure 2. The changes and additions are shown with heavy lines. The dissipation factor range of the older model was limited by the capacitance of the air condenser to slightly over 6%, corresponding to a change in capacitance of 500 μμf. A decade condenser, in which the unit steps are 398 μμf, has been added in the new model. This condenser is mounted in the upper part of the insulated compartment containing the dissipation factor condenser, as shown in Figure 3. The various steps are controlled by a TYPE 380 Switch shown at the right of the condenser case and which appears on the panel in Figure 1 just above the DISSIPATION FACTOR dial. Each step adds 5% to the dissipation factor reading, giving a maximum value of 56%. For this value, dissipation factor and power factor differ by 13%, and consequently the designation POWER FACTOR used on the older model has been changed to DISSIPATION FACTOR. This increased range in dissipation factor at 1 kc extends correspondingly the range at lower frequencies, being 5.6% at 100 cycles and 3.3% at 60 cycles.

When the bridge is used for substitution measurements, a balancing condenser must be placed across the UNKNOWN CONDENSER terminals and the unknown connected to the PARALLEL CONDENSER terminals in parallel with the internal precision condenser. The reading of the DISSIPATION FACTOR dial must then decrease, and the short negative scale of 0.15% is sufficient only for condensers with relatively small dissipation factor. The alternative of causing the DISSIPATION FACTOR dial to read up-scale initially by adding a suitable condenser across the *B* ratio arm has proved cumbersome. In the new model a reversing switch has been added, as shown in Figure 2, which transfers the dissipation factor condensers from the *A* to the *B* ratio arm and at the same time connects a small condenser across the *A* arm, equal to twice the zero capacitances of those transferred. This condenser and the reversing switch are shown in Figure 3.

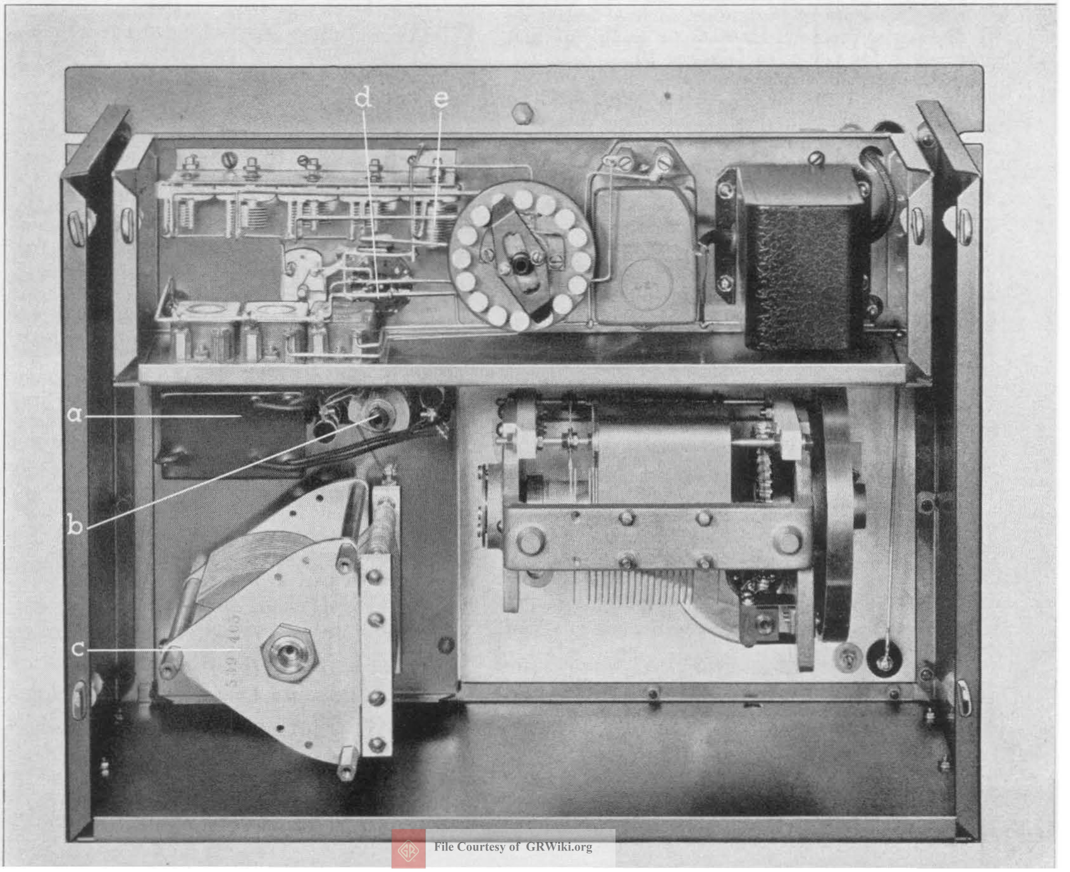
*Dissipation factor is the cotangent of the phase angle, while power factor is the cosine.

On the panel in Figure 1 the switch appears above the new DISSIPATION FACTOR switch as the METHOD switch with the two positions DIRECT and SUBSTitution. Schematic wiring diagrams of the bridge for these two positions of the METHOD switch are given in Figure 4. With the switch in the SUBSTitution position the dissipation factor range of the bridge at 1 kc is 56% multiplied by the ratio of the total capacitance to the unknown capacitance. The bridge can be balanced only for equal ratio arms with the CAPACITANCE MULTIPLIER switch set at 1.

Whenever in the Schering bridge circuit there are capacitances across both of the ratio arms, the simple bridge equations no longer hold, and the dial readings are in error for both capacitance and dissipation factors. These errors are approximately equal to the product of the dissipation factors of the two ratio arms. They are, therefore, proportional to the dissipation factor reading of the bridge and at the maximum reading of 56% can amount to almost 2%. Errors of a similar nature can occur even in substitution measurements.

The new edition of the operating instructions, Form 455-C, supplied with the Type 716-B Capacitance Bridge,

Figure 3. Rear view of the bridge with shields removed. The mica decade condenser, a, is controlled by a Type 380 Switch, b. The reversing switch, d, transfers the dissipation factor air condenser, c, together with the decade condenser, a, from the A to the B ratio arm for substitution measurements and at the same time places condenser, e, across the A ratio arm to make up for the zero capacitances of condensers, d, and, a.



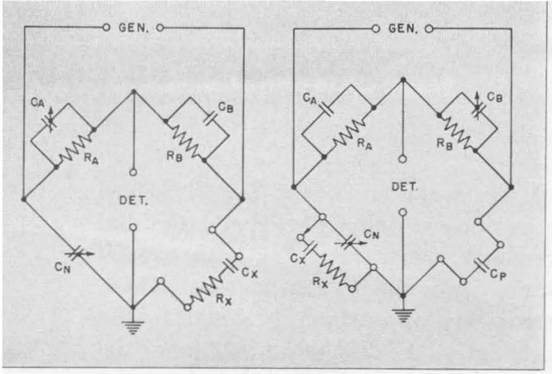


FIGURE 4. These diagrams show the bridge circuit for the two positions of the METHOD switch. The left-hand diagram is for the DIRECT position, the right-hand for the SUBSTitution position.

contains a complete discussion of the errors applying both to direct and substitution measurements. Conversion formulae for changing from series to parallel impedances are also included. Users of the older TYPE 716-A Capacitance Bridge will find the new booklet of considerable help in the measurement of a high resistance or any capacitance having very large dissipation factor. A copy will be sent on request.

— R. F. FIELD

SPECIFICATIONS

Ranges: Direct Reading — capacitance, 100 $\mu\mu$f to 1 μf; dissipation factor, 0.002% to 56% (0.00002 to 0.56 expressed as a ratio).

Substitution Method — capacitance, 0.1 $\mu\mu$f to 1000 $\mu\mu$f with internal standard; to 1 μf with external standards; dissipation factor, $56\% \times \frac{C'}{C_x}$ where C' is the capacitance of the standard condenser and C_x that of the unknown.

Accuracy: Direct Reading — capacitance, ±0.2% or ±2 $\mu\mu$f × multiplier reading (0.2% of full scale for each range) when the dissipation factor of the unknown is less than 1%; dissipation factor ±0.0005 or ±2% of dial reading, for values of D below 10%.

Substitution Method — capacitance ±0.2% or ±2 $\mu\mu$f; dissipation factor, ±0.00005 or ±2% for change in dissipation factor observed, when the change is less than 6%.

When the dissipation factor of the unknown exceeds the limits given above, additional errors occur in both capacitance and dissipation-factor readings. Corrections are supplied, by means of which the accuracy given above can be maintained over the entire range of the bridge.

Ratio Arms: The arm across which the dissipation factor condenser is normally connected has a resistance of 20,000 ohms. The other arm has four values, 20,000 ohms, 2000 ohms, 200 ohms, 20 ohms, providing the four multiplying factors 1, 10, 100, 1000. Suitable condensers are placed across these arms, so that the product RC is constant.

Standards: Capacitance, TYPE 722 Precision Condenser direct reading from 100 $\mu\mu$f to 1100 $\mu\mu$f; dissipation factor, TYPE 539-T Condenser with semi-logarithmic scale and decade-step condenser calibrated directly in dissipation factor at 1 kc.

Shielding: Ratio arms, dissipation-factor condensers, and shielded transformer are enclosed in an insulated shield. The unknown terminals are shielded so that the zero capacitance across them is not greater than 1 $\mu\mu$f. A metal dust cover and the aluminum panel form a complete external shield.

Frequency Range: All calibration adjustments are made at 1 kc and the accuracy statements above hold for an operating frequency of 1 kc. The bridge can be used, however, at any frequency between 60 cycles and 10 kc. Dissipation-factor readings must be corrected by multiplying the dial reading by the frequency in kilocycles.

Voltage: Voltage applied at the GENERATOR terminals is stepped up by a 1-to-4 ratio shielded transformer. A maximum of 50 volts can be applied to the transformer. If desired, power can be applied to the bridge between the junctions of the pairs of resistance and capacitance arms. With equal ratio arms, a maximum of 700 volts can be applied.

Mounting: The bridge is supplied for mounting on a 19-inch relay rack or for cabinet mounting.

Accessories Required: Oscillator, amplifier, and telephones or rectifier meter. TYPE 608-A Oscillator, TYPE 814-A Amplifier, and Western Electric TYPE 1002-C Telephones are recommended.

For substitution measurements, a balancing condenser is needed. This may be either an air-dielectric model, TYPE 539-C, or a fixed mica condenser of the TYPE 505 series.

Accessories Supplied: One TYPE 274-M Plug, one TYPE 274-NC Shielded Conductor, and one TYPE 274-NE Shielded Plug and Cable.

Dimensions: (Length) 19 x (height) 14 x (depth) 9 inches, over-all.

Net Weight: 41½ pounds, relay-rack model; 53¾ pounds, cabinet model.

Type		*Code Word*	*Price*
716-BR	For Relay-Rack Mounting.	BONUS	$335.00
716-BM	Cabinet Mounted........	BOSOM	360.00

SUBSTITUTE MATERIALS

● **RECENT CONSERVATION ORDERS** of the War Production Board particularly concerning aluminum have made substitutions for this metal imperative. New conservation regulations covering other materials are constantly being issued, and one of the principal jobs of our design staff is to find adequate substitutes for all of these critically scarce materials.

A primary objective of our substitution program must be to make use of the new materials with the minimum effect upon performance and with as little interference with the present rush production program as possible. This task is by no means a simple one, because the performance and utility of precision equipment frequently depend upon the materials used, and because even slight changes in design will occasion delays in production. Nevertheless, progress is being made, and substitutes are gradually being found which can be used without adversely affecting the performance specifications.

The problem is further complicated by the fact that shortages sometimes develop in substitutes, so that it may be necessary to use different materials at different times for a given instrument part. It should be recognized, therefore, that two instruments of the same type number made at different times may not weigh the same or even look exactly the same.

All of our facilities are devoted to war production. We must try in every way possible to keep vital material moving, but if your deliveries are delayed we hope that you will be patient while the technical and manufacturing problems connected with substitutions are being solved.

—A. E. Thiessen

PRIORITIES AND REPAIRS

● **BECAUSE PRACTICALLY ALL OF OUR MANUFACTURING FACILITIES**, as well as those of our suppliers, are devoted to war projects with high priority ratings, it is becoming increasingly difficult for us to repair instruments or to supply replacement parts under the repair rating of A-10. Reasonable delivery of materials and components such as wire, metal parts, meters, condensers, resistors, tubes, batteries, etc., cannot usually be had except under priority ratings much higher than A-10.

If equipment is being used directly or indirectly on war projects covered by a rating higher than A-10, then a properly executed preference rating extension will insure the repair being made within a time consistent with the higher rating. Otherwise, it will be necessary for us to order replacement parts or material under the A-10 repair rating, and it is quite probable that delays of many months will result. We must receive the proof of preference rating and purchase order before work is started on any repair.

The Service Department is equipped to handle repairs very promptly. Returned instruments being used on high priority projects are given preference in accordance with their individual ratings and dates, which, of course, are beyond our control. Whether or not you have a preference rating, remember that a set of our Service and Maintenance Notes will often enable you to make repairs and readjustments in your own plant.

— H. H. Dawes

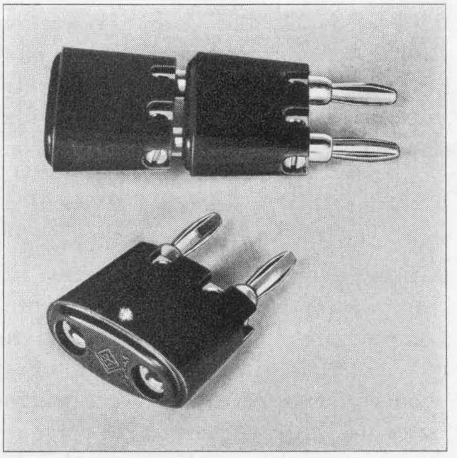
A NEW TYPE 274-M DOUBLE PLUG

● **ANOTHER FAMILIAR STAPLE** of the General Radio line takes on a new and streamlined form with the announcement of the new TYPE 274-M Double Plug.

The new plug, which replaces both the TYPE 274-M (black bakelite) and the TYPE 274-ML (yellow bakelite), is molded from polystyrene, a comparatively new molding material with greatly superior electrical characteristics.

The new TYPE 274-M has a power factor of about .07% at 1000 cycles per second as compared to a power factor of 13.2% for the black bakelite and 1.70% for the yellow bakelite of the previous type. The leakage resistance between pins is greater than 10^8 megohms. This is of the same order of magnitude as the earlier yellow bakelite plug but compares with only 65,000 megohms for the black bakelite.

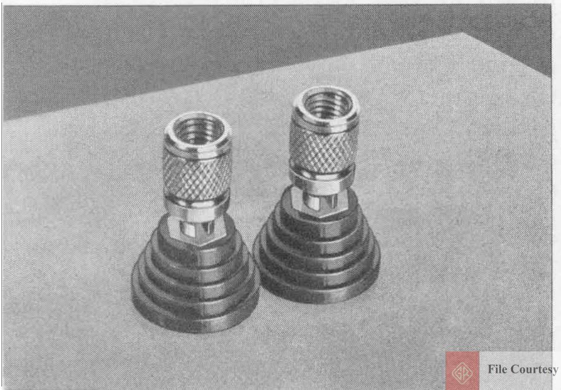
The capacitance between pins has also been somewhat reduced. It is now about 1 $\mu\mu$f compared with 1.5 and 1.75 for the yellow and the black bakelite respectively.



The new plug is of improved appearance, in conformity with present tendencies for simple forms, and is so shaped that it can be easily and positively gripped by the fingers. A dot is molded into the rim of one jack in order to make possible the identification of terminals.

Type		*Code Word*	*Price*
274-M	Double Plug	STANPARBUG	$0.50
	Package of 10		3.50

LOW-CAPACITANCE TERMINALS



● **A COMPANION ITEM** to the new TYPE 274-M Double Plug is the TYPE 138-UL Binding Post Assembly, designed for uses where low capacitance and low leakage conductance are required.

FIGURE 1. View of a pair of TYPE 138-UL Terminals mounted on a metal panel.

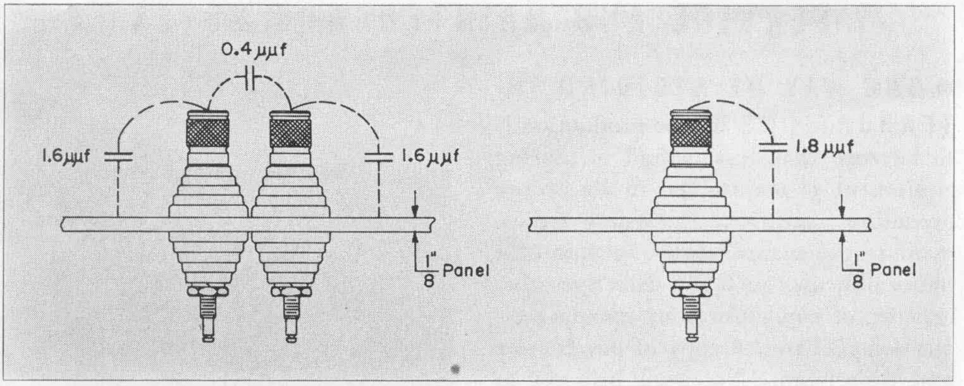


FIGURE 2. Capacitances associated with a pair of terminals and a single terminal mounted on a ⅛″ metal panel. The capacitance between binding post and panel includes the capacitance to free space.

The TYPE 138-UL Binding Post Assembly consists of a nickel-plated brass binding post (with knurled top), and two hollow conical insulators, molded of polystyrene. It is designed for mounting on panels from 1⁄16″ to ¼″ thick, through a 11⁄16″ hole. With this mounting hole the spacing between panel and stud exceeds ¼″, insuring a very low capacitance to panel. The power factor associated with this capacitance is low, because the only solid dielectric is low-loss polystyrene. The d-c leakage resistance through the polystyrene is also low, and the effect of surface leakage is minimized by the use of a stepped conical insulator, which increases the length of the leakage path over that of a smooth cone. Figure 1 shows a pair of terminals mounted on a panel, while Figure 3 shows the details of the various parts.

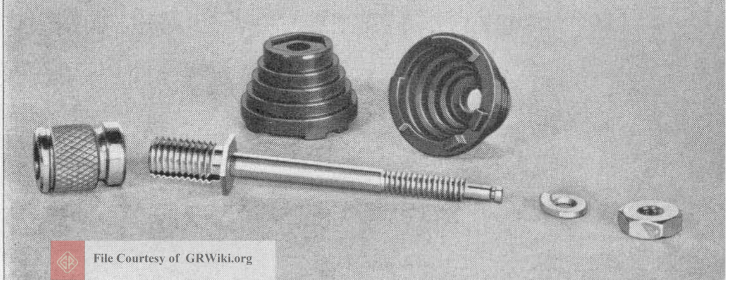
The sketches of Figure 2 show the capacitances associated with a pair of terminals and with a single terminal, mounted on a ⅛″ metal panel. The effective leakage resistance is greater than 10^8 megohms.

The shank is slotted to take a wire or a pin terminal and is drilled to take a TYPE 274 Plug. Two of these terminals can be mounted with ¾″ spacing to take the TYPE 274-M Double Plug.

Although they are not designed primarily for high-voltage use, these terminals can safely be used at voltages up to 5000 volts.

Type		*Code Word*	*Price*
138-UL	Binding-Post Assembly	STANPARULE	$0.50
	Package of 10		4.00

FIGURE 3. View of a TYPE 138-UL Terminal disassembled, showing the component parts. The projections on the base of the two polystyrene pieces interlock so that the terminal can be mounted on a panel as thin as 1⁄16″. The narrow slot is provided to take a projecting key on the mounting hole to prevent the insulator from turning.



SERVICE AND MAINTENANCE NOTES

● **ONE WAY OF AVOIDING INTERRUPTIONS** in war production is to be sure that testing and measuring equipment is maintained in its proper operating condition. Returning equipment to the manufacturer because of a minor difficulty such as a defective tube, resistor, or condenser, may mean a serious delay to you. A copy of our Service and Maintenance Notes will help you to avoid this situation with General Radio equipment.

These notes, which have been compiled from the records of the Service Department, Standardizing Laboratory, and Engineering Department, will, in most cases, enable the user to locate and remedy ordinary operating difficulties that do not require the use of elaborate equipment for testing and checking.

The notes are sent free of charge to users of General Radio equipment. They have already been distributed to many of our customers and have proved their value in obviating the return of instruments to our factory for minor repairs. We urge you to send us the type and serial numbers of your General Radio equipment so that your copy of Service and Maintenance Notes can be mailed promptly.

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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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

THE TYPE 727-A VACUUM-TUBE VOLTMETER

A PORTABLE BATTERY-OPERATED INSTRUMENT

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● **FOR A LONG TIME** there has been need for a general purpose vacuum-tube voltmeter which was battery operated and truly portable. The TYPE 726-A instrument, introduced in 1937, has filled the need for an instrument of the laboratory type where line power is available, but is often inconvenient for field work on account of its size and the required line connection. The new TYPE 727-A Vacuum-Tube Voltmeter is designed particularly for use in the field.

Both the new instrument and the older TYPE 726-A operate from the lowest audio frequencies up through the moderately high radio frequencies beyond 100 megacycles. Both are intended to cover as wide a voltage range as is reasonably practicable over such a wide frequency band. The different power supplies

FIGURE 1. View of the voltmeter with cover open.

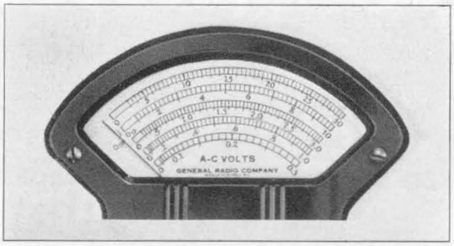


FIGURE 2. Close-up view of the meter scale.

and the size and weight considerations, however, result in somewhat different design compromises in the two cases.

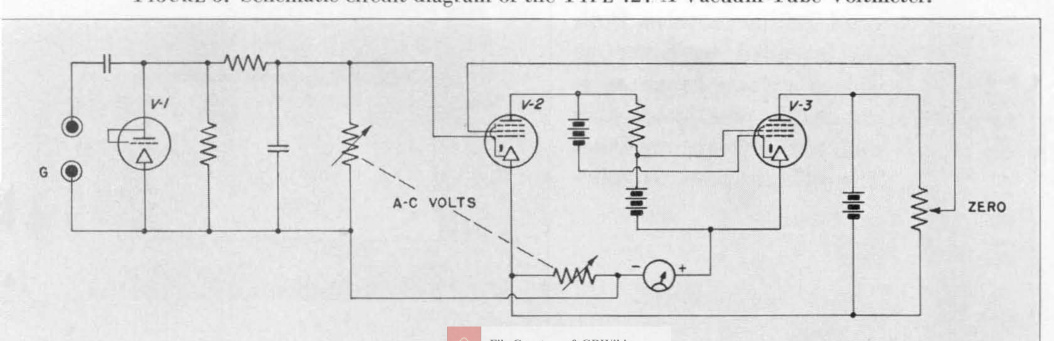
In the new portable meter the stability of the battery power supply, in comparison with that of the regulated power-line supply of the older instrument, makes it possible to increase the sensitivity substantially without fluctuations or zero drift becoming bothersome. The most sensitive range gives full scale deflection on only 300 millivolts with 50 millivolts easily readable. On the TYPE 726-A Meter the most sensitive range is 1.5 volts full scale, and 0.1 volt is the lowest calibrated point. The new instrument is particularly convenient where readings of the order of a few tenths of a volt are to be made.

In the high-voltage direction the range of the new portable instrument extends to 300 volts without the use of an external multiplier, in place of the previous 150-volt limit. This two-to-one increase is obtained with a possible slight loss in accuracy on the higher ranges, because a high-resistance voltage divider is employed in the d-c output circuit of the diode rectifier. Without this divider the voltage that can be measured is limited essentially by the B-supply voltage available, which must be made as low as possible in a battery-operated instrument to save weight. Since the expedient of a d-c voltage divider is made necessary in any event to reduce the battery requirements, it has been taken advantage of, within the limit set by the voltage rating of the rectifying diode, to increase the high-voltage limit of the instrument.

On the 0.3, 1, 3, and 10-volt ranges the sensitivity is largely determined by wire-wound resistances of relatively low value. The high-resistance voltage divider above referred to is used on the 30, 100, and 300-volt ranges only. The limits of accuracy of these upper three ranges are given as 5% of full scale instead of the 2% limit for the 1, 3, and 10-volt ranges, to allow for possible slow or seasonal drift in the divider ratio. Experience indicates that this allowance is conservative. In any event the maximum accuracy of 2% can be realized also on the high ranges if the setting of the internal calibration adjustments for these ranges can be checked occasionally.

In the new instrument the separate probe has been omitted to save space, and the rectifier circuit has been built into a compartment in the instrument, adjacent to the terminal posts. A con-

FIGURE 3. Schematic circuit diagram of the TYPE 727-A Vacuum-Tube Voltmeter.



struction has been worked out which gives a resonant frequency for the input loop only slightly lower than for the separate probe arrangement. The probe type of construction raises the upper frequency limit slightly and has also proved convenient in permitting the measuring circuit to be located at the point where the voltage is to be determined. On the other hand, the very small size of the new instrument makes this construction unnecessary for many applications, and many experimental set-ups can be arranged to make the reference point of the circuit at the meter terminals.

The transit time error is slightly less in the new instrument than in the TYPE 726-A. Consequently the frequency correction varies less with voltage over the range covered by the instrument. A single correction factor for frequency, therefore, can be applied for all voltage readings at a given frequency. The frequency correction curve for the new instrument is shown in Figure 4 in comparison with that for the TYPE 726-A Voltmeter. Although the latter instrument can be used at somewhat higher frequencies, the correction is less convenient to apply if a wide range of voltages is to be covered.

One sacrifice in the new instrument is in regard to the input capacitance. This is 16 $\mu\mu$f, or more than double that obtained with the separate probe construction. Figure 5 shows the resistive and reactive components of the input impedance as functions of frequency. It will be seen that the parallel resistive component has dropped to approximately 100,000 ohms at 20 megacycles. In many applications the parallel capacitance component can be taken care of by slight retuning. The losses are negligible for most ordinary applications but can be taken into account if desired.

Several improvements of design contribute materially to the convenience of operation of the instrument. Five scales are employed on the meter face for the seven voltage ranges, slight offsetting of the zero permitting the two highest voltage ranges to be read on the scales of the ranges below them. The five scales are printed alternately red and black, which materially reduces the eyestrain and effort involved in concentrating attention on any one scale.

Another improvement is that a circuit modification permits the zero of all four of the most sensitive ranges to be set by a single adjustment, thus obviating the use of any compensating arrangement which might get out of adjustment in time. The three highest

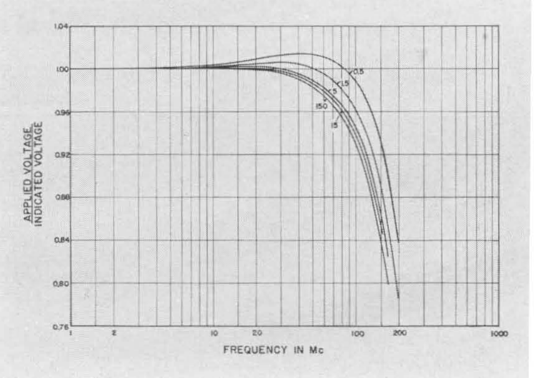
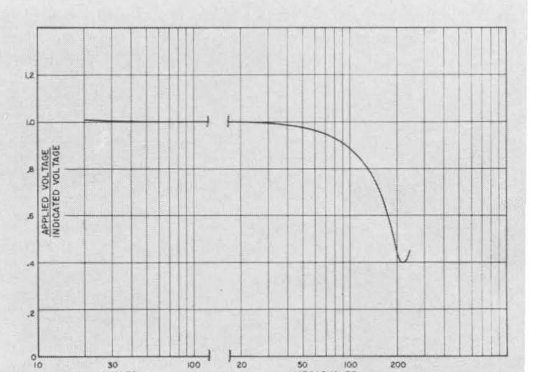


FIGURE 4a (*above*). Frequency characteristics of the older TYPE 726-A Vacuum-Tube Voltmeter.

