the GENERAL RADIO Experimenter

INDEX

TO

GENERAL RADIO EXPERIMENTER

VOLUME 32

JUNE, 1957 through DECEMBER, 1958

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☆

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Printed in U. S. A.

the GENERAL RADIO Experimenter

Since 1915 – Manufacturers of Electronic Apparatus for Science and Industry

VOLUME 32 No. 8 JANUARY, 1958

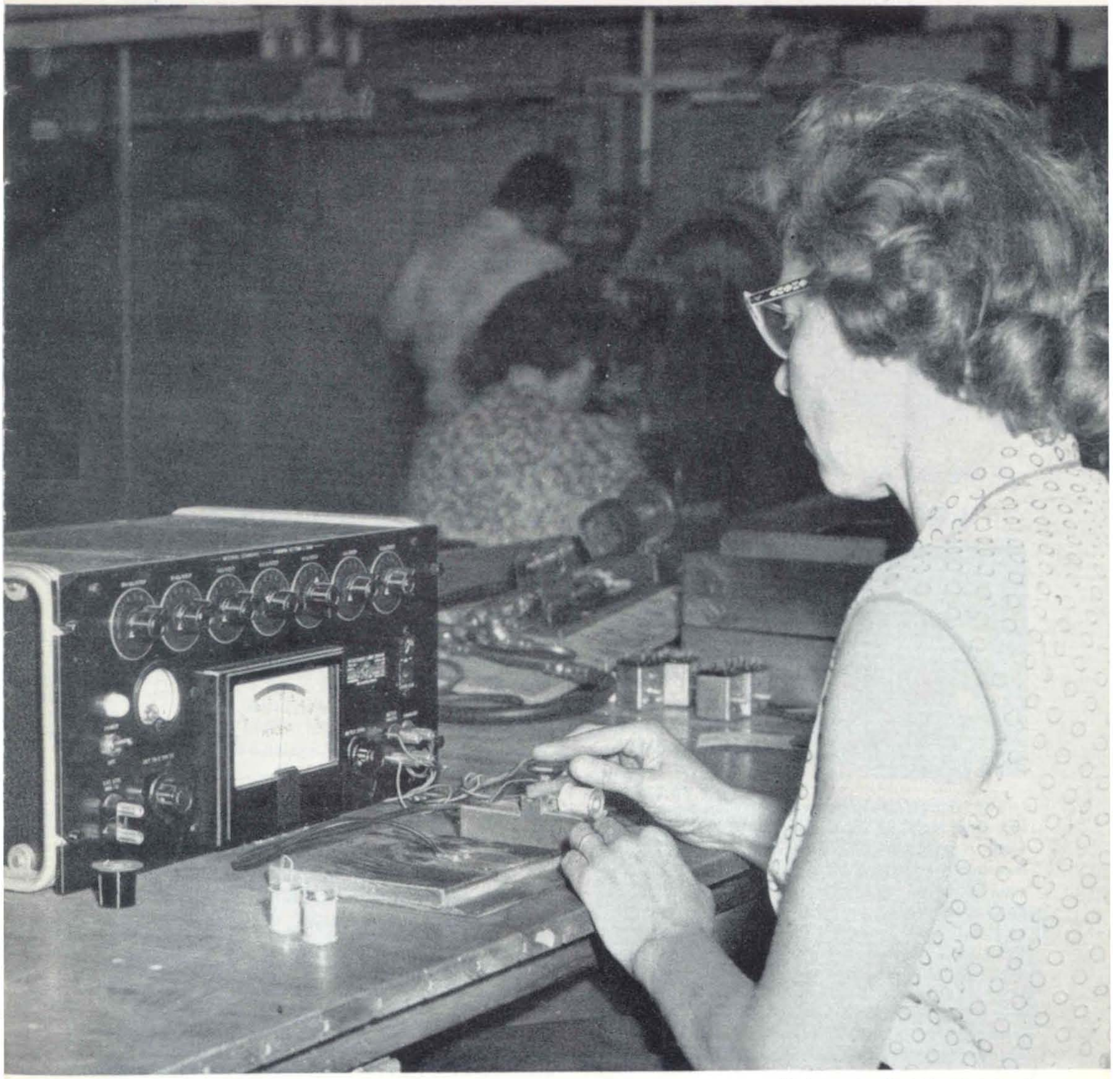

Photo courtesy Hart Manufacturing Company

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the GENERAL RADIO Experimenter

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The General Radio EXPERIMENTER is mailed without charge each month to engineers, scientists, technicians, and others interested in electronic techniques in measurement. 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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COVER

The Hart Manufacturing Company of Hartford, Conn., producer of "Diamond H" precision relays and switches, maintains a continuous check on relay coils with the General Radio Resistance Limit Bridge. Two bridges are used in their production operations, one in the coil winding department for initial d-c resistance measurements and one in the inspection department for final specification checks on the assembled relay.

A NEW UNIT OSCILLATOR, 180 to 600 Mc

Butterfly circuits with their wide tuning ranges and trouble-free operation at frequencies from about 100 to 1000 Mc are used in many General Radio Unit Oscillators. Their wide ranges and stable characteristics have helped to establish these inexpensive and versatile units as almost indispensable equipment in many laboratories and on many test benches.

Butterfly-type unit oscillators for 50 to 250 Mc and for 250 to 920 Mc have been available, but, in order to cover the military aeronautical bands between 225 and 420 Mc, both of these units are needed. The new TYPE 1209-BL Unit Oscillator covers this range in a single unit.

As the type number indicates, the new unit oscillator is a modification of the familiar 250-to-920 Mc TYPE 1209-B Unit Oscillator. By elimination of one of the two parallel inductance branches of the butterfly circuit, all frequencies are reduced by a factor of $\sqrt{2}$. At the same time the minimum output obtainable at any frequency is increased from 200 to 300 milliwatts. All other characteristics and the external appearance are unchanged.

SPECIFICATIONS

Frequency Range: 180 to 600 Mc.

Tuned Circuit: Modified butterfly, with no sliding contacts.

Frequency Control: 4-in. dial with calibration over 270°. Precision drive with 4½:1 reduction.

Frequency Calibration Accuracy: $\pm$ 1%.

Warm-up Frequency Drift: 0.2%.

Output System: Short coaxial line with adjustable coupling loop on one end and coaxial connector on other. Maximum power can be delivered to load impedances normally met in coaxial systems.

Output Power: Into 50 ohms, 300 mw at any frequency.

Modulation: Plate modulation of 30% at audio frequencies can be produced by external source of 40 volts. Input impedance is about 8000 ohms. When amplitude modulation without incidental f-m is required, the TYPE 1000-P6 Crystal Diode Modulator or TYPE 1000-P7 Balanced Modulator is recommended.

Power Supply Requirements: 330 v at 36 ma; 6.3 v at 0.4 amp.

Power Supplies Recommended:

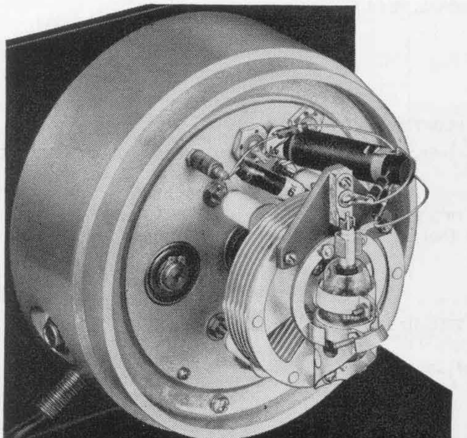
Standard: TYPE 1203-B Unit Power Supply, 115 volts, 50 to 60 cycles.

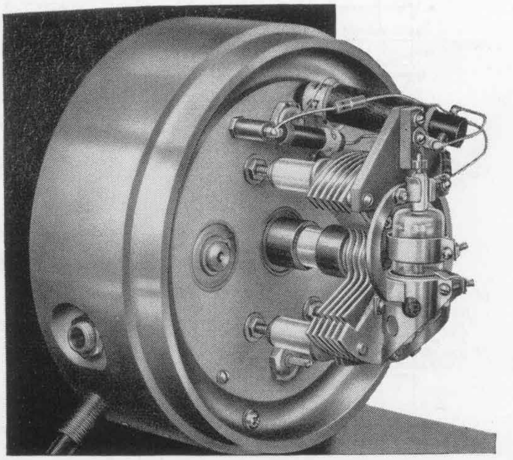
Stabilized Plate Voltage: TYPE 1201-A Unit Regulated Power Supply, 105 to 125 volts, 50 to 60 cycles.

Battery Operation: TYPE 1202-A Unit Vibrator Power Supply, 6- or 12-volt battery or 115 volts, 50 to 60 cycles.

Adjustable Plate Voltage: TYPE 1204-B Unit Variable Power Supply, 115 volts, 60 cycles.

Figure 1. (*Left*) View of butterfly circuit in the Type 1209-B Unit Oscillator. (*Right*) Rear view of Type 1209-BL, showing how one of the parallel inductance branches of the butterfly is removed to lower the frequency range.





Constant Output Level vs. Frequency: TYPE 1263-A Amplitude Regulating Power Supply with TYPE 874-VR Voltmeter Rectifier, TYPE 874-Q6 Adaptor, and TYPE 274-NF Patch Cord, 115 or 230 volts, 50 to 60 cycles.

Oscillator Tube: Sylvania Type RT 434.

Mounting: Aluminum casting surrounded by spun-aluminum shield. Assembly is mounted on L-shaped panel-and-chassis piece.

Accessories Supplied: TYPE 874-R22 Cable, TYPE 974-C58 Cable Connector, Jones socket, and telephone plug.

Accessories Available: Modulator, Sweep Drives, Relay Rack Adaptor Panels are available as listed in General Radio Catalog.

Dimensions: Height 6¼ in., width 9¼ in., depth 7 in., over-all.

Weight: 6¼ lbs.

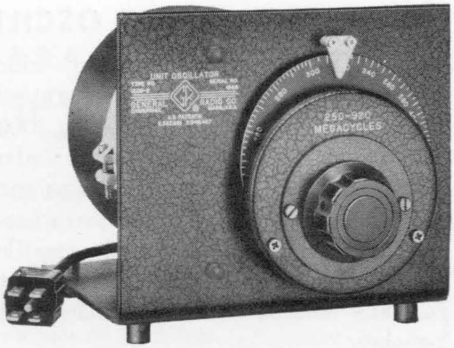


Figure 2. Panel view of the Type 1209-B Unit Oscillator. The new Type 1209-BL is identical in appearance except for the frequency calibration on the dial.

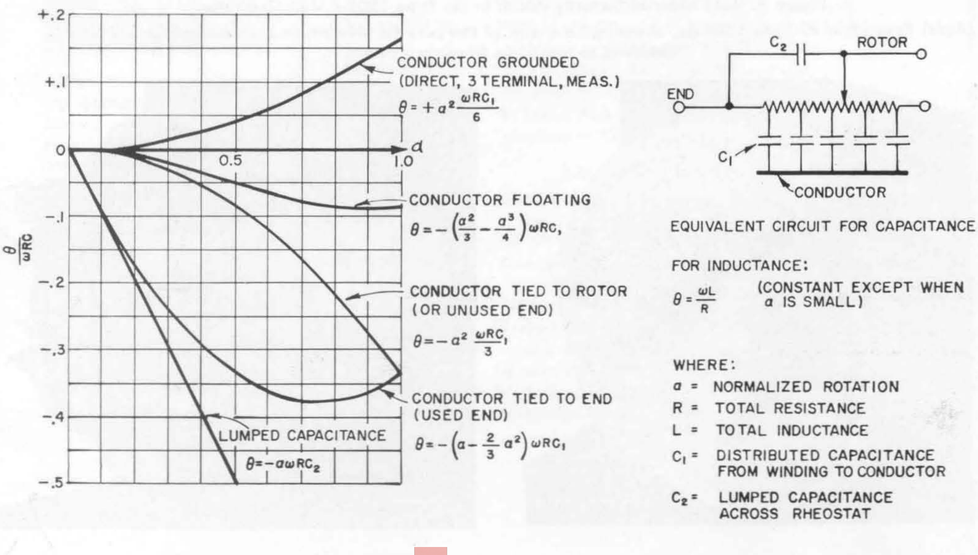
Type		*Code Word*	*Price*
1209-BL	**Unit Oscillator, 180 to 600 Mc**	ADMIT	**$245.00**

PHASE-ANGLE MEASUREMENT ON RHEOSTATS

An analysis of the effects of inductance and capacitance, both lumped and distributed, on the phase angle of a potentiometer used as a rheostat was presented in a previous article.[1] The distributed capacitance was assumed to be between the winding and a conducting surface (the housing, a metal mandrel, a mounting plate, or a shield), and the curve of the phase angle (θ) vs. fractional rotation (α) took various shapes, depending upon the potential of the conductor. These formulas and curves are summarized in Figure 1. The measured curves of the previous article were chosen to illustrate each effect separately. This article will describe the phase-angle characteristics of several potentiometers to show how their curves depend upon a combination of several effects.

[1] H. P. Hall, "The Phase Angle of Potentiometers used as Rheostats," *General Radio Experimenter*, 32, 4, September, 1957.

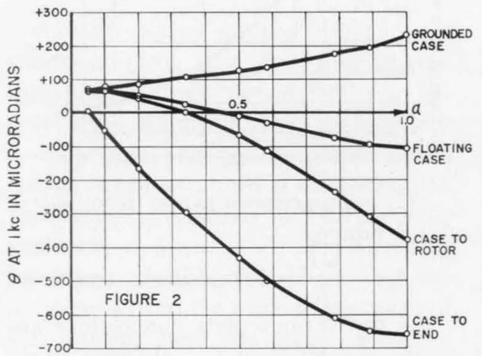
Figure 1.



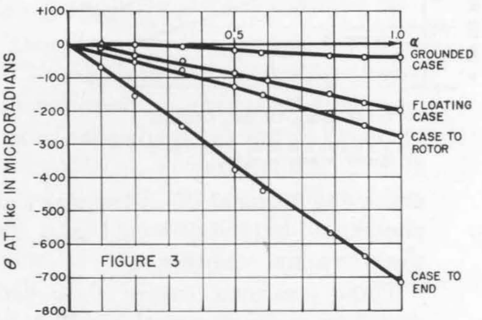
Measured Curves

Measured curves of θ vs. α for various potentiometers are given in Figures 2 to 8. The θ values are all given for 1 kc in order to facilitate comparison, although many of the measurements were made at higher frequencies in order to increase the sensitivity. The curves for the different types of potentiometers show quite a variation in magnitude and shape. The capacitances are more or less constant for all potentiometers of one type; a change in resistance should change only the resulting magnitude of θ. The inductance, however, depends upon the number of turns and wire size, so that it will vary as the resistance changes. The effect of resistance magnitude is illustrated in Figures 5 and 7, where two potentiometers of the same type are measured.

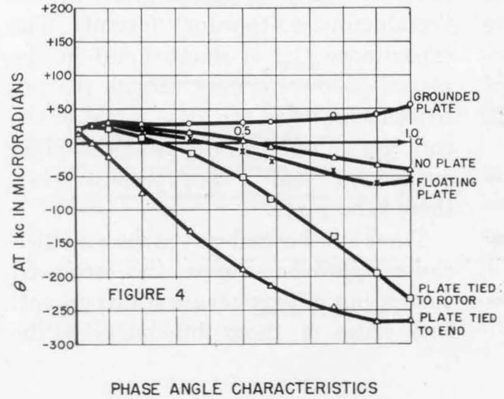
On potentiometers that do not have a metal case, such as the General Radio Type 970 Series, the "no plate" curve may be approached by using spacers to mount the potentiometer to the metal mounting plate. Figure 8 shows the behavior of a Type 977 Potentiometer spaced ¼ inch from the mounting plate, which gives substantial improvement over the curves of Figure 5. The "floating conductor" curve often gives the lowest θ for potentiometers with a metal case, but the case potential is affected by nearby objects. The "grounded conductor" connections usually result in an inductive phase angle, which can be easily compensated with a capacitor



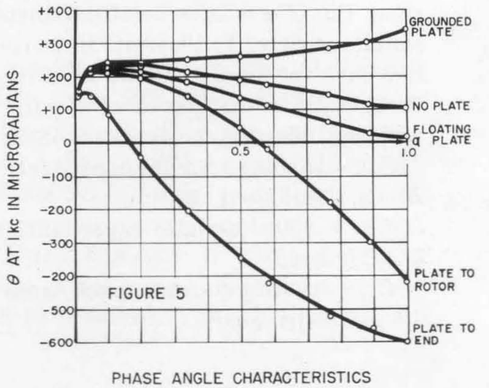
PHASE ANGLE CHARACTERISTICS FOR CLAROSTAT TYPE WX, 10KΩ, WIREWOUND POTENTIOMETER



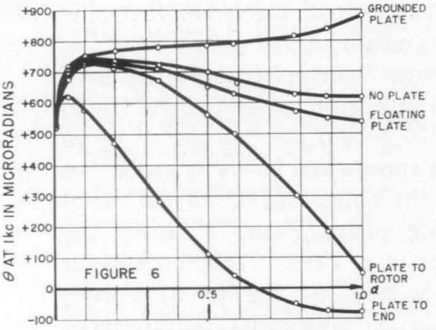
PHASE ANGLE CHARACTERISTICS FOR ALLEN BRADLEY TYPE JS, 10KΩ CARBON POTENTIOMETER



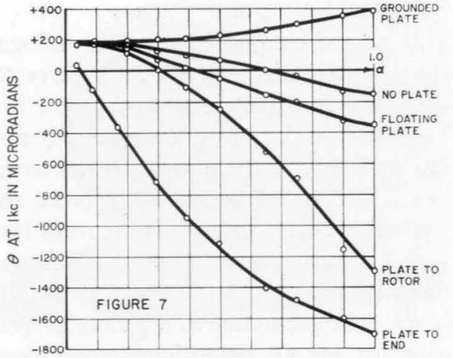
PHASE ANGLE CHARACTERISTICS FOR TYPE 971-N, 10KΩ WIRE WOUND POTENTIOMETER



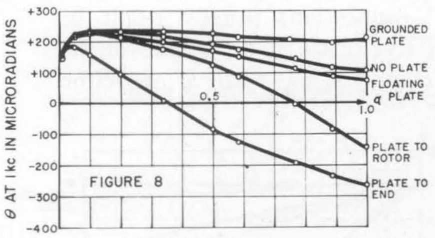
PHASE ANGLE CHARACTERISTICS FOR TYPE 977-N, 10KΩ, WIRE WOUND POTENTIOMETER



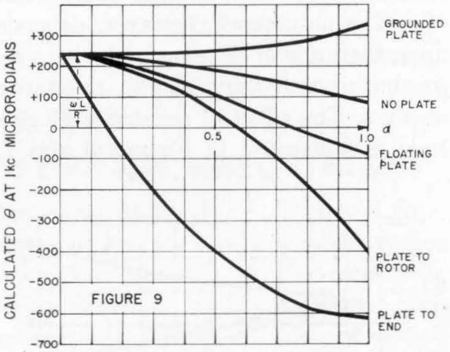
PHASE ANGLE CHARACTERISTICS
FOR
TYPE 978-N, 10KΩ WIREWOUND POTENTIOMETER



PHASE ANGLE CHARACTERISTICS
FOR
TYPE 977-P, 20KΩ WIREWOUND POTENTIOMETER



PHASE ANGLE CHARACTERISTICS
FOR
TYPE 977-N, 10KΩ WIRE WOUND POTENTIOMETER WITH 1/4" SPACERS TO MOUNTING PLATE.



SYNTHESIZED PHASE ANGLE CURVES

across the terminals. Not too many applications, however, would use this three-terminal connection.

These measured curves show some similarity to the curves of Figure 1, but large differences are apparent. The curves of Figure 1 show the effect of each parameter separately, whereas measurements on the potentiometer show the effects of several parameters simultaneously. To illustrate these combined effects, Figure 9 shows an attempt to synthesize the curves of Figure 5, with the following constants assumed:

C_{wp} = winding-to-plate capacitance = 24 $\mu\mu f$, distributed

C_{rp} = rotor-to-plate capacitance = 3 $\mu\mu f$, lumped

C_{wr} = winding-to-rotor capacitance = 4.5 $\mu\mu f$, distributed

C_A = capacitance across terminals = 1 $\mu\mu f$, lumped

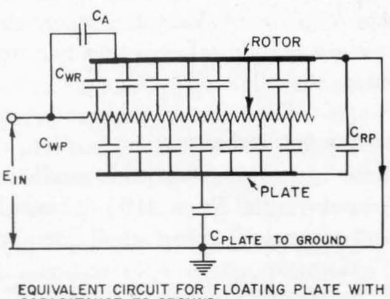
$$\theta_L = \frac{\omega L}{R} = 240 \ \mu\text{radians, constant}$$

As the various plate connections are made, the effect of C_{wp} changes as do the curves of Figure 1. However, C_A is always across the terminals, and the effect of C_{wr} is always as given by the "conductor-tied-to-rotor" formula. The capacitance C_{rp} is shunted out in the plate-tied-to-rotor case, shunts the potentiometer when the plate is tied to the end, has no effect in the grounded-plate connection, and is not present when there is no plate.

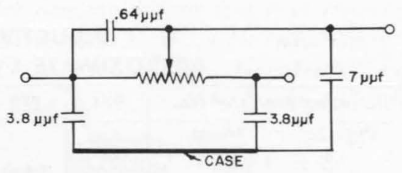
These synthesized curves show a much closer approximation to the measured curves, but differences are still apparent. One cause of these differences is the effect of capacitance from the winding and from the mounting plate to ground or to free space, which may be several micromicrofarads. This capacitance increases if grounded surfaces are brought near the plate. The effect of the winding-to-ground capacitance is the same as that of the capacitance from winding to plate in the "grounded-conductor" case when a direct measurement is made. This capacitance affects mainly the "no-plate" curve and makes it more positive. The plate-to-ground capacitance affects only the "floating-plate" curve, and, although the effect is more complicated, it makes this curve more positive also. Figure 10 shows an equivalent circuit for the "floating" case with the ground capacitance added, where the θ measured is that of the direct impedance, $\frac{E_i}{I_o}$.

Another reason for the differences is the assumption that the stray capacitances are evenly distributed. For example, the assumption that the winding-to-rotor capacitance is evenly distributed is only a fair assumption for Series 970 Potentiometers, which have a ring rotor structure, and would not hold at all for most potentiometers. If there were capacitances from the rotor to that part of the winding near the rotor contact, θ would vary as $-\frac{1}{\alpha}$ rather than as $-\alpha^2$

Figure 10.



EQUIVALENT CIRCUIT FOR FLOATING PLATE WITH CAPACITANCE TO GROUND.



EQUIVALENT CIRCUIT FOR POTENTIOMETER OF FIGURE 3

Figure 11.

as it does in the evenly distributed case. Not all the capacitances are included in the synthesized plots. In addition to the ground capacitance, there is probably some terminal-to-plate capacitance, which was not taken into account. Capacitances between various parts of the winding may also have a slight effect.

Still another effect is the presence of lumped capacitances which vary with α. For example, the capacitance from the terminal structure to the rotor will vary as the rotor is moved.

The curves for the carbon pot (Figure 3), which are almost straight lines, may be closely approximated if only the lumped capacitances shown in Figure 11 are considered. The area of the resistive element in this type of potentiometer is very small, so that distributed capacitances to the element are relatively small.

Summing up, we see that, although a combination of simple effects gives an approximation to the measured curves, differences are apparent, which are caused by additional factors that are much harder to analyze.

Phase-Angle Characteristics of General Radio Type 970-Series Potentiometers

Because several customers have inquired about the phase-angle characteristics of our 970-series pots, approximate specifications are given below. Usually it is not necessary to know the exact shape of the curve; the extreme values will be sufficient. The curve shapes are

TABLE I
INDUCTIVE PHASE ANGLE
APPROXIMATE θ_L in MICRORADIANS at 1 kc

Potentiometer Type No.		971	972	973	974	975	976	977	978
Type Letter	*Ohms*								
B	2	6500							
C	5	6300	9300	10600					
D	10	1800	3900	4900	9500				
E	20	740	1800	2200	4900				
F	50	620	910	1200	2800				
G	100	460	610	850	980				
H	200	340	430	730	760				
J	500	350	410	640	750	890			
K	1 K	130	440	330	640	730	900	1200	
L	2 K	72	102	147	275	360	610	820	980
M	5 K	63	73	120	170	215	420	610	990
N	10 K	30	68	80	120	165	210	240	730
P	20 K	16	15	35	62	130	130	200	250
Q	50 K	For these resistances, inductive effects are negligible compared to capacitive effects.							
R	100 K								
T	200 K								
U	500 K								

TABLE II
CAPACITIVE PHASE ANGLE at $\alpha = 1$
$\theta_{C\ max}$ in MICRORADIANS at 1 $k\Omega$ and 1 kc

Potentiometer Type		971	972	973	974	975	976	977	978
	No Plate	− 6.8	−11	− 8.6	− 9.4	− 7	−12	−13	−10
Plate Connected	to Ground	+ 3.4	+ .4	+ 5.7	+ 3.3	+11	+ 5.2	+12	+17
Plate Connected	to Winding End	−29	−33	−35	−36	−55	−57	−81	−81
Plate Connected	to Rotor	−26	−31	−30	−35	−46	−50	−66	−68
	Floating	− 8.3	−13	− 6.5	−12	−14	−18	−20	−17

These values depend upon the dielectric constant of phenolic, which varies with batch and with frequency.

quite similar for the various size pots, so that the complete curves can easily be approximated by drawing curves similar to those of Figures 4, 5, and 6.

Table I gives the approximate value for θ_L, the component of θ due to winding inductance for each size of each type of pot, except for the very large resistance values where the inductive effect is negligible compared to the capacitive effects. Table II gives $\theta_{C\ max}$, the θ value at $\alpha = 1$ caused by the capacitive effects. The value of $\theta_{C\ max}$ depends upon the connection used and is negative except for the grounded-plate connection. The mounting plates used for these measurements were square, and the edge dimension was one inch greater than the diameter of the pot.

To get the values at $\alpha = 1$, the appropriate value of θ_L and $\theta_{C\ max}$ should be chosen. The value of θ at $\alpha = 1$ is then:

$$\theta = f\,(\theta_L + R\theta_{C\ max})$$

where f is in kilocycles and R in kilohms. As α approaches zero, θ, for the synthesized curves, approaches θ_L as shown in Figure 9. Actually, as shown in Figure 5, the inductive component of θ drops off at low values of α, where the full extent of mutual coupling has not been established.

Measurement Technique

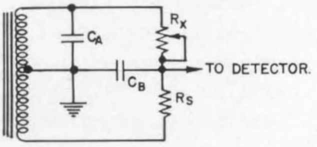
These measurements were made on a General Radio TYPE 1605-A Impedance Comparator[2], using small fixed re-

[2]Hall, H. P., and Holtje, M. C., "A High Precision Impedance Comparator," *General Radio Experimenter*, 30, 11, April, 1956.

sistors as phase-angle standards. This instrument presents on two meters the difference in magnitude and the difference in phase angle between the standard and the unknown impedances. The high sensitivity (one division = .0001 radian on the most sensitive range) and the wide frequency range (100 c, 1 kc, 10 kc, and 100 kc) make possible measurements over a very wide range of time constant (θ/ω).

In the measurement, the rheostat is adjusted for an approximate magnitude null, and then the $\triangle\theta$ value is read from the meter. For the most precise measurements, the standard resistors should have low capacitance and short leads. Because we are interested in the θ *difference*, the capacitance of leads to the rheostat can be balanced out by using similar leads to the standard.

It is interesting to see how direct measurements can be made if the rheostat is adjusted for a magnitude balance. The shunting capacitances C_A and C_B (Figure 12) have no effect in a direct measurement. C_A shunts half of the tightly coupled bridge transformer and causes negligible error if it is of reasonable size (less than .0001 radian error for $C_A = 0.1\mu f$ at 1 kc). The other stray capacitance, C_B, shunts the input and



BRIDGE CIRCUIT WITH STRAY CAPACITANCE

Figure 12.

causes an error of $\frac{1}{2}\triangle R\,\omega\,C_B$, where $\triangle R$ is the resistance difference between the rheostat and the standard resistor. It is of course impossible to set $\triangle R = 0$, particularly for wire-wound potentiometers of limited resolution. However, if C_B is of the same order of magnitude as the capacitance causing the phase angle, the error is usually negligible.

The author wishes to thank R. G. Fulks and J. M. Flower for making many of the measurements used in this article.

— H. P. Hall

THE NEW MIDWEST OFFICE

July 15, 1957, marked the opening of our new office in Oak Park and the closing of the old office at 920 South Michigan Avenue in Chicago, which had been in operation since 1943. During the intervening years, the office had grown in size more than threefold. This expansion in personnel together with our plans for a Midwest repair department made it necessary to move to larger quarters. A suburban location was desirable, because more and more of our customers were moving away from the downtown area.

Our new location at 6605 West North Avenue is in Oak Park about ten miles west of Chicago's Loop. Easy access to and from locations in Chicago is possible via North Avenue, the main east-west artery, or the Congress Expressway located nearby. Also, several north-south arteries, such as Harlem Avenue and Cicero Avenue, are in the vicinity. Convenient public transportation is also available, while those who drive will find the location easily and will have no difficulty in parking. There is a parking lot at the rear of the building as well as ample space on the adjacent streets.

There is a large demonstration room equipped with representative General Radio instruments and soon a good stock will be available for over-the-counter sales.

Sales engineers at the Midwest office. Left, Bob Bard; right, Bill Ihde.

The New Service Department

One of the major considerations in the new location was to provide a complete factory service department, and about one half the space is now devoted to servicing General Radio equipment. Alfred J. Guay, a factory-trained service engineer, is in charge of this operation and is ably assisted by expert service technicians. This department has been in operation since August 1. Facilities are available for standardization, recalibration, and repair of General Radio instruments.

View of the new front door — 6605 West North Avenue, Oak Park, Illinois.

The service laboratory at General Radio's new Midwest office. ***Left to right,*** **Al Guay in charge of service activities and George Hanson.**

For those customers interested in making their own repairs, a good stock of replacement parts is available.

Many service problems can be solved by discussing them with us. Please call on Bill Ihde or Bob Bard if you have a measurement problem, or Al Guay if you have a service problem in connection with GR equipment.

We cordially invite you to call or to visit us at any time.

Telephone: VIllage 8-9400.

— William M. Ihde

SCIENCE FOR THE BLIND

A unique publication, *Science Recorded,* entered its third year in October, 1957. *Science Recorded* is a scientific periodical, recorded on magnetic tape, and sent each month to some 200 blind subscribers in the United States, Great Britain, Australia, and Mexico. It is handled by a nonprofit organization, Science for the Blind, a subsidiary of the Philadelphia Association for the Blind. In the twelve months ending October, 1957, over 4000 tapes were sent out.

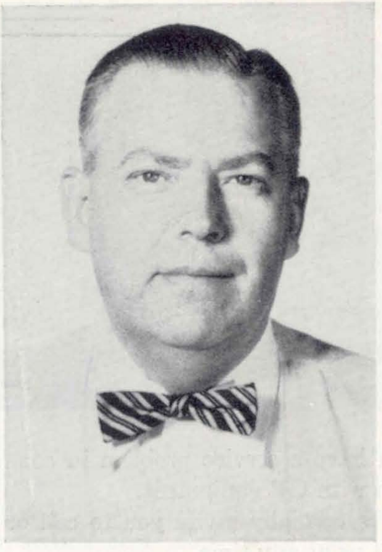
The staff is made up primarily of volunteers who serve as readers and mailers. The editor is Professor T. A. Benham of Haverford College, who is well known for his work in the development and adaptation of scientific instruments for use by the blind.

Among the sources of material for publication in *Science Recorded* is the *General Radio Experimenter*.

. . . IMPORTANT . . .

Be sure to save the new General Radio price list enclosed, and file with your Catalog O.

Effective date: **January 20, 1958**



Ira G. Mercer

GR REPRESENTATION FOR HAWAII

Radio-Television Corporation, Ltd., 777 Ala Moana, Honolulu, Hawaii, has been appointed General Radio's representative for the Territory of Hawaii. Ira G. Mercer, Manager of the Electronics Division and well known in electronic engineering circles both on the mainland and in Hawaii, is in direct charge of GR sales.

All interested in electronic test equipment and instrumentation questions are cordially invited to get in touch with our new representative. The telephone number is 50-2901.

NEW LOCKING SYSTEM FOR UNIT INSTRUMENTS

General Radio Unit Instruments have plug-in power supplies. This feature makes possible the most efficient use of power-supply units and also permits the user to select either a regulated or an unregulated supply, as the immediate requirements may dictate.

With the rectangular-cabinet type of unit instrument, it is possible to combine this feature with the advantages of a single structure. This is accomplished by the use of locking strips, which hold the power-supply cabinet and the instrument cabinet firmly together, so that they can be handled as a single assembly. The locking strips are now supplied with all rectangular-cabinet unit instruments. Installation is simple, as shown in Figure 1. Two locking strips are used, one at the top and one at the bottom.

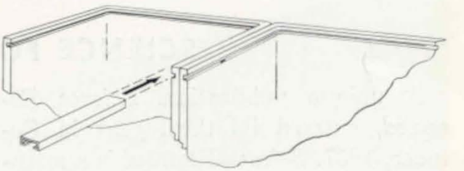


Figure 1.

the GENERAL RADIO Experimenter

Since 1915 – Manufacturers of Electronic Apparatus for Science and Industry

VOLUME 32 No. 9 FEBRUARY, 1958

Photo courtesy Vitramon, Inc.

In This Issue

Capacitance Bridges
Radio Engineering Show

the GENERAL RADIO Experimenter

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COVER

One of the earliest users of the Type 1605-A Impedance Comparator is Vitramon, Inc., of Bridgeport, Connecticut, manufacturers of ceramic capacitors. Vitramon uses Impedance Comparators for production testing, as shown in the photograph, and also for engineering tests to determine the effects of environmental factors such as temperature, humidity, immersion, and vibration.

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THE TYPE P-582 CAPACITANCE BRIDGE

AN INSTRUMENT DESIGNED TO CALIBRATE CAPACITIVE FUEL-GAGE TESTERS

The full range of an airplane can only be achieved if the plane lands with a dry fuel tank. Any fuel left over has not only reduced the range by not being used, but it has also added to the weight of the plane during the entire trip. A margin of safety is, of course, necessary, but an accurate fuel gage can keep this margin from being unnecessarily large.

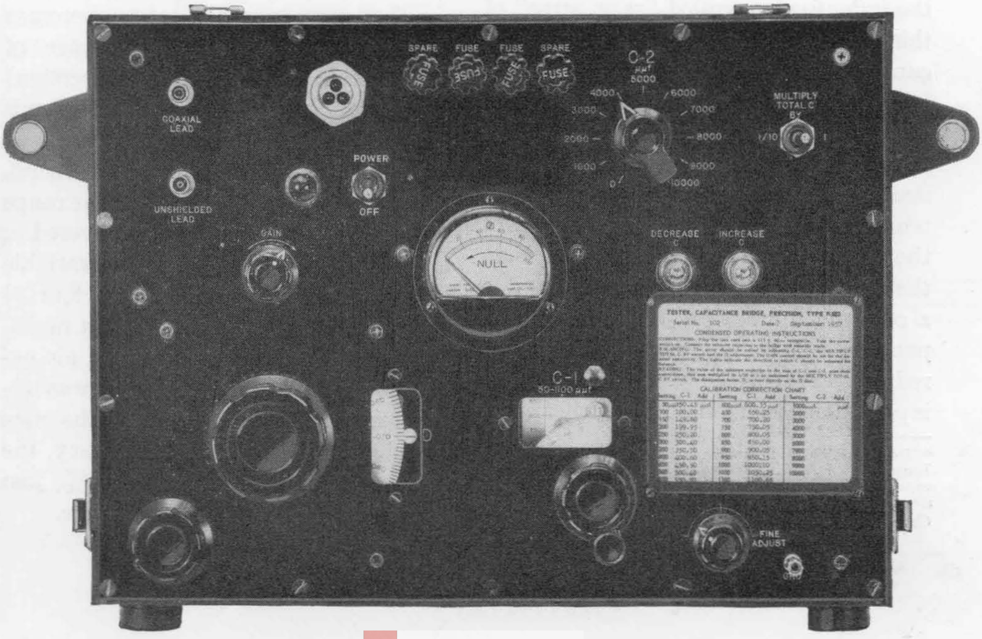
The fuel gage on a modern military aircraft measures the fuel quantity by measuring the capacitance of an element inserted into the fuel tank[1]. This element is, in effect, a capacitor whose capacitance changes as fuel is replaced by air. Periodic checks of the accuracy at full and empty gage readings are made by the substitution of a variable capacitor, such as the MD-1 (GR Type P-579) Field Variable Capacitance Tester,[2, 3] for the tank elements. The gage can be only as accurate as the test capacitor, which, although of high precision, should also be tested periodically, for an inaccuracy could result in a loss of life. Thus a bridge is necessary to test the tester which tests the fuel gage.

The GR Type P-582 Capacitance Bridge, shown in Figure 1, was developed to meet this need. Actually, original specifications (MIL-T-4778 (USAF)) called for a bridge of somewhat different design, but the Type P-582, which meets the essential requirements, is a smaller and lighter instrument, and yet provides greater range, higher accuracy, and has many convenience features.

After evaluation tests (including many environmental tests) the P-582 was given the militarily assigned commercial standard designation TEST SET, CAPACITANCE BRIDGE, TTU 24/E, PRECISION, THREE-TERMINAL, DEPOT.

[1] "Temperature Compensated Aircraft Fuel Gage," R. J. Levine, ELECTRONICS, 27, 9, September, 1954, pp. 160-1.
[2] "A Calibrator for Aircraft Fuel Gages," P. K. McElroy, *General Radio Experimenter*, September, 1955.
[3] "Electronic Fuel Gage Tester," P. Bishop, ELECTRONIC INDUSTRIES, May, 1957, p. 75.

Figure 1. Panel view of the Type P-582 Capacitance Bridge.



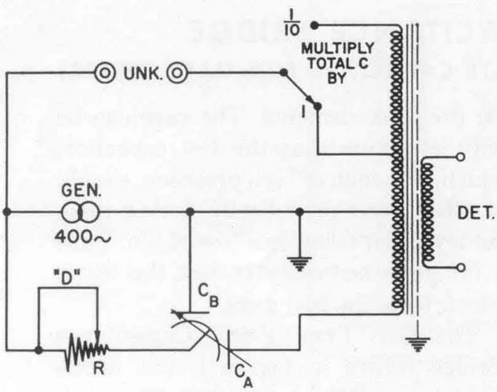


Figure 2. Basic bridge circuit.

This instrument is a self-contained bridge system which includes a 400-cycle oscillator and a sensitive null indicator. Since the fuel gage measures direct capacitance[1], a "transformer bridge"[4] was specified to make possible direct (three-terminal) capacitance measurements.[5] The basic bridge circuit is shown in Figure 2.

Capacitance from the unknown terminals to ground shunts either the oscillator or the tightly coupled transformer and has no effect unless it is very large (see specifications). In addition, the inductively coupled "ratio arms" of this type of bridge provide a ratio accuracy unattainable by other types of components. As a result, the accuracy of capacitance measurements depends almost entirely on the accuracy of the standard capacitor. The range is extended by changing the turns ratio of the transformer. The transformer is in the detector circuit which makes possible a constant voltage on the unknown capacitor as the range is changed, with no reduction in sensitivity. The transformer is potted inside of two, high-permeability shield cans to prevent stray fields from affecting the balance.

The important feature of this bridge is the T-network used in the "standard" side of the bridge. The direct impedance of this network balances out the direct impedance of the unknown at balance. The balance equations are:

$$C_X = C_A\,(M)$$

where M is a multiplier of 1 or 1/10 (as selected) and C_X is the series capacitance of the unknown.

$$D_X = \omega R\,(C_A + C_B)$$

The sum $(C_A + C_B)$ is kept constant, so that R is proportional to D_X and, therefore, is directly calibrated in dissipation factor. Other transformer-type capacitance bridges require a computation to obtain this quantity.

The standard capacitor, which forms the differential unit consisting of C_A and C_B, is actually a 50-to-1100 $\mu\mu f$ variable air unit and a 1000-to-10,000 $\mu\mu f$ decade of silvered-mica capacitors. The sum $(C_A + C_B)$ is kept constant by the addition of an extra set of stator plates to a TYPE 722 Capacitor, and by differential switching of the mica units. On the ×1/10 range the lower limit is extended down by a factor of 1/10, so that the over-all range becomes 5$\mu\mu f$ to 11,000 $\mu\mu f$. The accuracy of the variable unit (with dial correction) is ± 0.4 $\mu\mu f$ or ± 0.1%, whichever is larger. In the ×1/10 position, this means a bridge accuracy of ± .04 $\mu\mu f$ or ± 0.1% over the range of 5 to 110 $\mu\mu f$. The range from 110 to 1100 $\mu\mu f$ can be covered in either of two ways: (1) with the variable air capacitor and a multiplier of ×1; or (2) by use of the mica decade and a multiplier of ×1/10. When only the air capacitor is used, the accuracy of measurement is 0.4 $\mu\mu f$ or 0.1%; with the mica decade, which has a 0.1% accuracy, the over-all accuracy is close to 0.1%, just as on the 1000-to-11,000 $\mu\mu f$ range.

[4] "Double Ratio A-C Bridges with Inductively Coupled Ratio Arms," A. A. Clark and P. B. Vandeclyn. *Proceedings of IEE*, 92, Part III, 1949, pp. 189-198.

[5] "Direct Capacitance and Its Measurement," R. F. Field, *General Radio Experimenter*, November, 1933.

The percentage accuracy for capacitance measurement (with dial correction) is plotted against capacitance in Figure 3, which also shows the original specifications. The small jogs in the curve are the effect of the $\pm 0.4\ \mu\mu f$ specification of the variable capacitor and are of small consequence. The original specification called for a 1000 $\mu\mu f$ variable capacitor with an accuracy of $\pm 0.2\ \mu\mu f$ or $\pm 0.1\%$, which is unobtainable in present capacitors without a much more detailed correction chart than that called for by the specifications.

The dissipation-factor accuracy is $\pm 2\%$ of reading ± 0.0002. The small losses of the mica decade are balanced out by resistive networks on the unknown side of the bridge which are switched as the decade is adjusted.

The 400-cycle oscillator is thermistor stabilized and uses a Wien bridge selective *R-C* network in a three-stage feedback circuit, which has such a high loop gain that the frequency is practically independent of tube parameter changes. The frequency-determining components are GR precision resistors and capacitors. A buffer cathode-follower amplifier is added to prevent external loading from affecting the frequency.

