the GENERAL RADIO Experimenter

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TO

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COVER

CBS Laboratories' new Stereophonic Test Record, STR-100, is designed for use with the GR Type 1521-A Graphic Level Recorder. See page 15.

A TEST OSCILLATOR FOR S-BAND MEASUREMENTS

With the TYPE 1360-A Microwave Oscillator, the frequency range of General Radio continuously tunable oscillators is extended to 4 Gc. Although not itself one of the Unit Instruments, it supplements this popular line of test equipment that provides the user with versatility and quality at a reasonable price.

Every effort has been made to give this new oscillator, within the design limitations dictated by tube choice and price, maximum usefulness as a driver for slotted lines, as the local oscillator in a heterodyne detector, and as a general-purpose power source for measurements on components and systems at microwave frequencies.

The frequency range of the oscillator is 1.7 to 4.1 Gc. Output power is 100 milliwatts or more over most of the frequency range. Internal 1-kc square-wave modulation is provided, as is also a narrow-band sweep at both 1-kc and the power-line frequency. Modulation from external sources can be fm, square-wave, or pulse.

RF CIRCUIT

The microwave oscillator in the TYPE 1360-A is a TYPE 5836 Reflex Klystron in a coaxial cavity with a noncontacting tuning plunger. The frequency range is split in order to obtain a maximum range of interference-free operation and, at the same time, to provide maximum output power. For the higher frequency range (2.6-4.1 Gc), the cavity length is $\frac{3}{4}$ wavelength and the klystron is operating in its $2\frac{3}{4}$ repeller mode. In the lower frequency range (1.7-2.8 Gc), the numbers are $\frac{1}{4}$ and $1\frac{1}{4}$ respectively. The range switching is controlled automatically by the main frequency dial (center of panel, Figure 1), and, since the higher frequency range requires the longest cavity, the lower frequency range starts at the top end of the higher one. The two ranges are separated on the dial by different colored scales, and a

Figure 1. Panel view of the Microwave Oscillator.

pilot lamp indicates which scale is to be read. The tuning law of the repeller is matched to that of the cavity by a specially shaped, high-resolution potentiometer, and the final adjustment is made by trimmer rheostats. In series with the repeller potentiometer is also a small rheostat for fine frequency adjustment (△F knob), with a range of approximately 1 Mc. However, since adjustment of this rheostat may seriously harm the tracking of repeller voltage and cavity tuning when the oscillator is square-wave modulated, the △F control is disabled under that condition.

OUTPUT CIRCUIT

The output control is at the lower right of the panel. The output power is a function of frequency, as shown in Figure 2, and is more than 100 milliwatts over most of the frequency range. At the very low-frequency end of the range, it may be as low as 20 milliwatts, which corresponds to 1 v into a 50-ohm load. At the upper end of each frequency range, it is possible to overload the klystron oscillator, and an output monitor is provided to warn against this condition. The monitor enables the user to extract the maximum output power at any setting of the frequency dial. The output power is controlled by a retractable pick-up loop in the oscillator cavity. The dial plate on the attenuator (pick-up loop) knob is calibrated in arbitrary units (maximum coupling is 100), but the divisions are equivalent to decibels except in the nonlinear range of the attenuator where the output power is greatest. The area where over-coupling is possible at some frequencies, even with a 50-ohm load, is indicated on the attenuator dial by the legend WATCH OUTPUT MONITOR. The output monitor is fed from a directional coupler in such a manner that it is quite insensitive to load changes. A variable resistor in series with the meter serves as a sensitivity control.

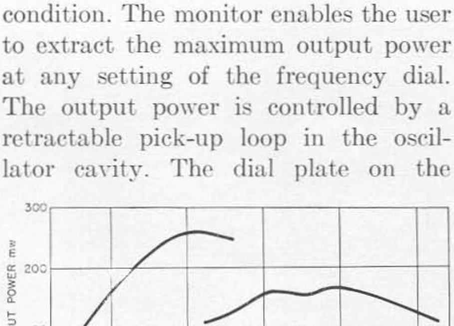


Figure 2. Typical variation in output with frequency.

The output connector is a 50-ohm TYPE 874 Locking Connector which will permit a semipermanent attachment of a cable or an adaptor to some other type of coaxial connector.

MODULATION

Since the most-used type of operation (other than CW) for a test oscillator at these frequencies is 1-kc square-wave modulation, this is provided internally. To facilitate matching the frequency to the filter in the detector system, a screwdriver adjustment on the front panel can vary the modulation frequency approximately ±5%. In the STANDBY position of the modulation switch, between the CW and 1-KC square-wave positions, the rf energy is shut off.

Narrow-band, linear sweep is provided at power-line frequency and 1 kc. This can be used for checking receivers and other narrow-band devices, and is also very useful for realigning the klystron oscillator after a change of klystron tube. When the klystron is being swept internally, oscilloscope synchronization can be obtained through negative trigger pulses from the oscillator.

Square waves for modulation at other

frequencies can be applied by an external source. Recommended sources are the Type 1210-C Unit RC Oscillator and the Type 1217-B Unit Pulse Generator. The latter is also recommended for pulse modulation. For external frequency modulation, the modulating signal is applied across a series resistor in the repeller lead.

The three block diagrams in Figure 3 illustrate how modulation is accomplished, and it should be noted that different methods are used for pulse modulation and square-wave modulation. For pulse modulation (Figure 3b), the klystron beam current is interrupted by application of a negative voltage to the normally positive biased grid. During the first couple of hundreds of milliseconds after the current is turned on

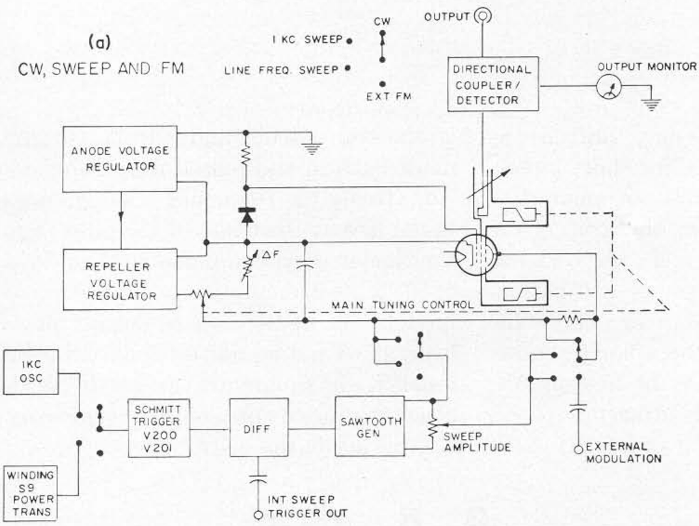


Figure 3a. Schematic for CW, sweep, and frequency modulation.

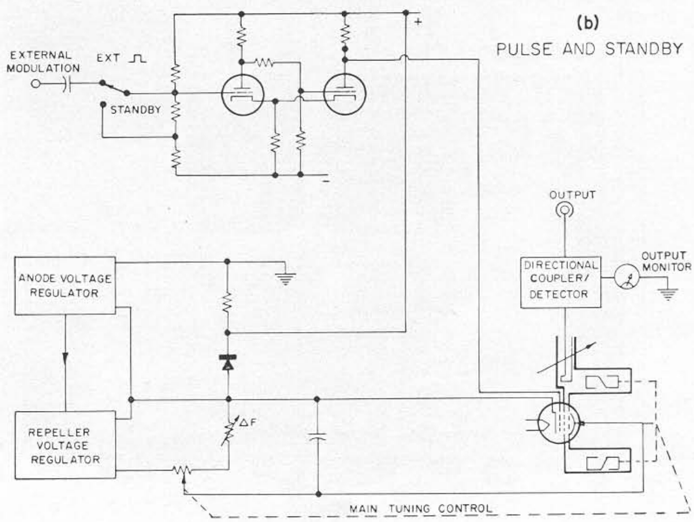


Figure 3b. Schematic for pulse modulation and standby operation.