FIGURE 4b (*below*). Frequency characteristics of the new TYPE 727-A Vacuum-Tube Voltmeter.



ranges of the instrument, however, still rely on the compensation method for zero alignment, but, since these ranges are the least sensitive, inconvenience seldom results.

The size and weight will be seen from the specifications to be very much reduced. The battery complement, giving a life of approximately 250 hours of intermittent operation, consists only of three 1.5-volt filament batteries and two 30-volt plate batteries. This results not only in light weight, but in usually low replacement cost.

It is felt that, where size and weight considerations are important, as well as for those cases where battery operation is required, the new instrument will fill an important need in the vacuum-tube voltmeter field.

In appearance and general construction, the new voltmeter resembles the Type 729-A Megohmmeter previously described.*

— W. N. Tuttle

*W. N. Tuttle, "A Portable Megohmmeter," *Experimenter*, Vol. XV, No. 2, July, 1940.

SPECIFICATIONS

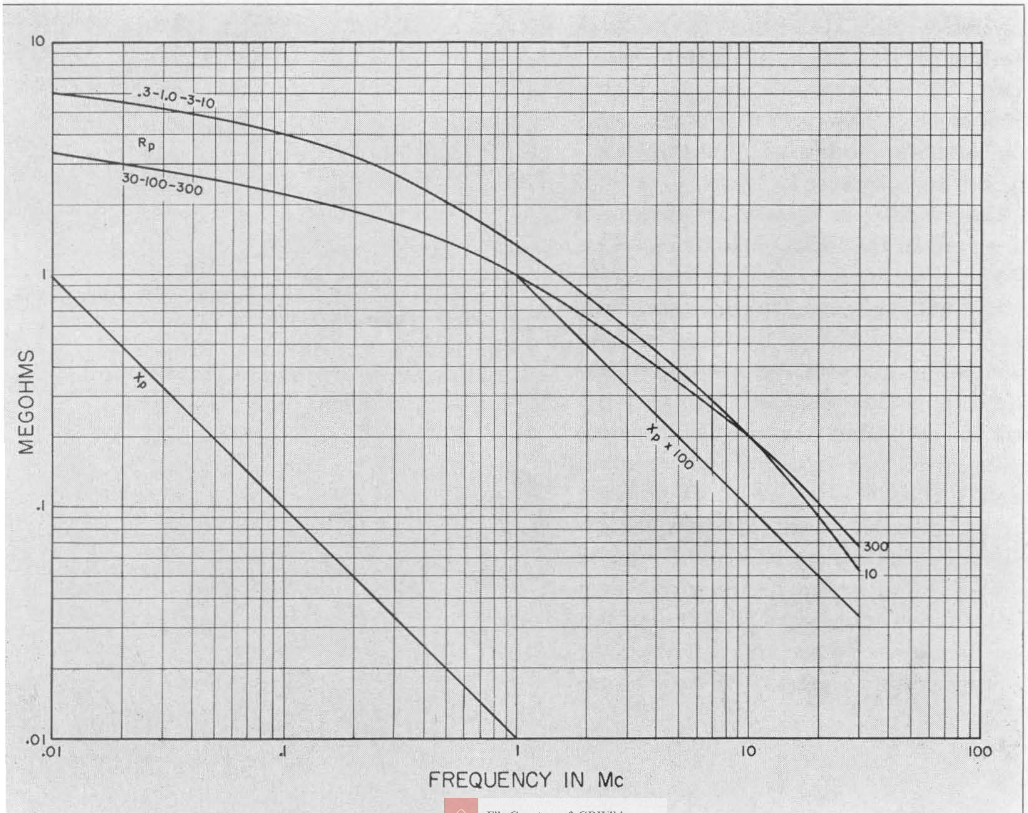
Range: 0.05 volt to 300 volts ac, in seven ranges (0.3, 1, 3, 10, 30, 100, 300 volts, full scale).

Accuracy: For sinusoidal voltages, ±3% of full scale on the 0.3-volt range;

±2% of full scale on the 1, 3, and 10-volt ranges;

±5% of full scale on the 30, 100, and 300-volt ranges (see text for a discussion of the accuracy of these ranges).

Figure 5. Input impedance of the Type 727-A Vacuum-Tube Voltmeter as a function of frequency.



Waveform Error: The instrument is calibrated to read the r-m-s value of a sinusoidal voltage. On the higher voltage ranges, however, it is essentially a peak reading device, calibrated to read 0.707 of the peak value of the applied voltage, and on distorted waveforms the percentage deviation of the reading from the r-m-s value may be as large as the percentage of harmonics present. On the lowest ranges the instrument approximates a true square-law device.

Frequency Error: Less than 1% between 20 cycles and 30 Mc. At higher frequencies, the error is about +5% at 65 Mc and about +10% at 100 Mc.

Input Impedance: The input capacitance is approximately 16 $\mu\mu$f. The parallel input resistance (at low frequencies) is about 5 megohms on the lower ranges and about 3 megohms on the 30, 100, and 300-volt ranges. The curves of Figure 5 give the variation of R_P and X_P with frequency.

Temperature and Humidity Effects: Over the normal range of room conditions (65° Fahrenheit to 95° Fahrenheit; 0 to 95% relative humidity) the accuracy of indication is substantially independent of temperature and humidity conditions. Somewhat reduced accuracy may be expected, however, if the instrument is subjected to extremes of temperature.

Zero Adjustment: A zero adjustment is provided on the panel. The setting is the same for all ranges.

Vacuum Tubes: Two 1S5 tubes and one 957 tube are used and are supplied with the instrument.

Batteries: Two Burgess W20P1, one Burgess W5BP, and three Burgess 2F batteries are required, and are supplied with the instrument. Battery life is approximately 250 hours of intermittent operation.

Mounting: The instrument is supplied in a walnut case with cover and is mounted on an engraved black crackle-finish aluminum panel.

Dimensions: 11 x 6⅝ x 5⅞ inches, over-all (cover closed).

Net Weight: 10⅞ pounds, including batteries.

Type		*Code Word*	*Price*
727-A	Vacuum-Tube Voltmeter..........	PIGMY	$115.00

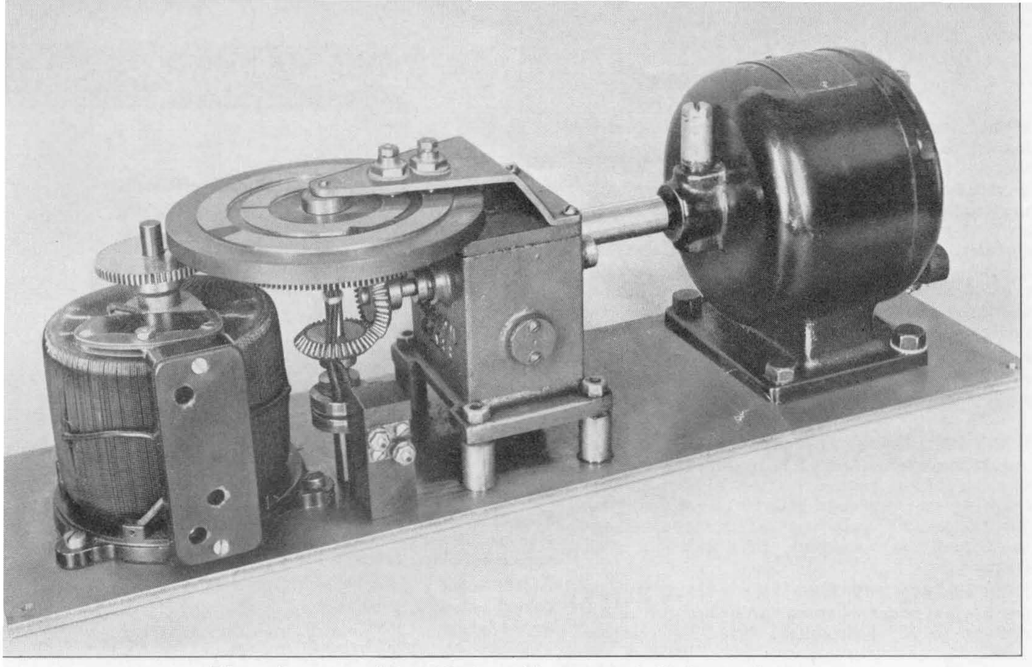
The Type 727-A Vacuum-Tube Voltmeter does not replace the Type 726-A. The two instruments are designed for different fields of application and both will be available. — Editor

DIELECTRIC STRENGTH TESTS WITH THE VARIAC

● **IN TESTING THE DIELECTRIC STRENGTH** (or "break-down strength") of electrical insulating materials, it has been found that the apparent strength varies greatly with the rate at which the test voltage is applied to the specimen. If it is desired to establish significant commercial specifications, it is obviously necessary to specify the rate and method of increase; otherwise no common basis for acceptance or rejection of materials can be had. Accordingly, one of the A.S.T.M. Standard Tests* for electrical insulating materials at standard power frequencies calls for a *uniform* increase of voltage until breakdown is reached. The actual rate of increase will depend on the type of insulation under test; for rubber-insulated cables, for instance, 3 kv per second is specified.

Of the various available means of varying the test voltage, a Variac in the primary of the high-voltage transformer is probably the most satisfactory and convenient. With the Variac a constant rate of rotation of the control wheel gives a constant rate of rise of voltage; furthermore, the voltage at any point is independent of the load, so that variations in the charging and leakage currents drawn by the specimen do not affect the voltage-vs.-time characteristic. All of the other commonly accepted methods of voltage control lack one or both of these features, making it vir-

*"Tests for Dielectric Strength (D149-40T)," *A.S.T.M. Standards on Electrical Insulating Materials*, December, 1941.



View of a motor-driven Variac with automatic limit attachment.

tually impossible to maintain the desired *uniform* increase of voltage. For these reasons the Variac has been widely used for voltage control in dielectric strength testing of all kinds.

Even with the Variac, however, a motor drive should be used, for tests have indicated that it is virtually impossible to rotate the Variac manually at a specified rate. The accompanying photograph, which is taken from the appendix to the specification mentioned above, shows a motor control arrangement incorporating automatic limit features.

A small direct-current motor (C) drives the Type 200-C Variac (A) through the reduction gears (B). The speed of the motor may be varied by means of a resistor in the armature circuit. For any given testing transformer, this resistor may be calibrated in terms of rate of voltage rise, thus facilitating the adjustment to the desired value. The reversing switch for the motor is so arranged that in the reverse position the armature resistor is out of circuit, and the Variac is returned to zero setting at maximum speed.

The motor-operating circuit is automatically opened at either end of the range of Variac rotation by means of segments mounted on the insulating disc.

The General Radio Company is not in a position to supply motor-driven Variacs at the present time. This article is published only to acquaint readers with an interesting use of the Variac.

BROADCAST EQUIPMENT

● **WE ARE VERY SORRY** to have to announce to our many customers in the broadcast engineering field that, owing to the increasingly rigid priority restrictions on both the buying and selling of raw materials and completed instruments, we have been compelled to restrict the sale of monitoring and measuring equipment to the broadcast industry very materially. This we do most reluctantly as we fully realize the importance of broadcasting to the war effort and to public morale, but the priority regulations are beyond our control.

Our policy on the sale of broadcast equipment is necessarily determined by these regulations. Instruments whose sole uses are in broadcasting will not be available after our present stock is exhausted, because the priority ratings available to most broadcasting stations are not sufficiently high to enable us to obtain materials for their manufacture. Broadcast instruments which are used also by the military services and for war production will still be manufactured, but they can be sold for broadcast use only in those instances where a high priority can be obtained.

Government policy about broadcast equipment is indicated by the following quotation from an FCC bulletin. It is by this policy, made after consideration of the over-all conditions, that we must all be governed.

"PRELIMINARY CURB ON BROADCAST ANTENNAS"

"At the request of the Defense Communications Board, pending the adoption of a specific policy by that Board and the War Productions Board with respect to curtailing standard broadcast construction to meet materials requirements by the military, the Federal Communications Commission will make no further grants for the construction of new standard broadcast stations or authorize changes in existing standard broadcast transmitting facilities where all or a substantial part of the primary area in either category already receives good primary coverage from one or more other stations.

"In general the Federal Communications Commission's Standards of Good Engineering Practice will be used as a guide in the determination of good primary service.

"National defense requires that there be adequate broadcast facilities, but this does not alter the fact that every economy in the use of critical materials for securing and maintaining these facilities must be practiced to the end that there will be the greatest possible saving in materials. Today's announcement concerns standard broadcast facilities only. It is understood that the Defense Communications Board is proceeding with studies looking toward the conservation of materials in all other radio services and will submit recommendations at the earliest practicable date."

HAVE YOU ANY IDLE INSTRUMENTS?

● **WHILE PRACTICALLY ALL GENERAL RADIO INSTRUMENTS** are urgently needed for war purposes, occasionally we are confronted with what might be termed a super-urgent need for a single instrument, the lack of which will delay the completion of a number of other projects. In these cases, we make every effort to speed up our own production, but this is not always possible. At times we have even borrowed an instrument from one customer to help out another temporarily.

If you have any current-model General Radio instruments in good operating condition that are not being used, you can help the war effort materially by letting us know about them. We can then refer prospective users directly to you. Most urgently needed are such items as standard-signal generators, oscillators, bridges, wave analyzers, noise meters, and other general-purpose instruments.

SERVICE AND MAINTENANCE NOTES — ERRATA

● **THE INEVITABLE SPRINKLING OF ERRORS** in the first printing of our Service and Maintenance Notes has been discovered, and those which have thus far been brought to our attention are listed below.

Please check your copy and make corrections if necessary.

Type 620-A Heterodyne Frequency Meter and Calibrator

Paragraph 4.2 should read as follows:

If the plate current milliammeter reads about 6.5 and 2.5 with the switch in the HET and CAL positions, respectively, both tubes are oscillating normally. Should this read higher, it would . . . .

Type 650-A Impedance Bridge

In the first mailing, pages 3 and 4 were missing. These will be sent on request.

Type 760-A Sound Analyzer

Paragraph 6.4, line 1; for R-28, read R-38.

Type 736-A Wave Analyzer

Paragraph 5.2, line 3; for V-7, read V-8.

Page 2, line 1; for 300, read 0.3.

MISCELLANY

● **A PAPER** entitled "Impedance Measurements from 1 to 100 Megacycles" was presented recently by R. F. Field at meetings of the Springfield (Mass.), Washington, and Detroit Sections of the Institute of Radio Engineers. In 1941, this paper was also presented at meetings of the Toronto, St. Paul, and Chicago Sections.

Another paper entitled "The Polarization Parameters of Several Solid Dielectrics and Their Changes with Temperature and Composition" was delivered by Mr. Field at the National Research Council Conference on Electrical Insulation at Williamsburg, Va., last October. This paper was also presented at two sectional meetings of the American Physical Society at Worcester, Mass., in March, 1942, and at Baltimore, Md., in May. A similar paper entitled "The Behavior of Dielectrics over Wide Ranges of Frequency and Temperature" was given by Mr. Field before the Boston Section of the I.R.E. on April 23, 1942.

We hope to publish these papers in forthcoming issues of the *Experimenter*.

● **WE STILL** have a supply of the cardboard Q-vs.-frequency templates mentioned in the article on iron-cored coils by McElroy and Field, which appeared in the March *Experimenter*. We shall be glad to send one to any reader who requests it.

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BRANCH ENGINEERING OFFICES
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THE General Radio EXPERIMENTER

VOLUME XVII No. 1 JUNE, 1942

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

Also IN THIS ISSUE

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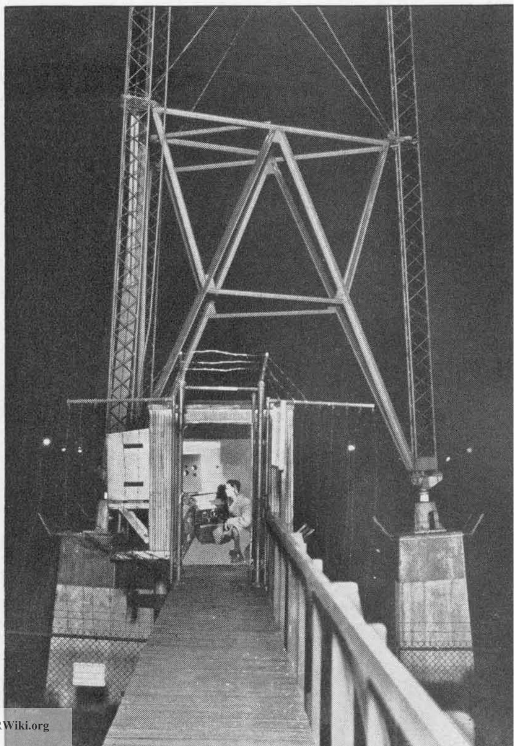
SERVICE AND MAINTENANCE NOTES...... 8

ANTENNA MEASUREMENTS WITH THE RADIO-FREQUENCY BRIDGE

● ALTHOUGH THE TYPE 516-C RADIO-FREQUENCY BRIDGE has been discontinued and will later be replaced by the TYPE 916-A, a considerable number of these bridges are in use in broadcasting stations, where they are quite satisfactory for measurements on antennas, lines, coupling networks, and other radio-frequency impedances, at standard broadcast frequencies. The instruction booklet supplied with this bridge covers quite completely the laboratory use of the bridge in measuring radio-frequency impedance, but it does not present the material in the most convenient manner for those who are interested solely in measuring antenna systems. It is the purpose of this article to supplement the operating instructions by outlining what has been found to be the most convenient procedure and by pointing out the precautions that must be observed if satisfactory results are to be obtained.

FIGURE 1. Measuring the impedance of an antenna tower as seen from the "dog house" at the base.



SETTING UP THE BRIDGE

Adequate shielding and grounding are important. Shielded conductors must be used for connecting the bridge to the generator and to the detector. The UNKNOWN binding post marked G must be grounded through as short a lead as possible, preferably not more than one or two feet. Grounding through the knurled panel screws is not satisfactory because the screws may not be in good electrical contact with the panel. To test for proper grounding, touch the panels of the bridge, detector, and generator after the bridge has been balanced. If the grounding is adequate, no effect upon the balance will be observed. If touching the bridge panel throws the bridge out of balance, the grounding is inadequate. This condition can sometimes be remedied by using individual ground leads for the bridge, the generator, and the receiver. A better remedy is to use coaxial terminals for the connection to the receiver input. General Radio Type 774-G Panel Plug and Type 774-M Cable Jack are satisfactory.

It is important that a well-shielded generator be used to prevent pickup not only from the generator to the detector, but also from the generator to the antenna under measurement. A standard-signal generator, such as the Type 605-B, is an excellent power source for these measurements.

Even at broadcast frequencies, it is essential to use a shielded receiver. The so-called communications type is recommended. The AVC, if any, should be disconnected, since its action tends to make the balance point difficult to locate.

BALANCING THE BRIDGE

For antenna measurements, the bridge is used as an equal-arm capacitance bridge, and so the balance point depends upon the adjustment of both the CAPACITANCE and RESISTANCE controls.

If one control is not set correctly, a *minimum* in the signal may be observed when the other is turned through its correct setting. Successive adjustments of both controls must be made until the signal in the detector is reduced to zero.

The balance point is frequently very sharp and may easily be missed if adjustments are made too rapidly. For a rough preliminary balance it may be desirable to use a modulated signal, with the receiver sensitivity turned well down. As balance is approached, the sensitivity can be increased. For a final balance, maximum accuracy, sensitivity, and signal-to-noise ratio are usually obtained by using an unmodulated signal, with a heterodyning oscillator to produce an audible beat tone.

METHODS OF MEASUREMENT

Although the Type 516-C is sufficiently flexible to permit measurements to be made by a wide variety of methods, it has been found by experience that one or more of the following three methods are most suitable for antenna measurements:

(1) Series Capacitor — The resistance range with this method is the same as with the direct method, but the capacitance range is greatly increased and, in addition, inductive reactance can be measured.

(2) Parallel Capacitor — The resistance range can be extended with this method but the actual range depends on the magnitude of the antenna reactance.

(3) Series and Parallel Capacitors — This is a combination of methods (1) and (2), and permits the measurement of reactance and resistance over wide ranges.

THE SERIES CAPACITOR METHOD

Of the above three methods, the series capacitor method is the one almost universally used for antennas whose resistive component does not exceed 311 ohms.[1] Since a large number of the antennas operating at standard broadcast frequencies fall into this category, this method is discussed first, with particular reference to operating procedure, and to reduction of errors caused by lead reactance.

Connections for this type of measurement are shown in Figure 2. A condenser, C_S, of such magnitude that the total series reactance presented to the UNKNOWN terminals is within the range of the balancing condenser, C_N, is connected in series with the unknown impedance. An initial balance is established with Z_a shorted; the short is then removed and the bridge rebalanced. Connections for the two balances are indicated in Figure 3.

The series resistance and reactance of the antenna are given by

$$R_a = R_2 - R_1 \qquad (1)$$[2]

$$X_a = \frac{1}{\omega} \frac{C_2 - C_1}{C_1 C_2} \qquad (2)$$

where the subscript 1 refers to the initial balance and the subscript 2 to the final balance. A positive value of X_a indicates an inductive reactance; a negative value, capacitive reactance.

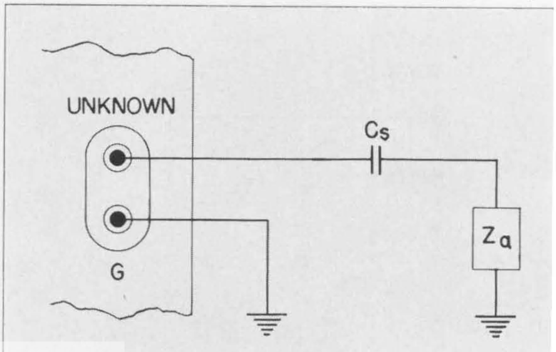
REACTANCE BALANCE

The best value of the capacitance, C_S, depends upon the reactive component of the antenna impedance. Since the value of the unknown reactance, X_a, is determined by the *difference* in capacitance settings between the two balances, fairly large errors can occur if this difference is small, and for maximum accuracy it is desirable to select a value of C_S to make $C_2 - C_1$ as large as possible. This is achieved by choosing the largest value of C_S for which the bridge will balance initially, if the antenna reactance is capacitive; and the largest value for which the bridge will balance with the antenna connected, if the antenna reactance is inductive. A good procedure to follow is to make a trial balance with a 1000 $\mu\mu$f series capacitor. If the unknown reactance is inductive, the setting of the CAPACITANCE dial will increase. Successive values of C_S then should be tried until the largest value is determined for which the CAPACITANCE control will balance with the unknown in circuit. If the unknown reactance corresponds to a capacitance greater than about 50 $\mu\mu$f, it will be possible to balance the CAPACITANCE control of the bridge, at a setting lower than the initial setting.

RESISTANCE BALANCE

In the discussion above it has been assumed that a balance can be obtained by adjustment of the RESISTANCE controls of the bridge. If the recommended fixed resistors are used, this

FIGURE 2. Diagram of connections for measuring an unknown impedance, Z_a, by using the series capacitor C_S.



[1] The RESISTANCE control of the bridge has a direct-reading range from 0 to 111 ohms, but provision is made for inserting external fixed resistors in series. The range can be increased to 311 ohms by inserting 200 ohms in series. If a resistance larger than 200 ohms is placed in series, the shunting effect of the ground capacitance of the condenser, C_N, introduces an appreciable error, particularly at frequencies above one megacycle

The TYPE 500-D (100 Ω) and the TYPE 500-E (200 Ω) are recommended for use as series resistors.

[2] R_1 can be made zero by establishing the initial balance with the POWER FACTOR and POWER FACTOR ADJUST controls.

will be true when the unknown resistance does not exceed 311 ohms. A balance should be attempted with the SERIES RESISTOR terminals short-circuited by means of the strap that is provided. If the RESISTANCE controls cannot be balanced, the 100-ohm, or, if necessary, the 200-ohm, resistor should be plugged in. Failure to obtain a balance at 311 ohms or below indicates that the parallel-capacitance method or a series-parallel connection must be used.

LEAD CORRECTIONS

The leads from the bridge to the unknown should be as short as possible, certainly less than three feet, and should be kept a reasonable distance away from grounded metal objects, in order to minimize capacitance to ground. The series capacitor should be connected at the antenna end of the lead, as close as possible to the point at which the antenna impedance is to be measured. By this method of connection, the inductance and resistance of the lead remain in series with C_S for the initial as well as the final balance, and drop out of the calculation for the antenna impedance.

The lead *capacitance* introduces small errors into the calculated values of resistance and reactance. The lead capacitance can be easily measured, however, and allowance made for it.

The lead capacitance, C_l, is determined by a substitution method, wherein initial balance is established with a fixed capacitor connected directly across the UNKNOWN terminals of the bridge. The lead is then connected to the ungrounded terminal of the bridge. The series capacitor is disconnected from the antenna but is still connected to the far end of the lead. A new balance is made, and the difference of the two capacitance readings is C_l which includes the capacitances to ground of the series capacitor, as well as that of the lead itself. The correction for lead capacitance is made by using the following expressions:

$$R_a = R_2\left(1 + \frac{C_l}{C_2}\right)^2 \quad (3)$$

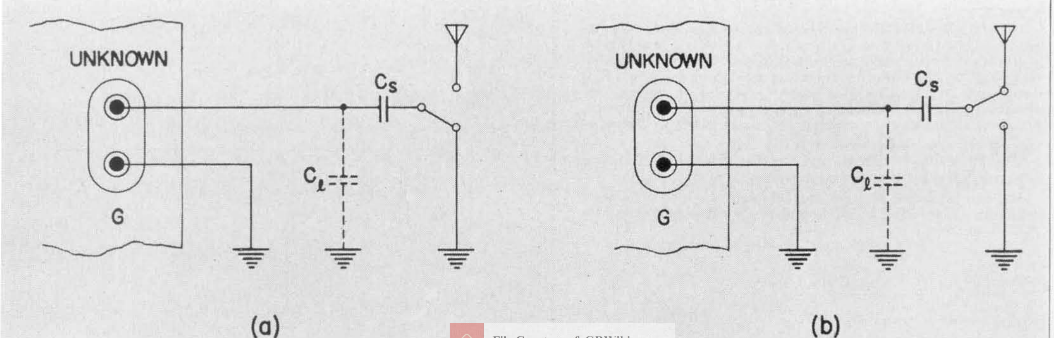
$$X_a = \frac{C_2 - C_1}{\omega C_1 C_2}\left(1 + \frac{C_l}{C_1}\right)\left(1 + \frac{C_l}{C_2}\right) \quad (4)$$

NUMERICAL EXAMPLE

The following data illustrate the procedure and data for a typical antenna measurement at a frequency of 1500 kilocycles.

