the GENERAL®RADIO Experimenter

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TO

GENERAL RADIO EXPERIMENTER

VOLUMES XXII AND XXIII

June, 1947 to May, 1949

GENERAL RADIO COMPANY
CAMBRIDGE MASSACHUSETTS
U. S. A.

I N D E X
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Volumes XXII and XXIII, June 1947 through May 1949

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THE General Radio EXPERIMENTER

VOLUME XXII No. 8 JANUARY, 1948

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

SENSITIVITY OF THE TYPE 916-A RADIO-FREQUENCY BRIDGE

Also IN THIS ISSUE

● **ALTHOUGH IN MOST APPLICATIONS** satisfactory performance is obtained from the Type 916-A Radio-Frequency Bridge when commercial signal generators and receivers are used, in some cases difficulty is encountered from apparently low bridge sensitivity or from noise and other extraneous signals picked up by the circuit under test. To those who have experienced this difficulty, the following analysis may be helpful in suggesting a solution.

In order to discuss bridge sensitivity, the term first must be defined. The definition used in this article is based on the following reasoning. With both the resistance and reactance dials set at the true balance positions, the voltage developed across the detector terminals is, of course, zero. However, if one dial is displaced from its true balance position while the other remains at its true balance position, the output voltage produced across the detector terminals is directly proportional to the deviation from the balance setting for small

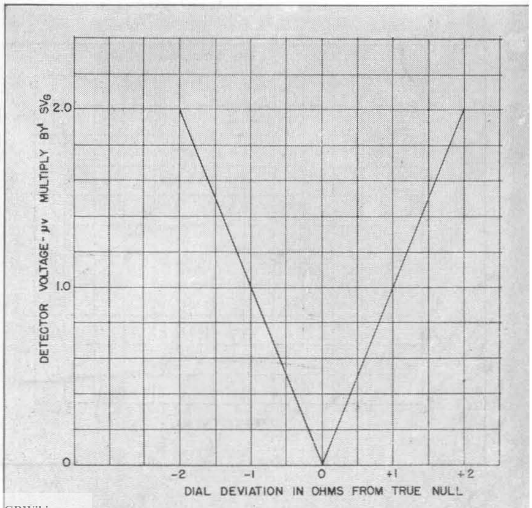


Figure 1. Voltage across the detector terminals of the r-f bridge as a function of the deviation of either dial from the true null position.

deviations. Therefore, a bridge sensitivity factor, S, can be defined as:

$$S = \frac{\frac{V_d}{V_g}}{\Delta}$$

where V_g is the generator voltage in volts, and V_d is the voltage in microvolts developed across the detector terminals when either dial is displaced from its true balance position by the amount Δ, in ohms, and the other dial is set at its true balance point.

The relationship between the output voltage produced across the detector terminals, in terms of the sensitivity factor and applied voltage, and the deviation from the true balance position in ohms is plotted in Figure 1. If the sensitivity factor, the applied voltage, and the detector sensitivity are known, the limitation on the bridge accuracy due to sensitivity can be determined from the graph.

The bridge sensitivity factor for an infinite detector impedance, S_o, is plotted as a function of frequency in Figure 2 for both the *L* and *C* positions of the *L-C* switch on the bridge. Of course, in the practical case, the detector impedance is not infinite, and in order to obtain the true sensitivity factor, S, the open circuit sensitivity factor, S_o, must be multiplied by $\left|\frac{Z_d}{Z_d + Z_o}\right|$ where Z_d is the detector impedance, and Z_o is the output impedance of the bridge. The resistive and reactive components of the output impedance are plotted as a function of frequency in Figure 3.

The voltage developed across the bridge generator terminals, V_g, is not equal to the open circuit voltage produced by the signal generator used as a signal source unless the generator output impedance is zero as the input impedance to the bridge is not infinite. To obtain the magnitude of the applied voltage, the open circuit generator voltage must be multiplied by the factor $\left|\frac{Z_i}{Z_i + Z_g}\right|$ where Z_i is the input impedance to the bridge and Z_g is the output impedance of the generator. The input impedance to the bridge is plotted as a function of frequency in Figure 3.

The sensitivity factor is not independent of the magnitude of the resistive component of the unknown impedance but decreases slowly as the resistive component increases. The sensitivity factor plotted in Figure 2 is for relatively small resistances; however, the effect of the uncertainty of balance upon the measurement of the resistive component expressed as a percent of the unknown resistance decreases as the magnitude of the resistive component increases.

In some applications, and particularly in the measurement of broadcast antennas, extraneous signals and noise picked up by the circuit under test and introduced into the bridge tend to decrease the measurement accuracy by partially masking the null. This effect can be serious when the maximum voltage available for application to the generator terminals is low. The use of a

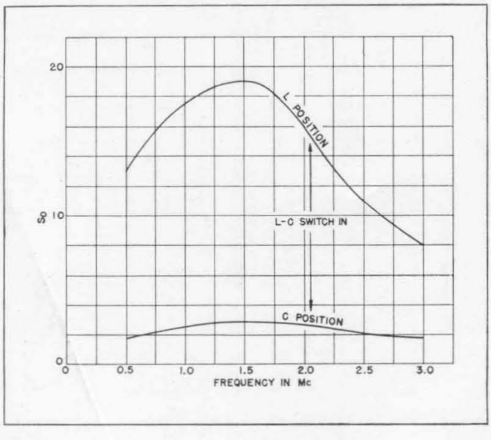


Figure 2. Open-circuit sensitivity factor, S_o, as a function of frequency.

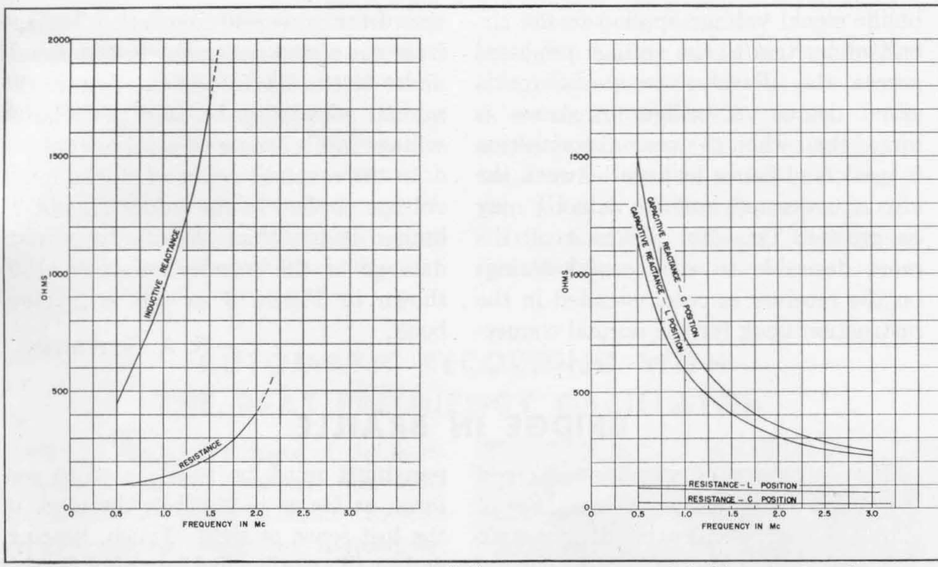
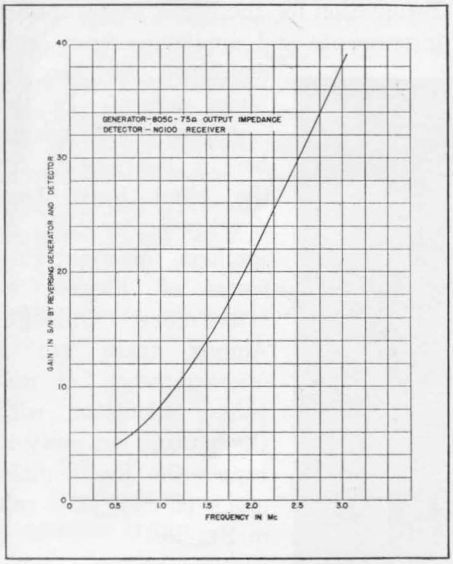


Figure 3. Resistive and reactive components of (*left*) input impedance and (*right*) output impedance of the bridge as a function of frequency.

highly selective receiver is helpful, since it amplifies a smaller amount of the undesired signal than does a receiver with poor selectivity. Ordinary communications receivers with crystal filters are desirable for this application.

Figure 4. Gain in signal-to-noise ratio as a function of frequency when generator and detector connections are interchanged.



Another method of reducing the effect of noise introduced into the bridge by the circuit under test is to reverse the generator and detector connections. In the normal connection, the bridge isolating transformer is connected between the generator and the bridge. In this position the insertion loss of the transformer attenuates only the signal applied to the bridge. In reversed connection, the transformer is between the bridge and the detector and both the desired signal and noise from the unknown are attenuated by it, resulting, in general, in an increase in the signal-to-noise ratio produced in the detector. The amount of increase is appreciably affected by the impedances of the signal generator and detector. In a typical arrangement, using a General Radio Type 805 Signal Generator and a National Type NC 100 Receiver, the increase in signal-to-noise ratio produced by reversing the connections is plotted as a function of frequency in Figure 4.

In the reversed connection, the ratio

of the signal voltage applied to the circuit under test to the voltage produced across the detector terminals for a given degree of bridge unbalance is larger than when the normal connection is used, and hence leakage between the circuit under test and the detector may be greater. This fact makes it all the more desirable to use coaxial fittings on the receiver as recommended in the instruction book for the normal connection. In the reversed connection, leakage from the signal generator to the circuit under test is less important than in the normal connection because the relative voltage levels are more nearly equal.

In the reversed connection, the input voltage applied to the bridge should be limited to less than 15 volts to prevent damage to the standard resistor, R-3, shown in Figure 8 in the instruction book.

—R. A. SODERMAN

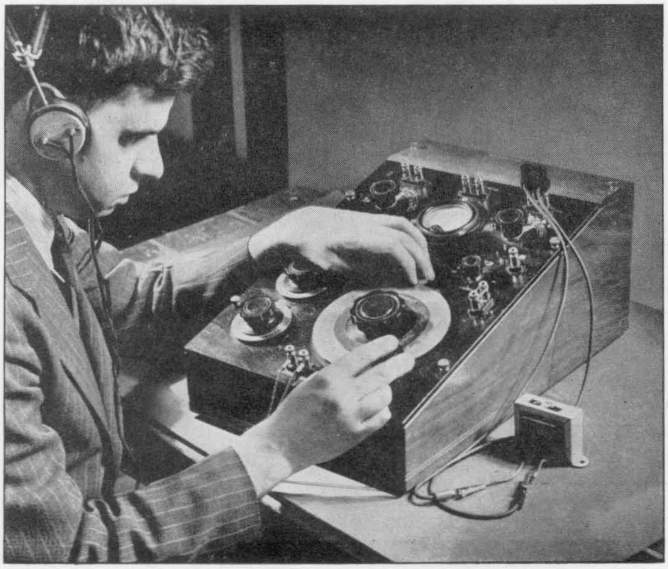
BRIDGE IN BRAILLE

The casualties of war have focused increasing attention on the problems of aiding the physically handicapped to find a satisfactory niche in industry and in society. With most types of disability, the solution is not too difficult: artificial limbs are being perfected for the legless and armless; plastic surgery restores lost features; and only the restriction to a sedentary life is necessary for many of those whose handicap is functional.

With the blind, however, the solution is more difficult The loss of what is, for many purposes, the most used of the five senses means that others of the four remaining must be re-educated to perform, as far as is possible, the work of the lost sense of sight. Touch, hearing, and smell are all called into play for this purpose, and in the education of the blind are developed to an amazing degree.

For written communication, and for utilizing the various instruments and machines that implement contemporary existence, the sense of touch is the one most used.

Under the auspices of the American Foundation for the Blind, many special instruments and appliances have been developed for use by blind persons. In the engineering field, a number of instruments for the blind have been developed by T. A. Benham, Assistant Professor of Physics at Haverford College. Among these are a microammeter, a machine calculator with the housing cut away to expose the Braille dials, and a circular slide rule in Braille.



One of Professor Benham's developments that is of particular interest to the General Radio Company is the modified Type 650-A Impedance Bridge shown in the photograph. This bridge has been fitted with celluloid dials with Braille figures, and the dials are so mounted that they do not interfere with the normal operation of the bridge.

Professor Benham has been blind since the age of two and has devised this dial system for his own use. A more complete discussion of the work being done in this field will be found in his article "Aids for the Blind," which appeared in the February, 1947, issue of Electrical Engineering.

AUTOMATIC RECORDING WITH THE BEAT-FREQUENCY OSCILLATOR

Of the many advantages of the beat-frequency type of oscillator, the outstanding is probably its ability to scan rapidly a wide range of frequencies. This feature is particularly valuable in connection with automatic graphic recording arrangements where frequency-response data or other information is to be obtained in the form of a permanent and graphic record. The single-control, single-sweep arrangement of the beat-frequency oscillator is readily adapted to mechanical drive from, or in conjunction with, a recorder mechanism.

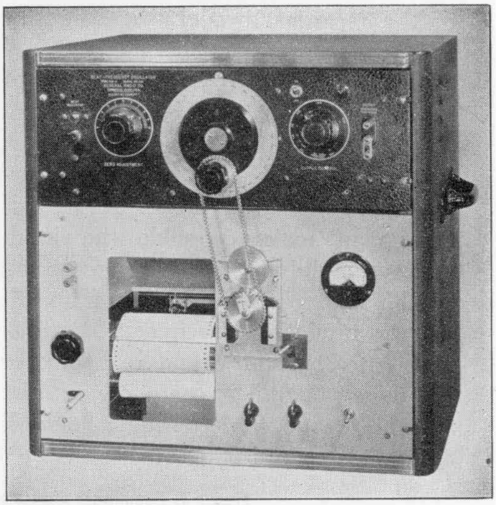
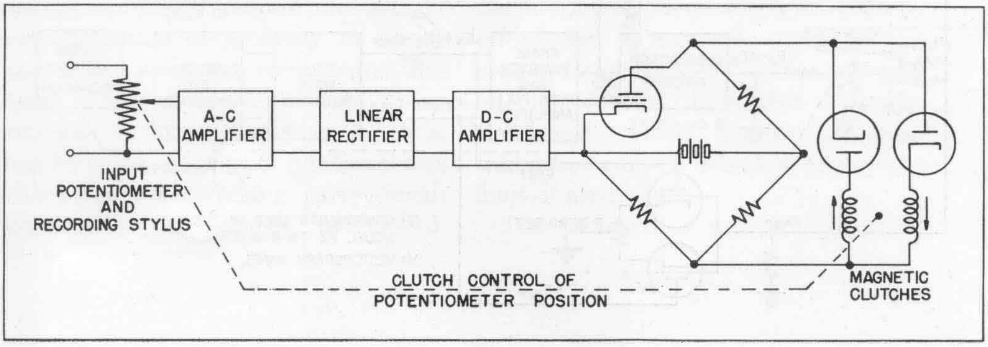


Figure 2. View of the Model FR-1 Graphic Recorder with the Type 913-C Beat-Frequency Oscillator.

For several years, engineers of the Sound Apparatus Company of New York have been engaged in a program of developing methods of graphic recording at audio frequencies. They have chosen the General Radio Type 913 Beat-Frequency Oscillator as the most suitable signal source, and a drive mechanism and chart are available for using this oscillator in conjunction with their recorders.

Figure 1. Functional diagram of the recorder.



The basis of operation of the recorder is readily understood with reference to Figure 1. It will be seen that the system is a servo mechanism, which maintains at a predetermined level the signal input to the a-c amplifier. The position of the potentiometer arm and recording stylus is thus proportional to the signal level impressed on the input potentiometer.

Figure 2 shows the combination of the 913-C and the Sound Apparatus Company's Model FR-1 Graphic Recorder. The oscillator dial drive is chain coupled to the paper-drive mechanism of the recorder through the link unit shown.

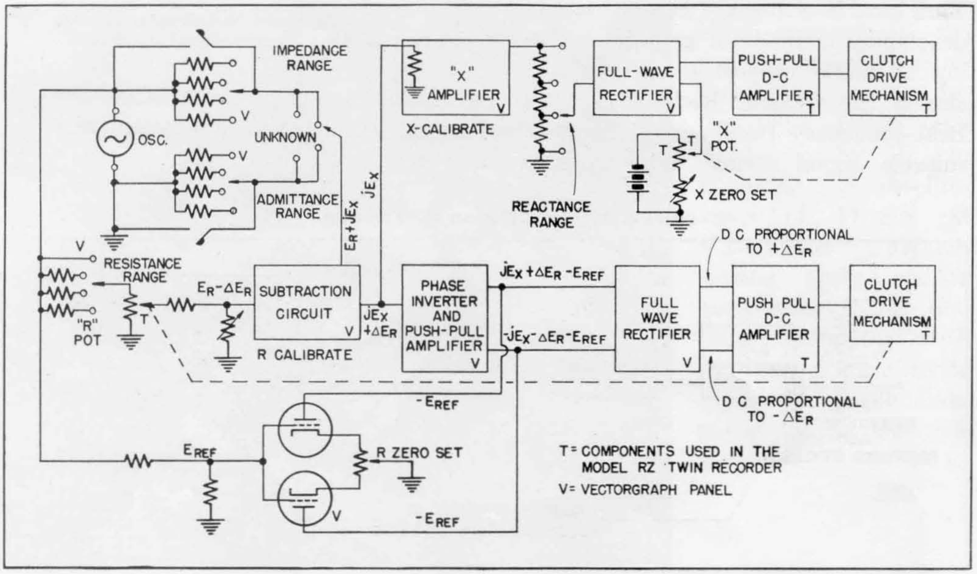
A recently announced development of particular interest is the "Impedance Vectorgraph." Incorporating a twin recorder, Model RX, and the beat-frequency oscillator, Model 913-C, this equipment makes possible the simultaneous and fully automatic recording of the separate components of impedance as a function of frequency. Equivalent series resistance and reactance or equivalent parallel conductance and susceptance of a two-terminal network or component can be plotted.

Essentially the method consists of maintaining a constant current through the unknown impedance and of recording the complex components of voltage developed across it. The real and imaginary components of this voltage are directly proportional to the resistance and reactance respectively of the unknown impedance. Referring to the functional diagram of Figure 3, the operation of the system is as follows: For impedance measurements, the Admittance Range Switch is short-circuited and current fed to the unknown through the Impedance Range Resistors. These are of such magnitude as to insure a current that is independent of the unknown impedance.

The "R" Potentiometer inserts into the subtraction circuit a signal which cancels the resistive component of voltage E_R with the exception of the error voltage ΔE_R. After phase inverting this error voltage ΔE_R and the reactive voltage jE_x, these voltages are added to the reference voltage E_{REF} and rectified.

The output of the rectifier yields a d-c voltage proportional to $+\Delta E_R$ and

Figure 3. Functional diagram of the Impedance Vectorgraph.



a d-c voltage proportional to $-\Delta E_R$. If the error voltage is reversed in phase, the voltages will be reversed, giving a push-pull error voltage output. This error voltage is amplified through a push-pull d-c amplifier, which operates a clutch mechanism to move the slider on the "R" Potentiometer to reduce the error voltage. Automatically, this "R" Potentiometer plots the resistive component of the impedance.

After the subtraction circuit there now remains only jE_x, the reactive voltage which is amplified and rectified. This rectified output is cancelled out by the "X" Potentiometer output. The error in cancellation is amplified through a d-c amplifier which operates a clutch mechanism to move the slider on the "X" Potentiometer to reduce the error. Automatically, then, this "X" Potentiometer plots the reactive component of the impedance.

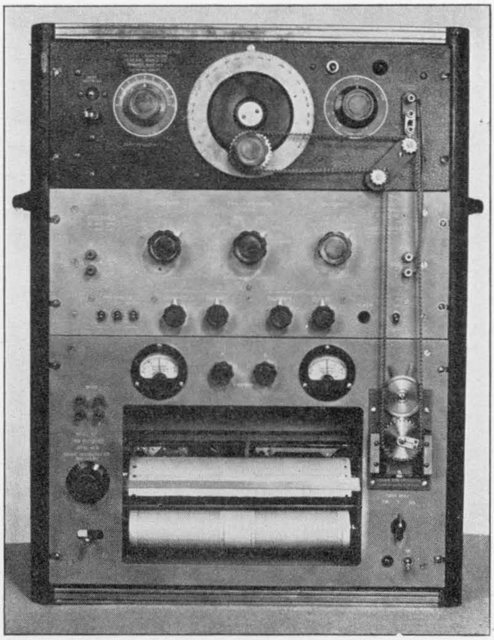


Figure 4. Panel view of the Impedance Vectorgraph.

The measurement of admittance is accomplished by shorting the impedance range resistors, A, and inserting the admittance range resistors into the circuit. Essentially constant voltage is now impressed across the unknown. The input to the subtraction circuit is now switched from the top of the unknown Z to the bottom of the unknown, placing the voltage across the small admittance range resistors into the subtraction circuit. This voltage is proportional to the current through the unknown both in magnitude and phase. These in-phase and out-of-phase components of voltage are then proportional to the conductance and susceptance respectively and are recorded similarly to the impedance components.

—L. P. Reitz
I. G. Easton

ECCENTRICITY EFFECTS IN PRECISION ROTARY DEVICES

Technical advances embodied in modern communication, laboratory, and process equipment require increasingly high standards of accuracy in rotary control and indicating components. Residual inaccuracies result from many contributing factors. Large among these may be an eccentricity of rotational axis with respect to working parts. Small eccentricities, sometimes overlooked, may often offset much or all of the care lavished on the construction of a precision control. Even when eccentricity errors are recognized, they cannot always be eliminated, as, for example, on the General Radio 722 Precision Condenser, which requires a worm calibration curve if it is to be used to the limit of accuracy.

The presence of eccentricity causes

the indicated rotation to differ from true rotation by an angle the size of which is a function of the eccentricity. In a device with working parts of true radius, R, with a shaft center displacement, E, from the origin of R, the indicated angle of rotation, β, will differ from the true angle, α, as expressed by the equation:

$$\beta - \alpha = \sin^{-1}\left[\frac{E}{R}\sin\beta\right].$$

Put into words, this says that the angle of error is that angle the sine of which is the product of the eccentricity to radius ratio multiplied by the sine of the indicated angle of rotation.

Since, with small angles, the sine of the angle closely approximates the angle, it is possible to write an approximate form of the equation, thus:

$$\beta - \alpha \approxeq 57.3\,\frac{E}{R}\sin\beta.$$

For values of E/R less than 0.1, the error introduced by the approximation is negligible, and no good design, certainly, should exceed a ten per cent eccentricity ratio.

The importance of the eccentricity factor can best be shown by a typical example. Consider a precision voltage-divider ("potentiometer") of two-inch radius, with an eccentricity of but twenty one-thousandths of an inch, and a total electrical rotation of 320°. Maximum eccentricity error occurs at 90° or 180° indicated rotation, when sine β is plus or minus one. This maximum error is:

$$\beta - \alpha = 57.3 \times \frac{.020''}{2.000''} \times \pm 1 = \pm 0.573^\circ$$

which yields a percentage error of

$$\frac{\pm 0.573^\circ}{320^\circ} \times 100 = \pm 0.18\%$$

instead of the 0.05% that might have been expected from the two thousand carefully spaced turns of the winding. Thus, in effect, the eccentricity error has nullified the accuracy inherent in the precision winding and contact arrangement.

Sometimes it is possible to use eccentricity deliberately to offset other nonlinearities; if such other errors are sinusoidal or approximately sinusoidal with rotation. Under these circumstances, the introduction of a calculated eccentricity of the proper phase and magnitude can improve over-all linearity materially. For instance, the regulation characteristic of a Variac, while not sinusoidal, can be partially offset by the deliberate introduction of eccentricity that is a function of load current. During World War II, this principle solved an annoying problem in connection with alternating current supply of a magnetic deflection coil operated from a Variac used as an alternating-current potentiometer. The corrective eccentricity offset the regulation error sufficiently to meet the narrow resolution specification required.

—Gilbert Smiley

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THE General Radio EXPERIMENTER

VOLUME XXII No. 9 FEBRUARY, 1948

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A BRIDGE-TYPE AUDIO-FREQUENCY METER

Also IN THIS ISSUE

● **DURING THE WAR** it was found necessary to drop from our line of instruments the popular Type 434-B Audio-Frequency Meter, along with a number of other items, in order to concentrate our productive capacity on fewer instrument types. The audio-frequency meter is now back with a four-digit type number, Type 1141-A, and changes in internal construction to improve its performance.

The accuracy of measurement of the Type 1141-A Audio-Frequency Meter is ±0.5% over the entire frequency range of the instrument. Since audio oscillators are usually calibrated to only 2%, the audio-frequency meter is extremely useful where more accurate calibrations and frequency measurements are necessary. The meter is also useful for measuring the audio beat between an unknown high frequency and a harmonic of a standard crystal oscillator. Since 10,000-cycle harmonics are provided throughout the low radio-frequency spectrum by means of multivibrators, the maximum audio beat frequency is 5000 cycles, and the maximum error in the frequency determination is 25 cycles.

The accuracy of measurement of

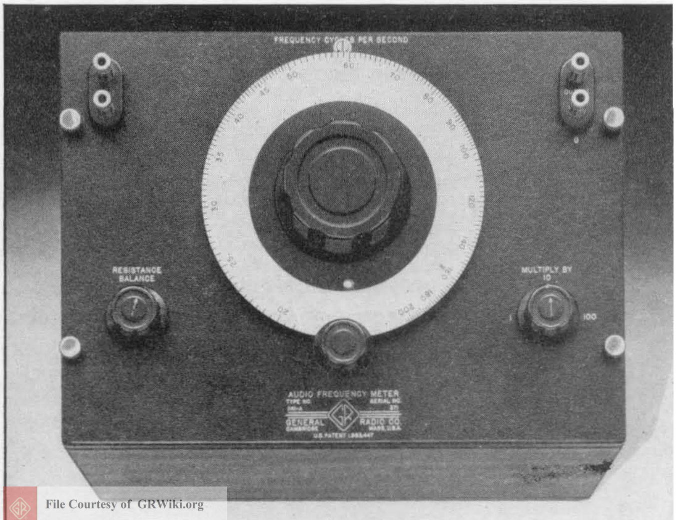

Figure 1. Panel view of the Type 1141-A Audio-Frequency Meter.

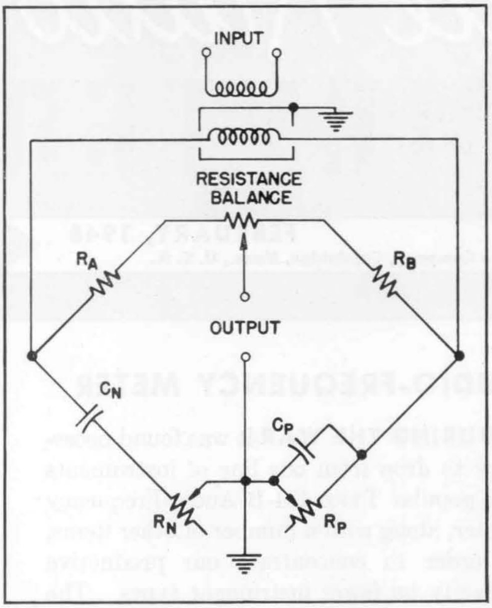

Figure 2. Elementary schematic circuit diagram of the frequency meter.

$\pm 0.5\%$ is made possible by individually calibrating the control dial, and by using extremely stable multiplier capacitors. These capacitors have mica dielectric for the smaller sizes and polystyrene tape dielectric for the large units.

The fundamental circuit for the Type 1141-A Audio-Frequency Meter is the Wien bridge, which uses only resistance and capacitance, and is frequency sensitive. The schematic diagram is shown in Figure 2, and the conditions of balance are given below:

$$f = \frac{1}{2\pi\sqrt{R_N R_P C_N C_P}} \tag{1}$$

and

$$\frac{C_P}{C_N} = \frac{R_A}{R_B} - \frac{R_N}{R_P}$$

In order to provide a single control upon which the frequency scale can be mounted, and also to maintain the second balance condition, the two resistors, R_N and R_P, and the two capacitors, C_N and C_P, are made equal, and the two ratio arms are made two to one, thus:

$$\frac{R_N}{R_P} = \frac{C_P}{C_N} = 1 \quad \text{and} \quad \frac{R_A}{R_B} = 2 \tag{2}$$

This arrangement* always fulfills the second balance conditions and reduces the first condition to:

$$f = \frac{1}{2\pi R_N C_N} \tag{3}$$

The two resistors, R_N and R_P, are wound on tapered cards of such shape that the frequency scale is logarithmic, and equal frequency ratios occupy equal intervals on the scale. Hence the fractional accuracy of reading is constant. There are fixed resistors in series with the variable parts of R_N and R_P having about one-tenth the value of the variable resistor, thus limiting the range of the frequency scale to a ratio of ten to one. The three frequency ranges, differing from one another by factors of ten, are obtained by the use of three sets of capacitors, C_N and C_P, which also differ by factors of ten, so the same engraved scale is used for all three ranges.

It is impractical to keep the resistors, R_N and R_P, and the capacitors, C_N and C_P, exactly equal as demanded by Equation (2). An auxiliary control, consisting of a small rheostat, is provided between the ratio arms *A* and *B* to whose sliding contact the null detector is connected. This control alters the effective ratio R_A/R_B and satisfies Equation (1). However, if this adjustment is not made, the null setting of the frequency dial is not altered, but merely dulled.

A shielded input transformer is provided in order to eliminate the effect of unbalanced capacitances to ground that may exist in the source being measured.

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* U. S. Patent No. 1,983,447.

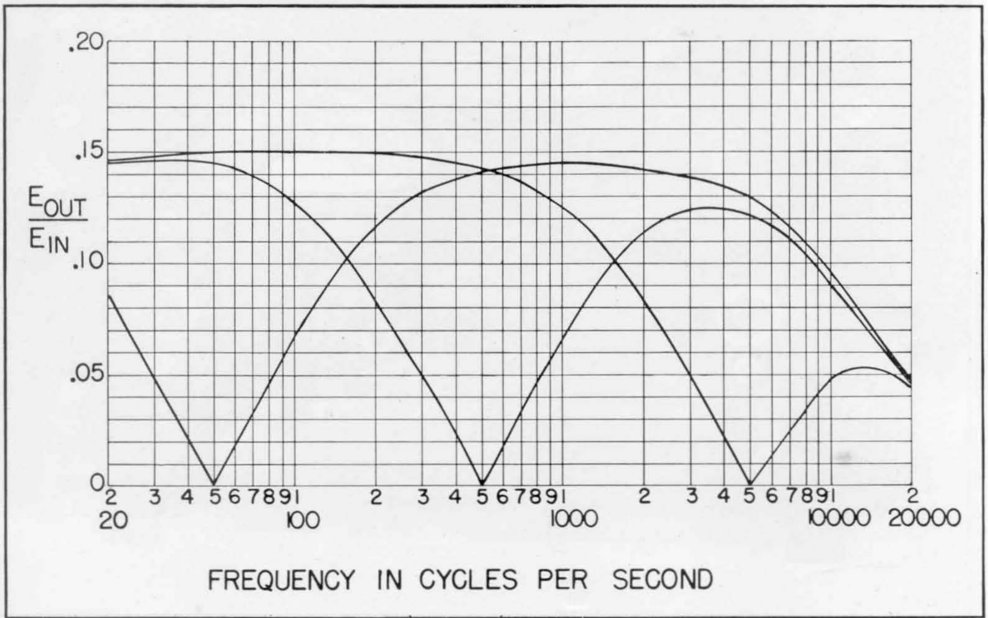
The null detector most often used for making measurements with the meter is a pair of head telephones. These are satisfactory in the frequency range from 300 to 5000 cycles, but for frequencies outside the range of head telephones a sensitive a-c voltmeter must be used. The TYPE 1231 Amplifier and Null Detector is well suited for this application and can also be used with head telephones to increase the sensitivity when the input voltage is low.

If the source of frequency to be measured contains harmonics, they will not be balanced out by the bridge and will be impressed on the head telephones or other detector. The human ear can discriminate against a considerable percentage of harmonics, but, if the harmonic content is high, the aid of a low-pass filter connected between the bridge and the telephones is necessary. If a voltmeter is used as the detector, the use of a filter is necessary when harmonics are present, since the voltmeter, unless tuned, lacks the power of discrimination between harmonics and fundamental.