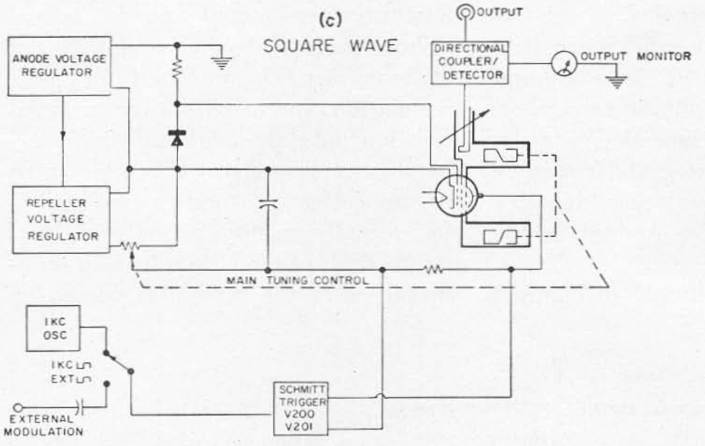


Figure 3c. Schematic for square-wave modulation.

again, the frequency may shift by as much as a megacycle. For short pulses, the frequency shift does not amount to very much, and, since the grid is the pulsing member that gives the best rise time, it is used for pulse modulation. For square waves, however, where the on-period may last for a longer time, the frequency shift may be undesirable, and it was found advantageous to repeller-modulate the klystron. The tube is tracked outside and parallel to the mode pattern and pulsed into the mode. Admittedly, a frequency change does take place at the edges of the pulse, but the time is short compared to the period of the most commonly used square-wave modulation. If the user so desires, however, short pulses can be applied to the repeller, or square waves to the grid, after readjustments of the symmetry control inside the instrument.

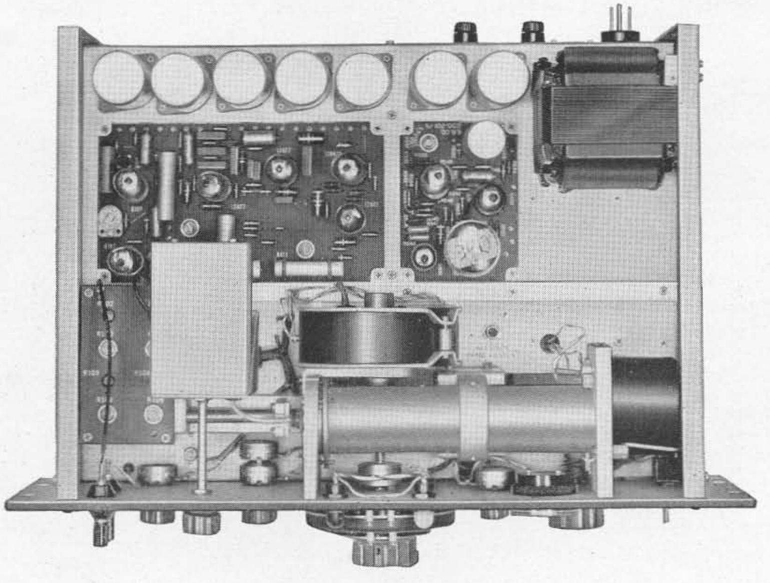


Figure 4. Interior view.

The internal narrow-band sweep at line-frequency or 1-kc rates is produced by application of an internally generated sawtooth voltage to the repeller. Owing to the inherent characteristics of the klystron, the sweep is practically linear in frequency.

External modulation voltage for fm is applied to the repeller through a 0.047-μf capacitor. The input impedance is 400 kilohms shunted with 70 pf.

POWER SUPPLIES

Both the cathode and repeller of the klystron are fed from well-regulated supplies. The repeller heater is fed from a dc supply, which, like the power supply for the modulator circuit, is unregulated but adequately filtered. The bias voltage for the klystron grid is taken from a Zener diode in order to make the voltage constant and independent of the grid current, which varies considerably from tube to tube.

MECHANICAL FEATURES

The Type 1360-A is packaged in a 7-inch relay-rack cabinet and can be obtained either with end frames for bench use or with support fittings for rack mount. Figure 4 shows the instrument with the cabinet removed. Almost all the electronic components are mounted on etched boards which are easily accessible from both sides. Tube replacements, including the klystron, do not require any tools, and precautions have been taken to prevent service personnel from accidentally touching high-voltage terminals.

The noncontacting tuning plunger is supported by a carriage with long-life reinforced Teflon bearings, and the rack and pinion drives for the tuning plunger and attenuator require a minimum of lubrication.

—Per A. Bergstad

CREDITS

The Type 1360-A Microwave Oscillator was developed by Per A. Bergstad, author of the foregoing descriptive article. William G. Cooper, Eduard Karplus, Charles S. Kennedy, Benedict O'Brien and Robert A. Soderman have all contributed to the final design. George A. Clemow was responsible for the mechanical design.

— Editor

SPECIFICATIONS

FREQUENCY

Range: 1.7 to 4.1 Gc in two ranges, 1.7 to 2.8 Gc and 2.6 to 4.1 Gc.

Fine Frequency Control (△F): Order of 1 Mc, but not functioning for square-wave modulation.

Accuracy: ±1%.

Stability: Warm-up drift is approximately 0.15% during the first hour, total drift approximately 0.25%. After warm-up, frequency is stable within approximately 5 ppm.

Residual FM: Approximately 0.5 ppm in the lower frequency range and 0.2 ppm in the higher. Dominant frequencies are 60 and 120 cps (with 60-cycle line frequency).

OUTPUT POWER

Typically more than 100 mw above 2 Gc. Total variation in maximum output with frequency is 20 to approximately 300 mw.

Attenuator: Relative calibration only.

INTERNAL MODULATION

Narrow-Band Sweep: 1 to 3 Mc maximum at 1 kc and power-line frequency. Negative trigger pulse supplied.

Square-Wave: 1 kc, adjustable approximately ±5%.

EXTERNAL MODULATION

FM: Sensitivity approximately 0.2 Mc per volt, input impedance, 400 kilohms and 70 pf (ac only).

Square-Wave: 50 cps to 200 kc, 12-v (rms) sine wave or 20-v (peak-to-peak) square wave; 20% minimum duty cycle from external source. Input impedance greater than 100 kilohms.

Pulse: Rise and fall times approximately 0.2 μsec, minimum length approximately 0.5 μsec, jitter may be 0.2 μsec. Input impedance 100 kilohms; driving-pulse amplitude, 20 v (peak-to-peak); maximum duty cycle 20%.

SPECIFICATIONS (*Cont.*)

GENERAL

Terminals: RF output, Type 874 Locking Connector. Modulation, binding posts.

Mounting: Bench or relay rack.

Power Input: 105 to 125 (or 210 to 250) volts, 50 to 60 cps, 85 watts. Instrument will operate satisfactorily (except for line-frequency sweep) at power-line frequencies up to 400 c.

Tube Complement: Two each 6197 and 12AT7, one each 6AN8, 6AV5GA, 12AX7, 12BH7A, 5651, 5836 (Reflex Klystron), 5965.

Accessories Supplied: Type 874-R22 Patch Cord, Type 874-C58 Cable Connector, Type CAP-22 Power Cord, and spare fuses.