To determine the most suitable value of C_S, a balance with $C_S = 1000$ $\mu\mu$f was attempted. One side of the capacitor was connected to the ungrounded UNKNOWN terminal through a two-foot length of insulated wire, and the other side connected to the antenna input through a few inches of lead. With a modulated signal from the generator, and the volume control of the receiver set to a low level, it was impossible to obtain a minimum with either bridge control. A slight decrease in intensity was observed when both controls were turned to maximum, however, indicating that the antenna reactance was inductive, and that R_x was greater than 111 ohms. Consequently, the 1000 $\mu\mu$f series capacitor was replaced by a 500 $\mu\mu$f unit, and a 100 Ω

FIGURE 3. Diagram of connections for the two measurements necessary to determine antenna impedance by the series capacitor method. Additional measurements, described in the text, are necessary to determine the connection for C_l, the lead capacitance.



resistor was inserted at the series resistance terminals. A definite minimum was now observed with the CAPACITANCE control of the bridge set at about 650 $\mu\mu$f and the RESISTANCE control set at a maximum. Accordingly, the 100 Ω external resistor was replaced by a 200 Ω unit. A true null was now obtained at $R \cong 200 + 39\ \Omega$, $C \cong 648\ \mu\mu$f. The CAPACITANCE dial setting suggested that a somewhat larger series capacitor might be used, and the 100 $\mu\mu$f unit was plugged in parallel with the 500 $\mu\mu$f already in circuit. This was tried and the CAPACITANCE control balanced at a setting over 900 $\mu\mu$f. With the proper value of series capacitor and SERIES RESISTOR determined, the data for calculation were obtained as follows:

(1) With the series capacitor grounded at the antenna end, the SERIES RESISTOR terminals shorted, and the RESISTANCE control set to zero, a balance was obtained by means of the POWER FACTOR ADJUST and CAPACITANCE controls. The balance occurred with the large CAPACITANCE dial set at 620, and the auxiliary dial set at −9.4. Thus

$$C_1 = 620 - 9.4 = 610.6\ \mu\mu\text{f}$$

(2) With the series capacitor connected to the antenna, and the 200 Ω resistor connected to the SERIES RESISTOR terminals, the bridge was balanced (leaving the POWER FACTOR controls untouched) and the following data obtained:

$$R_2 = 200.0 + 38.8 = 238.8\ \text{ohms}$$

$$C_2 = 930 + 3.3 = 933.3\ \mu\mu\text{f}$$

(3) A computation using the above data, and neglecting the lead capacitance, yields:

$$R_a = R_2 = 238.8\ \text{ohms}$$

$$X_a = \frac{1}{\omega}\frac{C_2 - C_1}{C_1 C_2}$$

$$= \frac{(933.3 - 610.6) \times 10^{-12}}{6.28 \times 1.5 \times 10^{6} \times 933.3 \times 610.6 \times 10^{-24}}$$

$$= +60.0\ \text{ohms (inductive)}$$

$$Z_a = 238.8 + j60$$

(4) The value of C_l, the lead capacitance, was determined by connecting the 1000 $\mu\mu$f capacitor across the UNKNOWN terminals, disconnecting the lead at the bridge, replacing the link across the SERIES RESISTOR terminals, and balancing the bridge. The CAPACITANCE controls balanced at 990 plus 8.8 $\mu\mu$f, but were rebalanced at 1000 minus 1.2 $\mu\mu$f, thus permitting the change in capacitance to be observed entirely on the AUXILIARY dial. With the lead connected to the high UNKNOWN terminal and the series capacitor disconnected at the antenna side, the balance was obtained with the auxiliary CAPACITANCE control set at plus 5.6 $\mu\mu$f. The value of C_l was thus $5.6 - (-1.2) = 6.8\ \mu\mu$f.

(5) The corrected values of R_x and X_x are determined by substituting the above values in Equations (3) and (4).

$$R_a = R_2\left(1 + \frac{6.8}{933.3}\right)^2$$

$$= (238.8)\ (1.0146)$$

$$= 242.3\ \text{ohms}$$

$$X_a = (60.0)\left(1 + \frac{6.8}{933.3}\right)\left(1 + \frac{6.8}{610.6}\right)$$

$$= (60.0)\ (1.0184)$$

$$= 61.1\ \text{ohms}$$

$$Z_a = 242.3 + j61.1$$

In this particular case, neglect of the lead capacitance causes an error of less than 2% in the calculation of either component.

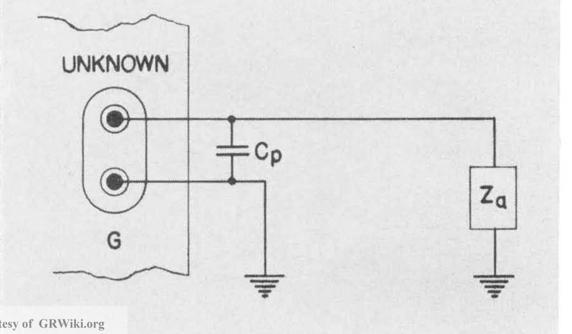
The above procedure may seem rather lengthy, but it should be borne in mind that, once a measurement has been made, additional measurements to study the effect of minor antenna adjustments, or of changes in frequency, can be made quite rapidly.

THE PARALLEL CAPACITOR METHOD

When the series resistance of the antenna exceeds 311 ohms, the parallel capacitor method should be tried. The technique of this type of measurement is quite similar to that for the series substitution method, already described. The computations, however, are somewhat more involved because of the transformation from parallel to series components that is required.

The parallel capacitor method consists essentially of connecting, in parallel with the unknown impedance, a capacitance of such magnitude that the impedance of the combination lies within the range of the bridge. Connections are shown schematically in Figure 4. The initial balance is made with the parallel capaci-

FIGURE 4. Connections for the parallel-capacitor method of measurement.



tor connected to the bridge, with the lead connected to the bridge but disconnected at the antenna end, and with the RESISTANCE decades set to zero. The lead is then connected to the antenna, and the bridge is rebalanced. The following equations give the *series* resistance and reactance of the antenna:

$$R_a = \frac{R_2}{(\omega C_1)^2} \frac{1}{R_2^2 + \left(\frac{1}{\omega} \frac{C_2 - C_1}{C_1 C_2}\right)^2} \qquad (5)$$

$$X_a = \frac{1}{\omega C_1} \frac{R_2^2 - \frac{1}{\omega C_2}\left(\frac{1}{\omega} \frac{C_2 - C_1}{C_1 C_2}\right)}{R_2^2 + \left(\frac{1}{\omega} \frac{C_2 - C_1}{C_1 C_2}\right)^2} \qquad (6)$$

where C_1 is the reading of the CAPACITANCE dial for the initial balance, and C_2 and R_2 are the CAPACITANCE and RESISTANCE setting for the second balance. The fact that the same terms appear in both expressions simplifies the calculations.

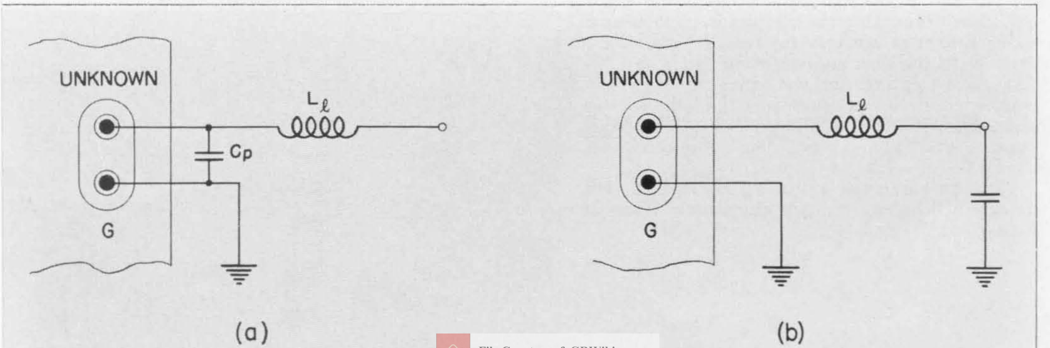
CHOICE OF PARALLEL CAPACITOR

The best value of parallel capacitance depends on the magnitude of the antenna impedance and can thus be determined only by trial. In general, the best value to use is the smallest capacitance with which a final CAPACITANCE balance can be obtained. Occasionally, however, it may be desirable to use a larger value, to permit the RESISTANCE controls to be balanced without the use of external SERIES RESISTORS. Probably the most rapid way to determine the proper value of C_P is to make a first trial balance using the 500 $\mu\mu$f parallel capacitor, and leaving the SERIES RESISTOR terminals short-circuited. If the change in capacitance is small, a smaller value of capacitance should be tried; if the resistance component of balance cannot be obtained with the internal resistance decades, an external series resistor should be added, or a larger value of parallel capacitance tried. If it is impossible to obtain a balance using the largest available capacitor (1000 $\mu\mu$f) and the largest permissible external resistor (200 ohms), the use of one of the other methods is indicated.

LEAD CORRECTIONS

Unlike the series capacitor method, the parallel capacitor method yields the most satisfactory results when the auxiliary capacitor is connected directly to the UNKNOWN terminals of the bridge. The lead to the antenna terminals should be left connected at the bridge, and the initial balance made with this lead disconnected at the antenna terminal. The lead capacitance thus remains always in parallel with the auxiliary capacitance, is measured as part of it, and introduces no error into the

FIGURE 5. Diagram of connections for the two measurements to determine the lead inductance L_l. Procedure and calculatious are outlined in the text.



measurement of the antenna impedance. The lead *inductance*, however, is included in the calculated value of antenna reactance. The true value of X_a is determined simply by subtracting the lead reactance from the value given by Equation (6).

The lead inductance (L_l) can be measured as shown in the two diagrams of Figure 5. Balance the bridge with the 1000 $\mu\mu$f capacitor connected to the UNKNOWN terminals, and with the lead to the antenna in place, but disconnected at the antenna terminals. Remove the 1000 $\mu\mu$f capacitor, connect it to ground at the far end of the lead in such a manner that the position of the lead remains unchanged, and rebalance the bridge. If the first and second CAPACITANCE readings are denoted by C' and C'', respectively, the inductive reactance of the lead is given by

$$X_l = \omega L_l = \frac{1}{\omega}\frac{C'' - C'}{C'C''} \qquad (7)$$

NUMERICAL EXAMPLE

The following procedure and data are representative of measurements of an antenna having a resistance greater than 311 ohms, at a frequency of 2000 kilocycles.

By following the general procedure outlined in detail in the numerical example for the series capacitor method, 300 $\mu\mu$f was determined to be a suitable value of parallel capacitance.

(1) The initial balance was established with 300 $\mu\mu$f plugged into the unknown terminals, and with the lead to the antenna in place but disconnected at the antenna end. The POWER FACTOR and RESISTANCE controls were set at zero, and the balance obtained by adjustment of the CAPACITANCE dials and the POWER FACTOR ADJUST knob. The balance occurred with the large CAPACITANCE dial set at 300 $\mu\mu$f, and the auxiliary dial set at +8.8 $\mu\mu$f.

(2) With the lead connected to the antenna the new positions of the main and auxiliary dials were 510 and −3.2 $\mu\mu$f respectively, while the setting of the RESISTANCE decades was found to be 98.5 ohms.

(3) The data and calculations were as follows:

$$C_1 = 300 + 8.8 = 308.8\ \mu\mu f$$
$$\omega = 2\pi f = 6.28 \times 2 \times 10^6$$
$$= 12.56 \times 10$$

$$\frac{1}{\omega C_1} = \frac{1}{12.56 \times 10^6 \times 308.8 \times 10^{-12}}$$
$$= 258 \text{ ohms}$$
$$\left(\frac{1}{\omega C_1}\right)^2 = 66{,}600$$
$$C_2 = 510 - 3.2 = 506.8\ \mu\mu f$$
$$\frac{1}{\omega}\frac{C_2 - C_1}{C_1 C_2} = 258 \times \frac{506.8 - 308.8}{506.8}$$
$$= 100.8 \text{ ohms}$$
$$\left(\frac{1}{\omega}\frac{C_2 - C_1}{C_1 C_2}\right)^2 = 10{,}160$$
$$\frac{1}{\omega C_2} = \frac{1}{12.56 \times 10^6 \times 506.8 \times 10^{-12}}$$
$$= 157 \text{ ohms}$$
$$\left(\frac{1}{\omega C_2}\right)^2 = 24{,}700$$
$$R_2 = 98.5 \text{ ohms}$$
$$R_2^2 = 9{,}700$$

Substituting in Equations (5) and (6)

$$R_a = \frac{98.5 \times 66{,}600}{9{,}700 + 10{,}160} = 331 \text{ ohms}$$
$$X_a = 258 \times \frac{9{,}700 - 157 \times 100.8}{9{,}700 + 10{,}160}$$
$$= -80.0 \text{ ohms (capacitive)}$$
$$Z_a = 331 - j80.0 \text{ ohms}$$

(4) The inductive reactance of the lead was determined by the method outlined under LEAD CORRECTIONS above. The CAPACITANCE readings C' and C'' were 510 + 1.2 $\mu\mu$f and 540 − 4.2 $\mu\mu$f. From Equation (7)

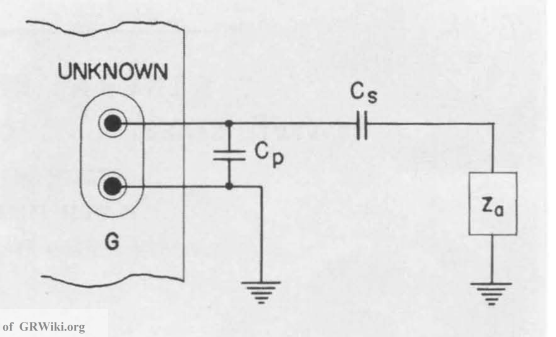
$$X_L = \frac{1}{\omega}\frac{C'' - C'}{C''C'}$$
$$= \frac{535.8 - 511.2}{12.56 \times 10^6 \times 535.8 \times 511.2 \times 10^{-12}}$$
$$= +7.15 \text{ ohms}$$

The corrected value of antenna reactance is thus

$$X_a = -80.0 - 7.15 = -87.2 \text{ ohms}$$
$$Z_a = 331 - j87.2 \text{ ohms}$$

In this case, neglect of the lead reactance would have resulted in an error of nearly 10% in the calculated antenna reactance.

FIGURE 6. For the occasional measurement where neither the series capacitor nor parallel capacitor method is satisfactory, a combination of the two can be used as shown here.



SERIES AND PARALLEL CAPACITOR METHOD

The methods described above generally suffice to measure antenna impedance over the ranges normally met in practice, but occasionally extreme values of reactance, or certain combinations of resistance and reactance, are encountered that do not permit a balance with either a series or a parallel capacitor alone. In such cases the use of both series and parallel capacitors, as shown in Figure 6, usually permits a measurement to be made.

USE OF CHARTS

When approximate results are desired, it is convenient to use a chart to determine the quantity $\frac{1}{\omega}\frac{C_2 - C_1}{C_1C_2}$. Copies of this chart can be obtained from the General Radio Company. Another convenient chart is a log-log plot of reactance vs. capacitance for some nominal frequency such as 1 megacycle. Conversion to other frequencies can be made by dividing the 1-Mc reactance by the frequency in megacycles.

SERVICE AND MAINTENANCE NOTES

CORRECTIONS

● **ADDITIONAL ERRORS AS LISTED BELOW** have been discovered in the first printing of Service and Maintenance Notes. Please check your copies and make corrections as necessary.

Type 544-B Megohm Bridge

Page 3: Paragraph 7.1, read 6K6G for 6K5G; paragraph 7.2, read 6J5G for 6V5G.

Page 4: Paragraph 9.1, read 10.0 and 11.0 for 3.0 and 4.0; paragraph 11.1, read 10.3 for 1.0.

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.*

GENERAL RADIO COMPANY

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BRANCH ENGINEERING OFFICES

90 WEST STREET, NEW YORK CITY

1000 NORTH SEWARD STREET, LOS ANGELES, CALIFORNIA

THE General Radio EXPERIMENTER

VOLUME XVII No. 2 JULY, 1942

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

Also

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RECENT PRIORITY ORDERS OF INTEREST TO BUYERS OF GR EQUIPMENT . . . 5

BRINGING THE BEAT-FREQUENCY OSCILLATOR UP TO DATE

● **OF THE THREE COMMON TYPES OF AUDIO-FREQUENCY OSCILLATOR** — the tuned circuit, the beat-frequency, and the degenerative or resistance-tuned—each has certain inherent advantages which make it preferable to other types for certain applications, while each has also certain disadvantages which must be eliminated as far as possible in a practical design.

The beat-frequency oscillator, in particular, is characterized by the relative simplicity with which the frequency can be changed over a wide range by varying a single control. Furthermore, with reasonable design precautions the output of such an oscillator can be made substantially constant as the frequency is varied. These two characteristics make the beat-frequency oscillator well suited for measuring the frequency response of amplifiers, filters, and other communication networks, and, for this application, it has almost completely superseded the older tuned-circuit type.

On the other hand, the beat-frequency oscillator has certain inherent disadvantages which must be overcome in any satisfactory design. In the first place, since the output frequency is obtained by heterodyning

FIGURE 1. Panel view of the TYPE 913-A Beat-Frequency Oscillator.

two higher-frequency oscillators, any given percentage drift in one of these oscillators with respect to the other will cause a much higher percentage drift in the output frequency. For instance, assuming that the high-frequency oscillators are operating around 100,000 cycles and one of them drifts 0.1%, or 100 cycles, the drift is then equivalent to 100% for an output frequency of 100 cycles. This was not an uncommon occurrence in earlier types of beat-frequency oscillators during the warming-up period.

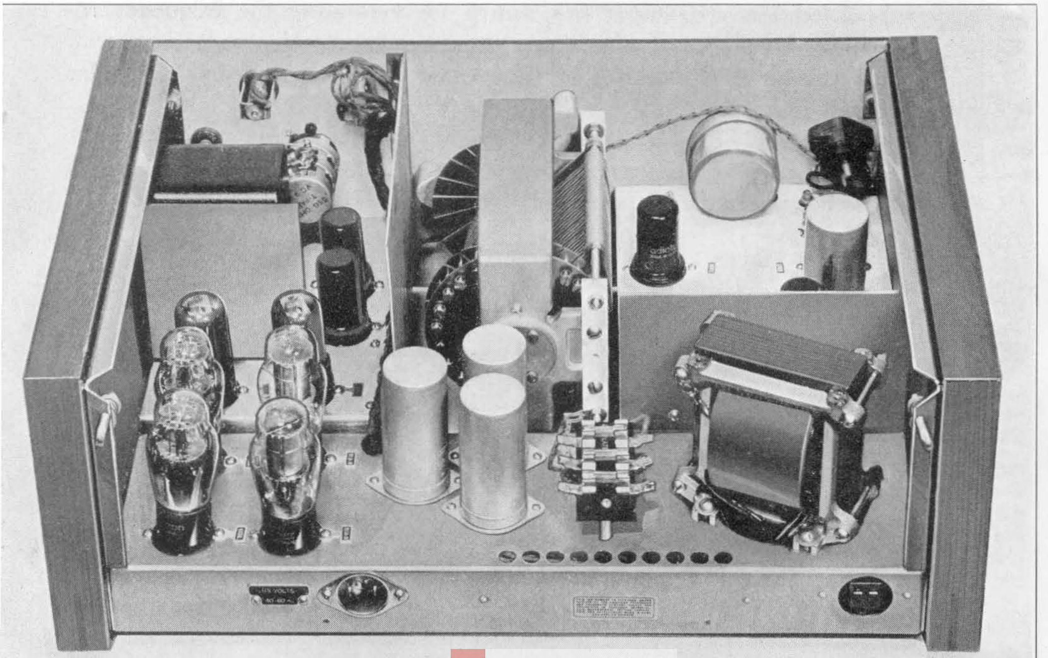
The second troublesome characteristic of the beat-frequency circuit is the "pulling-in" effect. As the beat frequency is lowered, any direct coupling between the two high-frequency oscillators becomes more serious, since one oscillator tends to pull the other into step with it. At frequencies above that at which the oscillators actually lock in together, this pulling in causes a serious distortion in the beat waveform.

A third disadvantage of the beat-frequency oscillator is that the beat frequency is generally obtained at a fairly low level and must be amplified. If the oscillator is to compete with other low-distortion types, such as, for instance, the degenerative, the design of the amplifier itself is an important problem.

In older designs these factors seriously limited the performance of the oscillator, but the tools available to the modern designer, such as improved circuit elements, new mixer tubes, and low-distortion amplifier circuits, have made it possible to eliminate them in the design of the new TYPE 913-A Beat-Frequency Oscillator.

This instrument was developed particularly with the idea of providing the best-balanced design possible at the present time, and in operating characteristics it bears little resemblance to the ordinary type of beat-frequency oscillator. Among its important features are convenience and ease of frequency control, constant output voltage, a high de-

FIGURE 2. Rear view of the oscillator with dust cover removed. Note that tubes and fuses are easily accessible.



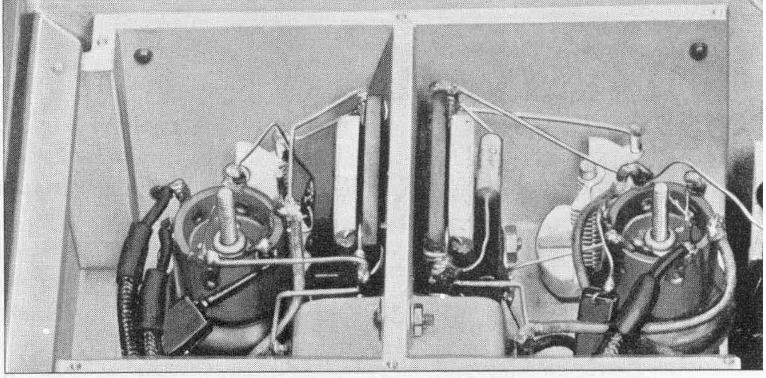


FIGURE 3. View of the cast-metal oscillator housing with cover removed to show the two oscillator circuits. The use of duplicate parts and the symmetry of construction about the center shield help to equalize the temperature coefficients of the two oscillators.

gree of stability, and low distortion. It is well adapted for running frequency characteristics (with a recorder, if desired) and also for all normal types of distortion measurements.

Since price and size are also of importance in instrument design, no attempt was made to utilize temperature control or other awkward, expensive systems for maintaining stability. Instead, the circuit elements themselves were designed to have minimum temperature coefficients. This necessitated first a new design for the oscillator coils. In these coils the dust core is supported by a continuous threaded rod, which is supported in the same manner at each end of the coil form. Hence any differential expansion between the coil form itself and the rod produces equal and opposite forces, so that the relative position of the core does not shift.

The fixed condensers are of the silver-mica type and are mounted directly on opposite sides of a shield separating the oscillator compartments. The two condensers are thus maintained at substantially the same temperature. The oscillator circuits are contained in adjacent sections of a cast-metal shield, so that their temperatures remain approximately the same. This is important, since it is relative drift between the two oscillators, and not the actual drift of either one, which changes the beat frequency.

Finally, in order to reduce so far as possible any remaining drift, each oscillator is checked in the calibration laboratory throughout a warming-up cycle, and a temperature-compensating condenser is added to the circuit to cancel out so far as possible any residual temperature drift. The result is that the TYPE 913-A Oscillator will, on the average, drift only a few cycles between the time it is turned on and the time it has reached equilibrium, when operating at normal ambient room temperature. For most purposes the drift is so slight that no readjustment of the calibration whatsoever is required. In stability, therefore, the TYPE 913-A Oscillator is second among commercially available types only to the best of degenerative oscillators* and for most purposes the stability of this beat-frequency oscillator is more than adequate.

In the past beat-frequency oscillator designs have included buffer amplifiers between the oscillator and the mixer or modulator, in an attempt to prevent pulling in. In the TYPE 913-A use has been made of one of the newer pentagrid converter tubes, which, when combined with suitable grid-bias circuits to provide the proper square-law characteristic, gives substantially distortionless heterodyning action and at the same time good isolation between the oscillator circuits. The circuit used in the TYPE 913-A is a considerable simplification over previous low-distortion beat

*Low-priced degenerative (resistance-tuned) oscillators, while useful for many purposes, do not, in general, have either the stability or low distortion which might be expected in view of the type of circuit used and which is characteristic of more expensive types.

oscillators, and yet provides substantially better performance.

To provide the required output voltage and power a highly degenerative amplifier is used. It includes a vacuum-tube phase inverter and a push-pull output stage, feeding a high-quality output transformer. Distortion in the amplifier is considerably less than 0.05% throughout most of the frequency range. The output impedance is 550 ohms. This value was chosen so that the oscillator could be used equally well with 500-ohm and 600-ohm equipment.

Other circuit refinements include plate voltage regulation, a constant-impedance volume control calibrated directly in output power in terms of db with respect to the standard reference level of 1 milliwatt into 600 ohms, and a simplified zero-beat indicator consisting essentially of a small gas-filled tube. The instrument is also equipped with a gear-drive dial, which can be connected readily to recording equipment when so required.

—H. H. SCOTT

SPECIFICATIONS

Frequency Range: 20 to 20,000 cycles.

Frequency Control: The main control is engraved from 20 to 20,000 cycles per second and has a true logarithmic frequency scale. The total scale length is approximately 12 inches. The effective angle of rotation is 240°, or 80° per decade of frequency.

Frequency Calibration: The calibration can be standardized within 1 cycle at any time by setting the instrument to zero beat. The calibration of the frequency control dial can be relied upon within ±2%±1 cycle after the oscillator has been correctly set to zero beat.

Zero Beat Indicator: A neon lamp is used to indicate zero beat.

Frequency Stability: Improved design of the oscillator circuits and the use of temperature-compensated capacitors and inductances result in an unusually high degree of stability.

Output Impedance: The output impedance is 550 ohms, either grounded or balanced-to-ground, and is essentially constant regardless of the output control setting. With load impedances of 2000 ohms or less, the output is balanced for all settings of the output control. With higher load impedances, unbalance may occur at low settings of the output control.

Output Power: 0.3 watt maximum.

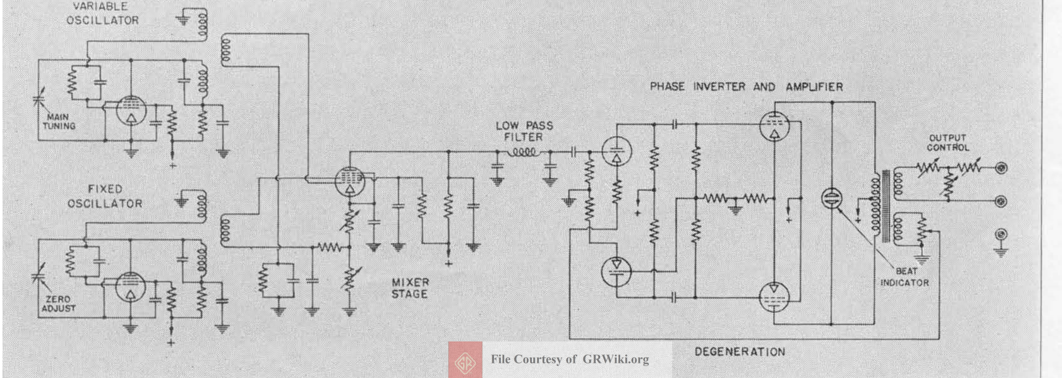
Output Voltage: Approximately 25 volts open circuit. For a matched resistive load the output voltage varies by less than ±0.25 db between 20 and 20,000 cycles.

Output Control: The output control is calibrated in db referred to 1 milliwatt into 600 ohms. The total range is from +25 to −25 db.

Waveform: When the oscillator is operating into a matched load or a load of higher impedance, the total harmonic content is approximately 0.2% from 150 to 7000 cycles. Below 150 cycles the harmonic content increases slightly, reaching about 2% at 50 cycles. With the volume control turned fully on, the harmonic content is approximately doubled when the oscillator is operated into an extremely low impedance. If, however, the volume control is turned 3 db or more below the maximum setting, the load impedance has no effect upon the waveform.

A-C Hum: The a-c hum is less than 0.05% of the output voltage at a line frequency of 60

FIGURE 4. Schematic circuit diagram of the TYPE 913-A Beat-Frequency Oscillator.



cycles, and is less than 0.1% at 42 cycles. Since the volume control is in the output circuit, the hum percentage does not increase for low output voltages.

Temperature and Humidity Effects: Large changes in ambient temperature and humidity necessitate a readjustment of the zero-beat setting. High temperatures and humidity cause a slight increase in distortion and a slight decrease in output.

Terminals: Jack-top binding posts with standard ¾-inch spacing are provided on the panel. A Jones socket and plug provide duplicate output terminals on the back of the instrument for relay-rack installation.