The fact that the audio-frequency meter passes harmonics while eliminating the fundamental makes it possible to use the meter for measuring the distortion in an audio signal. The instrument also forms a convenient band-elimination filter which is adjustable over the entire audio spectrum. A plot of the discrimination characteristic is shown in Figure 3. The maximum ratio of output to input voltage is .167 because the transformer steps down two to one and the ratio arms are also in the ratio of two to one. Decreases from this value are caused by the characteristics of the transformer.

— MARTIN A. GILMAN

Figure 3. Frequency discrimination characteristic of the bridge-type frequency meter.



SPECIFICATIONS FOR TYPE 1141-A AUDIO-FREQUENCY METER

Frequency Range: 20 to 20,000 cycles in three ranges, 20 to 200 cycles, 200 to 2000 cycles, and 2000 to 20,000 cycles.

Accuracy: ±0.5% over the entire frequency range. The null point is sharp enough so that the dial can be set to 0.1% provided the waveform is reasonably pure and the supply voltage or detector sensitivity is sufficiently high to provide the necessary over-all sensitivity.

Dial: The 6-inch dial, which has a slow-motion drive, turns through an angle of about 320° giving a scale length of about 17 inches for each 10 to 1 frequency range. The total scale length is thus over 4 feet.

Input Impedance: 3 to 10 kilohms, the smaller value corresponding to the higher frequencies.

Input Voltage: 110 volts rms, maximum.

Output Impedance: 1 to 4 kilohms, the smaller value corresponding to the higher frequencies.

Controls: Frequency dial, range selector switch, and resistance-balance control.

Accessories Required: A null detector is needed to operate the meter. Head telephones, such as the Western Electric 1002-C, or an amplifier-meter combination, such as the Type 1231-B Amplifier and Null Detector, can be used. Even with head telephones an amplifier and filter section will prove useful.

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

Dimensions: (Length) 12 x (width) 8¾ x (height) 9 inches over-all.

Net Weight: 15¼ pounds.

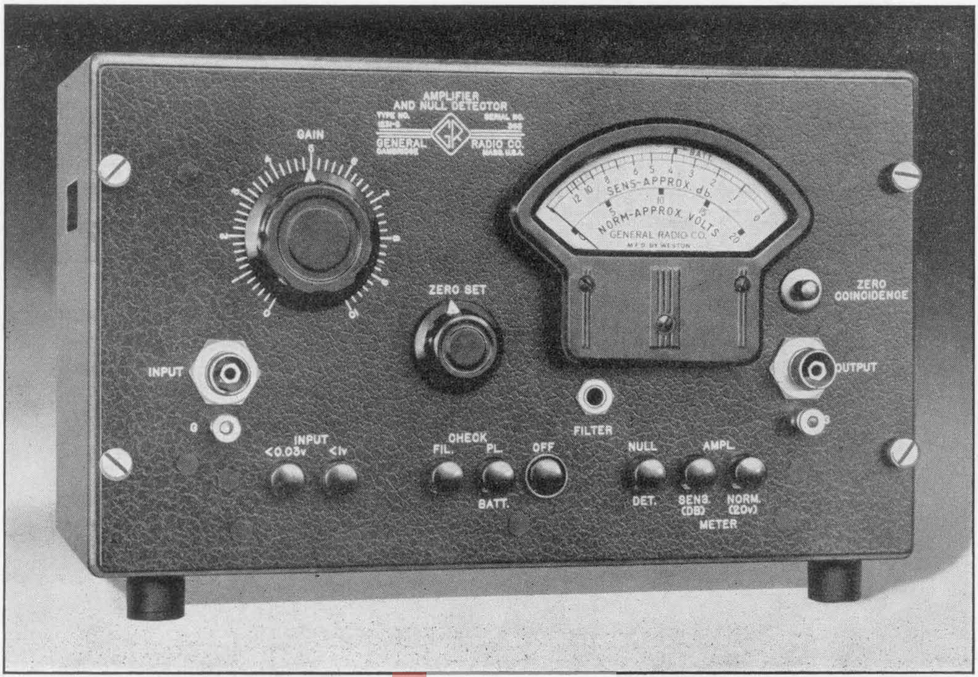
Type		*Code Word*	*Price*
1141-A	**Audio-Frequency Meter**	COLOR	**$215.00**

PATENT No. 1,983,447.

AN IMPROVED AMPLIFIER AND NULL DETECTOR

The Type 1231-B Amplifier and Null Detector is a battery-operated, resistance-coupled audio amplifier for use as a bridge detector, a standing-wave indicator, and a general-purpose laboratory amplifier. It has a built-in vacuum-tube voltmeter for measuring output voltage, and so for many applications no addi-

Figure 1. Panel view of the Type 1231-B Amplifier and Null Detector.



tional indicating device is needed. It can be operated either with the usual linear characteristic or with a semi-logarithmic characteristic to cover a wide range of voltage indication; during linear operation the voltmeter scale has two ranges, of 20 volts and about 2 volts respectively. A photograph of the instrument is shown in Figure 1.

The maximum open-circuit voltage gain is greater than 83 decibels at midband and has the frequency characteristic shown in Figure 2. Since the gain is greater than 70 decibels at 10 cycles and greater than 45 decibels at 100 kilocycles, the instrument is useful over this wide range for many bridge measurements. The input impedance is high, equivalent to 1 megohm in parallel with about 20 $\mu\mu$f, and the output impedance is about 50 kilohms resistive. Overloading of the last stage limits maximum output voltage to 20 volts for load impedances greater than 1 megohm and to 5 volts for a resistance load of 20 kilohms.

The simplified circuit diagram of Figure 3 shows that the amplifier has three stages and that the vacuum-tube voltmeter consists of a diode rectifier, a d-c amplifier, and the panel meter. A 30-decibel input attenuator selects a maximum input voltage rating of either 0.03 volt or 1 volt, and a tapered, wire-wound gain control allows continuous variation of gain over a wide range. During semi-logarithmic, null-detector operation, rectified d-c from the vacuum-tube voltmeter diode is applied as a gain-controlling bias to the last amplifier stage to produce the desired characteristic. An attenuator in the grid circuit of the d-c amplifier determines the scale range of the voltmeter, and a bridge-type plate circuit allows meter current to be zero with no signal at the instrument terminals.

The semi-logarithmic characteristic is intended primarily for null detector uses with bridges. Under this condition of operation less than 15 microvolts at the input terminals is required to produce a perceptible* movement of the meter needle at midband frequencies with the instrument set for maximum gain and no external filter. This input signal can then be increased by about 55 decibels before the meter reads full scale. Either the input attenuator or the gain control can be employed for higher input voltages. This sensitivity, combined with the wide range of voltage indication, means that the bridge balances can be made with precision and without gain adjustments in many instances.

In some situations, where the sensitivity of the null detector scale is not sufficient, the most sensitive range of the voltmeter can be used for a final

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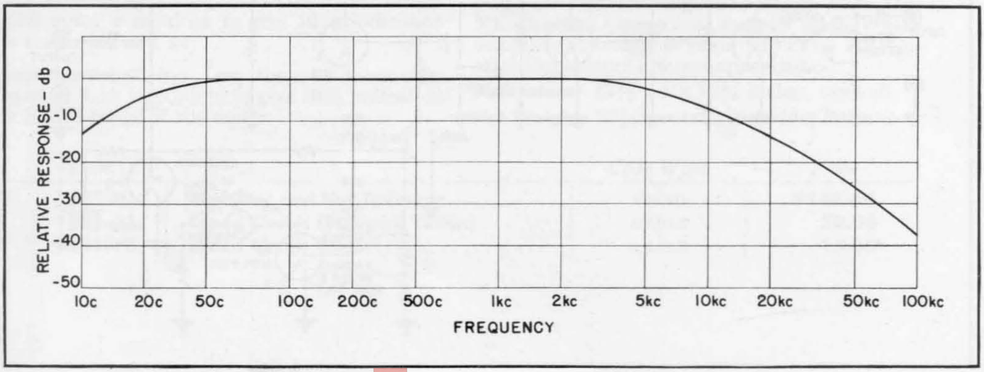
*"Perceptible" as used here means one-fifth of a one-sixteenth-inch division.

Figure 2. Frequency response characteristic of the amplifier.

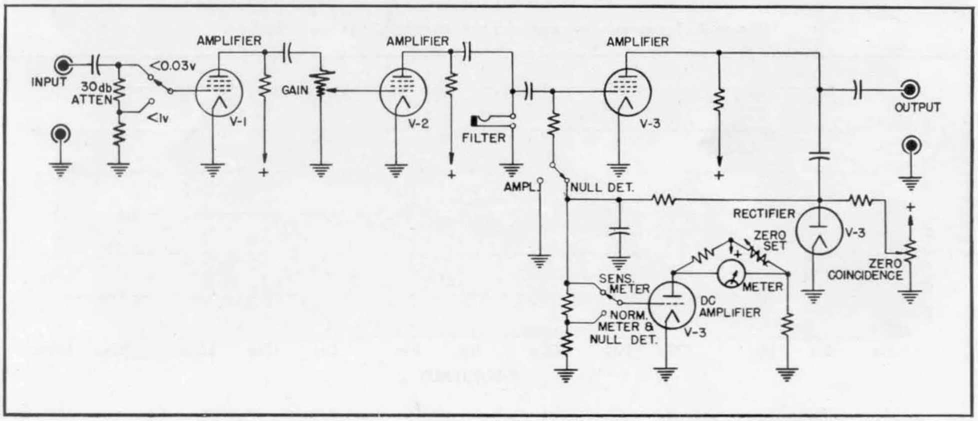


balance. The input signal required to produce a perceptible meter deflection is less than 5 microvolts for this range. For frequencies other than in the midband, sensitivity is less by the amounts shown in Figure 2. Head telephones can be used between 200 cycles and 10 kilocycles instead of the panel meter, according to individual preference, but there is no appreciable difference in sensitivity between the two methods.

One source of trouble in bridge measurements is the presence of unwanted voltages that obscure nulls. These voltages include random noise generated by the detector amplifier, power-frequency hum picked up by unshielded leads, and harmonics of the operating frequency that are produced either by non-linearity of the impedance being measured or by the generator itself. The noise level of the Type 1231-B is less than 15 microvolts referred to the input terminals, but this figure can, if necessary, effectively be reduced, along with hum and harmonics, by connection of an external filter through the panel jack provided. Filters thus connected are isolated by amplifier stages from the effects of varying input and output impedances and at the same time are at a high enough voltage level along the amplifying chain to be little affected by external fields. The Type 1231-P2 (400 and 1000 cycles) and Type 1231-P3 (60 cycles) Tuned Circuits are intended for use with the Type 1231-B. Any of these filters attenuates the second harmonic by about 20 decibels, and random noise by about 25 decibels; the Type 1231-P2 attenuates 60-cycle hum by about 35 decibels. The insertion loss of about 8 decibels caused by these filters is usually unimportant because of the high gain of the amplifier.

For measurements of impedance at ultra-high frequencies, a slotted, coaxial transmission line is often used with a traveling crystal detector to measure standing-wave ratios. If the u-h-f power source is pulsed, or otherwise amplitude modulated, at an audio-frequency rate, the rectified output of the crystal is an audio-frequency voltage. The Type 1231-B can be used to amplify and to indicate this voltage, and the sensitive range of the vacuum-tube voltmeter was incorporated into the instrument specifically for this application. The scale for this range has an approximate calibration in decibels. For more accurate measurements, a calibrated resistance attenuator can be inserted ahead of the Type 1231-B and the panel meter used only as a reference level indicator. Less than 100 microvolts is required at the amplifier input terminals to produce full-scale

Figure 3. Elementary schematic of the Type 1231-B Amplifier and Null Detector.



deflection on the sensitive meter range.

The Type 1231-B replaces the Type 1231-A and except for minor details is different only in the metering circuit. The use of a more sensitive, but still sturdy, 200 μa meter movement results in the high sensitivity of meter indication and makes the instrument suitable for standing-wave measurements, which is the application that initiated the redesign. However, the usefulness of the instrument as a bridge null detector has been greatly increased by the increased sensitivity, and an external indicating device is not needed.

The instrument is enclosed in a walnut cabinet, which also holds the battery. If desirable, the Type 1261-A Power Supply unit can be used to operate the Type 1231-B from 40 to 60-cycle lines and fits into the cabinet in place of the battery. Tubes are mounted on a shock-absorbing suspension to keep microphonic effects small. Push buttons are provided to operate the input attenuator, to set the condition of operation, and to select the meter-scale range. Other push buttons allow checking of the battery voltages on the panel meter. The input and output connections will take either General Radio Type 774-E Coaxial Connectors or the usual Type 274-M Plugs.

—W. R. Thurston

SPECIFICATIONS

Input Impedance: 1 megohm in parallel with 20 micromicrofarads.

Maximum Gain: Greater than 83 db at 1 kc with 1 megohm load.

Meter Scales: NORM scale: This scale is the one normally used to monitor the amplifier output voltage. It is calibrated approximately in volts with an accuracy of reading of ±5% of full scale.

SENS scale: This scale is used for determining ratios of voltages successively applied to the input terminals, as in standing-wave measurements. It is calibrated approximately in decibels with an arbitrary zero. Thus a ratio expressed in decibels is obtained by subtracting one meter reading from another. Ratios so obtained are accurate within 30% of the correct value in decibels, provided at least one of the readings is above half scale on the meter.

No separate scale is provided for NULL DET operation, since actual readings are not needed.

Null Detector Sensitivity: Less than 100 microvolts input is required to give 10% indication on the meter at 1 kc.

Amplifier Sensitivity: Less than 25 microvolts input at 1 kc is required to give 10% indication on SENS range of the meter.

Output Impedance: Approximately 50,000 ohms.

Maximum Output Voltage: 5 volts into 20,000 ohms; 20 volts into 1 megohm.

Noise and Hum Level: The open circuit noise level is less than 0.5 volt at full gain. When the Type 1261-A Power Supply is used, the open circuit noise and hum level is less than 1 volt.

Frequency Response: See Figure 2.

Tubes: The instrument requires two type 1L4 and one type 1D8-GT tubes, which are supplied in the instrument.

Power Supply: Burgess 6TA60 (Signal Corps BA48) Battery Pack is supplied in place in the instrument. When a-c supply is desired, the Type 1261-A Power Supply can be used.

Battery Life: Between 200 and 250 hours at 8 hours per day.

Accessories Available: Type 1231-P2 (400 and 1000 cycles) and Type 1231-P3 (60 cycles) Tuned Circuits are available for providing selectivity (see below). For facilitating connections to the input and output, two Type 274-M Plugs are supplied. Type 274-NC or Type 274-NE Shielded Connectors may be used. Where complete shielding is required, Type 774 Coaxial Connectors are recommended.

Dimensions: 12¼ x 8 x 10¾ inches, over-all.

Net Weight: 23¾ pounds, including batteries.

Type		*Code Word*	*Price*
1231-B	**Amplifier and Null Detector**	VALID	**$195.00**
1231-P2	**Tuned Circuit (400 and 1000c)**	AMBLE	**20.00**
1231-P3	**Tuned Circuit (60c)**	AMPLE	**15.00**

MISCELLANY

RECENT VISITORS to our plant and laboratories: Mr. Gunnar Hambraeus, Secretary of the State Research Council of Sweden, Stockholm; Professor E. K. Henriksen, Technical University of Copenhagen, Denmark; Professor Fu-Hsing Chu, National University of Chekiang, Hangchow, China.

THE AUDIO-FREQUENCY METER, TYPE 1141-A, was designed by Robert F. Field, of the development engineering staff, who designed its predecessor, the TYPE 434-B. The TYPE 1231-B Amplifier and Null Detector was designed by William R. Thurston, also of the development engineering staff, and author of the article in this issue.

I.R.E. CONVENTION. The 1948 National Convention of the Institute of Radio Engineers will be held in New York, March 22-25. Convention headquarters are at the Hotel Commodore and the Radio Engineering Show will be held at Grand Central Palace. Be sure to visit the General Radio exhibit in Booths 93 and 94. Many new products will be displayed, and General Radio engineers will be on hand to answer your questions. Among the completely new products that we plan to exhibit are a primary frequency standard and its associated frequency measuring equipment; capacitance test bridge; a standard signal generator for frequencies up to 50 Mc; a high-speed, high-intensity light source for stroboscopic work and photography; decade attenuators and decade inductors; and a line of high-quality parts, including switches, dials, terminals, connectors, and air capacitors.

A PAPER entitled "Recent Developments in Measuring Equipment" was presented by Ivan G. Easton of the New York District Office of the General Radio Company at I.R.E. and A.I.E.E. local sections in several southern cities last fall. His schedule included: Richmond, October 18; Memphis, October 21; Florence, Tennessee, October 22; Atlanta, October 24; Louisville, October 29; Charleston, West Virginia, October 30.

CORRECTION

The radio receiver used in obtaining data for the article entitled "Sensitivity of the R-F Bridge", by R. A. Soderman, in last month's *Experimenter* was incorrectly referred to as a National NC-100. The correct designation is NC-200.

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

ON THE EQUIVALENT CIRCUIT AND PERFORMANCE OF PLATED QUARTZ BARS

Also IN THIS ISSUE

● **ONE OF THE** lowest frequency modes of vibration in quartz is that of a long thin bar vibrating in the direction of its length. This led to the early use of this mode in constant frequency oscillators and in filters where frequencies of the order of 30 to 100 kc were desired. The early bars were all *X*-cut, that is, the faces of the bar were all parallel to the principal axes of the quartz crystal. Later developments have shown that rotation of the bar about the *X*-axis permits the choice of special properties with resulting improved performance in certain applications.

In the past, bars were generally operated at the fundamental frequency where the length of the bar is one-half wavelength. By choice of the lateral dimensions, the temperature coefficient of frequency could be reduced to a small value, averaging −0.75 parts per million per degree Centigrade for the TYPE 676-B bars. The use of adjustable baffles at the ends of the bar greatly reduced the damping caused by supersonic radiation and reduced changes in frequency caused by changes in air pressure. Since there was only one nodal region, at the center of the bar, mounting was somewhat difficult and there was a tendency for the bar to twist or displace in the mounting under shock.

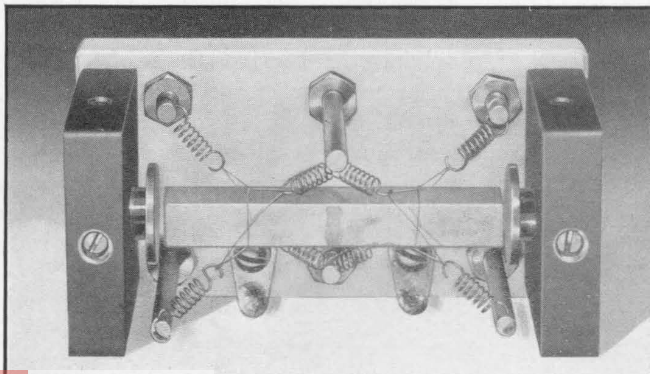

Figure 1. View of the Type 1190 Quartz Bar with cover removed. The bar is suspended in nylon filaments held in tension by coiled springs.

In the TYPE 1190-A Quartz Bar,[1] operation at the second harmonic, with a full wavelength, provides two nodal regions so that a much more stable mechanical arrangement results. Furthermore, utilizing quartz of about the same lateral dimensions as before, zero temperature coefficient of frequency can be obtained for a desired temperature. Finally, in a tension mounting, the mounting conditions can be maintained essentially unchanged over long periods, resulting in less frequency drift with time.

From the piezoelectric relations and the equations of motion, the electrical impedance and the equivalent electrical circuit can be derived for a bar, either plated or unplated. Since practically all such bars are used with plated electrodes, we will consider only that case.

First we must realize that quartz is anisotropic, that is, its properties are different in different directions. Second, that there are certain axes of symmetry in quartz about which the properties repeat. For rotation about any one of the three X-axes, the properties repeat every 180 degrees. For rotation about the Z-axis, or optic axis, the properties repeat every 120 degrees. A familiar example of a simple anisotropic material is wood, where the properties are different along and across the grain.

The six stress-strain equations of a perfectly anisotropic material involve 36 elastic constants. Because of the reciprocity relationship, the number of independent constants reduces to 21. Next, applying the considerations of symmetry for rotations about any X-axis, in quartz, 8 of the 21 constants become zero. Finally, applying the consideration of symmetry for rotation about the Z-axis, one more constant becomes zero and relationships are established between others such that the number of independent constants is reduced to six.

In general, additional stresses are introduced by the piezoelectric effect, so that the six general elastic equations are modified by additional terms involving 18 piezoelectric coefficients, which relate the stresses with the three component electric fields. Again applying symmetry conditions, the number of independent piezoelectric coefficients is reduced to two.

If we now consider an X-cut bar, with its length along the Y-axis, the piezoelectric equations reduce to:

$$-Y_y = s_{12}X_x + s_{11}Y_y + s_{13}Z_z - s_{14}Y_z + d_{11}E_x \qquad (1)$$

$$Q_x = E_xK/4\pi - d_{11}X_x + d_{11}Y_y - d_{14}Y_z \qquad (2)$$

where:

Y_y = longitudinal strain = change in length per unit length = dl/l.
X_x, Y_y, Z_z = The three longitudinal stresses = force per unit area = dynes/cm^2.
Y_z = shearing stress = dynes/cm^2.
d_{11}, d_{14} = piezoelectric constants = ESU charge/unit area/unit force = statcoulombs/cm^2/dyne or strain/unit field = cms/statvolt
E_x = applied field = voltage gradient = statvolts/cm.
s_{11}, etc. = elastic compliances = displacement/dyne = cm^2/dyne (These are the "zero field" values, sometimes written with a superscript E.)
Q_x = charge per unit area = statcoulombs/cm^2.
K = The "free" dielectric constant

To simplify these equations, we assume a bar having X and Z dimensions much smaller than the length, or Y, dimension, so that we can put

$$X_x = Z_z = Y_z = 0 \qquad (3)$$

Equations (1) and (2) then become

$$-Y_y = s_{11}Y_y + d_{11}E_x \qquad (4)$$

$$Q_x = E_xK/4\pi + d_{11}Y_y \qquad (5)$$

This obviously artificial simplification permits a solution to be obtained for the equivalent circuit. Since the length dimension Y is predominant, this theoretical solution will serve as a useful

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[1] Used in the new TYPE 1100-A Frequency Standards.

guide in practical cases. It would be expected that, as the lateral dimensions X and Z are increased, the departures from theory would become more pronounced. These departures are not so great as to limit the usefulness of the simplified interpretation until the lateral dimensions reach some 25 per cent of the length.

Equation (1) shows that, in addition to the desired lengthwise (Y) vibration, there will be vibrations in the directions of the width (Z) and thickness (X); also there will be a shearing vibration in the length-width plane (YZ). This condition may be thought of as a principal lengthwise vibration of the desired frequency having three other modes of vibration coupled to it. The effects of the other frequencies on the desired frequency will depend on the values of the frequencies (determined by dimensions and properties of the quartz in those directions) and on the magnitudes of the couplings between these different frequencies (determined by the properties of the quartz).

In general, if two modes of motion are in planes at right angles to each other, the coupling is weaker than if the motions are in the same plane. Also, if the modes of motion are such as to produce similar displacements at the boundaries, the coupling between them is much greater than when the boundary displacements are dissimilar.

In Equation (1), the frequencies of the thickness (X) and width (Z) longitudinal vibrations will be much higher than the desired length (Y) vibration. The frequency of the length-width (YZ) shear will decrease rapidly as the width is increased, approaching the length (Y) frequency, and the coupled effect becomes greater. Interpreted in this way, the simplification of Equation (1) into Equation (4) is accomplished by making the coupled frequencies so high compared with the lengthwise (Y) frequency that their effects are negligible.

It can be shown that the admittance of the crystal is:

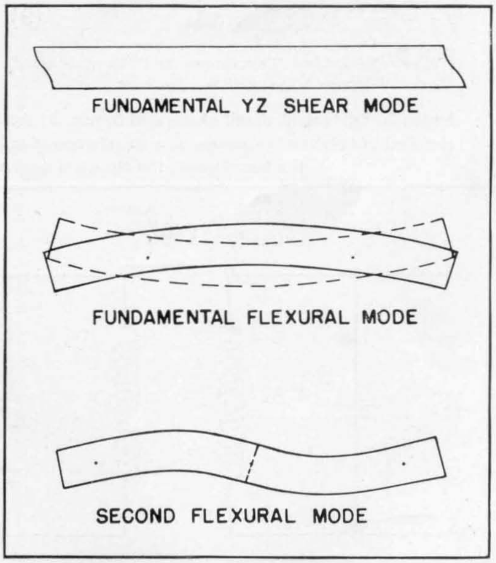
$$\frac{i}{E} = \frac{i}{E_x X} \tag{6}$$

$$= -j\frac{\omega E_x Y Z}{E_x X}\left\{\frac{K}{4\pi} + \frac{d_{11}^2}{s_{11}}\frac{\tan\frac{\omega Y}{2v}}{\frac{\omega Y}{2v}}\right\}$$

where v is the velocity of wave propagation in quartz.

This consists of two terms, representing two branches in the equivalent circuit. The first represents the capacitance C_0 of the two electrodes separated by the quartz. The second term could be represented by the reactance of a transmission line. Usually, however, the impedance of this branch is represented by an inductance L_1 and capacitance C_1 in series, resonant at the resonant frequency of the quartz and having a

Figure 2. Secondary modes of vibration in the bar. The shearing vibration is excited by the normal electrode. The second flexural mode is mechanically excited by the similarity of displacement at the ends. The fundamental flexural mode cannot be excited by these electrodes or displacements.



reactance slope at this frequency equal to that of the tangent curve.

The inductance value can be written

$$L_1 = \rho \frac{XYZ}{2}\left[\frac{s_{11}}{2d_{11}Z}\right]^2 \times 9 \times 10^{11} \text{ henries} \qquad (7)$$

The effective mass in the mechanical system, of a bar vibrating in the direction of its length, is one-half its actual mass. Consequently, in (7) the first factor represents the effective mass, or mechanical inductance, in the mechanical system. The second factor represents the transformation from the mechanical to the electrical system, and, to communications engineers, its representation as an electromechanical transformer with a certain transformation ratio is appealing.[2] The third factor is the conversion from electrostatic to practical units.

Similarly, for the capacitance:

$$C_1 = \frac{2}{\pi^2}\frac{Ys_{11}}{XZ}\left[\frac{2d_{11}Z}{s_{11}}\right]^2 \frac{1}{9 \times 10^{11}} \text{ farads} \qquad (8)$$

On this basis, the equivalent circuit can be drawn as shown in Figure Z.

If the mechanical elements are "taken through" the electromechanical transformer, we have

$$L_1 = \frac{XY}{Z} 106.8 \text{ henries} \qquad (9)$$

$$C_1 = \frac{ZY}{X} 0.00322 \times 10^{-12} \text{ farads} \qquad (10)$$

where the dimensions X, Y, Z are in cms.

Throughout the above no account has been taken of losses in the mechanical system. Due to factors such as supersonic radiation, dissipation in the mounting, and losses in the quartz, a resistance element should be included in the mechanical system.

For a bar mounted in air, free of reflecting obstructions, the greatest energy loss is due to supersonic radiation from the ends of the bar. If the smallest dimension is greater than one-half wavelength in air, this loss can be calculated as follows:

At each end of the bar a mechanical resistance $R_m = (\rho v)_a XZ$ mechanical ohms is effective, where $(\rho v)_a$ is the specific radiation resistance of air and is equal to 43 mechanical ohms per square centimeter. In the mechanical system the two ends are in parallel, giving an effective resistance $R_m/2$. Taking this mechanical resistance through the electromechanical transformer, we obtain the electrical equivalent

$$R_{\text{rad}} = \frac{X}{Z} 6920 \text{ ohms} \qquad (11)$$

The ratio of resonant reactance to

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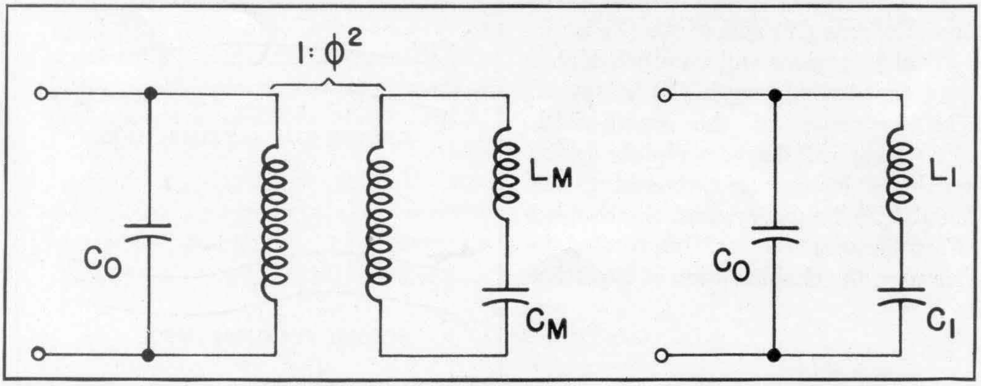
[2] "Electromechanical Transducers and Wave Filters," W. P. Mason, D. Van Nostrand Co., 1942.

Figure 3. Equivalent circuit of the quartz bar. At the left is shown the equivalent circuit with the electromechanical transformation shown as a transformer of ratio 1:ϕ^2. If the mechanical quantities are "taken through" the transformer, the electrical equivalent circuit shown at the right results.



resistance, or Q, of the crystal can be determined either in the mechanical system or the electrical equivalent. In the mechanical system, the resonant mechanical reactance and the mechanical resistance are:

$$Q_m = \frac{X_m}{R_m} = \frac{\omega_R Y \rho X Z}{4(\rho v)_a X Z} \qquad (12)$$

At resonance $\omega \; Y = \pi v_q$ so that

$$\begin{aligned} Q_m &= \frac{\pi(\rho v)_q}{4(\rho v)_a} \\ &= \frac{\pi\,(2.654 \times 5.44 \times 10^5)}{4 \times 43} \\ &= 26400 \qquad (13) \end{aligned}$$

In the electrical equivalent circuit we have from Equations (9) and (11) the resonant reactance and resistance giving

$$Q_e = \frac{182.5 \times 10^6 X/Z}{6920 X/Z} = 26400 \qquad (14)$$

Under these conditions it is seen that the Q is independent of the area of the ends of the bar, that is, a "thin" bar would have the same Q as a comparatively "fat" bar. However, operation under these conditions is of no practical interest since reflections would take place from objects near the crystal, and the losses are unnecessarily high.

One method of reducing the supersonic energy loss and of fixing the conditions of reflection near the crystal is to put baffle plates near the ends of the bar and adjust the position of the baffles for quarter-wave resonance.

For an X-cut 50-kc bar having the dimensions $X = 9.4$, $Y = 54.4$, $Z = 7.0$ mm, the resistance, when radiating freely, would be, from Equation (11), 9290 ohms. The actual measured resistance would be appreciably higher due to the effects of radiation from other surfaces, losses in the mounting, and the effects of coupled modes of motion.

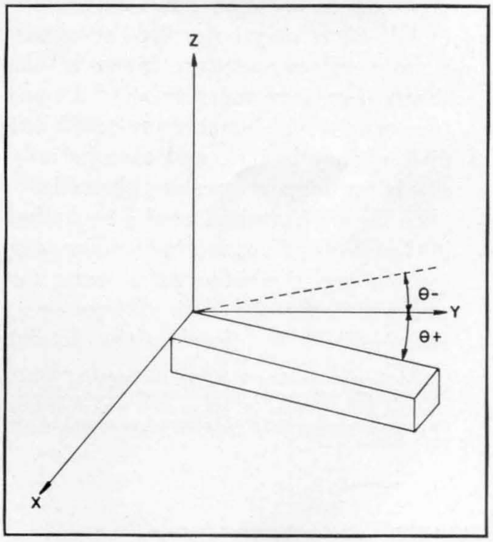
Using properly adjusted baffles, the above radiation resistance can be reduced by 90 or 95 per cent, becoming about 700 ohms. The actual measured resistance under these conditions would be about 2500 ohms, giving a Q of 98,000.

These figures indicate that vacuum mounting of such a bar will not result in an indefinite increase in Q. Here, vacuum mounting could at most remove the residual air-wave loss at the ends (700 ohms) plus a small amount due to radiation from other surfaces — perhaps 800 ohms in all, in the total of 2500 ohms. This would give a Q of about 98,000 × (2500/1700) = 147,000.

The residual loss is largely due to coupled modes of motion. It can be substantially altered by orienting the bar. While this reduction would be highly desirable from the point of view of obtaining the greatest possible Q, other factors are important, such as increased temperature coefficient.