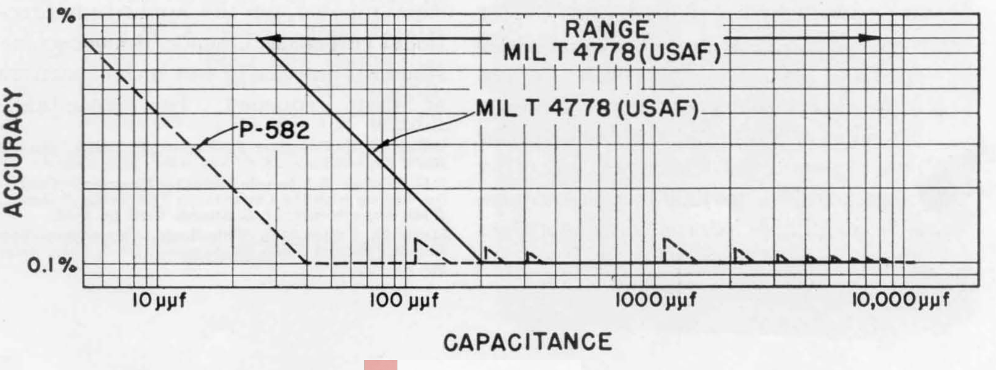
The detector circuit has several features that facilitate rapid and accurate balances. Two cascaded, selective, twin-T, feedback amplifiers provide high selectivity, and the sensitivity is more than adequate for a precise balance. The null indicator has a compressed response and uses a ruggedized meter. Two panel lights indicate the direction of capacitive unbalance when the meter is upscale. Thus it is not necessary to use a trial-and-error method of deciding which way to vary the standard capacitor when the unknown is first connected. This feature greatly reduces the time required to balance the bridge and usually makes gain-control adjustments unnecessary.

It should be noted that, although this bridge was designed for a particular purpose, it is also suitable for general capacitance measurements at 400 cycles. Its accuracy, range, portability, and self-contained nature make it a most useful instrument for precise, three-terminal measurements.

The author would like to acknowledge the contributions of the many people at General Radio whose ideas were incorporated in the instrument. The guiding hand behind the development was that of P. K. McElroy, whose aid in the design also assured compliance with the requirements of environmental tests. The ideas of I. G. Easton and R. A. Soderman were used in the circuitry, and the layout was largely the work of H. G. Stirling.

— Henry P. Hall

Figure 3. Plot of percentage accuracy versus capacitance measured.



SPECIFICATIONS

Range: Capacitance: 5 $\mu\mu$f to .011 μf.
Dissipation Factor: 0 to .11.

Accuracy: Capacitance: see Figure 3. Variable Capacitor: on ×1 Range, ± 0.4 $\mu\mu$f or ± 0.1%, whichever is greater; on ×1/10 Range, ± 0.04 $\mu\mu$f or ± 0.1%, whichever is greater. Decade Capacitor: ± 0.1% on both ranges.
Dissipation Factor: ± 2% of reading ± .0002.

Oscillator: Frequency: 400 cps ± 0.25%.
Output: 25 volts nominal.
Distortion: less than 0.5%.

Detector:
Sensitivity of amplifier alone: 10% scale deflection for 10 μvolts input.
Sensitivity of system: ×1 MULTIPLIER position —10% deflection for .05 $\mu\mu$f ΔC.
×1/10 MULTIPLIER position—10% deflection for .005 $\mu\mu$f ΔC.
Selectivity of amplifier alone: down 56 db at 800 cps, down 64 db at 60 cps.
Selectivity of amplifier and bridge transformer: down 50 db at 800 cps, down 80 db at 60 cps.

Effect of Impedance to Third Terminal (Chassis): Impedance from the unshielded lead to chassis shunts the oscillator and, therefore, causes no bridge error. The output voltage is reduced approximately 50% by shunt impedance of 5 kΩ or 0.1 μf.
Impedance from the coaxial lead to chassis shunts the bridge transformer. On the ×1 MULTIPLIER position, there is negligible effect from a shunt of 1 kΩ or 0.1 μf. On the ×1/10 MULTIPLIER position, there is negligible effect from 10 kΩ or 0.01 μf.

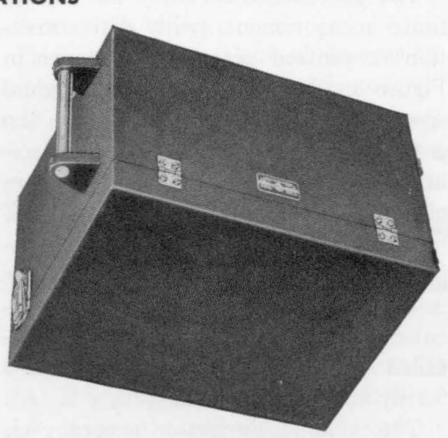


Figure 4. View of Type P-582 Capacitance Bridge with cover in place.

Accessories Supplied: One power cord.

Tube Complement: 5-5751; 1-12AT7WA; 1-6X4WA.

Power Supply: 105 to 125 volts, 50-60 cycles.
30 watts input at 115-volt line.

Dimensions: (Length) 22½ in. x (height) 14 in. x (depth) 12¾ inches over-all, including cover.

Net Weight: 55 lbs.

Type		*Code Word*	*Price*
P-582	**Capacitance Bridge**	SUPER	**Price on request**

CAPACITANCE TEST BRIDGE

60/120-CYCLE MODEL NOW STANDARD ITEM

The special 60/120-cycle model[1] of the popular Type 1611-A Capacitance Test Bridge[2] has enjoyed such widespread acceptance that it now becomes the standard model and will be known as Type 1611-B. This bridge will measure capacitance and dissipation factor over wide ranges; at a frequency of 60 cps from the power line and at 120 cps from an external generator. Filters are provided in the null-detector circuit for both frequencies. Operation at other frequencies, up to 1000 cps, is also possible, if suitable external generators and filters are used.

120-Cycle Measurements

Capacitance measurements at 120 cps are used primarily for the measurement of polarized electrolytic capacitors. In the majority of applications of these capacitors, a 120-cycle ripple current is superimposed on the applied unidirectional voltage and, hence, it has become standard practice to test such capacitors at that frequency. The capacitance

[1]"Electrolytic Capacitor Testing at 120 Cycles," *General Radio Experimenter*, 28, 6, November, 1953, page 8.

I. G. Easton, "A 120-cycle Source for Electrolytic Capacitor Testing with the Capacitance Test Bridge," *General Radio Experimenter*, 31, 3, August, 1956, pp. 9-12.

[2]Ivan G. Easton, "A Wide-Range Capacitance Test Bridge," *General Radio Experimenter*, 23, 2, July, 1948, pp. 1-8.

range of the bridge is 1 μf to 11,000 μf and the dissipation-factor range is 0 to 120% when the frequency of the test voltage is 120 cycles. The D dial is calibrated for 60-cycle measurements; at 120 cycles the correct value of dissipation factor is obtained by multiplying the dial readings by a factor of 2.

Measurements at 60 cps have many uses: In the laboratory and shop, for the testing of paper, mica, and other capacitors; in the electric-power industry, for the shop testing of insulators, particularly for the measurement of the dissipation factor of bushings and insulators, transformers, rotating machines, and cable; in the electronics industry it is used not only for measuring component capacitors, but also for measuring capacitance to ground of transformer windings, shields, and circuit elements.

The capacitance range for 60-cycle measurements is 0 to 11,000 μf; the dissipation-factor range is 0 to 60%. The extension of the capacitance range to zero is made possible by a unique compensating circuit[2], which eliminates the effect of stray capacitance at the bridge terminals. If, for practical purposes, we consider the lower limit of measurement to be one micromicrofarad, the bridge covers a capacitance range of 11 billion to one.

Other Frequencies

Measurements at other frequencies up to 1000 cycles can be made with this bridge, over a range of 1 μf to 11,000 μf. Both external generator and external filters must be used. The dissipation-factor range is 0 to f%, where f is operating frequency in cycles per second. Dissipation-factor dial readings are multiplied by $\frac{f}{60}$ to obtain the true dissipation factor, expressed in per cent.

Accessory Equipment

For 120-cycle measurements, the recommended external generator is the Type 1214-AS2 Unit Oscillator. For higher frequencies the Type 1210-B Unit R-C Oscillator is satisfactory, or for fixed-frequency measurements at 1000 cycles, the Type 1214-A Unit Oscillator. Polarizing voltage can be furnished by any convenient d-c power supply. For voltages up to 300 volts, the Type 1204-B Unit Variable Power Supply is recommended.

SPECIFICATIONS

Capacitance Range: 0 to 11,000 μf at 60 cycles. 1 μf to 11,000 μf at 120 cycles or other external frequency.

Dissipation-Factor Range: 0 to 60% at 60 cycles. Range proportional to frequency. (0 to 120% at 120 cycles.) Dial readings must be multiplied by the ratio $\frac{f}{60}$ for frequencies other than 60 cycles.

Accuracy: Capacitance, ±1%. Dissipation factor, ±(2% of dial reading +0.05% x $\frac{f}{60}$ dissipation factor).

Detector Filter: Tuned to 60 or 120 cycles, selected by switch. Jack provided for use of an external filter for other frequencies.

External Generator: Required for frequencies other than 60 cycles. Type 1214-AS2 Oscilla-

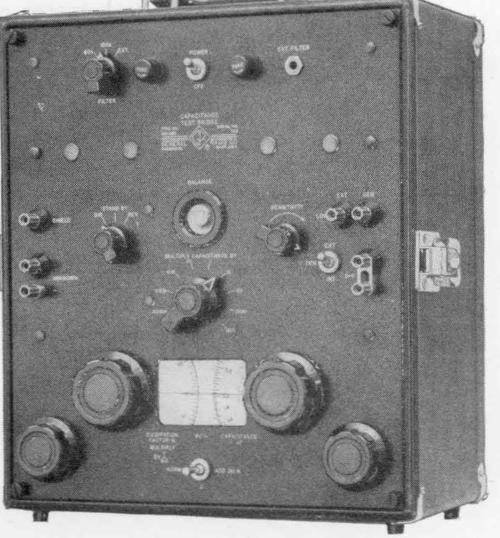


Figure 1. Panel view of Type 1611-B Capacitance Test Bridge.

tor listed below is recommended for 120-cycle measurements.

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: Type CAP-35 Power Cord and spare fuses.

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

Tube Complement: One each 6X5-GT/G, 6SJ7, and 6U5.

Net Weight: 30½ pounds.

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

Type		*Code Word*	*Price*
1611-B	Capacitance Test Bridge	FAVOR	$570.00
1214-AS2	Unit Oscillator (including power supply)	ABBOT	100.00

IRE—1958 RADIO ENGINEERING SHOW

New York Coliseum, March 24-27

Experimenter readers who attend the 1958 Radio Engineering Show are cordially invited to visit the General Radio display in Booths 3302 to 3310. These booths are on the third floor of the Coliseum, which is devoted almost completely to electronic instruments and test equipment.

Several new instruments will be on display, among them:

Slotted Line for Dielectric Measurements — A new model of the popular General Radio slotted line, developed particularly for measurements of dielectric constant and dissipation factor of solid insulating materials. Operating at frequencies between 200 and 5000 megacycles, this line will measure dielectric constants between 1 and 10 with an accuracy of approximately $\pm 2\%$, and dissipation factors between 0.0001 and 0.1 with an accuracy of $\pm$ (5% + 0.0001).

Transfer Function Meter — A new VHF-UHF instrument for the direct measurement of forward and reverse transfer functions of transistors, vacuum tubes, and networks.

Capacitance Bridge — A transformer bridge, which measures direct (3-terminal) capacitance over a range of 5 $\mu\mu f$ to 0.011 μf at 400 cycles. This bridge is fully described elsewhere in this issue of the *Experimenter*.

6-Gang Variac® Autotransformer — Six Type W50 Variac® Autotransformers on a single shaft, rated at 34½ KVA.

Other Important Displays include Pulse Equipment, Laboratory Standards, Voltage Regulators, Electrometer, Impedance Bridges, General Radio Parts and Connectors, and a complete line of the well-known General Radio Unit Instruments.

Plan to Visit the GR Booths. You'll find it worth your while. Our engineers will be on hand to greet you and to discuss your measurement problems.

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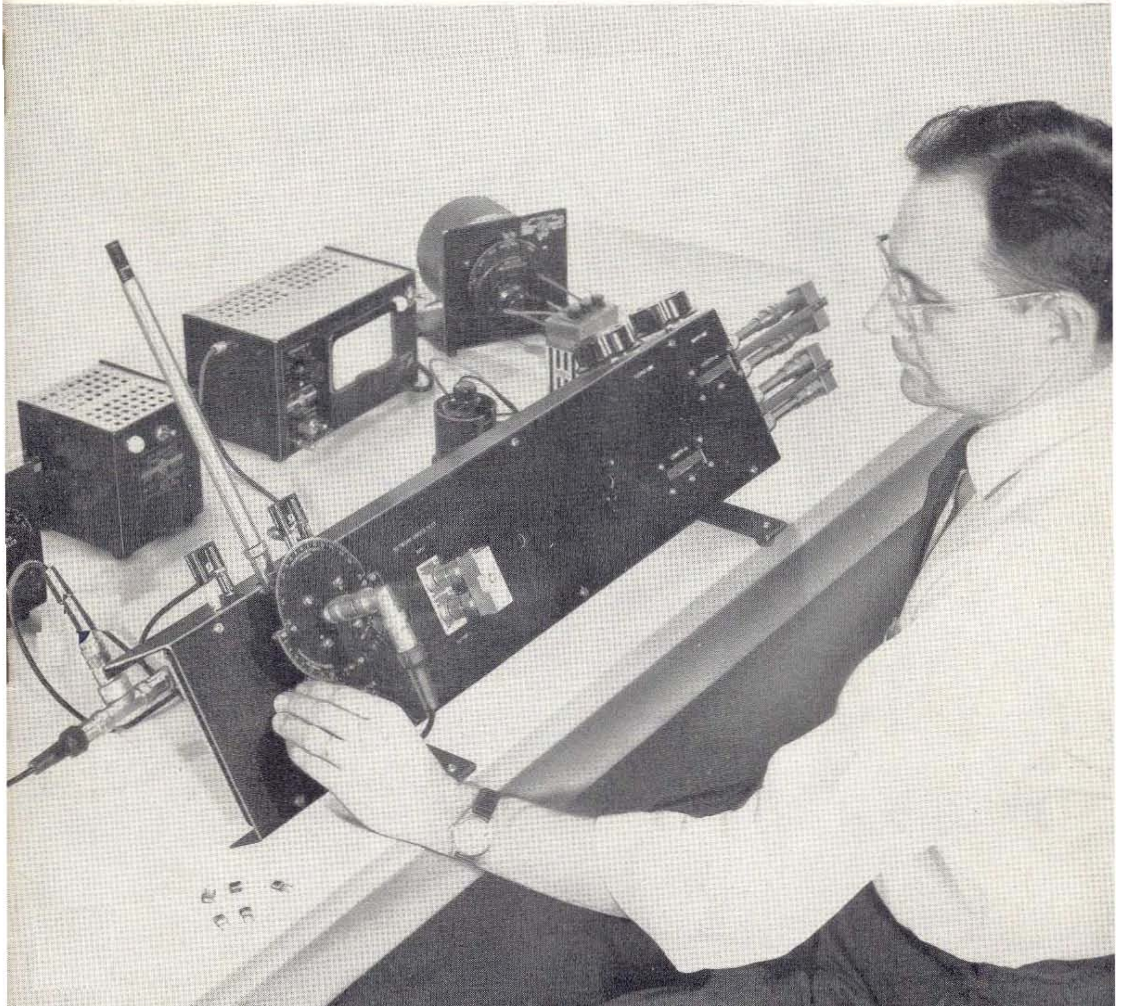
VOLUME 32 No. 10 MARCH, 1958



In This Issue

A Transfer-Function Meter for the VHF-UHF Range

the GENERAL RADIO **Experimenter**

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The General Radio EXPERIMENTER is mailed without charge each month to engineers, scientists, technicians, and others interested in electronic techniques in measurement. 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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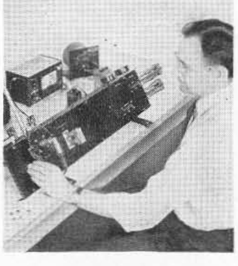
COVER



The Transfer-Function Meter described in this issue is a completely new, null-type device, which opens up a new field of measurement. The cover photograph shows the Transfer-Function Meter set up for the measurement of transistors at 400 Mc.

A TRANSFER-FUNCTION METER FOR THE VHF-UHF RANGE

The performance of most electrical devices and circuits can be described by specifying a transfer function, which is the ratio of an output to an input quantity, or vice versa. The "alpha" and "beta" current ratios of transistors, the transconductance of vacuum tubes, the gain of amplifiers, and the loss of attenuators and filters are common examples of widely used transfer functions. The new TYPE 1607-A Transfer-Function Meter[1] can measure these and many other transfer functions over the frequency range from 25 Mc to about 1500 Mc. Since grounded or ungrounded two-terminal impedances can be treated as four-terminal devices,[2] they can be measured as well, with appropriate adaptors. Answers, direct reading except for a multiplying factor, are obtained in terms of complex components by a null method. The phase information provided by measurement of complex components is especially valuable at these high frequencies, where effects of transit time, electrode resonances, and stray capacitances usually dominate the over-all performance of a device.

The Transfer-Function Meter is a basic measuring tool, well suited for laboratory measurements because of its versatility, accuracy, and wide frequency range. It can also be set up for rapid, routine, production tests on transistors, vacuum tubes, amplifiers, or networks, and a high degree of skill or knowledge on the part of the operator is not required. Several specific applications, with results of measurements, are described later in this article.

TYPES OF TRANSFER FUNCTIONS

For purposes of this discussion, four-terminal networks are considered to have separate, but grounded, input and output connections, and three-terminal networks are considered to have one terminal common to both input and output. There are several systems of parameters used to describe the electrical performance of these types of networks:

A. Open-circuit impedance parameters, which are the input and output impedances, Z_{11} and Z_{22}, and the forward and reverse transimpedances, Z_{21} and Z_{12}.

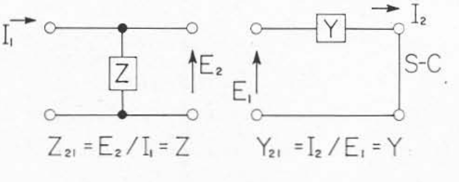
B. Short-circuit admittance parameters, which are the input and output admittances, Y_{11} and Y_{22}, and the forward and reverse transadmittances, Y_{21} and Y_{12}.

C. Hybrid combinations (such as the *h* parameters often used in transistor work) of impedance, admittance, open-circuit voltage-ratio, and short-circuit current-ratio parameters. For convenience, we define symbols to represent these voltage and current ratios as either *E* or *I* with appropriate subscripts, such as E_{21} to mean E_{out}/E_{in} with output open-circuited.

The Transfer-Function Meter can measure all of these functions.

[1]Originally described in a paper presented at the 1956 IRE Annual Convention and subsequently published in the 1956 IRE Convention Record, Part 5, pp. 3-7: "A Transadmittance Meter for VHF-UHF Measurements," by William R. Thurston. The name of the instrument has been changed to Transfer-Function Meter so as to indicate more completely its uses.

[2]Grounded, two-terminal impedance / Ungrounded, two-terminal admittance



THEORY OF OPERATION

To measure a transfer function of a network, it is necessary to supply to it an input driving signal and to measure the resulting output signal in terms of the input signal. It is also necessary to terminate the network output in an open circuit if the desired output signal is a voltage, or in a short circuit if the desired output signal is a current. If the network were terminated otherwise, the answer obtained would depend on the network output impedance or admittance as well as on its transfer functions[3] and would, consequently, be less useful for general calculations.

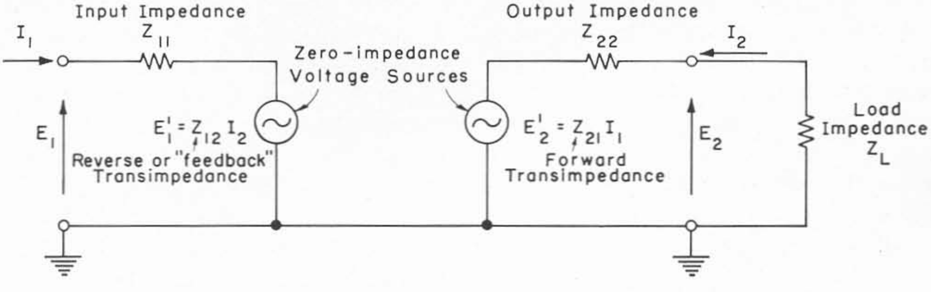
Nevertheless, there are undoubtedly applications where one wishes only to determine the over-all performance of a network working into a *specific* load impedance. In these cases, it is necessary to include the termination as a part of the network under test. Where output current is of interest, the termination must be in series, and where output voltage is of interest, the termination must be in shunt.

Forward transfer functions are, of course, measured by driving the normal input terminals of a network under test, while reverse, sometimes called "feedback," transfer functions are measured by reversing the network and driving the normal output terminals. Two-terminal impedances or admittances can be connected as four-terminal networks in accordance with the diagrams given in Footnote 2, depending on whether a grounded or an ungrounded element is to be measured.

In the Transfer-Function Meter there are three identical loops, as shown in Figure 1, driven in parallel by an external generator adjusted to the desired frequency of measurement. The currents, I_L, in all three loops are equal in magnitude and phase. Each loop is loosely coupled, through electrostatically shielding slots, to an associated coaxial line. In Figure 1, only the inner conductors of these lines are shown. Each loop can be rotated independently of the others so as to vary its coupling, or mutual inductance, to its associated line. The mutual inductances are designated M_G, M_B, and M_X. The series voltages induced in the three lines by virtue of the couplings to the associated loops are: $E_G = -j\omega M_G I_L$, $E_B = -j\omega M_B I_L$, and $E_X = -j\omega M_X I_L$.

The outer end of the left-hand line, called the G line, is terminated in a

[3]Example: Equivalent circuit using impedance parameters:



Measured forward transimpedance =

$$\frac{E_2}{I_1} = \frac{E'_2/I_1}{1+\frac{Z_{22}}{Z_L}} = \frac{Z_{21}}{1+\frac{Z_{22}}{Z_L}} \cong Z_{21}, \text{ if } Z_L \gg Z_{22}$$

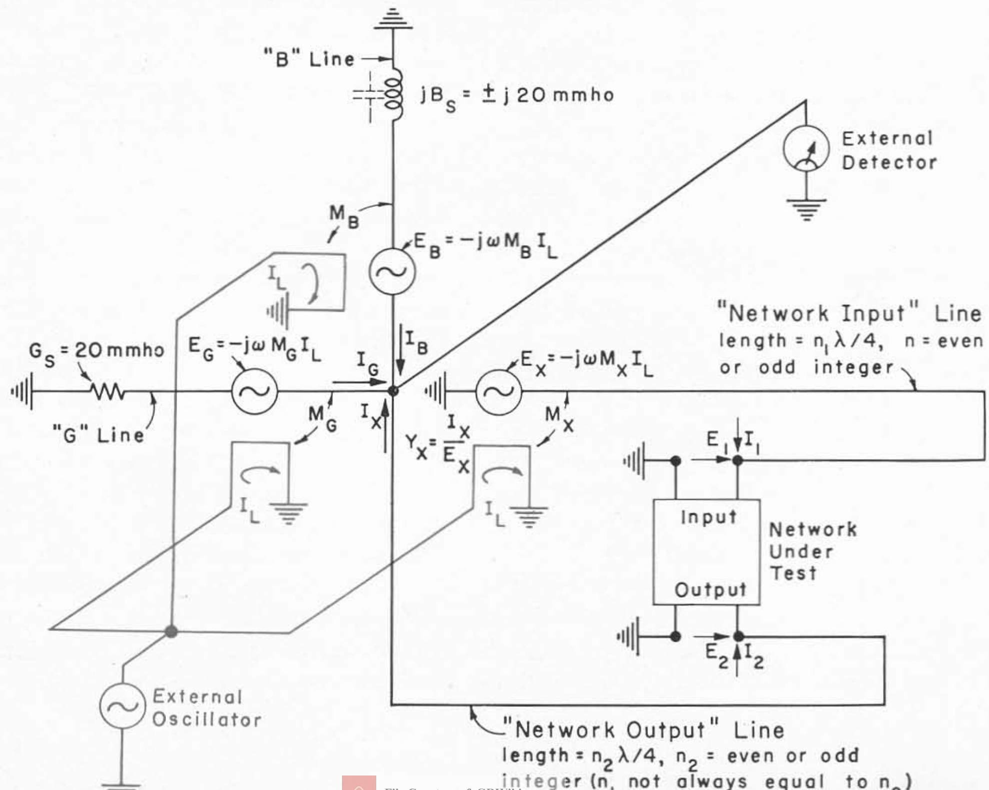
From circuit and equations given, it is seen that the measured transimpedance equals the value of Z_{21} only if the load impedance Z_L is very large compared to the network output impedance, Z_{22}. Otherwise the measured value is in error, and the error can be in phase angle, magnitude, or a combination of both, depending on the phase angles of Z_L and Z_{22}.

known, standard conductance, Y_0 (20 millimhos). The characteristic admittance of the coaxial lines and TYPE 874 Connectors used in the instrument and associated components is also equal to 20 millimhos (characteristic impedance, Z_0, is 50 ohms). The outer end of the upper line, called the B line, is terminated in a known, standard susceptance of $+jY_0$ at frequencies below 150 Mc (adjustable capacitor), $-jY_0$ between 150 Mc and 450 Mc (adjustable stub set to $\frac{\lambda}{8}$), and $+jY_0$ above 450 Mc (stub set to $\frac{3\lambda}{8}$). The far end of the right-hand line, which is adjustable in length and is called the *Network Input* line, is connected to the input of the network under test, and its electrical length is always set to equal either an odd or an even multiple, n_1, of a quarter wavelength, depending on which type transfer function is to be measured.

The near end of the *Network Input* line terminates in a short circuit. The inner ends of the B and G lines come together in a junction with two other lines not previously mentioned, as shown in Figure 1. One of these latter lines is connected to an external detector. The other, which is adjustable in length and is called the *Network Output* line, is connected to the output of the network under test. Its electrical length is always set to equal either an odd or an even multiple, n_2, of a quarter wavelength, depending on which type transfer function is to be measured, but not necessarily the same multiple as that to which the *Network Input* line is set.

The process of measuring complex quantities involves the balancing of the instrument by adjustment of the loop

Figure 1. Schematic diagram of r-f circuits of the Transfer-Function Meter.



couplings until the external detector indicates a null condition. At null, the voltage at the junction of the four coaxial lines is zero, and the three currents, I_G, I_B, and I_X, that enter the detector junction from, respectively, the G, B, and *Network Output* lines, must add up to zero. These line currents are readily calculated, because, for this purpose, the zero-voltage condition at the detector junction can be considered equivalent to a short circuit. For the purpose of simplifying the explanation, the lengths of the *Network Input* and *Network Output* lines will first be assumed to be zero. Under these conditions, $E_1 = E_X$ and $I_2 = I_X$.

The current, I_G, equals the induced voltage, E_G, times the admittance of the G line, which is the known, standard conductance, Y_0. That is,

$$I_G = Y_0 E_G = Y_0(-j\omega M_G I_L)$$
$$= Y_0 M_G(-j\omega I_L)$$

The current, I_B, equals the induced voltage, E_B, times the admittance of the B line, which is the known, standard susceptance, $\pm jY_0$. That is,

$$I_B = \pm jY_0 E_B = \pm jY_0(-j\omega M_B I_L)$$
$$= \pm jY_0 M_B(-j\omega I_L)$$

The current, I_X, equals the product of the induced voltage in the *Network Input* line, E_X, and the transadmittance of the network, Y_X. Therefore,

$$I_X = Y_X E_X = Y_X(-j\omega M_X I_L)$$
$$= Y_X M_X(-j\omega I_L)$$

When the sum of I_G, I_B, and I_X is equated to zero, which is the balance condition, the common $-j\omega I_L$ term is eliminated, and the basic balance equation for the instrument is obtained:

$$\frac{Y_X}{Y_0} = \frac{M_G}{M_X} \pm j\frac{M_B}{M_X}$$

The above equation is normalized with respect to the characteristic admittance of the line and corresponds to the dial calibration, which is normalized because impedances as well as admittances must be measured. As indicated above, the instrument actually measures the real and imaginary parts of the normalized transadmittance, $\frac{G_X}{Y_0}$ and $\frac{B_X}{Y_0}$, of the network connected directly between the input and output terminals of the instrument:

$$\frac{G_X}{Y_0} = \frac{M_G}{M_X}$$

$$\frac{B_X}{Y_0} = \frac{M_B}{M_X}$$

Since the connecting line lengths are assumed to be zero, $Y_{21} = Y_X$, $G_{21} = G_X$, and $B_{21} = B_X$. The mutual inductance, M_X, is the denominator in both the above equations and hence is a common multiplier. The values of the mutual inductances, M_X, M_G, and M_B, depend on the angular positions of the loops and hence can be adjusted from zero to a maximum value by rotation of the loops. The angular position of the G loop can therefore be calibrated directly in normalized transconductance, the B loop in normalized transsusceptance, and the X loop in a common multiplier. Figure 2 shows these calibrations, which are independent of frequency and which, by virtue of the positive and negative ranges for two of the three loops, allow measurements to be made in all four quadrants of the complex plane. The scale associated with the G loop is labeled the A scale and is calibrated from 0 to 1.5. The scale associated with the B loop is labeled the B scale and is calibrated from 0 to $\pm$ 1.5. The multiplier is calibrated from $\pm$ 1 to infinity.

The assumption of zero length of lines between the instrument and network made in the preceding analysis cannot be realized in practice, since the effective measurement points are located *within* the instrument. However, by the adjustment of the *Network Input* and *Network Output* lines to odd or even multiples of a quarter wavelength, the instrument can be made to indicate directly the transadmittance, transimpedance, complex transfer current ratio, and complex transfer voltage ratio of networks whose terminals are *not* directly the actual measurement terminals of the instrument. Each of the above measurements requires a different setting of the *Network Input* and *Network Output* lines and will be considered in detail in the following paragraphs.

In the following discussion, the term "half-wave setting" means that the line in question is set to an *even* multiple of a quarter wavelength, which is, of course, always a multiple of a half wavelength. A half-wave line has the property of "repeating" at one end all voltages, currents, and impedances appearing at the other end with 180 degrees of phase shift in voltages and currents for each half wavelength. Similarly, the term "quarter-wave setting" means that the line in question is set to an *odd* multiple of a quarter wavelength. A quarter-wave line has the property of "inverting" voltages into currents, impedances into admittances, and vice versa. The reversal of phase which occurs for each added half wavelength will be ignored, since it does not affect the basic theory of operation.

Transadmittance, Y_{21} and Y_{12}

The forward transadmittance of a network with its output terminals short-circuited is Y_{21}. In order to measure this parameter, the *Network Input* and *Network Output* lines are both adjusted to a half wavelength. Under these conditions the output terminals of the network under test are effectively short-circuited, because the half-wave *Network Output* line terminates at the detector junction, which under null conditions has zero voltage and can be considered to be a short circuit. The half-wave line produces a similar short circuit at the network terminals and makes $I_2 = I_X$. The input half-wave line makes $E_1 = E_X$. Therefore,

$$\frac{Y_{21}}{Y_0} = \frac{I_2/E_1}{Y_0} = \frac{I_X/E_X}{Y_0} = \frac{Y_X}{Y_0} = \mathrm{A} + \mathrm{jB}$$

where A and B are the *A* and *B scale* readings.

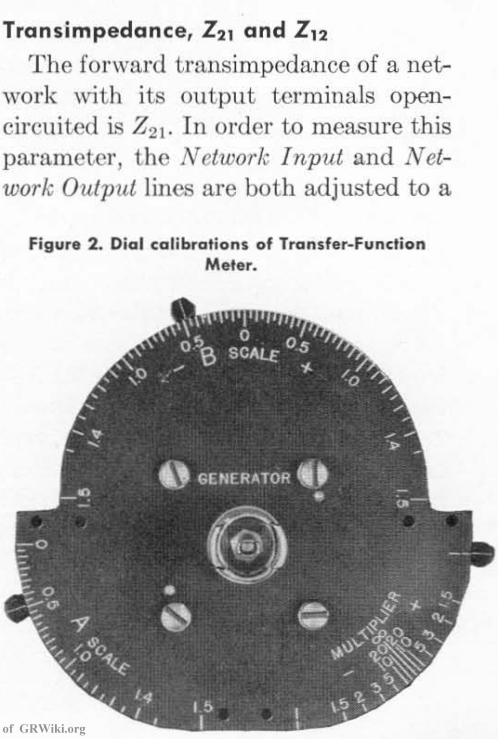
As previously shown, the instrument directly measures the normalized, real and imaginary components of Y_X, and from the above equation it is evident that it also indicates $\frac{G_{21}}{Y_0}$ and $\frac{B_{21}}{Y_0}$.

The reverse transadmittance, Y_{12}, can be measured by the same procedure as indicated for the forward transadmittance but with the input and output connections of the network interchanged.

Transimpedance, Z_{21} and Z_{12}

The forward transimpedance of a network with its output terminals open-circuited is Z_{21}. In order to measure this parameter, the *Network Input* and *Network Output* lines are both adjusted to a

Figure 2. Dial calibrations of Transfer-Function Meter.



quarter wavelength. Under these conditions the output terminals of the network under test are effectively open-circuited, because the quarter-wave *Network Output* line inverts the equivalent short circuit at the detector junction into an open circuit at the network. Also, the output quarter-wave line "inverts" the voltage E_2 into a constant times the current I_X, and the input quarter-wave line "inverts" the voltage E_X into a constant times the current I_1. It can be shown that

$$\frac{Z_{21}}{Z_0} = \frac{E_2/I_1}{Z_0} = \frac{I_X/E_X}{Y_0} = \text{A} + \text{jB}$$

where Z_0 is the characteristic impedance of the coaxial lines, 50 ohms. Thus the instrument reads directly the normalized transimpedance of the network under test. The readings are in terms of the normalized network transresistance, $\frac{R_{21}}{Z_0}$, read on the A scale, and the normalized transreactance, $\frac{X_{21}}{Z_0}$, read on the B scale.

Reverse transimpedance, Z_{12}, is measured in a similar manner with the input and output network connections reversed.

Transfer Current Ratio, I_{21} and I_{12}

The forward transfer current ratio of a network with its output terminals short-circuited is I_{21}. For this measurement the *Network Output* line is adjusted to a half wavelength and the *Network Input* line to a quarter wavelength. The output terminals of the network under test are effectively short-circuited, because the half-wave *Network Output* line "repeats" the equivalent short circuit at the detector junction as a short circuit at the network. The half-wave line also makes $I_2 = I_X$. The quarter-wave *Network Input* line makes $E_X = \frac{jI_1}{Y_0}$. Therefore,

$$I_{21} = \frac{I_2}{I_1} = \frac{jY_X}{Y_0} = \text{B} + \text{jA}$$

Thus the instrument reads directly the real and imaginary components of the complex transfer current ratio of the network. The "j" term in the above equation interchanges the real and imaginary scales.

The reverse transfer current ratio, I_{12}, can be measured by reversing the input and output connections to the network.

Transfer Voltage Ratio, E_{21} and E_{12}

The forward transfer voltage ratio, E_{21}, is measured with the network output terminals open-circuited. In this case the *Network Output* line is adjusted to a quarter wavelength and the *Network Input* line to a half wavelength The output terminals of the network are effectively open-circuited because the quarter-wave *Network Output* line "inverts" the equivalent short circuit at the detector junction into an open circuit at the network. Also, because of the quarter-wave *Network Output* line, $E_2 = \frac{jI_X}{Y_0}$, and because of the half-wave *Network Input* line, $E_1 = E_X$. Therefore,

$$E_{21} = \frac{E_2}{E_1} = \frac{jY_X}{Y_0} = \text{B} + \text{jA}$$

Here again, the instrument indicates the complex open-circuit transfer voltage ratio of the network under test with the real and imaginary component scales interchanged from those used for transadmittance measurements because of the "j" term in the above equation.

Reverse transfer voltage ratio, E_{12}, can be measured by reversing the input and output connections to the network.

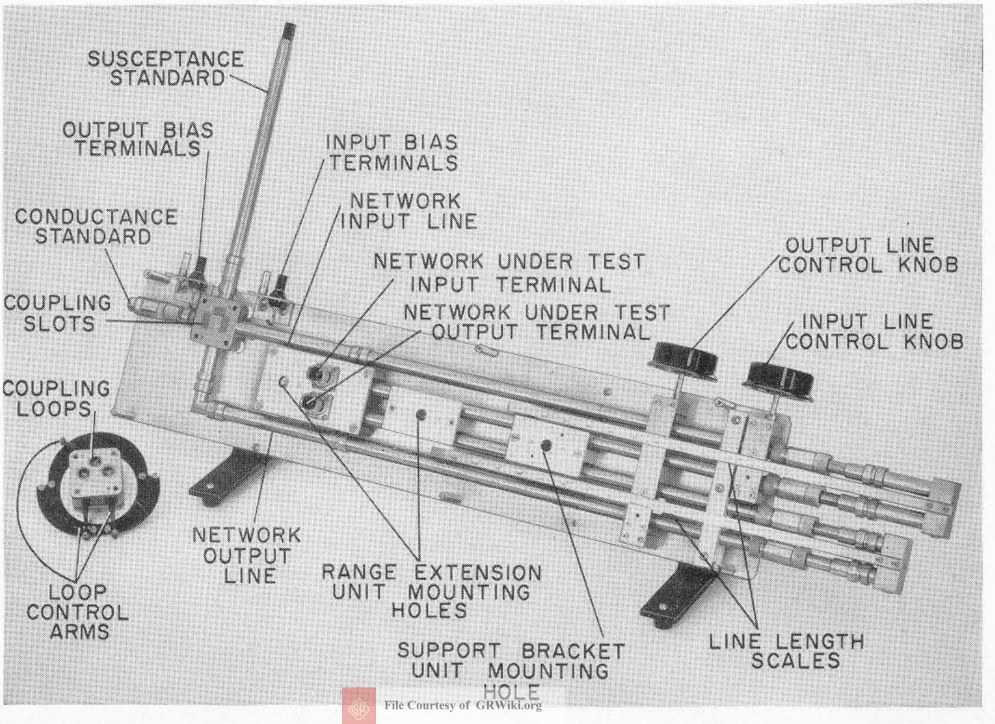
PHYSICAL DESCRIPTION

The physical arrangement of the parts of the Transfer-Function Meter corresponds closely to that shown in the schematic of Figure 1 and is illustrated further in Figure 3, in which the cover of the instrument has been removed and the coupling-loop assembly dismounted. At the left in Figure 3 can be seen the main junction block, in which coupling slots are cut into the coaxial lines within the block. When in place, each of the three loops in the coupling-loop assembly is centered over its respective slots and is coupled magnetically to the corresponding line. The *Network Input* and *Network Output* lines are of the constant-impedance, "trombone" type, driven independently by separate, rack-and-pinion drives having accurately calibrated scales to indicate total effective line lengths directly in cm. The lines are provided with locking sleeves to prevent accidental changes during prolonged work at a single frequency. All these parts are mounted on a heavy aluminum base plate.

In the measurement of active devices, especially transistors, it is important to keep the applied signal level low. In this instrument, the coupling loss of the loop between the external generator and the device under test is about 40 db at 500 Mc and decreases at a rate of 6 db per octave with increasing frequency. For tests on transistors, in which signal levels should be 5 millivolts or less, appropriate attenuators (874-G series) should be used to reduce the level of the signal supplied by the generator when necessary.

Since the external detector is usually of the heterodyne type with a local oscillator, it is important to prevent excessively high local-oscillator signals from appearing at the terminals of the unknown device. This problem is solved by the insertion of a tuned stub, or "trap," in parallel with the detector in-

Figure 3. The Transfer-Function Meter partially disassembled to show details of design and construction.



put and tuned to reject the local-oscillator frequency. This stub is supported horizontally behind the base plate of the instrument.

In measurements on active networks, d-c voltages or currents must be supplied without affecting the r-f circuits. In the Transfer-Function Meter, provisions are included for applying dc to both the input and the output of the network under test. The binding posts for connection to external power supplies are visible in Figure 3, and the internal filters and blocking capacitors are shown schematically in Figure 4. Built-in blocking capacitors isolate the measurement standards, the external detector, and the short circuit on the *Network Input* line. Filter networks, each comprising two chokes and two by-pass capacitors, allow insertion of d-c voltages and currents and prevent r-f leakage. Choke and capacitor ratings limit currents to 100 milliamperes and voltages to 400 volts. The loading effect of the input filter on the *Network Input* line is negligible, because of its proximity to the short-circuited end of the line. The only loading effect of the output filter on the detector line is a small reduction in detector sensitivity.

The range of adjustment of the *Network Input* and *Network Output* lines is such as to allow continuous coverage for all types of measurements above 300 Mc, plus separated bands of coverage below 300 Mc. In order to allow continuous coverage below 300 Mc, a set of extension lines is provided. When needed, these lines and their supports can be snapped into place by means of quarter-turn fasteners, as shown in Figure 5. This photograph also shows the shielded, variable capacitor used as the susceptance standard at low frequencies in place of the stub used at high frequencies.

Generator and Detector

General Radio Unit Oscillators are recommended for use as generators with the Transfer-Function Meter. The recommended detector is the General Radio TYPE DNT, a heterodyne type that combines high sensitivity with wide frequency range. Both generators and detectors are listed on pages 15 and 16.

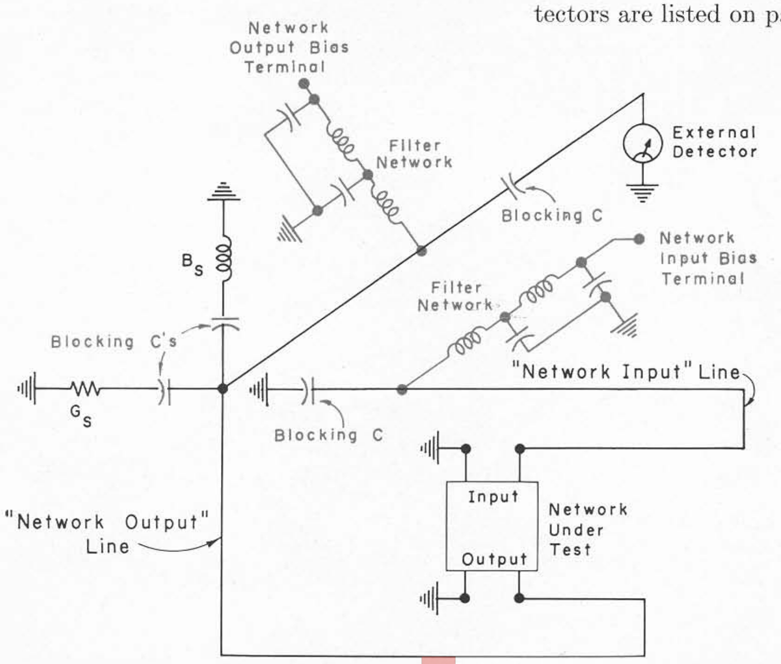


Figure 4. Schematic diagram of d-c circuits of the Transfer-Function Meter.

MEASUREMENT PROCEDURE

The equipment is set up by connecting generator, detector, and d-c supplies, if needed, to the Transfer-Function Meter, and making the necessary adjustments for desired operating frequency and d-c levels. The calibrated susceptance standard is also set to the operating frequency.

If isolation of the local-oscillator signal is desirable, as in measurements on transistors, the "trap" stub is included in the setup and is adjusted for maximum attenuation of the local-oscillator voltage.

Next, the *Network Input* and *Network Output* lines are set to the proper length, in accordance with the type of transfer function to be measured. An appropriate component mount is plugged into the *Network Under Test* connectors, and the unknown device or network is plugged into the mount. The three loop-control arms are then adjusted until the detector indicates a null, and the desired answer is read directly from the scale settings.