Dimensions: Width 19, height 7½, depth 15½ inches (485 by 195 by 395 mm), over-all; panel, 19 by 7 inches (485 by 180 mm).

Net Weight: 38 pounds (17.5 kg).

Type		*Code Word*	*Price*
1360-AM	**Microwave Oscillator, Bench Mount**	BURLY	**$1100.00**
1360-AR	**Microwave Oscillator, Rack Mount**	BASSO	**1100.00**

U.S. Patent No. 2,548,457

MORE AND BETTER PULSES FROM THE UNIT PULSE GENERATOR

The Type 1217-A Unit Pulser[1] was, like its companion instruments in the unit line, designed for maximum utility, minimum complexity, and low cost. The thousands of these compact, high performance devices that are now in use have shown that the design was indeed a successful blend of these often conflicting factors. Time has made available new circuits and components, and experience has shown where improvements would be both desirable and practical. In the redesign the goals set were simple: to make every possible improvement compatible with the two conditions of no increase in price and no increase in power supply requirements.

[1]R. W. Frank, "Pulses in a Small Package — A Pulse Generator for the Unit Line," *General Radio Experimenter*, 28, 10, March, 1954.

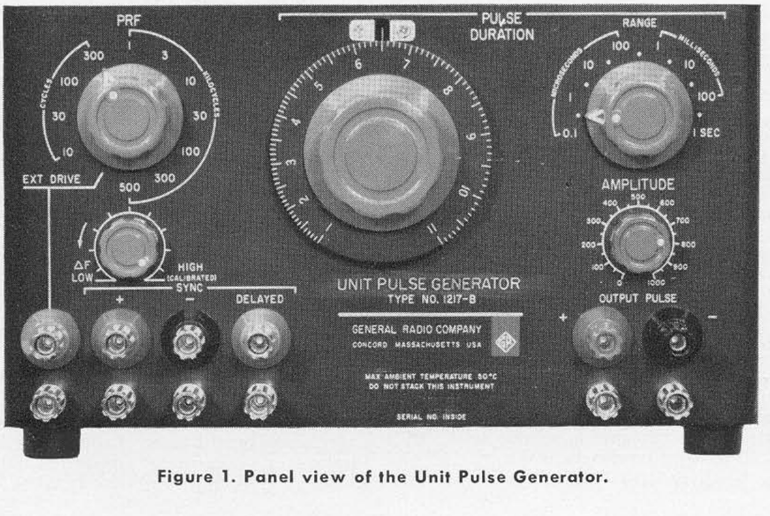


Figure 1. Panel view of the Unit Pulse Generator.

TABLE I

Comparison of Major Characteristics of the Types 1217-A and 1217-B

Characteristic	1217-A	1217-B
Pulse Rise Time	<50 nsec	<20 nsec (50 ohms)
Pulse Fall Time	<150 nsec	<10 nsec (50 ohms)
Pulse Duration	Continuous 150 nsec — 60 msec	Continuous 100 nsec — 1 sec
PRF (Internal)	Steps 30 cps — 100 kc	Continuous 2.5 cps to 500 kc
PRF (External)	Locked 30 cps — 100 kc	Continuous dc to 1 Mc
Pulse Amplitude (1-kilohm output impedance)	±20 v into 1 kilohm	±40 v into 1 kilohm
Input Sensitivity	30 v at 100 kc	0.3 v at 1 Mc
Accuracy PRF and Duration	±15%	±5%
Delayed Pulse	None	To trigger a second generator

Similarity between the new TYPE 1217-B Unit Pulse Generator and its popular predecessor goes little further than the four digits of its type number. Significant changes have been made in all performance specifications. The most important parameters are listed for comparison in Table I. It can be seen that, in every instance, the performance figures are increased by at least 2:1 and often by more than 10:1.

This performance is achieved in two ways:

(1) The TYPE 1217-B uses better devices; being neither wholly "transistorized" nor wholly "vacuum tube-ized" it takes full advantage of the best properties of both modern transistors and vacuum tubes.
(2) The TYPE 1217-B has completely unconventional circuitry for all functions—every component works full time. In fact, through a series arrangement of timing and output circuits, the 55-ma input current from the power supply is used to provide 40 ma of useful load current.

The new design has other new features, not clearly shown in Table I, which can be better appreciated after some of the new circuit characteristics are more completely explained. These will be discussed in the section on applications, below, after the circuits have been explained in some detail.

CIRCUITS

Block Diagram

Figure 2 is a block diagram of the circuit. In block form things look quite conventional. The input circuits consist of a Schmitt trigger circuit driven by an amplifier connected to the input terminals so that the pulse generator will be started by a triggering pulse once per cycle of any input waveform at any

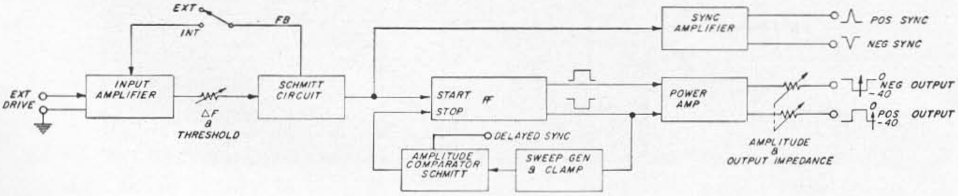


Figure 2. Block diagram of the circuit system.

frequency from dc to over one megacycle per second. Conventional—yes, but every active part of this input-triggering circuit is converted to a stable *RC*-controlled oscillator when internally produced pulse repetition frequency (prf) is desired. This oscillator will produce any desired recurrence frequency between 2.5 cps and 500 kc.

The trigger pulse from these input circuits: (1) operates a sync-pulse-producing stage to form both positive and negative pre-triggers, and (2) starts the pulse-generating and timing circuits.

A transistor-bistable circuit, set by the trigger from the input circuit, simultaneously operates the pulse output stage and the pulse-timing circuit. The output stages, producing both positive and negative pulses, are a pair of power pentodes acting as 40-ma current sources. The timing circuits are comprised of a switch tube, a high-speed clamp and a Schmitt trigger. When the transistor bistable switches, starting the pulse, the timing switch is turned off. A precision capacitor is charged to the point where the Schmitt trigger operates, producing a reset trigger for the bistable control circuit, thereby terminating the pulse.

The 40-ma current-source output pentodes are directly connected to the output terminals through a 1-kilohm amplitude control.[2] Forty-volt positive and negative pulses are thereby produced at full amplitude. Since the connection to the output terminals is direct, the dc component of the pulses is present, and ramp-off cannot occur, no matter how great the pulse duration.

Input Circuits

Figure 3 is a simplified schematic diagram of the input circuits and prf oscillator. The switching for the circuit is shown here in proper position for the aperiodic-input-circuit connection.

In this connection V_1 amplifies the input signal, and the voltage divider R_1 and variable resistor R_2 apply the amplified input signal to the Schmitt circuit, V_2. R_2 in this application permits an adjustment of the dc component of the input signal either to optimize the triggering sensitivity or to adjust the phase of the output pulse with respect to the input signal over a limited range.

When the PRF selector switch is thrown to any one of its other twelve positions

[2]This output circuit configuration is identical to that of the General Radio Type 1391-B Pulse, Sweep, and Time-Delay Generator.

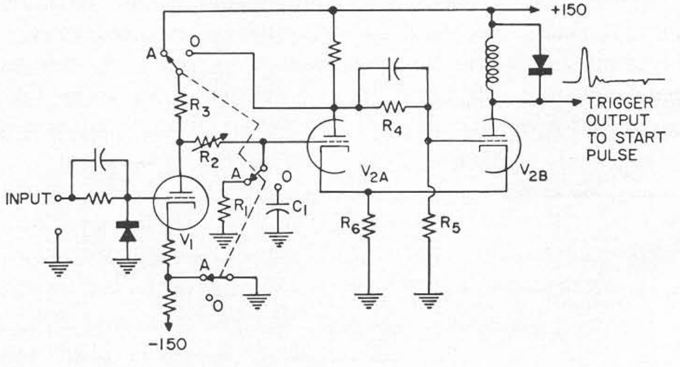


Figure 3. Elementary schematic of the input circuits.