An unexpected effect occurs, as a result of the YZ shear vibration, which is the excitation of a flexural vibration. This flexural vibration cannot be ex-

Figure 4. Showing the relation of the rotation angle, Θ, to the principal axis of the quartz.



cited piezoelectrically with the simple electrode system so far considered. A mechanical coupling exists between the flexural and shear modes by virtue of similar displacements at the boundaries. This may be visualized as shown in Figure 2. It is evident that there is a marked similarity of displacements at the ends of the bar, for the fundamental shear and the second flexural mode.

The flexural frequency is given by km^2Z/Y^2, where the factor m is a complicated function of the dimensions of the bar for all but extremely thin bars. An important property, however, is evident — that the flexural frequency *increases* with an *increase* in Z, or width, of the bar. This mode is the only one in which the frequency changes in this direction with change of dimension. When the width Z is made approximately 25 per cent of the length Y, the second flexural frequency is equal to the longitudinal frequency. Because of the comparatively strong coupling, the flexural frequency causes a marked disturbance of the longitudinal frequency and also causes a large increase in the temperature coefficient.

Further investigation of this region for bars having approximately square cross-section has shown that the coupled circuit effect may be utilized to obtain a frequency-temperature curve in the shape of an inverted parabola.[3] At one temperature, the temperature-coefficient of frequency is zero, and changes only slowly for temperatures on either side.

So far we have indicated a reduction in the effects of coupled frequencies only by making the frequencies very far removed from the desired frequency, accomplished by keeping the lateral dimensions of the bar very small compared with the length.

In Equation (1) it is possible to make certain of the elastic coefficients zero, by suitable orientation of the bar. Equation (1) is written with the particular values of the elastic and piezoelectric coefficients which apply when the three dimensions of the bar are taken along the three crystallographic axes.

Investigation of the properties of rotated bars shows that for a rotation of $-18.5°$, the coefficient s'_{24} vanishes. This means that for this rotation no shearing motion is excited and, consequently, no flexural motion will take place. Thus two of the disturbing frequencies are eliminated. The motion consists of the longitudinal vibration desired, accompanied by lateral motions only. A bar cut in this manner is found to have a higher Q than an unrotated bar. It has also a substantially higher temperature coefficient (approximately -24 parts per million per degree C. instead of -1 or -2), which is not too troublesome in some applications. At this angle the length of the bar is in the direction of minimum Young's Modulus in the YZ plane. For a given frequency this gives the shortest bar, some 6 per cent shorter than the unrotated bar.

At an angle of rotation of $+41.5$ degrees, not only is the coefficient s'_{24} zero, but also s'_{23}. The coefficient s'_{12} is much smaller than in the two previous cases. In this case a nearly pure longitudinal vibration is obtained, with a small lateral motion in the X-direction only.[4] A substantially higher Q is obtained but at the expense of a high temperature coefficient, approximately -77 parts per million per degree C. At this angle the length of the bar is very

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[3] "Low Temperature-Coefficient Quartz Crystals," W. P. Mason, *Bell System Technical Journal*, XIX, No. 1, January, 1940. On page 75, Mason gives the proportions for a square bar as 0.272. The bars here described are not square.

[4] "Some Experimental Studies of the Vibrations of Quartz Plates," R. B. Wright and D. M. Stuart, Research Paper No. 356, U. S. Bureau of Standards.

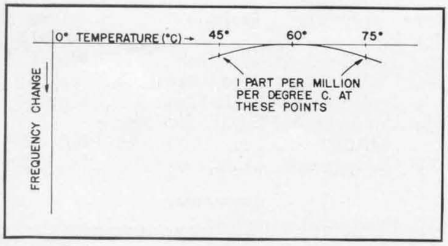

Figure 5. Plot of the change in frequency with the temperature for a Type 1190-A Quartz Bar.

nearly in the direction of maximum Young's Modulus in the *YZ* plane, giving very nearly the greatest length for a bar of given frequency, some 30 per cent longer than the unrotated bar.

The simple theory outlined above is not adequate to handle the case of bars the lateral dimensions of which are appreciable compared with the length. In general, experiment shows that the Q of bars whose lateral dimensions are of the order of 25 per cent of the length is somewhat greater and that the temperature coefficient can be made zero at a particular temperature. The bar described above for 50-kc fundamental operation can be operated in the second harmonic mode at 100 kc, thereby bringing the ratio of dimensions into the range giving low temperature coefficient, without using more quartz. Because of the mode of vibration, two nodal regions are obtained instead of one, making for a much better mechanical mounting with support at two points. A value of Q of 170,000 can be obtained, in a thread-suspension mounting with baffles, of the type shown in the photograph of Figure 1.

— J. K. Clapp

THE NEW ELECTRICAL UNITS

The change from the old "international" units to the new "absolute" units, which took place on January 1, 1948, was described under the above title in the *General Radio Experimenter* for July-August, 1947. This change involves a revaluation of the units of resistance, inductance, and capacitance as given in the following table:

1 international ohm = 1.000495 abs. ohms
1 international henry = 1.000495 abs. henries
1 international farad = .999505 abs. farad

To convert the values of existing standards to the new units, the present values should be multiplied by these factors. To adjust an existing standard to have its marked value in the new units, a resistor or inductor is decreased in value by 0.0495 per cent, while a capacitor is increased in value by 0.0495 per cent.

All instruments now being manufactured, together with all of their components, are adjusted and calibrated in terms of the new units. The difference between the old and new units is of importance only when the accuracy limits are 0.1 per cent and 0.25 per cent. In this accuracy classification there are 12 types of General Radio instruments and components, as shown in the accompanying table. All of these sold after the approximate dates given in the table will be calibrated in absolute units and will be marked with the abbreviation "abs." placed after or near the unit in which the instrument is calibrated. The spread in these dates from February 15 to May 1 has been caused by the fact that orders for parts for a production lot of instruments must be placed many months ahead of the final delivery to sales stock. It was necessary in such items as resistance units, which are manufactured in lots of several thousand and are used in almost all of this group of instruments, to arrange that all old "international" units be used up at about the same time and that new "absolute" units be then available.

New instruments with type numbers of 1000 and above and new components which may have type numbers below 1000 will not be marked "abs.", even though so calibrated.

Type	Resistors	Date
500	Resistor	April 15
510	Decade Resistance Unit	May 1
602	Decade Resistance Box	May 1
654	Decade Voltage Divider	May 1
668	Compensated Decade Resistance Unit	Feb. 15
670	Compensated Decade Resistor	Feb. 15
	Capacitors	
722	Precision Condenser	April 15
509	Standard Condenser	March 15
	Inductors	
106	Standard Inductance	March 15
	Bridges	
716	Capacitance Bridge	March 15
821	Twin-T Impedance Measuring Circuit	May 1
667	Inductance Bridge	March 1

READJUSTMENT OF OLD INSTRUMENTS

The difference of approximately 0.05 per cent between the old and the new units will be of little importance for much ordinary measurements work. On the other hand, in measurements where an accuracy approaching 0.1 per cent is desired, allowance for the difference between the two systems of units should be made. It is possible to readjust or recalibrate bridges and impedance standards in the absolute system of units, and our Service Department will quote prices upon request.

— Robert F. Field

MISCELLANY

Dr. M. A. El Said, Senior Lecturer at Fouad I University, Cairo, Egypt, at the request of the General Radio Company, visited the plant during January and February to demonstrate new electronic circuits particularly adapted to obtaining products and quotients of electrical voltages. One of these circuits is now under development for use in an electronic wattmeter.

The quartz bar illustrated in the article by J. K. Clapp is used in the new General Radio Type 1100-A Frequency Standard. Two models of this standard, the Type 1100-AP Primary Standard and the Type 1100-AQ Secondary Standard, are now in production and will be described in an early issue of the *Experimenter*.

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A VOLTAGE MULTIPLIER FOR THE VACUUM-TUBE VOLTMETER

Also

IN THIS ISSUE

● **A 10:1 VOLTAGE MULTIPLIER** is now available to extend the range of voltage measurement of the Type 1800-A Vacuum-Tube Voltmeter to a maximum of 1500 volts.

This multiplier is a capacitive voltage divider which provides a 10:1 reduction between the voltage applied to the multiplier and the voltage appearing across the voltmeter terminals. The multiplier screws on to the end of the voltmeter probe, adding about two inches to its length.

Since the input capacitances of the voltmeters differ slightly, an error in multiplier ratio of ±2% is possible, but an adjustment is provided by means of which the ratio can be adjusted to ±1% for any Type 1800-A Vacuum-Tube Voltmeter. When a multiplier and a voltmeter are ordered together, this adjustment is made at the factory.

The *effective parallel* input *resistance* of the multiplier is of the order of 100 times that of the voltmeter probe alone, and the effective parallel capacitance is 1.5 μμf. When the cap and center plug are used, approximately 0.5 μμf is added.

The resonant frequency of the probe, 1050 Mc, is not changed by the ad-

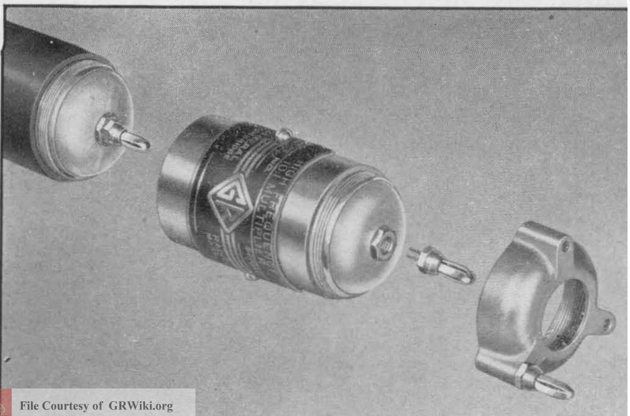


Figure 1. Exploded view of probe, multiplier, and cap.

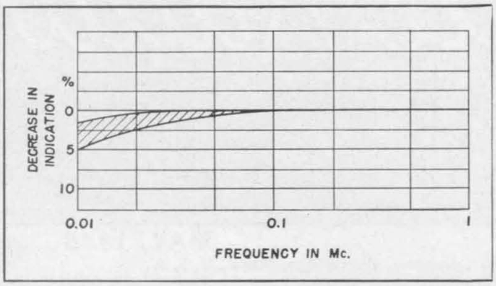


Figure 2. Plot of low-frequency error for the Type 1800-P2 Multiplier. The high-frequency correction for the voltmeter is unchanged by the addition of the multiplier.

dition of the multiplier. The multiplier frequency error is plotted in Figure 2. The multiplier is not recommended for use at frequencies below 100 kc.

SPECIFICATIONS

Multiplier Rates: 10 to 1.

Dimensions: (Length) 2⅝ x (diameter) 1⅝ inches, over-all.

Net Weight: 4 ounces.

Type		*Code Word*	*Price*
1800-P2	**Multiplier**	ABODE	**$18.00**

NEW STANDARD PARTS

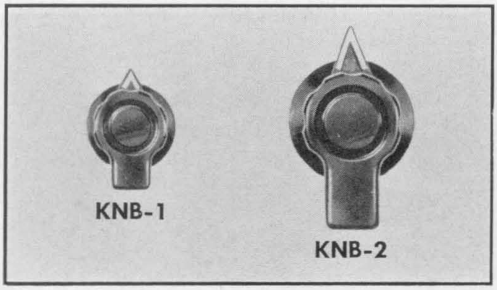
The new look that has lately been evident on General Radio instruments reflects the improvement in appearance of our standard parts. Designed to be attractive as well as useful, these knobs and dials are also available separately to those who make their own laboratory equipment.

KNOBS

Type KN Knobs, which replace the Type 637 series, are uniform in general appearance and application and were designed primarily for use on General Radio instruments. All are similarly fluted and have matching narrow skirts, so that a unity of design is achieved when different types are used on the same panel. Pointer models have large white V-shaped indicators for good visibility.

Two new types are now available, the bar knobs, KNB-1 and KNB-2, which are especially convenient for use on rotary switches, and the spinner knob, KNU-3, for rapid rotation of the control shaft on slow motion drives.

Each knob is made of black phenolic resin with a molded-in brass insert, and is fitted with two setscrews, 90° apart, which are threaded through the metal insert. The boring of the shaft hole is performed as a final operation on a precision machine, especially set up for the purpose, so as to insure an accurately sized hole which is concentric with and perpendicular to the molded portion. Holes are bored to fit a ⅜-inch diameter shaft and are equipped with removable bushings to adapt to ¼-inch diameter shafts.



1-INCH DIAMETER — WITH BAR

Type	*Net Weight for 5*	*Code Word*	*Package of 5*	*Package of 20*
KNB-1	**3¾ oz.**	BARKNOBONE	**$3.50**	**$13.00**

1⅜-INCH DIAMETER — WITH BAR

Type	*Net Weight for 5*	*Code Word*	*Package of 5*	*Package of 20*
KNB-2	**6 oz.**	BARKNOBTWO	**$3.75**	**$14.00**

2-INCH DIAMETER

Type	*Net Weight for 5*	*Code Word*	*Package of 5*	*Package of 20*
KNSP-8	**8 oz.**	NURLNOBATE	**$4.25**	**$16.00**

2-INCH DIAMETER — WITH SPINNER

Type	*Net Weight*	*Code Word*	*Unit Price*
ZKNU-3	**2¾ oz.**	SPINNOBTRE	**$3.00**

1⅜-INCH DIAMETER

Type	*Net Weight for 5*	*Code Word*	*Package of 5*	*Package of 20*
KNSP-6	**5½ oz.**	NURLNOBSIX	**$3.00**	**$11.00**

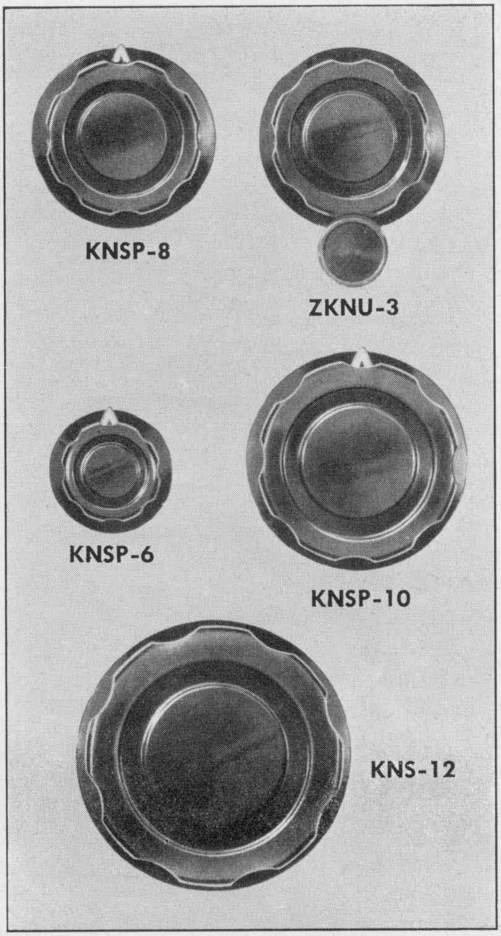
2⅜-INCH DIAMETER

Type	*Net Weight for 5*	*Code Word*	*Package of 5*	*Package of 20*
KNSP-10	**12½ oz.**	NURLNOBTEN	**$5.50**	**$21.00**

2⅞-INCH DIAMETER — WITHOUT POINTER

Type	*Net Weight for 5*	*Code Word*	*Package of 5*	*Package of 20*
KNS-12	**17 oz.**	NURLNOBDOZ	**$5.75**	**$22.00**

The Type KN Fluted Knobs are shown approximately one-half actual size in the illustration.



KNOB AND DIAL ASSEMBLIES

A new line of photo-etched dials with frosted-chrome surfaces replaces the older nickel-silver models. The Types 901, 902, and 904 dials are available with or without friction drives and in two or more scale lengths. These dials were designed for applications requiring simple, direct shaft positioning of moderate precision. Dials are assembled on standard Type KN knobs and therefore mount on the same size of shaft. The punched brass dial is accurately located on the knobs by bosses, from which the shaft hole is concentrically bored, and is also insulated from the shaft insert by the phenolic material of the knob.

Each dial is photo-etched and finished with black lines on a frosted-chrome plated background. This background finish has a silvery white color, furnishing excellent contrast with the black lines, and has diffuse reflecting properties, making it possible to view or illuminate the dial from any angle without objectionable glare.

The parallax-free indicator, which is designed to hug the rim of the dial and to remain flush with its surface, is finished in matching frosted-chrome plate with a black index line.

Friction-drive models are intended for use when the accuracy of a photo-etched dial is satisfactory, but when settings must be made more precisely than can be done with a direct drive. The speed-reducing drive grips the rim of a separate disc behind the dial, so that the dial surface is not marred, and the driving shaft is mounted on the panel in an eccentric bushing which permits lateral adjustment to the smoothest operating position.

2-INCH DIAMETER — TYPE 901 DIALS

	Dial					
Type	*Arc*	*Divisions*	*Drive*	*Net Weight*	*Code Word*	*Price*
901-HD	**180°**	**100**	Direct	2 oz.	DILOG	**$2.75**
901-JD	**270°**	**100**	Direct	2 oz.	DILAP	**2.75**
901-LD	**360°**	**100**	Direct	2 oz.	DILID	**2.75**

2¾-INCH DIAMETER — TYPE 902 DIALS

902-HD	**180°**	**100**	Direct	2½ oz.	DIMAP	**$2.75**
902-JD	**270°**	**100**	Direct	2½ oz.	DIMID	**2.75**
902-HF	**180°**	**100**	Friction, 3.3:1	4 oz.	DIMOB	**3.75**
902-JF	**270°**	**100**	Friction, 3.3:1	4 oz.	DIMUG	**3.75**

4-INCH DIAMETER — TYPE 904 DIALS

904-HD	**180°**	**100**	Direct	5 oz.	DIPAR	**$3.25**
904-JD	**270°**	**200**	Direct	5 oz.	DIPOD	**3.25**
904-HF	**180°**	**100**	Friction, 5:1	8 oz.	DIPEN	**4.25**
904-JF	**270°**	**200**	Friction, 5:1	8 oz.	DIPUT	**4.25**

FRICTION-DRIVE PRECISION DIALS

Types 905 and 906 Precision Dials replace the Types 704 and 706. They have fine black lines exactly positioned, by an automatic precision engraving machine, on a frosted-chrome background. The spacing of the lines is chosen to give the maximum accuracy of setting consistent with readability.

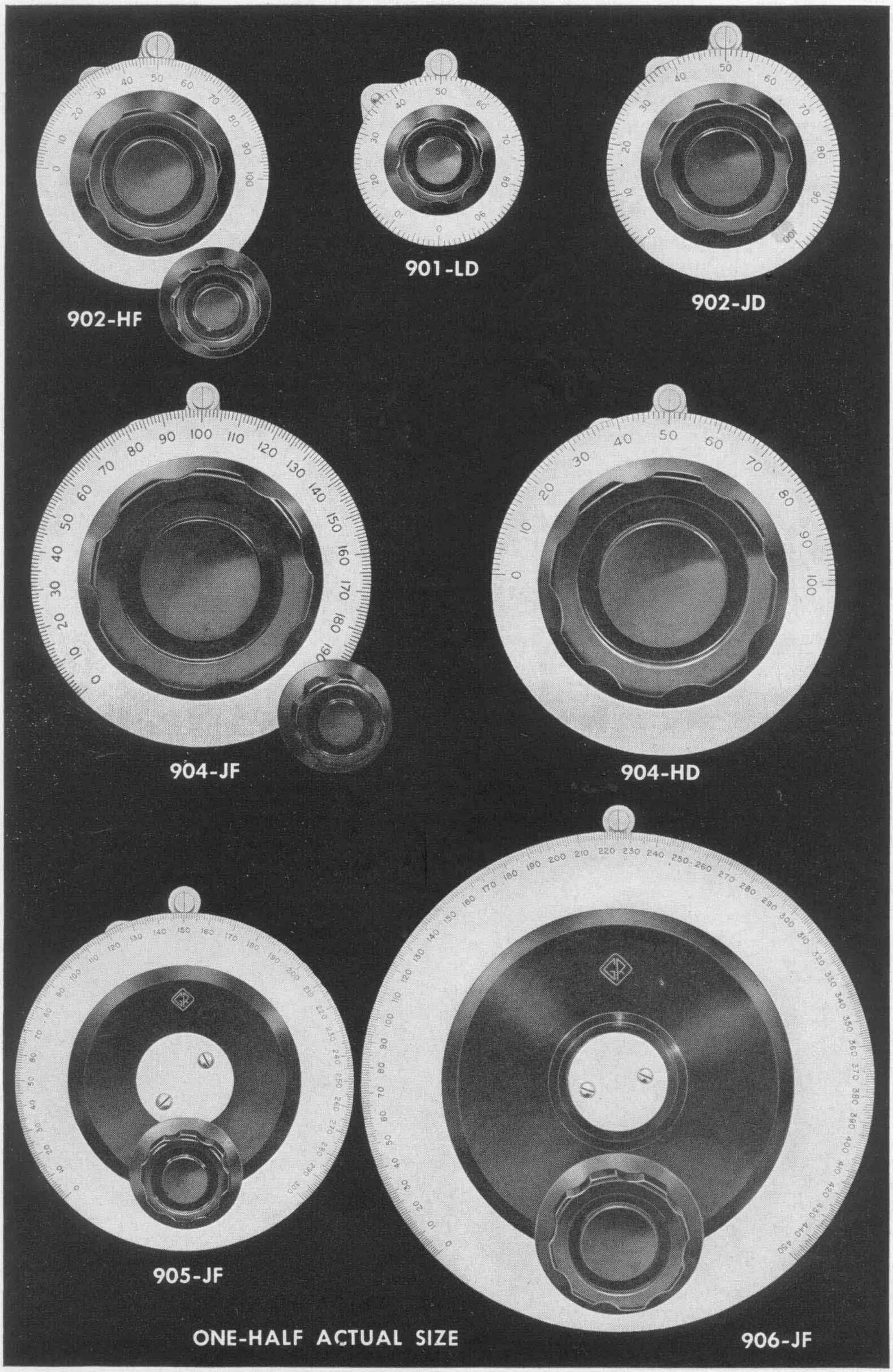
Each dial has a lathe-turned rim to insure concentricity, and is attached to a machined brass hub, which is fastened to the shaft by two setscrews, accessible from the front of the dial. The friction drive, which is fully adjustable, operates on the outer edge of a horseshoe-shaped slot in the dial, so that the slow-speed knob turns in the same direction as the dial and so that the drive mechanism all comes within the dial proper.

4-INCH DIAMETER — TYPE 905 PRECISION DIALS

905-HF	**180°**	**200**	Friction, 6:1	9 oz.	DIRUG	**$10.00**
905-JF	**270°**	**300**	Friction, 6:1	9 oz.	DIRIM	**10.00**

6-INCH DIAMETER — TYPE 906 PRECISION DIALS

906-HF	**180°**	**300**	Friction, 8:1	15 oz.	DIROT	**$12.00**
906-JF	**270°**	**450**	Friction, 8:1	15 oz.	DIRAP	**12.00**



ONE-HALF ACTUAL SIZE

GEAR-DRIVE PRECISION DIALS

Two entirely new units, the TYPES 907 and 908 Gear-Drive Precision Dials, were designed for the most exacting applications and combine the fine-line accuracy of a machine-engraved dial with a positive, fixed-ratio reduction drive. The dial is furnished in frosted-chrome plate with contrasting black lines, and its hub is attached directly to the shaft by two setscrews.

The gear has internal teeth, so that the knob, driving a stainless steel pinion, turns in the same direction as the dial. The pinion is held in a collet in the knob so that it may be readily adjusted to project through any panel, up to 5/16 inch thickness, by simply loosening the setscrews in the knob. This drive assembly runs in a floating bronze bushing that is spring guided to hold the gears in proper mesh without backlash and that also obviates the need for precise panel drilling.

The pinion and gear teeth are especially designed to run smoothly together and are cut on a precision gear shaper to insure uniformity. The 10:1 ratio permits calibrating the knob as a vernier.

The front-of-panel assembly requires only two 6-32 panel holes for mounting, and the large holes for the window and drive of the back-of-panel model are round, so that they can be readily put in with a punch, hole-saw, or flycutter.

The anti-parallax floating indicator is supplied with a dull-black back-up plate, which fills the window on the back-of-panel types.

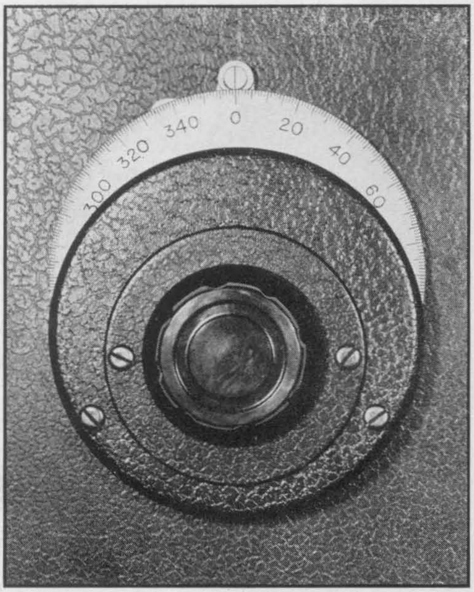
4-INCH DIAMETER GEAR-DRIVE PRECISION DIALS

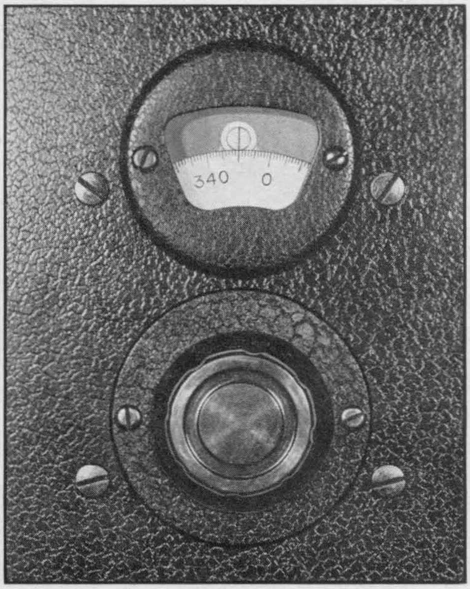
		Dial					
Type	*Mounting*	*Arc*	*Divisions*	*Gear-Drive Ratio*	*Net Weight*	*Code Word*	*Price*
907-LA	**Front-of-panel**	**360°**	**360**	**10:1**	11 oz.	DITAB	**$9.50**
907-LB	**Back-of-panel**	**360°**	**360**	**10:1**	11 oz.	DITOP	**9.50**

6-INCH DIAMETER GEAR-DRIVE PRECISION DIALS

908-LA	**Front-of-panel**	**360°**	**360**	**10:1**	21 oz.	DIVAT	**$11.00**
908-LB	**Back-of-panel**	**360°**	**360**	**10:1**	19 oz.	DIVIM	**11.00**

(*Left*) Type 907-LA; (*Right*) Type 907-LB, approximately half size.





VARIAC OPERATION WITH FIXED BRUSH SETTING

Variac adjustable transformers are devices for obtaining any required voltage within their range, quickly, conveniently, and efficiently. Variac voltage output is substantially independent of load current, as contrasted with resistive controls. Usually, Variacs are frequently adjusted to meet changing requirements of speed, illumination, line voltage, etc.

Occasionally, Variacs are required to act as fixed autotransformers, a service for which the latter are more properly suited. Under these conditions, the Variac brush setting remains fixed for long periods under load, and the turns under the brush are subjected to prolonged operation at high temperature.

At this point a brief discussion of the function of and need for "carbon" (actually a mixture of carbon, graphite, and, sometimes, metal) brushes in Variacs is in order. The brush is the secret of successful Variac design. To permit adjustment (i.e., switching) under load without interruption of current, the contacting device must establish contact with an adjacent turn of wire before breaking the contact already established. This means that the contacting device must, perforce, short circuit one or more turns in certain positions. In order to limit the short-circuit current to a safe and reasonable value, the contact must have a certain minimum resistance. Since this resistance is traversed by the load current as well as by short circuit currents, it must not be too high. In effect, we have a tug-of-war between short-circuit losses, which call for a high resistance, and load losses, which call for a low resistance. The compromise is best effected when the two losses are equal. Carbonaceous materials meet the resistance requirements nicely and have lubricating and wear characteristics that are excellent against the copper surface of the exposed brush track.

Because of the load losses which are unavoidably introduced by the brush requirements, the hottest spot on a Variac will always be directly under the brush. This localized heat source is in addition to the uniformly distributed heating of core and copper losses. These latter are maintained at conservative levels in Variac design, but the brush heat, even with generous heat radiator provisions, is still sufficient to raise the temperature of the turns in the immediate vicinity of the brush to a point where the bare copper oxidizes rapidly.

Copper oxide, unfortunately, is a poor electrical conductor, and its formation in the vicinity of the brush further increases the brush heating under load. Thus heat leads to oxide which leads to more heat in an extremely vicious circle. If failure is to be avoided, the circle must be broken.

Two good methods have so far been found to prevent oxide accumulation. One is obvious. Keep the brush track clean, using the fine crocus cloth and carbon tetrachloride cleaning technique outlined in the Variac instruction sheet, at frequent (semi-weekly) intervals. The other is to prevent or slow down oxidation by excluding air from the vicinity of the brush by the application of a thin layer of heat resistant, inert grease to the brush track. Dow-Corning D-C-44 Silicone Grease has been found to be excellent for this purpose. A combination of the two methods, a thorough cleaning and regreasing every two weeks, will permit indefinite operation of Variacs with fixed brush setting.

— GILBERT SMILEY

HONORS — Awarded to Melville Eastham, Chief Engineer and former President of the General Radio Company, the 1948 New England Award, at the annual meeting of the Engineering Societies of New England, at Boston, April 29. A scroll of illuminated parchment suitably inscribed, the New England Award is given each year to "a living engineer, resident in New England, who, by outstanding achievement, shall merit recognition of his accomplished work as well as of his character, by his fellow engineers of New England."

Melville Eastham.

RECENT VISITORS —

From Poland: Przemyslaw, J. Jaros, General Superintendent of Communications, Polish State Railways, Warsaw.

From Brazil: Captain Aldo V. da Rosa, Chief of Technical Division, Brazilian Air Force.

From Chile: William Feick, Professor of Communication Engineering, Universidad Tecnica Federico Santa Maria, Valparaiso.

From Holland: Mr. D. Goedhart and Mr. W. W. Storm of N. V. Philips Telecommunication Industries, Hilversum.

PAPERS presented at the I. R. E. National Convention, New York, March 21-24, 1948 — by H. B. Richmond, Chairman of the Board, "An Engineer in the Electronics Industry — Prospects, Preparation, Pay"; by W. N. Tuttle, Engineer, "Use of Diode Rectifiers with Adjustable Transformers for Motor Speed Control"; by R. F. Field, Engineer, "Losses in Air-Cored Inductors"; by J. K. Clapp, Engineer, "Frequency Measurement by Sliding Harmonics."

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

MAKING A GOOD INSTRUMENT BETTER

Also IN THIS ISSUE

● **THE TYPE** 913-A Beat-Frequency Oscillator[1], announced in 1942, established standards of performance not previously achieved by oscillators of this type. Its immediate widespread acceptance showed that frequency stability, low distortion, and constancy of output were truly important features in an audio oscillator. Experience with field use over the past six years, however, has shown several ways of improving the original design to provide better performance for the customer and greater convenience in manufacture. Changes have therefore been made from time to time that have been incorporated in the successive Type 913-B and Type 913-C models, culminating in the new Type 1304-A. These changes, in total, make the new instrument considerably different from the original and warrant mention as typical of the refinements that successively appear in instrument redesign.

---

[1] H. H. Scott, "Bringing the Beat-Frequency Oscillator Up to Date," *General Radio Experimenter*, Vol. XVII, No. 2, July, 1942.

Figure 1. Panel view of the Type 1304-A Beat-Frequency Oscillator. New features include an illuminated, precision-type, main dial, and an incremental frequency dial with a range of −50 to +50 cycles.

OSCILLATOR

The original oscillators used Type 6SK7 pentodes in tuned-plate tickler-feed-back circuits, with output pickup coils feeding the grids of the Type 6SA7 mixer directly. When properly constructed, these circuits gave excellent stability, but considerable difficulty was found in maintaining the high quality of fabrication necessary for satisfactory coils.