Mounting: The panel is designed for mounting on a 19-inch relay rack, but removable wooden ends are supplied so that it may be used equally well on a table.

Power Supply: 105 to 125 volts, 40 to 60 cycles ac. A simple change in the connections to the power transformer allows the instrument to be used on 210 to 250 volts. The total consumption is about 100 watts. Since the oscillator circuits are equipped with voltage regulators, the change in output with power-supply voltage is negligible.

Tubes:

2 — type 6SK7 2 — type 6V6-GT
1 — type 6SA7 2 — type 6X5-G
2 — type 6SF5 2 — type VR-150-30
1 — 139-949 Neon Lamp

All are supplied with the instrument.

Accessories Supplied: A seven-foot connecting cord, a multipoint connector, and spare fuses and pilot lamp are supplied.

Dimensions: 19⅜ x 14¼ x 7½ inches, over-all.

Net Weight: 35 pounds.

Type		*Code Word*	*Price*
913-A	Beat-Frequency Oscillator	CAROL	$260.00

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.

RECENT PRIORITY ORDERS OF INTEREST TO BUYERS OF GR EQUIPMENT

QUARTZ CRYSTALS

● **ON MAY 18** the Director of Industry Operations of the War Production Board issued General Conservation Order No. M-146 because of the shortage of quartz crystals which has been brought about by increasing demands for crystals and other forms of quartz in connection with the National Defense. Paragraph (g) of this order, given below, indicates the restrictions which have been placed on the sale and use of quartz by this Conservation Order and contains the certification which must be furnished to us before delivery of equipment containing manufactured forms of quartz crystals can be made.

Section 1218.1 General Conservation Order No. M-146:

"(g) *Restrictions on Use.* After the effective date of this Order, except as specifically authorized by the Director of Industry Operations, no Person shall consume or process Quartz Crystals except in the manufacture of:

(1) Radio oscillators and filters or other products for use in Implements of War,

(2) Radio oscillators and filters for use in radio systems to be owned, used, and operated by Federal Agencies, or by commercial airlines,

(3) Telephone resonators,

(4) Optical parts for use in Implements of War;

and no Person shall purchase or accept delivery of manufactured forms of Quartz Crystals except for use for purposes for which manufacture thereof is permitted under the foregoing provisions of this paragraph (g). Any Person who consumes or processes Quartz Crystals as aforesaid, shall require before the manufactured forms of Quartz Crystals leave his possession that the purchaser or transferee thereof make and deliver to him, or endorse on the purchase order a certificate, manually signed by the purchaser or transferee or a responsible official thereof, in substantially the following form, to wit:

The undersigned hereby certifies that he is familiar with the terms of Conservation Order M-146; and that the manufactured forms of Quartz Crystals covered by the accompanying Order of even date shall be used only for purposes permitted by the terms of said Order M-146.

Dated ....

Name ....

By ....

Such certificate shall constitute a representation by the purchaser or transferee to the consumer or processor and to the War Production Board of the facts stated therein. The consumer or processor of Quartz Crystals shall be entitled to rely on such representation unless he knows or has reason to believe it to be false."

Several instruments which General Radio manufactures, such as the TYPE 736-A Wave Analyzer, TYPE 620-A Heterodyne Frequency Meter, CLASS C-21-H and CLASS C-10 Frequency Standards, and other frequency monitors, use crystals as filters or oscillators. Accordingly it is necessary for our customers ordering these instruments to comply with the restrictions of M-146 and to furnish us with a signed certification before shipment can be scheduled.

END USE

Priorities Regulation No. 10 established an Allocation Classification System which has been designed to provide a means of identifying the ultimate uses and users of various products and materials as well as a means of transmitting such identifications down through industry to original suppliers. With the information so obtained it will be possible for the War Production Board to allocate materials more wisely and to reduce the rules and forms which now must be used for allocation purposes. *The Regulation provides that all purchase orders or contracts, other than retail, placed after June 30, 1942, must have indicated on them the appropriate Allocation Classification Symbol and Purchaser's Symbol.* Furthermore, all orders, regardless of when placed, which call for delivery after July 31 must also carry these symbols. Customers who have already placed orders for delivery after July 31 must, before that date, notify the supplier of the appropriate symbols applying on such orders.

For the convenience of our customers we are listing below the various Allocation Classification Symbols, with a short description of material covered.

ALLOCATION CLASSIFICATION

(NOTE: The symbol numbers have no relation to order of importance.)

Allocation Symbol — MILITARY

1.00 CLASS 1.00 — AIRCRAFT — PRODUCTION AND MAINTENANCE (complete except for armament and ammunition — as approved by the Joint Aircraft Committee)

CLASS 2.00 — SHIPS, PRODUCTION AND MAINTENANCE (complete except for armament and ammunition)

2.10 Battleships
2.20 Aircraft carriers
2.31 Escort vessels (aircraft), combat, loaded transports, and combat loaded cargo ships
2.32 Patrol vessels
2.33 Landing craft including the following types: APM, ATL, YTL, tank lighters, artillery lighters, landing boats, support landing boats
2.40 Light cruisers
2.50 Destroyers including escort vessels
2.60 Submarines
2.70 All other types of naval craft
2.80 Repairs to all naval vessels
2.90 Ships for Maritime Commission

CLASS 3.00 — VEHICLES — PRODUCTION AND MAINTENANCE (complete except for armament and ammunition)

3.10 Tanks and armored vehicles — all types
3.20 Vehicles, except rail — all other military types

CLASS 4.00 — ARMAMENT AND WEAPONS — PRODUCTION AND MAINTENANCE (complete mounts and related equipment)

4.10 Aircraft
4.20 Anti-aircraft, barrage balloon equipment, A. A. searchlights
4.30 Artillery including railway and seacoast
4.40 Fire control, all types
4.50 Machine guns — ground, hand arms
4.60 Naval, all types
4.70 Tanks and anti-tank
4.90 Weapons of all other types

CLASS 5.00 — AMMUNITION — PRODUCTION AND MAINTENANCE (complete items)

5.10 Ammunition 20 mm. and above
5.20 Ammunition, small arms below 20 mm.
5.30 Bombs, depth charges, mines, and torpedoes
5.40 Propellants, chemicals, explosives
5.50 Pyrotechnics

CLASS 6.00 — WAR EQUIPMENT AND SUPPLIES — PRODUCTION AND MAINTENANCE (complete with related equipment)

6.10 Chemical warfare equipment and supplies
6.20 Clothing, general supplies, and subsistence
6.30 Mapping, map reproduction, and photographic equipment
6.40 Medical equipment and supplies
6.50 Military field construction equipment
6.60 Military radio and wire communications, and Radar or electronic equipment — all types
6.70 Military railway including rail vehicles and bridge equipment
6.80 Supplies and equipment — all other military types
6.90 Supplies and equipment — all other

CLASS 7.00 — WAR FACILITIES — CONSTRUCTION AND/OR MAINTENANCE

7.10 Air fields, bases, camps, coast defense, depots, forts, navy yards, posts, stations — Continental U. S. A.
7.20 Air fields, bases, camps, coast defense, depots, forts, navy yards, posts, stations — outside Continental U. S. A.
7.30 Munitions manufacturing facilities and proving grounds — government owned
7.40 Panama Canal
7.50 Shipyards and ship repair facilities — government owned

INDUSTRIAL AND CIVILIAN

CLASS 8.00 — RAW MATERIALS, PRODUCTION, AND PROCESSING OF

8.10 All metals, production (including mining), smelting, and processing of
8.20 All chemicals, production, and processing of
8.90 All other raw materials, production, and processing of

CLASS 9.00 — POWER, LIGHT, AND HEAT

9.10 Electricity
9.20 Petroleum
9.30 Coal and coke
9.40 Gas

CLASS 10.00 — TRANSPORTATION

10.10 Railroad including urban and interurban
10.20 Automotive
10.30 Roads, streets, etc., construction and maintenance of
10.40 Water transportation, including construction of privately owned shipyards
10.50 Air transportation
10.90 All other transportation

CLASS 11.00 — COMMUNICATION

11.10 Telephone
11.20 Radio
11.30 Telegraph
11.90 All other communication

CLASS 12.00 — PUBLIC HEALTH AND SAFETY

12.10 Sanitary and health systems and facilities
12.20 Health equipment and supplies including personal care
12.30 Public safety equipment and supplies

13.00 CLASS 13.00 — AGRICULTURAL EQUIPMENT AND SUPPLIES

14.00 CLASS 14.00 — INDUSTRIAL FOOD PROCESSING

15.00 CLASS 15.00 — WEARING APPAREL

16.00 CLASS 16.00 — EQUIPMENT AND SUPPLIES FOR HOUSEHOLD USE

CLASS 17.00 — EDUCATION AND INFORMATION

17.10 Printing and publishing
17.20 Education

18.00 CLASS 18.00 — RECREATION AND AMUSEMENT

19.00 CLASS 19.00 — EQUIPMENT AND SUPPLIES FOR OFFICE USE

CLASS 20.00 — MACHINERY AND EQUIPMENT FOR INDUSTRIAL USE

20.10 Metal working machinery
20.20 All other — including mine, construction, special and general industrial

CLASS 21.00 — NEW BUILDINGS, CONSTRUCTION OF

21.10 Buildings for manufacturing and commercial purposes, construction of

(Continued on page 8)

21.20 All types of dwellings, construction of
21.90 All other types of building, construction of
22.00 CLASS 22.00 — OPERATING SUPPLIES AND BUILDING REPAIR AND MAINTENANCE
23.00 CLASS 23.00 — ALL OTHER END USES (excludes all sub-assemblies and parts going into finished products coming with the other classes)

Five different Purchaser's Symbols are used to indicate the end user of the material. They are:

United States Army .... USA
United States Navy .... USN
(Including Maritime Commission)
Lend-Lease LL
Other Foreign Purchasers .... FP
Domestic Purchasers .... DP

These symbols must be passed on frcm supplier to his suppliers, etc., even though the Allocation Classification Symbol may change. A complete copy of the Regulation, together with detailed information on the various symbols to be used by different users and industries, has been reproduced by the Chamber of Commerce of the United States in Washington under the title, "Allocation Classification System."

— MARTIN A. GILMAN

MISCELLANY

● **THE ABILITY TO SEE HUMOR** in the difficulties of doing business these days is rare, but refreshing. An instance of this turned up recently in one of the printing trade publications. According to this item a purchasing agent wrote to some of his regular sources of supply, requesting new catalogs, with the complaint that the last issues were out of date. He expressed the hope that, if no new catalogs were available, he might be given some schedule by which he could figure prices from the catalogs he had. A few of the replies that he received are quoted below :

"We are glad to advise the illustrations in our catalog are still O.K., only we have discontinued most of the items. If we sent you the whole list of what we are not making, our catalog would be useless."

"Forget the prices. Also forget the descriptions. By the time you get this letter we do not know ourselves how or what we will be making the stuff out of."

"Thank you for your note indicating you still have one of our complete catalogs. Please return it at once. You ought to see the prices we are receiving here for our waste paper."

"After reading your inquiry we are afraid you are thinking of sending us an order. It looks suspicious to us. Nevertheless, we will gladly meet you halfway by showing you how to calculate costs, if you will promise to send the order to some one else."

"The only part of that catalog we are still certain about is the line that says, 'Established in 1885.' All other information and prices have been withdrawn."

"We will answer your question if you will first answer one for us: What do you want with a price on things we do not have and cannot get?"

GENERAL RADIO COMPANY
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THE General Radio EXPERIMENTER

VOLUME XVII No. 3 AUGUST, 1942

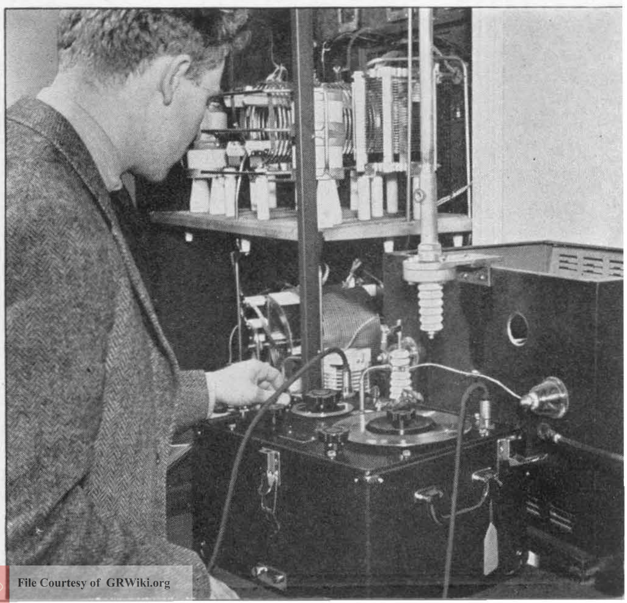
ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A NEW R-F BRIDGE FOR USE AT FREQUENCIES UP TO 60 MC

● THE NEED FOR A SIMPLE, ACCURATE INSTRUMENT to measure relatively low impedances in terms of their effective series resistance and reactance components has been pressing ever since quantitative information regarding antenna characteristics was first desired. The TYPE 516 Radio-Frequency Bridge,[1,2] first offered for sale in 1932, proved very satisfactory for such measurements at frequencies up to a few megacycles, and therefore found particular use in measurements of radiating systems in the standard broadcast band. However, as the upper frequency limit at which accurate measurements of impedance are required has increased, the limitations of this early bridge design have become increasingly apparent.

The TYPE 916-A Radio-Frequency Bridge,[3] described in this article, replaces the TYPE 516-C Radio-Frequency Bridge, supplying a wider range of direct impedance measurement in a more

FIGURE 1. Adjusting antenna coupling networks at Radio Station WHDH with the TYPE 916-A Radio-Frequency Bridge.



[1]Charles T. Burke, "Bridge Methods for Measurements at Radio Frequencies," General Radio *Experimenter*, Vol. 6, p. 1; July, 1932.

[2]C. E. Worthen, "Improvements in Radio-Frequency Bridge Methods for Measuring Antennas and Other Impedances," General Radio *Experimenter*, Vol. 8, p. 1; December, 1933.

[3]D. B. Sinclair, "A Radio-Frequency Bridge for Impedance Measurements from 400 Kilocycles to 60 Megacycles," *Proc. I.R.E.*, Vol. 28, p. 497; November, 1940.

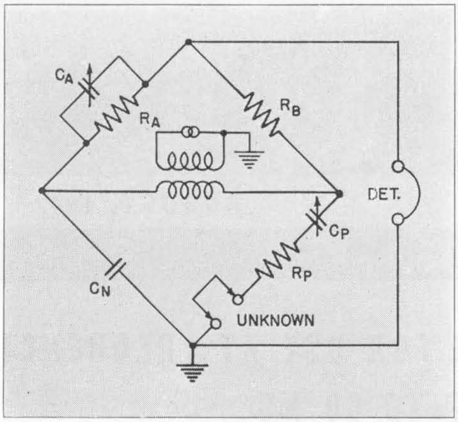


FIGURE 2. Elementary circuit diagram of the TYPE 916-A Radio-Frequency Bridge. The important feature distinguishing this bridge from the Schering bridge is the series substitution method of connecting the unknown impedance. The unknown reactance is determined from the change in setting of the condenser C_P and the unknown resistance from the change in setting of the condenser C_A.

convenient form over a wider frequency range. While useful as a general purpose instrument in the laboratory, the new bridge is particularly intended for measurements on radiating systems, and has been designed for maximum convenience in this application.

In addition to the greatly increased frequency range, the new bridge has two features that distinguish it from the older bridge, namely a considerably greater direct-reading resistance range, and a simplified dial for reading reactance. The resistance range, from zero to 1000 ohms, is covered on a single 8″ dial with a scale that is roughly linear from zero to 1 ohm and logarithmic from 1 ohm to 1000 ohms. The resistance-dial reading is independent of frequency. The reactance range, from zero to 5000 ohms, is covered on a single 4″ dial with a scale that is roughly linear from zero to 50 ohms and logarithmic from 50 ohms to 5000 ohms. The reactance-dial reading varies directly with frequency, the engraved scale being direct reading at a frequency of 1 Mc.

CIRCUIT AND THEORY

To achieve these greatly increased frequency and resistance ranges, the new circuit shown in Figure 2 has been developed.

Similar in appearance to the Schering bridge circuit, the new circuit differs in the method of connecting the unknown impedance and the method of measuring the reactive component. The balance conditions are:

$$R_P = R_B \frac{C_A}{C_N} \tag{1}$$

$$\frac{1}{j\omega C_P} = \frac{R_B}{R_A}\frac{1}{j\omega C_N} \tag{2}$$

When an impedance, $Z_x = R_x + jX_x$, is to be measured, the bridge is first balanced by means of the condensers C_A

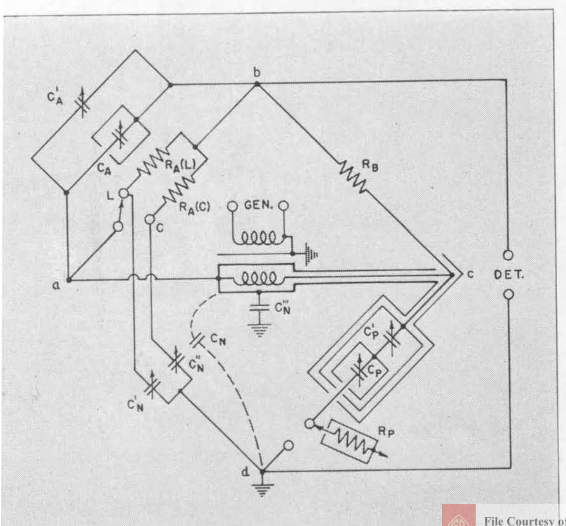


FIGURE 3. Complete circuit diagram of the TYPE 916-A Radio-Frequency Bridge. The L-C switch changes the value of the ratio arm R_A, thereby establishing the initial setting of the reactance dial at minimum or maximum for measuring inductive or capacitive reactance. The trimmer capacitances C_N', C_N'' are used to make the capacitance from point "a" to ground the same for the two positions of the L-C switch. The trimmer capacitance C_N''' is a part of the plug-in transformer assembly and is used to equalize the ground capacitances of the two transformers.

and C_P with a short circuit across the UNKNOWN terminals. The short circuit is then removed, the impedance connected, and the bridge rebalanced. This series-substitution method leads to the simple relationships:

$$R_x = R_B \frac{C_{A_2} - C_{A_1}}{C_N} \qquad (3)$$

$$X_x = \frac{1}{\omega}\left(\frac{1}{C_{P_2}} - \frac{1}{C_{P_1}}\right) \qquad (4)$$

in which subscripts 1 refer to initial balance values and subscripts 2 to final balance values.

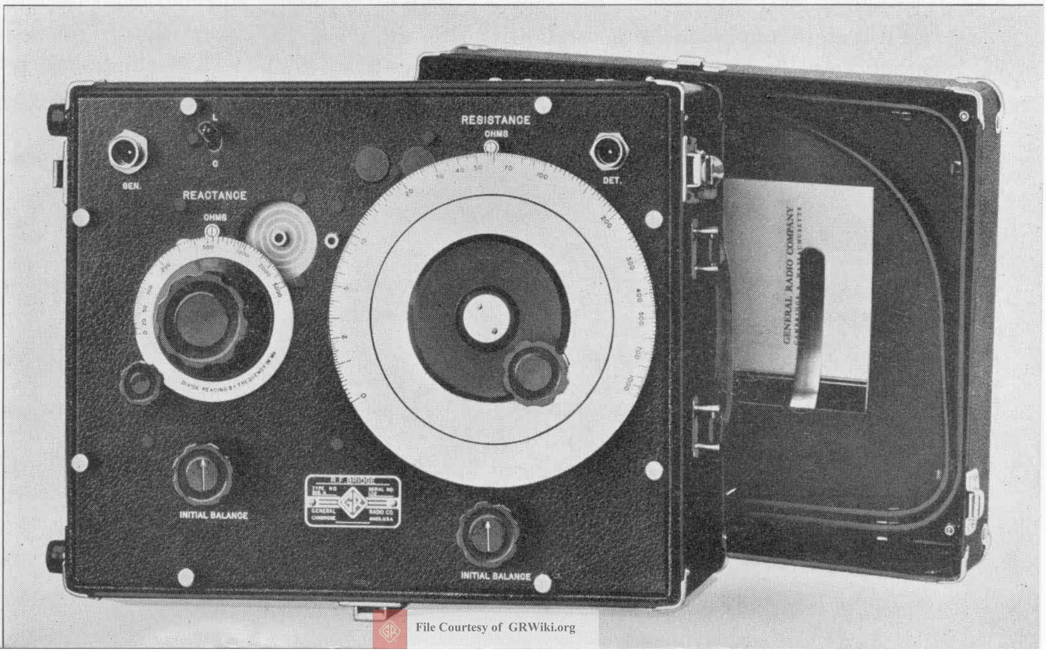
Equations (3) and (4) show that the resistance and reactance balances are independent and that each depends directly upon a change in setting of a variable air condenser. The absence of the cross terms that make the power-factor and capacitance balances interdependent in the Schering bridge results from the fact that the zero capacitance of the condenser C_A is balanced by the resistor R_P rather than by a trimmer capacitance C_B across the ratio arm R_B. The fact that the resistance component is measured in terms of the fixed resistance, R_B, and variable capacitance, C_A, is vital in establishing the wide frequency range, since these elements can be made with very small residual parameters. They combine, in this circuit, to form the equivalent of a continuously variable resistor having residual reactance much less than any variable resistor currently known.

DESCRIPTION

The complete circuit diagram of the bridge in Figure 3 shows the modifications in the basic circuit of Figure 1 necessary to meet the needs of a commercial instrument. The condensers C_A' and C_P' are trimmers for setting zero on the resistance and reactance dials when making the initial balance. The two ratio arms, $R_A(L)$ and $R_A(C)$, and the associated switch are provided so that the reactance dial can be set initially at zero for measuring inductive reactances, or at 5000 ohms for measuring capacitive reactances.

A panel view of the instrument is

FIGURE 4. Panel view of the bridge. The standard connecting leads supplied with the instrument plug into the jack adjacent to the reactance dial.



shown in Figure 4. Immediately below the RESISTANCE dial, at the right, is the INITIAL BALANCE knob controlling the condenser C_A'. Immediately below the REACTANCE dial, at the left, is the INITIAL BALANCE knob controlling the condenser C_P'. Immediately above the REACTANCE dial is the L–C switch for measuring inductive and capacitive reactances. The jack to which the unknown impedance connects is mounted in the center of the circular window above and to the right of the REACTANCE dial. A connecting lead, with a probe that plugs into this jack and that houses resistor R_P, is used to connect to the unknown impedance. Two of these leads, of different lengths, are supplied with the instrument.

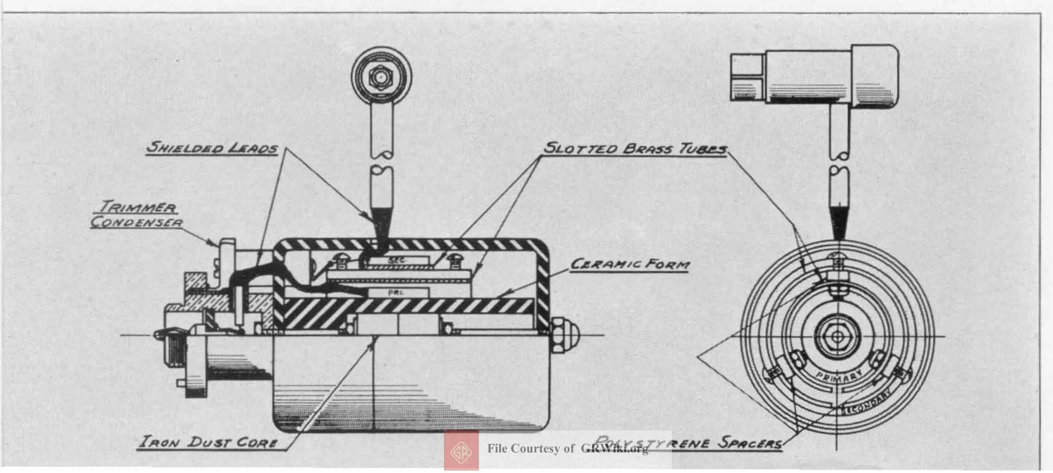
DESIGN FEATURES

While the circuit of Figure 2 is, in general, inherently more suitable for high frequency operation than previously used circuits, the actual characteristics obtained are directly affected by the mechanical and electrical design. Some of the more interesting features are described below.

In Figure 3 triple shielding is shown surrounding the reactance-measuring assembly comprising capacitances C_P and C_P'. The inner shield localizes the variable stray capacitance of the REACTANCE condenser, C_P, within the shield so that it cannot appear across the INITIAL BALANCE condenser, C_P', and cause interlocking of the settings of the two condensers. The middle shield eliminates any capacitance of the inner shield to ground and substitutes an intershield capacitance across the INITIAL BALANCE condenser. The outer shield eliminates the capacitance of the middle shield to ground and substitutes an intershield capacitance across the secondary of the transformer. The assembly as a whole therefore prevents any capacitance but that of the measurement jack itself from appearing across the measurement terminals, and eliminates any capacitance to ground between the measurement jack and the right-hand corner of the bridge. The capacitance of the outer shield to ground appears across the condenser, C_N, in the lower left-hand bridge arm. Actually the outer box dimensions and the spacing to the panel and cabinet shielding are so chosen that this residual capacitance forms the capacitance, C_N, itself.

Two 1:1 plug-in shielded transformers are supplied to cover the frequency ranges from 400 kc to 3 Mc and 3 Mc to

FIGURE 5. Sectional view of the shielded plug-in transformer. To make the shielding as complete as possible each winding is individually shielded with copper foil, in addition to the slotted brass tubes.



60 Mc. As shown in Figure 3, double shielding is required to complete the shielding system of the reactance-condenser assembly. The fundamental shielding requirements are that the grounded primary be located within a shield at ground potential and that the ungrounded secondary be located within a shield connected to the left-hand corner of the bridge. The shielding must prevent capacitive coupling between the windings, must be located so that the capacitance between the two shields is small compared with the capacitance to ground of the outer shield of the reactance-condenser assembly, and must not seriously impair the magnetic coupling between the windings. The design that has been found to furnish a satisfactory compromise is shown in Figure 5.

The most serious source of trouble in the design of these transformers was found to be the electromotive forces induced in the two split brass tubes used as shields between the primary and secondary. The potential difference taken along a radius between these tubes is practically zero at any point so long as the slots in the two tubes are lined up. If, however, the slots are not immediately opposite, over a sector between the two slots the radial potential difference is approximately equal to the electromotive force induced in a single turn in the magnetic field. Capacitive coupling between the two shields over this sector causes a residual component of voltage to be introduced from the left-hand corner of the bridge to ground. This component can cause very large errors in both resistance and reactance measurements at the higher frequencies. Locating the leads to the windings directly opposite the slots connecting the brass tubes to the coil at the same point, and lining up the slots carefully, makes the error from this source negligible. Rotation of the outer brass tube with respect to the inner, in fact, is used as an adjustment to eliminate it.

Residual unwanted impedances in the various circuit elements and in the wiring cause deviations from the idealized behavior so far tacitly assumed. Since the corrections that must be made for them generally depend upon frequency, they determine basically the frequency limits between which the instrument is satisfactory. One of the two corrections to be made is necessitated by the loss in the dielectric structure of the REACTANCE condenser, C_P. This loss causes an effective series resistance that varies inversely as the frequency and inversely as the square of the capacitance. It establishes the lower frequency limit for accurate measurements in the vicinity of 400 kc. A plot of the effective series resistance as a function of dial setting and frequency is shown in Figure 6. The other correction to be made arises from the effective series induc-

FIGURE 6 (*right*). Effective series resistance of reactance condenser as a function of dial setting and frequency.