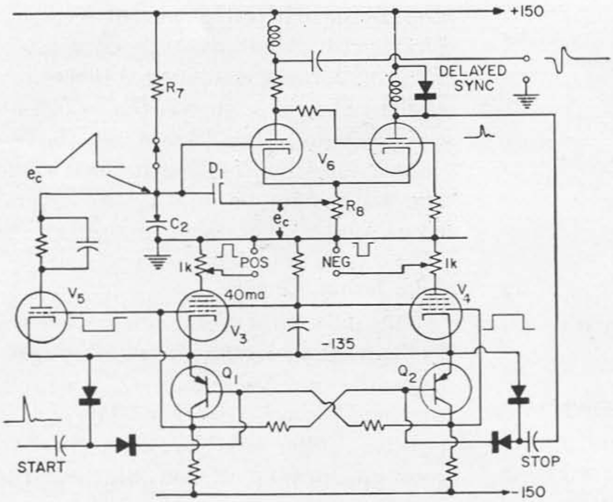


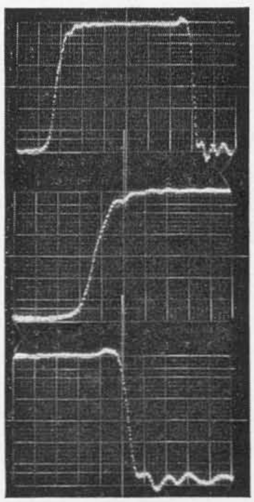
Figure 5. Elementary schematic of the pulse-timing and output circuits.

tion is set by the cathode current of the amplitude-comparator Schmitt in R_8. Any variations in this current will affect both the initial and final voltage values. Again, as in the input circuits, this comparator is stabilized by heavy current feedback and the triggering voltage is determined by precision resistors.

Output Circuits

Figure 5 also shows the output circuits. Before a start trigger pulse is received from the input circuits Q_1 is on and Q_2 off. V_3 is therefore conducting at (nearly) zero-bias and V_4 is off. When a trigger pulse is received Q_2 goes on bringing V_4 on. V_3 and V_4 are a pair of power pentodes which pass 40 ma when on at zero-bias. The interruption of plate current in V_3 produces a 40-volt positive pulse in its load resistor. Simultaneously V_4 turning on produces a 40-volt negative pulse across its load resistor. The extreme speed of Q_1 and Q_2 in the transistor flip-flop switches these plate currents on and off very rapidly. A typical positive current transition is of the order of 15 nanoseconds, while the negative transitions are typically 8 nsec. (See Figure 6.)

The very rapid current transitions are applied to the 1-kilohm output potentiometers and an internal stray-capacitance of approximately 30 pf. With no external loading the rise time of voltage is approximately 60 nsec. External capacitance will increase this rise time by

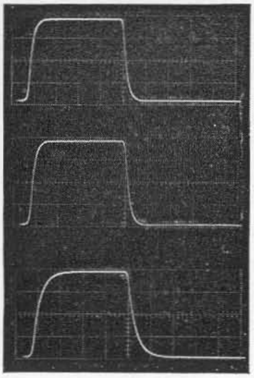


0.12-μ sec pulse in 1-nsec rise time sampling system; 20 nsec/cm.

Positive transition, 10 nsec/cm

Negative transition, 10 nsec/cm

Figure 6. Oscillogram showing typical output transitions in a 50-ohm system.

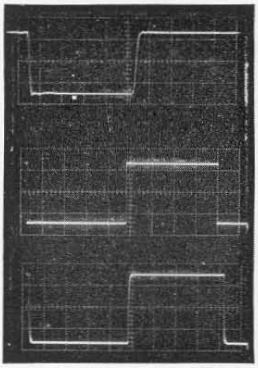


0.4 volt, peak-to-peak, into 10 ohms; 0.1 μsec/cm

4 volts, peak-to-peak, into 100 ohms, 0.1 μsec/cm

40 volts, peak-to-peak, into 1 kilohm; 0.1 μsec/cm

Figure 7. Open-circuit rise-and fall-time oscillograms; 'scope has 12-pf probe.



A
2-volt, 0.5-μsec pulse; 50-ohm termination; 0.1 μsec per division

B
2-volt, 5-μsec pulse; 50-ohm termination

C
As in (B), but with open-circuit termination, 40-volt pulse

Figure 8. 2-volt pulses into 50 ohms.

approximately 2 nsec/pf. With this output circuit form, no overshoot will ever be observed, and the rise and fall of output voltage is purely exponential (see Figure 7 where the output pulse is shown as presented on a Tektronix 543 oscilloscope with 12-pf probe).

When the ultimate current rise times are to be utilized it is necessary to terminate the pulse generator in an impedance appropriate to the coaxial cable (50 or 93 ohms) to be used. Fast 2-volt pulses in a 50-ohm system are shown in Figure 8.

APPLICATIONS

The extremely wide ranges of pulse duration and prf produced by this pulse generator fit it for almost any application in which a pulse is needed. There are so many applications that it is difficult to select a sample group to be included here. The new model has demonstrated itself to be far more useful than its predecessor because:

(1) Its duration control, being more accurately calibrated, can be used for quantitative measurement of maximum and minimum durations, for example, over which a flip-flop will function. The pulse duration can be established without the need to read an oscilloscope.

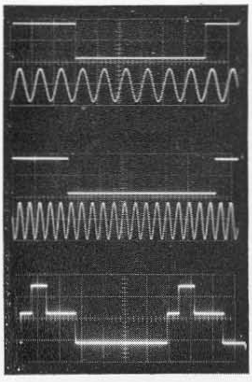
(2) Since the amplitude control varies output impedance, the instrument can be set to produce a correct driving-point impedance for any passive pulse network.

(3) Its linear current-source output system produces a clean pulse of easily adjustable and equal rise-fall time.

(4) Since the prf can be continuously varied it is possible, for example, to establish the resolution failure point of a flip-flop precisely.

(5) The aperiodic synchronizing circuit for external control of the prf makes it possible to drive the instrument from an *RC* or beat oscillator over the full range of that oscillator with no control adjustments on the pulse generator. Therefore, the prf accuracy and stability is that of the driving oscillator. It is also possible to produce pulses with a random frequency distribution.

(6) The stability of the internal prf oscillator makes it possible to use the Type 1217-B in systems as a precise frequency divider of high ratio (Figure 9, A and B).



A
9:1 division of 1-Mc input sine wave 1 μsec/cm; prf = 111.1 kc

B
20:1 division of 1-Mc input sine wave

C
Complex pulse from two generators in parallel

Figure 9.

(7) The presence of a threshold control for the external synchronizing circuit makes it easily possible to produce single pulses. A 1.5-volt cell and Micro Switch can also be used to produce single pulses from a hand-held trigger generator.

(8) The linear, dc coupled output permits paralleling to provide complex output pulses with no external adding networks, as shown in Figure 9C.

Beyond the general increases in applicability obtained through the design improvements listed above, experience has shown that the TYPE 1217-B is a useful source for measurements on transistor systems. It can operate saturated transistor switches, both *npn* and *pnp*, without coupling networks. Since the pulse generator is direct-coupled, the solid-state switches can be operated over its full duration-range. Figure 10 shows the connections for driving a *pnp* transistor switch. The low output impedance of the TYPE 1217-B is normally sufficient for hold-back during the pulse off-time. Figure 11 shows the direct connection for switching *npn* transistors.