The coils were wound on slotted catalin coil forms and were adjusted to the correct inductance by iron-dust cores on a mounting constrained at each end. It was found that dimensional changes of the form material with time and temperature caused relative motion of the dust cores and coils, resulting sometimes in excessive drift, and sufficient trouble on this score was ultimately encountered to make it desirable to look into other constructions.

In the course of this investigation it was found that a Hartley or Colpitts oscillator using a simple single-winding coil was more stable than the tickler-feed-back oscillator using the three-winding coil, because the circulating current in the tuned circuit flowed through the whole coil, causing uniform heat distribution, and because the close coupling between grid and plate circuits minimized phase shift. The final coil design, therefore, turned out to be a single multilayer universal-wound coil on a simple, unslotted ceramic form. As a further simplification, it was found that the inductance could be held to a close enough tolerance that no iron-dust-core trimming was necessary. Simplification of construction, improvement of performance, and reduction of manufacturing cost were therefore simultaneously achieved.

BUFFER AMPLIFIERS

In the original design, no buffer amplifiers were interposed between the beating oscillators and the mixer. A pentagrid mixer tube was used, and it was felt that the isolation obtained by using two individual grids, fed from the low-impedance pickup windings on the oscillator coils, was adequate to minimize low frequency distortion caused by incipient locking of the oscillators.

With the new, simplified oscillator coils, no low-impedance pickup coil was available. To permit high-impedance capacitance voltage dividers to be used to feed the mixer grids without increasing the cross-coupling between oscillators to an intolerable amount, buffer amplifiers were necessary. These were obtained without increase in the tube complement by changing the oscillator tubes from pentodes to twin triodes. One half of each twin triode was used as the oscillator section, the other half as a cathode-follower amplifier.

The plates of both sections of the twin triodes are operated at r-f ground, with the cathode of the oscillator section tapped on the oscillator coil and the grid connected to the end of the coil at the highest r-f potential. This particular method of operation leads to two desirable results: first, the stator of the main tuning capacitor operates at d-c ground and, second, the plates of the twin-triode sections act as shields, eliminating the need for additional shielding of the tubes.

The grids of the buffer-amplifier sections are supplied through capacitance voltage dividers, which minimize loading of the oscillator circuits and permit adjustment of the voltages supplied to the mixer grids by varying the setting of small adjustable ceramic trimmer capacitors. Once the proper grid voltages for the mixer have been established, it is therefore possible to set these voltages to the same levels in all instruments. This simple adjustment, combined with the excellent isolation afforded by the buffer amplifiers, has effected very considerable improvement in the low-frequency distortion below 100 cycles and has reduced locking of the oscillators to a point where beats of less than 1 cycle can be sustained without pull-in.

MIXER

The chief problem in adjusting the mixer for proper operation in the Type 913-A and 913-B Oscillators was in obtaining independent settings for output voltage and minimum distortion. It was, however, found that, if the former arrangement of bias controls was changed so that the common cathode-bias for both grids was adjusted by a rheostat, and the bias for the signal grid (grid No. 3) by a potentiometer in parallel, the two adjustments for output voltage and minimum distortion could be made with no observable interaction. The improvement in performance from this change, from the adjustable oscillator voltages, and from the improved isolation of the buffer amplifiers has been very great. In previous models careful selection of the mixer tube had been necessary to maintain the low distortion desired. In the new design, on the other hand, nearly any tube will perform satisfactorily, and large numbers of tubes previously rejected for high distortion have been found completely satisfactory.

AMPLIFIER

Changes in the amplifier circuit and shelf were made principally because of difficulty with hum introduced from the heaters of the Type 6SF5 phase-inverter tubes. Replacement of these two tubes with a single Type 6SL7-GT Twin-triode has resulted in no further difficulty and has made it possible to rearrange parts to provide more convenience in assembly.

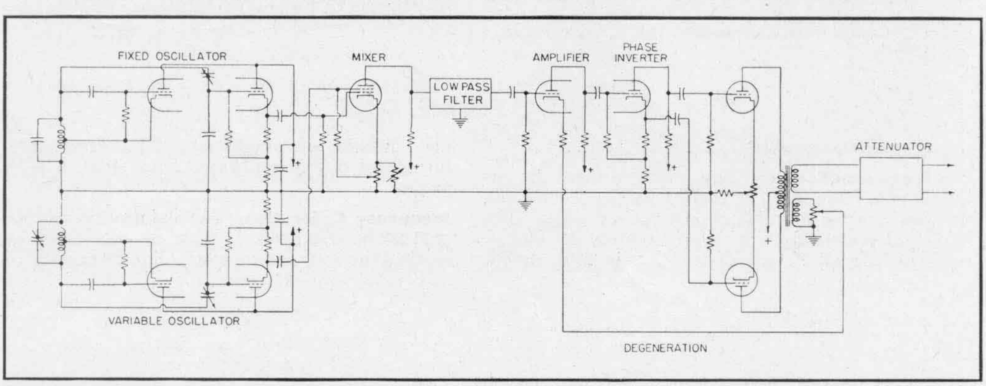
POWER SUPPLY

Rearrangement of the power-supply shelf and the substitution of a single 5V4-G for the two 6X5-GT tubes previously used have eliminated a difficult hum-reducing adjustment and improved heat-radiation conditions, the temperature rise in the oscillator compartment, for instance, being only 6 C.

MECHANICAL SIMPLIFICATION

Since these various changes necessitated a rearrangement of parts, considerable attention was given to utiliz-

Figure 2. Elementary schematic circuit diagram of the Type 1304-A Beat-Frequency Oscillator.



ing the available space most efficiently and to simplifying assembly. As a result, parts and circuit adjustments are more accessible, and wiring is simpler than in previous models.

NEW FEATURES

These variations have improved performance substantially, reduced the total number of tubes by two, simplified assembly and testing, and lowered manufacturing cost. As a result it has been possible to add new features at no increase in price.

The first of these is a cycles increment dial. This was first introduced in the Type 713-A Beat-Frequency Oscillator[2] and, since the discontinuance of this oscillator, many customers have requested that it be added to the Type 913 if it were ever redesigned. In the new Type 1304-A a better capacitor has been used than in the old Type 713-A, but the function and scale range remain the same. It is interesting to note that a considerable simplification in construction has been achieved by connecting the zero adjust capacitor across the variable-oscillator capacitor since no shield can for this capacitor is necessary and valuable panel space is conserved. The variation of capacitance required for zero adjustment is so small that no effect upon calibration accuracy results.

The second new feature is injection of line voltage into the beat-indicator circuit. As experience has been gained over the years with the adjustment of the variable capacitor used to give true logarithmic response over three decades of frequency, it has been found practical to furnish calibrations accurate to closer and closer tolerances. It has now become possible to adjust to ±(½ cycle + 1%), which is closer than zero beat can be established with the neon light. To translate this accuracy of adjustment into accuracy of readings, a more precise "zero" adjustment can be made at the line frequency by observing the waxing and waning of the neon light as the output of the oscillator beats with the line voltage. The improvement, of course, results from the fact that the beat is determined at a frequency where the amplifier has normal gain, as contrasted to the zero-beat condition where the amplifier has insufficient gain to operate the neon light until the deviation from zero beat is of the order of two cycles.

The third new feature is the improved gear-drive dial, patterned after the new Type 907-LA. This dial has better bearings than the old dial, smoother running gears and less backlash, and is simpler in construction and easier to make.

The Type 1304-A Beat-Frequency Oscillator furnishes an outstanding example of the painstaking development work that goes into instrument redesign over a period of years. As a commercial product in its own right, we believe this oscillator to be the finest now obtainable.

—D. B. Sinclair

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[2] L. B. Arguimbau, "Type 713-A Beat-Frequency Oscillator," *General Radio Experimenter*, Vol. X, No. 10, March, 1936.

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. The frequency-increment dial is calibrated from +50 to −50 cycles.

Frequency Calibration: The calibration can be standardized within 1 cycle at any time by setting the instrument to the line frequency or to zero beat. The calibration of the frequency control dial can be relied upon within ±(1%+ 0.5 cycle) after the oscillator has been correctly set to zero beat. The accuracy of calibration of the frequency-increment dial is ±1 cycle.

Zero Beat Indicator: A neon lamp is used to indicate zero beat at the line frequency or at zero scale.

Frequency Stability: The drift from a cold start is less than 7 cycles in the first hour and is essentially completed within two hours.

Output Impedance: The output impedance is 600 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 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. The open-circuit output voltage is approximately 40 volts with the output switch in the HIGH position.

Output Control: The output control is calibrated from +25 to −20 db, referred to 1 milliwatt into 600 ohms.

Output Power and Waveform: NORMAL output 0.3 watt maximum when operated into a matched load, with total harmonic content approximately 0.25% between 100 and 7500 cycles. Below 100 cycles the harmonic content increases, and may reach 0.5% at 50 cycles. A panel switch allows an increase in the output power to a maximum of 1 watt. For this HIGH position of the OUTPUT switch the distortion is less than 1% between 100 and 7500 cycles and increases to 2% at 50 cycles. With the OUTPUT control turned fully on, the harmonic content is approximately doubled when the oscillator is operated into a very low impedance. With the OUTPUT control turned 3 db or more below maximum load, impedance has very little effect upon the waveform.

A-C Hum: For NORMAL output the a-c hum is less than 0.1% of the output voltage.

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

Mounting: 19-inch relay rack panel; removable wooden ends are supplied so that it may be used equally well on a table.

Power Supply: 105 to 125 volts, 50 to 60 cycles a-c. 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.

Tubes:

3—6SL7-GT
1—6SA7
1—5V4-G
2—6V6-GT
2—OD3/VR150
1—991

All are supplied with the instrument.

Accessories Supplied: A seven-foot line connector cord and a multipoint connector.

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

Net Weight: 41½ pounds.

Type		*Code Word*	*Price*
1304-A	**Beat-Frequency Oscillator**	CAROL	**$450.00**

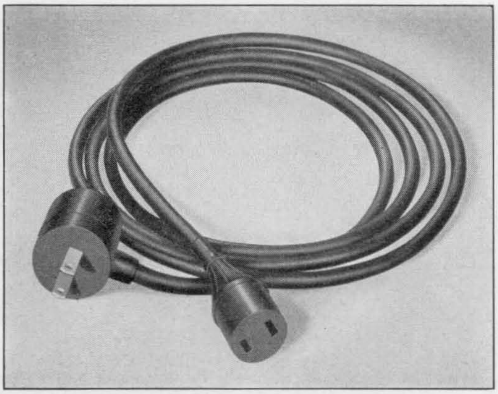
PATENT NOTICE: U. S. Patent No. 2,298,177. Licensed under patents of the American Telephone and Telegraph Company.

GR POWER CORD NOW AVAILABLE

There has been a considerable demand lately for replacement power cords of the type supplied with General Radio instruments and, because of this, we are making the power cord available as a standard catalog item.

The cord is 7 feet long and has the plug and socket ends molded in rubber directly to the two-conductor, No. 18, Type S5, stranded cord. Net weight is 8 ounces.

Type		*Code Word*	*Price*
CAP-35	**Power Cord**	CORDY	**$1.75**



A NEW MODEL OF THE MICROVOLTER*

The audio-frequency Microvolter is a very useful accessory for the electronics laboratory. Used in conjunction with an oscillator, the Microvolter converts it to a standard signal generator, capable of such measurements as gain or loss, frequency characteristic, overload level, and hum level on amplifiers, networks, and other low-frequency equipment.

This combination is also useful for the measurement of the generated voltage of microphones, vibration and phonograph pickups, and other transducers by the insert-voltage method. The Microvolter supplies the standardizing voltmeter and the calibrated adjustable attenuator which is necessary for providing an accurately known voltage over the range from 1 microvolt to 1 volt.

The Type 546-B Audio-Frequency Microvolter, which has been widely used for such measurements, has recently been replaced by a new model, the Type 546-C. Chief improvements in the new instrument are (1) a voltmeter which is more sensitive and has a better frequency characteristic, and (2) the adoption of the standard 600-ohm level for the input and output impedances.

The voltmeter uses a miniature copper-oxide rectifier designed, as in the earlier Microvolter, to have a negligible frequency error over the audio-frequency range. In addition, a compensation circuit corrects for frequency error (at the reference level labeled "0 db") up to 100 kilocycles per second to within ±5%. This compensation accordingly extends the usefulness of the meter to cover the full frequency range of the attenuator. The increased sensitivity reduces the non-linear loading of the rectifier on the input with a reduction in the waveform distortion introduced by the unit to less than 0.2% for a 600-ohm source. This low level of distortion brings the Microvolter in line with the low-level generally required from modern instruments, and, if the source impedance is reduced to much less than 600 ohms, its distortion is even less. The meter scale is calibrated in db with respect to the reference level so that an approximate correction for output level can be made when a reduced input level is desired.

The attenuator settings are made by

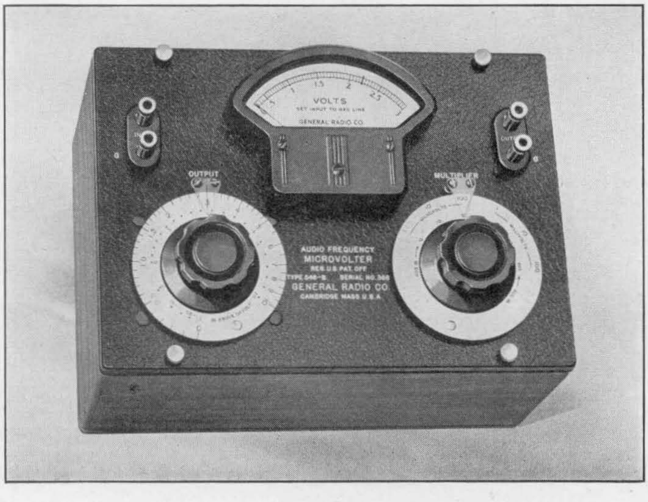


Figure 1. Panel view of the Microvolter. Photograph shows previous model, Type 546-B. New model differs very little in appearance.

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*Reg. U. S. Patent Office.

means of two panel controls, one a six-step decade multiplier and the other an individually calibrated dial which gives continuous variation over each decade. The dial carries both voltage and decibel calibrations. The voltage scale is approximately logarithmic and the decibel scale approximately linear.

The output voltage is expressed as the open circuit voltage in microvolts or millivolts, and the open circuit level in decibels is expressed in db above one microvolt. This reference level is used for the convenience of having almost all positive decibel readings over the range of the instrument. By subtracting 123.8 db, this level can be transferred to dbm (decibels with respect to one milliwatt) into a 600-ohm load.

The decibel calibration makes it possible to obtain gain or loss values directly in decibels for amplifiers, transformers, lines, and other networks without the necessity of converting voltage ratios.

—ARNOLD P. G. PETERSON

SPECIFICATIONS

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

Accuracy: For open-circuit output voltages the calibration is accurate within ± (3% + 0.5 microvolt) for output settings above 1 microvolt and for all frequencies between 20 and 20,000 cycles. For higher frequencies up to 100 kc the calibration is accurate within ±5% for output settings above 100 microvolts. These specifications apply only where waveform and temperature errors are negligible (see below).

In calculating ratios of output voltages, at a given frequency, the accuracy of any given reading can be considered to be within ± (2% + 0.5 microvolt), at frequencies up to 100,000 cycles. At the higher frequencies this accuracy applies only at levels above 100 microvolts.

The microvolter can be used on dc if an external meter is used or if the internal meter has been calibrated for dc.

Output Impedance: The output impedance is approximately 600 ohms and is constant with setting within ±5%. This impedance is sufficiently low so that no correction on the output voltage is necessary for load impedances of the order of 100,000 ohms and greater.

Input Impedance: Approximately 600 ohms, substantially independent of output setting on all but the highest multiplier position.

Waveform Error: The accuracy of the microvolter as a calibrated attenuator or voltage divider is independent of waveform. The absolute accuracy of the output voltage calibration depends on the characteristics of the input copper-oxide rectifier voltmeter, which has a small waveform error that depends in turn on both the phase and the magnitude of harmonics present in the input. This error in the voltmeter can, in general, be neglected when the microvolter is used with ordinary laboratory oscillators. The rectifier-type voltmeter itself introduces some distortion unless the source impedance is very low. With a 600-ohm source the distortion introduced is about 0.2%.

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

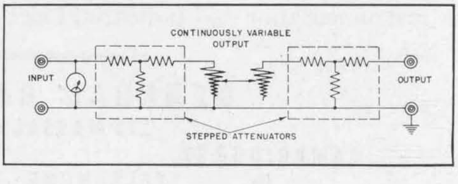
Power Source: The driving oscillator must be capable of furnishing about 2.2 volts across 600 ohms, or about 8 milliwatts.

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

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

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

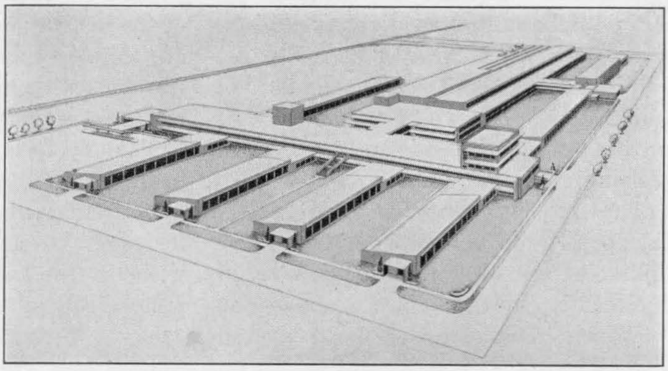
Net Weight: 6½ pounds.



Type		*Code Word*	*Price*
546-C	**Audio-Frequency Microvolter***	CROWN	**$110.00**

*Reg. U. S. Pat. Off.

MISCELLANY



NAVY DEDICATES NEW LABORATORY Aerial view of the new $8,000,000 Michelson Research Laboratory, dedicated on May 8, at the Naval Ordnance Test Station, Inyokern, California. Named in honor of the noted physicist, Dr. Albert A. Michelson, the new laboratory contains 410,000 square feet of floor space. It includes facilities for research and development in all branches of physical science, extensive modern shop facilities, and a mathematics section for theoretical work.

TECHNICAL PAPERS — "Determination of the Loudness of Noise from Simple Measurements," by Leo L. Beranek, Acoustics Laboratory, M.I.T., and Arnold P. G. Peterson, Engineer, General Radio Company; at the Washington, D. C., Meeting of the Acoustical Society of America, April, 1948.

—"A Standard-Signal Generator for F-M Broadcast Service," by Donald B. Sinclair, Assistant Chief Engineer; at the New England Radio Engineering Meeting, May 22, 1948.

— C. T. Burke spoke on the "Need for Improved Quality in Electron Tubes for Instrumentation and Industrial Use" at the Electron Tube Conference held by the A. I. E. E. Joint Subcommittee of Electronic Instruments, at Philadelphia, March 29.

RECENT VISITORS — Mr. Hayward C. Parish, Distributor of General Radio products in Australia; Mr. Ching Yi Sui and Mr. Hsiung Hsu, Central Broadcasting Administration, Shanghai; Mr. Mei-Lian Cheng, Central Radio Corporation, Nanking.

CREDITS — As is usual in instrument development, several engineers collaborated in producing the finished design of the TYPE 1304-A Beat-Frequency Oscillator. S. R. Larson was responsible for many of the new ideas and seeing the job through, while F. D. Lewis supplied the detailed criticism of the previous design as a guide to improvement. The project was carried out under the supervision of D. B. Sinclair, author of the article that appears in this issue.

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A WIDE-RANGE CAPACITANCE TEST BRIDGE

Also

IN THIS ISSUE

● **WIDE-RANGE** is a term that has no absolute meaning and is perhaps used a bit loosely on occasions in technical and promotional writing. Nevertheless, there is little risk of criticism of its use to describe the TYPE 1611-A Capacitance Test Bridge. This new bridge measures capacitance from 1 $\mu\mu$f to 10,000 μf, a range of ten billion to one. Over this entire range an accuracy of $\pm(1\% + 1\ \mu\mu\text{f})$ is maintained.

The new bridge combines the functions of the older TYPES 740-B* and 740-BG* but improves on the performance of each in several important respects. In accuracy, sensitivity, and convenience the performance equals or exceeds that of both previous models. In the very important range below 1000 $\mu\mu$f, the performance has been markedly improved by the use of a unique zero-compensating circuit. For the measurement of electrolytic capacitors, the new bridge permits the application of a polarizing voltage from a grounded power supply, an important convenience feature. The sensitivity of the detector is controlled by the bridge unbalance in such a manner that balance can be reached with a minimum manipulation of the manual gain control.

The panel controls and their location have been selected with ease of operation in mind. The less-used controls are near the top of the panel, the more used near the bottom where they are most accessible.

The scales of the capacitance and dissipation factor dials are direct reading in capacitance and dissipation factor, and the indexes for both scales are visible through a single window. Since at the index both scales are vertical, both increase upward. Hence increasing readings are obtained for counterclockwise rotation of the right-hand knob, and for clockwise rotation of the left-hand knob.

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*Although the TYPE 740-BG Capacitance Test Bridge has been discontinued, the TYPE 740-B is still available. Its limited field of usefulness, as compared to the new bridge, is offset by its lower price.

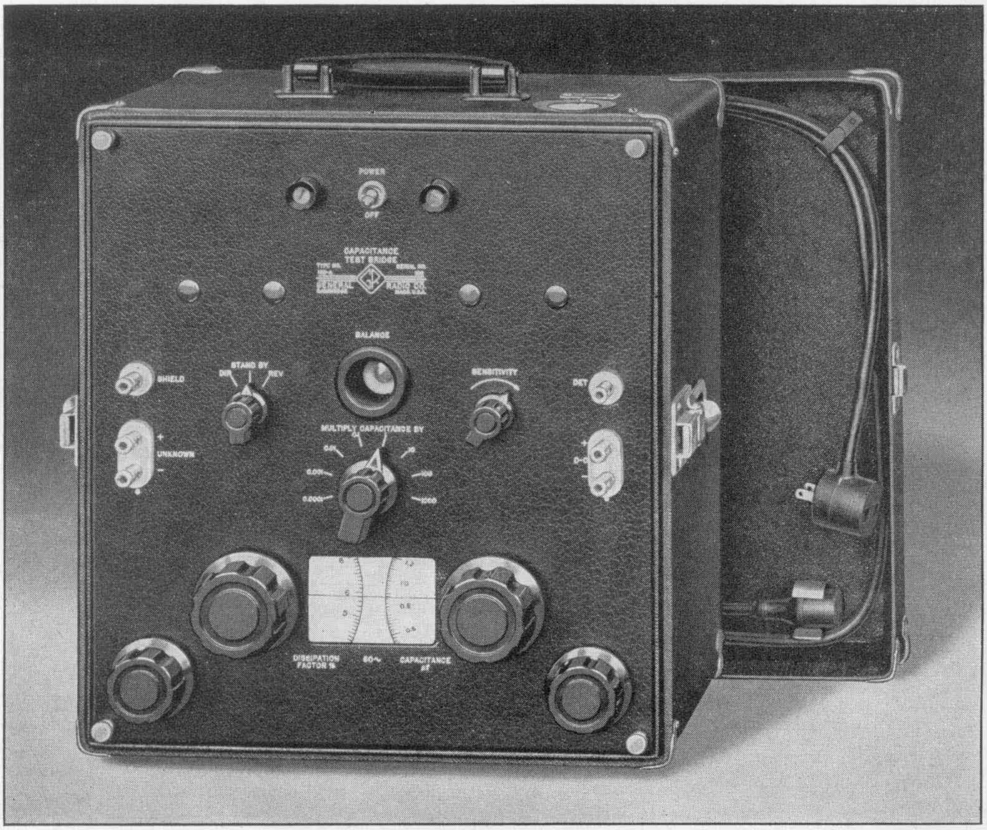
The Basic Bridge Circuits

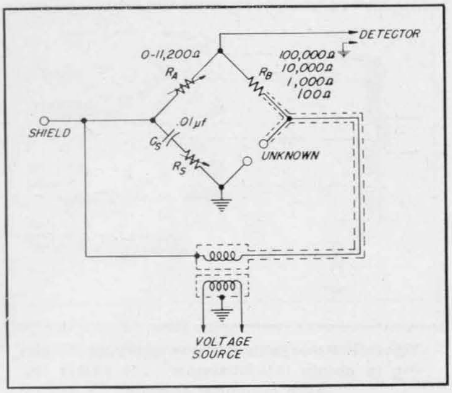
The series-resistance capacitance bridge circuit is used, in which the unknown capacitance is measured by variable resistance arms against a fixed standard capacitor, and the dissipation factor of the unknown is measured by a variable resistor in series with the standard capacitor. In order to cover effectively the extremely wide capacitance range, two bridge circuits are used, differing in the value of the standard capacitance used and in the method of connecting voltage source and detector. The necessary changes in circuit connection are all made by the CAPACITANCE MULTIPLIER control.

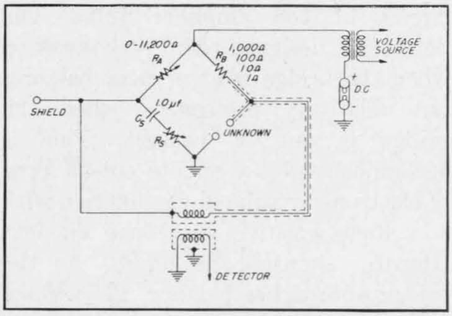
Figure 2a shows the basic circuit arrangement used for the four lower multiplier positions (0.0001, 0.001, 0.01, and 0.1). Note that the voltage source is connected across the similar arms of the bridge. With this method of connection, the sensitivity of the bridge depends upon the ratio of the two capacitances. The applied voltage and the detector sensitivity are such that satisfactory balance can be obtained for any ratio up to one hundred to one; that is, for unknown capacitances in the range from 100 $\mu\mu$f to 1 μf. Below 100 $\mu\mu$f, the absolute sensitivity falls off inversely with capacitance, which means that the precision of balance can be expressed as a constant capacitance, in this case of the order of 0.1 $\mu\mu$f. This precision of balance is adequate for the capacitance balance but does not permit precise dissipation factor determinations for very small values of capacitance.

Figure 2b shows the arrangement

Figure 1. Panel view of the Type 1611-A Capacitance Test Bridge with cover of cabinet removed.







(*Left*) Figure 2a. Basic bridge circuit used for the four lower multiplier positions.

(*Above*) Figure 2b. Circuit for the four higher multiplier positions.

used for the four higher multipliers (10, 100, 1000, and 10,000). Note that the voltage source and the detector have been interchanged as compared to the circuit of Figure 2a, and that the capacitance of the standard arm has been changed from 0.01 μf to 1.0 μf. In this method of connection the bridge sensitivity is a function of the ratio of the resistance R_A to the impedance in the standard arm. Unlike the circuit of Figure 2a, the sensitivity of the circuit in Figure 2b does not change as the ratio arm R_B is changed. Optimum sensitivity would be attained with the impedance of the standard arm equal to the resistance R_A. The latter is variable, however, being the balancing arm of the bridge, and, accordingly, the sensitivity varies with the setting of this arm. The reactance of a capacitance of 1 μf at 60 cycles is 2650 ohms, and maximum sensitivity is attained when the R_A equals this value. Actually the main decade of the A-arm rheostat goes from 1000 to 10,000 ohms, so that the sensitivity is optimum at about mid-scale and changes relatively little over the range.

Bridge Voltages

Because of the very wide range of input impedance presented by the bridge, no single source of voltage can be capable of efficiently supplying test voltage for all ranges. Ideally a separate source of proper impedance and voltage might be provided for each multiplier position, but an excellent compromise is obtained by providing four separate sources, one for each *pair* of the eight multiplier positions. For each pair of positions, a resistance is placed between the bridge and the voltage source such that the same power is delivered to the bridge for each position. The required value of resistance is, of course, the geometric mean of the two values of bridge impedance. The voltages for the four sources are so chosen that, with these series resistances, the maximum safe power is delivered to the bridge for any setting of the multiplier. Figure 3 shows schematically the arrangement. The proper source is selected by a switch mechanically connected to the switch that controls the multiplier ratio of the bridge.

The Detector

The detector system consists of a single stage of amplification and an electron-ray tube used as a visual null indicator. The amplifier is made selective to the operating frequency by a parallel resonant circuit in the plate

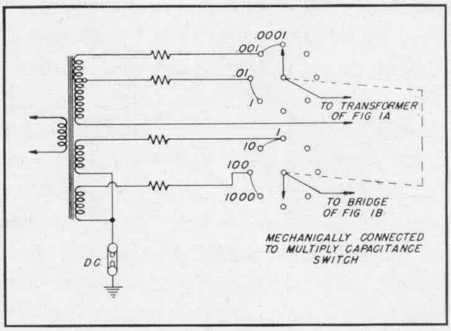


Figure 3. Arrangement of power supply switching to obtain the maximum safe power for each multiplier position.

circuit of the amplifier tube. The detector is designed to be very sensitive when the bridge is at or near balance, but relatively insensitive when the bridge is out of balance. This is accomplished by a remote cut-off type of electron-ray tube in conjunction with a voltage-sensitive resistance element (thyrite) shunting its input. As the bridge approaches balance, the voltage impressed on the thyrite element is reduced. The resistance-voltage characteristic of the latter is such that the resistance approaches a maximum value as the applied voltage approaches zero. Consequently, as the voltage is reduced, the gain of the system increases because the resistance element in question is located in the plate circuit of a high-impedance tube (see Figure 4).

The indicator tube (Type 6U5) is mounted in a slotted cylinder and held in position by a thumbscrew. When the bridge is used in very brightly lighted locations, or when the brilliance of the "eye" has been reduced with age, the tube may be slid back in its mounting to provide additional light shielding.

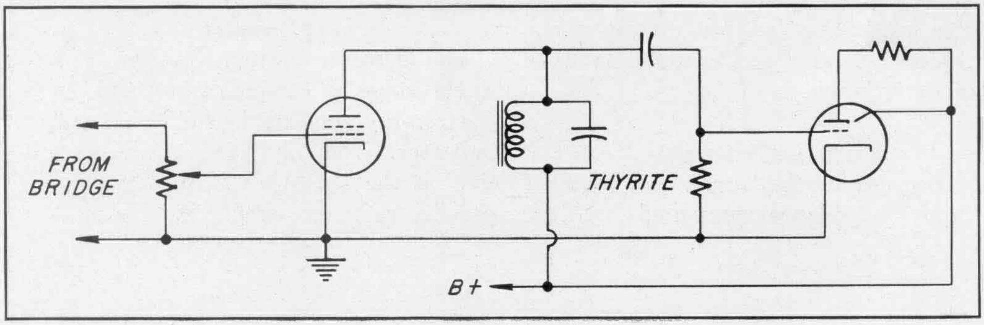
Compensating Circuits

Any loss in the standard capacitor causes the bridge to read low in dissipation factor by an amount equal to the dissipation factor of the standard capacitor. Although this value does not exceed 0.0003, it must be compensated for in order to realize the maximum accuracy of the bridge. Compensation is accomplished by connecting fixed capacitors across the resistors in the opposite arm, of such value that the product $R_B \omega C_B$ equals the dissipation factor of the standard capacitor.

A new method of compensation is used to eliminate the effects of the zero capacitance and losses across the unknown terminals. It is this compensation which makes possible accurate direct measurements of capacitance and dissipation factor obtained on the lowest multiplier position, without "zero" corrections.

As shown in Figure 5, voltages of adjustable magnitude are fed to the bridge output through a capacitor and a resistor. These voltages are adjusted to cancel the unbalance voltage produced by the zero capacitance. They

Figure 4. Elementary schematic diagram of the null detector.



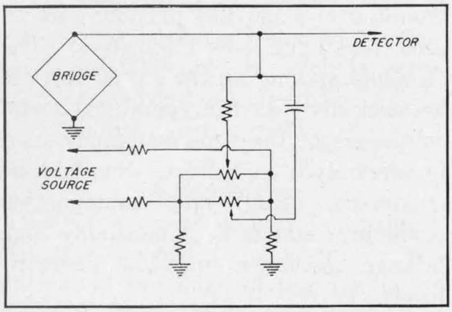


Figure 5. Schematic diagram of circuit used to cancel the unbalance voltage produced by the zero capacitance.

also serve to compensate partially for any leakage through the bridge transformer as well as for any small voltage induced in the amplifier through capacitance coupling to high-voltage leads.