If several units of the same type are to be checked at a given frequency, as in the case of production testing of transistors or tubes, each unit successively is plugged into the mount (with due precautions regarding the d-c supplies), the control arms are set for a null, and the answer is read off the scales. This operation can be performed very rapidly by relatively unskilled personnel.

Terminals

The terminals used on the Transfer-Function Meter are TYPE 874 Coaxial Connectors. General Radio oscillators and detectors are also equipped with these terminals. When generators and detectors having other types of terminals are used, the TYPE 874-Q series of adaptors provides a convenient means of connection.

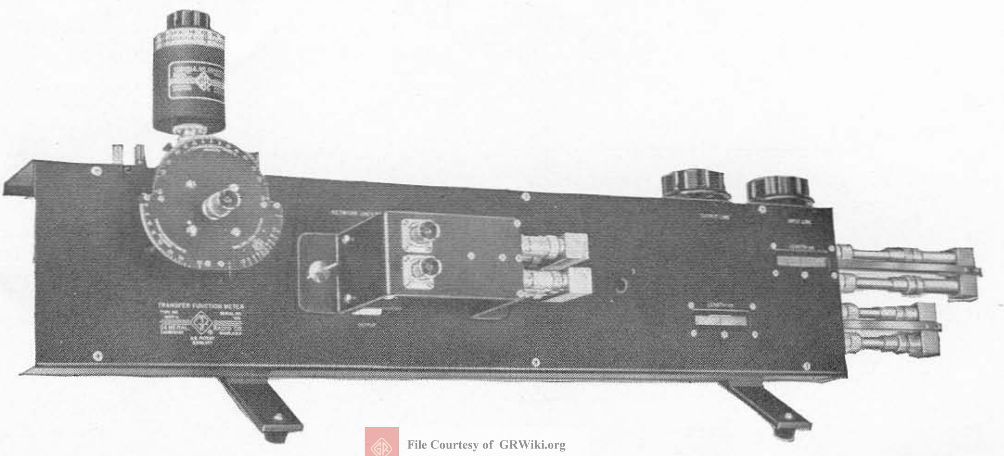
For most types of measurement, suitable mounts must be constructed (see Transistor Measurements, below) to connect the device being measured to the measuring terminals. TYPE 874 Coaxial Connectors to fit rigid line, panel, and cable are available for building into these mounts.

Both adaptors and connectors are listed on page 16.

Sources of Error

The major sources of error are incidental losses and small reflections in the *Network Input* and *Network Output* lines. The minor sources of error are similar to

Figure 5. View showing Range Extension Unit and low-frequency susceptance standard.



those in the Type 1602-B Admittance Meter[4, 5] and are spurious cross-couplings between the coupling loops and their associated lines, inductances between the junction center and the coupling points, incidental losses in the susceptance standard, and small reflections in the conductance standard.

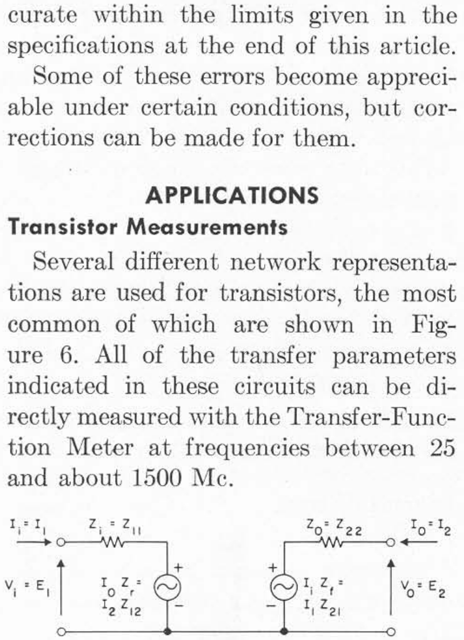
In most measurements, the instrument dial readings can be used directly without any corrections and will be accurate within the limits given in the specifications at the end of this article.

Some of these errors become appreciable under certain conditions, but corrections can be made for them.

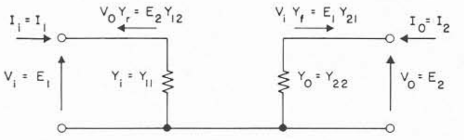
APPLICATIONS

Transistor Measurements

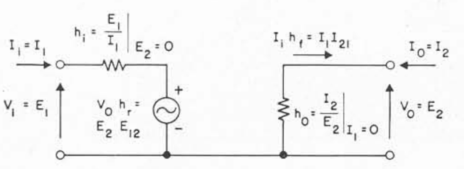
Several different network representations are used for transistors, the most common of which are shown in Figure 6. All of the transfer parameters indicated in these circuits can be directly measured with the Transfer-Function Meter at frequencies between 25 and about 1500 Mc.



A. OPEN-CIRCUIT IMPEDANCE PARAMETERS



B SHORT-CIRCUIT ADMITTANCE PARAMETERS



C. HYBRID PARAMETERS

Since many transistors operate at very low voltage levels, it is important that all applied signals be kept small during the measurements. As previously mentioned, the r-f signal level can be held below 5 mv, which has been found to be a satisfactory limit.

The high-frequency performance of a transistor or any other component can be greatly affected by the arrangement of the leads used to connect the element in a circuit. Therefore, for reproducible results, the mount used to connect the transistor to the measurement circuit must be standardized and must permit short leads. In a typical experimental transistor mount, the leads are connected to the measuring elements at a point about 1/32″ away from the case. This arrangement probably is reasonably close to that used in most practical high-frequency transistor circuits.

For measurement of the complex current ratios, α (or $-h_f$), the *Network Input* line is set to a quarter wavelength and the *Network Output* line to a half wavelength, as outlined in a previous paragraph. The local-oscillator trap is adjusted, with interchange of the generator and detector connections, by adjustment of the stub line until minimum output is observed on the meter of the detector. The normal connections are restored and the transistor mount plugged into the coaxial connectors on the panel of the instrument. The meter is then balanced by adjustment of the

[4]Thurston, W. R., "A Direct-Reading Impedance-Measuring Instrument for the UHF Range," *General Radio Experimenter*, May, 1950.

[5]Soderman, R. A., "Improved Accuracy and Convenience of Measurements with Type 1602-B Admittance Meter in VHF and UHF Bands," *General Radio Experimenter*, August, 1953.

Figure 6. Equivalent network representations of transistors. Left-hand set of symbols is from "IRE Standards on Electron Devices; Methods of Testing Transistors," *Proc. IRE,* Vol. 44, pp. 1542-1561, November, 1956. Right-hand set of symbols corresponds to those of this article.

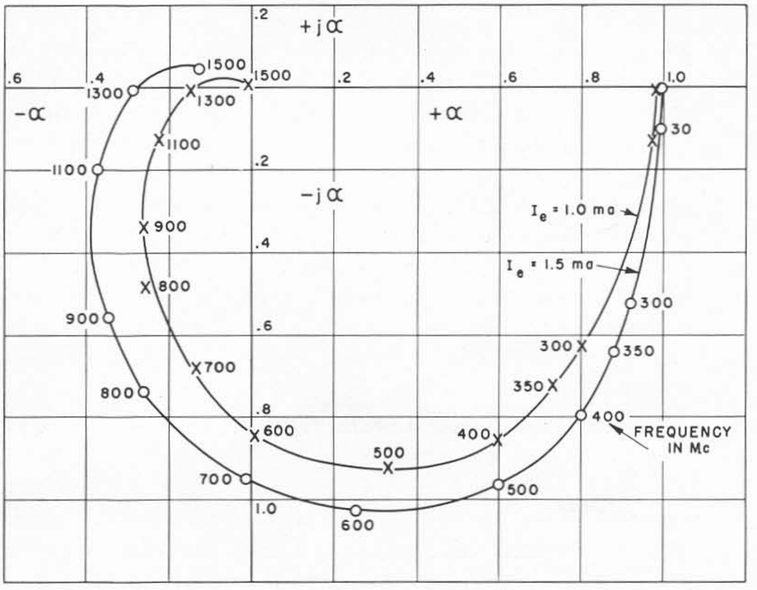
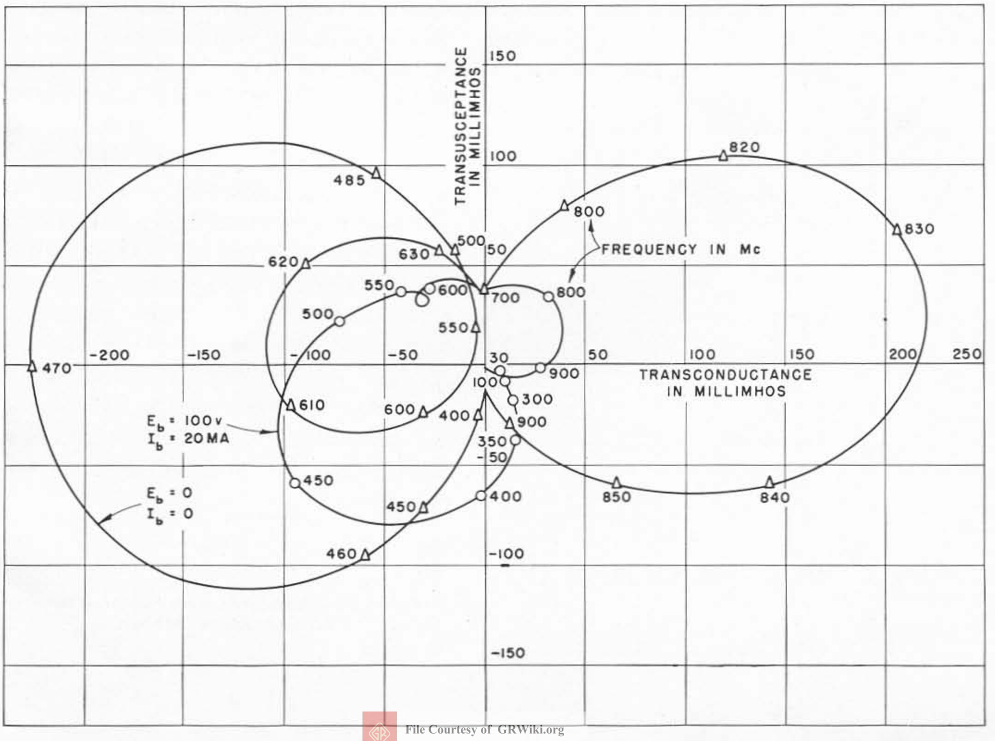


Figure 7. Plot of alpha versus frequency for an experimental high-frequency transistor supplied by Bell Telephone Laboratories.

three arms, and the real and imaginary components of the current ratio are indicated directly on the dial scales. The α-vs.-frequency characteristics of an experimental, Bell Telephone Laboratories, diffused-base, germanium transistor in a grounded-base connection are shown in Figure 7.

The hybrid feedback factor, $h_r (= E_{12})$, can be easily measured by reversal of the coaxial-line connections between the grounded-base mount and the instrument, which is accomplished by 180° rotation of the mount, reversal of the

Figure 8. Transadmittance versus frequency for a grounded-cathode 6AF4 vacuum tube.



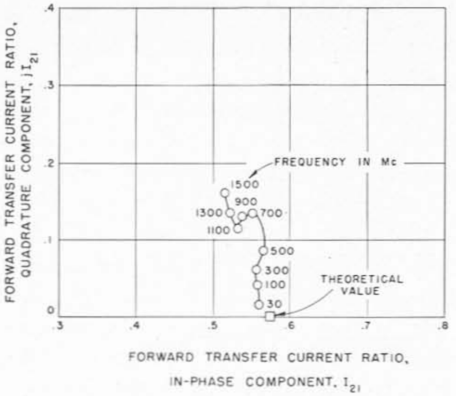


Figure 9. Forward transfer current ratio versus frequency for a Type 874-G10 Attenuator Pad.

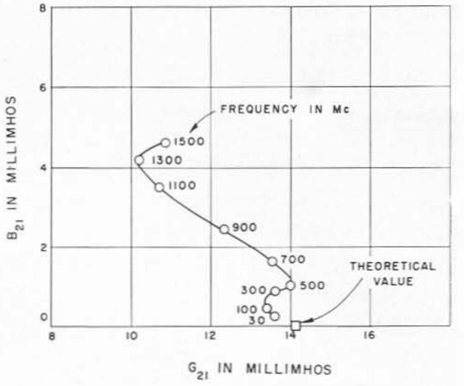


Figure 10. Forward transadmittance versus frequency for a Type 874-G10 Attenuator Pad.

d-c connections, and use of the procedure outlined for voltage-ratio measurements. Under these conditions the input is applied to the collector and the output is obtained from the emitter.

Tube Measurements

The high-frequency, complex, forward and reverse transadmittances of vacuum tubes can also be easily measured under dynamic conditions with the Transfer-Function Meter. D-C plate and bias voltages can be applied to the input and output terminals in the same manner as with transistors. Filters must be provided in the tube mount for heater and screen voltages. However, these filters are not so critical as are the filters associated with the input and output circuits. As with transistors, the mount must be carefully designed in order to give significant and reproducible results. The measured transadmittance of a 6AF4 in the grounded-cathode connection is plotted in Figure 8. The effective transadmittance first increases with frequency, apparently as a result of a resonance between the grid-cathode capacitance and cathode-lead inductance. At higher frequencies, other resonances are apparent, the largest one of which is probably a result of the grid-plate capacitance and plate lead inductance resonance. The large values of transadmittance shown do not result in correspondingly large magnitudes of gain when this tube is used in an amplifier, since the input impedance decreases rapidly as the resonances are approached.

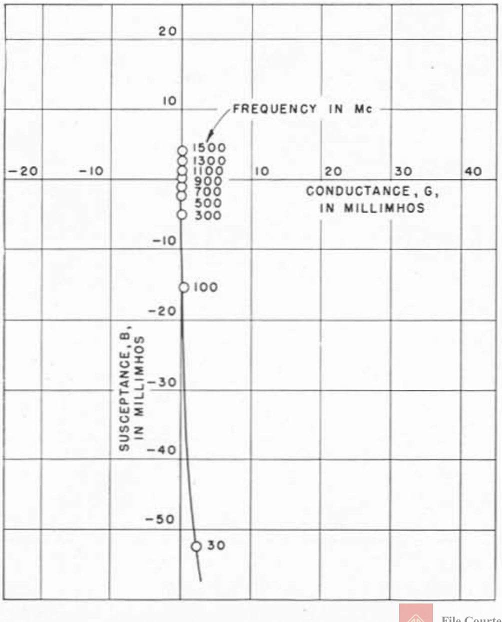


Figure 11. Admittance versus frequency for a 0.1-michrohenry inductor.

Coaxial-Component Measurements

The Transfer-Function Meter can measure the transfer admittance or impedance and attenuation of circuits fitted with coaxial connectors. Figure 9 shows the results of short-circuit, current-ratio measurements made on a Type 874-G10 Attenuator Pad, and Figure 10 shows transadmittance of the same pad. Other possible applications are for filters, coupling networks, amplifiers, and other four-terminal coaxial devices.

Lumped-Component Measurements

The direct admittance between the two ungrounded terminals of a circuit or component can be easily measured with this instrument. As is shown in the diagram in Footnote 2, the direct admittance measurement of a component with neither end connected to ground is not affected by the impedance from either side of the element to ground. This measurement is very useful for determining the characteristics of floating resistors, r-f chokes, capacitance between two ungrounded terminals, and many other three-terminal circuits. Figure 11 shows the direct admittance of one of the chokes used in the d-c supply filter in the Transfer-Function Meter.

—W. R. Thurston
R. A. Soderman

SPECIFICATIONS

Frequency Range: 25 to 1000 Mc, with reduced accuracy up to 1500 Mc.

Measurement Ranges and Accuracy:

Range	*Accuracy*
Voltage and Current Ratios (R) 0-30	$2.5\ (1+\sqrt{R})\%+0.025$
Transimpedance (Z_{21}) 0-1500Ω	$2.5\ (1+\sqrt{\frac{\lvert Z_{21}\rvert}{50}})\%+1.25\Omega$
Transadmittance (Y_{21}) 0-600 mmhos	$2.5\,(1+\sqrt{\frac{\lvert Y_{21}\rvert}{20}})\%+0.5\text{ mmho}$

D-C Bias: Terminals are provided for introducing d-c bias from external sources. Maximum bias current, 100 ma; maximum bias voltage, 400 volts.

Generator and Detector: Unit Oscillators and Type DNT Detectors (see list below) are recommended.

Other Accessories Available:
Type 1607-P101 Transistor Mount for JETEC-30 base arrangement, grounded base.
Type 1607-P201 Tube Mount, 7-Pin Miniature grounded-cathode, for 6AF4, 6AF4A, and other tubes with same connections.

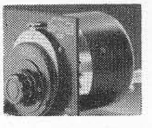
Case: The instrument, with accessories, is mounted in a wooden carrying and storage case 11¼ x 14½ x 40 inches.

Net Weight: 63 pounds.

Type		*Code Word*	*Price*
1607-A	**Transfer-Function Meter**	HYDRA	**$1525.00**
1607-P101	**Transistor Mount**		**Price on Request**
1607-P201	**Tube Mount**		**Price on Request**

GENERATORS*

Type	*Frequency Range*	*Code Word*	*Price*
1211-B	**0.5 — 50 Mc**	ATLAS	**$275.00**
1215-B	**50 — 250 Mc**	ADOPT	**190.00**
1209-BL	**180 — 600 Mc**	ADMIT	**245.00**
1209-B	**250 — 920 Mc**	AMISS	**245.00**
1218-A	**900 — 2000 Mc**	CARRY	**465.00**



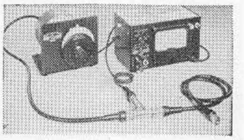
*Require power supply below.

POWER SUPPLY*

Type		*Code Word*	*Price*
1203-B	**Unregulated**	ALIVE	**$40.00**
1201-A	**Regulated**	ASSET	**85.00**

*Plug-in type; supplies power to any one of the above oscillators.

DETECTORS

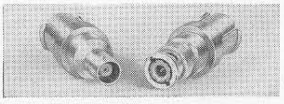


Type	*Frequency Range**	*Code Word*	*Price*
DNT-1	**40 — 530 Mc...**	NALTO	**$626.00**
DNT-2	**40 — 280 Mc...**	NERVO	**606.00**
DNT-3	**220 — 950 Mc...**	NULLO	**659.00**
DNT-4	**870 — 2030 Mc...**	NODDO	**879.00**

*Fundamental range. To cover a wider range than that listed for any one detector, harmonics of the local oscillator can be used. Thus Type DNT-2 will cover frequencies up to 1120 Mc if harmonics up to the 4th are used. Fundamental sensitivity is about 5 μv; 4th harmonic sensitivity, about 20μv. For this wide range, order also one Type 874-F1000 Low Pass Filter, price $14.00.

Another solution is to use the Type DNT-2 with an additional Type 1209-B Unit Oscillator (see Generators, above) and the Type 874-F1000 Filter. This covers the range with fundamental operation, which is, in general, more satisfactory. Harmonic operation is not recommended for measurement of active networks. Below 40 Mc, use a communications receiver.

COAXIAL ADAPTORS

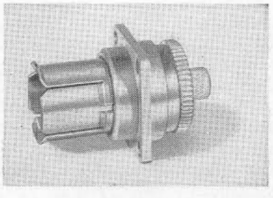


Types QBJ and QBP

Type	*Fits*	*Code Word*	*Price*
874-QBJ	**BNC Plug........**	COAXBOGGER	**$4.75**
874-QBP	**BNC Jack........**	COAXBUNNER	**4.75**
874-QCJ	**C Plug...........**	COAXCOGGER	**4.75**
874-QCP	**C Jack...........**	COAXCUFFER	**6.25**
874-QNJ	**N Plug...........**	COAXNAGGER	**3.75**
874-QNP	**N Jack...........**	COAXNUTTER	**4.50**
874-QUJ	**UHF Plug........**	COAXYUNDER	**4.00**
874-QUP	**UHF Jack........**	COAXYUPPER	**4.25**
874-QHJ	**HN Plug ....**	COAXHAWSER	**6.50**
874-QHP	**HN Jack ....**	COAXHANGER	**6.50**
874-QLJ	**LC Plug ....**	COAXLITTER	**19.50**
874-QLP	**LC Jack ....**	COAXLUGGER	**30.00**

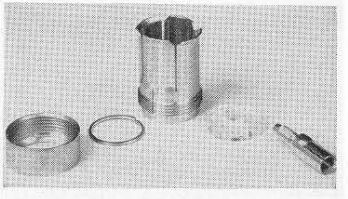
Adaptors are also available for connection to rigid air-dielectric lines.

COAXIAL CONNECTORS



Type 874-PB Panel Connector

Type 874-B Basic Connector, disassembled



PANEL CONNECTOR

The panel connector mounts with a flange and is available with rear fittings for commonly used RG-type cables. See the General Radio catalog for details. Connector listed below fits GR Type 874-A2 Cable, only.

BASIC CONNECTOR

The basic connector fits rigid, 50-ohm, air-dielectric coaxial line; ⅝″ OD and 9/16″ ID for outer conductor; 0.244″ rod for inner conductor.

Type		*Code Word*	*Price*
874-B	**Basic Connector**	COAXBRIDGE	**$1.25**
874-PB	**Panel Connector**	COAXAPPLER	**2.90**

CREDITS

The authors wish to express appreciation for the many helpful suggestions and other contributions of many individuals during the development of the Transfer-Function Meter, including Dr. J. M. Early and Mr. D. E. Thomas of the Bell Telephone Laboratories, and Mr. R. Wohl, Mr. S. Friedman, Mr. M. Zimet, and Mr. D. Youla of the Material Laboratory, New York Naval Shipyard.

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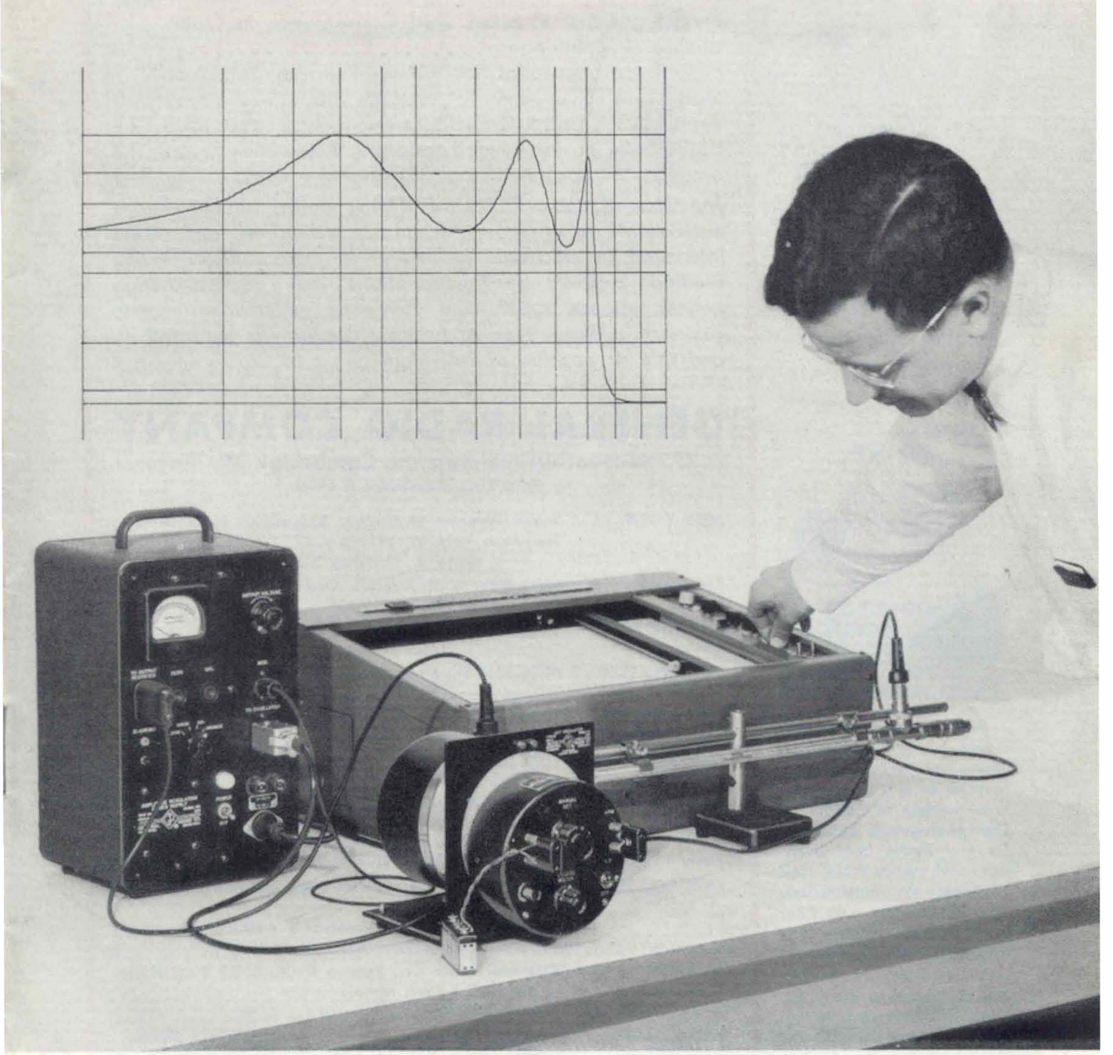

In This Issue

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R-C Oscillator
FCC-Approved TV Monitor

the GENERAL RADIO **Experimenter**

Published Monthly by the General Radio Company

VOLUME 32 · NUMBER 11 APRIL, 1958

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The General Radio EXPERIMENTER is mailed without charge each month to engineers, scientists, technicians, and others interested in electronic techniques in measurement. 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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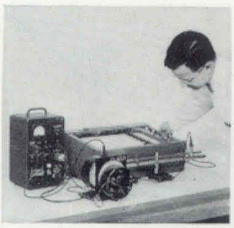


General Radio sweep drives make possible the display of data on an oscilloscope or a recorder over wide ranges of frequency. The photograph shows a General Radio Unit Oscillator with sweep drive, used to display the characteristic of a 185-Mc low pass filter on a two-axis plotter. The resulting plot, covering frequencies from 50 to 240 Mc, is shown in the inset. The horizontal and vertical coordinate lines are drawn by the plotter pen to correspond to the calibration of oscillator frequency dial and detector output.

NEW BROADCAST MONITORS ADD CONVENIENCE IN OPERATION AND MAINTENANCE

The conditions of continuous operation impose requirements on broadcasting station equipment that are more specialized and more stringent than those for laboratory equipment. In the event of failure of tubes or other components, maintenance procedures must be simple and rapid, and access to circuit elements must be as easy and direct as possible. This has long been recognized by transmitter and studio-equipment manufacturers, but for many years monitoring equipment design has lagged in its recognition of these important considerations.

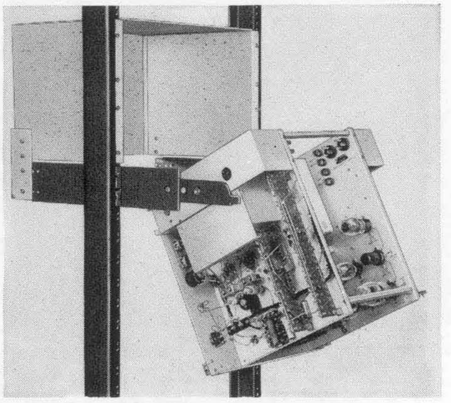


Figure 2. Type 1181-B Frequency Deviation Monitor pulled forward and tilted to give access to chassis.

The General Radio Type 1184-A Television Station Monitor[1] was the first tangible recognition of these requirements by a test-equipment manufacturer. This modern instrument, which has been type-approved by the Federal Communications Commission,[2] is designed for convenient access to all parts from the front of its rack while still in operation, and it has signal-flow lines printed on its chassis and color-coded adjustments for simple and rapid trouble shooting.

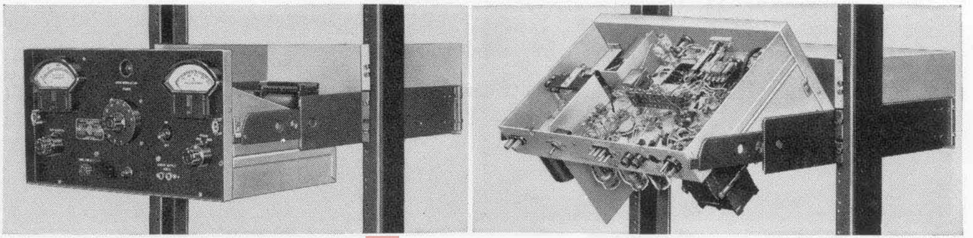
These same features have now been incorporated in the new models of the Type 1181 Frequency Deviation Monitor and the Type 1931 Modulation Monitor. At the same time, the external-meter circuits have been modified to permit the use of long telephone lines between the monitor and the external meter in remote monitoring applications.

Accessibility of chassis and components from the front of the rack is illustrated in Figure 1, which shows the Type 1931-B Amplitude Modulation Monitor pulled forward on its slides and then tilted to expose the underside of the chassis. The Type 1181-B Frequency Deviation Monitor is mounted in the same manner, as shown in Figure 2. Every operation — initial installation, operation, and maintenance — can be done from the *front* of the rack. Note that, for both instruments, as the assembly is drawn forward, the dust cover remains firmly attached to the rack. All power and other connections remain connected during these manipulations.

[1]Cady, C. A., "New Television Transmitter Monitor," *General Radio Experimenter*, Vol. 31, No. 4, September, 1956, pp. 1-10.

[2]See page 9.

Figure 1. *(Left)* Type 1931-B Amplitude Modulation Monitor pulled forward on its slides. *(Right)* Tilted to expose underside of chassis.



The new circuit for external meter connections permits the external loop resistance to be as high as 5000 ohms. A switch is provided at the rear of the monitor to disconnect the external meter and connecting lines and to substitute an internal 5000-ohm resistor, in which event the monitor functions independently of external connections. This feature aids in isolating any suspected difficulties in the external circuits. The circuit is shown in Figure 3.

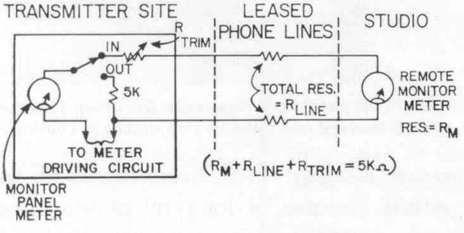


Figure 3. Circuit for external meter connections.

FREQUENCY DEVIATION MONITORS

Three models of the Type 1181 Frequency Deviation Monitor are now available, each designed for a specific type of service.

Type 1181-B Frequency Deviation Monitor 0.5 to 1.6 Mc

This model, which supersedes the Type 1181-A, is designed for use in the standard broadcast band, where, although the FCC requirements on transmitter stability are ± 20 cycles per second, transmitter frequencies must, as a practical matter, be held to a very few cycles. The crystal oscillator, temperature control, and other circuits are similar to those used in the A-model, and their stability and general reliability have been proved in over 1700 installations in all parts of the U. S. and Canada, and in countries overseas. The block diagram of Figure 4 illustrates the principle of operation.

The monitor can be used either at the transmitter site or at a location remote from the transmitter in accordance with the FCC rules permitting the unattended operation of transmitters. Remote operation up to several miles is possible with only a tuned antenna.

Type 1181-BT Color Subcarrier Monitor 3.579545 Mc

The Color Subcarrier Monitor uses circuits identical with those in the standard broadcast model, but the quartz crystal is operated at a very low oscillation amplitude in order to achieve maximum long-term stability.

In color-TV transmitters, the standard color-subcarrier-frequency signal of 3.579545 Mc is needed at any location where a color program originates, whether "live" or film. It is usually generated by an oscillator at that location, generally at or near a studio rather than at the

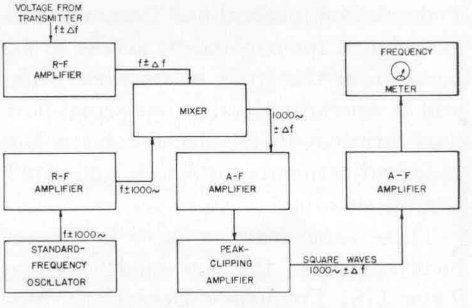


Figure 4. Block diagram of the Type 1181-B Frequency Deviation Monitor.

transmitter. Although there is no specific FCC requirement that this frequency be *monitored*, it must be held within ± 10.7 cps at all times. Good operating practice, therefore, makes continuous monitoring very desirable. The Type 1181-BT Color Subcarrier Monitor is ideally suited for this application. It indicates frequency deviation directly in cycles per second, and its stability of one cycle per month or five cycles per year obviates the necessity of frequency checks against an external source, while its price is less than half that of most counter-type frequency meters.

This monitor is intended for operation at the single frequency of 3.579545 Mc only and can be used only with an unmodulated signal input.

Type 1181-BH Frequency Deviation Monitor 1.6 to 15 Mc

Identical in its circuitry and general arrangement to the Type 1181-B, this higher-frequency model operates in a frequency range that includes such services as aeronautical, maritime, marine, public safety, and international broadcast. While the present FCC frequency-stability requirements for these services are 30 to 50 parts per million, the use of highly selective narrow-band receivers to minimize interference requires a considerably higher degree of carrier-frequency stability. The monitor is the least expensive and most reliable means of assuring the desired carrier accuracy.

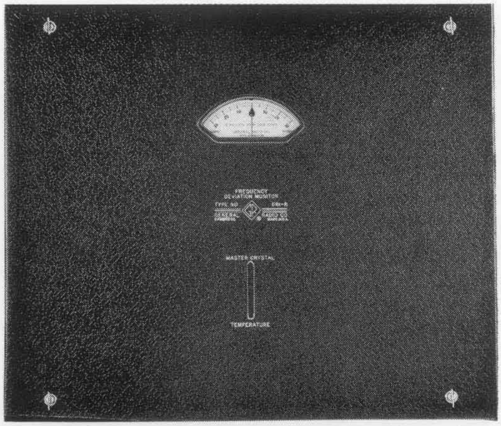


Figure 5. Panel view of the frequency deviation monitor.

General Construction

All three models are identical in construction: the Type 1181-B shown in Figures 5, 6, and 7 is typical. Figure 5 shows the front panel, from which all controls have been removed so that accidental misadjustment cannot occur. Only the indicators of frequency deviation and crystal temperature are visible.

The front panel is easily removed by means of the four fasteners shown in

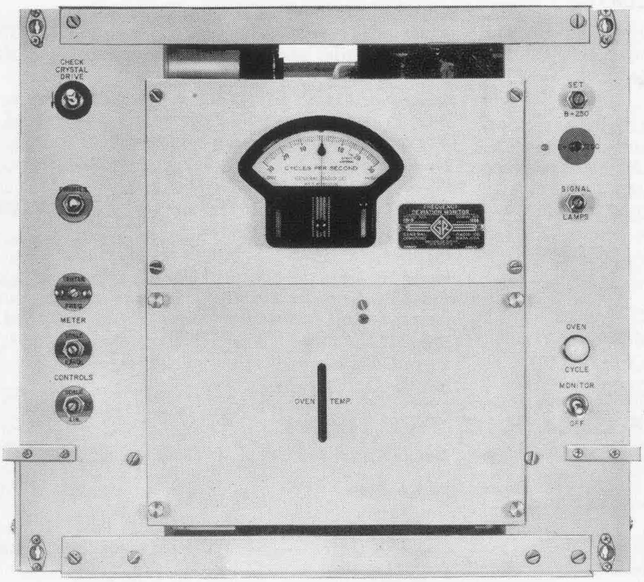


Figure 6. Monitor with dress panel removed giving access to adjustments and controls. Circles around ajustments are color coding marks.

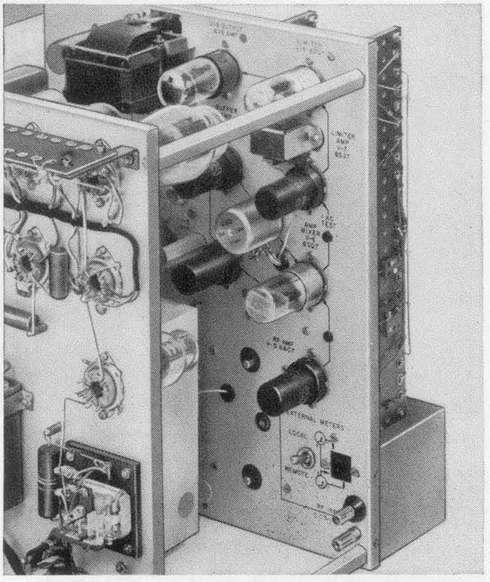


Figure 7. View of a portion of the chassis showing signal flow lines.

the corners, giving access to the adjustments and controls, as shown in Figure 6.

The inside face of the removable front panel carries a block diagram and condensed operating instructions, which, together with the signal-flow lines on the chassis, make frequent reference to the instruction book unnecessary either for operation or for maintenance. A portion of this marking is shown in the rear of Figure 7. All test points are clearly labeled, and adjustments are coded red, yellow, or green, according to their degree of importance. Thus RED means "STOP — do not change this setting without first consulting instruction book." YELLOW means "CAUTION — some external equipment (voltmeter, oscilloscope) is required to set." GREEN means "GO — can easily be set without the use of external test equipment."

In addition to these features, the ultimate in convenient accessibility is provided by the pull-forward and tilt feature shown in Figure 2.

SPECIFICATIONS, TYPE 1181-B

Deviation Range: ± 30 cycles, readable to one cycle.

Carrier Frequency Range: 500 to 1600 kc.

Accuracy: When received, within ± 5 parts per million. An adjustment is provided to bring the reading into agreement with monitoring station measurements.

Stability: Better than one part in a million under normal operating conditions for 6 months after an initial aging period. Adjustments are provided to correct the indicated frequency in terms of standard-frequency transmissions whenever necessary.

Quartz Crystal: Type 376-T.

Tube Complement:

3 — 6SJ7	1 — 5V4-G
2 — 6AC7	1 — 6B4-G
2 — 6H6	1 — OC3/VR105
2 — 6SQ7-GT	1 — 2050

Coupling to Transmitter: A few inches of wire serving as an antenna are usually sufficient. A minimum of 50 millivolts is required into a high-impedance grid circuit.

Accessories Supplied: Quartz crystal, 2 CAP-35 Power Cords, spare fuses, and plug for connecting an external meter.

Remote Indicator: External meter for local or remote deviation indication can be connected. Maximum external loop resistance: 5 KΩ.

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

Power Input: 25 watts for heater circuits, 100 watts for monitor circuits.

Mounting: 19-inch relay-rack panel.

Panel Finish: Standard General Radio black crackle. Certain standard finishes which can be processed in quantity can also be supplied.

Dimensions: Panel (length) 19 x (height) 15¾ inches. Depth behind panel, 13 inches.

Net Weight: 51 pounds.

SPECIFICATIONS, TYPE 1181-BT

Same as for Type 1181-B except as specified below:

Input Frequency: 3.579545 megacycles; unmodulated.

Frequency Stability: ± one cycle per second for 30 days; ± 5 cycles for one year.

Quartz Crystal: General Radio Type 376-R.

RF Sensitivity: .05 to 2.0 volts unmodulated r-f input.

Coupling to Transmitter: Shielded cable and plug provided.

SPECIFICATIONS, TYPE 1181-BH

Same as for 1181-B, except as noted below:

Frequency: 1.6 — 15 Mc.

Frequency Stability: ± 1 ppm for 30 days, or better; ± 5 ppm for 1 year.

R-F Input: 1.6 — 5 Mc, 0.1 — 2.5 volts, modulated or unmodulated.

5 — 15 Mc, 0.4 — 3.0 volts, modulated or unmodulated.

Quartz Crystal: General Radio Type 376-R.

Coupling to Transmitter: Shielded cable and plug provided.

Type		*Code Word*	*Price*
1181-B	Frequency Deviation Monitor	MALAY	$1025.00
1181-BT	Color Subcarrier Monitor	MAJOR	1025.00
1181-BH	Frequency Deviation Monitor	MADAM	1025.00

U.S. Patents 2,298,177 and 2,362,503. Licensed under patents of Radio Corporation of America and Dr. G. W. Pierce.

TYPE 1931-B AMPLITUDE MODULATION MONITOR

The Type 1931 Amplitude Modulation Monitor, like the Type 1181 Frequency Deviation Monitor, is a General Radio development and has become an industry standard. The principle of operation is shown in Figure 8. Per-cent modulation is indicated continuously on a panel meter, and a warning lamp glows whenever modulation percentage exceeds a level set by means of a calibrated dial.

The audio-frequency envelope of the modulation wave form is available at a pair of terminals for operating the Type 1932-A Distortion and Noise Meter. A 600-ohm program-monitoring output is also provided.

The front panel is shown in Figure 9. Only those controls necessary to normal operation appear on this panel, which is removable by means of the two fasteners shown. Other controls and indicators appear on the inner panel shown in Figure 10. The chassis-top view of Figure 11 shows the circuit-flow markings and the color-coded spots that identify the internal adjustments.

Constructional features of the modulation monitor are identical with those of the frequency deviation monitor and are clearly shown in Figures 1, 8, 9, and 10.

SPECIFICATIONS

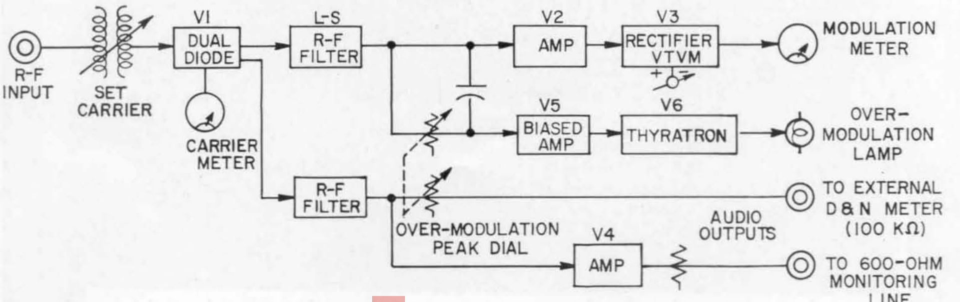
Range: Modulation percentage, 0 to 110%, indicated by meter on positive peaks, 0 to 100% on negative peaks. The flashing lamp is adjustable to operate from 0 to 100% on negative peaks.

Carrier-Frequency Range: The monitor will operate at any carrier frequency from 0.5 to 60 megacycles. A single set of coils (either 0.5 to 8 megacycles or 3 to 60 megacycles) is supplied with each instrument, unless both sets are specifically ordered.

Carrier-Frequency Input Impedance: About 75 ohms in the broadcast band, increasing slightly at higher carrier frequencies and varying somewhat with input tuning.

Accuracy: The over-all accuracy of measure-

Figure 8. Functional schematic of the amplitude modulation monitor.