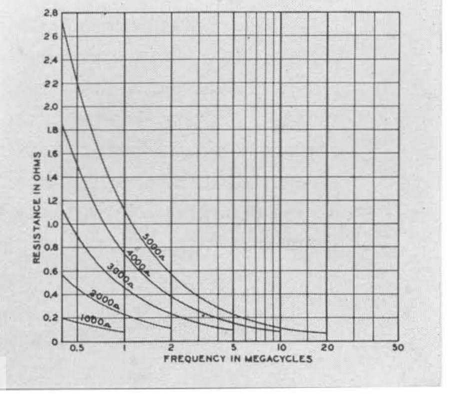
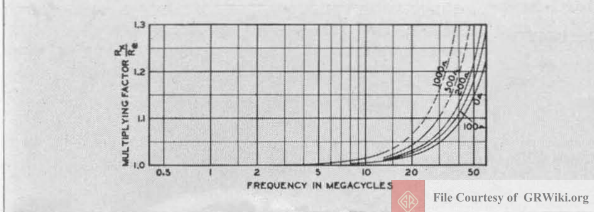


FIGURE 7 (*below*). Correction to be applied to resistance dial reading as a function of frequency.



tance of the RESISTANCE condenser, C_A. This residual inductance, L, causes the effective capacitance, $\hat{C}_A$, to differ from the static capacitance, C_A, according to the relation

$$\hat{C}_A = \frac{C_A}{1 - \omega^2 L C_A}$$

A plot of the correction to be made to the RESISTANCE dial reading as a function of frequency is shown in Figure 7. As the frequency is raised, the inductance is seen to reduce progressively the resistance range that can be measured and to establish an upper limit for accurate measurement in the vicinity of 60 Mc.

LEAD CORRECTIONS

In common with other types of impedance-measuring equipment, the bridge can only measure impedance at its own terminals. The residual impedances of the leads used to connect the unknown impedance to these terminals, however, often causes this impedance to differ from the impedance appearing at the terminals of the device under test. Under some circumstances the difference can be ignored and the measured impedance taken as the impedance of the device under test, including the leads. In most cases, however, the device will not be used with the same leads used to connect it to the measuring instrument and it is necessary to compensate for the effect of these leads to obtain the desired impedance.

To insure standard measurement conditions, two connecting leads are supplied, one about 5″ long and the other about 27″ long, over-all. Approximate capacitance and inductance figures are given in the instruction book and simplified procedures for making the corrections are outlined.

APPLICATIONS

The wide frequency range covered by the new bridge permits convenient and accurate direct measurements of low impedances at frequencies extending up through the f-m band to the top of television channel I. Two typical examples of measurements on an antenna and transmission line at frequencies between 2.5 and 60 Mc are shown in Figures 8 and 9. In addition to measurements of impedances that fall within the direct-reading ranges of the bridge, measurements can, of course, be made of higher impedances by indirect methods. The following examples show procedures to be followed in making typical measurements.

(*a*) *Measurement of 100 $\mu\mu f$ Condenser at 500 Kilocycles.*

The unknown impedance, in this example, is a small mica condenser of good power factor.

Plug short connecting lead (916–P3) into panel jack and fasten one lead of unknown condenser to panel binding post. Adjust location of unknown condenser so that clip of connecting lead can be transferred from ungrounded condenser lead to grounded condenser lead with minimum change in connecting-lead position. Reactance of condenser will be about 3200 ohms (1600-ohm change in dial reading) so balance cannot be made with switch in L position.

With switch in C position establish an initial balance. Set the REACTANCE dial at the lowest convenient reading, say 4000 ohms.

Transfer clip of connecting lead to ungrounded lead of unknown condenser and rebalance with RESISTANCE and REACTANCE dials. Suppose the respective readings are 2.3 ohms and 2450 ohms. Before corrections, the observed resistance, R_e, and reactance X_e, are:

$$R_e = 2.3 \text{ ohms}$$

$$X_e = \frac{2450 - 4000}{0.5} = -3100 \text{ ohms}$$

To correct for dielectric loss in the REACTANCE condenser look up in Figure 6 the effective resistances for dial settings of 4000 ohms and 2450 ohms at 0.5 Mc. The corrected value of R_e then becomes

$$R_e = 2.3 + 1.5 - 0.6 = 3.2 \text{ ohms}$$

To correct for the connecting-lead capacitance to ground, look up, in the lead reactance chart, the corresponding reactance X_a. It is $-114{,}000$ ohms. Applying Equations (5a) and (6a), which are given in the instruction book,

$$R_x = 3.2\left[1 + 2\left(\frac{-3100}{-114{,}000}\right) - \left(\frac{3.2}{-114{,}000}\right)^2\right]$$
$$= 3.4\ \Omega$$
$$X_x = -3100 + \frac{(-3100)^2 - (3.2)^2}{-114{,}000}$$
$$= -3184\ \Omega \text{ (capacitive)}$$

From these measurements, the capacitance, C_x, and dissipation factor,* $D_x = \frac{R_x}{X_x}$, are:

$$C_x = \frac{10^{12}}{2\pi \times 0.5 \times 10^6 \times 3184} = 100\ \mu\mu\text{f}$$

$$D_x = \frac{3.4}{3184} = 0.0011 = 0.11\%$$

This example is cited as an extreme case, in which failure to correct for the dielectric loss of the REACTANCE condenser leads to an error in resistance measurement of nearly 30%. For impedances in which the resistance component is larger compared with the reactance component the correction is of less importance.

(*b*) *Measurement of Broadcast Antenna at 1170 Kilocycles.*

In a typical case, the antenna terminal is located within a metal rack in a small house at the foot of the antenna tower. The bridge can be set up on packing boxes to come up to the front of the rack but cannot be brought close enough to the antenna terminal to use the short connecting lead (916-P3).

Plug long connecting lead (916-P4) into panel jack. Ground bridge to rack with short lead, preferably of copper strip 1″ or so wide. If this connection cannot be made conveniently to the clamp provided on the instrument case the panel can be loosened and a piece of copper foil slid onto the crack between the panel and the instrument case. *Do not ground to panel screws as they may not be making contact to the panel because of paint.* Arrange connecting lead so that it can be clipped to antenna terminal or to nearest ground point on rack with as little change in physical location as possible. The lead should be kept as far away from metal objects as possible throughout its length by any convenient means such as suspending it with string.

*This quantity is practically equal to the power factor (R_x/Z_x) for small values, and is often so miscalled.

Suppose the antenna to be about 0.6 wavelengths long, with an impedance having a capacitive reactance component. With the toggle switch set to the C position, and the connecting lead grounded to the rack, establish an initial balance. Set the REACTANCE dial to 5000 ohms pending further knowledge of the magnitude of the reactance.

Transfer clip of connecting lead to antenna terminal and rebalance with RESISTANCE and REACTANCE dials. Suppose the respective readings are 193 ohms and 4850 ohms. The resistance reading is adequate; the reactance reading is not as precise as might be desired because of crowding of the REACTANCE scale. To obtain a more precise reactance measurement, throw the toggle switch to the L position, set the REACTANCE dial to zero and rebalance the bridge with the two INITIAL BALANCE controls. Transfer clip of connecting lead to ground on rack and rebalance with RESISTANCE and REACTANCE dials. The RESISTANCE dial should rebalance at zero; suppose the REACTANCE dial reading is 160 ohms. Before corrections, the observed resistance, R_e, and reactance, X_e, are:

$$R_e = 193 \text{ ohms}$$
$$X_e = \frac{-160}{1.17} = -137 \text{ ohms}$$

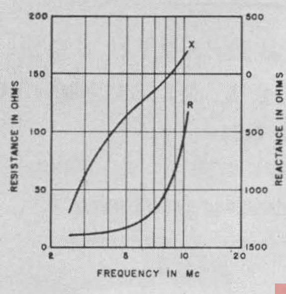
The corrections for dielectric loss in the REACTANCE condenser and inductance in the RESISTANCE condenser are seen, from Figures 6 and 7, to be negligible. To correct for the connecting-lead capacitance to ground, look up, in the lead reactance chart, the corresponding reactance, X_a. It is $-16{,}000$ ohms. Applying Equations (5a) and (6a), which are given in the instruction book,

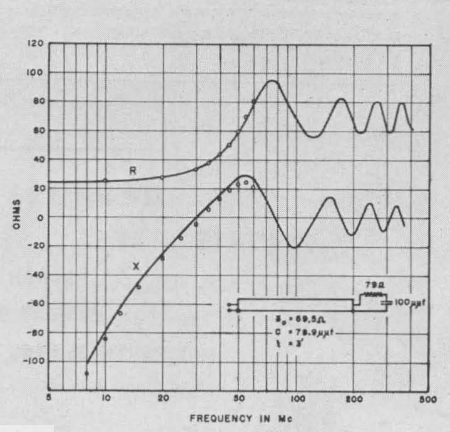
$$R_x = 193\left[1 + 2\left(\frac{-137}{-16{,}000}\right) - \left(\frac{192}{-16{,}000}\right)^2\right]$$
$$= 196\ \Omega$$
$$X_x = -137 + \frac{(-137)^2 - (193)^2}{-16{,}000}$$
$$= -136\ \Omega \text{ (capacitive)}$$

In this example, corrections are very small.

FIGURE 9 (*right*). Input reactance and resistance of a transmission line. The solid lines show calculated values; the circles, values measured with the bridge.

FIGURE 8 (*below*). Reactance and resistance of an antenna system as measured by the bridge.





The TYPE 916-A Radio-Frequency Bridge is particularly suited for such measurements.

(c) *Measurement of Terminated 72-Ohm Coaxial Line at 50 Mc.*

At very high frequencies, lead corrections become very important. It is, therefore, necessary to use the short connecting lead (916-P3). It is also desirable, if possible, to bring up the outer conductor of the coaxial line over the panel and make contact to it directly at the ground binding post on the panel.

Plug short connecting lead (916-P3) into panel jack. Clip to outer conductor of line or to ground binding post on panel, set toggle switch to the L position, and establish an initial balance. Set REACTANCE dial to as low a value as possible, say 500 ohms.

Transfer clip of connecting lead to center conductor of coaxial line and rebalance with RESISTANCE and REACTANCE dials. Suppose the respective readings are 64.5 ohms and 1450 ohms. Before corrections, the observed resistance, R_e, and reactance, X_e, are:

$$R_e = 64.5 \text{ ohms}$$

$$X_e = \frac{1450 - 500}{50} = +19 \text{ ohms}$$

To correct for inductance in the RESISTANCE condenser look up, in Figure 7, the correction for a dial reading of 65 ohms at 50 Mc. It is 1.17. The corrected value of R_e then becomes

$$R_e = 64.5 \times 1.17 = 75.4 \text{ ohms}$$

To correct for the connecting-lead capacitance to ground, look up, in the lead reactance chart, the corresponding reactance, X_a. It is -1150 ohms. Applying Equations (5a) and (6a), which are given in the instruction book,

$$R_x = 75.4\left[1 + 2\left(\frac{19}{-1150}\right) - \left(\frac{75.4}{-1150}\right)^2\right]$$

$$= 72.6\ \Omega$$

$$X_x = +19 + \frac{(+19)^2 - (75.4)^2}{-1150}$$

$$= +23.7\ \Omega \text{ (inductive)}$$

This example is cited as an extreme case, in which failure to correct for the inductance of the RESISTANCE condenser leads to an error in resistance measurement of the order of 12%.

— D. B. SINCLAIR

SPECIFICATIONS

Frequency Range: 400 kc to 60 Mc.

Reactance Range: 5000 Ω at 1 Mc. This range varies inversely as the frequency, and at other frequencies the dial reading must be divided by the frequency in megacycles.

Resistance Range: 0 to 1000 Ω.

Accuracy: For reactance, ±2% ±1 Ω.

For resistance, ±1% ±0.1Ω, subject to correction for residual parameters. At high frequencies the correction depends upon the frequency and upon the magnitude of the unknown resistance component. At low frequencies the correction depends upon the frequency and upon the magnitude of the unknown reactance component. Plots of both these corrections are given in the instruction book that is supplied with the bridge.

Accessories Supplied: Two input transformers, one covering the range from 400 kc to 3 Mc, the other from 3 Mc to 60 Mc; two leads of different lengths (for connecting the unknown impedance); two coaxial cables for connecting generator and detector.

Accessories Required: A radio-frequency generator and detector are required. The TYPE 605-B Standard-Signal Generator is a satisfactory generator. A well-shielded radio receiver covering the desired frequency range is recommended as the detector. The coaxial cable supplied for connection to the receiver is fitted with spade terminals at one end for connection to the receiver input terminals. For best results, however, it is recommended that the receiver be fitted with a TYPE 774-G Panel Plug and the cable with a TYPE 774-M Cable Jack.

Mounting: Airplane-luggage type case with carrying handles. Both input transformers are mounted inside the case. Coaxial cables, leads, and instruction book are stored in the cover of the instrument when not in use.

Dimensions: 17 × 13½ × 11⅛ inches, overall.

Net Weight: 35 pounds.

Type		*Code Word*	*Price*
916-A	Radio-Frequency Bridge	CIVIC	$350.00

Patent applied for.

GENERAL RADIO COMPANY
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90 WEST STREET, NEW YORK CITY
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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

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A GENERAL PURPOSE WAVEMETER

● **ONE OF THE INDISPENSABLE TOOLS OF THE RADIO ENGINEER** is the ordinary wavemeter. While precise frequency measurements are often necessary, there are many uses in the laboratory for an instrument that gives an answer within a few per cent quickly and conveniently. Among these are checking the frequency ranges of oscillator coils, setting and determining oscillator frequencies, and finding the frequencies of parasitic oscillations in r-f amplifiers For experimental work the low accuracy of the wavemeter, as compared to precise crystal frequency standards, is more than offset by the speed and convenience of measurement.

The two inexpensive, general-purpose wavemeters formerly carried in our catalog (Type 574 and Type 358) have now been replaced by a single instrument, Type 566-A, which combines, in an improved design, wide range, small size, good accuracy, and low price.

Figure 1. View of the Type 566-A Wavemeter showing how coils are stored when not in use.

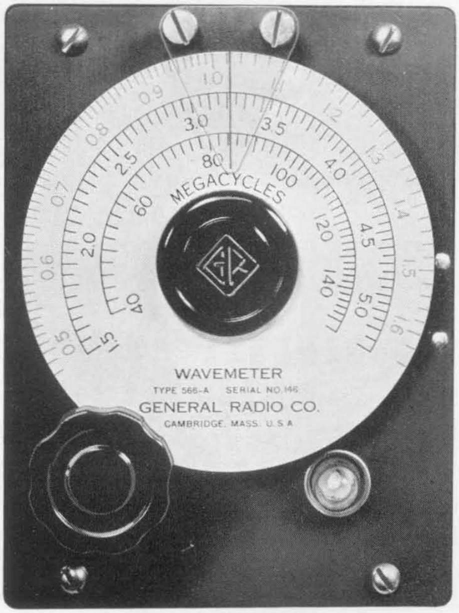


FIGURE 2. View of the wavemeter dial showing the frequency scales. The outer scale is engraved in red, the other two in black.

This new wavemeter is direct reading in frequency between 0.5 and 150 Mc. Only three frequency scales are used, as shown in Figure 2, although there are five plug-in coils. The outer scale is used with two coils, 0.5 to 1.6 Mc and 5 to 16 Mc. The middle scale covers the ranges 1.6 to 5 Mc and 16 to 50 Mc. The inside scale is used for the highest frequency coil, 50 to 150 Mc. These scales are accurate to ±2% up to 16 Mc, and to ±3% between 16 and 150 Mc.

The resonance indicator is an incandescent lamp. With low-power oscillators the reaction of the wavemeter on the oscillator tube currents can be observed.

Figure 3 gives an idea of the size of the wavemeter, and shows one of the features, that all coils except the highest frequency one can be rotated to secure the desired coupling. When not in use, coils are stored in the rack on the side of the instrument, as shown in Figure 1.

The slow-motion drive provided on the dial makes possible a fine adjustment of the condenser. The condenser itself is similar in construction to the TYPE 568, but has a longer stack. Figure 4 is an inside view of the instrument showing this condenser.

Four of the coils are wound on phenolic forms which enclose and protect the winding. The highest frequency coil, as shown at the left in Figure 3, is a straight bar.

The small size, accuracy, and low price of this wavemeter make it a particularly desirable instrument for the radio laboratory. Because all our facilities are devoted to war projects, this instrument is, at present, available only for war work. —E. KARPLUS

FIGURE 3. View of the TYPE 566-A Wavemeter in use. The coil can be rotated to secure optimum coupling to the source whose frequency is being measured.

SPECIFICATIONS

Frequency Range: 0.5 to 150 Mc (600 to 2 meters) using the five plug-in inductors furnished with the instrument. The condenser dial is direct reading in frequency. The precision with which the dial can be read is 2% or better.

Accuracy: The accuracy of dial indication is ±2%, 0.5 to 16 Mc; and ±3%, 16 to 150 Mc.

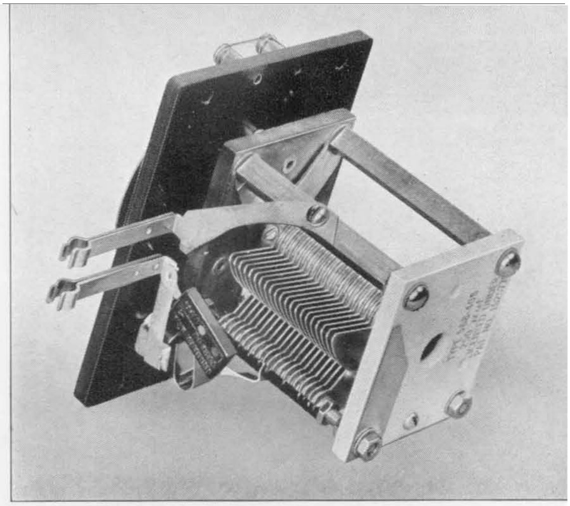


FIGURE 4. Interior view of the TYPE 566-A Wavemeter, showing the construction of the condenser.

Accessories Supplied: Two spare indicator lamps.

Dimensions: $4\frac{3}{4} \times 5\frac{7}{8} \times 5\frac{3}{4}$ inches, over-all.

Net Weight: 3 pounds.

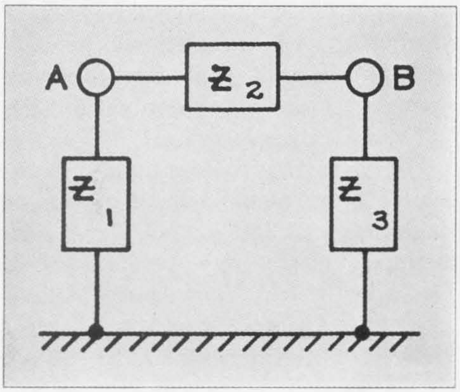
Type		*Code Word*	*Price*
566-A	Wavemeter....................	WAGON	$45.00

MEASURING BALANCED IMPEDANCES WITH THE R-F BRIDGE

INTRODUCTION

● **BECAUSE OF THE SPECIALIZED NATURE OF BALANCED IMPEDANCES**, equipment for their measurement has not received as much attention as has equipment for the measurement of impedances with one side grounded, and it is not as generally available. Consequently, the problem of measuring balanced impedances with equipment for measuring grounded impedances is often encountered. Measurements at radio frequencies of open-wire transmission lines and dipole antennas probably present the most common examples. Two methods by which these measurements can be accomplished are described here because of their particular usefulness with the TYPE 916-A and TYPE 516-C Radio-Frequency Bridges, and the TYPE 821-A Twin-T Impedance Measuring Circuit.

FIGURE 1. Equivalent circuit of a balanced line.



METHOD I

The first method[1] is similar to the well-known method of measuring the interelectrode capacitance of a triode by three capacitance measurements. The input impedance of the line is represented by the equivalent circuit of Figure 1. The measurement procedure is as follows:

(1) Short-circuit impedance Z_1 by grounding line A at point of measurement, and measure impedance from line B to ground. Call the measured value Z'.

[1]D. B. Sinclair, "Impedance Measurements on Broadcast Antennas," Part II, *Communications;* July, 1939.

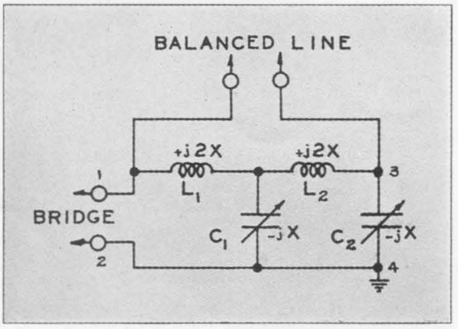


FIGURE 2. Circuit diagram of auxiliary network.

$$Z' = \frac{Z_2 Z_3}{Z_2 + Z_3} \qquad (1)$$

(2) Short-circuit impedance Z_2 by connecting line A to line B at point of measurement, and measure impedance from the junction to ground. Call the measured value Z''.

$$Z'' = \frac{Z_3 Z_1}{Z_3 + Z_1} \qquad (2)$$

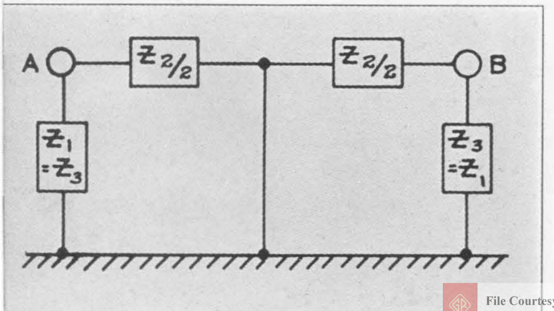
(3) Short-circuit impedance Z_3 by grounding line B at point of measurement, and measure impedance of line A to ground. Call the measured value Z'''.

$$Z''' = \frac{Z_1 Z_2}{Z_1 + Z_2} \qquad (3)$$

Combining Equations (1), (2), and (3) gives:

$$Z_1 = \frac{2Z'Z''Z'''}{Z'Z'' - Z''Z''' + Z'''Z'} = \frac{2}{-\frac{1}{Z'} + \frac{1}{Z''} + \frac{1}{Z'''}} \qquad (4)$$

FIGURE 3. Optional equivalent circuit of balanced line for balanced excitation voltages.



$$Z_2 = \frac{2Z'Z''Z'''}{Z'Z'' + Z''Z''' - Z'''Z'} = \frac{2}{\frac{1}{Z'} - \frac{1}{Z''} + \frac{1}{Z'''}} \qquad (5)$$

$$Z_3 = \frac{2Z'Z''Z'''}{-Z'Z'' + Z''Z''' + Z'''Z'} = \frac{2}{\frac{1}{Z'} + \frac{1}{Z''} - \frac{1}{Z'''}} \qquad (6)$$

Equations (4), (5), and (6) give each component of impedance, from which both the line impedance Z_{AB} and the unbalance can be found.

When the line is truly balanced, $Z_1 = Z_3$, $Z' = Z'''$, and

$$Z_1 = Z_3 = 2Z'' \qquad (4a)$$

$$Z_2 = \frac{2Z'Z''}{2Z'' - Z'} = \frac{1}{\frac{1}{Z'} - \frac{1}{2Z''}} \qquad (5a)$$

$$Z_{AB} = \frac{2Z_1Z_2}{2Z_1 + Z_2} = \frac{4Z'Z''}{4Z'' - Z'} \qquad (7)$$

This method, while highly accurate, is rather time consuming and is most useful when unbalance of the line is of particular importance.

METHOD II

A more rapid and direct method, suggested by Mr. John F. Byrne of Harvard University, is very useful when line balance is reasonably well maintained. It requires the use of the auxiliary network of Figure 2, which is a lumped-circuit single-frequency equivalent of a half-wave transmission line.

The open-circuit output voltage appearing across points 3 and 4 is equal in magnitude and 180° out of phase with the input voltage across points 1 and 2. The short-circuit impedance looking back from points 3 and 4 is zero. Thévenin's theorem therefore shows that, whatever the loading, the voltages between points 1 and 2 and between points 3 and 4 will be equal and 180° out of phase.

A balanced line connected between points 1 and 3 will, as a consequence, be excited with balanced voltages. Under this condition, the equivalent circuit of Figure 1 can be redrawn as shown in Figure 3 with the midpoint of the architrave impedance grounded.

When a balanced line is connected between terminals 1 and 3, it is therefore equivalent to connecting an impedance equal to Z_1 in parallel with $Z_2/2$ across the input terminals 1 and 2, and another identical impedance across the output terminals 3 and 4. Since the network simulates a half-wave line, the input impedance is equal to the terminating impedance. The effective input impedance, Z, is therefore

$$Z = \frac{1}{2}\,\frac{Z_1 Z_2/2}{Z_1 + Z_2/2}$$

$$= \frac{1}{4}\,\frac{2Z_1Z_2}{2Z_1 + Z_2} = \frac{Z_{AB}}{4} \qquad (8)$$

The experimental procedure in adjusting the circuit of Figure 2 is very simple. The two coils, L_1 and L_2, should be identical and should have a reactance at the operating frequency of the same order of magnitude as the impedance to be measured. The variable condensers, C_1 and C_2, should be of such size that their reactance can be set to half the coil reactances. The output terminals, 3 and 4, are first short-circuited and condenser C_1 adjusted to series resonance. This condition can be found by observing the input impedance on the bridge and setting to zero reactance. The output terminals are then open-circuited and condenser C_2 adjusted to parallel resonance. This is most easily observed by balancing the bridge with a capacitance across the UNKNOWN terminals and adjusting to make the change in reactance setting zero when the input impedance is connected in parallel. Losses in the condensers and coils used in the network will cause a small measurement error. By proper choice of circuit elements, however, these can ordinarily be made negligible. Substantial unbalance of the line will also cause error. Whenever suspicion arises that errors from these causes are significant, a measurement by Method I can be used for verification.

— D. B. Sinclair

TAKING THE PULSE OF TURBINES

● **AN EXCELLENT ILLUSTRATION** of the ever-increasing utility of electronic technique and equipment in the field of mechanical research is the method used by General Electric engineers for the study of vibration of high-speed-turbine buckets.

Figure 1. Type 713-B Beat-Frequency Oscillator mounted in assembly with power amplifier.

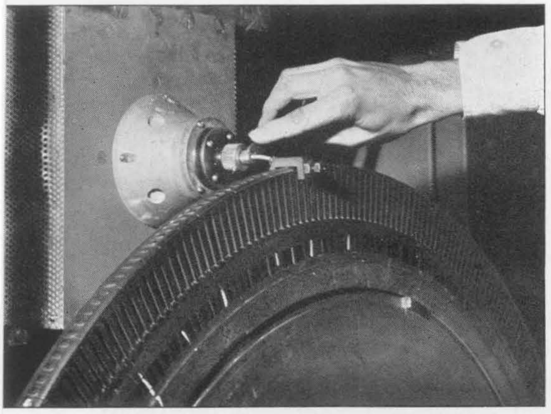
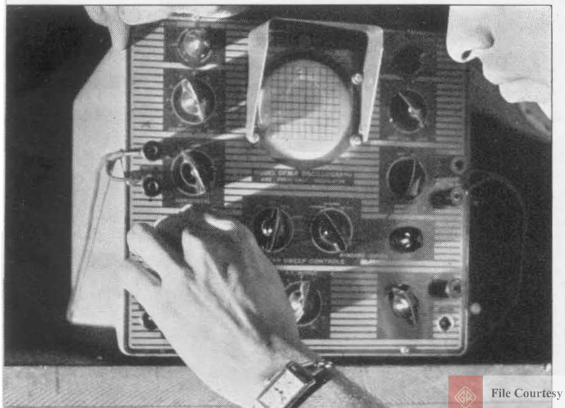


FIGURE 2. The oscillator supplies power to vibrate the turbine buckets through a power amplifier and an electro-mechanical transducer shown above.

FIGURE 3 (*above*). Engineers move the crystal pickup along the bucket tips. The pickup output is fed to the cathode-ray oscillograph shown below.

FIGURE 4 (*below*). Points of peak vibration are noted on the oscillograph as the pickup is moved along the buckets.