The theory upon which this compensating method is based is as follows. Consider the network of Figure 6 wherein a voltage E' is coupled through an admittance Y_5 to the detector terminals of the four-arm bridge network energized by the voltage E. The condition of balance (zero potential across terminals $A - A'$) is most easily determined by considering the shortcircuit current across the detector terminals. The bridge itself produces a current most conveniently expressed in admittance form as

$$i_{sc} = E\frac{Y_1Y_4 - Y_2Y_3}{Y_1 + Y_2 + Y_3 + Y_4} \quad (1)$$

The circuit E', Y_5 yields a current equal to $E'Y_5$. Equating the sum of the currents to zero and designating as α the ratio E'/E, we have the following expression:

$$Y_1Y_4 - Y_2Y_3 + \alpha Y_5(Y_1 + Y_2 + Y_3 + Y_4) = 0 \quad (2)$$

This can conveniently be rewritten in the equivalent form

$$(Y_1 + \alpha Y_5)(Y_4 + \alpha Y_5) = (Y_2 - \alpha Y_5)(Y_3 - \alpha Y_5) \quad (3)$$

Equation (3) states that the network of Figure 6 behaves at balance as if an admittance αY_5 were connected in parallel with each arm. Of these fictitious admittances, two are positive and two negative with the choice of sign depending on the polarity of the two voltages involved.

Referring specifically to the bridge network used in the TYPE 1611, it is seen that, by proper choice of the coupling admittance and the voltage, an effective negative capacitance and negative parallel resistance are produced across the unknown terminals to neutralize the real capacitance and resistance that exist there. Simultaneously, the same effective capacitances are placed across the standard arm of the bridge, but the circuit capacitance there is 10,000 $\mu\mu$f, and the effect of the introduced admittance is for practical purposes negligible. Across the resistance arm R_A, there is also produced an effective negative capacitance. This serves partially to neutralize the real capacitance that exists across this resistance arm and to this extent improves further the accuracy of the circuit. Across the fourth arm B, the introduced capacitance is positive and would act to produce an error in

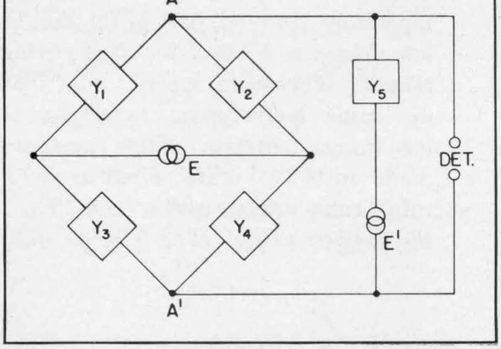


Figure 6. Equivalent circuit of the bridge and compensating circuit.

dissipation factor; but, as has been pointed out previously, capacitance is required across this arm to neutralize the losses in the standard capacitor. Therefore, in three of the four bridge arms, the effect of the added circuit is beneficial and in the fourth arm it is negligible.

The compensating circuit is effective for leakage through the bridge transformer and extraneous pickup in the amplifier to the extent that these can be represented by a voltage acting through a fixed impedance to the amplifier input. All of the extraneous effects together with the deliberately introduced voltage and admittance can be combined and represented as in Figure 6.

The equations as written have been in terms of voltage impressed on the bridge circuit, i.e., we have assumed a zero-impedance generator. Actually, of course, the magnitude and phase of the voltage impressed on the bridge change as the bridge arms are manipulated, thus making the compensation less effective. However, the shift of phase and magnitude is not significant except when the capacitance being measured is so large that the zero effects and the stray pickups are inconsequential.

Circuit Elements

Several of the components used in the bridge, in order to realize the accuracy and the direct-reading features, are a little unusual and are briefly described.

As previously noted, two standard capacitors are used, 0.01 μf for the four low ranges and 1.0 μf for the four high ranges. These are special units made up using polystyrene tape for the insulating material. They are each made up of two units paired to yield a total capacitance within ±0.25% of the desired value. The 0.01 μf unit is mounted in a low-loss phenolic case as used for TYPE 505 Capacitors. The elements making up the 1.0 μf unit are hermetically sealed in cylindrical metal containers of the type commonly used for electrolytic capacitors. Special heat treatment, aging, and impregnation result in a standard of unusually high leakage resistance and low dielectric losses.

Two rheostats (one for each standard capacitor) are used to balance the dissipation factor and are ganged to a common shaft. Each rheostat winding consists of two tapered sections with the resistances of these sections so chosen that the resulting scale permits precise readings at low values of dissipation factor while at the same time retaining the convenience of having the entire range on a single scale. The scale is pre-engraved and four adjustable shunt resistors are provided, one across each section of each rheostat to bring the actual resistance into agreement with the value required by the scale.

In some applications, as for instance in measuring many electrolytic capacitors in the range 1000–10,000 μf, dissipation factors in excess of 30% are encountered. For these values, provision is made in the bridge for switching into the standard arm additional fixed resistors which extend the range to 60%.

The variable resistor R_A, by means of which the capacitance balance is obtained, is the same unit that has been used previously in thousands of General Radio impedance and capacitance bridges. It is a tapered rheostat having a total resistance of approximately 11,000 ohms with the taper so chosen that the scale of the dial is essentially logarithmic. An adjusting plate and cam are built into the unit, which permits an adjustment of the position of the arm with respect to the dial at several points. As adjusted at the factory, the resistance in kilohms corresponds to the dial reading within ±0.5% over the main decade from 1.0 to 10.

Applications

This new bridge is suitable for use in the electric power industry for the testing, in the shop, of the dissipation factor of bushings and insulators and of the insulation of electrical equipment in general. In making measurements on such large, unshielded structures, voltages may be induced electrostatically which will shift the balance of the bridge. A switch is provided which permits reversing the test voltage with respect to the interfering voltage. The correct capacitance and dissipation factor can be computed from the two sets of readings taken for the two positions of the switch. In most cases the calculation consists merely of taking the arithmetic average of the direct and reverse readings.

In addition to its uses for the testing of insulators and of components, the Type 1611-A Capacitance Bridge should find wide application in chemical and plastics laboratories for measuring the dissipation factor and dielectric constant of both solid and liquid dielectric materials. The accuracy of dissipation-factor reading is adequate for all but extremely low-loss materials such as polystyrene, mica, and good electrical grade ceramics. Even for these materials, the dielectric constant can be evaluated accurately. It should be noted that the usefulness of electrical tests of this kind is not limited to insulating materials. Increasing use is being found for electrical measurements on materials destined for uses other than electrical. Product control and the checking of batch-to-batch uniformity of material on the basis of electrical constants is one application that shows increasing promise of usefulness.

—Ivan G. Easton

SPECIFICATIONS

Capacitance Range: 0 to 11,000 μf, covered by eight multiplier steps and an approximately logarithmic, direct-reading dial.

Dissipation-Factor Range: 0 to 60% (at 60 cycles), covered by a dial having an approximately logarithmic scale with a range of 30%, and a switch that adds a fixed value of 30%.

Capacitance Accuracy: ±(1% + 1 μμf) over the entire range of the bridge.

Dissipation Factor Accuracy: ±(2% of dial reading + 0.05% dissipation factor). Power Factor $= \frac{D}{\sqrt{1 + D^2}}$, where D = dissipation factor.

Sensitivity: The sensitivity is such that any capacitance in the range 100 μμf to 10,000 μf can be balanced to a *precision* of at least 0.1%.

Temperature and Humidity Effects: The readings of the bridge are unaffected by temperature and humidity variations over the range of room conditions normally encountered (65° F to 95° F, 0 to 90% RH).

A-C Voltage Applied to Capacitance under Test: The voltage impressed on the unknown capacitance varies from a maximum of approximately 125 volts at 100 μμf to less than 3 volts at 10,000 μf. The circuit is so arranged that a maximum of one volt-ampere of reactive power is delivered to the sample.

Polarizing Voltage: Terminals are provided for connecting an external d-c polarizing voltage. The maximum voltage that should be impressed is 500 volts.

One of the terminals is grounded so that any a-c operated power supply with grounded output can be used. The terminal capacitances of the power supply do not affect the bridge circuit.

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

Power Input: 15 watts.

Accessories Supplied: Line connector cord.

Mounting: Portable carrying case of luggage-type construction. Case is completely shielded to insure freedom from electrostatic pickup.

Vacuum Tubes: One each 6X5-GT, 6SJ7, and 6U5. All are supplied.

Net Weight: 30½ pounds.

Dimensions: (Width) 14½ x (depth) 16 x (height) 10 inches, overall, including cover and handles.

Type		*Code Word*	*Price*
1611-A	**Capacitance Test Bridge**	FORUM	**$375.00**

MISCELLANY

VACATION — During the weeks of July 26 and August 2 most of our employees will be vacationing. Manufacturing departments will be closed, and other departments will be manned by a skeleton staff. Every effort will be made to take care of urgent business, but repairs cannot be made, except in hardship cases. Our Service Department requests that shipments of material to be repaired be either scheduled to reach us well before this vacation period or delayed until afterward.

TECHNICAL PAPER — "Evaluation of Hysteresis Core Loss by Power Equations," by Horatio W. Lamson, at the 1948 Annual Meeting of the American Society for Testing Materials, Detroit, June 22.

RECENT VISITORS to our plant and laboratories include —

J. L. Tora, Instructor in Electrical Engineering, I. C. A. I., Madrid; Jose M. Rubiato, Assistant Professor, University of Madrid; Eugenio Méndez, Instructor, E. S. I. M. E., of Mexico, D. F.; and P. R. Desikochar, Engineer, All-India Radio, Bangalore.

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THE General Radio EXPERIMENTER

VOLUME XXIII No. 3 AUGUST, 1948

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

THE TYPE 1670-A MAGNETIC TEST SET

Also IN THIS ISSUE

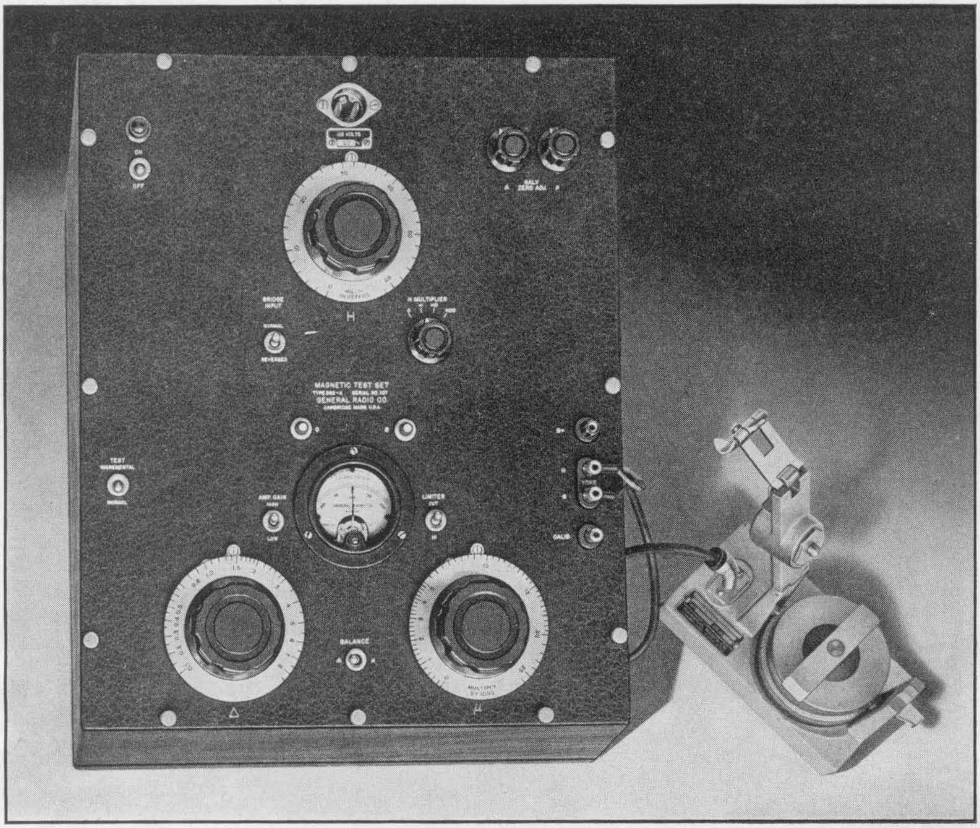
● **FOR A NUMBER OF YEARS** the accepted methods of determining the magnetic properties of laminated ferromagnetic alloys, used in transformer cores or in other electrical devices, have consisted of bridge or wattmeter measurements upon so-called Epstein squares, utilizing the various techniques described in the ASTM specification A34-46. The larger or 50-cm. Epstein square requires a sufficient number of strips of the specimen material, each 50 cm. by 3 cm., to aggregate 10 kilograms; while the smaller or 25-cm. Epstein square requires 2 kilograms of strips, each 28 cm. by 3 cm.

It is believed that many suppliers or users of these materials, together with other investigators, would find it convenient to make magnetic tests upon much smaller lamination samples, which, for example, might be cut in any desired direction from parental sheet stock or from small transformer core stampings already available. Furthermore, most of the ASTM procedures are not suited for measurements at the very low levels of induction, approaching initial permeability, which are frequently encountered in the cores of many inductors and transformers used in communication systems. The TYPE 1670-A Magnetic Test Set was developed to meet such a need by providing low-level 60-cycle measurements of permeability and core loss, using a single small strip or a few duplicate strips in parallel.

These midget test strips may have any uniform width up to ⅜″ and any length in excess of 2¼″. They are inserted into a helical coil, contained in the Test Yoke, where they interleave with, and become a diameter of, an assembly of laminated annular rings of a material having a high initial permeability. These rings have a sufficiently large composite cross-section so that a specific length of the specimen strip constitutes essentially the entire reluctance of the magnetic system. For specimens having an exceptionally high permeability, a simple correction can be made for the reluctance of the rings. By means of a spring-backed plunger in the hinged clamping bar, the laminated stack is subjected to a definite force which is more than sufficient to eliminate erratic air-gap errors. The assembly of this Test Yoke is a rapid and easy operation.

The inductor formed by the winding in this Test Yoke is inserted into a Maxwell bridge (see circuit diagram, Figure 2) which is energized at 60 cps through a high series resistance. This resistance insures that the current in the yoke and, hence, the magnetizing force H applied to the specimen is sinusoidal. Magnetic measurements are frequently made under conditions approximating a sinusoidal variation of induction or flux density B, rather than sinusoidal H. However, in the reference quoted (2), the author has advanced arguments favoring the latter procedure in bridge measurements. He has further demonstrated that, for a typical sample of silicon steel, the maximum discrepancy between the two methods occurred approximately at a level of normal induction corresponding to maximum μ and amounted only to about 8 per cent in the evaluation of μ and about 5 per cent in the evaluation of core loss. When the induction was less than one-half or exceeded twice the aforesaid level, the two procedures yielded essentially the same data. For comparative measurements between different specimens either procedure would be satisfactory. It should be noted that,

Figure 1. The Type 1670 Magnetic Test Set with its attached test yoke shown with clamp released.



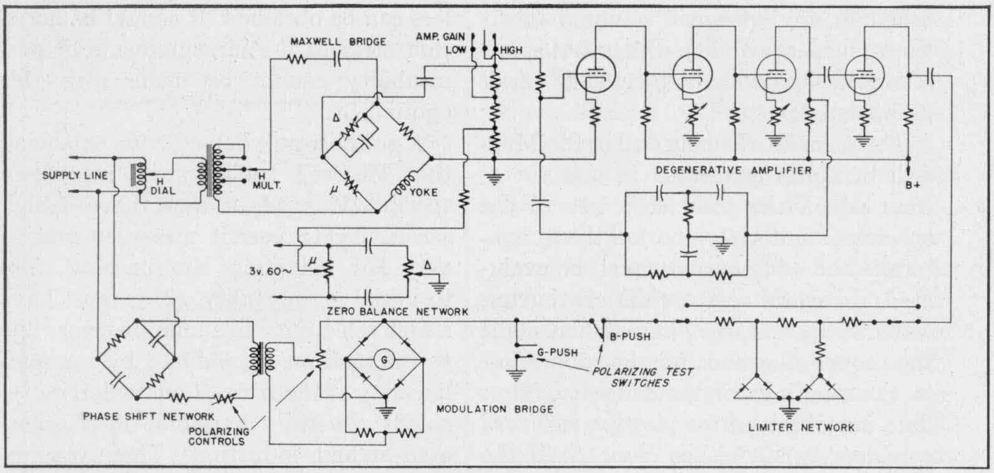


Figure 2. Schematic wiring diagram showing principal elements: Adjustable H supply, Maxwell bridge, degenerative amplifier, zero balance network, phase shift network, and modulation bridge.

while sinusoidal flux occurs in low-resistance power transformers energized by a low impedance source, if the combined resistance of the winding and the generator is appreciable compared to the reactance of the winding, then the flux will have a substantial harmonic content which may actually exceed the harmonic content of the exciting current, so that H becomes more sinusoidal than B.

By means of a Variac (H dial) and a supply transformer having 3 decade voltage steps (H multiplier), the magnetizing force applied to the specimen may be preset at any desired value from one millioersted to six oersteds, regardless of the existing impedance of the yoke. The H dial is calibrated in oersteds when the supply line voltage is 115 volts. Otherwise H is proportional to the existing supply voltage.

The two balancing components of the Maxwell bridge are rheostats. The μ-dial reading on one of these is proportional to the permeability of the specimen in terms of its arbitrary cross-section, which can be determined: (1) by micrometer measurements of width and thickness, or (2) from the length, mass, and density of the specimen strip or strips. In most cases it is the precision with which this cross-section can be evaluated that determines the accuracy of the absolute values of the magnetic data obtained. A knowledge of the cross-section is not required in *comparative* measurements between specimen strips having identical dimensions. The μ-dial is calibrated in terms of a specimen cross-section of 10 sq. mm., in which case the maximum scale reading indicates a permeability of 25,000. Higher permeabilities can be measured if an appropriate cross-section less than 10 sq. mm. is used.

With any preset value of H and the corresponding measured value of μ, the simultaneous induction B existing in the specimen may be computed as the product μH. In this manner data for the familiar B vs. H, μ vs. H, and μ vs. B curves may be obtained. The maximum induction which can be established in any specimen is equal to six times the permeability which that specimen possesses when subjected to an *H* of 6 oersteds.

The second balancing dial of the Maxwell bridge is calibrated in a factor Δ, from which the total core loss of the specimen material, due jointly to hysteresis and eddy currents, can be evaluated in watts per cubic centimeter, watts per pound, etc., having first made the proper allowance for the copper loss in the windings of the inductor. Thus data are obtained for plotting curves of core loss versus either *H* or *B*. If the resistivity of the specimen material is known, this core loss may be analyzed into its hysteresis and eddy-current components.

Provision is made for the introduction of a d-c or biasing current into the winding of the Test Yoke. Two additional components (not furnished) are then required — a rheostat of about 50 kilohms and a suitable milliammeter. The bias *H*, which is evaluated from the number of turns in the yoke winding and the measured biasing current, may have any value up to 2 oersteds. Sufficient d-c emf for producing a bias *H* up to 1.5 oersteds can be obtained directly from the rectifier system of the test set. For higher values this must be augmented by external batteries. Then, when any normal (peak a-c) value of *H* up to 6 oersteds is superimposed upon this bias *H*, the corresponding *incremental* values of permeability and core loss can be obtained. It should be noted that straight d-c measurements of permeability cannot be made with this equipment.

A suitable null detector for balancing this Maxwell bridge must meet two specifications: (1) it must have a high sensitivity to permit measurements at very low inductions approaching close to initial permeability, (2) it must have a high selectivity to eliminate errors due to the harmonics produced by the nonlinearity of the *B* vs. *H* characteristic — chiefly the third harmonic in *B* generated at high inductions. These requirements are met by the use of a four-stage amplifier tuned sharply to 60 cycles by a degenerative *R-C* network.

The use of a sensitive 60-cycle amplifier presents certain difficulties, especially when incorporated into equipment which is energized at the same frequency. This is due to an unavoidable stray pickup of minute 60-cycle voltages induced either by components of the equipment itself or by the 60-cycle electromagnetic fields present, although frequently unrealized, in all laboratories supplied with a-c power. Such stray voltages, greatly amplified, would, of course, result in false indications of bridge balance. Two auxiliary controls in a zero-balance network permit a small voltage of adjustable phase and amplitude to be introduced into the amplifier so as to counteract any stray pickup and to make the amplifier response dependent solely upon the output voltage of the unbalanced bridge.

The null balance indicator is a center-scale-zero galvanometer. This is in-

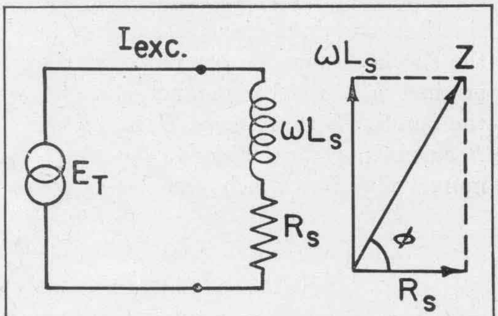


Figure 3. Misleading representation of an a-c iron-cored inductor. Due to core loss the full exciting current does not produce magnetizing force and flux.

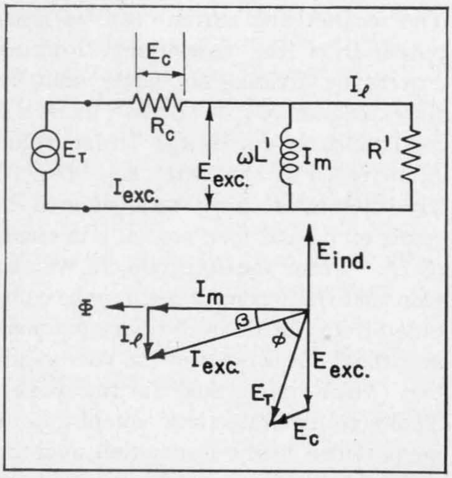


Figure 4. Correct representation of an iron-cored inductor showing the subdivision of exciting current into its magnetizing and loss components.

serted into a modulation bridge composed of four rectifier elements and polarized by one or the other of two voltages which have a phase displacement of 90 degrees between them. By the operation of a two-position switch, which changes the resistors of the phase-shift network, this *polarized* indicator can be made predominantly responsive to the manipulation of either the μ-dial or the Δ-dial in the Maxwell bridge. Such a polarized detector permits a more convenient and rapid balancing of the bridge than would be possible with the conventional non-polarized detector, especially with the "sliding zero" which this bridge exhibits in measuring an inductor having a relatively low Q. Furthermore, if only permeability and not core-loss data are desired, a precise setting of the Δ-dial is not required in obtaining an accurate balance of the μ-dial.

In addition, this indicator is *directional*, meaning that the displacement of the galvanometer needle, left or right of center, indicates the direction in which either the μ-dial or the Δ-dial should be turned to approach balance, which is analogous to the directional feature possessed by the galvanometer used in balancing a d-c Wheatstone bridge.

A limiter network, which contains non-linear elements in its shunt branches can be inserted, at will, between the amplifier and the polarized indicator to give the latter a quasi-logarithmic response. This prevents the galvanometer needle from going off-scale with a badly unbalanced bridge and eliminates the necessity for monitoring the gain of the amplifier as balance is approached. These several features combine to give a useful null balance system.

The Maxwell bridge, in common with many other bridges, evaluates an inductive impedance in terms of its series components R_s and $j\omega L_s$, see Figure 3. The μ-dial setting is determined by L_s, while the dissipation factor D, defined as the ratio of R_s to the series reactance ωL_s, determines the setting of the Δ-dial. If a resistance R_c, representing the copper losses of the winding (the d-c resistance when the frequency is as low as 60 cycles), is subtracted from R_s, the remainder multiplied into the square of the exciting current I^2_{exc} gives the core loss in the specimen.

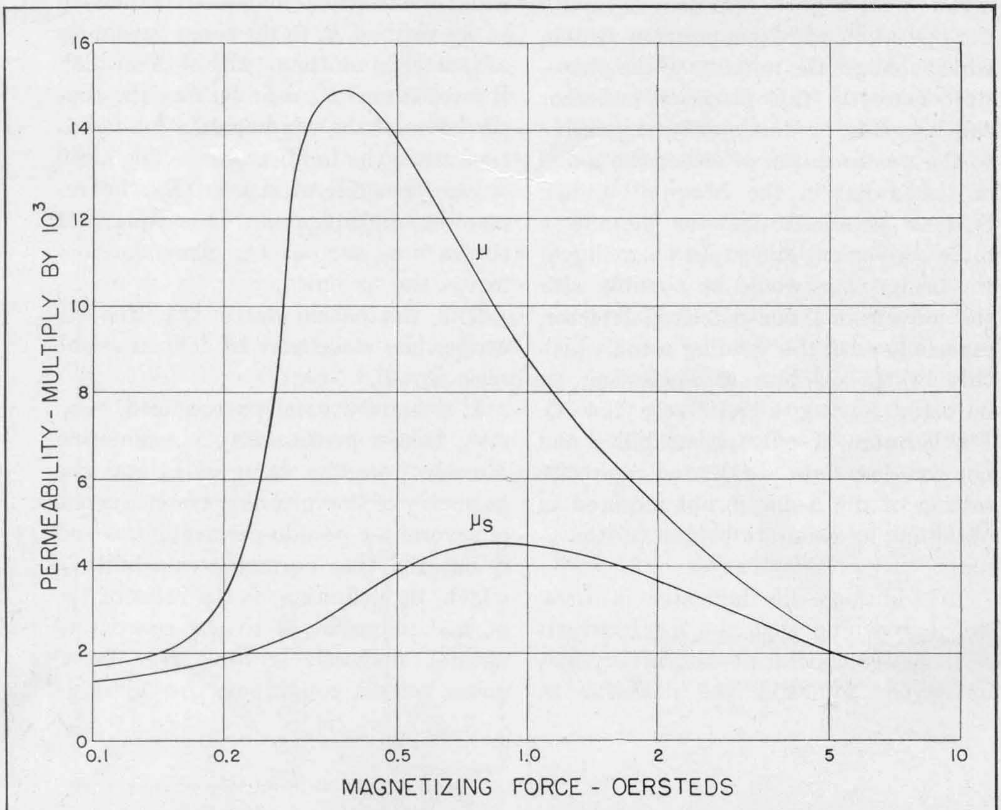
The dissipation factor D_c due to copper loss alone may be defined as the ratio $R_c/\omega L_s$.

It is not universally recognized, however, that a permeability μ_s, computed directly from the value of L_s and the geometry of the magnetic system, is one of several a-c pseudo-permeabilities and is not the true normal permeability μ which, by definition, is the ratio of the normal induction B to the co-existing normal magnetizing force H. While, under certain conditions, this μ_s value is a perfectly legitimate parameter for intercomparing similar ferromagnetic specimens, its magnitude is a function of core loss and, hence, depends upon the thickness of the specimen lamination even though the flux has complete penetration.

To evaluate the true normal permeability, which is a unique parameter of the specimen material, the inductor must be analyzed as depicted in Figure 4. The exciting current I_{exc}, flowing in the copper-loss resistance R_c, divides into two quadrature components: (1) magnetizing current I_m which flows in the inductance L' to create the H which produces the flux, (2) the loss current I_l which flows in the resistance R' representing the core losses I_l^2R' identical with the previous value I_{exc}^2 $(R_s\text{-}R_c)$. The magnetizing current can be computed from the measurable exciting current by dividing the latter value by the expression: $\sqrt{1 + (D\text{-}D_c)^2}$; while the inductance L' equals the bridge value L_s increased by the factor $1 + (D\text{-}D_c)^2$. True normal H must be evaluated in terms of I_m and true normal μ in terms of L'. From the foregoing it will be seen that these true values may be computed from the Maxwell bridge readings and the d-c resistance of the yoke winding (which is marked in the yoke). Tables to facilitate these computations are provided in the instruction manual.

The discrepancy which may exist between these pseudo and true normal values is illustrated in Figure 4, which gives data for a somewhat extreme case encountered in a sample of silicon steel

Figure 5. Comparison of true normal permeability μ with a pseudo-permeability μ_s evaluated in terms of series inductance.



having rather high core losses and, hence, requiring substantial values of the correction factor. The curve marked μ_s represents the pseudo permeability evaluated directly from L_s (Figure 3) and plotted against peak values of pseudo H_s evaluated from the full exciting current. This curve illustrates data obtained directly from the Maxwell bridge and which have been used extensively in the evaluation of "a-c permeability vs. H." The curve marked μ indicates the true normal permeability evaluated from L' (Figure 4) plotted against the true values of normal H computed in terms of the magnetizing current.

A detailed discussion of the theory of this Magnetic Test Set and a general analysis of a-c magnetic measurements can be found in two I.R.E. papers by the author.[1, 2] Reprints of these papers may be obtained upon request.

In conclusion it may be noted that this test set is a useful bridge for the 60-cycle measurement of *any* inductor (with or without d-c polarization) provided that its inductance does not exceed one henry and its 60-cycle Q value is less than 13.5. By simple formulae the μ and Δ-dial readings at balance may be converted into corresponding values of L_s and Q.

— HORATIO W. LAMSON

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[1] "A Method of Measuring the Magnetic Properties of Small Samples of Transformer Laminations," *Proc. I.R.E.*, Vol. 28, pp. 541–548, December, 1940.

[2] "Alternating Current Measurements of Magnetic Properties," *Proc. I.R.E.*, Vol. 36, pp. 266–277, February, 1948.

SPECIFICATIONS

Range of Magnetizing Force: The 60-cycle normal magnetizing force is adjustable from one millioersted to 6 oersteds (gilberts per centimeter) for a line voltage of 115 volts. A biasing magnetizing force (d-c) up to 2 oersteds can also be applied. The necessary d-c power, up to 1.5 oersteds, can be obtained from the internal power supply of the test set.

Permeability and Core-Loss Range: The range for permeability and core-loss measurements varies with the cross-section area. For a sample cross-section of 10 sq. mm. full scale on the μ dial is 25,000. The permeability and core loss of any ferromagnetic sample can be measured if a sample of proper cross-section is chosen. It may sometimes be necessary to calculate corrections for high-permeability materials.

Accuracy of Measurement: The accuracy of data obtained with this instrument is chiefly determined by the precision with which the cross-section of the specimen is known. Similar samples of identical cross-section can be compared, at any given H, with an accuracy of 1 to 2 per cent.

Power Supply: 115 volts, 60 cycles; by a change of connections on the power transformer primary, the instrument can be operated from a 230-volt line.

Power Input: 90 watts.

Tubes: 2 6C8-G, 1 6X5-G, and 1 0D3/VR150.

Accessories Supplied: Test yoke and line cord.

Accessories Required: When a d-c magnetizing force is applied, a milliammeter and a rheostat for varying the dc are required. The TYPE 371-A 50,000Ω Rheostat is suitable when the internal voltage source is used.

Mounting: The test set, exclusive of the test yoke, is housed in a walnut cabinet with sloping front panel.

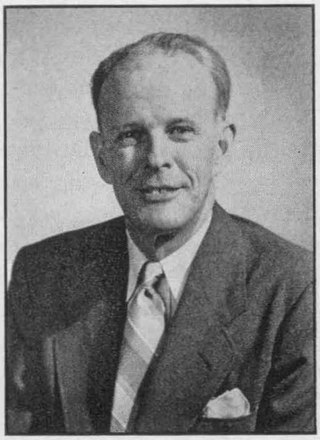
Dimensions: Test set, (width) 16 x (depth) 18 x (height) 10 inches over-all; test yoke, 8 x 4 x 5½ inches.

Net Weight: Test set, 44 pounds; test yoke, 10 pounds.

Type		*Code Word*	*Price*
1670-A	**Magnetic Test Set**	AFIRE	**$585.00**

MISCELLANY

Melville Eastham



John M. Clayton

Harold B. Richmond

HONORS — To Harold B. Richmond, Chairman of the Board, and to Melville Eastham, Chief Engineer, the Medal for Merit, for exceptionally meritorious conduct in the performance of outstanding services to the United States. The award to Mr. Richmond was made for his work as Chief of the Guided Missiles Division of National Defense Research Committee, in which capacity he was responsible for all wartime activities in this field, resulting in the development of the missiles Azon, Razon, and Felix.

The medal was awarded to Mr. Eastham for his work as a member of the Microwave Committee of the NDRC and later as Expert Consultant to the Office of the Secretary of War, and, in particular, for his guidance of the Loran development program.

—To John M. Clayton, Advertising Manager, the Meritorious Civilian Service Award, for Outstanding Service to the Navy. The award was made for the outstanding effectiveness with which he handled procurement problems of the Radio Division of the Naval Research Laboratory during the war.