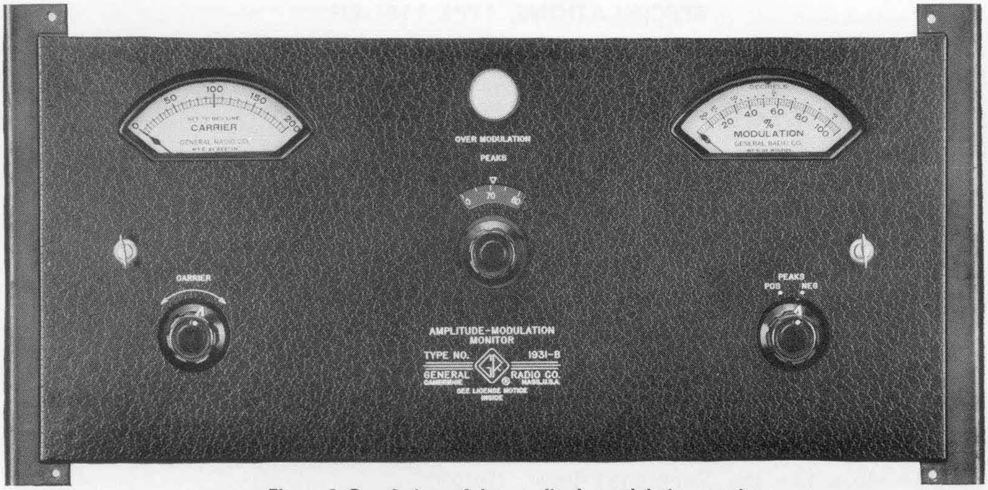


Figure 9. Panel view of the amplitude modulation monitor.

ment at 400 cycles is ±2% of full scale at 0% and 100%, and ±4% of full scale at any other modulation percentage.

Detector Linearity: The distortion in the diode detector is very low for frequencies up to 7500 cycles. Above this frequency, a small amount of negative-peak clipping occurs, reaching 5% at the extreme high end of the audio range at 15,000 cycles and 100% modulation.

R-F Power: In the broadcast range the maximum r-f power requirement is about 0.5 watt.

Tube Complement: The following tubes are used:

2 — 6SN7-GT	1 — 2050
2 — 6SJ7	2 — 0D3
1 — 6AL5	1 — 6X5GT

Warning Lamp Circuit: The OVERMODULATION lamp will flash whenever the negative modulation peaks exceed the setting of the MODULATION PEAKS dial by 2% or more modulation, for audio frequencies between 30 and 7500 cycles. For higher audio frequencies, the percentage overmodulation required to flash the lamp increases slightly.

The accuracy of the dial calibration is ±2% of full scale.

Meter Circuit: The response of the PERCENTAGE MODULATION meter circuit is flat, within ±0.25 db, between 50 and 15,000 cycles, and within ±0.1 db between 100 and 10,000 cycles.

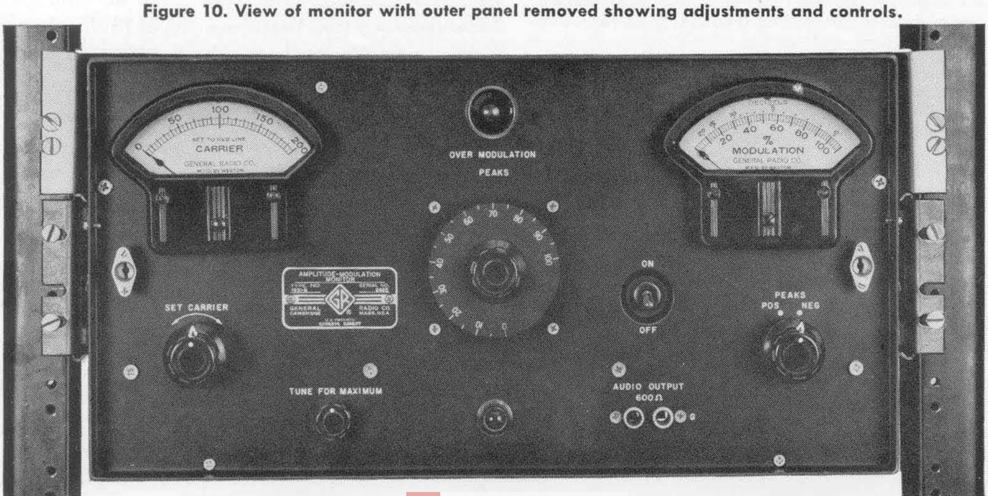
Either positive or negative modulation peaks may be read. Calibration in db below 100% modulation is provided.

The meter dynamic characteristic meets FCC specifications for modulation monitors.

Audio Monitoring Output: The audio output amplifier is flat, within ±1.0 db, from 30 to 45,000 cycles. The internal impedance is 600 ohms. Distortion is less than 0.2%. Open-circuit output voltage is about 300 millivolts.

Fidelity-Measuring Output: Flat within ±1.0 db between 30–30,000 cycles with Type 1392-A

Figure 10. View of monitor with outer panel removed showing adjustments and controls.



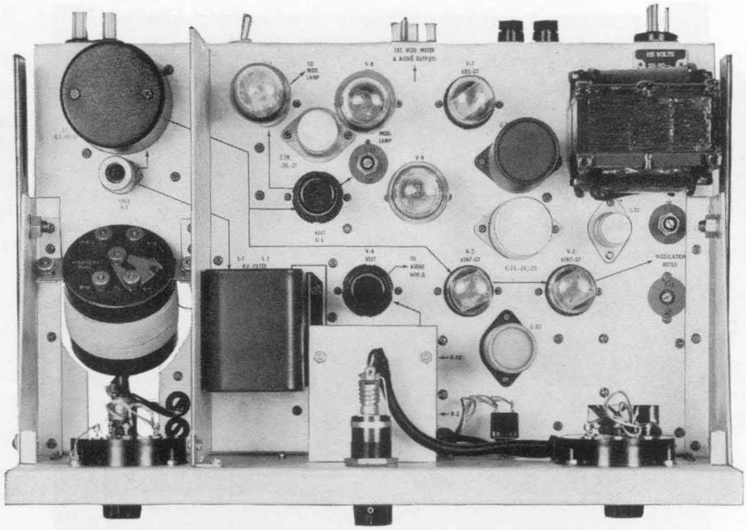


Figure 11. View of chassis showing signal flow lines.

Distortion and Noise Meter connected. Distortion less than 0.1%.

Output level varies inversely with setting of MODULATION PEAKS dial, thus providing reasonably uniform input to distortion meter at all modulation levels. Average output level, approximately 1.5 volts.

Residual noise and hum level will not exceed −80 db.

Auxiliary Output: A multipoint connector at the rear of the instrument provides a means of connecting:

1. An external Percentage Modulation Meter, either local, 800 Ω max., or remote, 5000 ohms max. loop resistance.
2. To a 600-ohm output for audio monitoring.
3. The Type 1932-A Distortion and Noise Meter.

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

Accessories Supplied: Multipoint connector Type ZCAP-5 Power Cord, spare fuses, and one set of input tuning coils (specify frequency range desired).

Mounting: The instrument is relay-rack mounted. End frames are available for table mounting. (See price list below.)

Panel Finishes: Standard General Radio black crackle. Certain standard grays which can be processed in quantity can also be supplied.

Dimensions: Panel (length) 19 x (height) 8¾ inches. Depth behind panel, 10 inches.

Net Weight: 32¾ pounds.

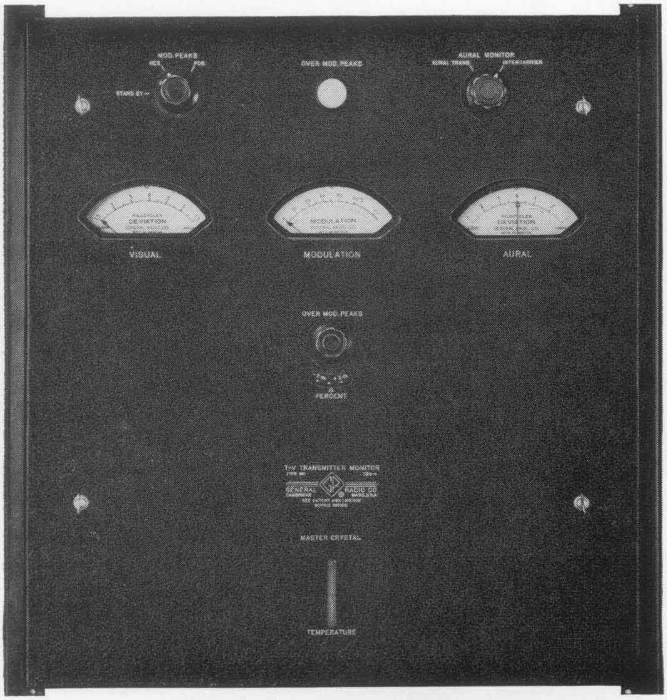
Type		*Code Word*	*Price*
1931-B*	**Modulation Monitor, 0.5 to 8 Mc**	TARRY	**$625.00**
1931-B*	**Modulation Monitor, 3 to 60 Mc**	TOPIC	**625.00**
1931-P5	**Extra Tuning Coils, 0.5 to 8 Mc**	TABBY	**23.00**
1931-P6	**Extra Tuning Coils, 3 to 60 Mc**	TOTEM	**23.00**
FRI-510	**End Frames**	ENDFRAMEAT	**13.00 pair**

*U. S. Patent 2,298,177.

FCC TYPE APPROVAL FOR GENERAL RADIO TELEVISION TRANSMITTER MONITORS

The Federal Communications Commission has issued type approval under Part 3 of the Commission's rules for the General Radio Type 1184-A-A Television Transmitter Monitor. The approval covers operation on all VHF and UHF channels, and the FCC type approval number is 3-105.

There are some Type 1184-A Monitors now in service that differ from the ap-



Panel view of the Type 1184-A-A Television Transmitter Monitor.

proved Type 1184-A-A in two minor respects, both easily changed in the field without interrupting operation. A bakelite disc in the overmodulation-warning lamp is to be removed, so as to make the light brighter, and a "Δf check" push button is to be installed in order to allow a quick check of over-all operation.

The Commission has also issued type approval for the older Type 1183-T-A Television Transmitter Monitor. The approval covers operation on VHF and UHF channels below 800 Mc, which covers all monitors now in service, and the FCC type approval number is 3-104.

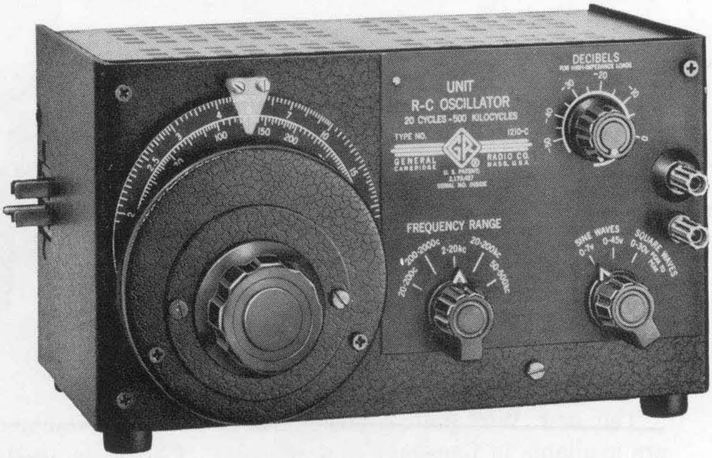
The existing Types 1183-T1, -T2, and T3 Monitors now in service differ from the approved Type 1183-T-A in the same two respects described above for the Type 1184.

We will supply, on request and at no charge, instructions and all necessary material for making the changes on all Type 1184-A and 1183-T1, -T2, and -T3 Monitors now in service. Please write to our Service Department, 22 Baker Avenue, West Concord, Massachusetts, giving type and serial number of your monitor, shipping address, and name of individual who will be responsible for handling the matter.

These General Radio monitors are the first *complete*, television frequency and modulation monitors, and the first UHF monitors of *any* type, to be granted FCC type approval under the current rules.

The older Type 1183-T Monitor was designed *before* precision-offset operation, color transmission, and the current FCC rules were in effect. The fact that it has been granted approval under the current rules illustrates one of the many advantages of buying General Radio equipment. The typical substantial margin of performance *beyond* current requirements is the purchaser's best available insurance against early obsolescence.

TYPE 1210-C UNIT R-C OSCILLATOR



The Type 1210 Unit R-C Oscillator,[1] one of the electronics laboratory's most useful general-purpose instruments, has several noteworthy features:

1. Small dimensions
2. Sweepable, with inexpensive drive
3. Sine- and square-wave output
4. Low-impedance, low-distortion output
5. High-impedance, high-voltage output

This versatile oscillator appears now in slightly different external appearance, necessitated by a change in internal construction.

The new model, Type 1210-C, remains unchanged in its electrical characteristics but has been redesigned to use etched circuits, a development that results in a greater uniformity of performance and greater reliability than can be achieved by individual wiring.

The etched circuit, like many manufacturing developments, has been some years in achieving maturity. Fluxing, soldering, and cleaning problems have been largely solved, and its use in equipment where uniformity and reliability are paramount considerations is steadily growing. For some years, the General Radio Company has been carrying on a continuous development program in etched circuit technology with the aims of (1) eliminating the "bugs" that inevitably crop up in a new process and (2) building a technique of manufacture and test.

The fruits of this program can already be seen in a number of General Radio instruments, the latest of which is the new Type 1210-C.

[1]Bousquet, A. G., Peterson, A. P. G., and Sinclair, D. B., "Unit R-C Oscillator — 20 Cycles to 500 Kc," *General Radio Experimenter*, Vol. 29, No. 12, May, 1955, pp. 1-11.

CONDENSED SPECIFICATIONS

Frequency Range: 20–500,000 c in 5 ranges.

Frequency Accuracy: ± 3%.

Output Control: Logarithmic, calibrated 0–50 db.

Output System: 3-position switch for square-wave, sine-wave low-impedance, or sine-wave high-impedance output.

Low-Impedance Output: (For loads of 500 ohms and higher) 0–7 v, ± 1 db up to 200 kc.

High-Impedance Output: (For loads of 10,000 ohms and higher) 0–45 v, ± 1 db from 200 c to 200 kc.

Square-Wave Output: 0–30 v peak to peak; rise time approximately ¼ μsec.

Tube Complement: One each 6BQ7-A, OB2; two 12AU7's.

Power Supply: 6.3 v ac or dc at 1 amp.; 300 v dc at 50 ma. Type 1203-A Unit Power Supply for operation from 115 v, 50–60 cycles.

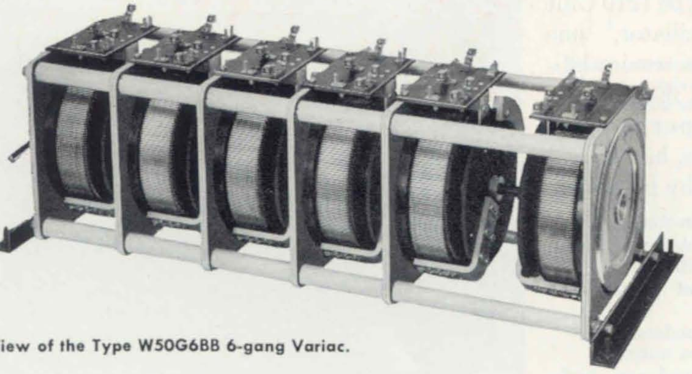
Mountings: Black-crackle finish aluminum panel and sides; aluminum cover finished in clear lacquer.

Dimensions: 10½ (width) x 5¾ (height) x 7 inches (depth) over-all. **Weight:** 6¼ lb.

Type		*Code Word*	*Price*
1210-C*	**Unit R-C Oscillator**	ABAFT	**$180.00**
1203-B	**Unit Power Supply**	ALIVE	**40.00**
480-P4U3	**Relay-Rack Panel (for mounting both 1210-B and 1203-A in one panel)**	UNIPANCART	**10.85**

*U. S. Patent No. 2,173,427.

HIGH-POWER VARIAC® AUTOTRANSFORMER ASSEMBLIES



View of the Type W50G6BB 6-gang Variac.

The new W50 and W50H Variacs® are available in four-gang and six-gang assemblies as well as the familiar two-gang and three-gang assemblies available in most Variac® types. This allows double the load capacity for single assemblies. The table gives the kva ratings for 4 and 6-gang assemblies in common circuits.

All 4-, 5-, and 6-gang models are equipped with ball bearings.

CHOKES

In order to avoid circulating current, it is recommended that Type 50-P Chokes be used with these gangs. Order chokes as follows:

Assembly	*Chokes Needed*
4-gang parallel	3 Type 50-P1
6-gang parallel	3 Type 50-P1 and 2 Type 50-P2
4-gang delta	2 Type 50-P1
6-gang delta	2 Type 50-P1 and 2 Type 50-P2
6-gang wye	3 Type 50-P1

Type	*Code Word*	*Price*
50-P1	PARALLCHOK	**$16.00**
50-P2	TRIPLECHOK	**16.00**

Type		*Load Ratings — KVA* *Parallel*	*Delta*	*Wye*	*Code Word*	*Price*
W50G4BB	**4-Gang Variac**	23.0	19.9		GATALGANKA	**$540.00**
W50G4BBM	**4-Gang Variac with case**	20.7	18.0		GATALBONKA	**600.00**
W50G5BB	**5-Gang Variac**				GATALGANFO	**670.00**
W50G5BBM	**5-Gang Variac with case**				GATALBONFO	**735.00**
W50G6BB	**6-Gang Variac**	34.5	29.9	40.0	GATALGANSA	**800.00**
W50G6BBM	**6-Gang Variac with case**	31.2	27.0	36.0	GATALBONSA	**870.00**
W50HG4BB	**4-Gang Variac**	29.9	25.9		NITALGANKA	**540.00**
W50HG4BBM	**4-Gang Variac with case**	28.5	24.6		NITALBONKA	**600.00**
W50HG5BB	**5-Gang Variac**				NITALGANFO	**670.00**
W50HG5BBM	**5-Gang Variac with case**				NITALBONFO	**735.00**
W50HG6BB	**6-Gang Variac**	44.9	38.9	52.0	NITALGANSA	**800.00**
W50HG6BBM	**6-Gang Variac with case**	42.7	36.9	49.4	NITALBONSA	**870.00**

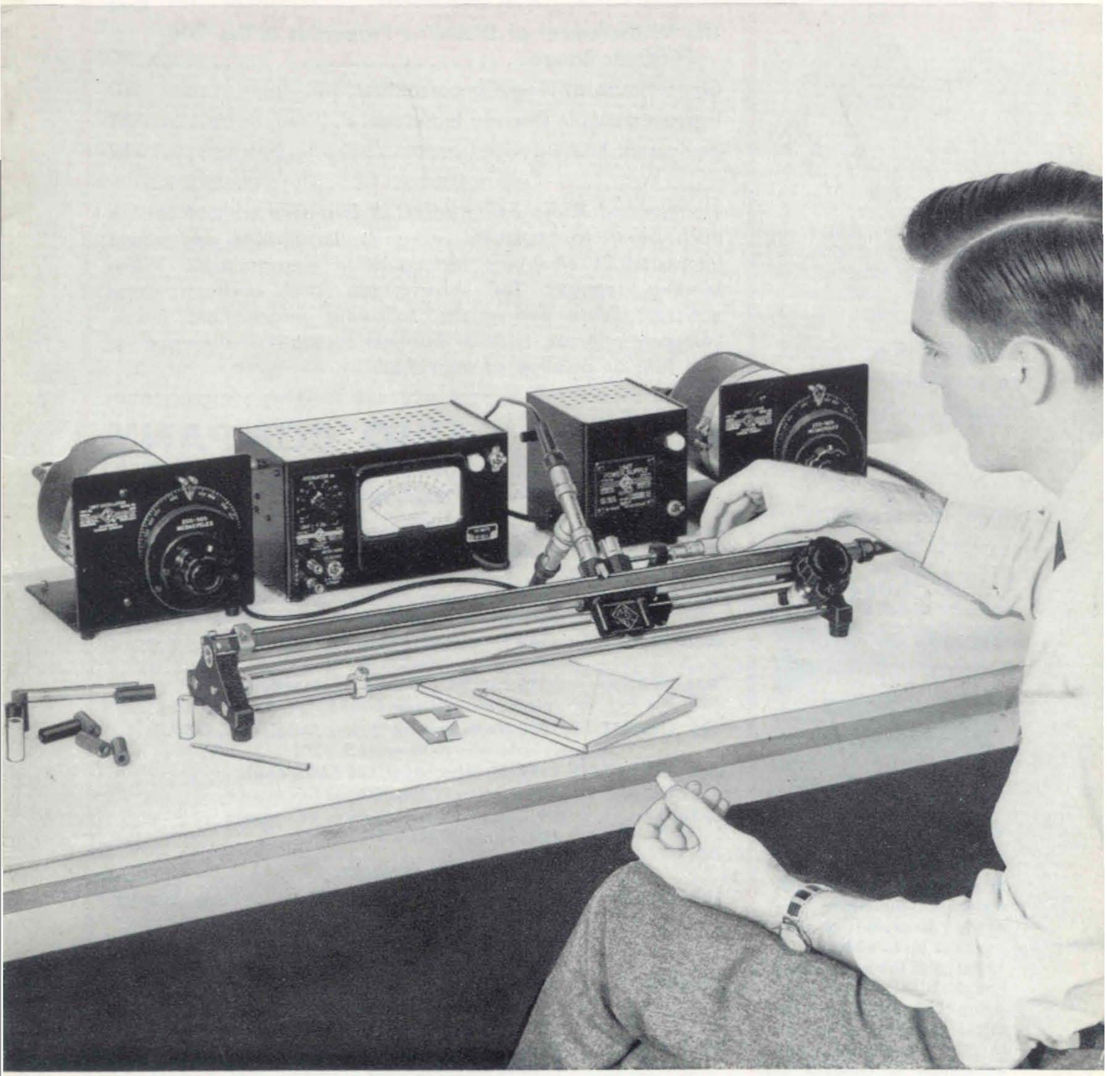
U. S. Patent applied for.

the GENERAL RADIO

Experimenter

Since 1915 – Manufacturers of Electronic Apparatus for Science and Industry

VOLUME 32 No. 12 MAY, 1958



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Experimenter

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VOLUME 32 • NUMBER 12 **MAY, 1958**

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COVER

Dielectric constant and dissipation factor of solid-dielectric materials are easily measured at frequencies up to 5000 megacycles with the new slotted line described in this issue. Using easily prepared cylindrical samples, as shown in the photograph, this equipment is capable of measuring very-low-loss materials, such as polystyrene and Teflon.

THE MEASUREMENT OF DIELECTRIC PROPERTIES IN THE 200-5000 MC RANGE

A simplified method and a new slotted line have been developed for the measurement of the dielectric constant and loss of low-loss solid insulating materials in the 200-5000 Mc frequency range. The basic method is a modification of one previously described by Robertson and von Hippel,[1] and Dakin and Works.[2]

Method

If a section of open-circuited transmission line is filled with dielectric as shown in Figure 1, the dielectric constant and dissipation factor can be determined from measurements of the input impedance to the section of line filled with the dielectric and from a knowledge of the frequency and the sample length. The input impedance, Z_i, can be calculated from measurements of the standing-wave pattern present on the air-filled section of line shown in the figure. The actual calculations are complicated by the fact that the equation which has to be solved is complex and transcendental, and they become particularly involved when a low-loss dielectric is being measured, in which case the resistive losses in the line are usually large compared with the dielectric losses in the sample. The whole procedure can be greatly simplified if either the sample length or the frequency is adjusted to make the electrical length of the sample an odd multiple of a quarter-wavelength, so that the voltage minimum on the air-filled section of the line will appear at the front face of the sample. Under these conditions the equations for the dielectric constant and dissipation factor are the following:

$$K = \left(\frac{N\lambda}{4l}\right)^2 \quad (1) \qquad D = \frac{\lambda}{\pi l \rho} - A \quad (2)$$

where
K is the dielectric constant
D is the dissipation factor
λ is the wavelength in free space
l is the physical length of the sample
N is an odd integer which is equal to the number of quarter-wavelengths in the sample
A is a constant which corrects for the resistive losses in the line conductors and
ρ is the VSWR on the air dielectric line.

Since a voltage minimum will appear at the dielectric face at frequencies at which the electrical length of the sample (equals $\frac{l}{\sqrt{K}}$) is ¼, ¾, 5/4, etc., wavelengths, one measurement is not sufficient to give a unique value of the constant. Measurements at two frequencies properly chosen will give a unique solution. In many instances the dielectric constant is approximately known, and, therefore, the proper integer for N can be chosen by inspection.

In practice, however, it is not possible to locate the voltage minimum accurately and to measure the standing-wave ratio when the minimum is directly at the front face of the sample. It has

[1] S. Robertson, A. von Hippel, Journal of Applied Physics, 17, 610 (1946).
[2] A. W. Dakin, C. N. Works, Journal of Applied Physics, 18, 789 (1947).
(For an excellent treatment of dielectric measurements, see also A. von Hippel, *Dielectric Materials and Applications*, John Wiley and Sons, New York, 1954.)

Figure 1.

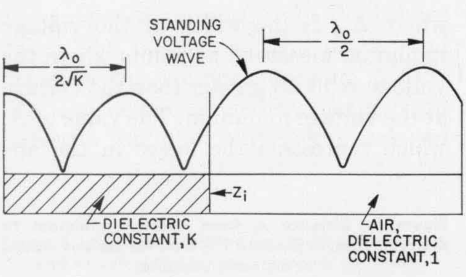
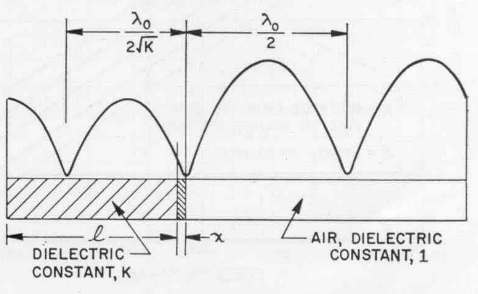


Figure 2.



been determined, however, that, with small modifications of the simple expressions previously presented, one can measure the dielectric properties with the voltage minimum in the vicinity of the sample and yet far enough away to make the selection of the frequency noncritical and to permit the measurement of VSWR by a width-of-minimum method.

Let us assume that the dielectric is exactly an odd multiple of a quarter-wavelength long and that the voltage minimum is directly at the front face of the sample as shown in Figure 2. Under these conditions, K and D can be calculated from Equations (1) and (2). Now suppose that a small amount of the dielectric near the voltage minimum is removed as indicated by the cross-hatched section in the figure, where the length of sample removed is indicated by x. If the sample has low loss, the voltage appearing at the voltage minimum relative to the voltage at other points along the line is very small and, therefore, the effect of dielectric near this point in the line is small compared to the effect at other points along the line. Consequently, the removal of the dielectric in the area near the voltage minimum has very little effect on the position of the minimum on the line. The dielectric constant can, therefore, be measured with the voltage minimum slightly ahead of the dielectric face with a negligibly small error. The modified equation for dielectric constant is:

$$K = \left[\frac{N\lambda}{4(l + x)}\right]^2 \qquad (3)$$

The above equation is based on the following approximation:

$$\frac{1}{\sqrt{K}} \tan (\sqrt{K}\, \beta x) = \tan \beta x \qquad (4)$$

where β is the propagation constant in free space. The error in the determination of K resulting from the use of this approximation is:

$$\frac{8}{3}\left(\frac{K - 1}{l}\right)\left(\frac{\pi}{\lambda}\right)^2 x^3 \qquad (5)$$

The value of x which results in a 1% error is plotted in Figure 3. As indicated in Equation (5), the error varies as the cube of x.

The same condition holds for measurements of the dissipation factor. Near voltage minimum the dielectric loss is low, because of the relatively low voltage and, therefore, the dielectric can be removed near the minimum without a significant change in the over-all dielectric loss. The resistive loss in the conductors in the area where the dielectric is removed is practically unaffected since the current distribution remains practically the same. The equation for dissipation factor becomes:

$$D = \frac{\Delta_{10}}{3(l + x)} - A \qquad (6)$$

where Δ_{10} is the width of the voltage minimum measured at points where the voltage is 10 db greater than the voltage at the voltage minimum. The value of A, which represents the losses in the air-

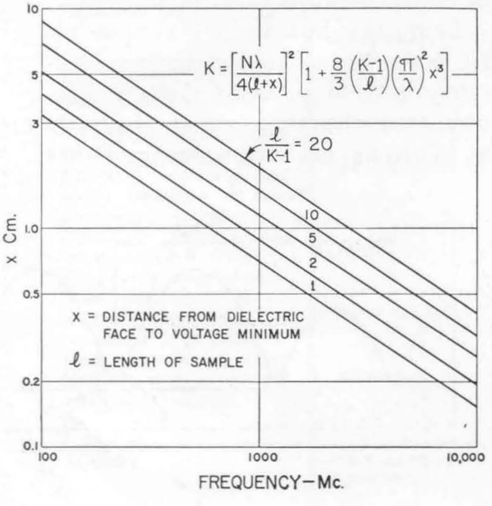


Figure 3. Distance x, from voltage minimum to dielectric face, to produce 1% error for various values of dielectric constant.

filled line, can be obtained from a measurement of the VSWR on the line with the sample removed. The equation for A is:

$$A = \frac{\Delta'_{10}}{3l'} \qquad (7)$$

where the width of the voltage minimum between 10-db points, Δ'_{10}, is measured with the sample removed, and where l' is the distance in centimeters between the voltage minimum at which the VSWR is measured and the open end of the line. For maximum accuracy, the length of line measured should be as long as possible.

THE TYPE 874-LM DIELECTRIC MEASURING LINE

As is obvious from the preceding description, in the proposed measurement method the slotted line and the sample holder must be one integral unit. The instrument designed for this function utilizes a coaxial transmission line which consists of an accurately machined bronze outer tube having an inside diameter of 0.562″ (7/16) with a 0.002″ overlay of silver on its inner surface and an inner conductor 0.250″ in diameter also with a similar overlay. The line is sufficiently long to accommodate samples up to 45 centimeters in length. The minimum frequency at which measurements can be made by the simplified method previously described depends upon the dielectric constant of the material being measured and is approximately $f_{min} = \frac{200}{\sqrt{K}}$ Mc. Of course, measurements can be made at lower frequencies by the more complicated method previously referred to. The actual low-frequency limitation is primarily determined by the sensitivity of the detector.

A movable carriage containing the probe rides on the accurately ground and lapped outer surface of the bronze tube. The position of the probe, and hence of the voltage minimum on the line, can be determined by means of an indicator on the centimeter scale to within ± 0.01 centimeter. A micrometer vernier drive on the carriage is provided for width-of-minimum measurements. The micrometer is capable of measuring the width of a minimum to within 0.0002 centimeter.

SOURCES OF ERROR

Line Losses

As previously mentioned, in measurements on very low-loss materials the line loss can be large compared to the dielectric loss. The effective dissipation of the line itself, A, caused by resistive losses is plotted in Figure 4 as a function of frequency. For purely resistive losses, the effective resistance per unit length increases as the square root of frequency as a result of skin effect and, hence, the effective dissipation factor per unit length decreases as the square root of frequency. Also plotted on Figure 4 is the calculated dissipation factor assum-

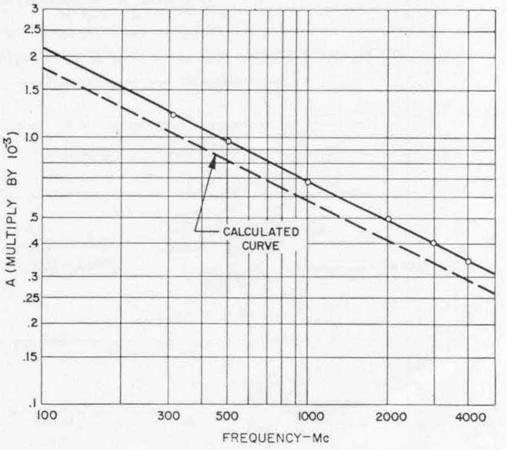


Figure 4. Losses in the slotted line itself, as measured, and as calculated for smooth silver surfaces.

ing smooth, pure-silver, inner and outer conductor surfaces. Note that the agreement is quite good.

The ratio of the dielectric loss to the resistive loss on the line varies with both the dissipation factor of the sample under test and the frequency. It is independent of the length of the sample when the VSWR measurement is made close to the dielectric face. Figure 5 is a plot of this ratio for a normalized dissipation factor 0.0001. In other words, at a frequency of 500 Mc, the ratio for a dielectric material having a dissipation factor of 0.001 would be 1.4. It is obvious that at lower frequencies the ratio becomes smaller, and, hence, an accurate measurement of line losses is very important. Small errors in this measurement can have a large effect on the measured dissipation factor of low-loss materials at low frequencies.

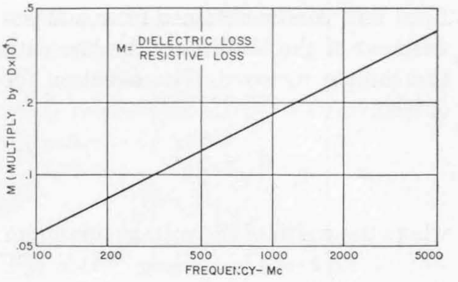


Figure 5. Ratio of loss in sample to loss in line normalized for a *D* of 0.0001.

Fringing

In an open-ended transmission line, fringing capacitance exists at the end of the conductors. This capacitance makes the transmission line appear longer electrically than it is physically. Figure 6 shows the results of measurements made on the fringing capacitance in the coaxial dielectric-measuring line with the dielectric removed. The added electrical length, resulting from the fringing capacitance of $0.15\mu\mu f$, is approximately 0.23 centimeter. When dielectric samples are present, the effective added length decreases approximately proportionally to the dielectric constant of the insulating material. The correction for fringing capacitance can be included in the dielectric-constant equation as indicated below or, if a sufficiently long dielectric sample is used, the error is negligible.

$$K = \left[\frac{N\lambda}{4(l + x)}\right]^2 - \frac{0.46}{l} \qquad (8)$$

Air Gaps

Air gaps between the sample and the dielectric conductors of the transmission

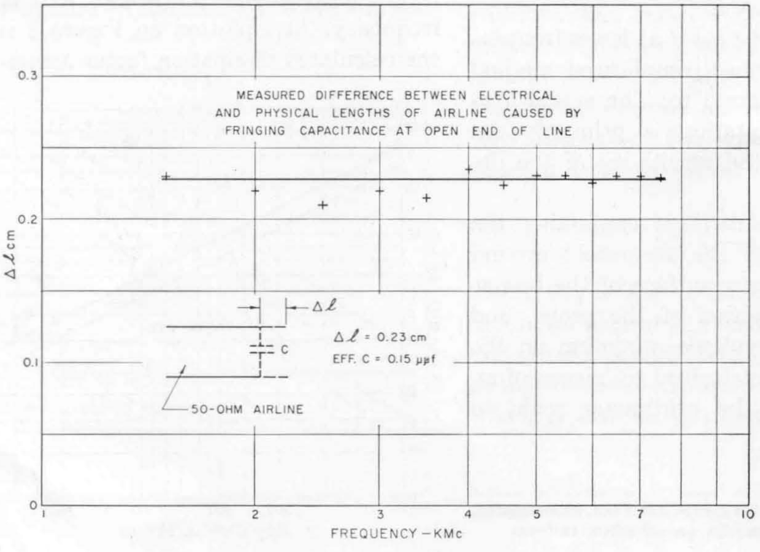


Figure 6. Effective increase in line length caused by fringing.

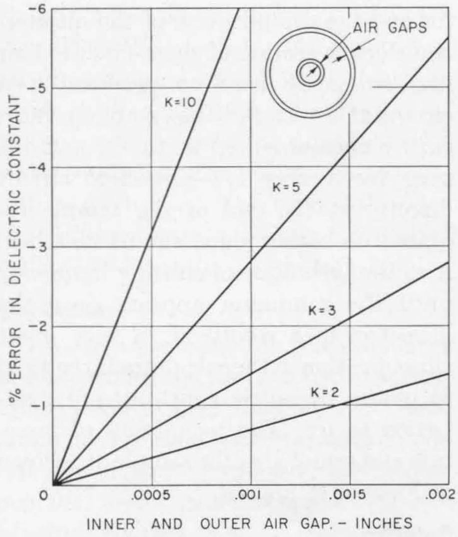


Figure 7. Error in *K* caused by air gaps between dielectric sample and inner and outer conductors.

line cause errors in the measurements. Figure 7 indicates how for a given gap those errors vary with the dielectric constant on the coaxial line described in this article. The minimum practical gap is one of the factors which set a limit on the maximum dielectric constant which can be measured to a given accuracy.

Probe Coupling

As a probe approaches the dielectric sample, its effective coupling to the center conductor will increase as a result of the dielectric effectively increasing the capacitance between the probe and line. The point at which the probe coupling is significantly affected is the minimum distance which can be tolerated between the dielectric face and probe. Figure 8 shows the variation in probe coupling as a function of distance measured in the line described later in this article. Note that, for relatively low-loss dielectric materials with dielectric constants even as large as 9,000, the increase in coupling is very small at distances greater than a millimeter. Therefore, the probe should not be allowed to approach closer than a millimeter when either the position of the minimum or the width of the minimum is measured. When high-loss materials are measured, and the minimum is very broad, an even larger minimum spacing is necessary, because even a slight variation in coupling will shift the apparent center of the broad minimum. Figure 9 shows the results of measurements made on various samples in order to determine the effect on the position of the minimum as a result of added probe coupling. In this figure the position of the voltage minimum with respect to the front face of the dielectric is plotted as a function of wavelength for samples having various dissipation factors. If probe coupling were constant, these curves would be straight lines. Note that, for the two relatively low-loss samples, the variation follows the linear law until a spacing of less than

Figure 8. Variation in probe coupling as a function of distance between probe and dielectric face.

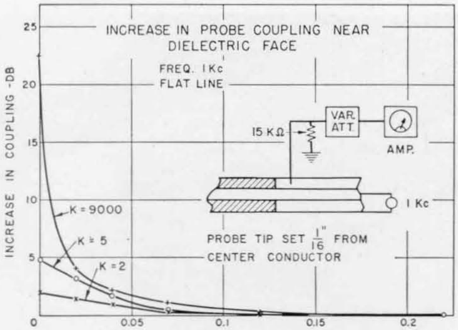
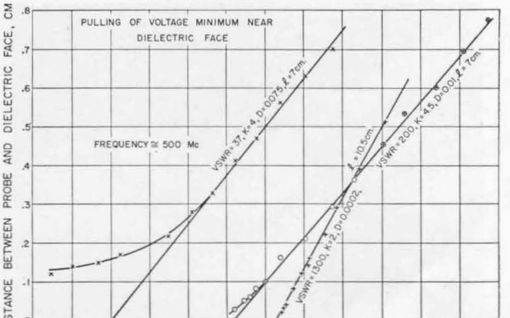


Figure 9. Effect of probe coupling of position of minimum.



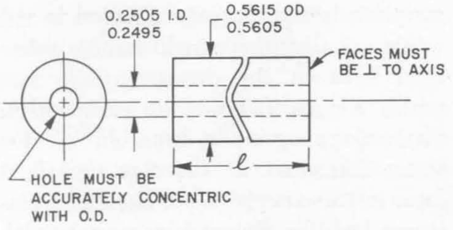


Figure 10. Dimensions of sample.

one millimeter is reached. The sample having a very high dissipation factor, however, shows that the position of the minimum is affected at spacings up to 0.3 centimeter.

The Sample

One limitation of the proposed method is that the samples must be at least a quarter-wavelength long, and at the lower frequencies this length may not be easy to obtain in a single sample. However, the quarter-wavelength section can be made up of a number of shorter lengths, which can be easily machined. Another limitation is that either the frequency or the length of the sample must be trimmed to obtain a voltage minimum within the limits previously described. If the measurements must be made exactly at a specified frequency, some cut-and-try work must be done on the samples themselves. The dimensions of the sample are shown in Figure 10.

The use of an open-ended line greatly simplifies the mechanics of the insertion of the samples and eliminates the need for and the complication of the quarter-wavelength section of short-circuited air line ordinarily used to produce open circuit at the end of the sample in most earlier open-circuited methods, and the need for a very low-loss short circuit directly at the end of the sample required on earlier short-circuit methods. Also the technique of shifting frequency until the minimum appears near the dielectric face results in a very great simplification in the calculations required to obtain dielectric constant and dissipation factor. The techniques of insertion and removal of the sample are shown in Figures 11 and 12.

Detector

The detector which produces the best results in this application is heterodyne type in which the signal from the probe and a signal from a local oscillator are applied to a crystal diode mixer and the difference frequency output is amplified by means of a fixed tuned I-F amplifier. The Type 1216-A Unit I-F Amplifier, Type 874-MR Mixer Rectifier and various Unit Oscillators are well suited to this application. These are available in complete combinations as the Type DNT Detectors. The r-f mixer is accurately linear over a voltage range from about 80 db and, hence, the relative level of the signal picked up by the probe can be easily measured by means of a calibrated step attenuator and calibrated

Figure 11. Sample is easily installed by means of tool furnished.

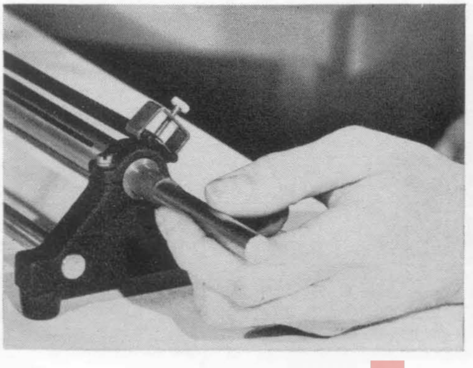
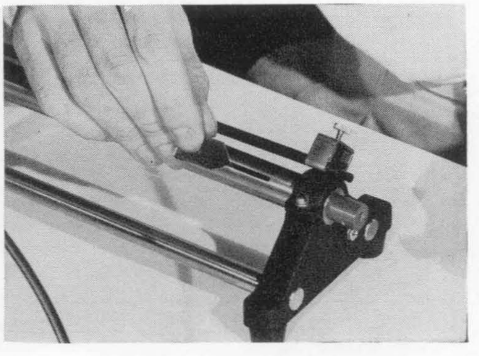


Figure 12. Tool for removing sample is also furnished.



output meter in the Type 1216-A I-F Amplifier. With a heterodyne detector, an unmodulated signal can be used to excite the line, thus eliminating errors resulting from incidental frequency modulation produced when the oscillator is amplitude modulated. This type of detector has high sensitivity, good linearity, and excellent discrimination against harmonics.

Oscillator

The requirements for the oscillator used to excite the line are not exacting. Simple, low-power, tunable, c-w oscillators having good stability, such as the GR line of Unit Oscillators, are recommended. Signal generators are also satisfactory sources. Since a high-sensitivity detector is used, the oscillator can be directly coupled to the slotted line, preferably through a resistive pad without the use of tuning elements.

The complete setup, including generator and detector, is shown in the block schematic of Figure 13.

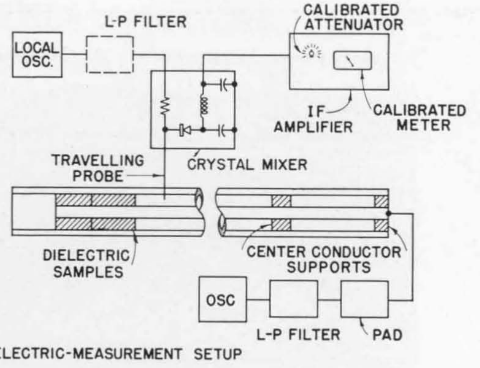


Figure 13. Block diagram of complete dielectric measuring system.