One of the problems confronting the designers of turbine wheels is that of obtaining a quantitative knowledge of the frequency and amplitude of resonant or other vibrations set up in the wheels, particularly in the buckets at the periphery (which may travel at speeds approaching a thousand miles per hour). In wheels designed for the high speeds prevalent in modern practice, the vibrations that occur are likely to be of high frequency and relatively small amplitude. Conventional mechanical means of reproducing and studying these conditions, while satisfactory with the larger structures and lower frequencies common at lower speeds, become practically useless. In the flexibility and versatility of modern electronic practice, however, a satisfactory solution has been found.

The buckets are vibrated by means of an electro-mechanical transducer, driven by a 1000-watt amplifier. A General Radio TYPE 713-B Beat-Frequency Oscillator was selected to drive the power amplifier. The relatively high-power output (1 watt), the low distortion, and the fine frequency control of the TYPE 713-B make it particularly suitable for this application. With the oscillator and power amplifier, the buckets can be driven at any frequency in the range from 30 cycles to 20,000 cycles.*

The resulting vibration of the driven buckets is picked up by a piezo-electric vibration pickup applied to the bucket tips, and applied to an amplifier feeding a cathode-ray oscillograph. The vibrations, amplified in magnitude approximately 100,000 times, can then conveniently be observed and measured. The resulting information assists in the design of wheels that are free from destructive resonances at normal speeds.

*The range of the TYPE 713-B is 20 cycles to 40,000 cycles, the power amplifier and driving system limiting the range to that given above.

DISCONTINUED INSTRUMENTS

● **IN ORDER TO PRODUCE CRITICAL WAR GOODS** at maximum efficiency, it is necessary for the manufacturer of specialties to eliminate from his line those items for which there is little demand, which can be easily produced by others, or which essentially duplicate other items in the line. In this way, production facilities, materials, and man power are conserved for the manufacture of more urgently needed items.

We have listed from time to time in the *Experimenter** instruments that are discontinued for the duration of the war. To those previously listed the following items have now been added:

Type 419-A Wavemeter
Type 714-A Amplifier
Type 672-A Power Supply
Type 673-A Power Supply
Type 755-A Condenser
Type 588-AM Meter

Except for frequencies between 150 Mc and 300 Mc, the Type 419-A Wavemeter can be replaced by the new Type 566-A described in this issue of the *Experimenter*. For frequencies above 150 Mc, the Type 758-A Wavemeter can be used.

The amplifier and the two power supplies can, for most uses, be duplicated easily in the laboratory or by other manufacturers.

The Type 588-AM Meter, formerly carried in our catalog for use with Type 493 Thermocouple, is no longer needed since this thermocouple was discontinued some time ago.†

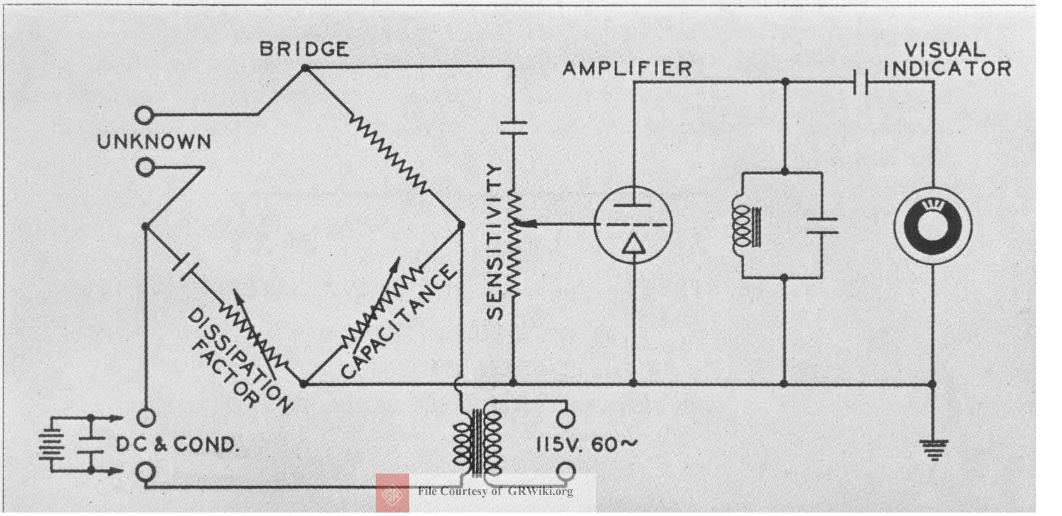
**Experimenter*, Sept., 1941, Feb., 1942.
†*Experimenter*, Dec., 1941.

USING A POLARIZING VOLTAGE WITH THE CAPACITANCE TEST BRIDGE

● **FOR MEASUREMENTS** of the capacitance and the power factor of electrolytic condensers with the Type 740-B Capacitance Test Bridge, it is usually desirable to apply a d-c polarizing voltage to the condenser under test, in order to simulate operating conditions.

Formerly, terminals for applying the polarizing voltage with this bridge were available on special order only, at an extra charge. Because of the growing de-

Figure 1. Schematic circuit diagram of the Type 740-B Capacitance Test Bridge.



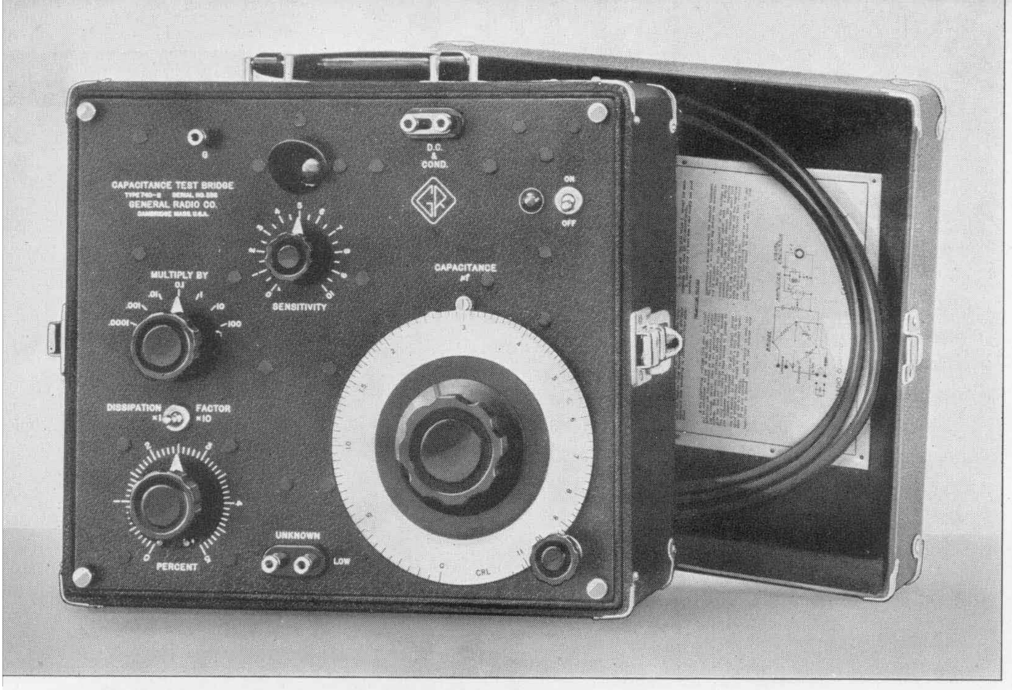


FIGURE 2. Panel view of the TYPE 740-B Capacitance Test Bridge. The polarizing voltage terminals are at the top of the panel.

mand for this feature, it is now included in the stock model of the bridge, at no increase in price.

Figure 1 is a schematic circuit diagram, and Figure 2 shows the position of the terminals on the panel.

The polarizing voltage is applied in series with the 60-cycle bridge supply. The condenser shown across the polarizing battery is usually necessary to avoid a reduction in bridge supply voltage resulting from the impedance of the battery in series with the supply. A rectifier-filter combination with a high-capacitance condenser in the filter output will obviate the need for the condenser as will also a storage battery, which usually has a high equivalent capacitance. The condenser should always be used with dry batteries.

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

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FREQUENCY CHARACTERISTICS OF DECADE CONDENSERS

● **IT HAS BEEN CUSTOMARY IN THE PAST** to think of the capacitance and dissipation factor of decade condensers, such as the TYPE 219 Decade Condensers and the TYPE 380 Decade-Condenser Units of which they are composed, as fixed quantities independent of frequency, at least to their stated accuracy. This is by no means the case and, since decade condensers are now being used over increasingly wide frequency limits, it becomes important to know the limits within which these condensers may be safely used.

The general way in which both capacitance and dissipation factor vary with frequency is shown by the various curves of Figure 6. There is always a minimum, but the increases at the high and low ends are produced by quite different causes. The low frequency rise is caused by dielectric polarization, a property of the solid dielectric of which the condensers are made, while the high frequency rise comes from the effect of the residual impedances in the leads from the terminals through the switches to the individual mica or paper condensers. Each of these causes will be discussed in considerable detail, first for TYPE 380 Decade-Condenser Units and then for TYPE 219 Decade Condensers.

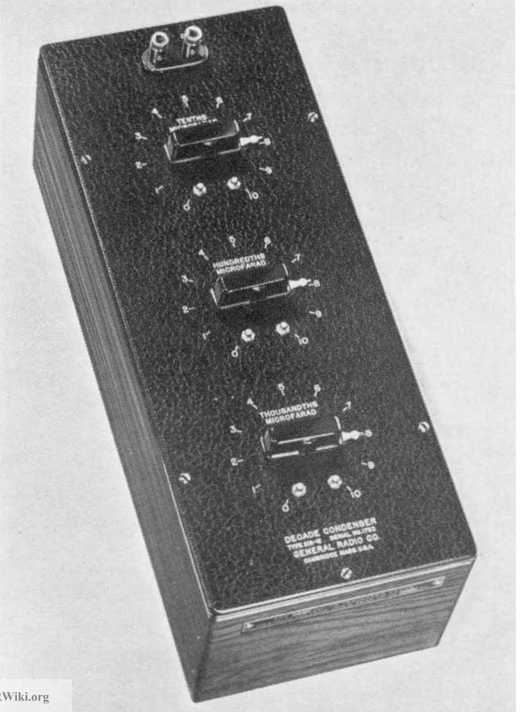

FIGURE 1. Panel view of a TYPE 219-M Decade Condenser.

DIELECTRIC POLARIZATION

The behavior of any dielectric as the frequency is varied depends upon the polarization existing in that dielectric. There are two distinct kinds of polarization, dipole and interfacial. Dipole polarization, as first described by Debye, occurs in dielectrics having polar molecules. In an alternating electric field these dipoles tend to oscillate with the field and the degree with which they succeed determines the increase in dielectric constant and capacitance. The way in which this increase occurs as the frequency is decreased is shown in Figure 1.[1] Dissipation factor also increases at first but reaches a maximum and then decreases in a symmetrical curve. Interfacial polarization, first described by Maxwell, occurs in composite dielectrics. The heaping up of the charged carriers, ions or electrons, at the interfaces of the components during each alternation of the electric field, serves to increase the dielectric constant and capacitance just as effectively as the oscillation of the dipoles. In fact the two types of polarization cannot be distinguished by the way either dielectric constant or dissipation factor varies with frequency. In that respect they differ only in the frequency ranges in which they occur, as indicated roughly in Figure 2. Interfacial polarization has also been called dielectric absorption and volume charge, these names being particularly appropriate when referring to the fact that the total charge stored in such a dielectric by the long-time application of a steady voltage can be several times that expected from the audio-frequency dielectric constant.

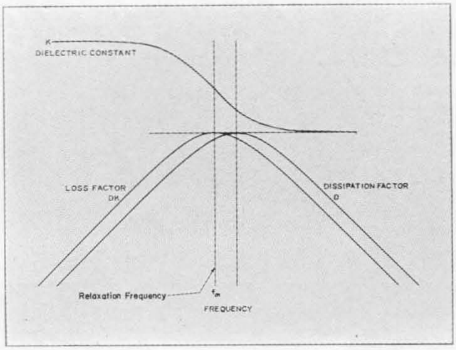

Figure 3. Change of dielectric constant, loss factor, and dissipation factor with frequency for a single polarization.

The frequency at which maximum loss factor [2] occurs, called the relaxation frequency, and the rates at which both loss factor and dielectric constant change with frequency depend on the kind of polarization, the kind of dielectric, and the temperature. For mica there is no dipole polarization and the relaxation frequency for its interfacial polarization is well below 1 cycle. The interfacial relaxation frequency for paper is also well below 1 cycle. Paper has also a dipole polarization in the neighborhood of 100 Mc, which is of little consequence be-

[1] E. J. Murphy and S. O. Morgan, The Dielectric Properties of Insulating Materials, *Bell System Technical Journal*, Vol. 16, Oct. 1937, pp. 493–512.

[2] Dissipation factor $D = \frac{R}{X}$; loss factor $= DK$, where K is the dielectric constant.

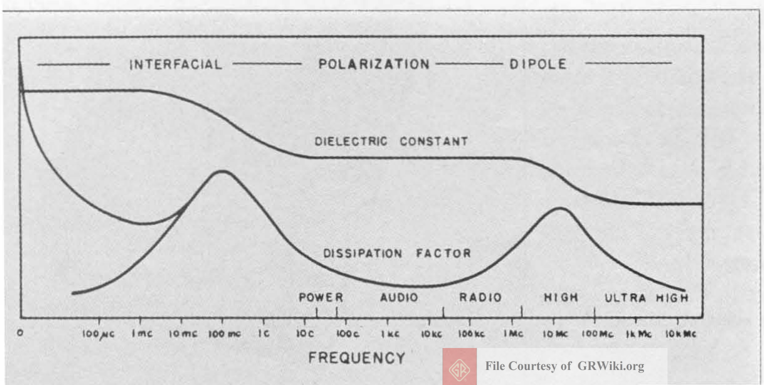

Figure 2. Typical curves showing how dielectric constant and dissipation factor change with frequency due to dielectric polarization (after Murphy and Morgan). The upper branch of the dissipation factor curve shows the effect of d-c conductivity.

cause paper condensers are rarely used at high frequencies. The use of wax or mineral oil for impregnation adds little to the polarization. Some of the recent synthetic oils, particularly those containing chlorine, have dipole polarizations with relaxation frequencies around 1 Mc.

The exact shapes of the curves of Figure 2 are shown in Figure 3, where frequency, loss factor, and dissipation factor are plotted logarithmically and dielectric constant arithmetically. At the relaxation frequency f_m, where loss factor is a maximum, dielectric constant is changing at the greatest rate and has a value just half way between its low and high frequency values. The shoulders of the loss factor and dissipation factor curves at a considerable distance from the relaxation frequency are linear. Similarly, a plot of the fractional increase in dielectric constant, expressed in terms of the total increase, at frequencies considerably higher than the relaxation frequency is also linear. The slopes of the shoulders of both of these lines are equal.[3]

Representative values for the four different kinds of Type 380 Decade-Condenser Units are shown by the lines of Figure 4, which start down from the left. In the case of dissipation factor this would imply very low values at high frequencies. Actually for mica and other low loss dielectrics, such as quartz and polystyrene, there appears to be an underlying polarization which provides an almost constant dissipation factor. The dipole polarization in paper causes a minimum in dissipation factor beyond which, at higher frequencies, the dissipation factor rises.

RESIDUAL IMPEDANCE

At high frequencies the small inductance and resistance of the leads of a solid dielectric condenser produce an increase in its apparent capacitance and dissipation factor.[7,8,9] The fractional increase in capacitance of the circuit shown in Figure 5, which represents this case, is

$$\frac{\Delta C}{C} = \frac{\omega^2 LC}{1 - \omega^2 LC} = \omega^2 LC \text{ approx.} \quad (1)$$

[3]The slopes are related to the depression angle of the center of the circular arc obtained by plotting loss factor against dielectric constant. [3,4,5] The theories of both Debye and Maxwell demand that such a plot result in a semicircle with its center on the dielectric constant axis. The fact that data obtained from all solid and most liquid dielectrics is well represented by a circular arc with depressed center was pointed out by K. S. and R. H. Cole[4,5,6] in 1941.

[4]K. S. and R. H. Cole, Dispersion and Absorption in Dielectrics, *Journal of Chemical Physics*, Vol. 9, Apr. 1941, pp. 341–351.

[5]R. F. Field, The Basis for the Non-destructive Testing of Insulation, *AIEE Transactions*, Vol. 60, Sept. 1941, pp. 890–895.

[6]W. Kauzmann, Dielectric Relaxation as a Chemical Rate Process, *Reviews of Modern Physics*, Vol. 14, Jan. 1942, pp. 12–44.

[7]R. F. Field and D. B. Sinclair, A Method for Determining the Residual Inductance and Resistance of a Variable Air Condenser at Radio Frequencies, *Proceedings of the Institute of Radio Engineers*, Vol. 24, Feb. 1936, pp. 255–274.

[8]D. B. Sinclair, Parallel-Resonance Methods for Precise Measurements of High Impedances at Radio Frequencies and a Comparison with Ordinary Series-Resonance Methods, *Proceedings of the Institute of Radio Engineers*, Vol. 26, Dec. 1938, pp. 1466–1497.

[9]D. B. Sinclair, The Behavior of Type 505 Condensers at High Frequencies, *General Radio Experimenter*, Vol. 12, Apr. 1938, pp. 4–8.

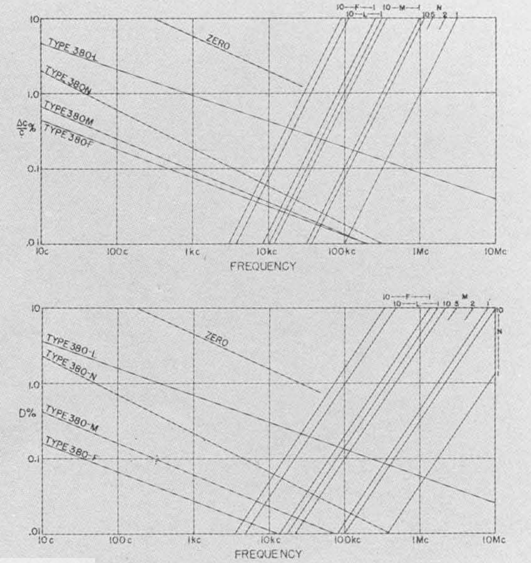

Figure 4. Components of fractional capacitance and dissipation factor as a function of frequency. The lines slanting down from the left come from interfacial polarization, those slanting up to the right come from residual inductance and resistance.

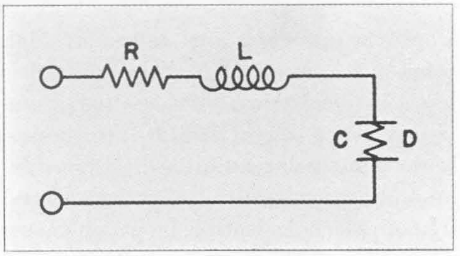

FIGURE 5. The residual impedances L and R are in series with the solid dielectric condenser having a capacitance C and dissipation factor D.

The fractional increase in capacitance varies as the square of the frequency. The increase in dissipation factor is

$$\Delta D = R\omega C \tag{2}$$

But the actual value of resistance to be used is the a-c resistance at a given frequency, not the d-c resistance. The leads are usually of sufficient diameter so that all the current flows on the surface.[10] Under this condition the resistance increases with the square root of the frequency,

$$R = R_1\sqrt{f} \tag{3}$$

where R_1 is the resistance at unit frequency, usually 1 Mc. Hence

$$\Delta D = 2\pi R_1' f^{3/2} C \tag{4}$$

where R_1' is the value that the resistance would have at a frequency of 1 cycle, if the square law held to that frequency. Thus, the increase in dissipation factor varies as the three-halves power of the frequency.

[10]H. A. Wheeler, Formulas for the Skin Effect, *Proceedings of the Institute of Radio Engineers*, Vol. 30, Sept. 1942, pp. 412–424.

Representative values of the residual impedances for the first four steps of a TYPE 380 Decade-Condenser Unit are 0.25 μh and 0.07 Ω at 1 Mc for the L, M, and N types and 0.35 μh and 0.01 Ω for the F type. The d-c resistance of the switch is only about 0.02 Ω, so the skin effect ratio at 1 Mc is 3.8. This corresponds to a No. 18 copper wire. The increases in capacitance and dissipation factor caused by these residuals are shown by the lines slanting upward on the right in Figure 4. The complete frequency characteristics of these decade condensers are the sums of the slanting lines for capacitance and dissipation factor, as shown by the curves in Figure 6.

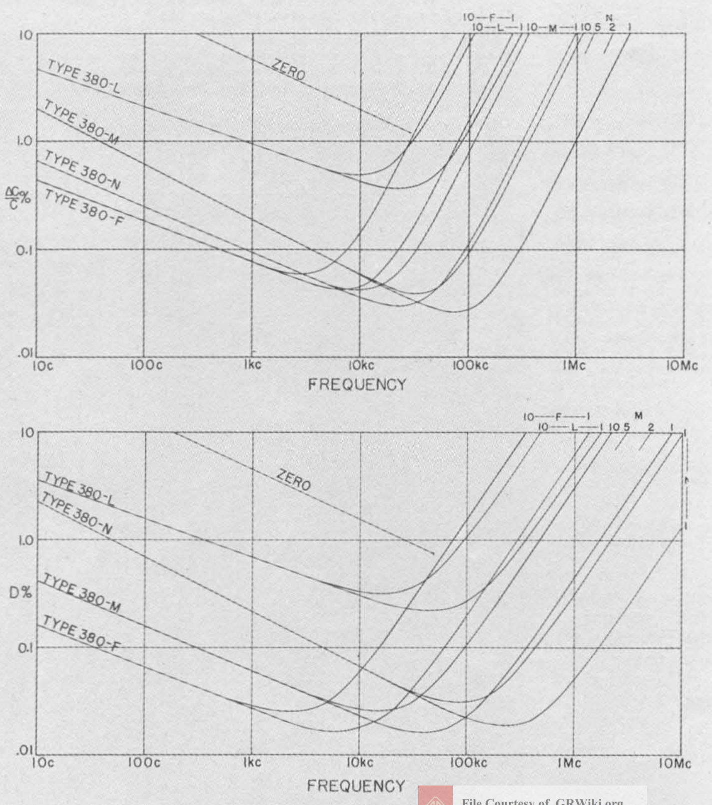

FIGURE 6. Frequency characteristics of TYPE 380 Decade Condenser Unit for switch positions 1 and 10, obtained by adding the components shown in Figure 4. The distribution of the curves for the other switch positions is shown for the TYPE 380-N Unit. The curves for the zero capacitance are shown in the upper left corner of each plot.

CONDENSERS IN PARALLEL

In a Type 380 Decade-Condenser Unit four separate condensers are successively connected in parallel groups by means of the drum switch shown in Figure 7 to give the ten steps of a decade. The four units have capacitances in the ratio 1 : 2 : 3 : 4, giving the first four steps directly and are then combined in the manner shown in Table I to give the other steps. For the small values of dissipation factor which these condensers have, capacitances add directly,[11] while dissipation factor is calculated from the rule that the products, dissipation factor times capacitance, add.[12]

$$C = C_1 + C_2 + \cdots = \Sigma C$$

$$D = \frac{D_1C_1 + D_2C_2 + \cdots}{C_1 + C_2 + \cdots} = \frac{\Sigma(DC)}{\Sigma C} \quad (5)$$

To a good approximation all four units have the same dissipation factor, which thus becomes the value for all settings. Limiting values of the *DC* products for the various decades are given in Table II.

The Type 380 Switch and the wiring to the four condensers has a capacitance of about 11 $\mu\mu$f. This is the zero capacitance of the switch and must be added to the sum of the capacitances of the condensers used at any setting to get the total capacitance. The dissipation factor of this zero capacitance is about 0.05 at 1 kc, thus giving a *DC* product of 0.55

TABLE I

Step	*Units Used*	*L/L$_C$ and R/R$_C$*
1–4	1 to 4	1.000
5	4 + 1	.680
6	4 + 2	.556
7	4 + 3	.510
8	4 + 3 + 1	.406
9	4 + 3 + 2	.358
10	4 + 3 + 2 + 1	.300

[11]Actually there are mutual capacitances among the four units which are successively shorted by the switch in its different positions. This causes the actual capacitance to be less than the sum of the separate capacitances. This difference is small and is significant only in the Type 380-N Decade.

[12]Note the analogy with loss factor which is dissipation factor times dielectric constant.

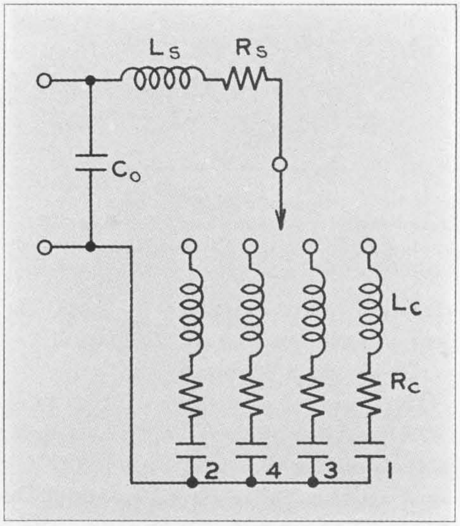

Figure 8. Schematic diagram of a Type 380 Decade Condenser Unit showing the residuals L_C and R_C associated with the separate condensers, residuals L_S and R_S of the leads, and the zero capacitance C_O with its dissipation factor D_O.

$\mu\mu$f. This is sufficiently large so that it must be included in calculating the dissipation factor for all settings of the Type 380-N Unit and the first several settings of the Type 380-M Unit. The

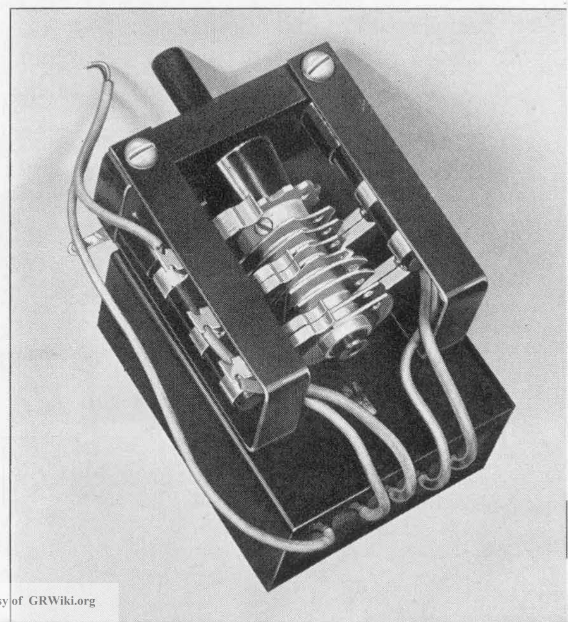

Figure 7. View of a Type 380 Decade Condenser Unit showing the switch and the metal container which holds the condensers.

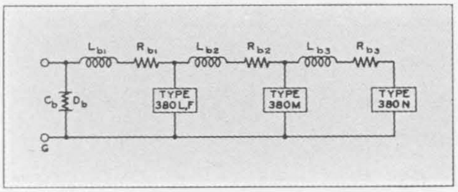

FIGURE 9. Schematic diagram of a TYPE 219 Decade Condenser, showing the added residuals L_B and R_B of the leads to the TYPE 380 Units and the added zero capacitance C_P and D_P.

variation with frequency of both the zero capacitance and its dissipation factor is shown in Figure 6.

The residual impedances of a TYPE 380 Decade-Condenser Unit are divided in the manner shown in Figure 8. Each condenser has the residuals L_C and R_C in its own leads to the switch, with the residuals L_S and R_S common to all for the L, M, and N types. Approximate values of these residuals are 0.10 μh and 0.15 μh for the inductances and 0.03 Ω and 0.04 Ω at 1 Mc for the resistances, respectively. Actually, because the leads from switch to condensers are of different lengths, the residuals L_C and R_C for the four condensers differ slightly. For a particular switch the measured values for the four lead inductances were 0.100, 0.076, 0.095, 0.076 μh in the order 1, 2, 3, 4. These values follow roughly the areas embraced by the leads as shown in Figure 7. These differences are not significant and it will be sufficient to use the rounded value 0.10 μh. There are also similar small differences in the resistance values. The residuals for the F type are 0.10 μh and 0.25 μh, 0.03 Ω and 0.07 Ω.