— R. W. FRANK

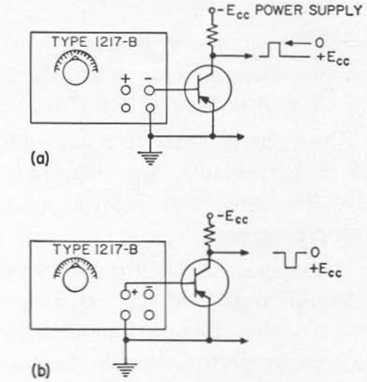


Figure 10. Control of *pnp* transistor switch.

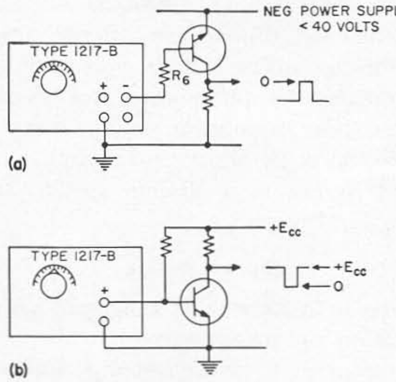


Figure 11. Control of *npn* transistor switch.

SPECIFICATIONS

PULSE REPETITION FREQUENCY

Internally Generated: 2.5 cps to 500 kc with calibrated points in a 1-3 sequence from 10 cps to 300 kc, and 500 kc, all ±5%. Continuous coverage of the range from 2.5 cps to 500 kc with an uncalibrated control lowering the frequency of the calibrated points.

Externally Controlled: Aperiodic, dc to 1 Mc with 1-v rms input (0.5 v at 500 kc and lower); input impedance, at 0.5 v rms, approximately 100 kilohms shunted by 50 pf.

OUTPUT PULSE CHARACTERISTICS

Duration: 100 nsec to 1 sec in seven decade ranges, ±5% of reading, or ±2% of full scale or ± 25 nsec, whichever is greater.

Rise Time:

a. Into terminated 50- or 100-ohm cables all transitions will have rise times less than 20 nanoseconds (typically 12 nsec).

b. On high-voltage output (40 v at 1 kilohm) rise time will be limited by load capacitance.

SPECIFICATIONS (*Cont.*)

Rise and fall times typically 60 nsec + 2 nsec/pf external load capacitance.

Voltage: Positive and negative 40-ma current pulses available simultaneously. DC coupled, with dc component negative with respect to ground. 40 volts peak into 1-kilohm internal load impedance for both negative and positive pulses. Output control marked in approximate output impedance.

Overshoot: Overshoots and noise in pulse, less than 5% of amplitude with correct termination.

Ramp-off: less than 1% everywhere.

Synchronizing Pulses:

Pre-pulse: Positive and negative 10-volt pulses of 150-nsec duration. If positive sync terminal is shorted, negative pulse can be increased to 50 v. Sync-pulse source impedance:

positive — approx 300 ohms
negative — approx 1 kilohm

Delayed Sync Pulse: The delayed sync pulse consists of a negative-going transition of approximately 5 volts and 100-nsec duration coincident with the late edge of the main pulse. The duration control reads the time between the pre-pulse and the delayed sync pulse. The delayed sync-pulse negative transition is immediately followed by a positive transition of approximately 5 volts amplitude and 150-nsec duration to reset the input circuits of a following pulse generator. (See oscillogram above.)

1-μsec pulse into 50 ohms with delayed sync pulse

STABILITY

PRF and pulse-duration jitter are dependent on power-supply ripple and regulation.

a. With Type 1201 Power Supply (recommended), input terminals short-circuited,

PRF Jitter	0.01%
Pulse-Duration Jitter	0.01%

b. With Type 1203 Power Supply

PRF Jitter	0.05%
Pulse-Duration Jitter	0.05%

POWER REQUIRED

300 v at 55 ma, 6.3 v at 3 amp. Type 1203-B Unit Power Supply or Type 1201-B Unit Regulated Power Supply is recommended.

DIMENSIONS

Width 9½, height 5¾, depth 6½ inches (240 by 150 by 165 mm), over-all.

NET WEIGHT

4½ pounds (2.1 kg).

Type		*Code Word*	*Price*
1217-B	**Unit Pulse Generator**	AMASS	**$250.00**

AUTOMATIC MEASUREMENT OF PHONOGRAPH REPRODUCERS

By B. B. BAUER, *Vice President*
CBS Laboratories, Stamford, Connecticut

Among the latest of manual procedures to yield to automation is the measurement of phonograph reproducer characteristics. This is made possible by development of the new CBS Laboratories Type STR 100 Stereophonic Frequency Test Record, which is adapted for use with General Radio Type 1521-A Graphic Level Recorder.

A stereophonic record contains two related program channels which are identified with orthogonal modulations of the walls of a single groove. The left channel corresponds to the inner groove wall, the one closest to the center, and the right channel to the outer groove wall (away from the center). The positive directions of these modulations are at

+45° and −45° to the record surface, respectively (Figure 1). The pickup has a single stylus acting on a pair of transducers, arranged in an orthogonal fashion and intended to generate independent voltages when driven by the respective groove-wall modulations (Figure 2). The voltages e_l and e_r generated in the transducers drive two amplifiers and loudspeakers to reproduce the recorded information.

The only feasible way of testing the performance of a phonograph pickup is by use of a frequency-test record, on which tones have been recorded at various frequencies. Stereophonic frequency-test records have separate recordings for the left and the right channels. The response vs frequency of any given pickup channel produced by the corresponding record channel is known as "response-frequency characteristic" or simply "response" of the channel on the particular record. The response from the opposite channel is known as "crosstalk-frequency characteristic" or simply "crosstalk." The customary units for both characteristics are db re 1 volt rms. The difference at any one frequency (or average over a group of frequencies) between response and crosstalk is known as "channel separation," expressed in db.

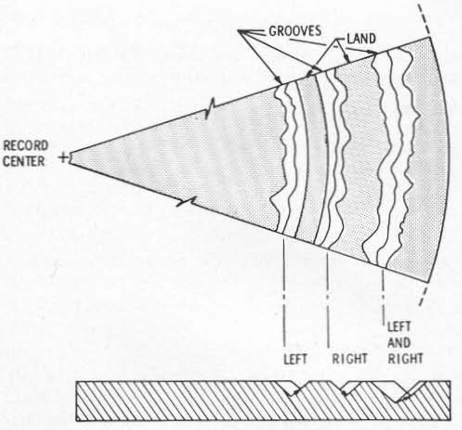


Figure 1. "Pie" representation of a stereophonic record portraying modulation associated with left channel, right channel, and combined left and right channels.

The STR 100 record has a frequency sweep band for each channel, and its frequency varies logarithmically with time, at a rate of 1 decade each 24 seconds. This corresponds to a chart speed on the General Radio Type 1521-A Graphic Level Recorder of thirty ¼-inches per minute. The sweep band starts with a 1000-cycle tone of sufficient duration to permit the recorder pen to be set to the 40 cps ordinate and the recording level to be adjusted to a convenient value. Upon cessation of the 1000-cycle tone the frequency drops immediately to 40 cps and

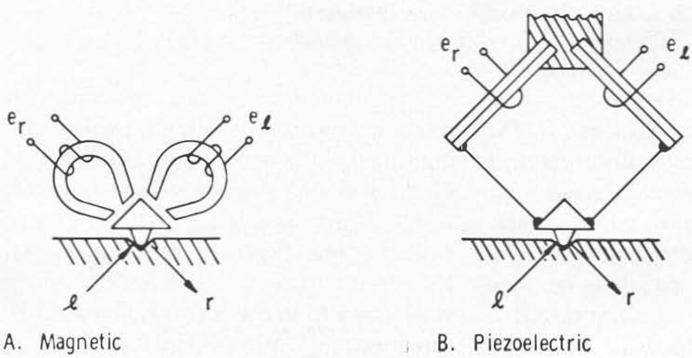


Figure 2. Typical arrangements of stereophonic pickups.