Frequency Limitations

The maximum frequency at which reliable measurements can be made is that at which the first higher-order mode can be propagated in the dielectric under test. In the air-filled line, the first higher-order mode, which is a circumferential mode, has a cutoff frequency of about 9000 Mc. This cutoff frequency decreases as the square root of the dielectric constant and hence the upper frequency limit is $9000/\sqrt{K}$.

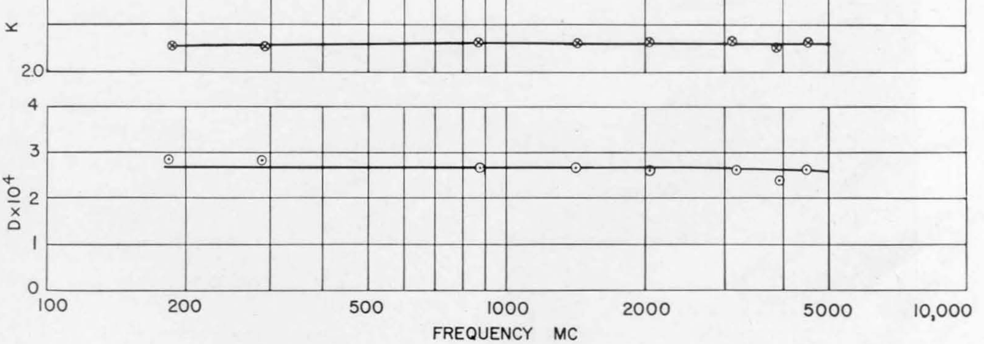
Accuracy

With this instrument, accurate straightforward measurements on materials having dielectric constants between 1 and 10 and dissipation factors between 0.0001 and 0.05 can be made with an accuracy of about $\pm$ 2% in dielectric constant and $\pm$ (5% + 0.0001) in dissipation factor over a frequency range from about 200 to 5000 Mc. The dielectric-constant accuracy is satisfactory for most applications, and the dissipation-factor resolving power makes possible reasonable measurements on even the lowest loss materials presently available.

Applications

Figure 14 shows the results of measurements made on Teflon over a wide range of frequency. — R. A. Soderman

Figure 14. Dielectric constant and dissipation factor of Teflon as measured on the Type 874-LM Dielectric Measuring Line.



SPECIFICATIONS

Frequency Range: Minimum, $\frac{200}{\sqrt{K}}$ Mc; Maximum 5000 Mc or $\frac{9000}{\sqrt{K}}$ Mc, whichever is the smaller.

Measurement Ranges and Accuracies: Dielectric Constant $(K) \pm 2\%$ for values of K between 1 and 10; Dissipation Factor, $\pm$ (5% + 0.0001) for values of dissipation factor between 0 and 0.05.

Sample: Cylindrical; O.D., 0.561 inches; I.D., 0.250 inches; length depends upon dielectric constant and frequency; long samples can be made up of a number of short sections.

Micrometer Scale: Can be read to 0.0002 cm.

Accessories Required: Generator and detector; General Radio Unit Oscillators (with TYPE 1201-A Unit Regulated Power Supply) and Type DNT Detectors, respectively, are recommended. A complete listing of these will be found in the March, 1958, issue of the *Experimenter*. Also required, one TYPE 874-G6 6-db Pad and one TYPE 874-G3 3-db Pad. Appropriate TYPE 874-F Filters are also recommended.

Dimensions: 26 x 4½ x 3½ inches, over-all.

Net Weight: 9½ pounds.

Type		*Code Word*	*Price*
874-LM	**Dielectric Measuring Line**	COAXFACTOR	**$400.00**

OPEN HOUSE AT NEW CONCORD PLANT

On Friday, June 6, 1958, from 1:00 P.M. until 4:00 P.M. the General Radio Company will hold open house at its newly enlarged plant in West Concord, Mass., located between Massachusetts highways Routes 2 and 62. We invite all of our customers and friends who are interested to attend. Ample parking facilities are located immediately adjacent to the buildings.

Open for inspection:

☆ Many operating displays of the latest GR instruments.

☆ The engineering development laboratories.

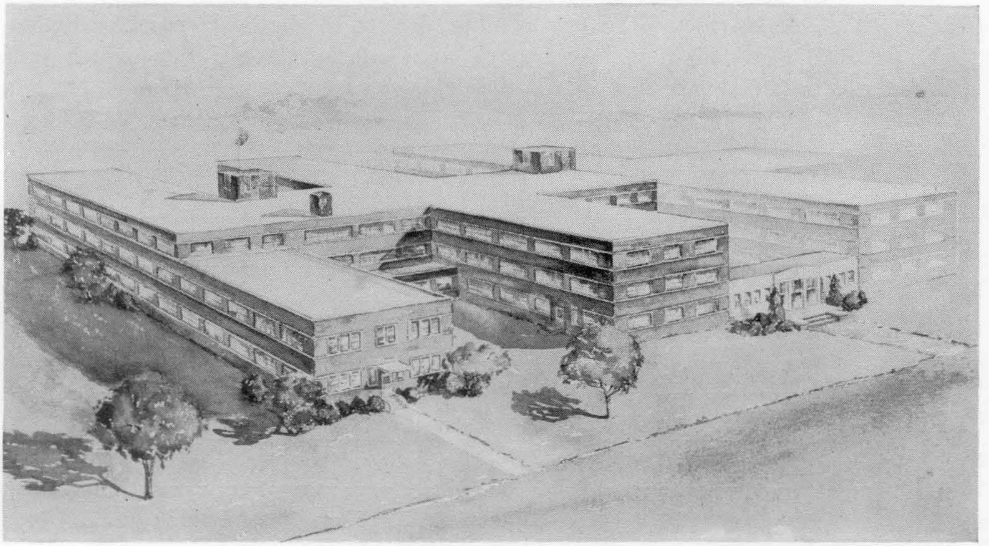
☆ The well-equipped model shop, in which the first models of new items are made.

☆ Standardizing laboratory, where all GR instruments are calibrated, checked and certified prior to shipment.

☆ Instrument- and component-assembly departments.

☆ Variac® winding, assembly and testing department.

☆ Supporting departments, such as production planning, purchasing and shipping.



IMPROVEMENTS IN DECADE INDUCTORS

The Type 1490 Decade Inductors and the Type 940 Decade Inductance Units have recently been redesigned to use a newer type of switch. As a result, the Q's of the low-inductance units have been improved and the resonant frequencies of the high-inductance units have been raised. These two improvements come about from the lower and more stable contact resistance of the new switch and from a change in the switching method.

The new decade switch has high-quality, ceramic stator-and-rotor members and utilizes a well-defined ball-and-socket detent. All contacts are made of a solid-silver alloy and have a positive wiping action. This switch is inherently reliable in extensive use and should not require bothersome cleaning or adjustment in service.

The d-c resistance at zero setting of a four-decade Type 1490 Decade Inductor with the older switch is over 400 milliohms, while with the new switch it is approximately 30 milliohms. The former value is a substantial part of the total resistance in the lowest-inductance decade, where the improvement ranges from 6.2:1 for the 1-mh step to 1.6:1 for the 10-mh step. There is a corresponding improvement in Q at low frequencies, as shown for the 1-mh step in Figure 2.

A different method of switching has lowered the stray capacitance across the active inductors, with a consequent increase in their resonant frequency. This is most important on the high-inductance decades but is also significant for the lower decades of a Type 1490 Assembly. Originally, four inductors of unit values 1-2-2-5 were connected in series, and those inductors not required for any setting were shorted out by switches. This added the resistance of the closed switches in series with the active inductors and placed the ground capacitance of the unused inductors across the active inductors below them in the series circuit.

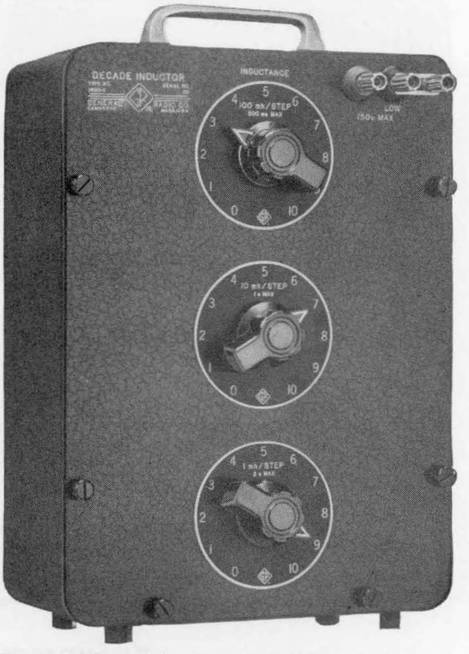
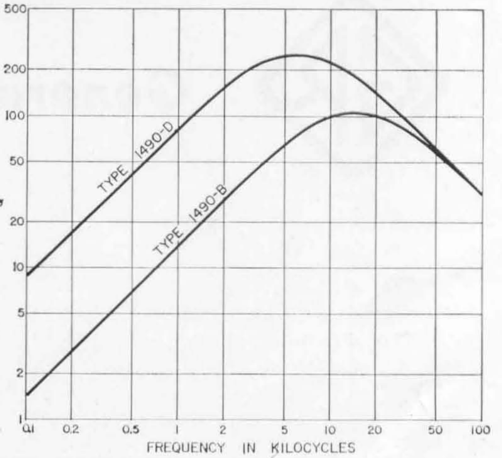


Figure 1. View of the Type 1490-C Decade Inductor.

With the new switching sequence the number of switch contacts in circuit is minimized, and unused inductors are

Figure 2. Plot of Q versus frequency for the 1-millihenry step in the old and new decade inductors.



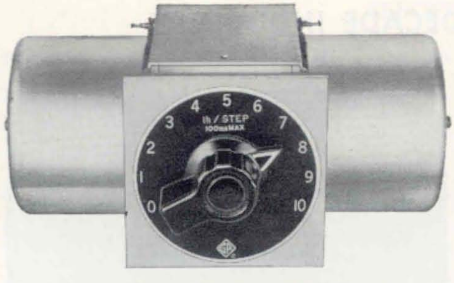


Figure 3. View of the Type 940-H Decade Inductance Unit.

completely disconnected. The resultant increase in resonant frequency for the low setting in each decade runs between 1.5 and 1.9, so that the correction factors for inductance at the high frequencies are reduced by these ratios.

The new switch requires somewhat more space, which increases the length of the Type 940 Decade in a direction perpendicular to the switch shaft and likewise the panel width of the Type 1490 Cabinets. New type letters have been assigned as follows:

New Type	*Replaces Old*	*Inductance*	*Code Word*	*Price*
940-E	940-A	1 millihenry/step	INDUCTOANT	$100.00
940-F	940-B	10 millihenrys/step	INDUCTOBOY	100.00
940-G	940-C	100 millihenrys/step	INDUCTOCAT	100.00
940-H	940-D	1 henry/step	INDUCTODOG	110.00
1490-C	1490-A	1.11 henry max, 1 mh/step	CLUMP	330.00
1490-D	1490-B	11.11 henry max, 1 mh/step	COACH	440.00

EQUIPMENT LEASING

Some users have found it desirable to lease test equipment rather than to purchase it outright.

We do not have arrangements for direct leasing, but there are a number of concerns which make a business of leasing. We should be glad to suggest to those interested the names of several firms which specialize in the leasing of electronic equipment.

General Radio Company

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the GENERAL RADIO

Experimenter

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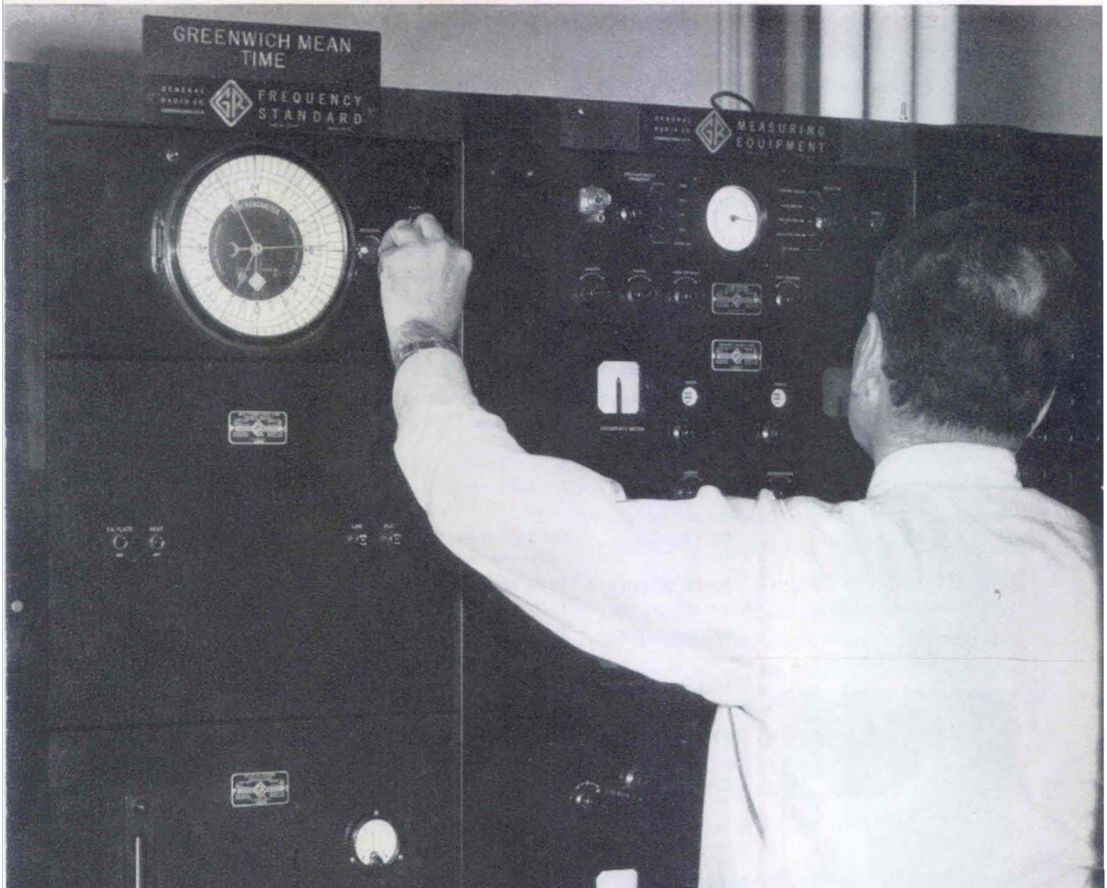
VOLUME 32 No. 13 JUNE, 1958



In This Issue | Improving the Accuracy of Time Comparisons

the GENERAL RADIO Experimenter

Published Monthly by the General Radio Company

VOLUME 32 • NUMBER 13 **JUNE, 1958**

CONTENTS

The General Radio EXPERIMENTER is mailed without charge each month to engineers, scientists, technicians, and others interested in electronic techniques in measurement. 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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COVER

Comparing time signals with the Type 1100-AP Primary Frequency Standard in the General Radio laboratories. The Comparison Oscilloscope is at the top of the rack directly in front of the observer, who adjusts the Microdial as he watches the oscilloscope pattern.

STANDARD TIME SIGNALS

Improving the Accuracy of Comparison Between Radio and Local Time Standards

Many of the activities associated with the International Geophysical Year require accurate time and frequency standardization. It is frequently desirable to have a local time-and-frequency standard, for use when radio reception is poor or when local timing equipment is to be operated. Standardization can be accomplished by a comparison of the time indicated by the local standard, such as the Type 1100-AP Primary Frequency Standard, with the standard radio time signals transmitted by national agencies in the United States and in other countries. These signals consist, in part, of pulses, dots, or dashes at intervals of one second. This article describes a method of measurement of the time of arrival of time signals with a precision of ± 1 millisecond.

The Local Standard

The Type 1100-AP Primary Frequency Standard comprises a quartz crystal oscillator operating at 100 kc and a frequency divider chain with output frequencies at 100 kc, 10 kc, 1 kc, and 100 cycles per second. The 1-kc output drives a precision clock (the Type 1103-A Syncronometer), in which is incorporated a contactor that opens and closes once per second. This contactor is adjustable in phase, or time of operation with respect to the clock shaft, in such a way that the closing time can be set at any value with respect to an arbitrary zero, from 0 to 999 milliseconds and on around, through zero, into the next second. A calibrated control, called the Microdial, is provided, which is graduated in 10-millisecond increments from 0 to 100 in 360°, or 0 to 1000 milliseconds.

Calibration

The calibration of a time or frequency standard by reference to radio time signals requires a series of measurements of the time of arrival of standard time signal pulses over an extended period of days or weeks in order to reduce the errors in time-signal reception times to negligible proportions. In the measurement the audio frequency dash or tick in the radio receiver output is compared with a timed reference signal from the local source to be calibrated. For example, let us assume that the standard time signals are received at a setting of 26.3 on the first day, 26.4 on the second — to 27.3 on the eleventh day. This represents an increase of the Microdial

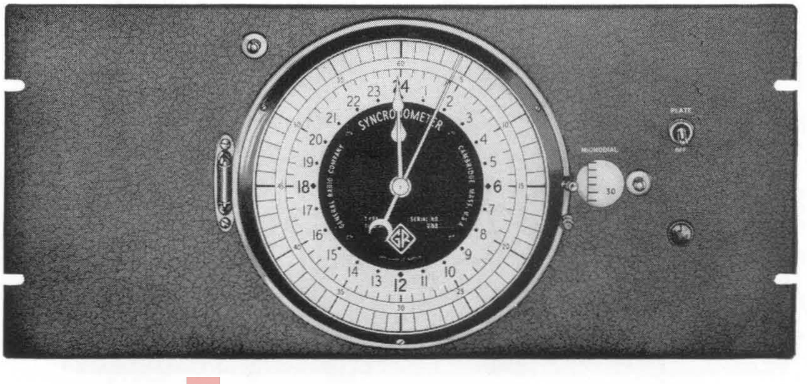


Figure 1. Panel view of the Syncronometer. The Microdial is at the right of the clock face.

reading of 1 division in 10 days or 10 milliseconds, which is 1 millisecond per day gain. Each day consists of a total of 86,400 seconds (mean solar time). This figure is approximately 10^5 seconds, or 10^8 milliseconds, per day. Since the frequency-standard clock is gaining at the rate of 1 millisecond per day, the clock-oscillator frequency (local frequency standard) is running approximately 1×10^{-8} too high in frequency, *and is constant in frequency.* If the Microdial setting had gained, or lost, time at a rate proportional to the square of the elapsed time, the *frequency* of the clock oscillator would have been changing at a *constant rate of increase* per day. Thus it is possible to compare local frequency standards with standard time signals for calibration of both frequency and time indications. In order to achieve the desired precision in the use of this method to calibrate the TYPE 1100-AP Primary Frequency Standard, it is necessary to use the Microdial contactor with an oscilloscopic indication of the time signal.

Use of the Microdial

As originally contemplated, the Microdial contactor was used as a phaseable gate, which let a time signal through when it was open and shorted it out when closed. The operator determined the time of arrival of the time signal by finding the setting of the Microdial for which the "nose" of the time signal was just barely perceptible, in head telephones or a loudspeaker. The precision

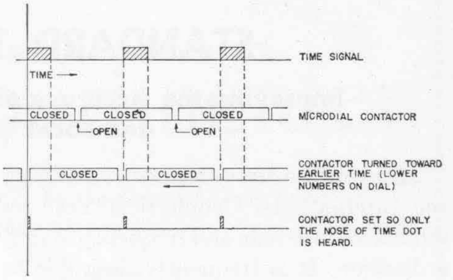


Figure 2. Diagram of time comparison using the Microdial.

of setting by this method is approximately ± 5 milliseconds. A time diagram of this operation is shown in Figure 2. For greater precision in this comparison method, a detector, or display device, is needed for indicating more precisely the relationship between the Microdial setting and the time of arrival of the signal. The aural method is subject to confusion from interfering noises and signals, and it depends upon the hearing of the operator for establishing a constant interval in milliseconds between the start of the time pulse and the closing of the Microdial contactor.

The Comparison Oscilloscope

A precision time-interval indicator is readily available in the TYPE 1105-A Frequency Measuring Equipment in the form of the TYPE 1109-A Comparison Oscilloscope, shown in Figure 3. The circular sweep of this cathode-ray indicator can be driven by the frequency standard at a sweep rate of 100 cycles per second, which gives a time scale of 10 milliseconds per revolution, with no

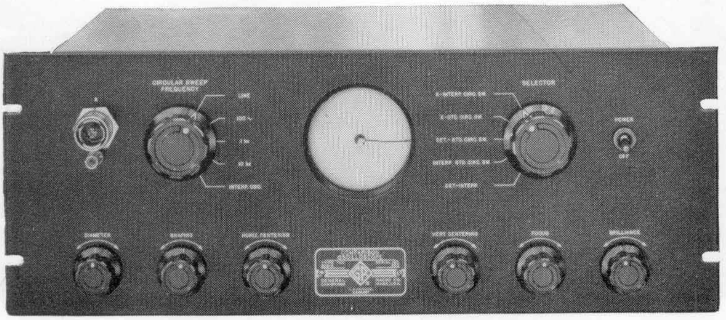


Figure 3. Panel view of the Type 1109-A Comparison Oscilloscope.

starting transient since it runs continuously. This oscilloscope thus provides the precision time scale required for accurate determination of the time of arrival of the time signals. All that is required is a knowledge of the characteristics of the time signals and of their appearance as seen on the oscilloscope, so that a suitable method may be chosen for using the display to calibrate the time of arrival of the time pulses.

Such a method has been in use here at the General Radio Company for several months with entirely satisfactory results. This method requires the following settings of the selector switches of the TYPE 1109-A Comparison Oscilloscope:

(CIRCULAR SWEEP FREQUENCY) to 100 cycles, and SELECTOR to DET.-STD. CIRC. SW. The radio-frequency time signal must then be received by a detector unit, which may be a TYPE 1106-A, -B, or -C Frequency Transfer Unit or a radio receiver. The resulting audio signal is supplied to the oscilloscope. In the TYPE 1105-A Frequency Measuring Equipment Assembly, this is accomplished by setting the switch on the TYPE 1108-A Coupling Panel to select the DETECTOR at L(ow), M(edium), or H(igh) frequency, as required, or EXT. (external) for a separate receiver. The audio signal will produce a radial deflection of the scope pattern. Adjustment of the Microdial setting will then vary the closing time of the gate so that, with the MICRODIAL switched ON (on the Coupling Panel), the start of the time signal can be seen, and the closing time of the contactor can be set to a selected constant time delay after the start of the time signal. Although time signals from different sources have slightly differing characteristics, it is interesting to note that it is possible to reduce many of them to a completely standardized display, as shown in Figure 4.

In this diagram, the constant time-delay is 3 milliseconds and the audio signal is 1000 cycles per second. These are convenient values and are used in the examples that follow. The choice of 1000 cycles for the audio tone is convenient because (1) it is the modulation tone on the time signals from WWV, and (2) it bears an integral relationship to the 100-cycle circular sweep frequency, thus assuring a stationary pattern. The 3-millisecond delay interval is chosen because (1) the Microdial contactor must short out at least a portion of the time-signal pulse or there will be no way to tell when it is set correctly, and (2) the 3-cycle (or 3 millisecond) tick which remains is long enough to give an audible pulse, which can be distinguished aurally from most impulse noise, and (3) it provides a long enough period for the signal to build up if filters are added to reduce noise.

THE CALIBRATION METHOD IN USE

NSS Time Signals (U. S. Naval Observatory)

The time signals controlled by the U. S. Naval Observatory are transmitted from stations NSS, NBA, and NPG on a variety of frequencies (see current listing of frequencies and transmission schedules in U. S. Naval

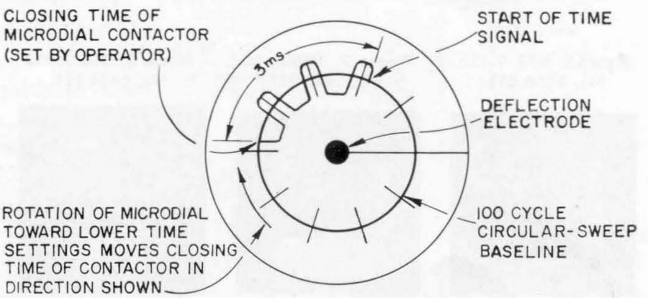


Figure 4. Diagram of time-signal display on Type 1109-A Comparison Oscilloscope with Microdial contactor set to close 3 milliseconds after start of time signal.

Observatory bulletins). These transmissions consist of a series of dashes at one-second intervals, each dash being a keyed continuous-wave signal, unmodulated. These time signals are radiated during the five minutes preceding the hour at the scheduled time.

For the display of this type of time signal on the circular-sweep scope, a heterodyning frequency must be added to provide an audio-frequency beat tone. For ease of calibration, this beat should be set as near to 1 kc as possible by reference to the pattern on the circular-sweep scope with the Microdial contactor not operating. Each cycle of the beat frequency is then equal to one millisecond, but the exact phasing of the various cycles is usually not steady enough to be used as a direct time calibration. The display presented by this method appears identical to that of the modulated signal from WWV (see next section) when the Microdial contactor is operating with a 3-millisecond interval between "nose" and closing time of the contactor. A photograph of the oscilloscope display of the time signal from NSS is shown in Figure 6. Note the apparently exact duplication of the display of Figure 7 with the notable exception that the signal-to-noise ratio is considerably better in the case shown in Figure 6. The heterodyning frequency was provided by the heterodyne frequency-meter oscillator in the Type 1106-B Frequency Transfer Unit. A Hammarlund SP-600-JX receiver was used as an external detector.

WWV Time Signals (National Bureau of Standards)

The time pulses from WWV comprise five complete cycles of 1-kc modulation on the carrier of the standard-frequency transmission, as shown in Figure 5. During the last two minutes of most of the five-minute intervals in any given hour (except for the various interruptions scheduled — see the current bulletin of National Bureau of Standards), the carrier frequency and the time ticks are the only signals transmitted. The Microdial contactor remains open for approximately 50 milliseconds, or 5 revolutions of the circular sweep. Therefore, the baseline of the sweep is visible throughout the entire circle even though the deflection leaves no baseline during the pulse. If the contactor is not set to short out part of the time signal pulse, the display is similar with the Microdial contactor in or out of operation, except that there is more noise and a brighter baseline display when the contactor is not in use.

When the phasing of the contactor is adjusted close to the desired value, the counter-clockwise end of the 5-cycle pulse will be shorted out first if the Microdial is rotated

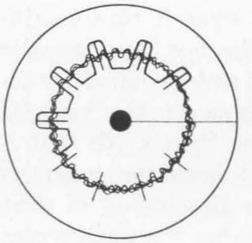


Figure 5. Diagram of time signal from WWV with Microdial contactor not operating.

from the higher numbers toward the lower numbers on the dial, as shown in Figure 2. If the Microdial is set to an improper setting, the contactor will always be closed when a time pulse is received and no deflection will be displayed.

One of the advantages of the use of this method of time-signal reception-time calibration is that relatively accurate results can be obtained in the presence of interference strong enough to prevent use of the carrier frequency of WWV as a standard frequency. A photograph of the Type 1109-A Comparison Oscilloscope display taken under such conditions is shown in Figure 7. The interference was a combination of several signals, at least one of which was an experimental pulse-train generator in an adjoining laboratory. The receiver was the detector in the Type 1106-B Frequency Transfer Unit.

CHU Time Signals (Dominion Observatory, Ottawa, Canada)

Time signals are transmitted by the Dominion Observatory, Ottawa, Canada, by the keying of a 1000-cycle tone on the carrier of station CHU. The signals are long dashes of 1000-cycle signals once each second, the time being announced in voice after the fiftieth second of each minute. The 1000-cycle modulating frequency is a standard frequency, and the keying device selects a constant starting phase for the modulating pulse. Hence each modulation cycle represents one millisecond. The display of this signal is again practically identical with the preceding two examples, as shown by the oscilloscope photograph of Figure 8. Some interference is apparent in this photograph, the beat note shown arising from a commercial communications transmitter on an adjacent channel. This signal was received on the Type 1106-B Frequency Transfer Unit, a regenerative detector. A sketch of the oscilloscope display during the "on" period of the CHU time signal is shown in Figure 9, the Microdial contactor being switched off. Familiarity with this display enables easy recognition and tuning-in of the time signal.

Figure 6. NSS, 9.425 Mc, 1456 EST.

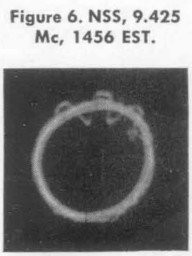


Figure 7. WWV, 10 Mc, 1437 EST.

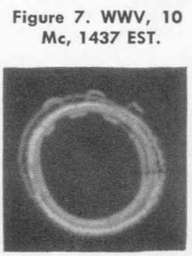
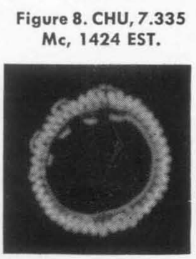


Figure 8. CHU, 7.335 Mc, 1424 EST.



Photographs of time-signal displays on Type 1109-A Comparison Oscilloscope, at Cambridge, Mass., December 19, 1957. Receiver was the Type 1106-B Frequency Transfer Unit except for Figure 6 where Frequency Transfer Unit supplied heterodyning frequency for SP-600-JX receiver.

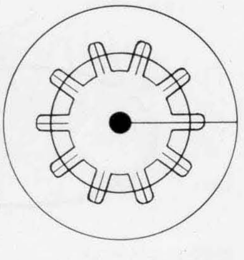


Figure 9. Diagram of display during "on" period of time signal from CHU (Dominion Observatory, Ottawa, Canada), showing 10 equally spaced intervals of 1 millisecond each, corresponding to the modulation frequency of 1000~ (100~ scope sweep frequency). Microdial switched OFF. The phasing shown here is arbitrary.

Calibrated Scale on Oscilloscope

A useful addition to the Type 1109-A Comparison Oscilloscope is a calibrated scale engraved on a sheet of transparent plastic material, which is placed next to the face of the oscilloscope tube. This scale, shown in Figure 10, has 10 equiangular divisions and a complete circle, which is used as a target on which to align the circular sweep.

Each tenth of the circular sweep circle represents one actual millisecond of time as counted off by the frequency standard, so that this scale can be used directly as a millisecond vernier scale to read between the 10-millisecond calibration points on the Microdial. Since the sweep voltage is not adjustable in phase for synchronism with the calibrated divisions of the Microdial, an arbitrary zero point must be established.

Use of Selective Filters

When electrical noise and interference are heavy, the use of selective filters may improve the signal-to-noise ratio. Filters can also be used to select the time-signal component of the WWV transmission during transmission of the standard audio tone signals. In each case, it is essential to determine the time-delay introduced by the use of the filter in order to be able to remove the additional error from this source. This problem is important mainly in the establishment of accurate time for time-of-occurrence measurements, since the time-delay of a given filter would subtract out in a time-interval measurement such as is used in frequency standardization.

The circular-sweep oscilloscope display provides a simple means for checking the filter time-delay. If the time signals can be received without the filter, then a quick check of the display time with and without the filter switched into the circuit gives the desired delay-time calibration directly.

For example, the crystal filter of the SP-600-JX receiver in use in our frequency standard room appears to introduce approximately 0.5 millisecond delay when the selectivity is switched from 3 kc (xtal out) to 1.5 kc (xtal in). This crystal filter can be used to suppress the single-sideband tone transmission of WWV, leaving the time signals in the clear during the entire five-minute period.

The use of a selective audio filter ahead of the Microdial contactor may result in some noise-pulse-induced ringing of the filter. This condition is improved slightly by placement of the audio filter in the circuit following the Microdial contactor, since the number of large noise pulses coming through the open "gate" is less than the total number of noise pulses ahead of the "gate."

Discussion of Results of Use of This Method of Calibration

The dial readings of the Microdial can be estimated to 0.1 division of the dial. The accuracy of the estimate de-

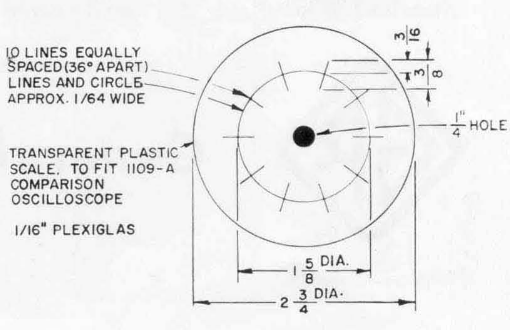


Figure 10. Diagram of transparent scale used on Type 1109-A Comparison Oscilloscope.

pends upon the use of a standard direction of approach to the setting point and the use of the vernier scale provided by the oscilloscope as described above. In general, the dial readings made by this method have exhibited a consistency of better than ± 1 millisecond. This degree of reliability is adequate, since the time signals show a propagation-caused variation sometimes as great as ± 1 millisecond. In general, the stability of the arrival time of the time signals is better than ± 0.3 millisecond.

Over the past few months, calibration of the General Radio working frequency standard has been carried out by this method with a precision entirely adequate for precise frequency measurement of ± 1 x 10^{-8} without correction for the variations in the transmission time of the time signals. By taking account of the corrections provided by the U. S. Naval Observatory, Washington, D. C., it is possible to improve this figure. In any case, this calibration method provides a local time standard, independent of radio propagation conditions, with a simple, accurate, checking method for use when radio propagation conditions are favorable, to allow direct calibration by comparison with externally available standard time signals.

An "Emergency" Method

The display method described above assumes possession of a complete TYPE 1100-AP Primary Frequency Standard with TYPE 1105-A Frequency Measuring Equipment Assembly. When only the TYPE 1100-AP Primary Frequency Standard is available, an oscilloscopic display can be made, independent of the circular sweep, by a setting of the Microdial to chop off the required part of the time signals as shown on a conventional oscilloscope with, for example, a 60-cycle sine-wave sweep on the horizontal plates and a small amount of 60-cycle sine wave added to the vertical deflection signal along with the time-signal input. The "nose" of the time pulse will then be visible, as it is on the circular sweep display with standard sweep rate, the difference being that there is no accurate vernier "gain-or-loss of time" scale on the scope. The modulated time signals from WWV and CHU permit easier use of this display than do the c-w pulse type of signals. A sketch of an oscilloscope display of this "emergency" display method is shown in Figure 11.

— FRANK D. LEWIS

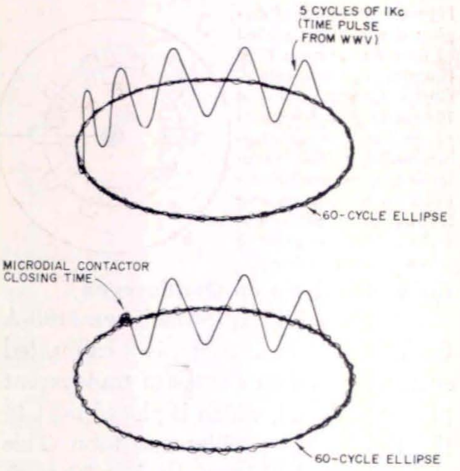


Figure 11. Diagram of oscilloscope display on conventional oscilloscope, showing 3-millisecond interval of modulated pulse between "nose" of pulse and closing of contactor.

the GENERAL RADIO Experimenter

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Photograph Courtesy Underwood Corporation

In This Issue

Standard Frequency Multipliers
Measuring Impact Noise

the GENERAL RADIO Experimenter

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CONTENTS

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COVER

Measuring typewriter noise with the General Radio Impact Noise Analyzer. See article on page 10.

NEW STANDARD FREQUENCY MULTIPLIERS

The Type 1112 Standard Frequency Multipliers provide sine-wave signals of 1, 10, 100, and 1000 megacycles when driven from a 100-kilocycle source. Thus they greatly extend the useful range of conventional crystal-controlled frequency standards, such as the General Radio Type 1100-A, and facilitate accurate measurement of microwave frequencies. These multipliers are characterized by low noise and by almost complete freedom from submultiple-frequency spurious signals. In addition, the phase stability of the output signal at each desired carrier frequency is maintained at a high value instead of being rapidly degraded as it may be in some conventional multiplier circuits.

The multiplier chain consists of two units, the first providing 20 milliwatts at 1, 10, and 100 megacycles from three phase-locked quartz-crystal oscillators; the second, 50 milliwatts at 1000 megacycles from a phase-locked klystron oscillator. The input to the first unit, Type 1112-A, is normally 100 kilocycles, but alternatively, 1, 2.5, or 5 megacycles[1] can be used. The second unit, Type 1112-B, is driven from the 100-megacycle output of the first unit. The input frequency can vary a few parts in 10^6 from the nominal value without loss of control. The multiplier stages can be manually detuned on either side of the nominal standard frequency by an additional few parts per million if desired, provided that the operation of the multiplier is carefully monitored to prevent improper operation. Figures 1 and 2 are panel views of the two units.

Principle of Operation

The underlying principle on which the phase-stability and noise-reduction properties of these multipliers are based is the use of a narrow-band filter to select only the desired output harmonic at each output frequency. The filters used at the three lowest frequencies are quartz crystals since they afford the highest possible Q, and hence narrowest bandwidth, in this frequency range. In order to maintain these crystal filters at the correct resonant frequency to pass the desired harmonic, each crystal is incorporated in an oscillator circuit whose frequency is phase locked to the desired harmonic frequency by an automatic-phase-control loop.

[1]When 2.5 Mc or 5-Mc input is used, the 1-Mc output cannot be used.

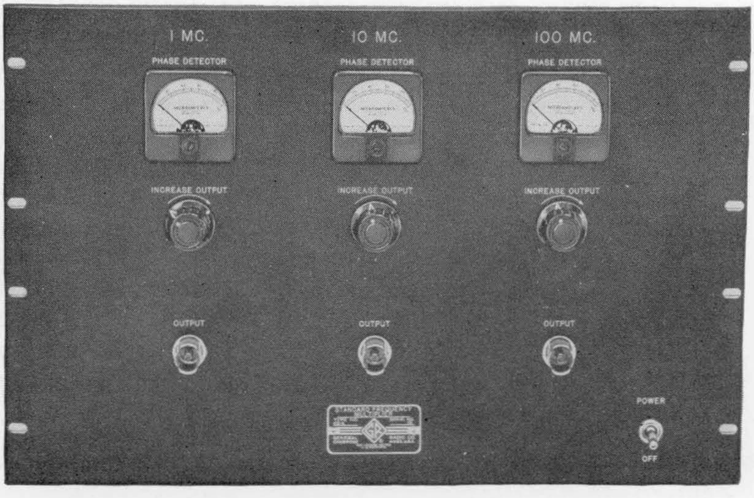


Figure 1. Panel view of Type 1112-A Standard Frequency Multiplier.

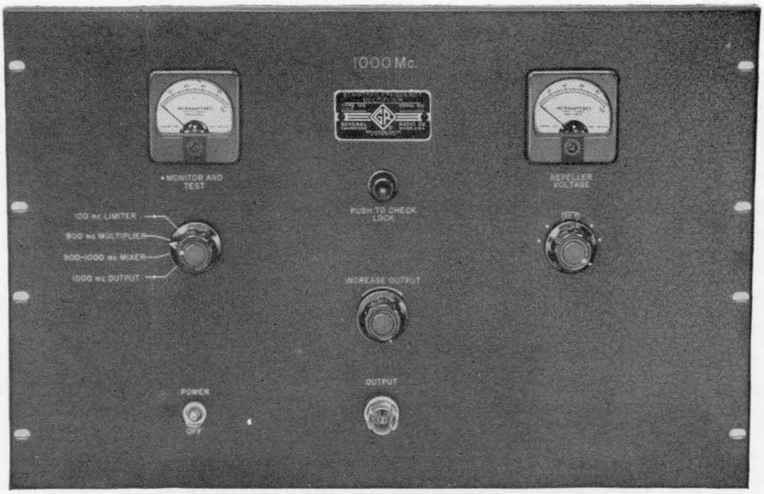


Figure 2. Panel view of Type 1112-B Standard Frequency Multiplier.

At the 1000-Mc output frequency, a phase-locked klystron oscillator is used as a selective filter to eliminate unwanted harmonics of the control frequency, thus operating in much the same manner as the locked crystal oscillators at lower frequencies. Since, however, the Q of the klystron resonator is not extraordinarily high, the phase modulation noise inherent in klystrons is reduced by means of negative feedback. The automatic-phase-control loop for the 1000-Mc klystron feeds back phase noise over a wide frequency band to reduce phase instability, the reference standard in this case being taken as the multiplied harmonic of the crystal-controlled 100-Mc driving signal.

Block Diagrams of Multiplier Units

The operation of the multiplier chain units can be understood easily with the aid of the block diagrams of Figure 3 (Type 1112-A) and Figure 4 (Type 1112-B).

The Type 1112-A Standard Frequency Multiplier receives a driving signal from the 100-kc frequency standard and multiplies it to 1000 kc. The signal from the 1000-kc crystal oscillator is then compared with the 1000-kc multiplied signal in a tuned, balanced, phase detector. The output signal from the phase detector is a d-c control voltage, which is applied to the grid of a reactance tube connected to the crystal-oscillator circuit. When the crystal-oscillator frequency is adjusted close to the frequency of the 1-Mc harmonic of the standard frequency, the phase-detector signal drives the reactance tube to the proper value of reactance to synchronize the crystal frequency exactly to that of the standard-frequency harmonic.

A small residual phase error remains, but, as long as the crystal-oscillator frequency is within the lock-in range of the system, the crystal oscillator stays locked directly at the harmonic frequency of the standard. The phase error provides the controlling voltage which holds the crystal oscillator in lock, or, putting it in different terms, the servo loop is closed with a small static error. If the crystal tends to drift, the phase error changes, and the change in the controlling voltage readjusts the reactance tube to hold the phase error to a minimum. Since the crystal has a high Q, the instantaneous phase stability of the crystal oscillator is high, and the locked oscillator then has good short-term stability, while its long-term stability is identical with that of the frequency standard used as the driving source.

If the servo circuit, including the reactance tube, phase detector, and reference harmonic generator, introduces noise into the reactance-tube grid circuit, frequency- or phase-modulation noise can be generated in the crystal-oscillator signal. The bandwidth of the feedback signal is kept narrow to minimize this noise. Thus a clean, crystal-controlled signal is available at the exact harmonic frequency desired.

In Figure 3, the functional diagram of the 10-Mc and 100-Mc stages is essentially identical with that of the 1-Mc stage. There are only minor differences in the circuits to take care of the different operating conditions at the different frequencies. A balanced phase detector is used at 10 Mc, but an unbalanced phase detector is used at 100 Mc to simplify the circuit. The 100-Mc crystal-oscillator circuit uses a fifth-overtone-mode crystal, and requires circuit refinements to insure stable operation at the fifth overtone.

The Type 1112-B Standard Frequency Multiplier provides output at 1000 Mc from a locked klystron oscillator. In order to prevent unnecessary conversion of amplitude-modulation noise to frequency modulation, several departures from the straightforward arrangement of the lower frequency stages are incorporated.

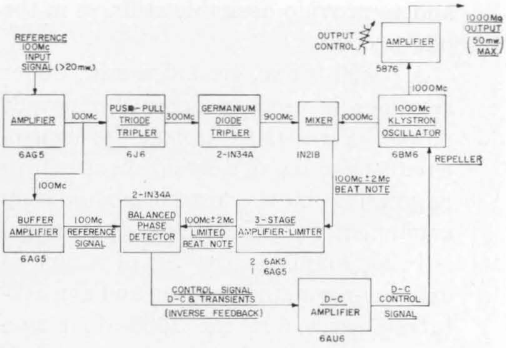


Figure 4. Block diagram of Type 1112-B Standard Frequency Multiplier, 100 to 1000 Mc.