RECENT VISITORS to our plant and laboratories include Mr. Jack Smith of Warburton-Franki, Ltd., distributors for General Radio instruments and manufacturers of Variacs in Melbourne, Australia; Mr. J. A. Jones and Mr. Raoul Rago, of the Direction General de Fabricaciones Militaires, Buenos Aires, Argentina; Mr. Antonio Millan, Instituto L. Torres Quevedo, Madrid, Spain; and Mr. Ronald E. Burgess, National Physical Laboratory, England.

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VOLUME XXIII No. 4 SEPTEMBER, 1948

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A FREQUENCY MONITOR FOR TELEVISION VIDEO TRANSMITTERS

Also

IN THIS ISSUE

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● **TELEVISION TRANSMITTERS** need two types of frequency monitor, an f-m type for the audio channel and an a-m type for the video. Monitoring requirements for the audio transmitter are met by the Type 1170-A F-M Monitor[1], which is specifically designed for both television and f-m broadcast applications. For video transmitters, however, agreed standards of performance have only recently become available.

When only a very few stations were on the air, the general requirements were met reasonably well by the general-purpose Type 1175-A

Figure 1. Panel view of the complete video frequency monitor.

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[1] "Type 1170-A F-M Monitor for Broadcast and Television Services," *Experimenter*, October, 1947.

Frequency Monitor[2], and from the experience gained in early installations, changes in design were made to improve performance and to extend the range to all television channels[3]. Recently, however, monitoring requirements for television have been more definitely established, since the Federal Communications Commission has prepared a proposed draft of "Parts 15 and 16 of the Rules and Standards Concerning Television Broadcast Stations." To meet these requirements, the General Radio TYPE 1182-T Television Video Monitor has been designed.

This video monitor, like its general-purpose predecessors, consists of a monitor unit and a deviation indicator unit. The over-all principle of operation is the same, except that the frequency-offset method of indication is used, in order that the direction of the deviation be indicated. All modifications have been worked out in cooperation with transmitter manufacturers, who have been supplying these monitors as standard equipment with television transmitters.

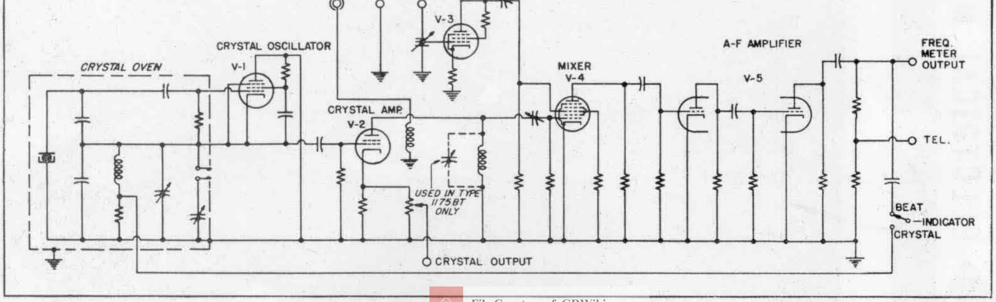
The monitor unit is a TYPE 1175-BT Frequency Monitor[3], which consists of a stable crystal oscillator, a harmonic generator, a mixer, and an output amplifier. The elementary circuit is shown in Figure 3.

The frequency indicator is a TYPE 1176-AT Frequency Deviation Meter, a modification of the standard TYPE 1176-A Frequency Meter. The beat frequency to be measured is passed through a series of amplifiers and clippers to develop a square wave, which is then applied to a pulse counter circuit. Unidirectional pulses are applied to a d-c microammeter, whose deflection is proportional to the number of pulses per unit time, and hence to the frequency. Figure 4 shows the basic circuit.

ZERO BEAT VS. OFFSET-FREQUENCY MONITORING

When a monitor is designed for maximum flexibility to cover a wide range of applications, the so-called "zero-beat" method of frequency monitoring offers many advantages. This method consists essentially of measuring the difference frequency between a crystal fundamental (or harmonic) frequency, which has been adjusted to the desired channel frequency, and the actual transmitter frequency. The difference frequency can be indicated on a multi-range, direct-reading frequency meter, which provides a means for determining the transmitter frequency error over a considerable range. This system is particularly useful when the applications involve monitoring over a wide range of carrier frequencies and where the transmitter frequency tolerance is expressed as a varia-

Figure 2. Elementary schematic diagram of the monitor unit.



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[2] "A Versatile Monitor for Use from 1.6 to 150 Megacycles," *Experimenter*, February, 1947.

[3] "A Frequency Monitor for Television Video Transmitters and Other A-M Services," *Experimenter*, September, 1947.

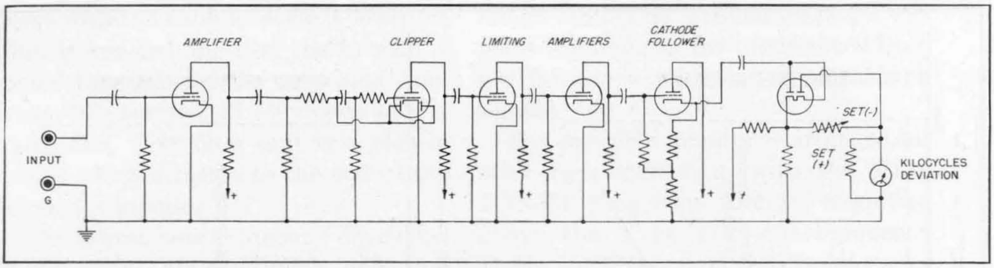

Figure 3. Elementary schematic diagram of the frequency deviation meter.

ble percentage of the carrier frequency.

Disadvantages of such an arrangement are (1) the lack of a direct indication of the direction of the deviation, that is, whether the transmitter frequency is high or low, and (2) the limited region near zero frequency error, where the system becomes inoperative since it lacks a d-c response. The first of these limitations can be taken care of by providing for a temporary frequency shift of the crystal oscillator, and the second by limiting the low frequency cutoff of the frequency measuring circuit to a value comparable to the instantaneous stability of the transmitter.

For specific monitoring applications such as standard broadcast transmitters, continuous monitoring is essential. A direct-reading meter scale, calibrated in plus and minus channel frequency error, has become the accepted practice. This can be achieved by employing the "offset crystal-frequency" principle, which differs from the zero-beat method only in that the monitoring crystal frequency is adjusted to give a known frequency difference between one of its harmonics and the required transmitter channel frequency. Thus, when the transmitter is exactly on channel frequency, the beat-frequency produced within the monitor detector circuit will be this predetermined value. If the crystal offset frequency amounts to slightly more than the half-scale value of the calibrated meter scale, then the beat frequency will not be required to pass through zero over any part of the normal range of the monitor. A comparison between the two monitoring systems is shown in Figure 4.

In the zero-beat method, the frequency meter indicates the beat frequency directly and hence cannot distinguish between positive and negative transmitter frequency shifts. The offset method employs a meter scale calibrated in plus and minus deviations about a zero center. The actual beat frequency corresponding to the zero center point of the deviation meter is mainly determined by the deviation range, the sensitivity requirements of the system, and the low frequency response. As the offset frequency increases, the sensitivity decreases because the $\Delta f/f$ becomes smaller. Very low beat frequencies should be avoided to minimize interference from power-line hum or television synchronizing signals.

By setting the crystal frequency on the *low* side of the desired transmitter channel frequency, the direction of the beat frequency will follow the direction of the actual transmitter shift. Should the transmitter frequency be incorrectly set on the wrong side of the crystal frequency, this condition will be reversed. This provides a check to determine proper operation. An alternative test can be made by shifting the crystal

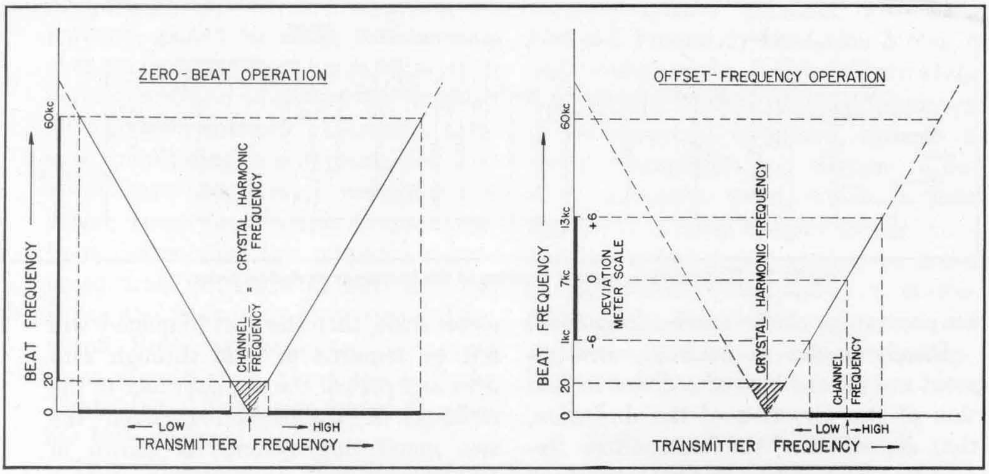

Figure 4. Graphical comparison of zero-beat and offset-frequency monitoring.

frequency itself, by a small amount. In this case, *decreasing* the *crystal* frequency should produce an *increase* in the beat frequency.

DETAILS OF THE 1182–AT TELEVISION FREQUENCY MONITOR

The Type 1176-AT Frequency Deviation Meter is designed especially for television video frequency monitoring. When used in conjunction with the Type 1175-BT Frequency Monitor as a part of the Type 1182-AT Television Monitor, it provides a direct indication of the video transmitter frequency deviation from the assigned channel frequency. In accordance with FCC proposals, two alternative scale ranges are available, 3-0-3 kc for TV channels 2-6 inclusive, and 6-0-6 kc for TV channels 7-13 inclusive. The scales, shown in Figure 2, are identical to those used on the Type 1170-A F-M Monitor[1], which has found wide application as a frequency and modulation monitor for television aural transmitters.

The 3-0-3 kc range employs an offset crystal harmonic frequency of –3500 cycles; the 6-0-6 kc range has just twice this value, or –7000 cycles. To obtain a full-scale range of 6 kc, electrical suppression is used in the d-c meter circuit to balance out the meter current corresponding to an input signal of 500 cycles. A standard d-c microammeter meter is used, having normal left deflection for zero d-c current, but the meter is calibrated with a zero center scale. The scale zero is made to coincide with the crystal harmonic offset frequency, thus indicating correct transmitter channel frequency when this beat frequency is obtained. Internal adjustments are provided to calibrate the meter at both ends of the scale. Scale ranges can be easily interchanged by reversing the meter scales which have alternative calibrations on either side, and making one internal connection change. Because of the different offset frequency, a change in the monitor crystal frequency is also required.

Since the highest beat frequency developed by the monitor is 13 kc, obtained when operating at the +6 kc deviation

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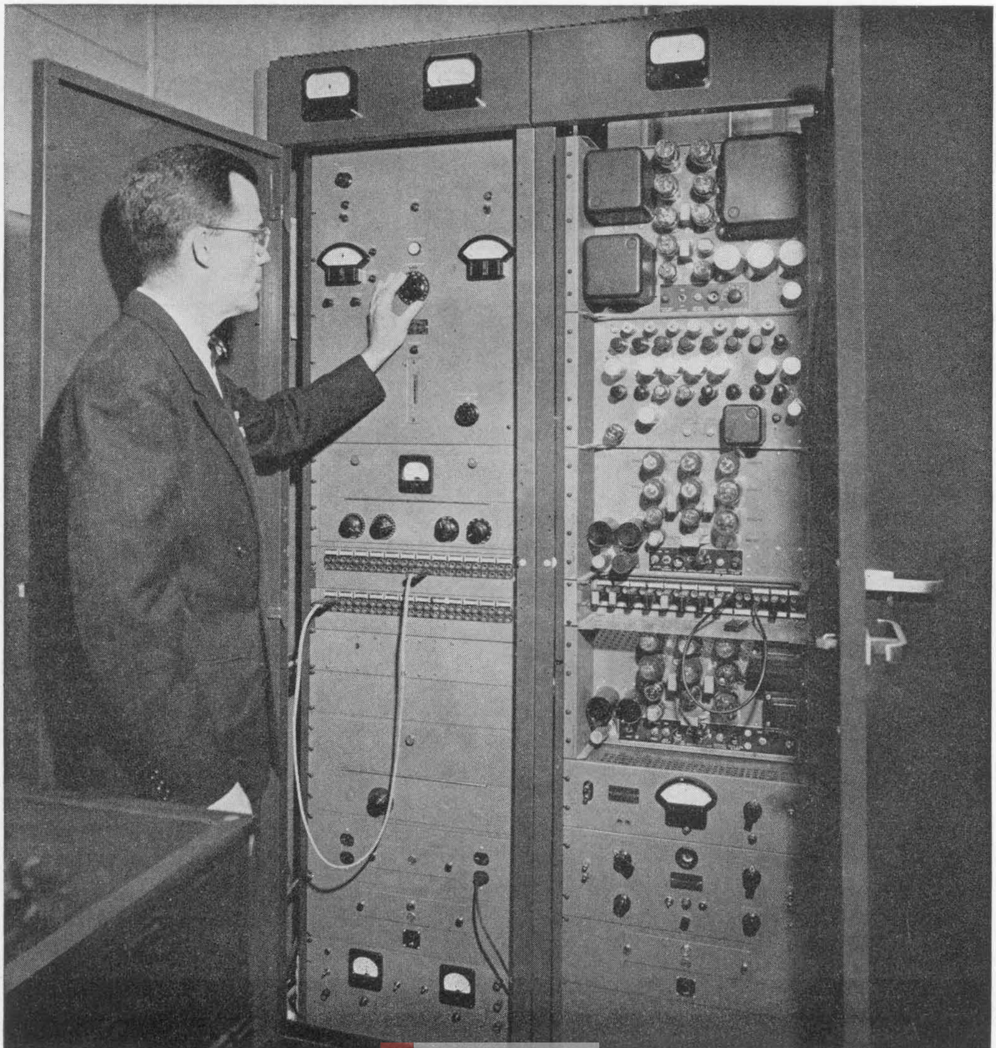
[1] Loc. cit.

point with the 6-0-6 kc scale, a low-pass filter is inserted ahead of the frequency meter to eliminate the television line-frame frequency of 15,750 cycles and its harmonics. This filter unit is a plug-in device which attaches to the rear of the frequency monitor.

The lowest beat frequency developed is 500 cycles, obtained when operating at the –3 kc deviation point with the 3-0-3 kc scale; hence the response of the frequency deviation meter is made to fall off rapidly for frequencies below this point to minimize the interference from low frequency television synchronizing signals.

The complete monitor is arranged for relay rack mounting, with the Type 1175-BT Frequency Monitor mounting above the Type 1176-AT Frequency Deviation Meter, and thus provides for easy access to the crystal oven. Spare crystals can be mounted within the oven and selected by means of a panel switch.

Figure 5. General Radio television monitors installed at WBAL-TV, Baltimore. The video monitor is at the lower right, the f-m monitor at the upper left. Shown operating the latter is R. S. Duncan, Chief Engineer.



Three tuning adjustments are reached through holes in the chassis. It is not necessary to change these adjustments, except possibly during the initial installation, and the settings are not critical.

To bring the monitor readings into agreement with an external frequency check, a small adjustment has been provided which will shift the crystal frequency by a small amount. For reasons of maximum stability, this control is located within the temperature-controlled crystal oven and can be reached through an access hole in the side of the oven.

—C. A. Cady

FEATURES OF THE TELEVISION VIDEO MONITOR

1. High sensitivity, requiring negligible power from transmitter.
2. High-stability, temperature-controlled, crystal oscillator.
3. Operates directly from transmitter-modulated signal. Indication is unaffected by television video modulation.
4. Direct indication of frequency drift from assigned channel.
5. Panel switch for stand-by operation.
6. Provision for remote frequency deviation meter.
7. Panel indicator to check crystal oscillation.
8. Panel indicator to check monitor output.
9. Adjustment to set monitor in agreement with frequency measuring service.
10. Spare crystal positions (3) selected by panel switch.
11. Switch for check of crystal offset frequency.
12. No critical adjustments.

SPECIFICATIONS

Transmitter Frequency Range: 160 to 220 Mc.

Deviation Range: 3-0-3 kc for television channels 2 to 6 inclusive; 6-0-6 kc for channels 7 to 13 inclusive.

Accuracy: Crystal frequency, when monitor is received, is within ±10 parts per million (0.001%) of specified channel frequency. Center-frequency reading can be adjusted to bring monitor into agreement with frequency measuring service.

Stability: ±0.001%.

Input Impedance: High-impedance circuits for channels 2 to 6, coaxial line for channels 7 to 13. Complete coupling directions are included in the operating instructions.

Vacuum Tubes:

1—6AC7
1—6AG7
1—6E5
1—6SN7GT
1—6H6
2—6SJ7
2—6X5
1—6BE6
1—OD3/VR150
1—6SQ7
1—6J5
1—OA3/VR75
1—6V6
1—Amperite 3-4

Accessories Supplied: All tubes, connecting cable, and power cords; plug-in filter; one quartz plate.

Mounting: 19-inch relay rack panel. Walnut end frames are available for table mounting; see price list below.

Dimensions: Panel, 19 x 12¼ inches, overall; depth behind panel, 11¼ inches, overall.

Net Weight: 42 pounds.

Type		*Code Word*	*Price*
1182-T	**Video Frequency Monitor**	ALERT	**$675.00***

*For General Radio Black Crackle Panel Finish.

This instrument is covered by U. S. Patents Nos. 1,967,185; 2,012,497; 2,362,503; and is licensed under patents of the American Telephone and Telegraph Company, and under patents and patent applications of G. W. Pierce.

VARIACS USED WITH INCANDESCENT LAMPS AND OTHER RESISTIVE LOADS

Variac adjustable transformers should never be subjected to load currents in excess of five times their rated current. Unfortunately, certain loads draw large inrush currents when cold. A modern 120-volt incandescent lamp measured by the writer showed a ratio of cold-to-hot current of 13.9 to 1. Heating devices exhibit similar though less severe characteristics.

Variac instructions specifically state, "Always set Variac to zero before switching to avoid surges." Apparently, some Variac users have interpreted this as having reference to reactive loads (predominantly inductance or capacitance), but it applies just as fully to resistive loads of the type under discussion. If the instructions are followed, full voltage will not be applied to the cold load, and, as the Variac is turned up to line voltage from zero, the load will have time to heat sufficiently to limit the inrush current to a reasonable value.

If the load must be switched rapidly off and on at full voltage, it should be disconnected from the Variac and connected to the line before so doing. A "snap-action" toggle switch or relay may be used to change the load from Variac to line. Such switching introduces no perceptible flicker and avoids throwing a cold load across the full Variac output. The Variac setting should, of course, be zero whenever the load is cold.

A choke coil in series with an "inrush" load will serve to limit the surge without seriously reducing the load operating voltage. The choke should have an inductance sufficient to limit the load current to five times the Variac rated current even with a zero resistance load. Under the worst conditions such a choke will reduce the load voltage by not more than 4 per cent. The inductance may be calculated by the formula

$$L = \frac{\text{Load Volts}}{31.4\, f\, I_R}$$

where L = inductance in henrys

f = frequency in cycles/second

I_R = Variac rated current in amperes

The choke must be large enough to maintain its inductance at five times the rated current and must be capable of carrying the load current continuously. Unfortunately, chokes that will effectively limit the output of the larger Variacs are, themselves, bulky and expensive, a factor that often makes their use impracticable.

Another alternative is to reduce the load until its inrush current falls within the Variac limitation, but this reduces the load unduly as compared with either the choke or switch method.

— Gilbert Smiley

MISCELLANY

TELEVISION MODEL OF THE F-M MONITOR

The General Radio Company now supplies complete monitoring equipment specifically designed to meet the requirements of television broadcasting. For monitoring the video transmitter, use the Type 1182-T Television Video Monitor described above. For the audio transmitter, use the Type 1170-AT F-M Monitor, the television model of the F-M Monitor described in the *Experimenter* for October, 1947.

RECENT VISITORS to our plant and Laboratories — *From India:* Mr. Gopal Chandra Sen, Lecturer in Mechanical Engineering, College of Engineering and Technology, Bengal; Mr. Ram Vepa, Indian Institute of Science, Bangalore; and Mr. P. Chawla, New Delhi.

— *From Switzerland:* Dr. Julius E. Weber, Swiss Aluminum Co., Ascona; Dr. E. Bruno Siegrist, Geneva; and Mr. Gerald Nenn, Geneva.

ERRATUM — In our August issue it was stated that Warburton-Franki, Ltd., manufacture Variacs in Melbourne, while the city that should have been mentioned is Sydney. The editor offers his apologies to Australian readers of the *Experimenter* and hastens to assure them that he knows these two cities are some 400 miles apart.

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 address, type of business company is engaged in, and title or position of individual.

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VOLUME XXIII No. 5 OCTOBER, 1948

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A RADICALLY NEW COAXIAL CONNECTOR FOR THE LABORATORY

Also IN THIS ISSUE

●THE NEW TYPE 874 Coaxial Connector has been demonstrated to a number of engineers during the past two years both at our laboratories and at radio engineering conventions[1]. Their approval has been almost unanimous. Two important characteristics are combined in this connector, unparalleled convenience of use and excellent electrical performance. Intended for use at all frequencies from 0 to over 4500 Mc, it is applicable wherever a shielded connection is needed. The characteristic impedance is 50 ohms, and the electrical uniformity, important at ultra-high frequencies, is excellent. It is available for use in coaxial parts and systems of all kinds, such as slotted lines, stubs, attenuators, bolometers, and certain flexible cables; and on panels of instruments such as oscillators, signal generators, receivers, bridges, and amplifiers.

The unique feature of the design is that two identical TYPE 874 Connectors plug smoothly into each other without any intermediate elements, and the complication of male and female assemblies is completely avoided. A strong friction grip is made by the multiple, spring-loaded contacts, so that no locking means is required, and connections

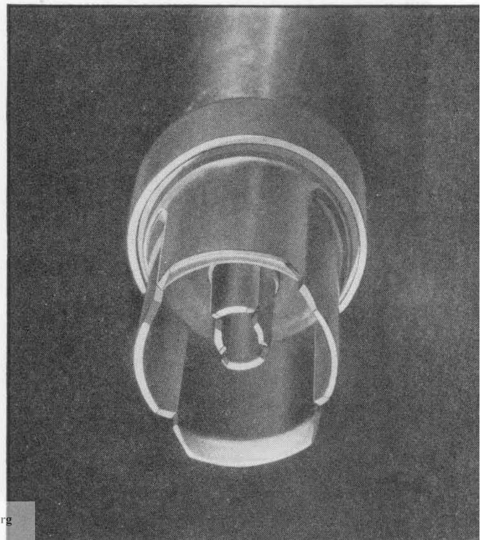

Figure 1. Type 874 Coaxial Connector (approximately 1½ times normal size).

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[1] W. R. Thurston, "Coaxial Elements and Connectors," Proceedings of the National Electronics Conference, 1947, pp. 97–108.

can be made and broken very quickly and conveniently. The basic elements of the connector are an inner conductor, an outer conductor, and a supporting polystyrene bead. Figure 1 shows one of these connectors assembled at the end of a rigid, 50-ohm, air-dielectric, coaxial line. The inner and outer conductors are similar in principle; each is essentially a tube with four longitudinal slots in the end and with two opposite quadrants displaced inward. To make a joint, two connectors are plugged together so that the undisplaced quadrants of one connector overlap the displaced quadrants of the other. Figure 2 is a cross-section sketch of a joint in which the elements of one connector are shaded dark and those of the other light. The mutual overlapping referred to can be seen, as well as the resultant circularity of the joined conductors.

Some connector parts and the basic, cable, and panel forms of the connector are shown in Figure 3. Shown at the left is the inner conductor, the polystyrene bead, the outer conductor, and the coupling nut, which holds the connector together. Next in order are a rigid, 50-ohm, air-dielectric line prepared as described in the specifications below to take a connector, two connectors assembled at the ends of a 10-cm. air line, and a connector joint between two air lines. The cable and panel forms of the connector are on the right of the figure. The cable connector is made from the basic connector elements by the addition of two tapered pieces for the cable connection and a rubber guard, and is shown in detail at the left of Figure 4. The panel connector is obtained by the addition of the panel adapter, which holds the basic connector elements on a

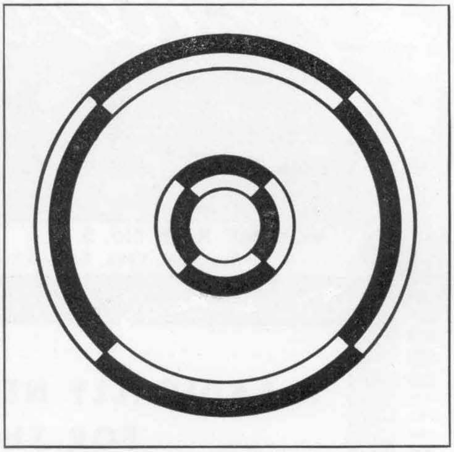

Figure 2. Cross-section sketch of overlapping joint.

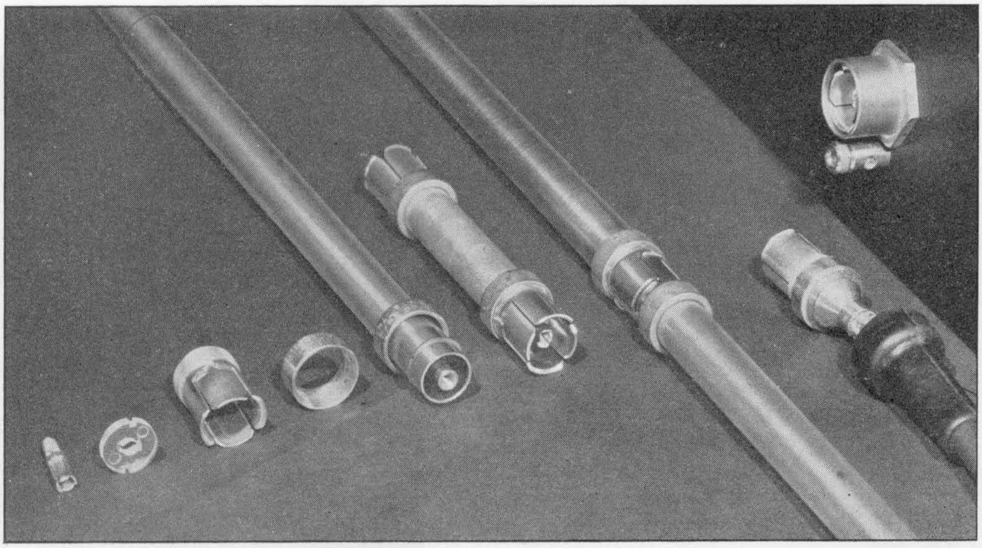

Figure 3. Array of connectors and connector parts.

panel. The panel connector also includes two tapered pieces for a cable connection. A useful feature of the Type 874 Connector is that the axial hole in the inner conductor (see Figure 1) will receive a Type 274 Plug, the so-called "banana" type. Therefore, the addition of a binding post beside a panel connector, as shown at the right of Figure 3, makes it possible to plug a Type 274 Double-Plug Connector directly into the Type 874 Panel Connector without an adapter. This feature is particularly valuable for wide-frequency-range instruments, because at high frequencies a shielded coaxial connection is necessary, while at low frequencies an unshielded, two-wire connection is often allowable and more convenient.

One of the most interesting electrical characteristics of a coaxial connector is its performance at ultra-high frequencies, as measured by the reflections it causes in an otherwise matched line. To make reflections small, the characteristic impedance of the Type 874 Connector has been made the same at all points, and discontinuities are minimized. Figure 4 is a cut-away view of two connectors plugged together. The joint portion of the connection, at the right of the figure, can be seen to be a fairly uniform coaxial section, as also shown by Figure 2, and the effective diameter ratio of the conductors in the bead sections is so chosen as to maintain the 50-ohm impedance level. The input standing-wave ratio of a connector joint terminated by a 50-ohm matched line was measured for each of twenty separate connections made from among a group of twelve pilot-production-lot connectors over the frequency range from 1000 Mc to 4500 Mc. The solid curve of Figure 5 shows for each frequency the average of the twenty measured values of standing-wave ratio expressed in decibels[2], and the dashed curve shows for each frequency the maximum of the twenty values. The average is less than 0.3 db over the entire range. and the worst single value measured is 0.8 db for one pair of connectors at 3500 Mc. The maximum error of any single meas-

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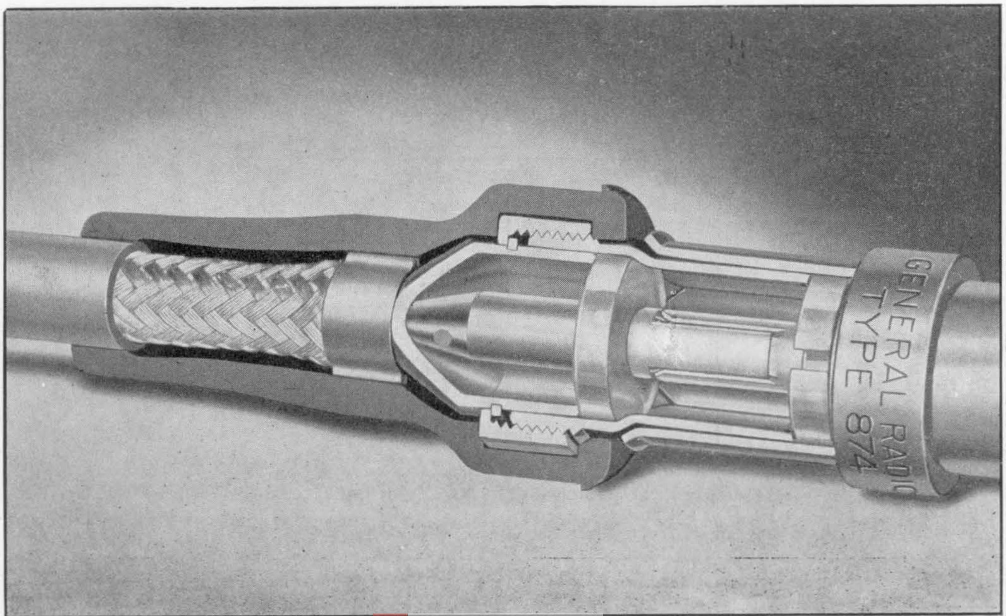
[2] Standing-wave ratio expressed in decibels = 20 $\log_{10} S$, where S is the numerical voltage-standing-wave ratio.

Figure 4. Cut-away view of basic connector joined with cable connector.



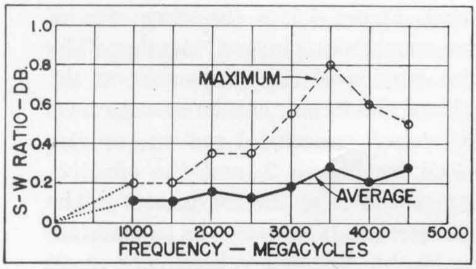


Figure 5. Type 874 Connector standing-wave ratio as a function of frequency.