When capacitances with residual impedance are connected in parallel, the impedances add according to the following rules:

$$L = \frac{L_1C_1^2 + L_2C_2^2 + \cdots}{(C_1 + C_2 + \cdots)^2} = \frac{\Sigma(LC^2)}{\Sigma^2C}$$
$$R = \frac{R_1C_1^2 + R_2C_2^2 + \cdots}{(C_1 + C_2 + \cdots)^2} = \frac{\Sigma(RC^2)}{\Sigma^2C} \quad (6)$$

When the residuals are all equal,

$$L = L_C \frac{\Sigma C^2}{\Sigma^2 C}$$
$$R = R_C \frac{\Sigma C^2}{\Sigma^2 C} \quad (7)$$

which yield the ratios L/L_C and R/R_C for the ten positions of the switch given in Table I. For this case the residuals at maximum capacitance are only three-tenths of their values at minimum capacitance. The effect of this change in the apparent value of the residuals on the capacitance change and dissipation factor of the various TYPE 380 Units is shown in Figure 5. Curves for the maximum and minimum capacitance settings are shown for all four TYPE 380 Units.

TYPE 219 DECADE CONDENSERS

When several TYPE 380 Decade-Condenser Units are assembled to form a

TABLE II

Decade	*Material*	*Case*	*C* μf	*DC* $\mu\mu f$
F	Mica	505	.1 –1.0	20–200
M	Mica	Moulded	.01 – .1	5–50
N	Mica	Moulded	.001– .01	1–10
L	Wax paper		.1 –1.0	500–5000

TABLE III

Position	*Unit*	*L* μh	*R at 1Mc* Ω	*Total* *L* μh	*Total* *R at 1Mc* Ω
1	380–F & L	.16	.025	.16	.025
2	380–M	.10	.015	.26	.040
3	380–N	.10	.015	.36	.055

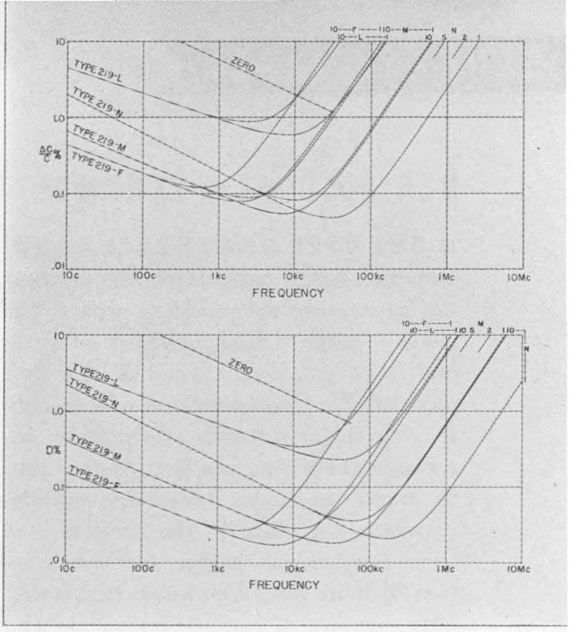

FIGURE 10. Frequency characteristics of TYPE 380 Decade Condenser Units when mounted to form TYPE 219-M and -K Decade Condensers. As compared to Figure 6, the high-frequency parts of the curves are shifted to lower frequencies.

TYPE 219 Decade Condenser, a zero capacitance common to all the units and residual impedances between the units is added as shown in Figure 9. The largest unit is placed next to the terminals and, therefore, has the smallest residual impedances. The added zero capacitance is about 8 $\mu\mu$f with a dissipation factor of about 0.05 at 1 kc. Their variation with frequency is the same as those of the separate units as shown in Figure 5. It is most convenient to add together all of the zero capacitances, for the box and for the three units. This gives a total of 41 $\mu\mu$f for the TYPE 219-M and 46 $\mu\mu$f for the TYPE 219-K Decade Condensers. Their respective *DC* products are 2.05 $\mu\mu$f and 2.30 $\mu\mu$f. These values are sufficiently large so that they must be considered in calculating the dissipation factors of all settings for both of the two lower decades.

Values of the residuals indicated in Figure 9 are given in Table III. It is also most convenient here to add together all of the residuals associated with each unit. These totals are given in the last two columns. The increases in capacitance and dissipation factor caused by these larger residuals are shown in Figure 10. These curves differ from those of Figure 6 mainly by having larger increases at the higher frequencies.

When more than one decade unit is used it is generally sufficient to assume that all of the capacitance is associated with the highest decade. This assumption will be safe except when the capacitances in the two higher decades are nearly equal, 18 or 19, or in the TYPE 219-K Decade Condenser where the TYPE 380-F Unit is so much better than the TYPE 380-M Unit. In any case, however, it must be recognized that the data and curves given in this article are average values, considerably smaller than the catalog limits, which may at times be approached. Divergencies from the given average values will be least for residual inductances, and most for the dissipation factors from dielectric polarization. High relative humidity can greatly increase the dissipation factor of the zero capacitances and to a lesser extent the capacitances and dissipation factors of the condensers themselves. The dissipation factors of the separate condensers may vary by a factor of two or three among themselves, and their capacitances are only adjusted within $\pm 1\%$, $\pm 2\%$ for TYPE 380-L. It follows, therefore, that accurate values of any particular decade condenser can be obtained only by extensive measurements on that particular instrument.

— R. F. FIELD

PERIODIC MAINTENANCE WILL KEEP THEM IN SERVICE

● **ONE USER OF GENERAL RADIO** standard-signal generators reports that the generators are working around the clock — three eight-hour shifts a day — and have not been turned off for four months! We know, and the user knows, that these instruments are built for intermittent service, and that their ability to stand up under these unusual demands is a tribute to the excellence of their mechanical design and construction. But we both also know that a definite schedule of periodic inspection and maintenance is what makes this performance possible.

If your General Radio instruments are being used in production testing, set up a maintenance schedule and see that it is rigidly followed. The small cost and bother will be saved many times over in eliminating the hours that are lost when instruments must be returned to us for repair.

Get a set of Service and Maintenance Notes for your General Radio instruments from our Service Department. These will help you to locate trouble and to identify defective components. Inspect, lubricate, and clean moving parts regularly, and if gears, slide wires, or other parts show serious wear, get a replacement from our Service Department and have it ready to install before actual failure. If major failures occur, write to our Service Department, which is equipped to give prompt service to war plants.

Our Service and Maintenance Notes are compiled from Service Department records (we keep a complete case history file for all major instruments), from our standards laboratory test specifications, and from engineering department suggestions. They have already helped hundreds of General Radio customers to get better service from their equipment, and to save valuable production time.

If you do not already have our Service and Maintenance Notes, write to our Service Department. You will receive in return a form on which to list the type number, serial number, and approximate date of purchase of your General Radio instruments. When the form is returned, the notes will be mailed to you promptly.

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.

GENERAL RADIO COMPANY

30 STATE STREET - CAMBRIDGE A, MASSACHUSETTS

BRANCH ENGINEERING OFFICES

90 WEST STREET, NEW YORK CITY

1000 NORTH SEWARD STREET, LOS ANGELES, CALIFORNIA

THE General Radio EXPERIMENTER

VOLUME XVII No. 6 NOVEMBER, 1942

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

Also

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PROGRESS IN SIGNAL GENERATOR DESIGN

● **FOURTEEN YEARS IS A LONG TIME** in radio history. These years have seen the adoption by the industry of complete a-c operation, the single tuning control, the superheterodyne circuit, short-wave bands, the high-powered output stage, the dynamic loudspeaker, the diode detector, and the automatic volume control, not to mention the countless other refinements of circuit and design which have been added from year to year and which brought the pre-war home-type receiver to a point of perfection and low cost undreamed of fourteen years ago. Similar progress has been made in commercial, military, and communications equipment, including tremendously improved sensitivity with lower noise level and high selectivity. Such rapid development has been made possible only by the availability of suitable measuring instruments, which have allowed definite and accurate evaluation of the various factors involved in receiver performance. Exact measurements of receiver performance have eliminated the guesswork from receiver

Figure 1. Panel view of the Type 805-A Standard-Signal Generator.

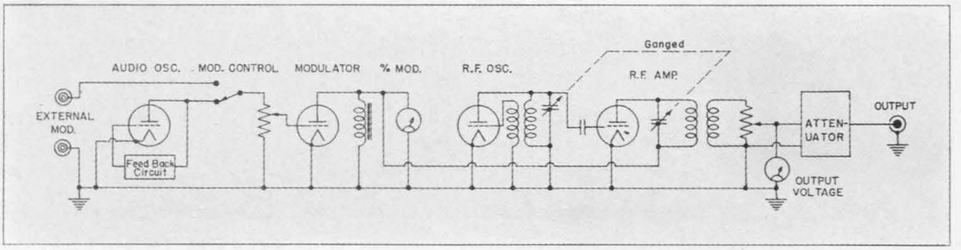


FIGURE 2. Elementary schematic circuit diagram of the TYPE 805-A Standard-Signal Generator.

design and resulted in a high degree of standardization and uniformity in the industry.

Just fourteen years ago the General Radio Company announced the TYPE 403 Standard-Signal Generator, the first commercial instrument of its type. To radio manufacturers, this early signal generator was the first commercially available means of measuring quantitatively the performance of their receivers. It was one of the first standardizing influences in an industry whose advertising claims were not exactly modest. To quote from the original announcement of the TYPE 403, "The inadequacy of such ratings as 'coast-to-coast reception every night' becomes apparent as the Barnum era passes."

In succeeding years the range, versatility, and accuracy of signal generators have been progressively improved, and models have been developed to meet definite requirements of frequency range, accuracy, etc. For some time, however, there has been a need for an amplitude modulated signal generator in the frequency range up to 50 megacycles, that would be suitable for testing all types of receivers, ranging from special military equipment to high-fidelity broadcast sets, and that could be sold at a reasonable price.

At the beginning of the war in Europe, one of America's largest receiver manufacturers asked the General Radio Company to undertake the design and manufacture of such an instrument. This was done, and a number of the new generators were in use in war production long before Pearl Harbor. A modification of this generator with further refinements is now available as the TYPE 805-A Standard-Signal Generator.

Since the generator was designed for use in the quantity production of receivers for the armed forces, it was necessary that high-quality performance be combined with ease of operation, ruggedness, and freedom from unnecessary frills. Modern design and quantity manufacturing have made it possible to price this instrument considerably below what has hitherto been considered normal for an instrument of this class.

Among the important features of this new generator are the following:

(1) Wide frequency range (16 kilocycles to 50 megacycles).

(2) High-ratio dial drive for accurate selectivity curves.

(3) Wide range of output voltage (0.1 microvolt to 2 volts).

(4) Constant impedance over entire output voltage range.

(5) Terminated cable to reduce reflection errors.

(6) Panel meters reading percentage modulation and output voltage directly and continuously.

(7) Degenerative modulating oscillator and modulator.

(8) High-level high-power modulation.

(9) Practical elimination of frequency modulation.

(10) Electronic voltage regulation of the power supply.

ELECTRICAL CIRCUIT

In order to provide the highest degree of performance with the simplest equipment, the signal generator is of the master-oscillator, power-amplifier type with a tuned output circuit. To eliminate frequency modulation without the use of a buffer stage, the oscillator runs at a relatively high-power level, nearly as high as the amplifier, and is coupled to the amplifier only very loosely. Amplitude modulation is accomplished in the plate circuit of the amplifier, thus providing high-quality modulation over the entire range of the instrument. Proper damping of the output circuit eliminates side-band cutting.

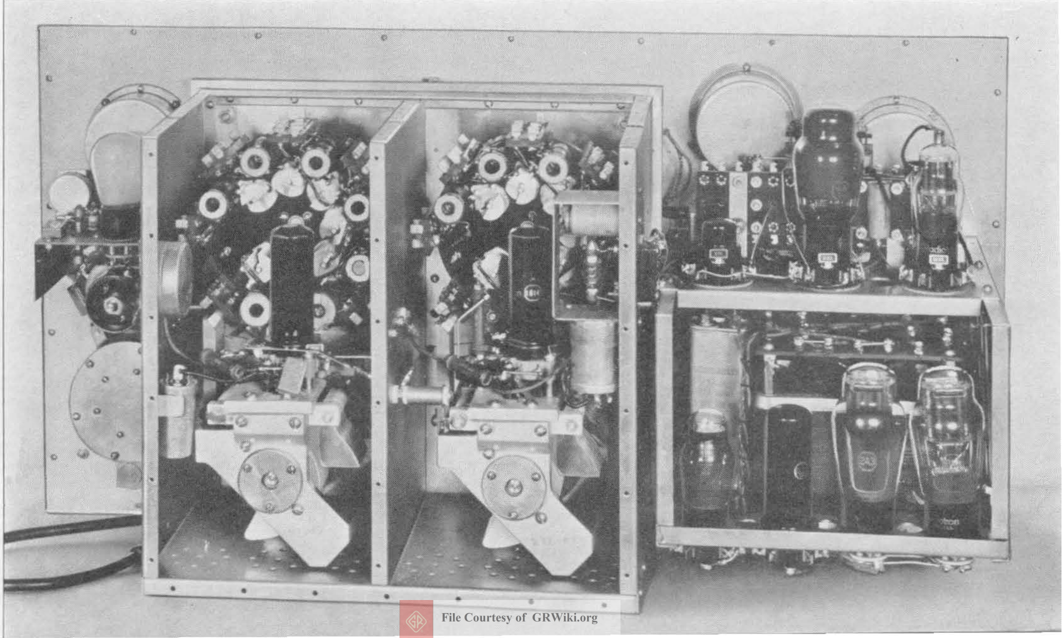
Figure 2 is a simplified block diagram showing the elemental operation of the circuit. In designing the generator, an attempt was made to keep everything as simple as was compatible with the desired performance. The generator, therefore, does not involve elaborate or new circuits, but rather represents a combination of sound engineering principles used to the best advantage.

TUNED CIRCUITS

Figure 3 shows a back view of the instrument with the shield removed. The oscillator and amplifier sections are identical except for a few minor differences. Each uses a type 1614 beam power tube, a condenser which provides a logarithmic frequency variation, and a coil turret.

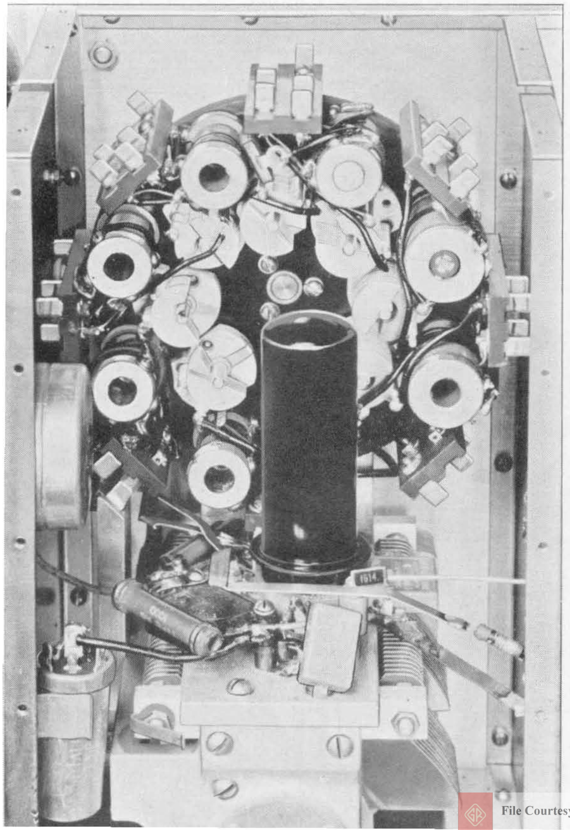
Associated with the coil for each range is an individual trimmer condenser and an adjustable iron-dust core, so that each range can be adjusted to track with a pre-engraved dial scale. One blank coil position is provided on each of the turrets so that an extra range can be added for any particular purpose.

FIGURE 3. Interior view of the signal generator.



All of the ranges excepting that for the highest frequency are direct multiples of lower-frequency ranges, so that a minimum number of dial scales is required. The total frequency range is covered on seven sets of coils. All frequencies are direct reading from the engraved panel dial, no calibration charts being necessary.

The two main tuning condensers have cast frames and ball bearings, and the plates are shaped to give a logarithmic frequency scale. These condensers are driven through a gear train which may be operated by either of two knobs. The first is used to change quickly from one point on the dial to another. The second provides a slow-motion reduction of 100 : 1 and allows frequency increments as small as 0.01% to be read directly.



MODULATION

The modulating oscillator is of the negative-feedback R-C type, providing good waveform and driving through an amplifier including a 6L6G tube in a degenerative circuit. Oscillator frequencies of 400 and 1000 cycles are provided, and provision is also made for modulating from an external source. The modulation is continuously variable from 0 up to 100% and for the broadcast and higher frequencies the characteristic is substantially flat from 50 cycles to 7000 cycles. Modulation at somewhat lower levels can be obtained at audio frequencies above and below these limits. The frequency modulation is so low as to be undetectable by ordinary means.

Percentage modulation is indicated directly by a vacuum-tube voltmeter connected to the output of the modulator. The plate voltages are held constant by an electronic, series-type regulator, utilizing two type 2A3 tubes controlled by a VR-150-30 regulator tube. This provides accuracy in the modulation calibration and also eliminates the effects of any fluctuations of line voltage within the range from 105 to 125 volts. Any power-line frequency from 40 to 60 cycles may be used, and the power transformer can also be reconnected for use on 210-250 volt lines.

OUTPUT CIRCUIT

The output circuit of the generator represents a considerable improvement over previous models. The tuned plate circuit of the amplifier feeds directly through a coupling coil to a constant-impedance type, Ayrton-Perry-wound volume control. The constant impedance feature eliminates any reaction of the

FIGURE 4. View of the coil assembly for one of the tuned circuits.

control on the r-f amplifier, but the output-voltage calibration does not depend upon this control, since the output voltage meter follows it in the circuit.

The output voltage is indicated directly on a panel meter having shaped pole pieces giving a semilogarithmic scale. This meter is driven by an average-type vacuum-tube voltmeter circuit which measures directly the input voltage to the step-by-step attenuator. In practice, then, the volume control is used to adjust this voltage, and the attenuator is calibrated as a multiplier. No slide-wire calibrations enter into the output reading whatsoever. Since the tube voltmeter circuit reads average voltage, it is unaffected by modulation and reads correctly whether or not the signal is modulated.

The step-by-step attenuator, enclosed in the cast "mousetrap" housing, is a further development of those used in the Types 603 and 605 Signal Generators and has seven steps, providing successive dividing factors of 10 : 1 in output voltage. For all settings of this attenuator the output impedance at the panel jack is 75 ohms. This matches the 75-ohm connecting cable, which is terminated in a resistance of 75 ohms, thus eliminating reflections and their consequent errors. The net output impedance at the termination unit is, accordingly, 37.5 ohms, and the panel meter reads directly in terms of the voltage across this impedance.

The terminating unit is also provided with a voltage divider for providing output impedances of 7.1 and 0.75 ohms, with 1/10 and 1/100 the normal output voltage, respectively. These low impedances are particularly useful in testing loop receivers. The terminating unit is also equipped with a standard IRE and RMA dummy antenna. The maximum output voltage available from the signal generator — 2 volts at 37.5 ohms — is sufficient for testing most high-level detector circuits and similar types of equipment. The minimum voltage, 0.1 microvolt, represents the best that the present state of the art requires. Leakage from the generator is kept well below this level.

MOUNTING

The complete signal generator is enclosed in a black wrinkle finish, steel cabinet, with full provision for proper ventilation and shielding. The modulation input terminals, power-supply leads, etc., are completely filtered.

OPERATING FEATURES

The advantages of the Type 805-A Standard-Signal Generator can only be appreciated through actual operation of

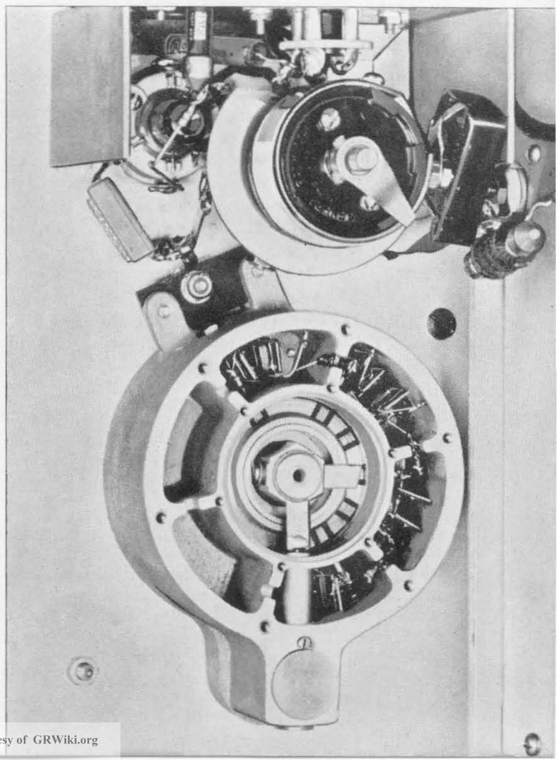


Figure 5. View of the attenuator with the cover of the multiplier housing removed. At the top is the Ayrton-Perry-wound output control.

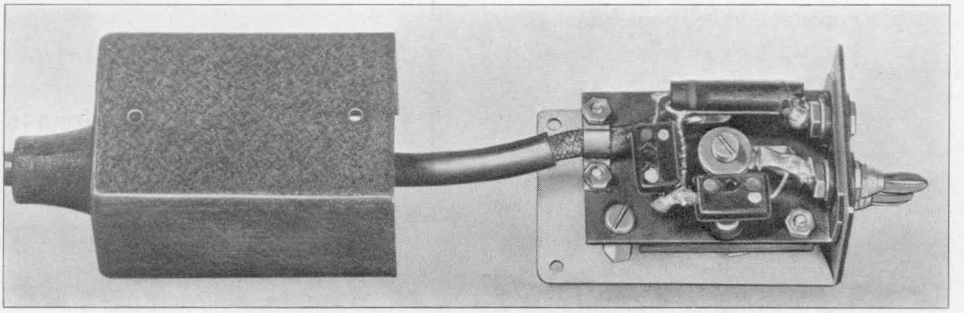


FIGURE 6. View of the cable terminating unit removed from its housing.

one of these units. The usual difficulties due to frequency modulation, cable errors, etc., are entirely absent for all practical measurements. The range switch is easily and quickly operated, the main frequency dial can be turned quickly to any desired point and increments measured with a high degree of accuracy, power-line fluctuations do not affect the operation of the generator, the percentage-modulation and output-voltage settings are continuously visible and direct reading, and the attenuator impedance does not change with setting. The modulation percentage and the output voltage do not change rapidly as the tuning is varied. While these generators are of a type which in pre-war days would be considered only for high-quality laboratory use, the strict requirements of war production have necessitated their use in a wide range of actual production work. In such applications, of course, ease and simplicity of operation are of prime importance.

— H. H. SCOTT

SPECIFICATIONS

Carrier Frequency Range: 16 kilocycles to 50 megacycles, covered in seven direct-reading ranges, as follows: 16 to 50 kc, 50 to 160 kc, 160 to 500 kc, 0.5 to 1.6 Mc, 1.6 to 5.0 Mc, 5.0 to 16 Mc, 16 to 50 Mc. A spare range position is provided so that a special set of coils can be installed if desired.

Frequency Calibration: Each range is direct reading to an accuracy of $\pm 1\%$ of the indicated frequency, except for the lowest frequency range, where the accuracy is 2%.

Incremental Frequency Dial: A slow-motion vernier drive dial is provided, by means of which frequency increments as small as 0.01% may be obtained.

Output Voltage Range: Continuously adjustable from 0.1 microvolt to 2 volts. The output voltage (at the termination of the 75-ohm output cable) is indicated by a panel meter and seven-point multiplier.

Output System: The output impedance at the panel jack is 75 ohms, resistive. A 75-ohm output cable is provided, together with a termination unit that furnishes constant output impedances of 37.5, 7.1, and 0.75 ohms. The calibration of the panel voltmeter-multiplier combination is in terms of the voltage across the 37.5-ohm output. When the 7.1 and 0.75-ohm positions are used, the indicated output voltage must be divided by 10 and 100, respectively. A standard dummy antenna output is also available at the termination unit.

Accuracy of Output Calibration:

Below 3 Mc	$\pm$ 3%	$\pm$0.1 microvolt
3 to 10 Mc	$\pm$ 5%	$\pm$0.2 microvolt
10 to 30 Mc	$\pm$ 10%	$\pm$0.4 microvolt
30 to 50 Mc	$\pm$ 20%	$\pm$0.8 microvolt

Modulation*: Continuously variable from 0 to 100%. The percentage of modulation is indicated by a panel meter to an accuracy of $\pm 10\%$ of the meter reading up to 80%.

Internal modulation is available at 400 cycles and 1000 cycles, accurate in frequency within $\pm 5\%$.

The generator can be modulated by an external oscillator. Approximately 5 volts across 500,000 ohms are required for 80% modulation. The over-all modulation characteristic is flat

*By means of a minor modification, it is possible to modulate the instrument externally with signals having steep wave fronts.

within ±2 db from 50 cycles to 7000 cycles, at carrier frequencies above 0.5 megacycle.

Frequency Modulation: Negligible for all practical purposes.

Distortion and Noise Level: The envelope distortion at a modulation level of 80% is less than 5% at 1 Mc carrier frequency. Carrier noise level is at least 40 db below 80% modulation.

Stray Fields: Radio-frequency leakage fields are completely negligible with respect to the calibrated output voltage, at all levels down to 0.5 μv. At the higher frequencies, and for output settings below 0.5 μv, a very small amount of leakage may be detected within a few inches of the panel, but the 3-foot output cable allows the receiver under test to be kept well beyond this field.

Power Supply: The instrument operates from any 40 to 60 cycle, 115-volt (or 230-volt) line. An electronic voltage regulator compensates for line voltage fluctuations from 105 to 125 volts (or from 210 to 250 volts). A maximum input power of 180 watts is required.

Tubes: Supplied with instrument:

2 — type 1614	1 — type 6SF5
1 — type 6C8-G	1 — type VR-150-30
1 — type 6L6-G	1 — type 955
1 — type 5T4	1 — type 6H6
2 — type 2A3	1 — Sylvania Ballast Lamp No. 2

Accessories Supplied: Seven-foot line connector cord, spare pilot lamps and fuses, shielded output cable and termination unit, and one Type 274-M Plug.

Mounting: The panel is black crackle finished and the cabinet is black wrinkle finish.

Dimensions: (Height) 16 x (width) 33 x (depth) 12 inches, over-all.

Net Weight: 120 pounds, approximately.

Type		*Code Word*	*Price*
805-A†*	Standard-Signal Generator........	LEPER	$850.00

†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.

*Although this signal generator has not been publicly announced before, many are already in service, mainly as the result of word-of-mouth information passed along from one user to another. A large number of orders are on our books awaiting shipment, so deliveries of new orders at the present time are necessarily delayed. Users who are going to need the instrument are urged to apply the best possible priority rating to their orders because under present conditions shipment on priority ratings lower than AA-1 are very uncertain.

ORDERS FOR REPLACEMENT PARTS

● **IN ORDER TO ASSIST CUSTOMERS** who service their own General Radio instruments, we maintain a small stock of replacement parts for most major instruments. Replacement parts are just as scarce as new instruments. Therefore, if this stock of replacement parts is to be of maximum service to industry, it is necessary that each customer order only enough for his immediate needs. Over-ordering only prevents some other user from getting badly-needed replacements.