The frequency multiplication is obtained by multiplying 3 x 3 and adding 1 in order to obtain 10 times the input frequency. This circuit arrangement permits the use of small receiving-type vacuum tubes up to the 1000-Mc stages and makes possible the introduction of a limiter to insure a constant drive level for the phase detector.

The reflex klystron has a built-in "reactance tube" in its repeller, which allows the operating frequency and phase to be adjusted by variation of the repeller voltage. A d-c amplifier is incorporated in the repeller circuit to isolate the phase detector from the repeller

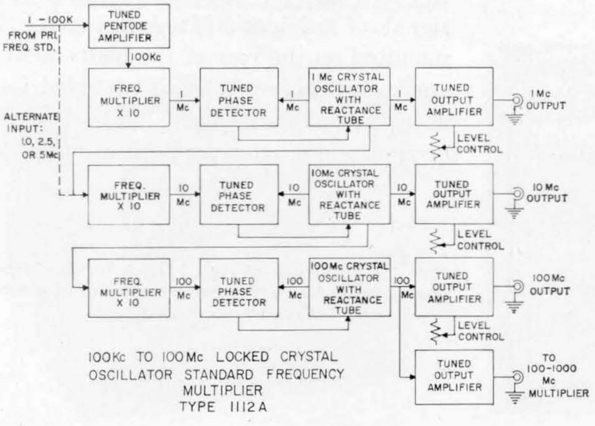


Figure 3. Block diagram of Type 1112-A Standard Frequency Multiplier, 100 kc to 100 Mc.

and to provide desirable stiffness in the phase lock.

A pencil-triode, grounded-grid, buffer amplifier is used to raise the output power level and to protect the control circuit from outside signal disturbances originating in the external measuring equipment to which the output signal is being supplied. The use of a plug-in external-resonator klystron and a pencil-tube amplifier in the 1000-Mc stages keeps tube replacement cost down. D-C heater power is supplied by a rectifier to the klystron to reduce hum modulation, and regulated plate supplies are used throughout.

Performance

Output power from each of the amplifiers at 1, 10, and 100 Mc is a maximum of 20 mw into a 50-ohm load. At 1000 Mc a maximum of 50 mw is available into 50 ohms.

The spurious signals at harmonics of the lower-frequency control signals are all at least 100 db below the desired output signal, except for higher harmonics of the desired signal. This means that a signal at 1000 Mc, for example, is not accompanied by a family of 100-kc or 1-Mc sidebands, unless they are specifically added by external mixing. For many measurements, such as marker generation applications, this feature alone is a great time saver and sometimes makes the difference between a practical measurement setup and an impractical one. The power levels directly available are adequate for a large percentage of measurement applications. The phase jitter, or phase-modulation noise, is low, being equivalent to that of a free-running crystal oscillator at 1, 10, and 100 Mc. The output signal at 1000 Mc appears to have the same stability as the crystal-controlled harmonic of the 100-Mc control signal supplied at the input connection. The amplitude modulation of the output signals is likewise low, as a result of the electronically regulated power supplies and the use of high-Q tuned circuits for the oscillator stages.

Design Features

The Type 1112-A Standard Frequency Multiplier consists of three sections, each of which receives a driving frequency and emits an output signal at a harmonic of this input frequency. A rear view of this instrument, Figure 5, shows the manner in which the circuits are constructed. The three multiplier sections are arranged in three horizontal rows of shielded compartments. The top row receives the input signal at 100 kc and supplies an output signal at 1000 kc. The center row gives 10-Mc output, and the bottom row 100 Mc. The input connection is normally attached to the input amplifier in the top row, at the upper right-hand corner, but may be moved to the input amplifier of the second row at the left side for operation with input signals of 1, 2.5, or 5 Mc. A test meter is mounted on the rear of the instrument, and a series of switches is provided for energizing either the harmonic multiplier or crystal oscillator, or both at once, in

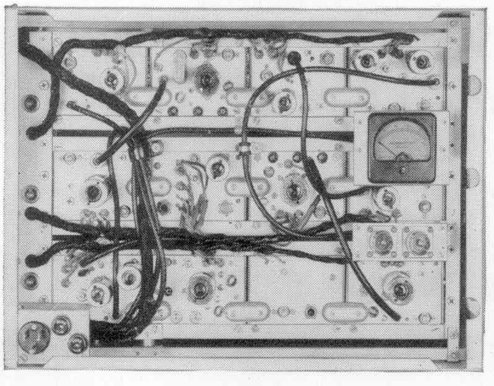


Figure 5. Rear view of Type 1112-A Standard Frequency Multiplier showing arrangement of stages in three horizontal rows.

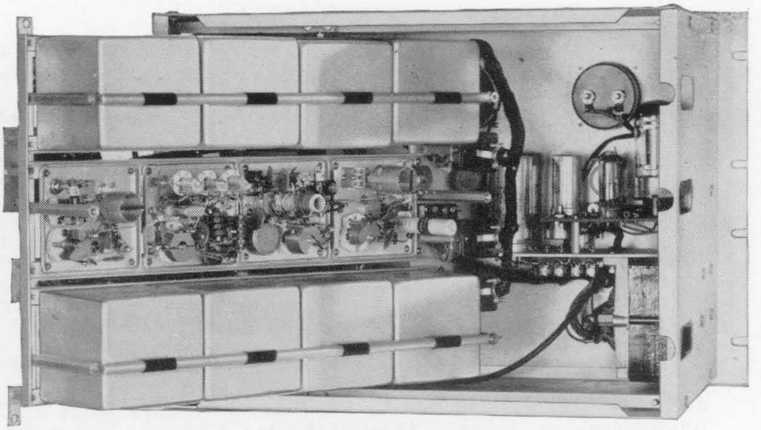
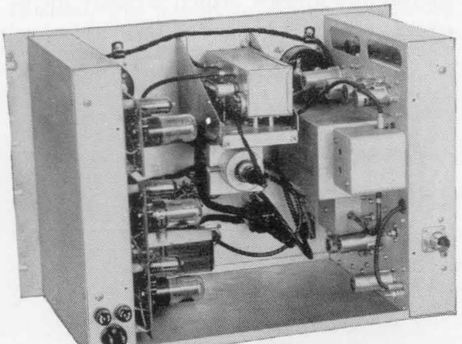


Figure 6. View of Type 1112-A Standard Frequency Multiplier with chassis assembly swung open for maintenance, covers removed from 1 to 10-Mc multiplier stages.

each row. These switches are visible at the left side of the rear view. The entire assembly swings out on hinges to allow access to the components in each compartment (Figure 6). The meters on the front panel are permanently connected in one side of the balanced phase-detector circuits for monitoring operating level and for indicating loss of lock.

The Type 1112-B Standard Frequency Multiplier is constructed with vertical chassis mounting of the power supply components and control circuit elements, the klystron and associated buffer amplifier being mounted in a removable subassembly or "r-f head." A rear view of this instrument, with the shielding partially removed, is shown in Figure 7. The r-f head in the center of the unit is removable to facilitate replacement of the r-f amplifier tube and to allow easy replacement of other tubes in the side-mounted chassis assemblies. A view of the r-f head, with the cover plate of the amplifier removed, shows the grounded grid amplifier using a pencil triode (Figure 8). This view also shows the detuning "button" for introducing a small deviation in the frequency of the klystron resonator to check for locking of the oscillator. It is also possible to mount this r-f head assembly on an external "storage" resonator to improve the effective Q of the klystron resonator, if such an application is ever considered necessary.

Figure 7. Rear view of Type 1112-B Standard Frequency Multiplier showing vertical chassis construction and r-f head (center).



Applications

Two separate and distinct applications have been kept in mind during the development of this equipment. The first, and most obvious, is the generation of stable, low-noise, microwave, standard frequencies free from unwanted sideband frequencies, which are often present in multiplier chains of more conventional design. However, in order to generate the desired marker frequencies or microwave standard frequencies for various measurement purposes, it is sometimes necessary to mix or add signals of different frequencies.

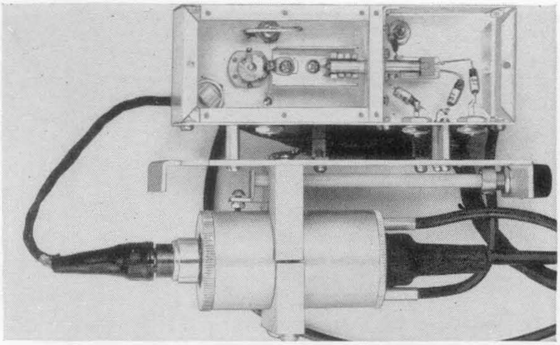


Figure 8. R-F head of Type 1112-B Standard Frequency Multiplier showing klystron installed in resonator (below) and grounded-grid 1000-Mc amplifier with cover removed.

The output circuits of the 1-, 10-, and 100-Mc stages of the TYPE 1112-A Standard Frequency Multiplier contain coupling networks that allow all three of these circuits to be connected in parallel without short circuiting each other. This arrangement is indicated in Figure 9, which also indicates a possible method of adding the 1000-Mc signal from the TYPE 1112-B Standard Frequency Multiplier. Another, and perhaps better, method of adding the 1000-Mc signal is indicated in Figure 10. In this arrangement, the signal from the low-frequency unit is added to that from the high-frequency unit in a tee, the branches being isolated by mismatching. A line stretcher in each branch allows adjustment of the mismatch for maximum isolation. By use of these paralleling schemes, it is possible to generate a marker frequency or harmonic series of marker frequencies at will in the microwave system. For some applications, additional power will be desirable, but for many uses the output of the multipliers will be adequate directly.

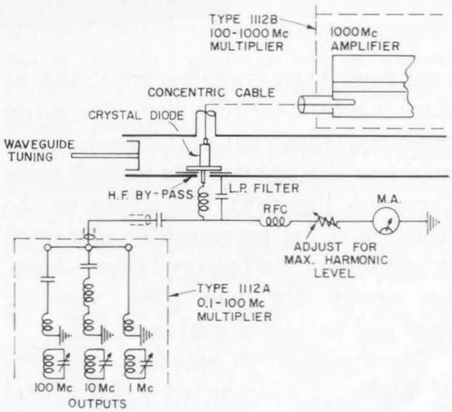


Figure 9. Schematic diagram showing parallel connection of 1, 10, and 100-Mc output circuits for application to generation of microwave harmonics. The 100-Mc signal is fed in separately.

The second application is in the intercomparison of standard-frequency oscillators for stability measurements. For example, a pair of 100-kc oscillators may be compared at 100 Mc, the frequency variations between them being then multiplied by 1000. A frequency difference of one part in 10^8, for instance, would appear as a one-cycle difference at 100 megacycles, which is easily measurable.

— FRANK D. LEWIS

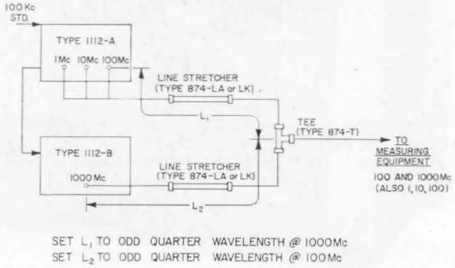


Figure 10. Arrangement for paralleling output signals of Type 1112-A and Type 1112-B Standard Frequency Multipliers using coaxial-line adding network.

SPECIFICATIONS

TYPE 1112-A STANDARD FREQUENCY MULTIPLIER

Input: 1 volt, 100-kc sine wave from standard-frequency oscillator. Can also be driven at input frequencies of 1, 2.5, and 5 Mc. Will run free with no input signal, but absolute frequency will be in error by several parts per million unless standardized.
Output: Four channels; one each of 1 Mc and 10 Mc, and two of 100 Mc; all sine wave; all 50 ohms; 20 milliwatts, max., into 50 ohms.
Terminals: TYPE 874 Coaxial Connectors; adaptors are available to fit all commonly used connector types. See *Experimenter* for March, 1958.
Power Supply: 105 to 125 (or 210 to 250) volts, 50 to 60 cycles, 100 watts. Power input receptacle will accept either 2-wire (Type CAP-35) or 3-wire (Type CAP-15) power cord. Type CAP-35 2-wire cord is furnished. Type CAP-15 3-wire cord can be purchased separately at $2.25.
Mounting and Dimensions: Relay-rack panel, 19 x 12¼ inches; over-all depth, 11½ inches.
Net Weight: 25 pounds.

TYPE 1112-B STANDARD FREQUENCY MULTIPLIER

Input: 20 milliwatts, 100 Mc, sine wave from TYPE 1112-A Standard-Frequency Multiplier; 50-ohm input impedance.
Output: 1000-Mc sine wave; 50 mw into 50-ohm load; 50-ohm output impedance.
Net Weight: 35 pounds.

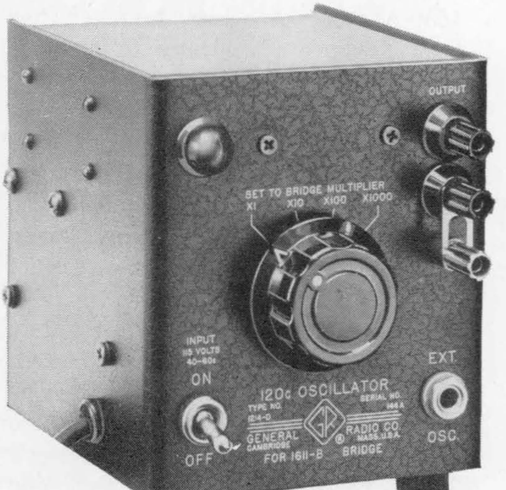
Other specifications are identical with those for TYPE 1112-A, above.

Type		*Code Word*	*Price*
1112-A	**Standard Frequency Multiplier**	EPOCH	**$1450.00**
1112-B	**Standard Frequency Multiplier**	EPODE	**1360.00**

TYPE 1214-D UNIT OSCILLATOR

The TYPE 1214-AS2 Unit Oscillator, which has been supplied for use with the TYPE 1611-B Capacitance Test Bridge[1] in the measurement of electrolytic capacitors at 120 cycles, has now been given the type number 1214-D. Specifications, which were originally published in the August, 1956, issue of the *Experimenter*, remain unchanged, and are reprinted below.

[1] "Capacitance Test Bridge," *General Radio Experimenter*, 32, 9, February, 1958, pp. 6-8.



SPECIFICATIONS

Frequency: 120 cycles ±2%.
Output Impedance: Four impedances to match the impedance of the TYPE 1611-B Capacitance Test Bridge at four multiplier positions.
Output: At least 2000 mw into matched load.
Distortion: Less than 3% into a matched load.
Terminals: The output terminals are jack-top binding posts with standard ¾-inch spacing; a ground terminal is provided, adjacent to one of the output terminals. Jack is provided for connecting external oscillator.
Power Supply: Unlike most instruments of the Unit line, the power supply is built into the instrument; 115 volts, 40–60 cycles; power consumption is about 16 watts.
Accessories Supplied: Spare fuses; the power cord is integral with the unit.
Tube: One 117N7-GT, which is supplied with the instrument.
Mounting: Aluminum panel and sides finished in black-crackle lacquer. Aluminum dust cover finished in clear lacquer. Relay-rack adaptor panel available.
Dimensions: (Height) 5¾ x (width) 5 x (depth) 6¼ inches, over-all, not including power-line connector cord. **Net Weight:** 4½ pounds.

Type		*Code Word*	*Price*
1214-D	**Unit Oscillator**	ABBOT	**$100.00**

SHORT DECAY-TIME IMPACT-NOISE MEASUREMENT

Perhaps the greatest causes of noise in offices (other than people) are typewriters, calculators, and other business machines. Aware of this, Underwood Corporation's General Research Laboratory in Hartford, Connecticut, has, for over twenty years, carried on a program of measurement and interpretation of business-machine noises.

Noise from a typewriter is made up of a series of short-duration bursts of sound. Continuous spectrum and band-spectrum analyzers are not suited to the measurement of these extremely short impact noises. The General Radio TYPE 1556-A Noise Analyzer, however, has been designed specifically to measure this type of noise.

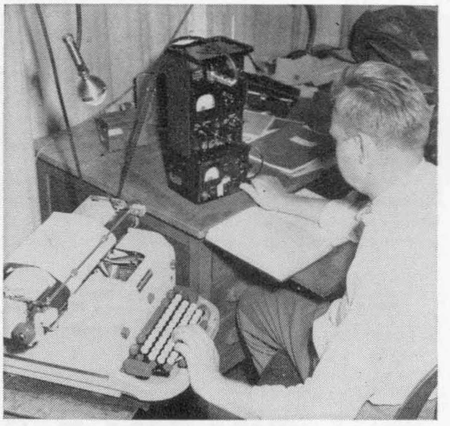


Figure 1. Setup used by Underwood for noise analyses of their "Golden Touch" typewriters. The transients are picked up by the suspended microphone, amplified by the Type 1551-A Sound-Level Meter, and then fed into the Type 1556-A Impact-Noise Analyzer.

The decay times of the impact noises from typewriters are very short, and, consequently, the Underwood Corporation found it convenient to modify the Impact-Noise Analyzer to have lower time constants for the time-averaging circuit.

The modification for short decay-time measurement is easily accomplished. The time constant of the analyzer is determined by an R-C circuit whose series resistance is set with the analyzer's TIME CONSTANT switch. Any one of seven different resistances can be selected to provide charging times from 2 milliseconds to 0.2 second. The recommended modification procedure is to change the capacitive element to a lower value. For example, if this capacitance is halved, all the time constants of the circuit are divided by 2, or if the value is reduced by a factor of 5, the time constants are then divided by 5.

Figure 1 shows an over-all view of the measuring setup used at the General Research Laboratories of the Underwood Corporation. The ease and rapidity of measurement possible with the TYPE 1556-A Impact-Noise Analyzer permitted extensive investigation into the nature of impact noises. Measure-

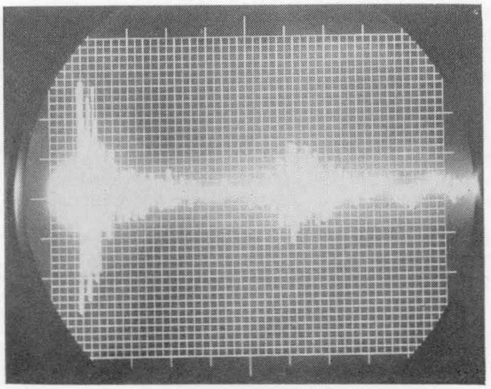


Figure 2. Oscillogram of typical typewriter noise. First impact occurs when bar strikes the platen — peak value is approximately 115 db and lasts about 3 milliseconds. A second peak occurs some 50 msec later when type bar returns to its rest position.

ments could be made quickly and conveniently to ascertain the effects of various typewriter modifications. No elaborate test equipment other than the Impact-Noise Analyzer and the Sound-Level Meter were required for preliminary measurements. Oscilloscopes, tape recorders, and other test equipment were necessary only when detailed investigations were required (Figure 2).

Underwood Corporation found the Impact-Noise Analyzer well suited to their measurement needs and, with its help, were able to make their "Golden Touch" typewriter the quietest that they have ever built.

VARIAC® USED IN THE JETCAL ANALYZER AND TESTER

In jet aircraft, exhaust-gas temperature and engine speed are vital to best engine life, efficiency, and safe operation. Engine temperature and engine speed must be maintained within close limits and be indicated accurately on the pilot's cockpit instruments during flight. To test and to calibrate the systems that measure and indicate these quantities, the B & H Instrument Company of Fort Worth have developed the Jetcal Analyzer and Tester.

The Jetcal is a rugged, portable instrument, which contains various check circuits, potentiometer, temperature regulator, meters, switches, and the necessary probes, cables, and adaptors for performing all tests. For convenient movement from one aircraft to another along a flight line, the Jetcal has wheels, adjustable handle, and support post.

Unique features of the Jetcal are: (1) It is a precision instrument of laboratory accuracy that is taken to the aircraft, performing its functions anywhere the aircraft is parked; (2) the Jetcal test system is isolated from the aircraft's system and thus provides means for checking and calibrating engines and cockpit instruments free of errors or limitations in the aircraft systems.

The Jetcal is used to determine the accuracy of the aircraft exhaust gas temperature system without the engine running and to read engine speed accurately during engine run-up. In checking the EGT system, the Jetcal heater probes apply precisely measured heat to thermocouples in the engine tail pipe.

The Type V-20 Variac® Autotransformer is an important component of every Jetcal Analyzer and Tester — the Variac is the Jetcal's temperature regulator. It controls the temperature of

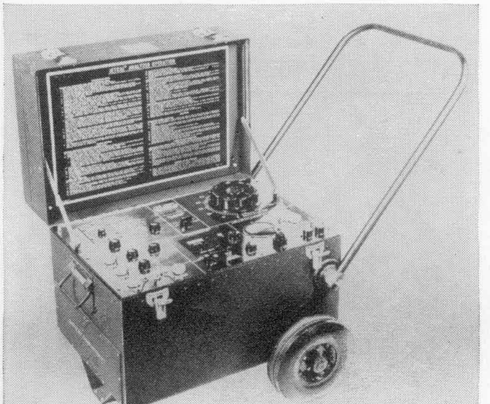


Figure 1. View of the Jetcal Analyzer and Tester. Wheels and handle permit the instrument to be moved conveniently to the aircraft to be tested.

resistance heaters in the heater probes, which apply accurately measured heat to jet engine thermocouples (or to other thermal systems).

Depending upon line voltage available, the Variac will vary the temperature of the Jetcal's wire-wound resistance heaters from 0 to approximately 900 degrees centigrade.

The Jetcal operates on any 95-to-135 volt, 50-400 cycle, AC power supply, in temperatures of −54° C. (−65° F.) to 71° C. (160° F.).

Since the aircraft exhaust gas temperature system is functionally checked without the engine running, the Jetcal achieves important savings — savings of fuel, savings of maintenance man-hours, savings of engine operating life. Here's just one example of savings: Prior to Jetcal, one major aircraft manufacturer had to make an average of three flights — at an average $752.57 per hour — to check a jet plane's cockpit indicator of tail pipe temperature. With the Jetcal, the EGT system is calibrated to within ± 4° C., without flight or engine running.

If test by Jetcal shows that an engine's EGT system is not functioning within allowed tolerance, the Jetcal will trouble-shoot and isolate errors in the system — still without the engine running.

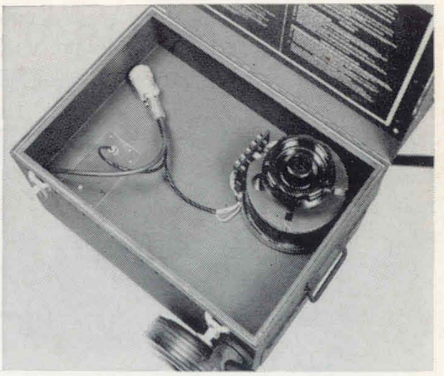


Figure 2. View of interior of Jetcal, showing the Type V-20 Variac® Autotransformer in position.

The Variac was selected for the Jetcal by B & H design engineers because it fully meets B & H standards to assure reliability and durability in sustained field use.

Jetcals, with Variacs and other highest quality components, have been proved in world-wide operational use. More than 3600 Jetcals are being used by the United States Air Force, Army and Navy, the NATO forces, jet engine and aircraft manufacturers, and airlines.

The August and September numbers of the *Experimenter* will be combined in a single issue, to be published about September 1.

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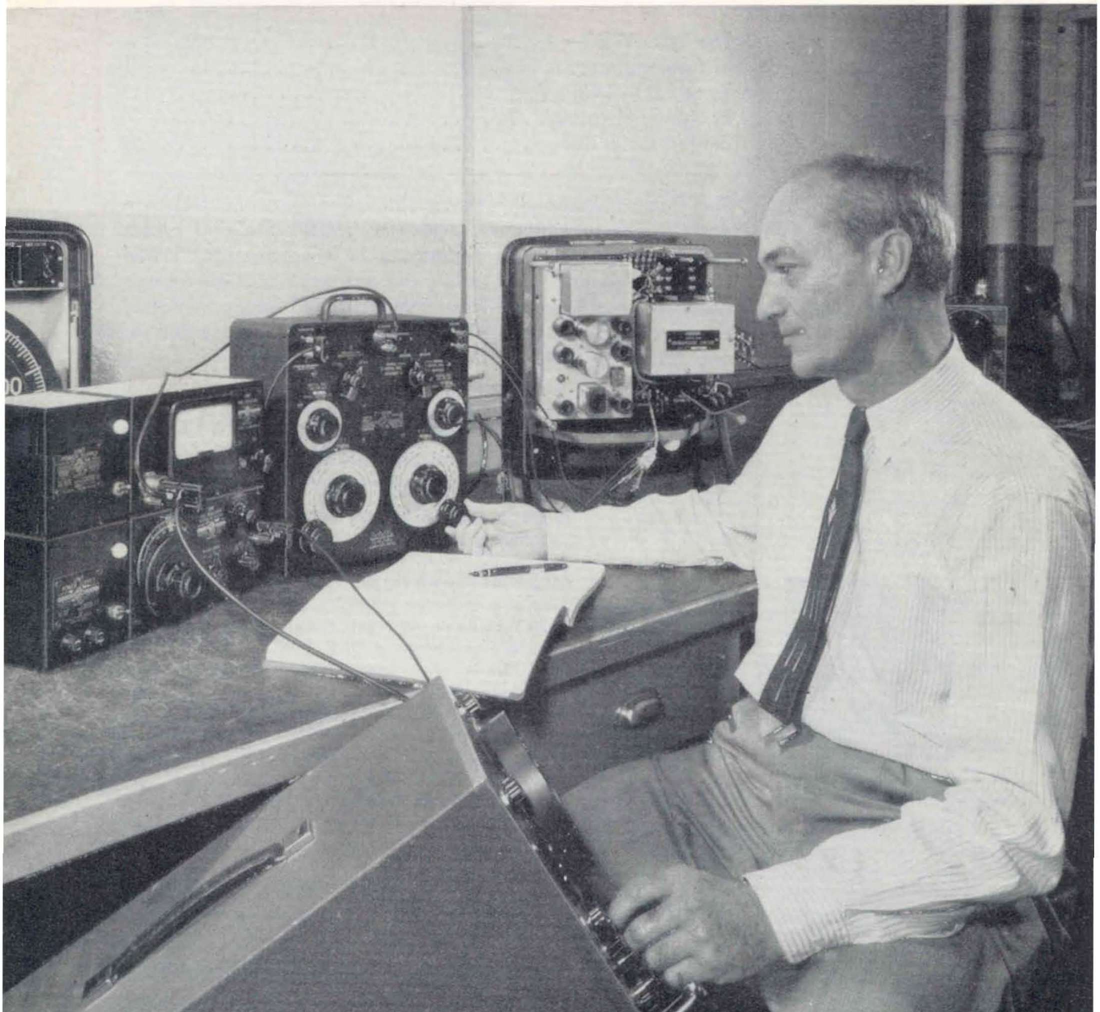

Photo courtesy The Foxboro Company

In This Issue

New 20-Ampere Variac®
Equipment Leasing
Regulated Power Supply

the GENERAL RADIO Experimenter

Published Monthly by the General Radio Company

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COVER

The Type 1603-A Z-Y Bridge will balance for any impedance, real or imaginary, positive or negative, over the entire audio-frequency range. This photograph shows the bridge in use in the Electronic Design Department of The Foxboro Company, makers of *Dynalog* self-balancing recorders, to measure impedance characteristics of a "Dynapoise Drive," self-balancing recorder element.

THE TYPE W20—A NEW 20-AMPERE VARIAC® AUTOTRANSFORMER

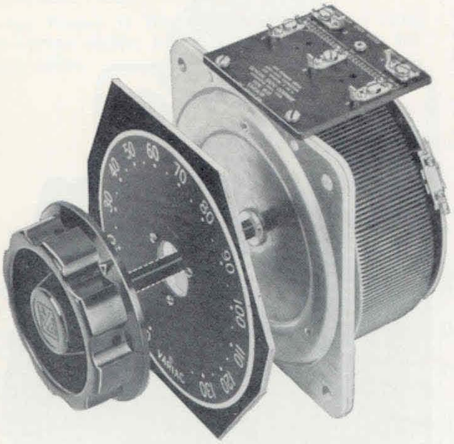


Figure 1. Uncased, 115-volt model, Type W20.

Variacs, unlike automobiles, are *not* restyled each year to meet the whims, real or fancied, of hypothetical purchasers. Their design is the result of long-term, frequently reviewed planning, the purpose of which is to supply the best possible continuously adjustable autotransformers for today's dollar. Barring radical improvements in economically justifiable core material or a competitively priced superconductor, Variacs closely approach the optimum design criteria.

The design features of W-model Variacs, already proved in the W2, W5, and W50 sizes, have now been incorporated in the new Type W20 series. These features, which were discussed in a previous article,[1] make the W-model Variac® a more rugged, more adaptable, and more durable device than its predecessors.

You will find wrought metallic parts substituted for castings in the interest of improved mechanical properties. Heat transfer between coil and base, brush and radiator has been improved. Ball-bearing models for motor-drive and other demanding applications are stock items. Totally enclosed and portable models are included in the new line.

All W20 models have *Duratrak* contact surface, developed by General Radio, which minimizes brush-track deterioration under adverse environmental or load conditions and assures long and trouble-free life.

Basic Uncased Models — Types W20, W20H

The two uncased models, Type W20 for 115-volt service and Type W20H for 230-volt service, are normally used for back-of-panel mounting on switchboards or built into electrical equipment, as shown in Figure 1, but, when the shaft

[1] "The Type W5 Variac® — A New and Better Variable Autotransformer," *General Radio Experimenter*, 30, 7, December, 1955.

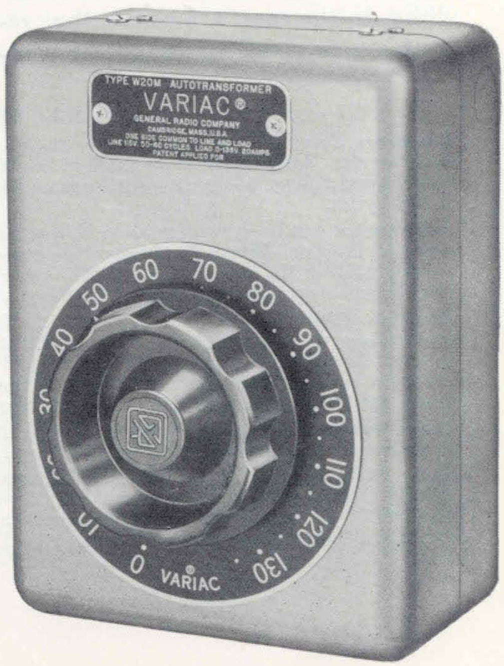


Figure 2. Type W20M, with case.

is extended on the winding side rather than the base plate side and the dial attached to the knob, they can also be used on a bench or table. This arrangement, however, affords no protection against electric shock or against accidental damage to the Variac from abrasion.

Cased Models — Types W20M, W20MH

These models, each consisting of a basic Type W20 or W20H Variac® mounted in an aluminum case, are finished in attractive gray enamel. For conduit or armored cable, cases are provided with four knockouts, two on the end and one on each side. There is ample space inside the case for wiring. The front half of the case is easily removable for access to terminals, mounting holes, and brush. These models are easy to install — on wall, on bench, or behind panel.

Figure 3. Portable model, Type W20MT3, with case, overload breaker, 3-wire output receptacle, and heavy-duty cord and plug.

Portable Models — Types W20MT3, W20HMT3

The portable models include, in addition to the case, a 3-wire output receptacle, an on-off switch and an overload circuit breaker, which trips on either excessive current or excessively-high prolonged operating temperature. Heavy-duty, 3-wire cord and plug are permanently attached.

GENERAL SPECIFICATIONS

Frequency: Specifications are for 50- to 60-cycle service. Variacs can be operated at rated current and voltage at line frequencies from 50 to 400 cycles.

Rated Current can be drawn from the Variac at any dial position. When the overvoltage connection is used, the load should not take more than rated current at maximum output voltage.

Maximum Current can be drawn at the input line voltage when the line-voltage connection is used. At any lower setting the Variac will control a constant-impedance load drawing no more than the maximum current at line voltage.

Output Voltage is the range of voltage available at the output terminals with rated input voltage.

Dials: Portable models (MT3, HMT3) are wired for overvoltage connections and have corresponding dial scales but can be supplied on special order with line-voltage connections and dials.

Dial plates for all other models are reversible, with line-voltage scale on one face and overvoltage on the other. Angle of rotation is approximately 320 degrees.

Line-Voltage or Overvoltage Output Connections: "Line-Voltage Connection" refers to the connection for output voltage range of zero to line voltage. "Overvoltage Connection" refers to the connection for output voltage range of zero to 17% above line voltage.

KVA Rating is the maximum current multiplied by normal input line voltage. A Variac can handle, at any lower setting, a constant-impedance load that draws at rated input voltage a current no greater than the maximum current.

Temperature Rise: Variac ratings are based on operation at ordinary room temperatures, with an average temperature rise of not more than 50°C. When ambient temperature exceeds

50°C., kva ratings should be decreased as shown in the chart, Figure 6.

No-Load Loss: 27 watts at 60 cycles with rated input voltage. Losses are guaranteed not to exceed this value.

Driving Torque is the torque required to turn the Variac shaft: 55 to 110 ounce inches for single units.

Panel Thickness is the maximum thickness of the panel on which the Variac can be mounted, with the shaft normally supplied: 15/32 inch.

Dimensions: Uncased model, base $7\frac{1}{2}$ x $7\frac{1}{2}$ inches, depth behind panel $4\frac{5}{8}$ inches; case dimensions, (width) $8\frac{5}{8}$ x (height) $11\frac{5}{16}$ x (depth) $5\frac{3}{8}$ inches.

Weight: See individual specifications.

		Line-Voltage Connection				*Overvoltage Connection*				
Type and Mounting	*Input Volts*	*Rated Output Amps.*	*Output Voltage*	*Max. Output Amps.*	*Output KVA*	*Output Volts*	*Rated Output Amps.*	*Net Weight Pounds*	*Code Word*	*Price*
W20 Uncased	115	20	0–115	26	3.0	0–135	20	$21\frac{3}{8}$	FEDAL	**$45.00**
W20M With case	115	20	0–115	26	3.0	0–135	20	$24\frac{1}{2}$	FEDER	**58.00**
W20MT3 Portable	115					0–135	20	$28\frac{1}{8}$	FEDOM	**87.00**
W20H	230	8	0–230	10.4	2.4	0–270	8	$20\frac{1}{4}$	MEPAL	**47.00**
Uncased	115	4				0–270	4			
W20HM	230	8	0–230	10.4	2.4	0–270	8	$23\frac{3}{8}$	MEPER	**60.00**
With case	115	4				0–270	4			
W20HMT3 Portable	230					0–270	8	27	MEPOM	**85.00**

BALL BEARINGS

All W20-model Variacs can be supplied with ball bearings. When ordering, add the suffix BB to the type number and add to the price the amount shown in the following table:

Model	*Surcharge for Ball Bearings*
Single	**$8.00**
2-gang	**10.00**
3-gang	**12.00**

Type W20 and Type W20H Variacs are approved by the Underwriters' Laboratories

GANGED MODELS

2- and 3-gang assemblies of W20-model Variacs are available either uncased or with cases.

Type		*Load Rating KVA* Parallel	Series	Delta	Y	*Input Line Volts*	*Code Word*	*Price*
W20G2	**2-gang Type W20......**	6				115	FEDALGANDU	**$100.00**
			6			230		
				5.2		115		
W20G2M	**2-gang W20 with case.**	Same as TYPE W20G2					FEDALBONDU	**125.00**
W20G3	**3-gang Type W20......**	9				115	FEDALGANTY	**147.00**
					10.4	230		
W20G3M	**3-gang W20 with case.**	Same as TYPE W20G3					FEDALBONTY	**175.00**
W20HG2	**2-gang Type W20H....**	4.8				230	MEPALGANDU	**104.00**
			4.8			460		
				4.2		230		
W20HG2M	**2-gang W20HM.......**	Same as TYPE W20HG2					MEPALBONDU	**129.00**
W20HG3	**3-gang Type W20H....**	7.2				230	MEPALGANTY	**153.00**
					8.3	460		
W20HG3M	**3-gang W20H with case**	Same as TYPE W20HG3					MEPALBONTY	**181.00**

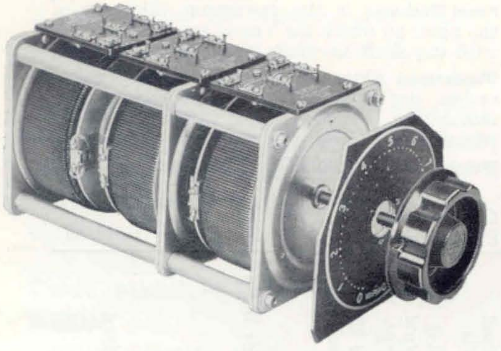


Figure 4. Uncased three-gang assembly, Type W20G3.

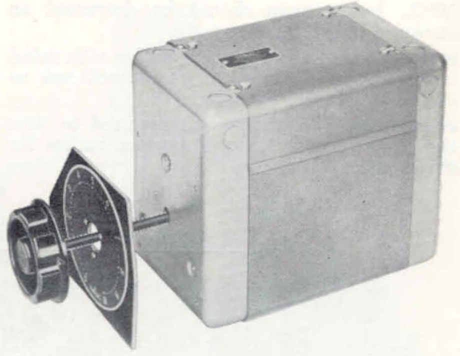


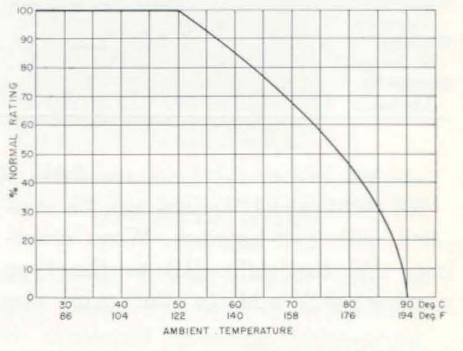
Figure 5. Three-gang assembly with case, Type W20G3M.

MOTOR DRIVE

As with other W-model Variacs, motor-driven units and assemblies can be supplied. These simple, relatively inexpensive drives are available for both servo and remote positioning applications. Cases similar to those used on ganged assemblies (Figure 2) can be supplied for applications where complete enclosure is desired.

The price for motor drive varies with the quantity ordered. We shall be glad to quote prices and to recommend the model best suited to your requirements.

Figure 6. For ambient temperatures above 50°C., Variacs should be derated according to this curve.



TYPE 1201-B UNIT REGULATED POWER SUPPLY

The Unit Regulated Power Supply has been redesigned for reduced ripple voltage and more nearly constant output voltage. The new model number is Type 1201-B. The d-c output voltage of the new instrument is constant within $\pm\frac{1}{4}\%$ for all values of load current and line voltage, and the ripple voltage is less than 1 millivolt (120 cps) at full load. The Type 1201-B Unit Regulated Power Supply is recommended for use with General Radio Unit Instruments as well as with other equipment in applications where line-voltage fluctuations are serious.

SPECIFICATIONS

Output: 300 v dc ($\pm1\%$) at 70 ma.
6.3 v ac at 4 amp. (unregulated).

Regulation: D-c output voltage is constant within ±¼% for all values of load current and line voltage.

Ripple: Less than 1 mv (120 cps) at full load.

Internal Impedance: 0.4Ω + 10 μh (max).

Input: 105–125 volts, 50-60 cps, 87 w, full load at 115 v.

Connectors: Line cord permanently attached to instrument. Standard 4-point connector mounted on cabinet side for other Unit Instruments.

Accessories Supplied: Line cord; mating plug for equipment other than Unit Instruments.

Mounting: Black-crackle-finish aluminum panel and sides. Aluminum cover finished in clear lacquer.

Dimensions: Width 5 in., height 5¾ in., depth 6¼ in. over-all, not including power cord.

Weight: 6 pounds.

Type		*Code Word*	*Price*
1201-B	Unit Regulated Power Supply	ASSET	$85.00

EQUIPMENT LEASING

In our dynamic economy, capital equipment users often face the problem of acquiring large quantities of equipment for modernization, expansion, replacement, and for Government contracts. This demand for capital equipment can tie up substantial amounts of working capital that might be more profitably used otherwise. Because of this, there is a strong trend towards leasing, which has been sparked by the growing realization that it may be better to pay for the use of equipment out of current income than to pay for ownership out of past profits, debt, or equity financing. The following benefits may be realized by leasing test equipment:

(1) Liquid capital can be most profitably employed.

(2) Equipment costs can be pinpointed to specific projects or contracts.

(3) Government contracts can be partially financed on a lease basis.

(4) Monthly payments can be handled as an operating expense.

(5) The normal credit pool is not dried up, leaving cash and borrowing facilities intact for other purposes.

(6) New instruments are acquired on a pay-its-way basis.

(7) Because equipment can be leased long before funds can be made available through inflexible capital budgets, expensive delays can be avoided.

(8) By not making heavy demands on liquid cash, leasing encourages management to provide their valuable engineering talent with the most efficient up-to-date instrumentation. The result is an increase in output per man-hour.

To meet this need for leased instruments, a number of reputable concerns provide leasing services for credit-worthy companies in the electronic industry. It is important to note that these companies *lease* equipment and *do not rent* equipment. The difference is that leased equipment is purchased by a leasing company on specific orders from the user and placed in the user's plant under leasing arrangements, whereas rented equipment implies that an inventory of instruments is available for short-term sporadic demands. Because of the extreme diversity of modern electronic instrumentation, the capital demands for maintaining an instrument inventory are so great that rental-service companies have not evolved. As far as we know, no leasing company stocks equipment for rent.

How does one go about leasing an instrument? The first step is to determine what instrument you need and who manufactures it. This step is no different than the making of a regular capital purchase. You look over the field and select the instrument and manufacturer who has the desired instrument. You then get in touch with his Sales Department to discuss the technical details and inquire about delivery and cost. Do *not* go to a leasing company for technical details or specifications of the instrument. Once you have established, in cooperation with the equipment manufacturer, the particular instrument desired, *then* go to the leasing company. The leasing company will furnish you with a quotation for the leasing of the specific instrument that you have selected. If the leasing terms are acceptable to you, a lease will be drawn, and upon execution the leasing company will place a purchase order with the equipment manufacturer directing shipment to the user. All equipment warranties offered by the manufacturer are extended to the user. With General Radio Company, all instruments carry a two-year warranty. Should you have a service problem after purchasing the instrument, all negotiations would be directly between you and the manufacturer. Most electronic equipment is leased on a 36-month basis. Thereafter, annual renewals are made available at nominal rates.

A typical example is the purchase of a General Radio Type 1100-AP Primary Frequency Standard. The catalog price for this instrument is $2440. The cost per month for the Frequency Standard will be approximately $80.50 for 36 months. If at the end of the 36-month period you wish to renew the lease, it can be done for $146.50 per *annum*. When a larger-dollar-volume is involved, the rate may be reduced somewhat, depending upon the complexity of the order.