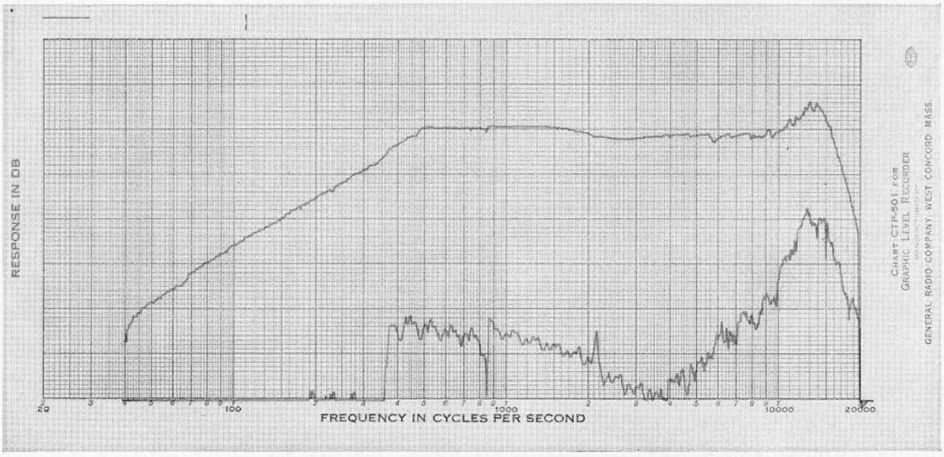


Figure 3. Automatically plotted response of typical high-grade magnetic pickup. Upper curve is direct response of one channel; lower curve is crosstalk from other channel.

then rises continuously to 20,000 cps in 24 log (20,000/40) = 64.8 seconds. Next, an interval is allowed for the operator to reset the recorder chart back to the 40 cps ordinate. Then, the right-channel 1-kc tone is heard, being followed by the right-channel sweep. If one channel of a pickup under test is connected to the recorder (through a suitable amplifier), then the response-frequency and the crosstalk-frequency characteristics for the particular channel will be successively recorded. A typical set of curves for a magnetic phonograph pickup is shown in Figure 3. Two such sets, one for each channel, are required to describe the performance of the pickup.

Previously available test records used spot-frequency tones not adapted for automatic recording. The STR 100 record also has such fixed-frequency tones for the left and right channels, but each tone is preceded by a voice announcement of frequency, so that there is no doubt as to which tone is being played. The results of the spot-frequency tests are similar to those obtained with the sweep-frequency bands, but the process is far more laborious, and the information between the spot-frequency tones is not revealed. A system that appears flat when measured with spot-frequencies often has resonant peaks or dips in between the spot frequencies.

The spot-frequency bands in the STR 100 record above 500 cps were cut with the same recorder setup as were the sweep-tone bands. This permitted an absolute calibration of the sweep-tone bands by measurement of the spot-frequency bands with microscope and by diffraction-of-light patterns.[1]

Test Record Characteristics

As indicated above, the reproducing characteristics of a pickup or system under test are referred to the characteristics of a particular test record. These may be defined in terms of displacement or velocity of groove modulation.

For constant-displacement recording, a displacement-responsive (piezo-elec-

[1]58-IRE-19-S1, *Standards on Recording and Reproducing: Methods of Calibration of Mechanically Recorded Lateral Frequency Records*, Institute of Radio Engineers, 1 East 79th Street, N.Y. 21, N.Y.

tric) pickup will produce a constant output at all frequencies, while a velocity-responsive pickup (magnetic, moving coil) will produce an output directly proportional to frequency. On the other hand, a constant-velocity recording will produce a constant output with a velocity-responsive pickup, and an output inversely proportional to frequency with a displacement-responsive pickup.

The STR 100 record has a constant-displacement modulation up to 500 cps and constant-velocity modulation above 500 cps. This explains the response of a magnetic pickup in Figure 3, which is a rising straight-line in db vs log frequency at 6 db per octave below 500 cps and constant above 500 cps. The deviation of response from two straight lines denotes the departure of the pickup from ideal performance. The crosstalk curve, which is 20 to 30 db below the principal channel output, is typical of the channel separation that may be expected in present-day high-grade pickups.

Similar measurements can readily be performed on displacement-responsive pickups if the generated voltage is differentiated by connection across the pickup terminals of a resistance which is small compared to the capacitive reactance of the pickup. Usually a 10,000-ohm resistor will suffice. The response-frequency characteristic of an ideal piezo-electric pickup terminated in this manner is similar to that of a magnetic pickup.

Testing Pickup Preamplifiers

Modern 33⅓ rpm records are recorded with a frequency characteristic that is a composite of several constant-displacement and constant-velocity segments as follows:

Up to 50 cps — constant velocity.

From 50 to 500 cps — constant displacement.

From 500 to 2120 cps — constant velocity.

Above 2120 cps — constant displacement.

The transitions between these segments are not sharply defined, but instead they are blended together, in a manner defined by the RIAA.[2] The velocity (db)-vs-frequency character-

[2] *Standard Recording and Reproducing Characteristic*, Record Industries Association of America Inc., 1 East 57th Street, N.Y. 22, N.Y.

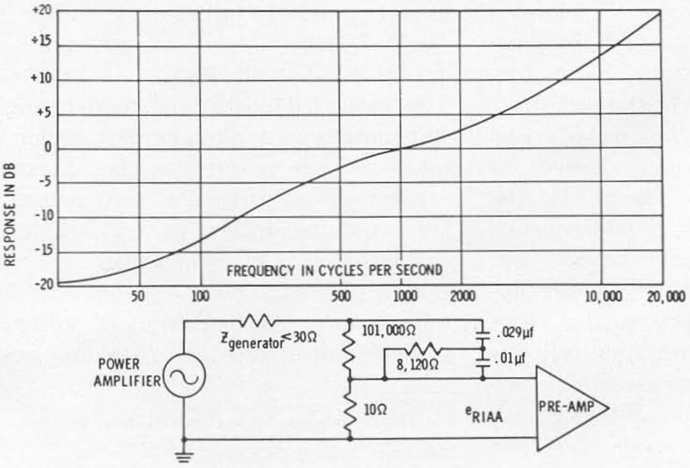


Figure 4. RIAA recording characteristic and R-C RIAA network.

istic which is obtained is shown in Figure 4.

When the pickup is a true velocity-responsive device, it follows that the pickup preamplifier should have an inverse RIAA characteristic. This characteristic of the preamplifier is most conveniently verified by the insertion of an RIAA generator network between the oscillator and the amplifier under test. One such network is shown in Figure 4. If the measured response is uniform with frequency, then the preamplifier is properly designed for reproducing records with a magnetic pickup. A Type 1521-A Recorder and Type 1304-B Beat-Frequency Audio Generator can be advantageously used in this test.

Over-all Measurements of Response Characteristics

The STR 100 record and the Type 1521-A Recorder can also be used to measure the over-all response-frequency characteristics of a playback system regardless of the type of pickup or amplifier employed. The ideal response-frequency characteristic of a properly adjusted reproduction system playing an STR 100 record will be simply the difference between the response-frequency characteristic of the record and the RIAA characteristic shown in Figure 4. Subtracting these two results in the set of values in the accompanying table.