Many replacements are of standard manufacture. Among these are vacuum tubes, pilot lamps, and fuses. Whenever possible, these should be purchased locally, because our stock is limited and is earmarked for use in new instruments. Where tubes are critical and must be selected, they should, of course, be ordered from us.

Replacement output cables and interconnecting cables can no longer be supplied, because of the shortage of rubber and copper. When cables fail, attach the old end fittings to whatever substitute conductors you can get to do the job adequately. If the end fittings are damaged or missing, we can supply replacements.

For the same reason, we are no longer able to supply replacement power cord-and-plug assemblies for our instruments. If your power cord fails, replace it with any of the types available on the market. The male plug is readily obtainable, but the small female plug may be difficult to obtain locally. A satisfactory replacement is the type 2173 manufactured by the General Electric Company, Bridgeport, Connecticut.

Our stock of replacement batteries is exhausted, and no more can be obtained. If the correct type cannot be obtained locally, substitutes must be used. While some instruments can be operated satisfactorily from external batteries, without any particular precautions, in others the problem is complicated by the necessity for complete shielding of the power supply. We hope to have more on this subject in a forthcoming issue of the *Experimenter*.

Electrical indicating instruments (meters) are hard to get. New production of the instrument manufacturers is practically all allocated to new war equipment. Repairs, however, can still be handled in a few weeks, and we urge that damaged meters be returned to us for repair.

We hope that our customers will cooperate with us in keeping their replacement part orders within reasonable figures. At present, every order is checked against our serial-number records and orders for excessive quantities are necessarily reduced or refused. Only in this way can replacements be supplied to all who need them with a minimum delay.

— H. H. Dawes

SELL US YOUR UNUSED PARTS

● **WHEN GENERAL RADIO** circuit components such as Variacs, rheostats, decade resistors, and decade condensers are built into another manufacturer's equipment, knobs or dials are sometimes removed by the manufacturer and replaced by items which conform to his standard design. We can use these General Radio parts. They are built of scarce materials. If you have accumulated surplus parts in this way, return them to us and credit will be allowed, provided of course that the parts are in good condition. Please write our Service Department for shipping instructions and credit allowances.

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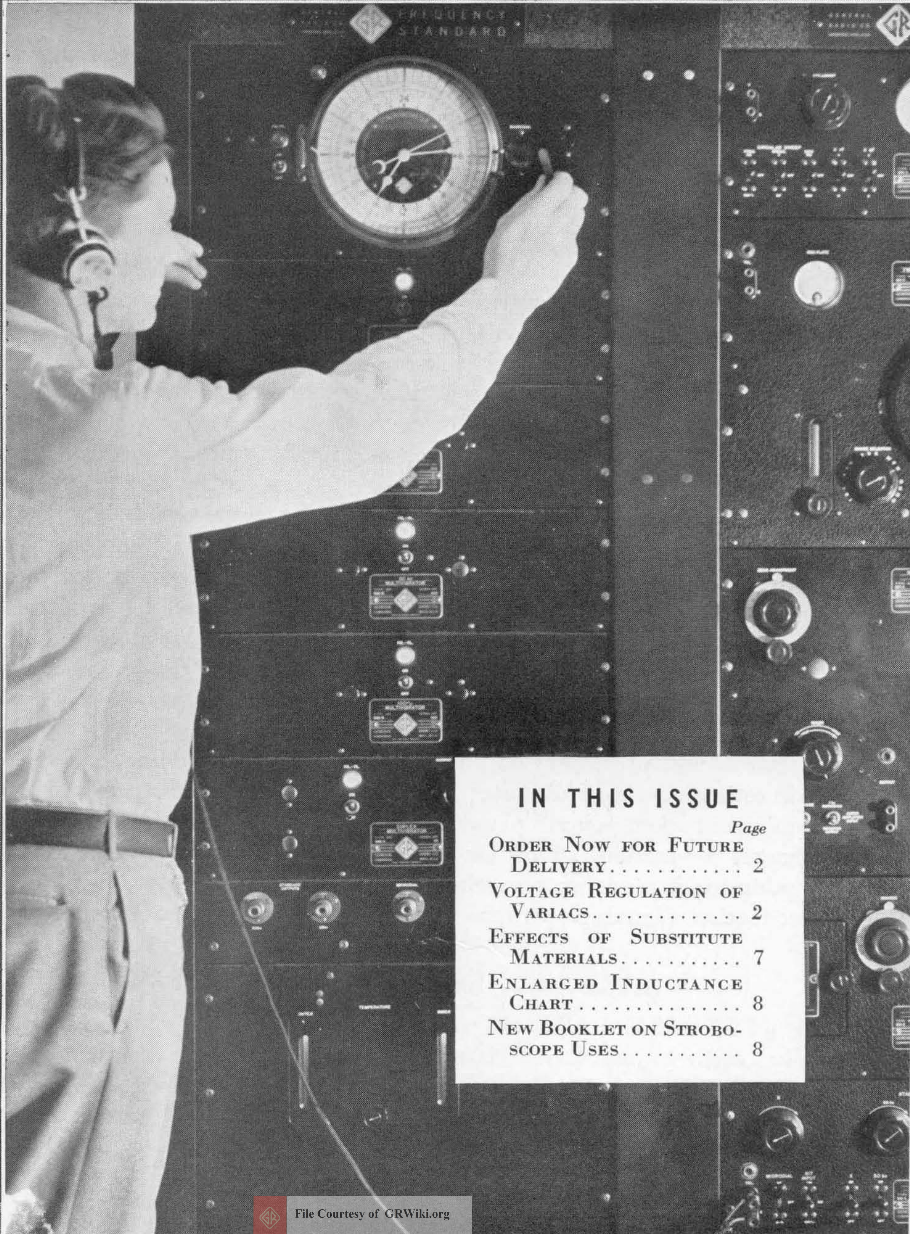
THE General Radio EXPERIMENTER

VOLUME XVII No. 7 DECEMBER, 1942

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS



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Cover Photograph.
Checking the Class C-21-HLD Primary Frequency Standard against radio time signals.

ORDER NOW FOR FUTURE DELIVERY

● **TO MANUFACTURE** the several hundred items of test equipment, instruments, and component parts that are in continuous production at General Radio we must have an unbelievable variety of materials. Nearly all of them are, unfortunately, very scarce, and, for the most part, there are no known substitutes. They are available only on orders bearing the highest priority ratings, and even then they must be ordered months in advance of actual production.

A complete production cycle including the procurement of materials is planned many months in advance. This same procedure is, of course, used in peacetime, too, but normally the cycle takes only a few months, rather than eight to ten months as is now required. In peacetime the material goes on the shelf and is available for immediate delivery. In wartime things are different, and it is impossible to produce equipment fast enough to fill all needs from "off the shelf."

In order to include your requirements in future production, we naturally must know what they are. You can be much better assured of delivery on time if you will estimate your requirements for test equipment when you are planning your general production. Unfortunately test equipment is sometimes overlooked in production planning, perhaps because for so many years it has been available for immediate delivery. If you will give us all the advance notice that you can of your test equipment requirements and will specify delivery at approximately the date when you will need the equipment, we will, as a general rule, be able to meet your needs.

Please do not specify early delivery if later delivery will do. This only prevents someone else from getting his equipment on time and will interfere with the general war effort.

If your order is for a quantity of parts such as Variacs, rheostats, plugs, etc., which are to be built into your equipment, specify the rate at which the material will be needed, so that we can schedule our production of parts correctly to meet your requirements. Please ask for continuing deliveries as you need the material so that it will not pile up in your stock room and thus be idle for a long time before it is finally used. We are sure that careful planning along these lines will help immeasurably to make deliveries on *required* schedules and will eliminate delays both in your production and in ours.

All of our production facilities are now completely allocated for delivery during the balance of this year and for early in 1943. Your order should be placed now for material that you will require in the late spring, summer, and fall of 1943.

VOLTAGE REGULATION OF VARIACS

● **USERS OF VARIACS** occasionally desire quantitative information regarding the variation of output voltage with changes in load, and so we are publishing here a rather complete picture of the regulation of all the current models.

The accompanying curves show the variation of output voltage as obtained from actual measurement with resistive loads drawing the currents indicated on the various curves. Plots are given of voltage drop versus dial setting, which is the no-load voltage when the nominal value of input voltage, 115 or 230 volts, is used.

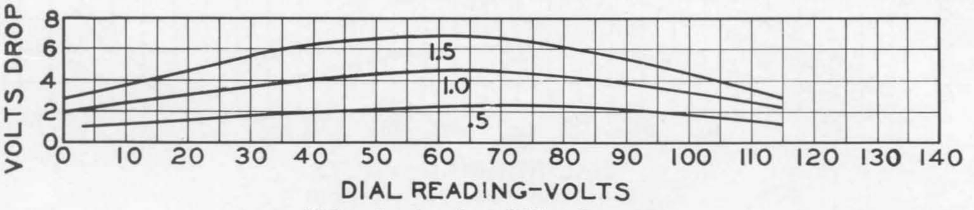
Some inaccuracy in the data is to be expected since the voltage drop is a difference reading of two relatively large voltages. Furthermore, the brush drop varies not only with current but also with particular position and number of turns bridged. The brush drop is naturally most important at low currents and at voltages near zero and line voltage. At other settings and for the higher currents the leakage reactance and resistance of the Variac winding cause the major part of the drop.

When 230-volt models are operated at 115 volts input, very high reactance drops occur near maximum output voltage. Consequently, this connection should not be used for high output voltages except at very low output current.

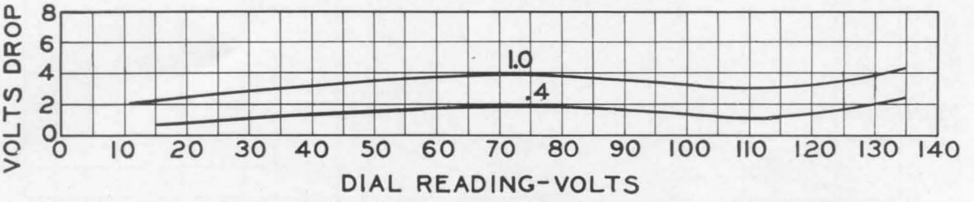
It should be pointed out that for some conditions data are given for currents in excess of the rated current of the Variac. This is done to give a more complete picture of performance, especially since some users may have applications where maximum current will be drawn near the mid-point for short periods of time only.

— Martin A. Gilman

TYPE 200-B

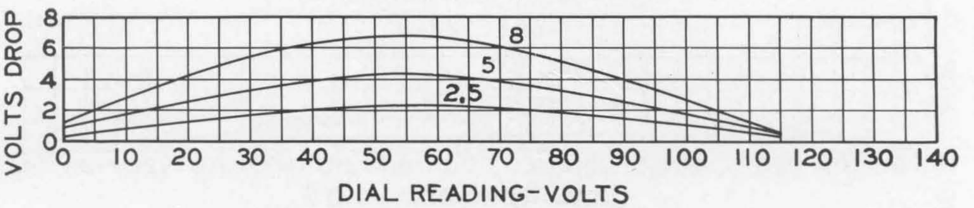


115 volts in; 0 to 115 volts out



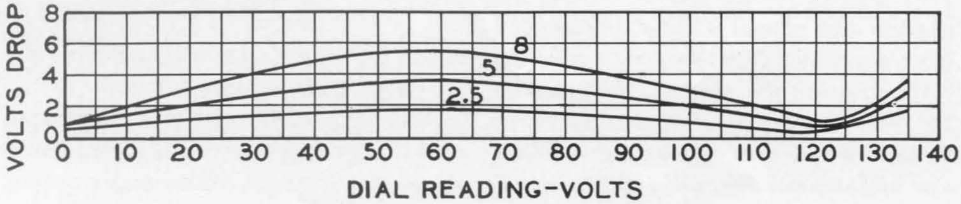
115 volts in; 0 to 135 volts out

TYPES 200-CM AND 200-CU



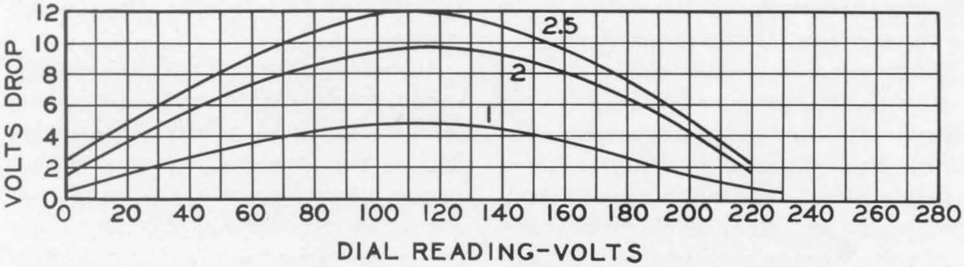
115 volts in; 0 to 115 volts out

TYPES 200-CM AND 200-CU (continued)

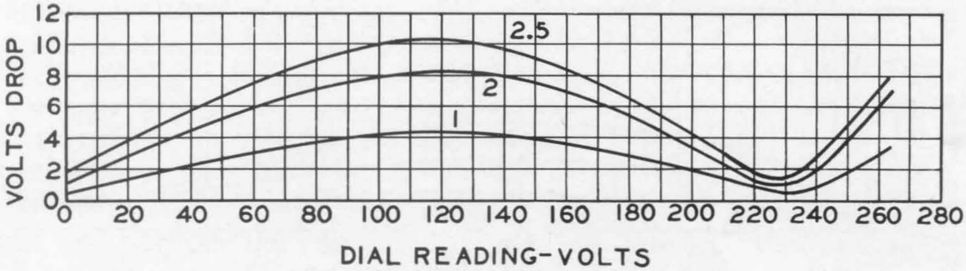


115 volts in; 0 to 135 volts out

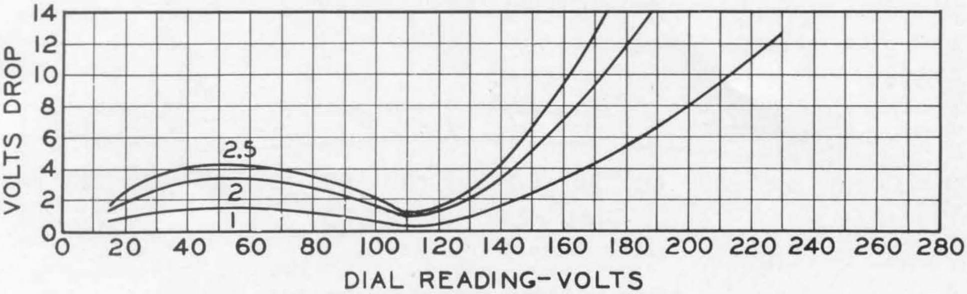
TYPES 200-CUH AND CMH



230 volts in; 0 to 230 volts out

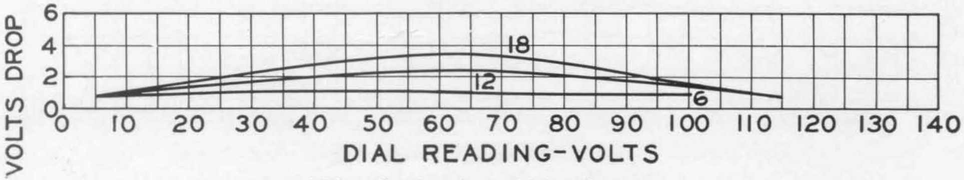


230 volts in; 0 to 270 volts out



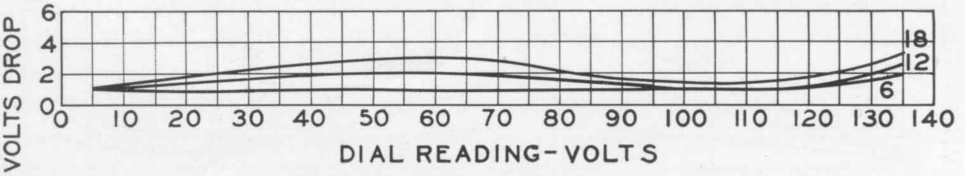
115 volts in; 0 to 230 volts out

TYPE 100-Q



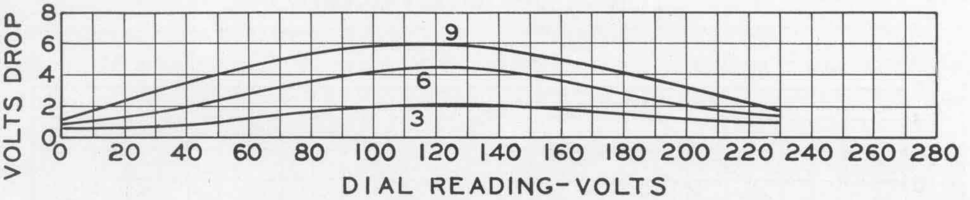
115 volts in; 0 to 115 volts out

TYPE 100-Q (continued)

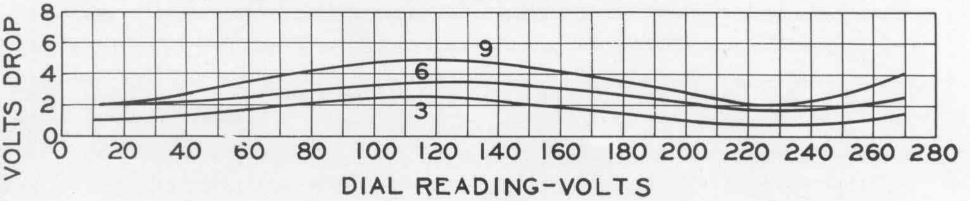


115 volts in; 0 to 135 volts out

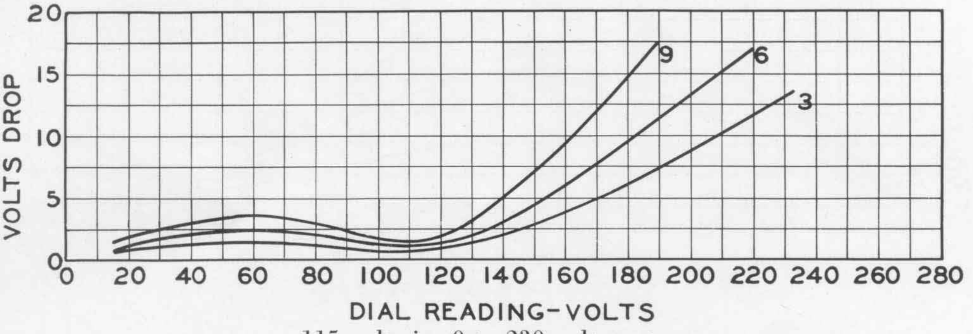
TYPE 100-R



230 volts in; 0 to 230 volts out

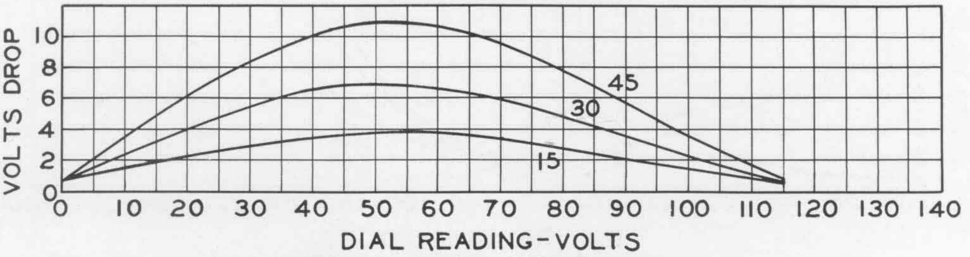


230 volts in; 0 to 270 volts out



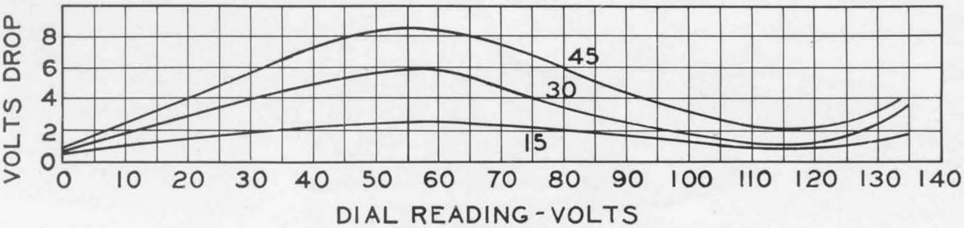
115 volts in; 0 to 230 volts out

TYPE 50-A



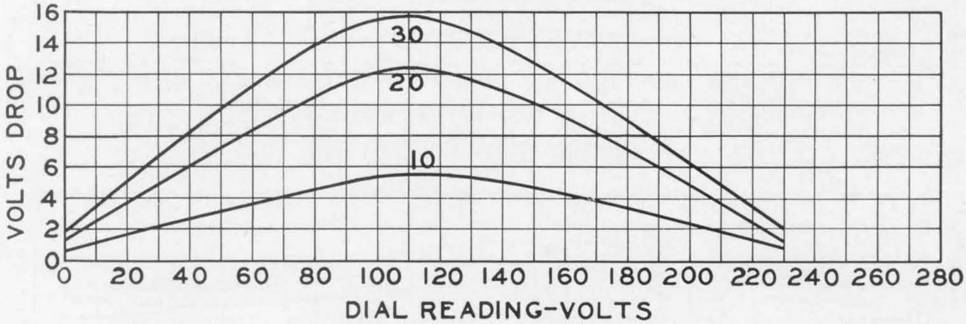
115 volts in; 0 to 115 volts out

TYPE 50-A (continued)

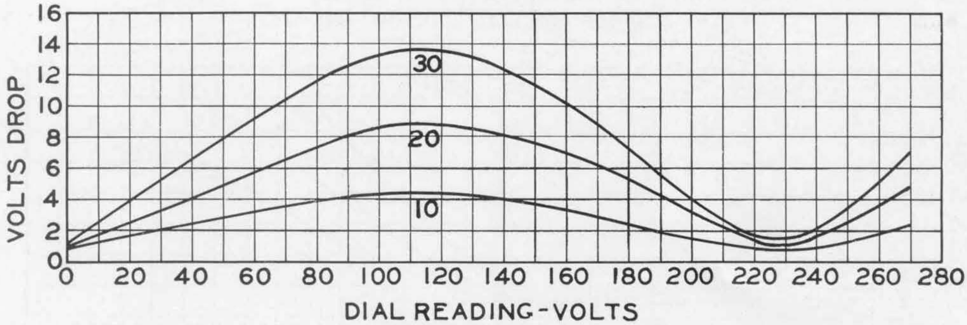


115 volts in; 0 to 135 volts out

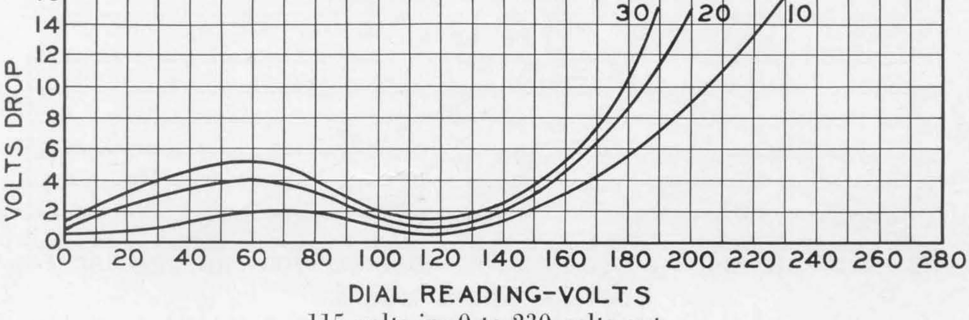
TYPE 50-B



230 volts in; 0 to 230 volts out



230 volts in; 0 to 270 volts out



115 volts in; 0 to 230 volts out

EFFECTS OF SUBSTITUTE MATERIALS

INSTRUMENT PERFORMANCE

● **IN THE DESIGN** of laboratory and communication test equipment the use of certain metals of construction is necessary for engineering reasons. Some of these metals, notably aluminum and brass, have become extremely scarce because of war conditions. Frequently, there is no substitute for the scarce materials, but in other cases zinc can be used for aluminum, and sheet steel for brass. Whenever possible the General Radio Company, in order to conserve the critical metals, uses the more plentiful substitute. These substitutes are selected to have as little effect upon the electrical performance characteristics of the equipment as possible. Generally speaking, the substitution of construction metals makes no difference in the electrical performance, but increases the weight substantially.

To achieve the maximum electrical performance the electrical components of General Radio instruments are carefully selected and are held to close manufacturing tolerances. The quantity production of component parts for war purposes has resulted in an inevitable loosening of production tolerances. Because of these wartime conditions and governmental regulations we are not always able to obtain components like resistors and capacitors to the close tolerances that we would like and must use them with wider tolerances. Vacuum tubes can no longer be selected or paired as they have been for some instruments. These factors sometimes adversely affect performance, but not enough to limit seriously the usefulness of the instrument.

A substantial part of our current engineering work is directed toward keeping instrument performance up to the highest standards in spite of the inferior materials that must be used even on the most vital war jobs. We have so far succeeded in maintaining General Radio performance standards and hope to be able to continue to do so, but if minor defects in appearance or performance occur, we hope our customers will understand that the causes are beyond our control.

FEWER ACCESSORIES

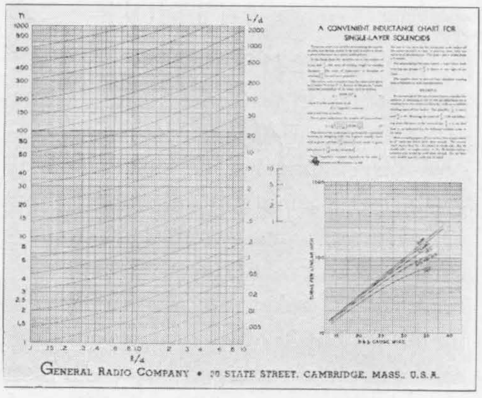
One obvious step in the conservation of scarce materials is the elimination of unnecessary spare parts and accessories.

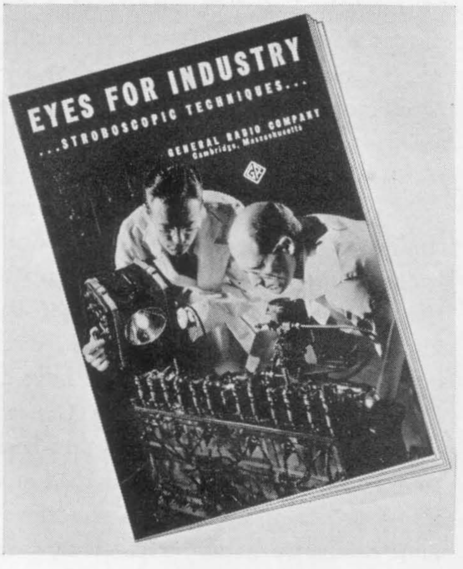
The Type 274-M Double Plug, for instance, is made from polystyrene, brass, and beryllium copper, all of which are scarce materials. A convenient but not essential accessory, this plug is no longer included in shipments of instruments with which it was formerly supplied free.

Other items no longer supplied are the metal shield cans for Type 510 and Type 668 Decade Resistance Units, the spare brushes formerly included as standard accessories with shipments of Variacs, and the replacement cables and power cords mentioned in last month's *Experimenter*.

ENLARGED INDUCTANCE CHART

The inductance chart for single-layer solenoids, originally published in the August, 1940, issue of the *Experimenter*, is now available in enlarged form, 17 x 22 inches overall, suitable for wall mounting. This chart indicates the number of turns required for a given inductance in terms of the length and diameter of the winding form. Write for a copy if you can use it. No charge, of course.





NEW BOOKLET ON STROBOSCOPE USES

This recently published 32-page booklet illustrates the use of General Radio stroboscopes in many branches of American industry. Slow-motion observations, speed measurements, and stroboscopic photography are discussed in detail. A copy is yours for the asking. This booklet will help you to get the maximum usefulness from your Strobotac and Strobolux.

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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