It may be seen from these rates that it is possible through leasing to keep your laboratory and production test equipment completely modernized at a nominal cost and still have capital available for other investments.

— Robert B. Richmond

EXHIBIT CALENDAR

During the months of September and October, many of the newest General Radio instruments will be on display at the technical meetings and conventions listed below. A cordial invitation is extended to all *Experimenter* readers who attend these conventions to visit the General Radio booths and to talk over your measurement problems with our engineers.

IRE CONFERENCE ON COMMUNICATIONS — **CEDAR RAPIDS, IOWA**
September 12 and 13, 1958 — Booths 111 and 112 — Sheraton Montrose Hotel

IRE CANADIAN CONVENTION — **TORONTO, CANADA**
October 8-9-10, 1958 — Booth 450 — Exhibition Park

NATIONAL ELECTRONICS CONFERENCE — **CHICAGO, ILLINOIS**
October 13-14-15, 1958 — Booths 172, 173, 174 — Hotel Sherman

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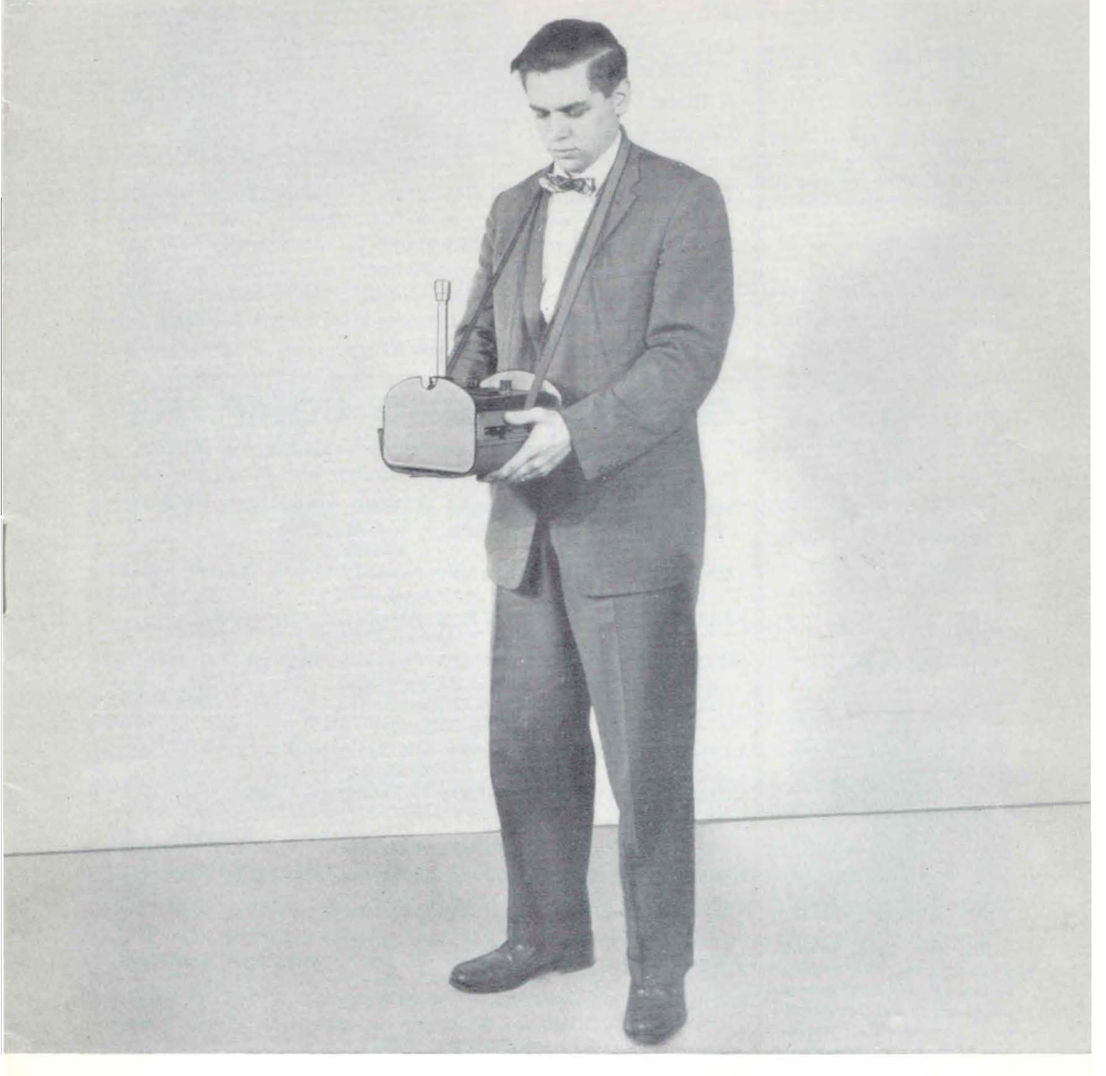

In This Issue

New Sound-Level Meter

Three-Terminal Precision Capacitor

the GENERAL RADIO

Experimenter

Published Monthly by the General Radio Company

VOLUME 32 • NUMBER 17 **OCTOBER, 1958**

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COVER

The new Type 1551-B Sound-Level Meter described in this issue sets a new high in both accuracy and convenience. Photograph shows instrument in carrying case, supported by shoulder strap, and ready for use.

IMPROVED PERFORMANCE PLUS A NEW LOOK FOR THE SOUND-LEVEL METER

Figure 1. Panel view of the Sound-Level Meter, with microphone folded down in OFF position. In operating position, microphone can be either erect, as shown on front cover, or extending outward from case.

The primary function of the General Radio Sound-Level Meter is noise measurement — noise generated by machines and appliances, environmental noise, and the propagation of noise by materials and structures. With its many accessories, however, it also comprises a complete sound-measuring system, capable of accurate sound-pressure measurements, sound-spectrum analysis, and other acoustic measurements.

The latest model of this versatile instrument, TYPE 1551-B, is smaller, lighter, and easier to hold than its predecessor[1,2] and, in addition, has many worthwhile new technical features and improvements. Among these are:

1. A new microphone, for better all-round performance.

2. A new meter circuit, which more closely approximates rms response.

3. A new calibration circuit for amplifier gain standardization, which does not require a power line connection.

4. A new adjustment for microphone sensitivity, which facilitates the use of special-purpose microphones.

5. Improved signal-to-noise ratio and dynamic range. These make possible the use of spectrum analyzers over a greater amplitude range so that noises with a steeper spectrum slope can be analyzed.

6. Improved frequency response, which permits increased accuracy of measurement.

7. Improved performance at high sound levels; the upper limit of measurement has been raised and microphonics have been reduced.

8. Improved stability, requiring less frequent calibration adjustments.

These improvements, which are discussed in detail below, result from General Radio's continuous program of development in the field of acoustic measurements and also from the many constructive suggestions received from users of the TYPE 1551-A Sound-Level Meter.

MICROPHONE

Choice of the proper microphone to serve as the acoustic pickup for a high-

[1] E. E. Gross, "TYPE 1551-A Sound-Level Meter," *General Radio Experimenter*, Vol. 26, No. 10, March, 1952.

[2] E. E. Gross, "The Sound-Level Meter as an Audio Frequency Voltmeter and Amplifier," *General Radio Experimenter*, Vol. 31, No. 8, January, 1957.

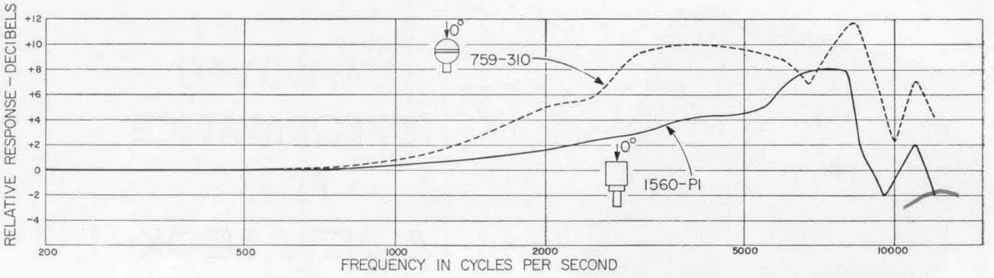


Figure 2. Free-field, perpendicular incidence (0 degrees) frequency response of Type 1560-P1 (98B99) and Type 759-310 (9898) Microphone.

quality but moderately priced sound-level meter is no easy task. Many months of testing and evaluating the many types of "hi-fidelity" microphones available must go into the process. After taking all of the factors into account, we feel that the Shure Brothers TYPE 98B99 microphone[3] is a "Best Buy" for this use. This microphone, which we list as the TYPE 1560-P1, is a small Rochelle-salt-crystal type of modern design. It has the same sensitivity as the older TYPE 9898, which was used on the TYPE 1551-A Sound-Level Meter. Its new design and small size contribute to a response that is smoother and more nearly nondirectional at high frequencies than that of the older microphone and retains the same flat response to low frequencies. Figure 2 illustrates the improvement in high-frequency response that has been obtained. The solid curve is the free-field frequency response for one of the new microphones to sounds arriving along its axis or perpendicular to the plane of its diaphraghm. The dotted curve is a similar frequency response for one of the older microphones. The new microphone response extends fairly smoothly out to 8 kc, with no major peaks or valleys. Measurements on production quantities of these microphones confirm the uniformity exhibited by the eighteen microphones that gave the results pictured in Figure 3. The solid curve starting at 0 decibel shows the C weighting, or flat, response of the TYPE 1551-B with the average microphone. The shaded area within the dotted lines shows the measured variation in response. The upper and lower solid curves are the current ASA[4] limits for the C weighting response of a sound-level meter.

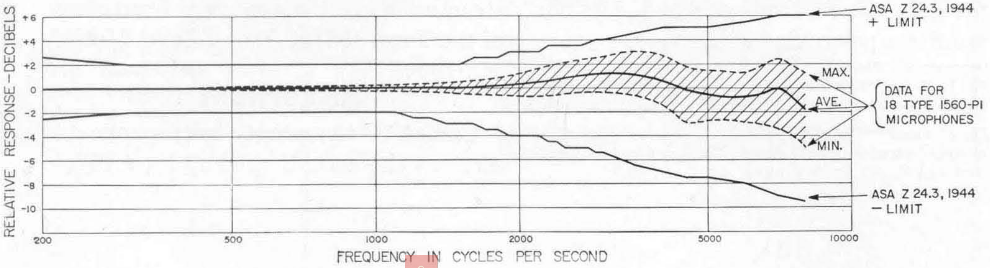
METER CIRCUIT

For many years American sound-level meters built to conform to the standard

[3]John Meddill, "A Miniature Piezo Electric Microphone," *Transactions of the IRE Professional Group on Audio*, Vol. AU 1, No. 6, November-December, 1953.

[4]American Standards Association, American Standard for Sound-Level Meters for Measurement of Noise and Other Sounds, Z24.3-1944.

Figure 3. Free-field frequency response of Type 1551-B Sound-Level Meter to sounds of random incidence (C weighting). Average and extreme values for 18 microphones are shown, together with ASA tolerances.



referred to before have used, as indicators, full-wave rectifier meters, with rectifiers operating in the low-density region to approximate square-law operation. These meters were basically average-reading indicators and, although they met the two-tone test for rms response, their indications were much closer to average than rms on most other tests.

The new meter circuit developed for use in this sound-level meter, while it is not a true rms indicator, indicates much more closely to the rms value than did the previous meter. Simply stated, the new meter circuit shown in Figure 4 combines an average-reading circuit with a peak-reading circuit to create what we have called the "quasi-rms" indicating circuit. Since the rms value of most waveforms falls above the average value but below the peak value, it was reasoned that a portion of the peak indication added to the average value of a given waveform should result in an rms indication.[5, 6] This is true, but different waveforms require different amounts of peak indication to give the rms value. In the circuit shown in Figure 4, the ratio of R-2 to R-1 determines the amount of the peak contribution, and one set of values will give an indication on many signals that is very close to rms. Table I below shows the decibel difference between the indication of a true rms meter and the new meter circuit, an average meter, and the meter in the older Type 1551-A for several test signals.

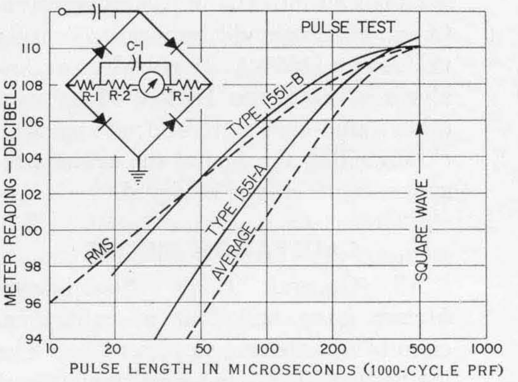


Figure 4. Quasi-rms meter circuit used in Type 1551-B, and response of meter circuits in both -B and -A models to pulses of constant amplitude and varying duration.

In Column 2, the signal consisted of two tones, one at 1000 cps set to give a convenient indication on the meter. The other tone differs from 3000 cps by a few cycles and has 30% the amplitude of the first. A true rms meter will show no fluctuation with this type signal. Column 3 refers to the two-signal test outlined in the Sound-Level Meter Standard.[4]

The solid curves shown in Figure 4 show the response of the new meter and its predecessor to pulses of constant height but varying length. It can be seen that the new meter indicates the rms value (upper dotted curve) within ± 1 db until the pulse duration becomes as short as 1/25th that of a square wave.

[5] R. L. Frank, "Harmonic Insensitive Rectifiers for AC Measurements," *Proceedings National Electronics Conference*, Vol. 8, 1952, pp. 495-505.
[6] Arnold P. G. Peterson, "Response of Peak Voltmeters to Random Noise," *General Radio Experimenter*, Vol. 31, No. 7, December, 1956, pp. 3-8.

TABLE I

Type Meter	*db Fluctuation with Phase Changes at 30% 3rd Harmonic*	*Difference in Meter Indication and RMS Decibels*		
		For Two-Signal Addition	*For Square Waves*	*For Noise*
1551-B	0.45	+.05	+0.1	−0.25
AVE.	1.8	−1.0	+1.0	−1.0
1551-A	1.7	−0.4	+0.6	−1.0

This is well into the impact-type waveforms, which should be measured using the TYPE 1556-A Impact-Noise Analyzer.[7] The TYPE 1551-A Meter and other sound-level meters do not approach closely either the rms or the average response for this type of signal.

CALIBRATION CIRCUIT

All General Radio Sound-Level Meters have had built-in calibration circuits for checking amplifier gain.[8] The method used in previous models required a 115-volt, 60-cycle line, which often was not convenient and sometimes not possible. The new method, like the old, equates the amplifier gain to a stable, resistive attenuator, but the method of comparison is simplified. The output of the amplifier with its internal attenuators is connected back to the input through a frequency-determining network and a preset calibrated attenuator. When the net loop gain is unity (i.e., when gain equals attenuation), the system starts to oscillate.

One might expect that the sudden start of oscillation when the critical, unity-gain condition was reached would cause the meter to slam from off down-scale to off up-scale. This is prevented by an amplitude limiter in the feedback loop, which controls the build-up of oscillation level as gain is increased. The slope of the build-up is purposely left fairly steep, however, to magnify the

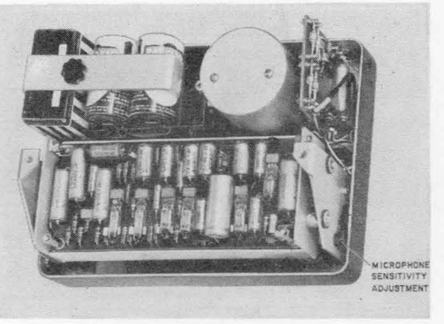


Figure 6. Location of microphone sensitivity adjustment.

precision of setting. As illustrated in Figure 5, adjustment of oscillation amplitude within the 4-db span AOA′ on the meter sets the amplifier gain to its proper value within ± 0.2 db as indicated by the small triangle BOB′.

MICROPHONE SENSITIVITY ADJUSTMENT

The versatility of the sound-level meter is increased greatly by the use of different types of microphones for different applications. The Rochelle-salt type supplied with the instrument is the best all-round microphone for general use; but dynamic types are useful when long cables are required, and condenser types are better when extended frequency response is required or very high sound levels are encountered.

The attenuator in the new calibration circuit is calibrated directly in terms of microphone sensitivity from —49 to —61 db re 1 volt per μbar. This attenuator is normally preset in our laboratory to match the microphone supplied. The usual setting is in the region of —58 db.

[7] Arnold P. G. Peterson, "The Measurement of Impact Noise," *General Radio Experimenter*, Vol. 30, No. 9, February, 1956.

[8] W. R. Thurston, "The Basis for Field Checking Sound Meter Calibration," *General Radio Experimenter*, Vol. 27, No. 6, November, 1952.

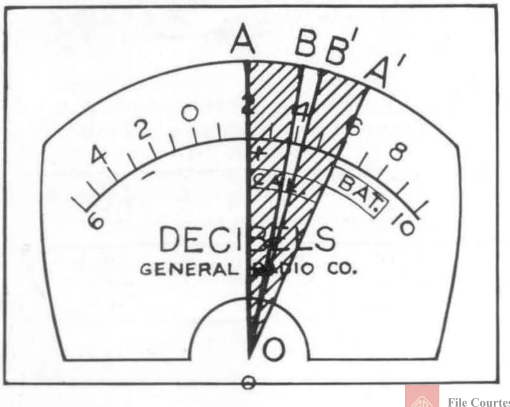


Figure 5. Showing the magnifying action of the calibration check circuit.

To use a microphone with a —55 db sensitivity, one merely sets the microphone sensitivity adjustment (see Figure 6) to —55 and readjusts the amplifier gain until the meter needle falls in the area AOA′ of Figure 5; the sound-level meter is then direct reading with that microphone.

SIGNAL-TO-NOISE RATIO AND DYNAMIC RANGE

One of the operating characteristics of a sound-level meter which becomes important when an analyzer[9] is used is the internal noise level with respect to the desired signal. This is particularly true when the noise being analyzed has a sloping spectrum. This signal-to-noise ratio has been substantially improved in the new instrument. The attenuator controlled by the panel DECIBELS switch has been distributed throughout the amplifier to keep all tubes operating at as high a signal level as possible and hence to increase the signal-to-noise ratio. The improvement in signal-to-noise ratio and, hence, in dynamic range is apparent from the figures in Table II below.

The signal-to-noise ratio is slightly better at all higher attenuator settings than those shown below. The amplifiers will handle signal levels 10 db above full scale on the meter, so that one can say that the total dynamic range of the instrument for all attenuator settings above 60 db is greater than 70 db in the octave bands and of the order of 68 db over the frequency range of 20 to 10,000 cps.

FREQUENCY RESPONSE

Five design improvements contribute to a better over-all frequency-response characteristic:

A. The new microphone has a smoother high-frequency response. Manufacturer's tolerances on the allowable variation in the response of the microphone have been appreciably reduced.

B. The frequency responses of the range attenuators have been improved and these characteristics are more closely controlled.

C. Modifications in the amplifier design have resulted in a greater bandwidth. Figure 7 shows the response of the new amplifier over the frequency range of 10 cps to 50 kc.

D. New weighting networks have been designed to give *A*, *B*, and *C* responses that conform more closely to the ASA design centers. The improved frequency characteristics of the new microphone justify this effort.

E. The new indicating meter in the TYPE 1551-B has no observable frequency error in the usable range.

PERFORMANCE AT HIGH SOUND LEVELS

The maximum attenuator setting has

[9]As, for instance, the General Radio TYPE 1550-A Octave-Band Analyzer and the TYPE 760-B Sound Analyzer.

TABLE II

OCTAVE-BAND NOISE LEVELS DB BELOW FULL SCALE ATTENUATOR SETTING = 70 DB (C Weighting)

BAND – CPS	*20–10,000*	*20–75*	*75–150*	*150–300*	*300–600*	*600–1200*	*1200–2400*	*2400–4800*	*4800–10,000*
Type 1551-B	58	66	71	72	71	69.5	67	65	62
Type 1551-A	37	56	56	55	52	49.5	46.5	44	40
Improvement	21	10	15	17	19	20	20.5	21	22

been increased from 130 to 140 db, making the direct-reading range of the instrument 24 to 150 db above the reference pressure level of .0002 μbar. The numerous factors that contribute to the improved signal-to-noise ratio also improve the operation of the instrument in high sound fields. Much attention has been given to the problem of keeping the microphonic signals at a minimum in this new instrument. Microphonics due to vibrations and airborne sound are reduced by the use of low-noise cable throughout the instrument, by improvement in the shock and vibration mounting of the amplifier compartment, and by the mounting of all tubes between soft rubber pads. Measurements made in high sound fields, however, should be carried out with great care, and one should exercise the usual precautionary measures to determine that microphonics are at a safe level.

STABILITY

Increased stability of the amplifier was an important objective in the redesign of the sound-level meter. The change effecting the greatest part of the improved stability is the addition of a voltage-reference diode, which maintains constant plate voltage on the first two amplifier stages over the full operating range of the B+plate battery. Other contributing factors are increased negative feedback around the main amplifier, made possible by an increase in the gain of the second amplifier stage, and conversion of the output amplifier from a vacuum tube to a transistor connected as an emitter follower.

Figure 7. Frequency response of amplifier in Type 1551-B Sound-Level Meter (weighting switch set at 20 kc).

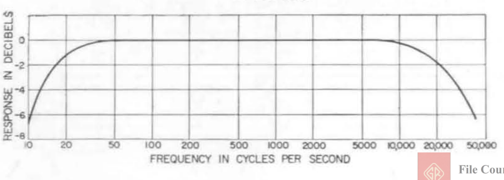


Figure 8. View of Sound-Level Meter with A-C Power Supply attached.

A-C POWER SUPPLY

Batteries are the normal power supply of the sound-level meter, as listed in the specifications. A-C power is more convenient for laboratory use, however, and the Type 1262-B A-C Power Supply is available for such applications. This small power unit mounts directly on the end of the Type 1551-B and connections are made between the two by means of a 6-terminal connector when the supply is so mounted. The supply can be operated from either a 105-125-volt, 50-60 cycle line or a 210-250-volt, 50-60-cycle line. Two well-filtered filament supplies and a plate supply are provided. Stability of the amplifiers in the Type 1551-B is such that normal line voltage variation causes negligible change in meter reading.

CARRYING CASE

The shielded aluminum cabinet, finished in dark-gray crackle, is a rugged and durable carrying case which affords adequate protection to the interior elements. To make field use of the new meter as convenient as possible, however, an ever-ready type leather carrying case is now available. This case, the TYPE 1551-P2, is illustrated on the cover, protects the sound-level meter in transit, and provides a handy carrying handle. With the long strap placed over the shoulder, the instrument can easily be carried and can be operated without removal from the case.

ACCESSORIES

The full line of accessories for the A-model sound-level meter can be used equally well with the new model. These include acoustic calibrator, dynamic and condenser microphones, vibration pickup, and spectrum analyzers. The new sound-level meter uses a different type of plug-in connector for the microphone, and an adaptor is furnished with each accessory instrument that requires it.

Much thought and effort have gone into the design of the TYPE 1551-B Sound-Level Meter. A new look and improved performance have resulted. Mechanical design and handling ease have been carefully considered as well as electrical and acoustical performance. The TYPE 1551-B is a big step forward in sound-level-meter design.

—E. E. GROSS

SPECIFICATIONS

Sound-Level Range: From 24 db to 150 db (re 0.0002 microbar).

Frequency Characteristics: Any one of four response characteristics, A, B, C, and 20 kc, can be selected by a panel switch.

The A and B weighting positions approximate the response of the ear to pure tones referred to a 40-db level and a 70-db level, respectively, at 1000 cps.

The C weighting provides uniform response to all frequencies within the range of the microphone. This characteristic is used for measuring high sound levels, for measuring sound-pressure levels, or when the instrument is used with an analyzer.

The 20 kc position allows the use of the complete frequency response of the sound-level meter's amplifier, which is flat from 20 c to 20 kc, so that complete use can be made of wide-range microphones such as the General Radio TYPE 1551-P1 Condenser Microphone Systems.

Microphone: The microphone is a high-quality Rochelle-salt-crystal diaphragm type. Condenser and dynamic microphones are available as accessories.

Sound-Level Indication: The sound level is indicated by the sum of the readings of the meter and attenuator switch. The clearly marked, open-scale meter covers a span of 16 db with calibration from −6 to 10 db. The attenuator is calibrated in 10-db steps from 30 to 140 db above the standard reference level.

Output: An output of 1 volt across 20,000 ohms (when the panel meter is at full scale) is available at an output jack. The output can be used to drive frequency analyzers, recorders, and oscillographs.

Meter Damping: The panel meter is provided with two different damping characteristics, selected by a switch. In the FAST position, the meter ballistics agree with the current ASA standards. In the SLOW position, the meter is heavily damped and indicates, for easy reading, the average level of rapidly fluctuating sounds.

Calibration: Internal means are provided for standardizing the sensitivity of the electrical circuits in the sound-level meter. After standardization, the accuracy of sound-level measurements, as specified in ASA standards, is within ± 1 db for average machinery noise. The TYPE 1552-B Sound-Level Calibrator is available for making periodic acoustic checks on the over-all calibration, including microphone.

Temperature and Humidity Effects: Readings are independent (within 1 db) of temperature and humidity over normal ranges of room conditions.

Power Supply: Two 1½-volt size D flashlight cells (Rayovac 2LP or equivalent) and one 67½-volt Burgess XX45 battery or equivalent are supplied. The TYPE 1262-B Power Supply for a-c operation is available.

Tube Complement: Four Raytheon CK-512-AX and two Raytheon CK 6418 tubes; one RCA 2N105 transistor.

Accessories Supplied: Telephone plug.

Cabinet: Shielded aluminum cabinet, finished in gray crackle, which serves as a convenient and rugged carrying case. A leather carrying case is available, which permits operation of the instrument without removal from the case.

Dimensions: Sound-level meter only, (depth) 6⅛ x (length) 9¼ x (width) 7¼ inches, overall; with a-c power supply attached, 6⅛ x 12⅜ x 7¼ inches.

Net Weight: 7 pounds, 10 ounces with batteries; 9 pounds, with a-c power supply; a-c power supply only, 2 pounds, 8 ounces; leather case, 2 pounds.

Type		*Code Word*	*Price*
1551-B	Sound-Level Meter*	MIMIC	$395.00
	Set of Replacement Batteries	MIMICADBAT	3.90
1262-B	Power Supply	MANLY	70.00
1551-P2	Leather Carrying Case	CALYX	20.00

*Licensed under patents of the American Telephone and Telegraph Company.

A THREE-TERMINAL PRECISION CAPACITOR

The normal construction of the TYPE 722 Precision Capacitor is with the rotor grounded to the frame and panel. Such construction gives what is normally referred to as a two-terminal capacitor, with one terminal connected to the ground of the unit and, in turn, to the ground of the measuring circuit.

In some measuring circuits it is necessary or desirable to have both terminals of the variable capacitor at other than ground potential. This is the so-called three-terminal construction, ground being the third terminal. We have built a great many capacitors of this type on special order in the past, and in recent years the interest in this type of construction has been increasing. The TYPE 722-CB, a three-terminal capacitor having a 1050-$\mu\mu$f range, is now offered as a stock item.

For capacitance measurement, three-terminal standards have certain definite advantages, the most obvious of which is the relative freedom from the problem of lead capacitance. In the three-terminal construction, the capacitance of leads *to ground* does not affect the direct capacitance between rotor and stator. The capacitance *between leads* can be eliminated by shielding of at least one terminal and lead. When this is done, complete freedom from lead capacitance problems is insured. Theoretically only one terminal and lead need be shielded, but in certain instances it may be advantageous to shield the low terminal as well as the high terminal. For example, in a circuit where the impedance of both terminals to ground is relatively high, electrostatic shielding to guard against pickup will be necessary on both leads. For this reason coaxial terminals (TYPE 874) are provided for both rotor and stator connections in the TYPE 722-CB.

A second equally important advantage of a three-terminal capacitor is that losses are definitely lower than in a two-terminal capacitor of similar construction. The losses in the dielectric supporting structure, which in the two-terminal construction are in parallel with the desired capacitance, appear in the terminal-to-ground capacitances in the three-terminal construction. In theory all the dielectric losses can be placed in terminal capacitance, but actually in typical construction there may be some residual fringing field between rotor and stator which passes through solid dielectric material. In the TYPE 722-CB the rotor insulation is completely shielded and the stator insulation is sufficiently removed from the main field so that residual losses from this cause are ex-

Panel view of the Type 722-CB Precision Capacitor.

tremely small. In addition to these extremely low losses from dielectric supports, there may be some losses at the surface of the capacitor plates, and there are even detectable losses in air, both of which depend greatly upon relative humidity.

In the two-terminal construction, the Type 722 Precision Capacitor has a D_0C_0 product of 0.03 $\mu\mu$f, corresponding to a dissipation factor of 30×10^{-6} at 1000 $\mu\mu$f and 300×10^{-6} at 100 $\mu\mu$f. Ideally, such a capacitor in the three-terminal construction should have losses at all settings smaller than the loss at the 1000-$\mu\mu$f setting in the two-terminal construction. Measurements show that there is actually a slight tendency for the dissipation factor to rise at low capacitance setting. The increased losses, which result when the edges of the rotor and stator plates are in close proximity, are presumably caused by the localized high voltage gradients at the relatively sharp edges of the plates.

The measurement of dissipation factors in the order of a few tens of microradians is fairly difficult, and most conventional measuring techniques have uncertainties of the same general order of magnitude. We believe, however, that under favorable conditions our measurement of dissipation factor, using a substitution technique with the Type 716 Bridge, has an absolute accuracy of the order of 10 microradians. Allowing for this uncertainty of measurement, the Type 722-CB is rated as having a dissipation factor less than 50 microradians at all settings in the linear capacitance range.

Since, in the three-terminal construction, the stator-to-ground capacitance is eliminated from the calibrated value, a lower minimum value of capacitance results, and linear calibration to a lower value is possible. The useful direct-reading linear range of the Type 722-CB is from 50 $\mu\mu$f to 1100 $\mu\mu$f.

—Ivan G. Easton

SPECIFICATIONS

Capacitance Range: Direct reading on drum and dial, 50 $\mu\mu$f to 1100 $\mu\mu$f.

Accuracy: ± 1 $\mu\mu$f or $\pm 0.1\%$, whichever is the greater.

Correction Chart: A correction chart is supplied, giving corrections at multiples of 100 $\mu\mu$f. The accuracy obtained through the use of this chart is $\pm 0.1\%$ or ± 0.4 $\mu\mu$f, whichever is the greater, for direct measurements. For capacitance differences, the accuracy is $\pm 0.1\%$ or ± 0.8 $\mu\mu$f, whichever is the greater.

Worm Correction Calibration: A correction for the slight residual eccentricity of the worm drive can be supplied at extra charge. The accuracy after worm correction is applied is $\pm 0.1\%$ or ± 0.1 $\mu\mu$f for total capacitance and $\pm 0.1\%$ or ± 0.2 $\mu\mu$f for capacitance differences.

Dissipation Factor: Less than 50×10^{-6}.

Maximum Voltage: 1000 volts peak.

Residual Inductance: Effective series inductance is approximately 0.06 μh.

Series Resistance: Approximately 0.02 ohm at 1 Mc.

Temperature Coefficient of Capacitance: Approximately 0.002% per degree Centigrade for small temperature changes.

Accessories Supplied: Two Type 874-C58 Cable Connectors.

Terminals: Type 874 Coaxial Panel Connectors. In addition, a jack-top binding post is provided as an alternate ground connection.

Mounting: Capacitor is mounted on an aluminum panel finished in black crackle lacquer and enclosed in a shielded hardwood cabinet. A wooden storage case with carrying handle is supplied.

Dimensions: Panel, 8 x 9⅛ inches; depth, 8⅛ inches.

Net Weight: 9 pounds, 10 ounces; 17 pounds with carrying case.

Type		*Code Word*	*Price*
722-CB	Precision Capacitor	CAROM	$205.00

COMING EXHIBITS

PATENT OFFICE SPONSORS EXHIBIT

To promote the role played by invention and the patent system in the development of the electrical and the electronics industries, the U. S. Patent Office has invited sixteen companies to participate in an exhibit to be held in the Main Lobby of the Department of Commerce Building, Washington, D. C., October 13 through November 7, 1958.

Among those exhibiting will be the General Radio Company. Our display emphasizes the invention, patenting, development and improvement, by General Radio, of the variable autotransformer, which this Company manufactures and markets under the trade name VARIAC.®

We cordially invite *EXPERIMENTER* readers to visit this and the many other interesting displays at this exhibit.

NEREM — THE NEW ENGLAND RADIO ELECTRONICS MEETING

NEREM — New England's annual electronics convention and show, sponsored by the Boston and Connecticut Valley Section of the IRE, will be held at Mechanics Building, Boston, November 19 and 20, 1958. An outstanding technical program has been arranged, and more than 100 exhibitors will show their products. You will find it worth while to attend this outstanding meeting.

General Radio, in Booths 153 and 154, will have on display the new instruments that you have read about in the pages of the *EXPERIMENTER* this year. We hope you will drop in to see us.

General Radio Company

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the GENERAL RADIO Experimenter

Since 1915 – Manufacturers of Electronic Apparatus for Science and Industry

VOLUME 32 No. 18 NOVEMBER, 1958

In This Issue

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The General Radio EXPERIMENTER is mailed without charge each month to engineers, scientists, technicians, and others interested in electronic techniques in measurement. 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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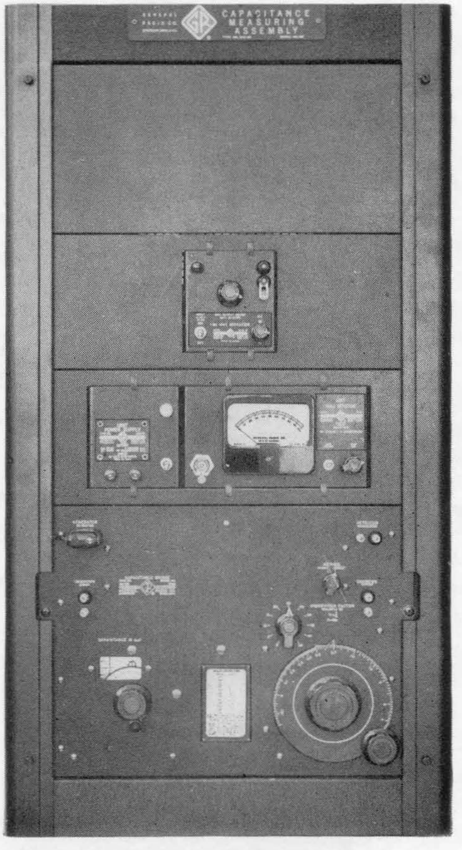
COVER

Testing of coaxial devices can be speeded up greatly through the use of the General Radio Motor-Driven Slotted Line. The photograph shows this line in use in the General Radio laboratories to test coaxial elements for standing-wave ratio.

COMPLETE ASSEMBLY FOR CAPACITANCE MEASUREMENTS AT ONE MEGACYCLE

Both commercial and military specifications for capacitors of 1000 $\mu\mu$f and less call for measurement of capacitance and dissipation factor at a frequency of one megacycle. The TYPE 1610-AH Capacitance Measuring Assembly is a complete set of equipment for making these as well as other 2-terminal measurements.

Accessories are available which enhance the usefulness and convenience of the assembly for specific measurements. For the measurement of small capacitors, particularly disc-ceramic types, the TYPE 1691-A Capacitor Test Fixture is recommended. With the TYPE 1690-A Dielectric Sample Holder, specimens of dielectric materials in the form of ASTM 2-inch or smaller discs can be measured.



The TYPE 1610-AH Capacitance Measuring Assembly consists of:

TYPE 716-CS1 Capacitance Bridge
TYPE 1214-M 1-Mc Unit Oscillator
TYPE 1212-A Unit Null Detector
TYPE 1212-P1 1-Mc Filter
TYPE 1203-B Unit Power Supply
TYPE 480-P5UC1 and TYPE 480-P4U3 Adaptor Panels
Relay Rack Cabinet
Connection Cables and Power Cord

Although calibrated for a frequency of one megacycle, the bridge itself can be used at any frequency between 0.1 and 5 megacycles. A variable-frequency oscillator such as the TYPE 1211-B and a tunable selective detector (or radio receiver) are then required. If corrections are applied, rated accuracy can be attained.

SPECIFICATIONS

Capacitance Range: Direct Method, 100 to 1150 $\mu\mu$f; Substitution Method, 0.1 to 1050 $\mu\mu$f.

Dissipation Factor Range: Direct Method, 0.00002 to 0.56; Substitution Method, 0.00002 x $\frac{C'}{C_x}$ to 0.56 x $\frac{C'}{C_x}$ where C' is the capacitance of the internal standard capacitor at initial balance and C_x that of the unknown.

Accuracy (at one megacycle):

Direct Reading: Capacitance, ±0.1% ±1 μμf when the dissipation factor of the unknown is less than 0.01; Dissipation Factor, ±0.0005 or ±2% of dial reading, whichever is larger, for values of *D* below 0.1.

Substitution Method: Capacitance, ±0.2% or ±2 μμf, whichever is larger; Dissipation Factor, ±0.00005 or ±2% of the change in *D* observed, when the change is less than 0.06.

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

A correction chart for the precision capacitor is supplied, giving scale corrections to 0.1 μμf at multiples of 100 μμf. By using these data, substitution measurements can be made to ±0.1% or ±0.8 μμf, whichever is the larger. For capacitances less than 25 μμf, the error will decrease linearly to ±0.1 μμf. It is also possible to obtain, at an extra charge, a worm-correction calibration with which substitution measurements can be made to an accuracy of 0.1% or ±0.2 μμf, whichever is the larger.

This same accuracy can be obtained with the bridge at other frequencies between 0.1 Mc and 3 Mc, if corrections are made for the effects of residual impedance, and if adequate selectivity is provided for the null detector.

Accessories Available: Type 1690-A Dielectric Sample Holder and Type 1691-A Capacitor Test Fixture.

Power Supply: 105 to 125 volts, 50 to 60 cycles, 100 watts input at 115 v line. Instrument will operate satisfactorily on power-supply frequencies up to 400 cycles, provided that the supply voltage is at least 115 volts.

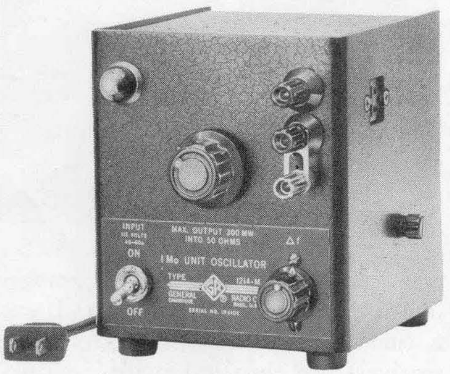
Power input receptacle will accept either 2-wire (Type CAP-35) or 3-wire (Type CAP-15) power cord. Two-wire cord is supplied.

Dimensions: (Height) 43 x (width) 22½ x (depth) 20 inches, over-all.

Net Weight: 150 pounds, approximately.

Type		*Code Word*	*Price*
1610-AH	**Capacitance Measuring Assembly**	SIREN	**$995.00**
	Worm-Correction Calibration for Internal Precision Capacitor	WORMY	**50.00**
1690-A	**Dielectric Sample Holder**	LOYAL	**435.00**
1691-A	**Capacitor Test Fixture**	EDICT	**22.50**

A NEW, 1-MC UNIT OSCILLATOR



The Type 1214-M Unit Oscillator is another of General Radio's line of compact, inexpensive Unit Instruments. This oscillator generates a frequency of 1 megacycle per second and is a useful power source for bridge measurements at that frequency. The Type 1214-M is the oscillator supplied with the Type 1610-AH Capacitance Measuring Assembly.

This oscillator uses a Hartley circuit and is designed for low distortion and high output level. The output can be isolated from ground. A continuous rotary control on the panel of the instrument varies the frequency over a range of ± 1%, while a second control varies the output from zero to maximum.

The built-in transformerless power supply operates from a 115-volt, 40- to 60-cycle ac line.

SPECIFICATIONS

Frequency: 1 Mc.

Frequency Accuracy: ±1%.

Maximum Output: 300 mw into 50 ohms.

Open Circuit Output Voltage: 7 volts.

Distortion: 3.5% with 50-ohm load.
Power Input: 115 v, 40-60 cps.
Power Consumption: 12 watts.

Dimensions: (Height) 5¾ x (width) 5 x (depth) 6¼ inches.
Weight: 2¾ pounds.

Type		*Code Word*	*Price*
1214-M	**Unit Oscillator, 1 Mc**	ATONE	**$75.00**

TYPE 1421 VARIABLE AIR CAPACITORS

The Type 1421-J and Type 1421-K Variable Air Capacitors are new units in the General Radio line of instrument-type, parallel-plate air capacitors. These capacitors are machined from solid metal, in the same manner as the Type 1420 Capacitors.[1] This unique method of construction has many advantages, both electrical and mechanical, over conventional construction methods. Some of these added features are:

Better control of thickness and straightness of plates, since machining is a more precise operation than rolling.

Elimination of cumulative spacing errors by gang milling.

Better concentricity insured by turning and boring a single piece.

Sturdier structure with high mechanical stability, resulting from the integral-plate construction.

[1] H. M. Wilson, "Type 1420 Variable Air Capacitor," *General Radio Experimenter*, Vol. 31, No. 2, July, 1956, pp. 7-10.

Good linearity and control of capacitance magnitude, insured by precise machining.

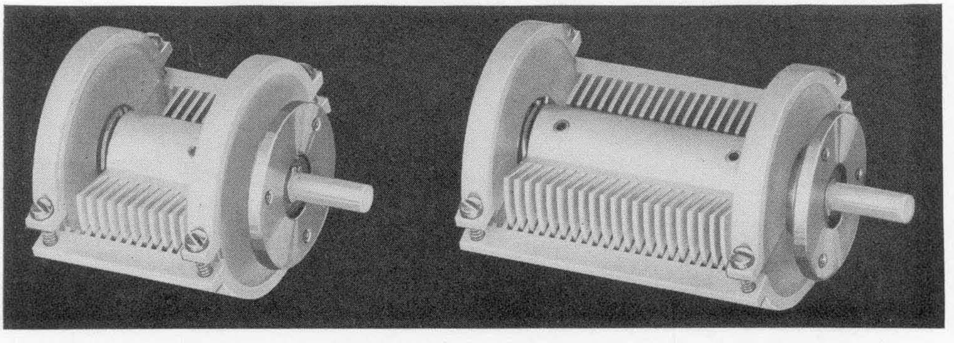
Lower metallic resistance and inductance and low thermal drift, because all the conducting material is homogeneous.

Minimum microphonic tendencies, due to the ruggedness of the plates.

Elimination of irregularities in the capacitance-vs.-rotation curve, because the stator plates are completely joined on their outer peripheries.

Elimination of crazing or other structural failure of the polystyrene dielectric, because the insulators are machined from a cast bar of cross-linked polystyrene for thermal adequacy and are stressed entirely in compression over a wide area.

Insured concentricity of the rotor, since the plates are milled and turned on a centered arbor held by the setscrews exactly as the shaft is secured in assembly.