Ideal System Response — RIAA Equalized

Frequency	db	Frequency	db
1,000	0	800	+0.7
20,000	—19.5	600	+1.8
18,000	—18.8	500	+2.6
16,000	—17.7	400	+1.9
14,000	—16.6	300	+1.1
12,000	—15.3	200	+0.2
10,000	—13.7	150	—0.6
8,000	—11.9	100	—0.9
6,000	—9.6	80	—1.3
5,000	—8.2	60	—2.3
4,000	—6.6	50	—3.0
3,000	—4.8	40	—4.2
2,000	—2.6	30	—5.8
1,500	—1.5	25	—7.0
1,000	0	20	—8.6

Tone-Arm-Resonance Test

To test resonance of tone arms, loudspeakers, etc., the STR 100 record provides sweep-tone bands, left and right, from 200 to 10 cps. These are synchronized also to the Type 1521-A Recorder. The recorder must be set to operate in reverse, as the sweep tone begins at 200 cps and the frequency decreases during the glide.

Automatic-Start Circuit

The 1000-cycle tones at the beginning of each glide serve not only for level adjustment, but also for keying an automatic-start circuit for the recorder. This circuit, developed by Messrs. A. Schwartz and A. Gust of CBS Laboratories, is shown in Figure 5.

The 1000-cycle keying tone preceding the sweep initiates the cycle. All relays are initially de-energized as shown in the schematic diagram. Left and right channel inputs are combined in the cathode of *V1* insuring that the keying tone will be present for either direct or crosstalk measurements. The cathode follower output is fed to the *V2* high gain amplifier through a high-*Q* *LC* 1000-cycle filter allowing only 1000 cps to feed through. Following this stage is a cathode follower *V3* employed as a power amplifier to drive a sensitive relay *K1* (Elgin Advance) after rectification by the two 1N2482 diodes. A Zener diode and clipping-range control prevent high signal levels from overheating the sensitive relay.

With *K1* energized, relay *K2* is energized and locks itself across the power

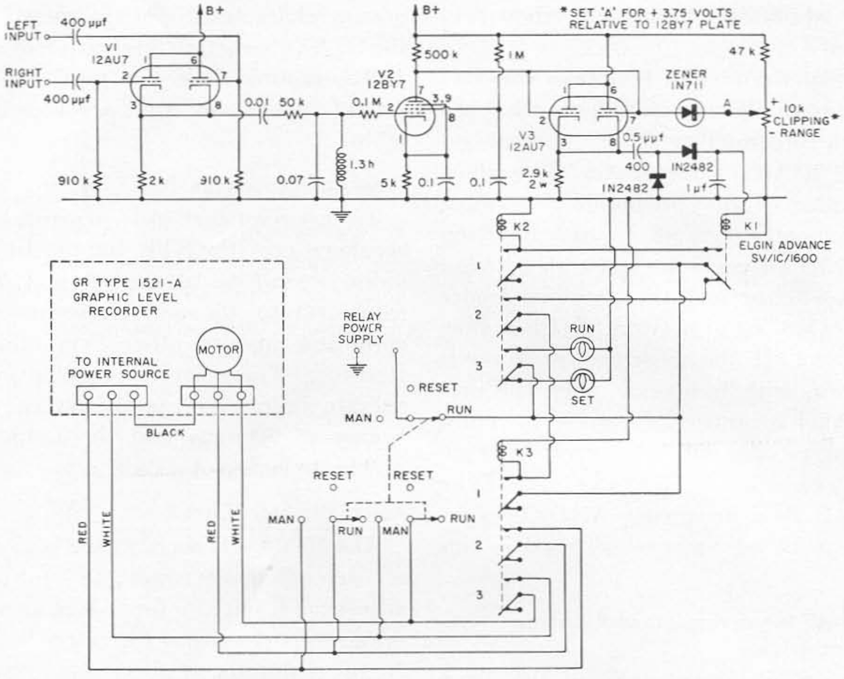


Figure 5. Automatic control circuit for the recorder.

supply through contact 1. At the cessation of the keying tone, *K1* is de-energized — thereby actuating *K3*, which starts the recorder motor at the instant the sweep begins. At the termination of the sweep band, the reset switch is manually set at RESET momentarily to de-energize *K2*, and the circuit is then ready for the next sweep.

Economics of Automation

The economic value of automation in testing phonograph reproducers deserves particular mention. The development of a phonograph pickup or complete player involves considerable experimental work in which the device under test is modified by successive rearrangement or modification of components until the desired performance is obtained. Each modification is followed by response and separation tests. Spot-frequency tests require the better part of one hour. It has been estimated that one-third to one-half the manpower used in pickup and phonograph development is expended in this tedious endeavor. Delegating these measurements to an automatic device which does not mind tedium and makes no error releases trained manpower for more creative tasks and saves thousands of dollars annually. In one instance, changing from manual to automatic recording has been shown to save the cost of a recorder in a single month, and the cost of the new record in a single test.

Use by Audiophiles

While the STR 100 record has been designed especially for the professional user, the requirements of the audiophiles have also been kept in mind. Complete instructions are included with each record for testing frequency range of the reproducing system, channel separation, pickup compliance and tracking, tone-arm resonance and stylus wear.

The STR 100 record is priced at $8.50 and can be obtained from CBS Laboratories, Stamford, Connecticut, or from Columbia Records distributors and dealers.

MORE TALENTS, NEW DRESS, FOR THE OUTPUT POWER METER

In the nearly thirty years since its first announcement, the General Radio Type 583-A Output Power Meter has served as a work-horse of the audio-frequency industry. During this time, significant advances in material and techniques have made possible its replacement by a new instrument, based on the same general theory, but improved in all respects.

Both instruments are basically multi-tapped audio-frequency transformers that, by transformation ratio, reflect an essentially fixed secondary load as a variable primary impedance. They differ, however, in several respects, which make possible the new instrument's greater frequency, impedance, and power ranges and its improved accuracy on complex waveforms.

The Type 583-A used a mu-metal core to secure high initial permeability (necessary for impedance accuracy at low power input), but was limited to a 5-watt maximum input by the low saturation level of mu-metal. The Type 1840-A secures high initial permeability through the use of grain-oriented silicon steel in a lamination specifically designed to take advantage of grain-orientation, and thereby increases its maximum input to 20 watts with but little increase in core size, since grain-oriented silicon-steel is a true "power" material.

The Type 583-A had ten secondary taps to yield ten impedance values

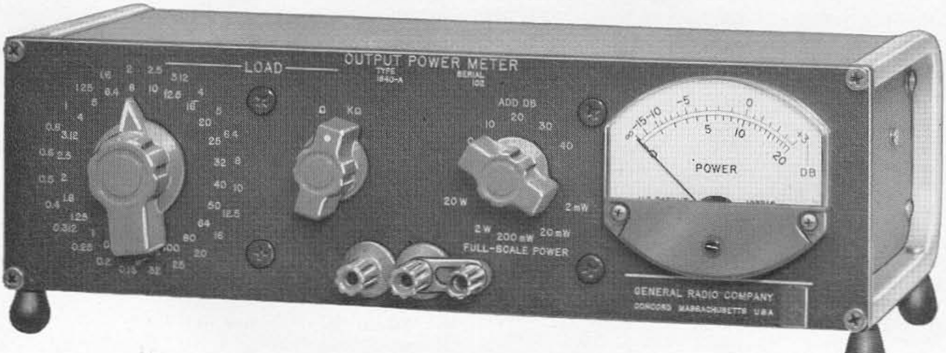


Figure 1. Panel view of the Output Power Meter.