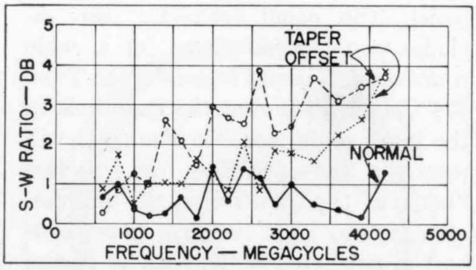
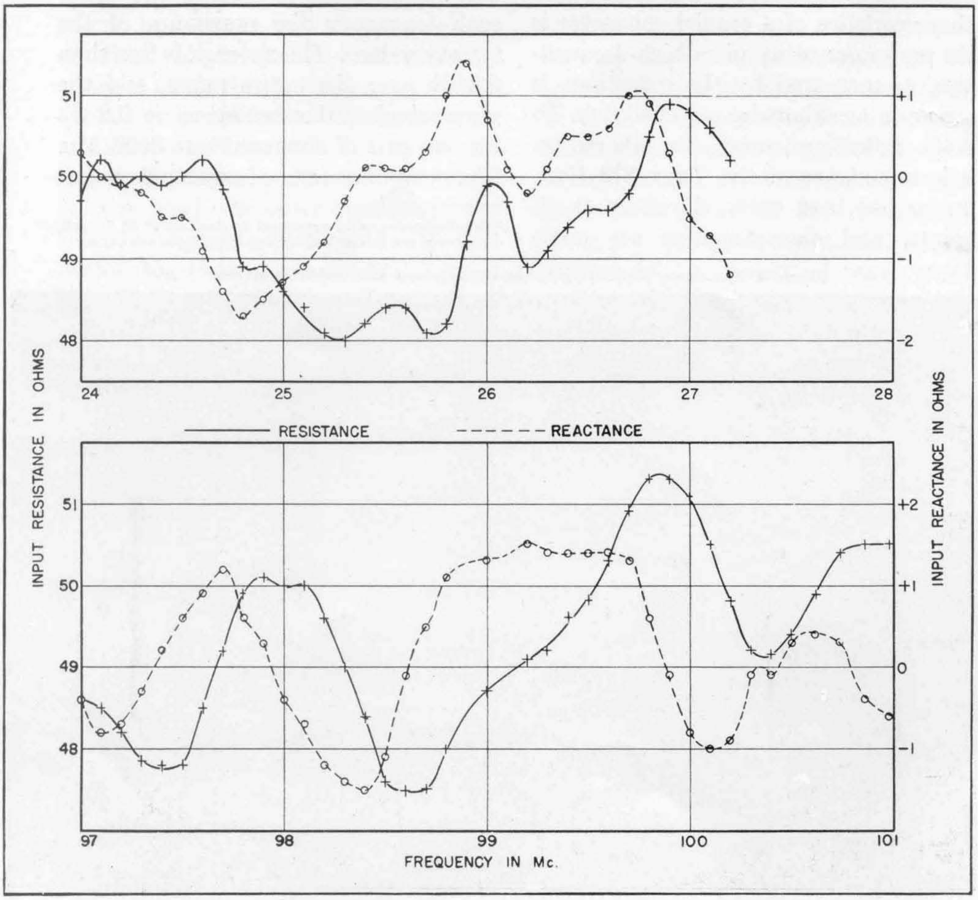


Figure 7. Input standing-wave ratio of Type 874-C Cable Connector terminated by long cable.

urement was less than ±0.2 db, except at 4500 Mc where it was ±0.3 db. These reflections are so small that they can be neglected in most measurements of moderate accuracy, and so this connector is well suited for use on ultra-high-frequency measuring equipment. External fields from the connector are so low as to be negligible in all the usual applications.

The basic connector fits a ⅝-inch O.D. tube, and the coupling nut determines the maximum diameter of $1\frac{3}{16}$ inch. The inner conductor is made of hardened beryllium copper and has the

Figure 6. Input impedance as a function of frequency for 2400 feet of Type 874-A2 50-ohm coaxial cable. Measurements were made on a v-h-f bridge.



necessary flexibility for receiving a banana plug as mentioned above. The outer conductor is of hard brass. The polystyrene supporting bead is molded with keyways, one of which can be seen in the right-hand bead in Figure 4, for positive alignment of the inner and outer conductors. All metal parts are finished in bright-alloy plate.

No special tools are needed to assemble these connectors.

For use with the connectors, a flexible, low-loss, coaxial cable of 50-ohm characteristic impedance is available. The TYPE 874-A2 Polyethylene Cable is designed for the high-frequency laboratory, where mechanical flexibility is a great convenience. Flexibility greater than that of other commonly used cable types of approximately the same impedance level has been achieved by using an inner conductor made up of many strands of small diameter wire and by choosing a medium over-all diameter for combining sturdiness and uniformity with flexibility. Low electrical leakage is achieved by means of two braided shields. Figure 6 shows the results of bridge measurements of input impedance of a 2400-foot length of TYPE 874-A2 Cable in the vicinity of 25 Mc and 100 Mc. With this length of cable the effects of reflections from the far end are negligible. The characteristic impedance of a flexible cable varies from point to point due to dimensional changes and produces the observed variations in input impedance with respect to frequency. The solid curve of Figure 7 shows the results of a similar cable measurement up to 4200 Mc, except that in this case a slotted line was used to measure the input standing-wave ratio of a very long cable connected to the line by means of a TYPE 874-C Cable Connector. The characteristic variation with frequency is present as at lower frequencies, and the lack of any definite trend with frequency shows that reflections caused by the connector and its tapered transition section are small compared to reflections caused by the cable itself. For comparison, the dashed and the dotted curves show the gradual trends caused by deliberate offsets of the outer taper with respect to the inner taper by distances of plus and minus 0.1 inch.

While it is planned eventually to use the TYPE 874 Connector on all new instruments requiring shielded connections, the TYPES 874-Q2 and 874-Q7 Adapters shown in Figure 8 are available for use with existing General Radio equipment.

The TYPE 874 Connector is presented as a useful element for simplifying laboratory work, with special advantages at ultra-high frequencies. A group of coaxial elements for making many kinds of ultra-high-frequency measurements, and using the TYPE 874 Connector throughout, is at present under development. At the present time the items listed below are available, and other coaxial elements will be announced later.

— W. R. THURSTON

Development of the TYPE 874 Coaxial Connector was carried out under the supervision of Eduard Karplus, who conceived the original idea of a universal connector and worked out the original design. In the development of the final design, the mechanical engineering was done by Harold M. Wilson, and the electrical engineering by William R. Thurston, author of the foregoing article.

— EDITOR.

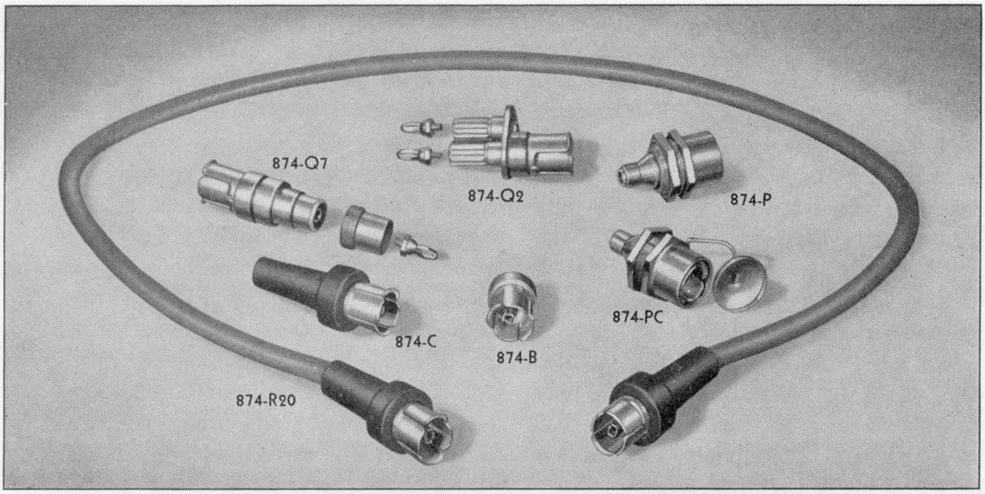


Figure 8. Coaxial connectors and parts now available.

SPECIFICATIONS

Type 874-B Basic Connector consists of inner and outer conductors, insulating bead, coupling nut, and retaining ring. This connector is designed for attachment to rigid, 50-ohm, air-dielectric, coaxial line made from ⅝″ O.D., 9⁄16″ I.D. tubing, and 0.244″ D rod. The inner conductor is to be screwed into an 8-32 tapping in the end of the rod, and the retaining ring for the coupling nut is to be snapped into a groove cut in the ⅝″ tubing.

Type 874-C Cable Connector contains the basic connector parts plus inner and outer transition pieces, a soft copper ferrule, and a rubber guard. The transition pieces are designed to attach to Type 874-A2 Polyethylene Cable and are tapered so as to maintain the 50-ohm characteristic impedance of the connector and cable throughout the change in diameters. The cable inner conductor is to be soldered to the inner transition piece, and the cable braid is attached to the outer transition piece by crimping the ferrule. The rubber guard provides a protective handle.

Type 874-P Panel Connector is similar to the cable connector, including transition pieces for Type 874-A2 Cable, except that a panel adapter, clamp ring, and nut are supplied in place of the rubber guard. The panel adapter is designed to clamp the connector in any desired orientation.

Type 874-PC Panel Connector with Cap is similar to the Type 874-P except that the panel adapter is equipped with a captive, hinged, spring cap that effectively shields the open connector when not in use.

Type 874-Q2 Adapter is designed for making the output of a coaxial system available at a pair of ¾-inch-spaced binding posts or plugs.

Type 874-Q7 Adapter is used to connect from a Type 874 Coaxial Connector to any Type 774 Coaxial Connector.

Type 874-A2 Polyethylene Cable consists of a No. 14 stranded inner conductor, separated from a double-braid tinned-copper shield by 0.244″ O.D. Polyethylene N1 insulation, and with an outer gray Plastex jacket 0.365″ O.D. The characteristic impedance is 50 ohms ±5%, and the nominal capacitance is 32 $\mu\mu$f per foot. The attenuation at 100 Mc is about 2.6 db per 100 feet, and at 1000 Mc about 10.5 db per 100 feet.

Type 874-R20 Patch Cord consists of three feet of Type 874-A2 Polyethylene Cable with a Type 874-C on each end.

Type		*Net Weight*	*Code Word*	*Price*
874-B	**Basic Connector**	⅔ oz.	COAXBRIDGE	**$1.50**
874-C	**Cable Connector**	1¼ oz.	COAXCABLER	**2.00**
874-P	**Panel Connector**	2¼ oz.	COAXPEGGER	**2.25**
874-PC	**Panel Connector with Cap**	2¼ oz.	COAXCAPPER	**2.75**
874-Q2	**Adapter**	2¼ oz.	COAXTIPPER	**3.00**
874-Q7	**Adapter**	2¼ oz.	COAXPASSER	**3.75**
874-A2	**Polyethylene Cable**	1½ oz./ft.	COAXCUTTER	**0.50/ft.***
874-R20	**Patch Cord**	7 oz.	COAXHATTER	**6.25**

*In lengths of 100 feet or more, $27.00/100 ft.

Type 874 Connectors are licensed under U. S. Patent No. 2,125,816; patent applied for.

VARIAC RATINGS

Important to All Present or Future Variac Users

Variac ratings are sometimes misunderstood or ignored, often with unfortunate results. Previously published material covering ratings may, perhaps, have lacked proper emphasis and clarity.

The several Variac ratings are, at first glance, confusing, yet they are all required to define exactly the limits of Variac performance. Each should be thoroughly comprehended for optimum results.

As with all continuously adjustable autotransformers, the current that may be drawn from a Variac through the brush is not constant (as in fixed transformers) but varies as a function of the Variac brush setting. Near zero and line voltages, winding heat is low and brush heat becomes the principal current limitation. At fifty and one-hundred-seventeen per cent of line voltage, winding heat is the dominant factor. Both factors operate at intermediate settings.

Figure 1 is a graphic illustration of these conditions. *The figure is typical, not specific;* 230-volt models differ from the curve shown. *Consult individual Variac instruction sheets for particulars.* The curves are:

1. Rated Current, solid line, determined by maximum winding heating. *Rated Current may be drawn at any setting.*

2. Maximum Allowable Current, dashed line, determined jointly by winding and brush heating.

3. Maximum Current, dotted line, determined by brush heat. *Maximum Current may be drawn at, or near (within approximately ten per cent of line voltage), line voltage and zero.*

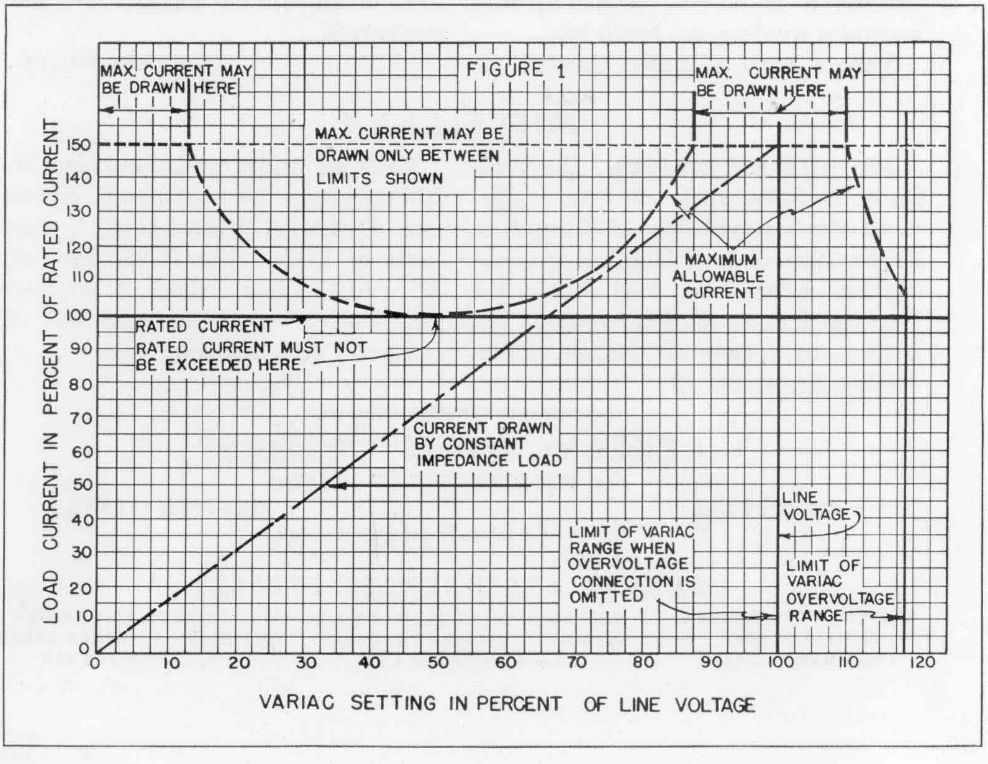


FIGURE 1

In any doubtful application, play safe; keep within Rated Current at all times!

The straight dot-dash line in Figure 1, which connects the point of intersection of Maximum Current and Line Voltage with the zero-zero point, represents the current versus voltage curve of a typical load. (Technically, the curve is for a "constant impedance" load, and most Variac loads approximate this curve.) Since the line lies entirely below the Maximum Allowable Current curve, this means that a load which draws Maximum Current at line voltage can be regulated between zero and line voltage (not higher) without overloading the Variac. Variac kva ratings are derived from this concept. We believe load ratings based on other than standard line voltages are meaningless as all readily procurable electrical devices are designed for operation on these same standard voltages. Ratings are on a volt-ampere, or kilo-volt-ampere (kva), basis because it is current (amperes) that generates winding and brush heat.

Variacs have adequate thermal capacity to withstand overloads up to five times Rated Current for short periods (five seconds), as in motor-starting. Larger currents rapidly generate brush temperatures exceeding the softening temperature of the copper wire winding, with consequent damage to the wire surface beneath the brush.

Certain loads draw inrush, or surge, currents greatly beyond Variac capacities. Typically, cold incandescent lamps momentarily may draw more than fifteen times their normal, hot, operating current. To avoid such serious surges, *always set the Variac brush to zero before switching load or mains.*

Variacs are commonly used as automatic or manual line-voltage regulators. If the line-voltage fluctuation is appreciable (greater than ten per cent), load current should be limited to Rated Current to avoid overloading.

We repeat: *When in doubt, play safe; keep within Rated Current at all times!* Best of all, *use an ammeter and know you're safe!*

—GILBERT SMILEY

MISCELLANY

ARTHUR E. THIESSEN, Vice-President for Sales, spoke on "An Instrument Man's Impressions of Europe" at the luncheon of the Scientific Apparatus Makers' Association, held during the National Instrument Conference of the Instrument Society of America, Philadelphia, September 10.

RECENT VISITORS to our plant and laboratories — Mr. Krishan L. Khandpur, Education Department, Government of India, New Delhi; Professor Augusto Condom, School of Engineering, University of Havana; and M. Roberto Corbo of Rome.

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THE General Radio EXPERIMENTER

VOLUME XXIII No. 6 NOVEMBER, 1948

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A COMPACT RADIO-FREQUENCY CAPACITANCE MEASURING INSTRUMENT

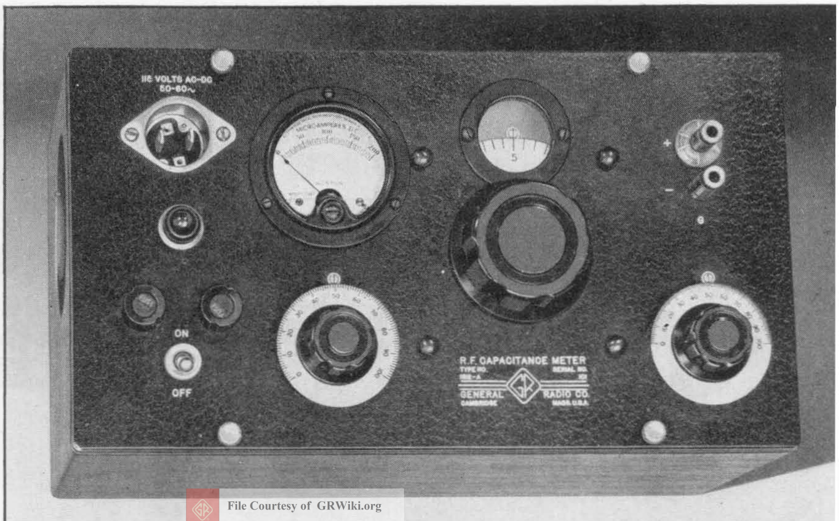
Also

IN THIS ISSUE

Page

●**THE TYPE 1612-A** Radio-Frequency Capacitance Meter is a device for conveniently measuring capacitance of magnitudes usually encountered in radio-frequency applications, where the accuracy required is not extreme, and where the dissipation factor of the capacitance to be measured is reasonably low at one megacycle. This instrument is complete in itself and requires no auxiliary equipment except a source of 115-volt power. The operation is extremely simple, the capacitance value being read from a calibrated dial after it has been set to give the proper visual indication on a panel meter. In addition to indicating capacitance values, the capacitance

Figure 1. Panel view of the R-F Capacitance Meter.



meter provides a relative indication of losses in the capacitance measured. An interesting application of this feature is in the intercomparing of losses in dielectric samples at the operating frequency of one megacycle.

The capacitance range is covered in two parts, zero to 80 micromicrofarads and zero to 1200 micromicrofarads. The ranges are calibrated in separate 180-degree sectors of the dial and range switching is accomplished automatically as the dial is rotated. Thus any ambiguity is avoided in capacitance indications since only one scale at a time appears under the dial indicator. The capacitance scale distribution is nearly logarithmic so that accuracy and readability are maintained down to low capacitance values.

The functional diagram illustrates the principles of operation. The instrument consists simply of a one megacycle oscillator whose output is loosely coupled to a resonant detector circuit. The resonance indicator includes a crystal rectifier and a d-c microammeter loosely coupled to a resonant circuit by means of a small pickup coil.

Measurement is made by a substitution method in which the capacitance of a calibrated condenser is reduced to reestablish resonance after an unknown capacitance is placed across the terminals. Resonance is indicated by maximum deflection of the microammeter.

In the instrument, the calibrated capacitor is a straight-line-capacitance variable capacitor, carrying the capacitance dial, in series with a fixed padding capacitor. The relative magnitudes are so chosen that the capacitance change with dial rotation of the combination is small near the maximum capacitance end of the variable capacitor (minimum capacitance on dial) and much larger at the minimum capacitance end (maximum capacitance on dial). Thus, as the dial is rotated, the effective capacitance removed varies slowly at first, increasing in rate with angular rotation. The resulting capacitance scale approaches a logarithmic variation with angle for the low capacitance range.

An extension of the series padder principle is used to obtain the high capacitance range. In this case an additional padding condenser is switched in series with the unknown terminals. Thus incremental changes of capacitance in the unknown present decreasing capacitance changes to the resonant detector circuit as the value of the unknown capacitance increases. The net effect is an increased angular spreading of the low capacitance end of this range with respect to the high end.

In order to permit the angular rotation of the dial required to operate the range switch, the dial is provided with separate zeros for the two ranges. A panel trimmer permits standardizing the circuit at zero capacitance and can be used to balance out the capacitance of leads that might be used to connect the unknown capacitance. Approximately 5

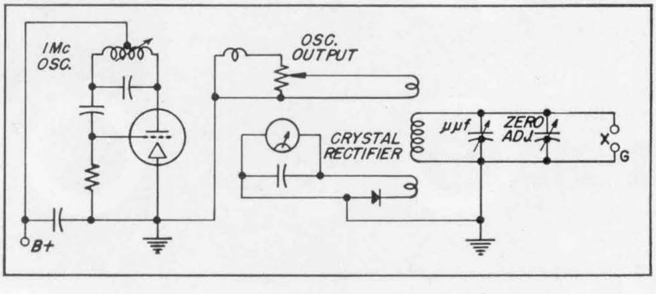


Figure 2. Elementary schematic circuit diagram of the R-F Capacitance Meter.

micromicrofarads of external capacitance can be balanced out on the low range and 120 micromicrofarads on the high range.

Once the circuit has been standardized at the zero point, any capacitance can be easily measured by connecting it, turning the dial to point of maximum indication on the microammeter, and reading the capacitance value from the calibrated dial. Re-standardization of the zero for successive measurements of capacitance on the same range is not required if the zero of the panel trimmer is not disturbed between measurements. If the unknown capacitances are connected directly to the terminals of the instrument, the zero settings of the trimmer are coincident for the two ranges and, therefore, no attention to zero standardization is required in going from one range to the other.

For production testing of a number of capacitors of the same type, a test jig can be devised for facilitating connections.

On the low capacitance range, losses in the unknown capacitance show up on the indicating meter as a lower indication at resonance with the unknown connected than at the zero-standardization resonance. The difference in resonant peak indications is a function of the shunt conductance in the unknown and consequently is a relative indication of the losses at a given value of capacitance. An oscillator output control permits setting the meter to full scale during the zero-standardization process in order to provide a reference for loss comparisons. This indication of losses is intended only for rough intercomparisons and not for precise measurements.

If the terminals of the instrument are provided with two electrodes between which can be placed dielectric samples, intercomparison of the losses can be made. The oscillator output control should be set so that the microammeter gives a full-scale indication at resonance before a sample is placed between the electrodes. A setting of the capacitance dial should be found on the low range where resonance can be obtained with the various samples in place by merely changing the position of one of the electrodes on the sample. The amount the microammeter indication differs from the full-scale value at resonance, with a sample in place, is an index of the losses in that particular sample. These intercomparisons should be made at the same capacitance setting because the indication depends upon shunt conductance rather than upon dissipation factor.

The oscillator used in the 1612-A uses a pentode section of a type 117N7GT tube, connected as a triode in a simple Hartley circuit. The oscillating frequency is preset to one megacycle by means of an adjustable powdered iron core in the oscillator coil. The type 117N7GT also contains a rectifier section which is used to supply power through an *R-C* filter to the plate of the oscillator section from a 115-volt a-c supply. The instrument can also be operated from a 115-volt d-c supply provided the power plug is inserted to give the proper polarity. This is simply determined by reversing the power plug if the instrument fails to function. The heater in the 117N7GT is supplied directly from the 115-volt supply.

The oscillator circuit is not connected by any conducting path to the metal panel or chassis so that the use of a transformerless power supply places absolutely no limitation on the use of the instrument.

— W. F. Byers

SPECIFICATIONS

Capacitance Range: 0 to 1200 $\mu\mu$f in two ranges —0 to 80 $\mu\mu$f and 0 to 1200 $\mu\mu$f. Ranges are switched automatically as capacitance dial is rotated.

Capacitance Accuracy: Low Range: From 0 to 50 $\mu\mu$f, $\pm(3\% + 0.3\ \mu\mu\text{f})$. Between 50 and 80 $\mu\mu$f, ±6%. High Range: From 0 to 1200 $\mu\mu$f, $\pm(3\% + 5\ \mu\mu\text{f})$.

Dielectric Losses: Relative meter indications with different dielectric samples give a comparative measure of dielectric loss.

Oscillator Frequency: 1 megacycle ±1% adjusted at factory. Frequency can be readjusted if necessary by means of a movable dust core.

Resonance Indicator: A 1N34 crystal rectifier is used with a microammeter to indicate resonance.

Tube: A 117N7-GT tube is used in the oscillator circuit, and is supplied.

Power Supply: 115 volts, 50 to 60 cycles, ac or dc.

Power Input: 12 watts at 115 volts, ac; 11 watts at 115 volts, dc.

Dimensions: (Length) 12 x (height) 6⅝ x (depth) 7½ inches, over-all.

Net Weight: 11 pounds, 10 ounces.

Type		*Code Word*	*Price*
1612-A	**R-F Capacitance Meter....................**	AFTER	**$155.00**

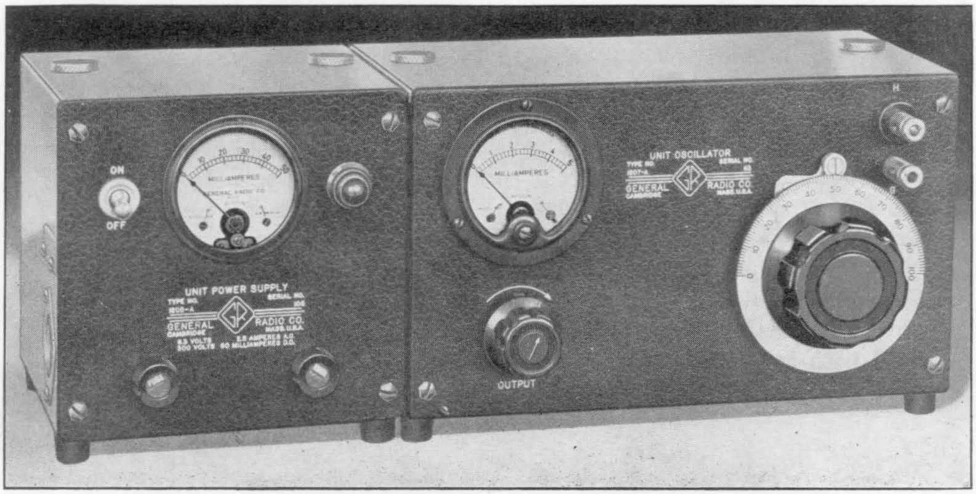
INEXPENSIVE, BASIC INSTRUMENTS FOR THE LABORATORY

Believing that there exists a need for simple, wide-range, basic laboratory tools at low cost, the General Radio Company has developed a series of unit instruments, designed to be versatile in application and lower in price than any heretofore available.

Three units are now available; others will be added in the future. The three currently available instruments, which can collectively be purchased for a total of $316, include (1) a 3-watt amplifier covering the frequency range from 20 c to 200 kc, whose gain is adjustable from 0 to 45 db; (2) an oscillator operating over the range from 400 c to 80 Mc, and delivering ½ watt up to 5 Mc; and (3) a compact a-c power supply that plugs into either the oscillator or the amplifier. Each unit can be purchased separately, if desired.

Such instruments have many applications in the laboratory. They supplement more expensive equipment in the large

Figure 1. View of the Type 1205-A Power Supply and the Type 1207-A Oscillator plugged together.



organization and fill a basic need in the small laboratory whose budget is limited. A particularly important field of application is in the college laboratory, where the cost of providing several oscillators and amplifiers for student use is often prohibitive. The low cost of each of these units permits several to be purchased at the same price that would be paid for a single, more elaborate instrument.

A further advantage for the undergraduate laboratory lies in the fact that the wiring in these instruments is open and accessible, permitting the student to modify the circuit in many ways and to study the effect of such modification on circuit performance. Laboratory practice can thus be closely correlated with theory and the student's understanding of the circuit operation greatly enhanced.

DESCRIPTION OF THE UNITS

1. TYPE 1206-A UNIT AMPLIFIER

The unit amplifier measures 10¼ x 5⅞ x 6½ inches, over-all. Plug terminals on the end, as shown in Figure 2, directly engage a jack in the power supply. Hence it is small, compact, and easily handled by laboratory personnel. Input and output terminals are on the front panel. Its performance, as detailed in the specifications below, is completely adequate for most laboratory uses, and considerably better than might be expected from its size and price.

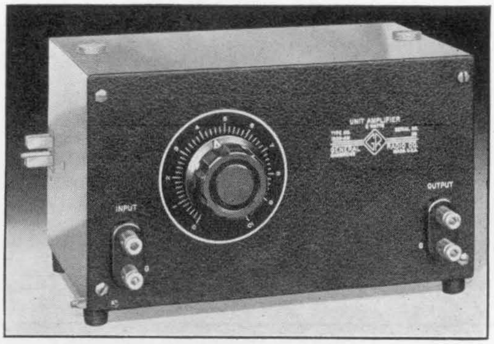


Figure 2. Panel view of the Type 1206-A Amplifier, showing, at the left, the plug that engages a socket in the power supply.

The amplifier covers both audio and supersonic frequencies. Its 45-db gain is sufficient for use in the detector circuits of impedance bridges, and its maximum output of 3 watts is adequate for driving low-power laboratory devices.

The circuit uses two triode direct-coupled voltage-amplifier stages and a resistance-capacitance-coupled output stage. Cathode degeneration is employed on the input stage and additional degeneration is provided between the last two stages. As a result, the gain stability is excellent, distortion is low, and the phase-shift characteristic is good.

SPECIFICATIONS

Voltage Gain: Continuously adjustable from 0 to 45 decibels.

Load Impedance: 7500 ohms optimum. Blocking capacitor is 1 μf.

Maximum Output: 3 watts (150 volts) into 7500 ohms can be obtained with less than 5% distortion.

Input Impedance: The input resistance is 200,000 ohms. Blocking capacitor is 0.05 μf.

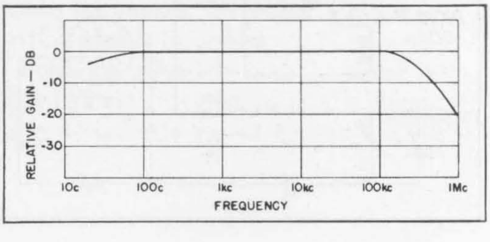
Frequency Response: Essentially constant from 100 cycles to 100 kc. Response drops 6 db per octave above 200 kc. Response at 20 c is down 6 db.

Tubes: One 6SN7-GT and one 6V6-GT are supplied.

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

Distortion: The distortion when delivering 1 watt into a load of 7500 ohms is less than 2% at frequencies above 100 cycles. At lower frequencies the distortion increases, but is less than 3% at

Figure 3. Frequency characteristic of the Type 1206-A Amplifier.



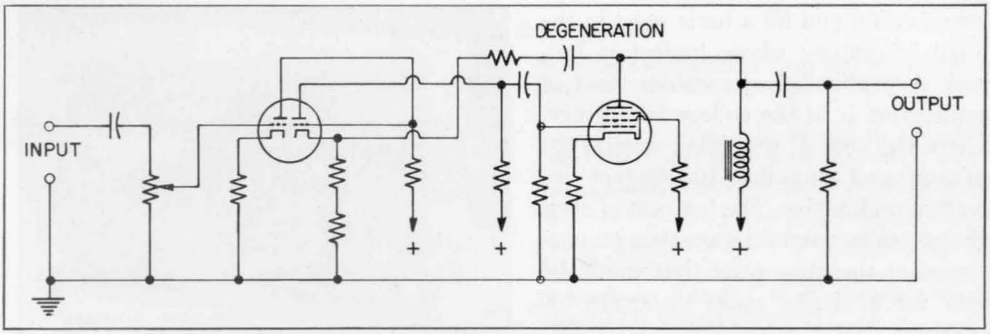


Figure 4. Elementary schematic circuit diagram of the Type 1206-A Amplifier.

50 cycles. At an output of 3 watts, total distortion is under 5% above 100 cycles.

A-C Hum: The 60-cycle hum level in the output is about 125 millivolts.

Power Supply: The Type 1205-A Unit Power Supply plugs directly into the amplifier.

Dimensions: (Width) 10¼ x (height) 5⅞ x (depth) 6½ inches over-all.

Net Weight: 8¾ pounds.

Type		*Code Word*	*Price*
1206-A	**Unit Amplifier**	ARBOR	**$65.00**

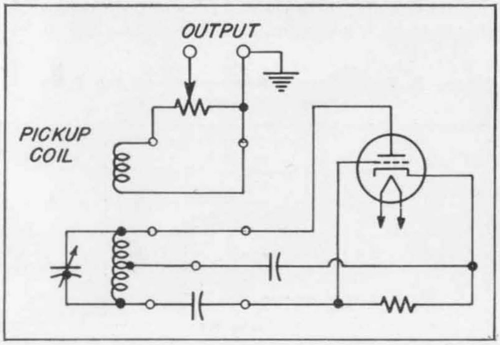
Licensed under Patent of the American Telephone and Telegraph Co.