SPECIFICATIONS

Capacitance Range:

	Nominal *Max.*	*Min.*	*Range for Linear variation*
J	575	22	540 $\pm$20 $\mu\mu$f
K	1120	29	1025 $\pm$25 $\mu\mu$f

The rotor-to-ground capacitance is about 2.5 $\mu\mu$f, and the stator-to-ground capacitance is about 4 $\mu\mu$f, for all sizes. The data in the above table are for the capacitor used as a two-terminal device, with rotor grounded. If stator is grounded, maximum and minimum capacitance values will be decreased by about 4 $\mu\mu$f.

Independent Linearity: The variation of capacitance with angle of rotation does not exceed $\pm$0.3% of full scale. The angular range of linear variation is 160°.

Typical linearity is better than $\pm$0.15%.

Dielectric Losses: For the grounded-rotor connection, the dielectric losses correspond to a D_oC_o product of less than .01 x 10^{-12}. The rotor-to-ground capacitance has a D_oC_o product of 0.1 x 10^{-12}. This loss component is in parallel with the main capacitance only for the grounded-stator connection.

Insulation Resistance: Greater than 10^{11} ohms under standard ASTM laboratory conditions (23°C, 50% RH).

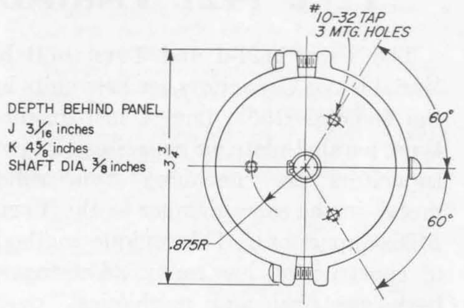
Temperature Coefficient of Capacitance: Approximately +.003% per degree C.

Maximum Voltage: 700 volts peak.

Torque: 2 ounce-inches, maximum, with shaft vertical.

Net Weight: TYPE 1421-J, 1 pound, 8 ounces; TYPE 1421-K, 1 pound, 14 ounces.

Dimensions: See sketch.



Type		*Code Word*	*Price*
1421-J	**575 $\mu\mu$f, max.**	NABIR	**$46.50**
1421-K	**1120 $\mu\mu$f, max.**	NABOB	**50.00**

AN AUTOMATIC-BALANCING CAPACITANCE BRIDGE

The Automatic Capacitance Indicator (Figure 2) produced by Barnes Development Company is a self-balancing unit, which automatically indicates capacitance and dissipation factor in an average time of 7 seconds. This instrument will measure capacitance from 100 $\mu\mu$f to 1.1 μf in four ranges, and dissipation

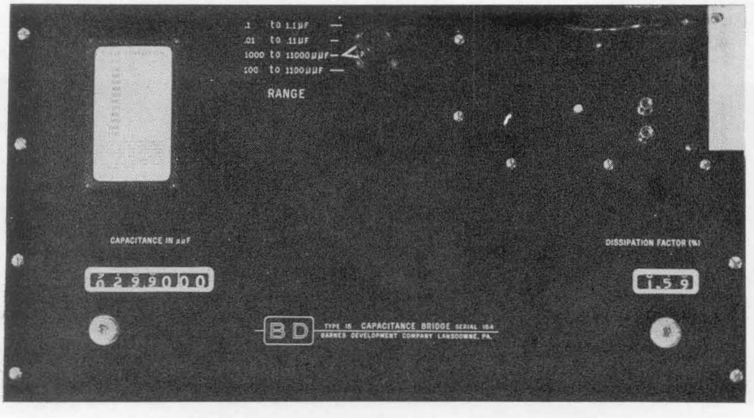


Figure 1. Panel view of the Barnes Automatic-Balancing Capacitance Bridge.

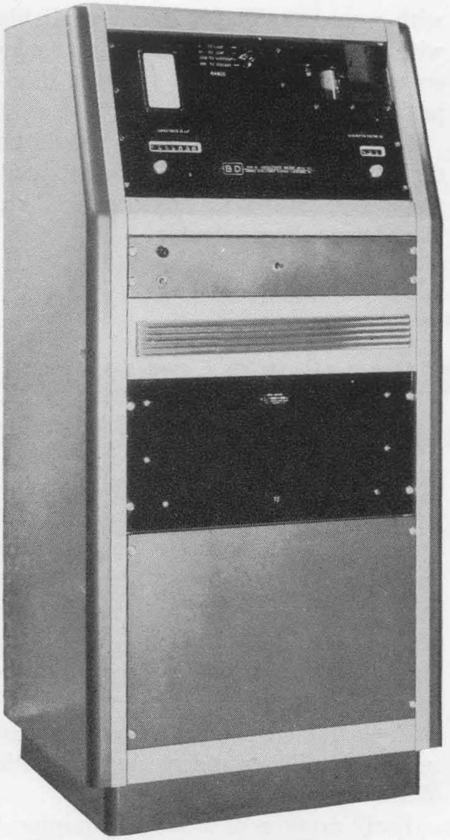


Figure 2. View of the complete bridge assembly.

factor from 0 to 16% in three ranges.

The basic bridge of the indicator is a General Radio TYPE 716-C Capacitance Bridge. Servo motors operating through gears accomplish the automatic balancing. The accuracy of the automatic unit is equal to that of the capacitance bridge (capacitance, ± 0.1% ± 1 μμf × capacitance multiplier setting) when the dissipation factor of the unknown is less than 0.01; dissipation factor, ± 0.0005 or ± 2% of dial reading, whichever is the larger, for values of D below 0.1.

The block diagram of the Barnes Automatic Capacitance Bridge is shown in Figure 3. The error signal from the detector terminals is separated to feed the two channels for "C" and "D" balances. The "C" channel is driven through an attenuator; the "D" channel is fed directly.

After passing through voltage amplifiers the signals are fed to phase detectors. Phase discrimination is achieved by determination of the phase reference voltage fed to the phase detectors. The "C" detector receives an in-phase reference while the "D" detector receives a quadrature voltage.

The outputs of the phase detectors are filtered and then inverted into phase-reversible 60-cycle voltages by synchronous choppers. After power-amplification, these signals are impressed on the control winding of the two-phase servo motors controlling the "C" and "D" channels.

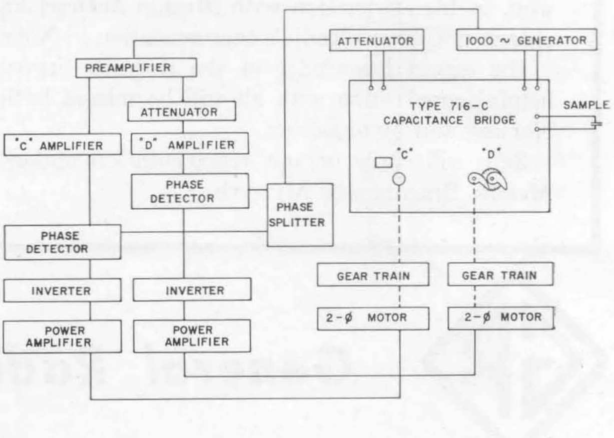


Figure 3. Block diagram of the automatic capacitance bridge.

The balance point is indicated by pilot lights, and the reading is presented directly as a row of figures, with automatically placed decimal points. The Automatic Capacitance Indicator is available with digital readout for use with any punched card or tape data system, or with servo (voltage) readout for producing a graphic record of values with recorders. The bridge is available with either manual or automatic range switching of the capacitance and dissipation-factor ranges.

Inquiries about this bridge should be directed to the Barnes Development Company, 213 West Baltimore Pike, Lansdowne, Pennsylvania.

SMITHSONIAN INSTITUTION TO COLLECT EARLY COMMUNICATIONS EQUIPMENT

As many of our readers know, the famed Smithsonian Institution in Washington has begun an interesting collection of early communications equipment. A good start has already been made.

The Institution is now especially interested in considering for permanent display donations of early wireless equipment dating from the World War I period or earlier and limited to an arc transmitter, an induction coil spark transmitter and all kinds of measuring equipment. In the latter category, we are planning to make several contributions from our own collection. Any of our readers who own or know of the existence of any of these types of pioneer apparatus are urged to communicate with the Editor or directly with the Curator, Division of Electricity, Smithsonian Institution, Washington 25, D. C.

The exhibit will eventually be housed in a new building under construction in Washington.

Early catalogs of old radio equipment are also desired for the historical library.

It is with deep regret that we report the death of Mr. G. Hammerik who, in his association with Maskin-Aktieselakapet Zeta of Oslo, was in charge of General Radio's representation in Norway.

His expert knowledge of the scientific instrumentation field and his helpful cooperation with all will be missed both by his many friends in Norway and by ourselves.

Zeta will carry on our representation under the capable direction of Messrs. Braenne and Myrseth.

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Experimenter

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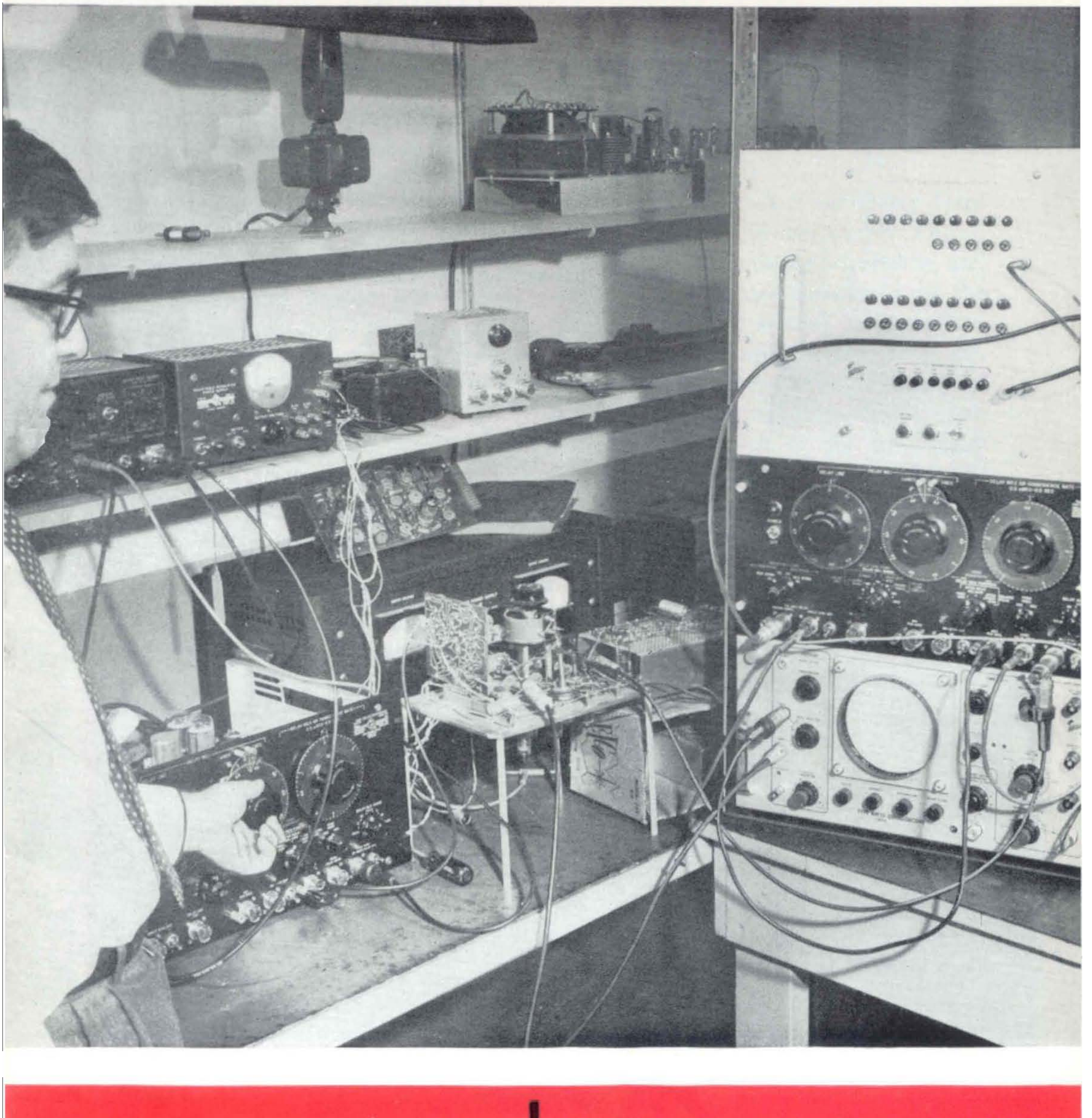
VOLUME 32 No. 19 DECEMBER, 1958



In This Issue | A Precise Time-Delay Generator

the GENERAL RADIO **Experimenter**

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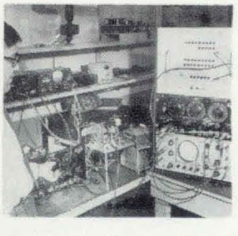
COVER



The new Type 1392-A Time-Delay Generator, described in this issue, brings a new degree of precision and adaptability to delay generation and measurement. Here, two of these generators are used in an experimental system for producing time-coherent tone bursts.

HOW TO KILL TIME — ACCURATELY!

A Generator of Precise Time Delays

The immediate acceptance accorded the Type 1391-A Pulse, Sweep, and Time-Delay Generator[1] upon its introduction in 1955 made it apparent that there was a general desire for precise, versatile, and accurate time-domain instrumentation. The Type 1392-A Time-Delay Generator to be described in this article uses portions of the Type 1391-A as a point of departure and extends its original circuitry to produce a new degree of precision and adaptability in a time-delay system.

The development of circuits for amplitude comparison by Holtje[2] and of highly stable and precise distributed-parameter variable delay lines by Lewis and Frazier[3] came about almost simultaneously with the availability of an abundance of application information resulting from the introduction of the Type 1391-A Pulse, Sweep, and Time-Delay Generator. This combination of new circuits, new components, and field experience has resulted in the development of an instrument that will, we are certain, become a standard for accuracy and reliability in the field of analog time measurements. It is an excellent range calibrator for radar, sonar, and radio navigation systems.

The Type 1392-A Time-Delay Generator produces an accurately known time delay, continuously adjustable, and with 1% or 10-mμsec accuracy over the range from 0 to 1 second. It contains a second, slightly less accurate, delay circuit adjustable over the time range from 0.5 μsec to 0.5 second. This circuit, completely independent of the first delay circuit and capable of being calibrated by it, can be used:

1. As a second "parallel" delay circuit.

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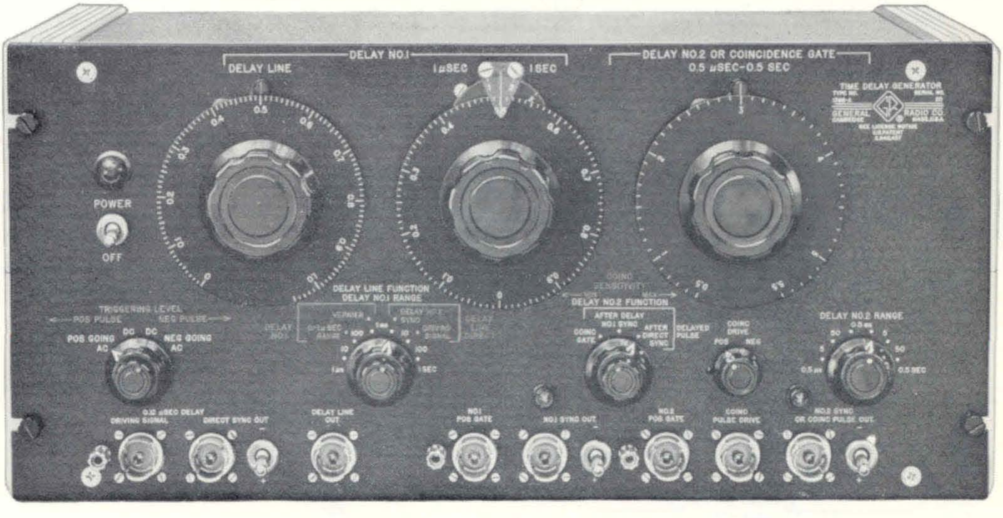
[1] R. W. Frank, "A Versatile Generator for Time-Domain Measurements," *General Radio Experimenter*, 30, 12, May, 1956.

[2] M. C. Holtje, "A New Circuit for Amplitude Comparison," *General Radio Experimenter*, 30, 6, November, 1955.

[3] F. D. Lewis and R. M. Frazier, "Distributed-Parameter Variable Delay Lines Using Skewed Turns for Delay Equalization," *Proc. IRE*, Feb., 1957.

Figure 1. Panel view of the Time-Delay Generator.



2. As a second delay cascaded after the first delay in time.

3. As a coincidence "and" circuit enabling-gate started by the first delay. Since the gate can be reduced in duration to 0.5 μsec, timing marks of PRF in excess of 1 Mc can be selected.

We believe that this is the first commercially available coincidence system capable of delay steps below 1 μsec in an analog system.

The fact that this instrument is an analog generator capable of producing precise delays should be emphasized, because only in an analog system can the delay be produced from a start command pulse without some sort of gating transient or quantization error. With this new delay generator, it is also possible to establish delays with quartz-crystal-oscillator-controlled precision in any time unit, as in range (yards or miles). The advantage of this system should be immediately obvious to those working with radar, with rocks, with water, or with bats.

Another advantage accruing to the analog delay-circuit system is that the delay can be varied by an external control voltage. The amplitude comparator developed for this instrument (and described below) has a very high input impedance for the reference voltage and permits one to inject an external signal to control the magnitude of the delay over a decade. This makes the unit a linear time modulator for the production of pulse duration modulation, pulse position modulation, or plain jitter of known and controlled magnitude over modulating frequencies ranging from 3 c to 100 kc.

DELAY LINE

Figure 2 shows the circuit system reduced to its simplest form. The delay line is shown connected in one of its four possible positions in the system. Here it acts as a 0 to 1-μsec vernier on the electronic portion of the Delay 1 circuit. It can also be switched to:

1. Act as a first (0 to 1-μsec) range to produce a delay between the direct sync pulse and the Delay No. 1 sync pulse.

2. Delay by 0 to 1 μsec the Delay No. 2 sync pulse (see page 7).

3. Delay directly any waveform applied to the PRF drive terminals.

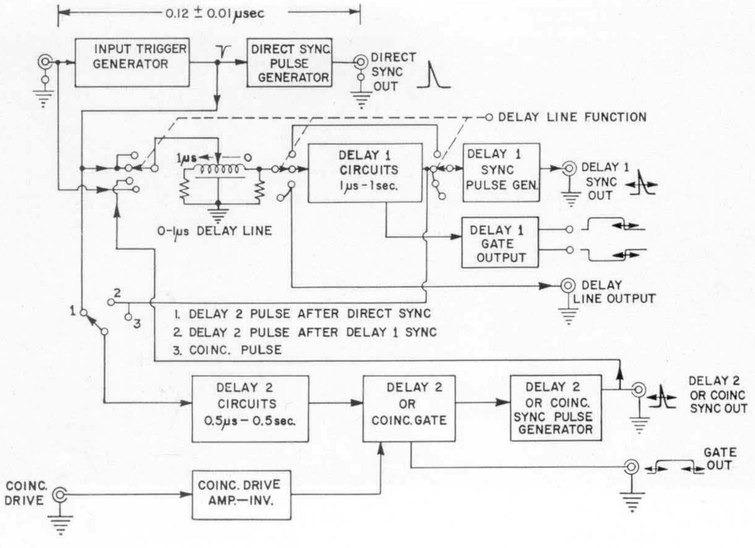


Figure 2. Simplified block diagram.

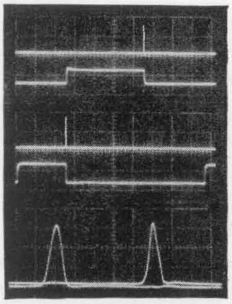


Figure 3. Typical waveforms.

INPUT CIRCUIT

The input circuits comprise a trigger generator, which will accept almost any waveform having a peak amplitude of about 0.3 volt at frequencies ranging from dc to over 300 kc. This circuit produces a pretrigger, which is fed to the two independent delay circuits and an output sync pulse of standard shape, having a base duration of approximately 0.1 μsec, an amplitude of 20 to 25 volts, positive or negative, and a source impedance of 93 ohms (Figure 3c). The sync pulses produced by the Delay No. 1 and Delay No. 2 circuits are similar.

DELAY NO. 1 CIRCUIT

Electronic circuits which will produce accurate, linear, and precisely controlled delays of less than 1 μsec are difficult to build and even if realizable are very expensive and complex. The first range, up to 1 μsec, is timed by a Type 1477-J1000 delay line. This line has an accurate, individually calibrated, four-inch dial graduated in 0.01-μsec intervals and produces a delay between the direct sync pulse and the Delay No. 1 sync pulse of 0 to 1 μsec with an accuracy of ± 0.01 μsec.

To produce delays over 1 μsec, the electronic circuits are switched in cascade with the line as shown in Figure 2. These circuits are shown in some detail in Figure 4. They comprise a high-speed bistable gate, a precise linear sweep generator, an amplitude comparator, and a reset trigger generator followed by the Delay No. 1 sync-pulse generating circuit. When opened by the trigger from the delay line, the bistable gate starts the sweep and causes the associated buffer to produce an output pulse (positive 5 volts and negative 60 volts), which marks the delay interval. The sweep circuit, a highly linear "bootstrap," produces a saw tooth whose slope is determined by the decade range switch. This sweep of predetermined slope rises from a minimum of 10 volts to a maximum of 110 volts. The sweep is fed to one grid of the amplitude comparator, while the reference voltage from the 10-turn range potentiometer is fed to the other. When the sweep and reference voltages are equal, the comparator triggers and produces a reset pulse for

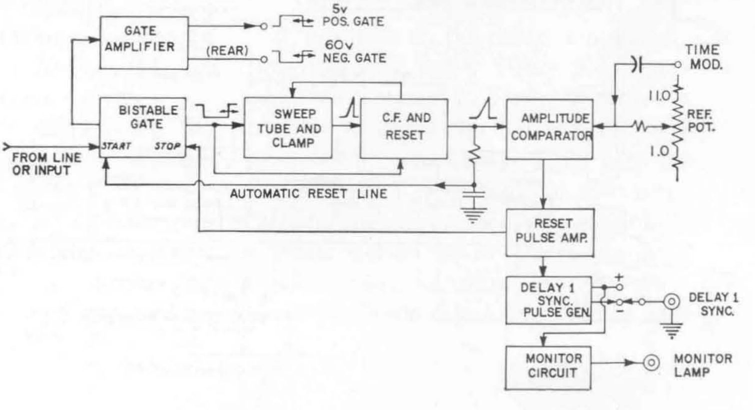


Figure 4. Functional diagram of Delay No. 1 circuits.

the gating bistable multivibrator and at the same time produces the Delay No. 1 synchronizing pulse. Thus, the loop consisting of the bistable gate, sweep generator, amplitude comparison circuit, and resetting circuits assumes a monostable characteristic.

Stability

Both long-term and short-term stability of delay has been assured by the use of precision resistance-capacitance components to establish the sweep slope; by the use of highly regulated and filtered dc voltages for the comparison reference voltage, plate supply, and heater voltage in the sweep generator and amplitude comparator; and by careful circuit design, which minimizes the effects of tube changes and tube aging on the accuracy of the circuit (Table I).

Convenience Features

Both sweep and reference grids of the amplitude comparator are at high impedance, and provision has been made to permit the superposition of a modulating voltage on the reference voltage. A jack on the rear panel of the instrument provides for the injection of this voltage. The low-frequency cutoff of this coupling network is 3 cycles, while the high-frequency performance will, of course, be determined only by the slope of the sweep voltage.

Several other features have been provided to increase the convenience of the Delay No. 1 circuit group in application. A 5-volt positive-gate waveform, corresponding to the delay interval set, is available at low impedance from a coaxial connector on the front panel, so that the delay interval can be observed on an oscilloscope over the entire range of delays where the 0.1-μsec sync pulse may not be visible because of its brief duration (Figure 3b). This waveform, with a negative polarity and a 60-volt magnitude, is provided on the rear panel to drive auxiliary pulse-generating equipment such as the General Radio TYPE 1219-A Pulse Amplifier. In addition to this gate waveform a neon indicator lamp monitors the presence of the Delay No. 1 sync pulse.

DELAY NO. 2 CIRCUITS

The second group of delay circuits is identical in principle with the first (Figure 5). It lacks only the features of high accuracy and resolution. The delay reference voltage is provided by a 4″ single-turn potentiometer with a resolu-

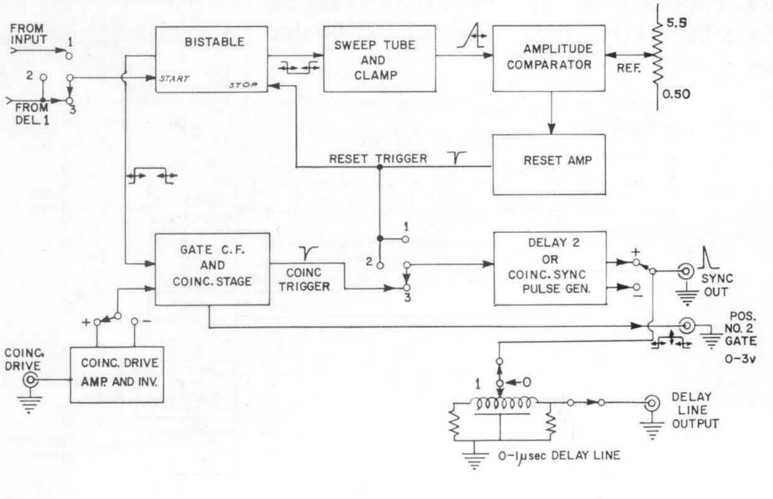


Figure 5. Functional diagram of Delay No. 2 circuits.

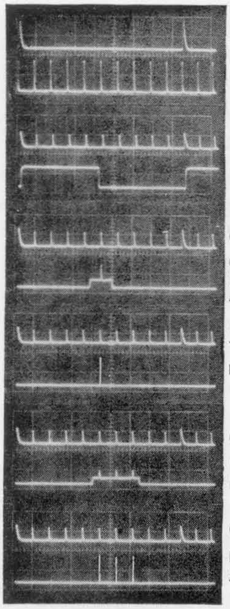


Figure 6. Coincidence circuit used for multiple pulses.

tion of 1 part in 2,000, and the accuracy is 3% everywhere over the six decade ranges from 0.5 μsec to 0.5 second.

The action of this circuit can be initiated either by the direct sync pulse or by the sync pulse of the first delay circuit, so that the delay is either concurrent with the first delay or is initiated after the first delay interval. It should be pointed out here that, since the delay circuits can always be started concurrently, the first, more accurate delay can be used to calibrate the second.

Outputs from the second delay circuit are similar to those of the first: A 5-volt positive gate and, marking the end of the delay interval, a 20-volt 0.1-μsec pulse at 93-ohm impedance level.

When the function switch for the second delay circuit is in "coincidence" position,

1. The delay gating circuits are started after the Delay No. 1 sync pulse.

2. The Delay No. 2 sync-generating circuits are connected to be triggered by the action of an "and" circuit.

The "and" circuit requires the simultaneous existence of the Delay No. 2 gate and an externally applied pulse to provide a trigger for the sync circuits. Photographs of the gate and output sync pulses are in Figure 6. An over-all photo of a precision delay-generating system involving the Type 1392 Time-Delay Generator and the Tektronix 180-A Time-Mark Generator, as a source of crystal-controlled time-coherent synchronizing pulses, is shown in Figure 7.

Since the shortest duration of the Delay No. 2 gate has been set at 0.5 μsec, it is possible to select a single pulse from a timing pulse train whose frequency may be as high as 1.7 Mc. Precise 1-μsec steps of delay are easily obtainable using the Tektronix 180-A as a crystal-controlled time-mark generator. If the gate produced by the Delay No. 2 circuits is made longer than the interval between the timing pulses from the marker generator, a group of pulses will be produced at the Delay No. 2 synchronizing-pulse terminal (Figure 6, (e) and (f)). The maximum rate at which pulses can be produced in the burst is to some extent determined by the amplitude and rise time of the trigger pulses provided by the timing source. The pulses produced by the Tektronix 180-A, for example, will typically produce a 200-kc burst.

DELAY LINE CONNECTION

In addition to the delay-line connection described under Delay No. 1, the connection shown in Figure 5 makes it possible to delay the No. 2 sync pulse produced by coincidence. Thus, with the combination of a precise coincidence circuit and the passive line, it is possible to generate delays up to several hundred microseconds, accurate to 0.01 μsec. Here, we should inject one word of cau-

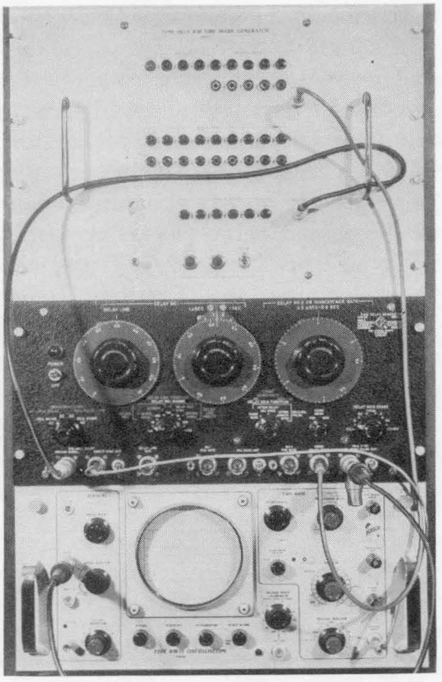


Figure 7. Precision time-generating system, consisting of the Type 1392-A Time-Delay Generator and the Tektronix 180-A Time-Mark Generator

tion. Timing generators like the Tektronix 180-A derive their lower frequency outputs by feeding each pulse-divider circuit from the next higher order divider. Since the trigger pulse has a finite and somewhat variable rise time (depending upon the bandwidth of the divider involved), there is an accumulation of delay between pulses within the timing chain. (One possible use for the coincidence circuit in the TYPE 1392-A Time-Delay Generator is, of course, the removal of this delay.) In order to make the steps of delay produced by the coincidence system exactly correct, some delay must be inserted in the path of the high-frequency pulses,[4] or the error must be measured and subtracted from the delay readings.

[4]The General Radio TYPE 314-S86 0.5 μsec Variable Delay Line is useful in equalizing these residual delays (independently of the Time-Delay Generator). F. D. Lewis and R. M. Frazier, "A New and Better Variable Delay Line," *General Radio Experimenter*, 31, 7, October, 1956.

The delay line connection shown in Figure 2, in which the 0 to 1-μsec line is connected directly from the input terminal to the line output terminal, is also useful in many ways. In this case, the line will pass whatever waveform is available at the input terminals. This will, of course, make the input impedance of the instrument equal to the impedance of the line at about the 250-ohm level. Although the simplified switching of Figure 2 does not so indicate, the electronic portions of the TYPE 1392 are still operative. One use for the delay line in this connection is to permit a check on the delay generated in the input circuits. This delay is nominally about 0.12 μsec between a fast driving pulse and a direct sync pulse.

THE DELAY NO. 1 AMPLITUDE COMPARATOR

The amplitude comparator used in the Delay No. 1 circuit is new[5] and certainly useful elsewhere, so it will be described in some detail. This circuit was designed primarily for operation at repetition rates higher than those at which the comparator described by Holtje[2] will operate and for as great a dynamic range.

In the TYPE 1392-A, the amplitude comparator will operate at rates as high as 300 kc and over a range from 0 to 120 volts. It has the additional advantage of a very high input impedance at both signal and reference terminals. The complete circuit is shown in Figure 8. It is convenient to view the circuit as a bistable multivibrator (V-1), in which a differential amplifier (V-2) replaces the usual shunt-arm resistors of the cross-coupling network. Tube V-3 is a current source for V-2 and is only necessary

[5]Patent applied for.

when the maximum dynamic range at a given supply voltage is desired.

The circuit operates in the following manner: Suppose initially that the signal voltage is more negative than the reference voltage, then the current through the left-hand side of V-2 will be less than that in the right, and the grid voltage on V-1, left, will be more positive than V-1, right. V-1, left, will be on. This situation will persist until the reference and signal voltages are equal. At that point, the grid voltages of V-1 would be equal if it were not for the fact that V-1 is a bistable using R_{C1}-R_{C2} and the plate resistance, R_p, of V-2 as cross-coupling resistors. Hence, at equal grid voltages on V-2, V-1 will remain stable as initially. Now, as the grid of V-2 is raised above the reference voltage, V-2, left, conducts more heavily and, by the differential connection of V-2, V-2, right, conducts less heavily. At low frequencies the gain to the bistable grids is[6]:

$$K = \frac{R_c}{r_{p2} + R_c}$$

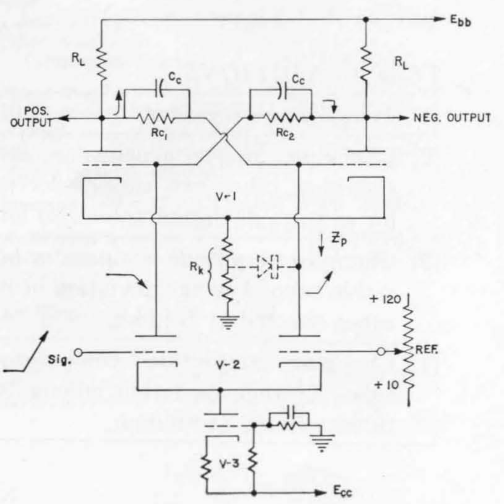
and only a sufficient voltage needs to appear on the bistable grids to cause the bistable circuit to switch. The increase in grid voltage on V-1, right, reduces the amount by which this tube is below cutoff while the decreasing voltage on the left-hand grid of V-1 increases its plate voltage by the gain of this side of the flip-flop tube. It is therefore the left-hand side of the bistable going off which primarily contributes to the switching action.

The hysteresis which the circuit exhibits at the grid of the differential section is controlled by the magnitude of the excess grid voltage of the off flip-flop tube. This quantity is determined by the current in the differential amplifier and, since this current is set by V-3, a well-stabilized stage, only very small excess voltages need be tolerated. Typical values of hysteresis for the circuit shown in Figure 8 are 10 to 100 mv.

An additional advantage of this circuit over a single-ended design is apparent at this point. The circuit is balanced with respect to changes in plate current caused by heater voltage fluctuations. Drift of the trigger voltage with line-voltage changes is reduced by about an order of magnitude over more conventional single-ended designs.

The action of the circuit at high frequencies is more complex than that described above. Space will not permit a detailed analysis here, but the limits on performance are clear. The cross-coupling capacitors will cause the loading of the differential amplifier to increase until, at very high frequencies, the load will be the plate load resistors, R_L, and the stray capacitances. Comparators for the megacycle range must be built with a differential amplifier capable of attaining high g_m and will, of course, have lessened sensitivity.

Figure 8. Functional schematic diagram of the amplitude comparator.



[6]Valley and Wallman, "Vacuum-Tube Amplifiers," Radiation Laboratory Series, Number 18, McGraw-Hill Book Co., 1948.

For high-frequency operation, overshoots of the grid waveform due to the flip-flop switching action should be eliminated both by reduction of the cross-coupling capacitors and by the application of positive and negative grid clamps. In the unit designed for the TYPE 1392-A, the flip-flop does not draw grid current, so only a negative clamp is used and, since the input waveform is unidirectional, only one clamp is necessary to improve recovery time.

Some characteristics of the comparator used in the Time-Delay Generator are shown in Table I. The hysteresis of this unit is of the order of 0.1 volt. The differential amplifier is heated with regulated dc both to minimize drift with changing line voltage and to reduce the hum-jitter in the delayed signal. The presence of the dc reduces the effects of 20% of variation in line voltage to an error in delay of 1 part in 10,000 and aids in attaining the specified jitter figure of 1 part in 30,000 at the worst. (Jitters lower than 1 part in 100,000 have been observed.) On the 1 to 11 msec range, the TYPE 1392-A has exhibited an absolute accuracy of delay of better than 0.1% over periods in excess of a week in continuous operation.

APPLICATIONS

The TYPE 1392-A can be applied anywhere that accurate delays are desired, either for the precise measurement of delay or for its generation. Its range of available delay plus the fact that it can be triggered by almost any sort of signal should make it an acceptable range calibrator for radar, sonar, and radio navigation systems. The ease with which the time of occurrence of the Delay No. 1 sync can be modulated renders it an ideal device for the simulation of a "noisy" target. The longer ranges of delay and the presence of the second delay with its associated gate signal should make the TYPE 1392-A a very useful tool for physiological nerve-transmission studies.

The use of the coincidence circuitry has been described in detail in connection with the processing of coherent signals. This system has another entire field of application in processing signals which are not coherent. Counting, time-vernier measurements, and the production of partial coherence between otherwise incoherent signals are examples of these applications. In the latter application, for example, coherent tone bursts in the megacycle range have been timed

TABLE I

TEST CONDITIONS	*ERROR AT 5 msec DELAY*
(1) Temperature transient.	.005%/° C.
(2) Changing sweep generating and clamping tube, two tubes selected for average deviation among 25 lot.	0.15% (worst tube .45%)
(3) Changing amplitude comparator bistable tube. Average deviation in 10 tubes selected at random.	.24% (worst tube 0.3%)
(4) Changing amplitude comparator tube. Average deviation among 10 tubes selected at random.	.47% (worst tube 0.9%)

by feeding the high frequency desired within the burst to the coincidence drive terminals and the burst recurrence rate signal to the PRF drive terminals. Specific uses are:

1. Calibration and precision sweep delay for oscilloscopes.
2. Radar target simulator.
3. Sonar target simulator.
4. Loran received pulse simulator.
5. Geophysical echo simulator.
6. Precision pulse timer — used with GR Type 1219-A Pulse Amplifier.
7. Delay and gate simulator for use with electronic computing systems — used with Tektronix 180-A Time-Mark Generator to produce jitter-free pulses of crystal-controlled accuracy for any of above applications.

ACKNOWLEDGMENT

The development of this instrument resulted from the collaboration of many people. The author wishes to acknowledge specifically the aid of M. C. Holtje, on general design and power supply problems; of Carl Alsen, in finish model testing; of H. C. Sterling, in mechanical design; of H. T. Anderson, Production Engineer; and of William Pote, Laboratory Test Engineer.

— R. W. Frank

SPECIFICATIONS

INPUT SYSTEM

Input Signal:
Sine Wave, 0.1 volt rms.
Square Wave, 0.3 volt, peak-to-peak.
Pulse (negative or positive), 1 volt peak.
Ac or dc, input trigger threshold control provided.

Frequency: dc to over 300 kc.

Time Delay (Input to direct sync): 0.12 ±0.02 μsec.

Direct Sync Pulse:
Amplitude, 15 volts or more, positive or negative.
Duration, 0.13 ± .02 μsec.
Impedance, 93 ohms or less.

DELAY NO. 1

Delay Range: 0-1.1 sec in seven ranges.

Accuracy: 1 μsec-1.1 sec range, ±1% of dial reading; 0-1 μsec range, ±0.01 μsec.

Stablity: Jitter, 1:30,000 at worst
Drift, 1:10,000 with 20% line variations

Resolution: 0-1 μsec, 0.004 μsec
1 μsec, 1 sec 1:8800

Duty Ratio Effects: Less than dial accuracy to 60%
5% at duty ratio of 80%

Delay One Sync:
Duration, 0.1 ±0.2 μsec
Amplitude, 25 volts or more, positive or negative
Impedance, 93-ohm output impedance
Monitor Lamp
Max. PRF, 0-1 μsec range, 300 kc;
1 μsec-1.1 sec range, 250 kc (at 1 μsec).

DELAY NO. 2 OR COINCIDENCE CIRCUIT

Range: 0.5 μsec to 0.5 sec (six decade ranges).
Accuracy: ±3% of dial reading.

Stability: Jitter, 1:20,000
Line Drift, 1:5000 for 20% line change

Resolution: 1:2,000

Delay Two Sync:
0.13 μsec ±0.02 μsec
20 volts or more, positive or negative
93 ohms
Monitor Lamp

Duty Ratio Effects: Full scale, less than dial accuracy at 60% duty ratio; bottom of scale, less than dial accuracy at 20% duty ratio. Maximum PRF, 300 kc.

Coincidence:
Input, positive or negative pulse, 5 volts or over
Input frequency, 1 cps to 1.7 Mc (for single pulse selection)
Input pulse rise time, .1 μsec or less at 5v.

Power Supply: 105 to 125 (or 210 to 250) volts, 50-60 cycles, 180 watts at 115 volts. Power receptacle will accept either 2-wire (Type CAP-35) or 3-wire (Type CAP-15) Power Cord. Two-wire cord supplied.

Dimensions: 9 x 19 x 12 inches (rack mount).

Accessories Supplied: Type CAP-35 Power Cord; spare fuses; test lead; 4 Type 874-C58 Cable Connectors.

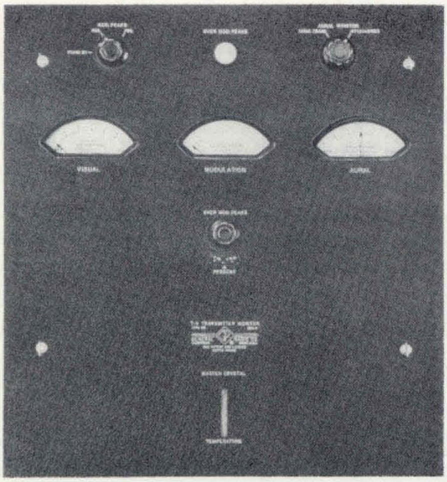
Net Weight: 35 pounds.

Type		*Code Word*	*Price*
1392-AM	**Bench Model**	ENTRY	**$985.00**
1392-AR	**Relay-Rack Model**	EXTOL	**985.00**

U. S. Patent No. 2,548,457.

2-WEEK DELIVERY ON ALL GENERAL RADIO MONITORS

FCC Type Approved



The Type 1184-A Television Station Monitor.

With the continuing trend to higher technical standards in broadcasting, a monitor that barely meets *today's* requirements is likely to become obsolete too soon to give good value. General Radio Monitors are designed *beyond* mere legal minimum requirements for today's use. Thus protected against early obsolescence, these new instruments promise *long-term* value that far outweighs initial cost consideration.

Now is an excellent time for stations which have moved or anticipate moving or modernizing to purchase monitoring equipment.

General Radio Monitors express a wholly new concept in mechanical design that gives convenience never before attained in an instrument of this type. Every operation in the installation, use, and maintenance of these new monitors can be handled from the front. You never need to go behind the instrument rack. The monitor slides out and tilts forward and back for easy access to tubes and adjustments. Chassis pin jacks permit a rapid check of current and voltage at critical points in the circuit. The unique chassis marking is so comprehensive that most maintenance can be performed without need for an instruction book or a circuit diagram.

Type		*FCC Type Approval Number*	*Code Word*	*Price*
1184-A-A	Television Transmitter Monitor — Channels 2-13	3-105	GIANT	$3200.00
1184-A-A	Television Transmitter Monitor — Channels 14-83	3-105	GIANT	3450.00
1181-BT	TV Color-Subcarrier Monitor	not required	MAJOR	1025.00
1181-B	AM Frequency Deviation Monitor	3-106	MALAY	1025.00
1931-B	Amplitude-Modulation Monitor	3-107	TARRY	625.00

CHOICE OF PANEL FINISH

General Radio Black Crackle is standard. Other standard finishes to match transmitter equipment can also be supplied.

General Radio Company

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