2. TYPE 1207-A UNIT OSCILLATOR

Wide frequency range and adequate output were the design objectives for this oscillator. Its ability to produce a test signal at frequencies ranging from 400 cycles to 80 megacycles makes it a most useful laboratory instrument. It fills the need for a simple, wide-range signal source for measurement and testing, without expensive refinements, but quickly adaptable to a variety of uses.

The shielding is adequate for bridge measurements at audio frequencies. At radio frequencies, for bridge and resonant-circuit measurements, additional external shielding must be provided by the user.

Figure 5. Elementary schematic circuit diagram of the Type 1207-A Oscillator.



A conventional Hartley circuit is used in this oscillator. Seven plug-in coils are available, which collectively cover a continuous frequency range from 70 kc to 80 Mc. Plug-in coil forms are available and can be used to meet special requirements. Three plug-in tuning units are available to operate at fixed frequencies of 400, 1000, and 20,000 cycles, respectively. Jacks across the tuned circuit permit extending the range of any coil by plugging in external capacitors or inductors, so that the entire range from 100 cycles to 100 megacycles can be covered.

The output circuit is coupled inductively to the tuned circuit, and the output is controlled by a voltage divider. The oscillator plugs directly into the Type 1205-A Unit Power Supply.

In addition to the various plug-in coils and tuning units listed below, note that a blank coil form with plugs and case is furnished and that others are available, so that a coil for any particular frequency or range of frequencies can be wound by the user.

SPECIFICATIONS

Frequency Range: Seven plug-in coils are available to cover the range from 70 kc to 80 Mc, and a plug-in coil form, also available, may be wound to meet special requirements. Three plug-in tuning units are listed to provide within ±2% fixed audio frequencies at 400, 1000, and 20,000 cycles. Coils, coil forms, and plug-in units must be ordered separately (see price list).

Frequency Control: With the seven tunable coils, continuous frequency adjustment is provided by a variable air capacitor, having a uniform scale from 0 to 100.

The audio frequency tuning units are inductor-capacitor combinations, and the internal variable air capacitor has little effect on the frequency. The frequency can be changed, however, by connecting a suitable capacitor or inductor to the jacks provided.

Frequency Stability: The frequency stability is adequate for most laboratory applications, except those involving highly selective tuned circuits. Variations of load impedance cause some shift in frequency.

Output Impedance: Approximately 75 ohms at full output for the coils (70 kc to 80 Mc) and 750 ohms for the tuning units (400, 1000, and 20,000 cycles). A 2000-ohm voltage divider at the output terminals provides an output control.

Output Power: At least 0.5 watt into a matched load up to 5 Mc, and 10 milliwatts at 80 Mc.

Modulation: Jacks are provided for connecting a modulating audio source in series with oscillator plate supply. The oscillator can be amplitude modulated to 50% from 0.5 to 15 Mc. The maximum modulating frequency is 10 kc over this carrier range. The modulating audio oscillator must be capable of delivering about 115 volts to yield 50% modulation.

Terminals: Jack-top binding posts with standard ¾-inch spacing are provided for the output connection.

Power Supply: The Type 1205-A Unit Power Supply plugs directly into the oscillator.

Tube: One 6C4 is used and is supplied.

Accessories Supplied: One plug-in coil form with case; multipoint connector; one shorting bar, 274-SB.

Accessories Available: Tuning units, coils, and coil storage rack are listed in price list below.

Dimensions: 1207-A: (Width) 10 x (height) 5⅞ x (depth) 6⅝ inches over-all.

P1, P2, and P3: 4¼ x 2½ x 3 inches each over-all.

P4, P5, P6, P7, P8, P9, and P10: 2⅜ inches diameter x 3½ inches over-all.

Net Weight: Coils: 5 oz. each; Tuning Units: (P1, 1¾ lbs.) (P2, 1¼ lbs.) (P3, 1½ lbs.); 1207-A: 5¼ lbs.

Type		*Code Word*	*Price*
1207-A	**Unit Oscillator**	ARGON	**$73.00**
1207-P1	**Tuning Unit, 400 c.**	ARGONSAWAY	**19.50**
1207-P2	**Tuning Unit, 1 kc.**	ARGONSAFIT	**17.50**
1207-P3	**Tuning Unit, 20 kc.**	ARGONSAPIS	**16.50**
1207-P4	**Coil, 70 kc to 180 kc**	ARGONSALOE	**9.00**
1207-P5	**Coil, 180 kc to 500 kc**	ARGONSABLE	**9.00**
1207-P6	**Coil, 0.5 Mc to 1.5 Mc**	ARGONSACRE	**9.00**
1207-P7	**Coil, 1.5 Mc to 5 Mc**	ARGONSANTY	**9.00**
1207-P8	**Coil, 5 Mc to 15 Mc**	ARGONSARTY	**9.00**
1207-P9	**Coil, 15 Mc to 50 Mc**	ARGONSAQUA	**9.00**
1207-P10	**Coil, 50 Mc to 80 Mc**	ARGONSAYAH	**9.00**
1207-P11	**Extra coil form with case and plug-in base.**	ARGONSAXLE	**6.50**

3. TYPE 1205-A UNIT POWER SUPPLY

This power pack is designed primarily for use with the Type 1206-A Unit Amplifier and the Type 1207-A Unit Oscillator. Connections to the oscillator or amplifier are made through multipoint connectors mounted in the ends of the instruments. When so assembled, the combination of units is very compact, occupying a minimum of bench space. The power supply can also be used separately as a general-purpose source of heater and plate power for other electronic equipment, for which purpose a mating plug is furnished.

Fuses are accessible from the panel. A pilot light is provided, as is a milliammeter to indicate output current.

SPECIFICATIONS

Output Voltages: 6.3 volts, ac, at 2.5 amperes, maximum. Approximately 300 volts, dc, at 50 milliamperes maximum. No-load voltage is about 390 volts. No regulation is provided.

Approximate Hum Level: 0.8 volt at 300 volts and 50 milliamperes, 60 cycles.

Input Power Supply: 115 (or 230) volts, 50 to 60 cycles.

Input Power: Approximately 12 watts, no load; approximately 50 watts, full load.

Rectifier Tube: One 6X5-GT/G which is supplied.

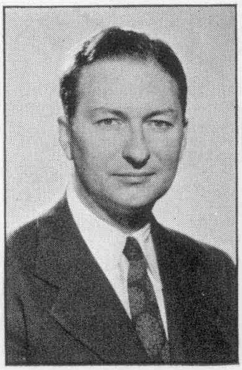
Output Terminals: A standard multipoint connector is arranged for plugging directly into either the Type 1206-A Unit Amplifier or the Type 1207-A Unit Oscillator. A mating plug is provided for use with other equipment, and a line connector cord is supplied.

Dimensions: (Width) 5⅝ x (height) 5⅞ x (depth) 5⅞ inches over-all.

Net Weight: 6¾ lbs.

Type		*Code Word*	*Price*
1205-A	**Unit Power Supply**	APPLY	**$55.00**

MISCELLANY



Frank D. Lewis

Donald B. Sinclair

HONORS—Awarded, to Donald B. Sinclair, Assistant Chief Engineer, and to Frank D. Lewis, Engineer, the President's Certificate of Merit for outstanding services. Dr. Sinclair's certificate was awarded for his work in the Counter-measures Division and the Guided Missiles Division of the NDRC, Mr. Lewis's for his work in the field of international liaison with the NDRC, OSRD, and as Expert Consultant to the Secretary of War.

PAPERS—"Problems Relating to Research Personnel in Industrial Laboratories," by H. B. Richmond, Chairman of the Board, at the Conference on Research Administration, at Pennsylvania State College, September 14.

—"Catalog Design and Distribution," by A. E. Thiessen, Vice-President for Sales, at the Mid-Year Meeting of the Scientific Apparatus Makers' Association, French Lick, Indiana, October 12.

RECENT VISITORS to our plant and laboratories—Dr. Pierre Marie of Societe Sadir-Carpentier, Paris; Mr. Jacques Franeau, Faculte Polytechnique, Mons, Belgium; Mr. L. Webster, Chief Radio Engineer, and Mr. A. Birrell, Assistant Radio Engineer, G.P.O., Victoria, South Africa; and Mr. R. S. Medlock, of George Kent, Ltd., Sutton, Bedfordshire, England.

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

THE INTERPOLATING FREQUENCY STANDARD

Also

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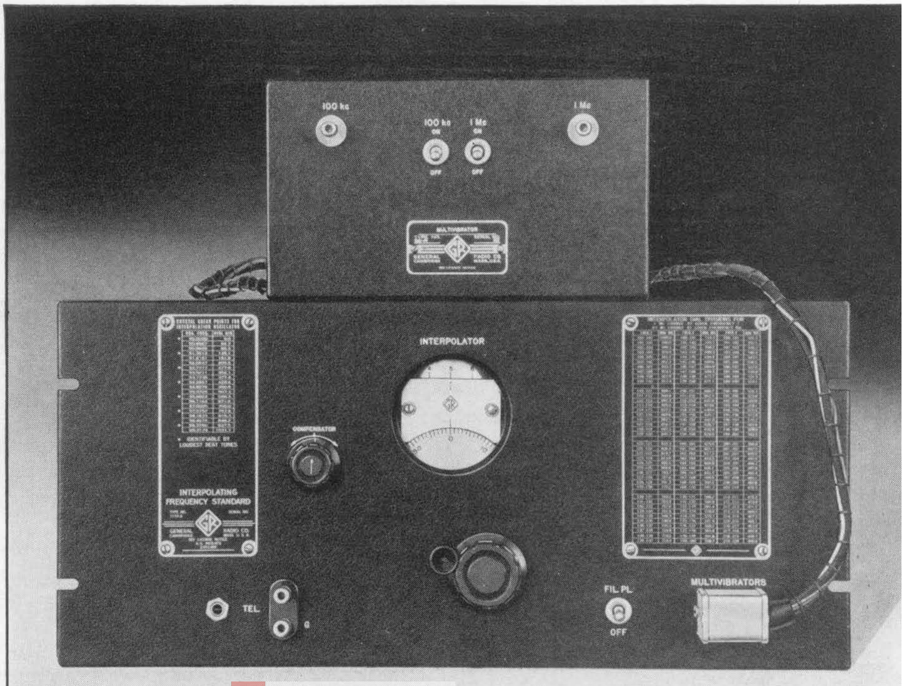
● **THE INCREASING IMPORTANCE** of the higher frequencies in radio communication has emphasized the need for more accurate means of frequency measurement between 100 megacycles (the nominal upper limit obtainable with commercial frequency measuring equipment based on a primary standard) and a few thousand megacycles. This range is at present covered by heterodyne frequency meters with an accuracy in the range between 0.01 and 0.1 per cent.

Since the heterodyne frequency meter provides a means of detecting

Figure 1. Panel view of the Type 1110-A Interpolating Frequency Standard with Type 1110-P1 Multivibrator.



the signal and identifying its frequency fairly closely, an additional instrument that provides a reference standard and a means of precise interpolation, and which is designed to work with the frequency meter, is a logical, as well as the least expensive, means of achieving an increased accuracy of measurement.

The Type 1110-A Interpolating Frequency Standard meets the requirements for this type of instrument. It is designed to be used primarily with the Type 720-A Heterodyne Frequency Meter and the Type 620-A Heterodyne Frequency Meter and Calibrator. It can also be used for frequency measurement with high-frequency receivers if their frequency calibrations are sufficiently good to identify frequencies separated by as little as one per cent.

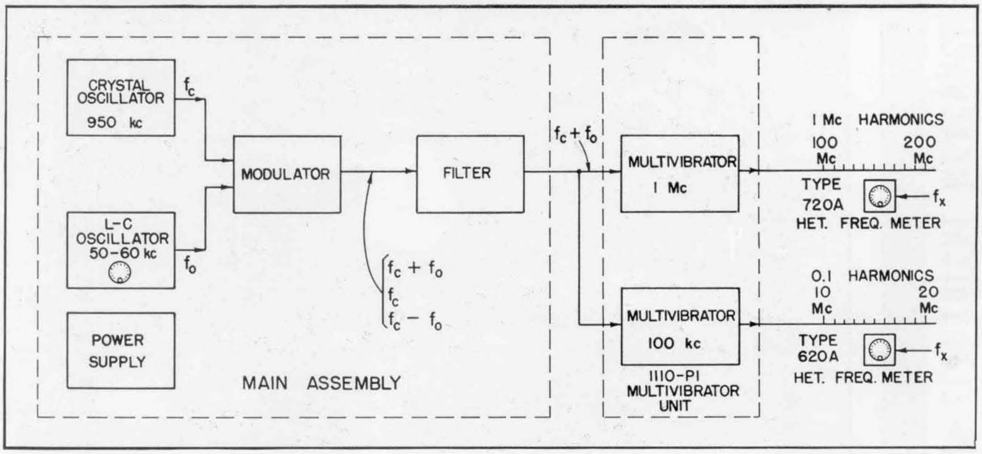
DESCRIPTION

As shown in the functional diagram of Figure 2, the Interpolating Frequency Standard contains a 950 kc quartz crystal oscillator, a 50 to 60 kc bridge-controlled L-C Oscillator, with a mixer, filters, and output amplifiers for obtaining the sum of the two oscillator frequencies as the output frequency. The control of the 50 to 60 kc oscillator is a worm-drive with a scale of 1000 divisions. The output frequency is variable from 1.000 to 1.010 Mc by varying the 50 to 60 kc oscillator over its full range. Means are provided for checking the 50 to 60 kc oscillator frequency against the crystal oscillator frequency at a number of points over the dial. Any error in the 50 to 60 kc oscillator calibration can be corrected by a control provided for the purpose.

The Type 1110-P1 Multivibrator Unit contains two multivibrators with output amplifiers. One multivibrator operates at 1 Mc, for use with the Type 720-A Heterodyne Frequency Meter, while the other operates at 0.1 Mc for use with the Type 620-A Heterodyne Frequency Meter. With either multivibrator, the output frequency of the Interpolating Frequency Standard is used as the control frequency. The multivibrators remain in control over the full range of variation of the control frequency.

Since the range of control frequency is just one per cent, it follows that, as the control frequency is varied over its range, the frequencies of all the harmonics of the multivibrators will likewise be varied over a range of one per cent. For the 100th harmonic this means that the total range of adjustment is from 100 to

Figure 2. Block diagram showing the functional arrangement of the interpolating frequency standard.



101 Mc for the 1 Mc Multivibrator and from 10.0 to 10.1 Mc for the 0.1 Mc Multivibrator. For harmonics higher than the 100th, the one per cent range is greater than 1 Mc or 0.1 Mc, so that *continuous coverage* is obtained.

METHOD OF MEASUREMENT

In making frequency measurements it is usually desired to obtain the result on a cycles basis rather than a percentage basis. To do this, it is convenient to think of the frequency of any multivibrator harmonic as its unaltered value *plus* an increment in frequency. The unaltered value is the true standard frequency value obtained when the dial of the Interpolating Frequency Standard is set at zero or, more precisely, when the 50 to 60 kc oscillator is set at 50 kc in terms of the crystal oscillator frequency. The increment in frequency is the fraction of a megacycle (or of 0.1 Mc) required to obtain the actual frequency. This increment can be calculated from the number of dial divisions required to advance any harmonic by just 1 Mc (or 0.1 Mc). These values are prepared in the form of a table mounted on the panel of the instrument. The fraction of 1 Mc (or 0.1 Mc) in any given case is then the number of dial divisions divided by the number of dial divisions for 1 Mc (or 0.1 Mc). This fraction is then *added* to the unaltered frequency of the used harmonic to obtain the final result.

NUMERICAL EXAMPLE

An illustrative example may be easier to follow. Suppose the reading of the Type 720-B Heterodyne Frequency Meter at zero beat with the frequency being measured is 162.3 Mc. The dial of the Interpolating Frequency Standard is advanced, say, 249.0 divisions to obtain zero beat with the frequency meter. The used harmonic is 162 (the next integral value below 162.3); entering the table for the range 162-163 Mc, it is found that the dial of the Interpolating Frequency Standard must be advanced by 617.3 divisions to cover the 1 Mc range from 162 to 163 Mc. Since the dial was actually advanced 249.0 divisions, the fraction 249.0/617.3 of a megacycle was actually covered, amounting to 0.404 Mc. The final result is then 162 + 0.404 = 162. 404 Mc.

It is evident from the above that if the dial of the Interpolating Frequency Standard is set at zero (or, more precisely, when the 50 to 60 kc oscillator is set to 50 kc in terms of the crystal oscillator), the output harmonics are all multiples of 1 Mc (or 0.1 Mc) and can be used as standard frequencies for checking calibrations of receivers, oscillators, heterodyne frequency meters, etc.

It also should be evident that, in making frequency measurements, the dial reading of the Interpolating Frequency Standard would normally never exceed the value given in the table. For, if the frequency increment exceeds 1 Mc (or 0.1 Mc), it would be natural to use the next higher harmonic and obtain a small frequency increment.

HARMONICS OF THE FREQUENCY METER

When the equipment is used to measure frequencies lying in the fundamental frequency range of the heterodyne frequency meter, all of the steps required have been covered above. Harmonics of the frequency meter can be used, however, for extending the range of measurement to higher frequencies. In such cases, the number of the *heterodyne frequency meter* harmonic must be determined and the result obtained in measuring the frequency meter fundamental (as detailed above) must be multiplied by this harmonic number to obtain the value of the unknown frequency. In the above numerical example, if the fifth harmonic of the heterodyne frequency meter was used to beat with the unknown frequency, the final result would be 5 × 162.404 = 812.020 Mc. Identification of the harmonic order can easily be made by receiver calibration, wavemeter, or other approximate means.

ACCURACY

The interpolator dial has 1000 divisions, corresponding to 0.001 per cent, or 10 parts per million, per division. Setting the dial to one-fifth division, therefore, gives a precision of two parts per million. If the frequency error of the dial is carefully corrected by means of the trimmer dial on the panel, the over-all accuracy of measurement is limited mainly by the accuracy of the crystal oscillator. Since the frequency of the crystal oscillator can be checked and adjusted in terms of standard-frequency radio transmissions to well within one part in a million, measurements to two or three parts per million are possible after these checks and adjustments are made. If the oscillator is used without trimming, the over-all accuracy is about ± 25 parts per million (0.0025 per cent).

MOUNTING

The interpolating standard is mounted on a standard 19-inch relay rack panel, and the multivibrator unit, TYPE 1110-P1, is mounted in a smaller cabinet attached to the interpolator by a plug-in cable. This arrangement facilitates coupling the multivibrator output to the heterodyne frequency meter or receiver, so that maximum response to weak higher harmonics can be obtained. This flexibility of mounting and the general simplicity of operation make the Interpolating Frequency Standard a convenient and easy instrument to use.

— J. K. CLAPP

SPECIFICATIONS

Frequency Range: The output frequency range of the 1110-A Interpolating Frequency Standard is from 1000 to 1010 kc. The output frequencies of the 1110-P1 Multivibrator Unit are 1.0- and 0.1-Mc fundamentals with harmonics up to 200 or more.

Calibration: The variable frequency oscillator dial has 1000 divisions corresponding to 0.001 per cent or 10 parts per million per division.

A list of check settings is provided on the panel. This check can be made at any time by simply plugging a set of headphones into the jack or binding posts provided on the panel. A trimmer control on the panel provides for adjusting the oscillator to agreement with the crystal.

To facilitate conversion of the dial readings from their basic percentage or parts per million values of frequency increment to fractions of a megacycle or of 0.1 Mc (100 kc), a table listing the number of dial divisions for frequency increments of 1.0 Mc and 0.1 Mc at each harmonic from 100 to 220 is given on the panel. A simple slide-rule ratio then gives the desired frequency increment.

Crystal Oscillator: The crystal oscillator is adjusted to within one part in a million of correct frequency at room temperature. It should be reliable to within ±10 parts per million at ordinary room temperatures. The crystal frequency can be checked and adjusted in terms of standard frequency transmissions from WWV using an external receiver, maintaining the variable oscillator at exactly 50 kc in terms of the crystal.

Accuracy of Measurement: The over-all accuracy of measurement is ±25 parts per million using the oscillator dial directly. If the oscillator is carefully trimmed in terms of the crystal, the over-all accuracy is limited principally by the error of the crystal.

Vacuum Tubes: The following tubes are supplied:

2 — 6AC7	3 — 6J5GT/G
4 — 6SN7-GT	1 — 5R4GY
1 — 6SJ7	1 — 9001
1 — 6SA7	1 — 2LAP-430 (Bridge Circuit Lamp)

Power Supply: Either 105-125 or 210-250 volts, 50-60 cycles.

Power Input: 85 watts from 115-volt, 60-cycle line.

Mounting: Type 1110-A Relay Rack; Type 1110-P1 (attached to 1110-A by cable) small metal cabinet.

Accessories Supplied: Line connector cord and Type 1110-P1 Multivibrator Unit with connecting cable.

Accessories Required: Head telephones.

Dimensions: 1110-A Panel (length) 19 x (height) 8¾; behind panel, (length) 17¼ x (height) 8⅜ x (depth) 14 inches. 1110-P1 (length) 9¼ x (height) 5¼ x (depth) 5¼ inches.

Net Weights: Type 1110-A assembly, 40 pounds; Type 1110-P1 Multivibrator Unit 7½ pounds.

Type		*Code Word*	*Price*
1110-A	**Interpolating Frequency Standard* .....**	RAVEN	**$725.00**

*U.S. Patent 2,012,497.

Licensed under patents at the American Telephone and Telegraph Company and under patents and patent applications of G. W. Pierce.

Note: The method of frequency measurement described in the foregoing article was also discussed in a paper by J. K. Clapp entitled "Frequency Measurement by Sliding Harmonics," which appeared in the October, 1948, issue of the *Proceedings of the Institute of Radio Engineers.* — Editor.

A LOW-FREQUENCY MULTIPLIER FOR THE VACUUM-TUBE VOLTMETER

The Type 1800-P3 Multiplier extends the range of the Type 1800-A Vacuum-Tube Voltmeter to 1500 volts for both d-c and low-frequency a-c measurements. This multiplier plugs into the binding posts on the panel of the voltmeter. For d-c use the multiplier consists of a fixed resistance voltage divider giving a 10:1 reduction in the voltage applied to the voltmeter, while for a-c measurements a capacitance-resistance voltage divider is used.

View of the Type 1800-P2 Low-Frequency Multiplier.

The multiplier is not intended for use at frequencies above 5 megacycles. For higher frequencies the Type 1800-P2 Multiplier[1] is recommended.

SPECIFICATIONS

Multiplier Ratio: DC, 10:1 ±1 per cent for all Type 1800-A Voltmeters.

AC, Adjustable to 10:1 ±0.5 per cent. A multiplier ordered with a Type 1800-A Voltmeter is adjusted at 1500 cycles and 15,000 cycles to that voltmeter at our factory. When the multiplier is ordered for use with a Type 1800-A Voltmeter already in the hands of the user, the necessary adjustments can be made by the user. If these adjustments are not made, an additional error of ±5 per cent is possible.

Input Impedance: DC, 10 megohms.

AC, 10 megohms parallel resistance; 10 $\mu\mu$f parallel capacitance.

Frequency Error: 20 — 20,000 c ±2 per cent.
20 kc — 5 Mc ±4 per cent.

Waveform Error: As with the Type 1800-A Voltmeter, the deviation from r-m-s reading can be as large as the percentage of harmonics present in the signal, but the error will not necessarily be the same as that of the voltmeter used without the multiplier because of the different impedance presented to the voltmeter by the multiplier.

Dimensions: 5 x 2 x 2 inches.

Net Weight: 8 ounces.

Type		*Code Word*	*Price*
1800-P3	**Low-Frequency Multiplier ....**	ABEAM	**$25.00**

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[1] "A Voltage Multiplier for the Vacuum-Tube Voltmeter," *General Radio Experimenter*, May, 1948.

MORE VARIAC WATTS FOR YOUR DOLLAR

With the new V-5, V-10, and V-20 series now in stock, it seems advisable to point out to our friends and customers that the judicious selections of the proper Variac or Variac combination may often result in substantial savings for a given power requirement.

Reference to Table I will quickly reveal that certain load requirements may be met by several Variac assemblies, of which one is an outstanding bargain when considered on the basis of KVA rating per dollar of list price. In this respect, parallel combinations of the V-20 series have considerable advantage over the older 50 series Variacs.

Note particularly the following comparisons: For 115-volt service in the 10 KVA load range, a V-20G3 parallel gang delivers 10.35 KVA for $204 versus a 50-AG2 parallel gang delivering 10 KVA for $320. Here the V-20 series delivers 62 per cent more KVA per dollar. In 230-volt service a V-20G2 series gang delivers 6.9 KVA for $126 versus a 50-B, which delivers 7.0 KVA for $140, a 9.6 per cent gain for the V-20 over the 50-B. Other examples will be apparent on ex-

TABLE I

SINGLE-PHASE VARIAC UNITS AND COMBINATIONS

Listed by Common Line Volts and Increasing KVA

115-Volt Service						
Model	*Connection*	*KVA*	*Cost of Variacs*	*Cost of Chokes*	*Total $*	*KVA/$*
200-B	———	.170	12.50	———	12.50	.0136
V-5	———	.862	18.50	———	18.50	.0466
V-10	———	1.725	33.00	———	33.00	.0523
V-20M	———	3.45	55.00	———	55.00	.0627
50-A	———	5.00	140.00	———	140.00	.0357
V-20G2	Parallel	6.90	126.00	10.00	136.00	.0507
50-AG2	Parallel	10.00	310.00	10.00	320.00	.0313
V-20G3	Parallel	10.35	182.00	22.00	204.00	.0507
50-AG3	Parallel	15.00	460.00	22.00	482.00	.0311
230-Volt Service						
V-5H	———	0.575	21.00	———	21.00	.0274
V-10H	———	1.15	34.00	———	34.00	.0338
V-5G2	Series*	1.725	49.00	———	49.00	.0352
V-20HM	———	2.30	55.00	———	55.00	.0418
V-10G2	Series*	3.45	79.00	———	79.00	.0437
V-20HG2	Parallel	4.60	126.00	10.00	136.00	.0338
V-20G2	Series*	6.90	126.00	———	126.00	.0548
50-B	———	7.00	140.00	———	140.00	.0500
50-BG2	Parallel	14.00	310.00	10.00	320.00	.0438
50-GB3	Parallel	21.00	460.00	22.00	482.00	.0436
460-Volt Service						
V-5HG2	Series*	1.15		———	54.00	.0213
V-10HG2	Series*	2.30		———	81.00	.0284
V-20HG2	Series*	4.60		———	126.00	.0365
50-BG2	Series*	14.00		———	310.00	.0452

*Cannot be used where a common connection between input and output is required.

The trade name VARIAC is registered at the U. S. Patent Office. Variacs are patented under U. S. Patent No. 2,009,013 and British Patent No. 439,567 amination of Table I. For series operation of 230-volt units on 460-volt circuits, however, the Type 50-BG2 gives the greatest KVA per dollar.

There are practical limitations on the use of such assemblies. It should be noted that ganged units do not mount in the same space as a single unit, and that series connections cannot be used when a common ground is required between line and load. Further, chokes are necessary with parallel-connected Variacs, and choke equipment has to be mounted and connected, so that the ease and simplicity of use that characterizes the single-unit Variac is partially lost. A 2-gang or 3-gang unit with chokes cannot, for instance, be used conveniently on a laboratory bench, unless a special mounting is devised with proper plug-in terminal equipment.

On the other hand, for building into permanent equipment, where mounting and grounding requirements can be controlled, ganged assemblies can often be selected that will meet the power requirements at considerably lower cost than that of the next larger single unit.

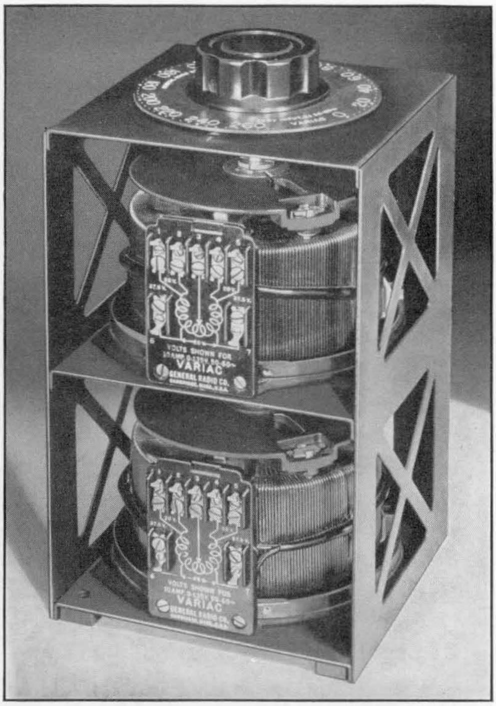
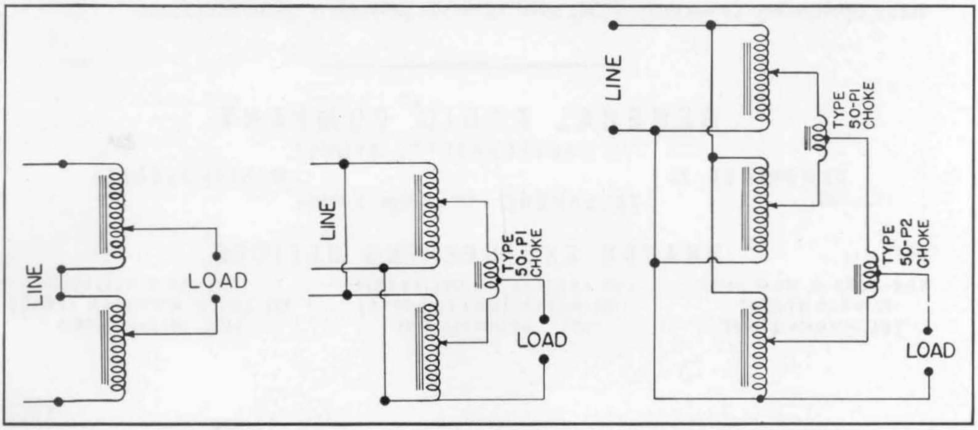


Figure 1. View of a Type V-10G2 Ganged Variac Assembly. The supporting structure is of welded steel and can be mounted either vertically or horizontally.

While standard gang assemblies of Variacs are limited to a maximum of three units, two or more gangs can be mechanically coupled by the user to obtain the same advantage in three-phase delta or wye circuits.

— Gilbert Smiley

Figure 2. Connection diagrams for ganged Variacs in single-phase circuits. Left, 2-gang series connection; center, 2-gang parallel; and, right, 3-gang parallel.



MISCELLANY

PAPERS — By Donald B. Sinclair, Assistant Chief Engineer, "A New Method of Measuring the Product of Two Voltages Using a Single Vacuum Tube," at the West Coast Convention of the I.R.E., Los Angeles, October 1, and at a meeting of the Boston Section, I.R.E., November 18. The paper was based on the work of Dr. M. A. H. El-Said, Senior Lecturer at Fuad I University, Cairo, Egypt. Dr. El-Said has been in this country on a Fellowship Technical Mission provided by the Egyptian Government. Some of the experimental work covered by the paper was performed by Dr. El-Said in collaboration with R. A. Soderman in the General Radio Laboratories.

RECENT VISITORS to our plant and laboratories — Dr. Jean Mercier, Dr. Rene Musson-Genon, and Mr. Ernest Rostas of the Hyperfrequency Research Laboratories, Cie. Francaise Thomson-Houston, Paris; and Mr. D. R. Austin, Consulting Engineer, Bartle and Co., Johannesburg, South Africa, distributors of General Radio products in South Africa.

NORWAY — It is with pleasure that we announce for the first time on these pages our representation in Norway by the firm Maskin-Aktieselskapet ZETA, Drummensveien 26, Oslo 22. ZETA has been the distributor of our products in Norway for several years and was appointed on an exclusive basis last spring. As our Norwegian friends know, the management is in the capable hands of Messrs. Braenne, Elligers, and Hammerik. Mr. G. Hammerik is in charge of the department that handles our products.

DENMARK — We also take pleasure in announcing the appointment of Mogens Bang & Co., Copenhagen/Skodsborg, as our representatives for Denmark. Mr. Mogens Bang is equipped to provide our Danish customers with all information concerning our products.

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 address, type of business company is engaged in, and title or position of individual.

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