spaced, approximately, at $\sqrt[10]{10}$ (tenth root of 10) intervals and four-primary taps to provide multiplication by powers of 10 (0.1, 1, 10, 100). This yielded forty impedance values between 2.5 and 20,000 ohms. The TYPE 1840-A has six secondary taps to yield six impedance values, spaced, approximately, at $\sqrt[6]{4}$ (sixth root of 4) intervals. Note that the two intervals are comparable

$$\sqrt[10]{10} \cong \sqrt[6]{4} \cong 1.26.$$

The TYPE 1840-A, however, has eight identical primaries, each tapped to provide a multiplier of 250/1, which primaries are switched from all parallel to all series in four configurations, each configuration introducing a multiplier of 4/1. The six secondary taps and four primary connections are all controlled by a single twenty-four position rotary switch yielding six times four, or twenty-four, discrete impedance values. The taps on all eight primaries are switched, simultaneously, by the ohms-kilohms switch. The ohms range lies between 0.6 and 128 ohms. The kilohms range multiplies these values by 250, to yield 0.15 to 32 kilohms. The net result of all these shenannigans is more efficient use of the transformer windings. While, in the older instrument, up to forty percent of the input power was dissipated in the windings, in the new model it is reduced to less than eight percent. This fivefold decrease in winding dissipation so reduces the winding's contribution to input impedance that all TYPE 1840-A's use the same accurate resistors in contrast to the hand-tailored resistors required for each TYPE 583-A.

A further advantage of primary switching is apparent in the improved frequency response. Since all primaries are always active, and are interleaved with the secondary windings in two pi's, an octave improvement in both high and low frequency response has been achieved.

The quasi-rms meter in the TYPE 1840-A tolerates second and third harmonics up to 20% in the signal without departure from a true rms indication. The new General Radio rack-bench-instrument cabinet provides convenience, access and an adjustable tilt for easy reading.

A T-network attenuator, described in the instruction book, permits extension of the power level to 200 watts for any particular impedance setting.

— GILBERT SMILEY

SPECIFICATIONS

Power Range: 0.1 milliwatt to 20 watts. Auxiliary db scale reads from —15 to +43 db re 1 milliwatt.

Power Accuracy: Maximum error in full-scale power indication does not exceed 0.5 db from 50 to 10,000 cps; does not exceed 1.5 db from 20 to 20,000 cps.

Impedance Range: 0.6 ohm to 32 kilohms in two ranges; yielding 48 individual impedances spaced $\sqrt[6]{4}$ apart.

Impedance Accuracy: Maximum error does not exceed ± 5% from 100 to 10,000 cps or ± 50% from 20 to 30,000 cps.

Waveform Error: A quasi-rms meter is used which will indicate true rms with as much as 20% second and third harmonics.

Cabinet: Rack-bench instrument cabinet, aluminum panel. Cabinet has extension legs to permit instrument to be used in a tilted position. Panel extensions are available for relay-rack mounting.

Dimensions: Panel, width 12, height 3½ inches (305 by 89 mm); depth behind panel, 6½ inches (170 mm).

Net Weight: 10¾ pounds (4.9 kg).

Type		*Code Word*	*Price*
1840-A	**Output Power Meter**	BELOW	**$210.00**

THE MEASUREMENT OF THE BALLISTICS OF INDICATING INSTRUMENTS

The ballistic characteristics of moving-coil electrical indicating instruments are important in many applications, for example, in VU meters, sound-level meters, and modulation monitors. They can be uniquely defined in terms of (1) dc resistance, (2) time to first rise to a specified point, when energized from a high-impedance source, (3) overshoot, and (4) time to fall from one point to another with the instrument terminals short-circuited.

Stroboscopic light offers a means of determining (2), (3), and (4), while (1) is easily measured by a conventional bridge. One method of measurement previously described[1] used a continuous-film camera in conjunction with a stroboscope.

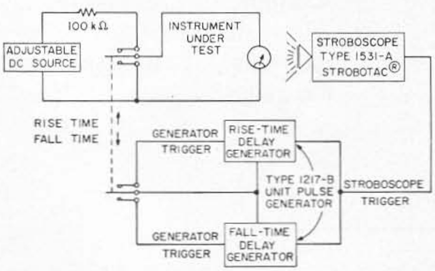


Figure 1. Block diagram of the meter test system.

A new technique of measurement, permitting direct observation,[2] has been made possible by the short, high-intensity flash of the General Radio Type 1531-A Strobotac® Electronic Stroboscope. Figure 1 is a block diagram of the test setup. The necessary time delays are supplied by two Type 1217-B Unit Pulse Generators.

The electrical indicating instrument is energized simultaneously with the triggering of one pulse generator. The pulse generator fires the Strobotac at a time later than the initial event determined by the setting of the pulse-duration dial. The pointer is seen by a single, microsecond flash of light from the stroboscope after the pointer has moved for the length of time that the stroboscope firing impulse was delayed. A series of such observations at different delay times will provide data for plots such as that in Figure 2. The details of the initial acceleration, the rate of rise and the overshoot with its decaying oscillation are readily apparent. The ballistic parameters of the instrument mechanism can be derived by analysis of curves plotted by this method.

Production quality control is an important use of this delay-generator-stroboscope technique. For our own instrument specifications we have defined rise time as "the time, in seconds, for the pointer first to reach 0.9, ± a specified tolerance, of the end scale when constant

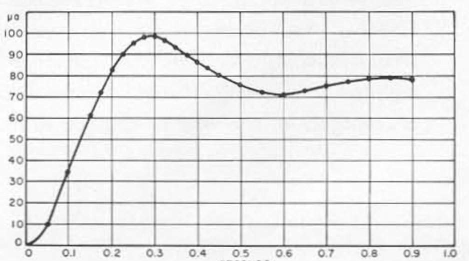


Figure 2. Plot of meter deflection vs time from data taken with the test system of Figure 1.

[1] "A Note on the Measurement of Meter Speeds," *General Radio Experimenter*, 10, 6, November, 1935.

[2] R. G. Fulks and H. C. Littlejohn, "A Direct Observation of Instrument Ballistics," AIEE Paper No. CP62-370, presented at the AIEE Winter General Meeting, N. Y., N. Y., January 28 – February 2, 1962.

electric power is suddenly applied from a high impedance source"; likewise, fall time as "the time, in seconds, for the pointer to reach 0.1, ± a specified tolerance, of end scale from a steady end-scale deflection when the instrument is short circuited."[3]

For production testing two delay generators are used, one energized when the electrical instrument is energized and the other when the electrical instrument is short circuited. One generator is set to the specified rise time, the other to the specified fall time. An operator then simply places the instrument to be tested in a fixture, which makes the connections and indicates the spread allowed in rise-time deflection and fall-time deflection. By the flipping of a switch once for rise time and again for fall time the inspection is completed.

Our system of direct observation of instrument ballistics will serve the research engineer and the quality-control man alike.

— H. C. Littlejohn

[3]Definitions to be incorporated in proposed revision of ASA C39-1 *American Standard Requirements for Electrical Measuring Instruments.*

CREDITS

This method of test was developed by R. G. Fulks, A. E. Sanderson, and H. C. Littlejohn.

— Editor

QUANTITY DISCOUNTS FOR COAXIAL CONNECTORS AND ADAPTORS

Prices of Type 874 Adaptors listed in the table on page 10 of the October, 1961, *Experimenter* are subject to domestic quantity discounts as listed below.

This schedule also applies to quantities of 10 to 99 of the connectors listed in the table on page 9 of the same issue.

Quantity	*Discount*
10 — 19	5%
20 — 99	10%
100 and over	15%

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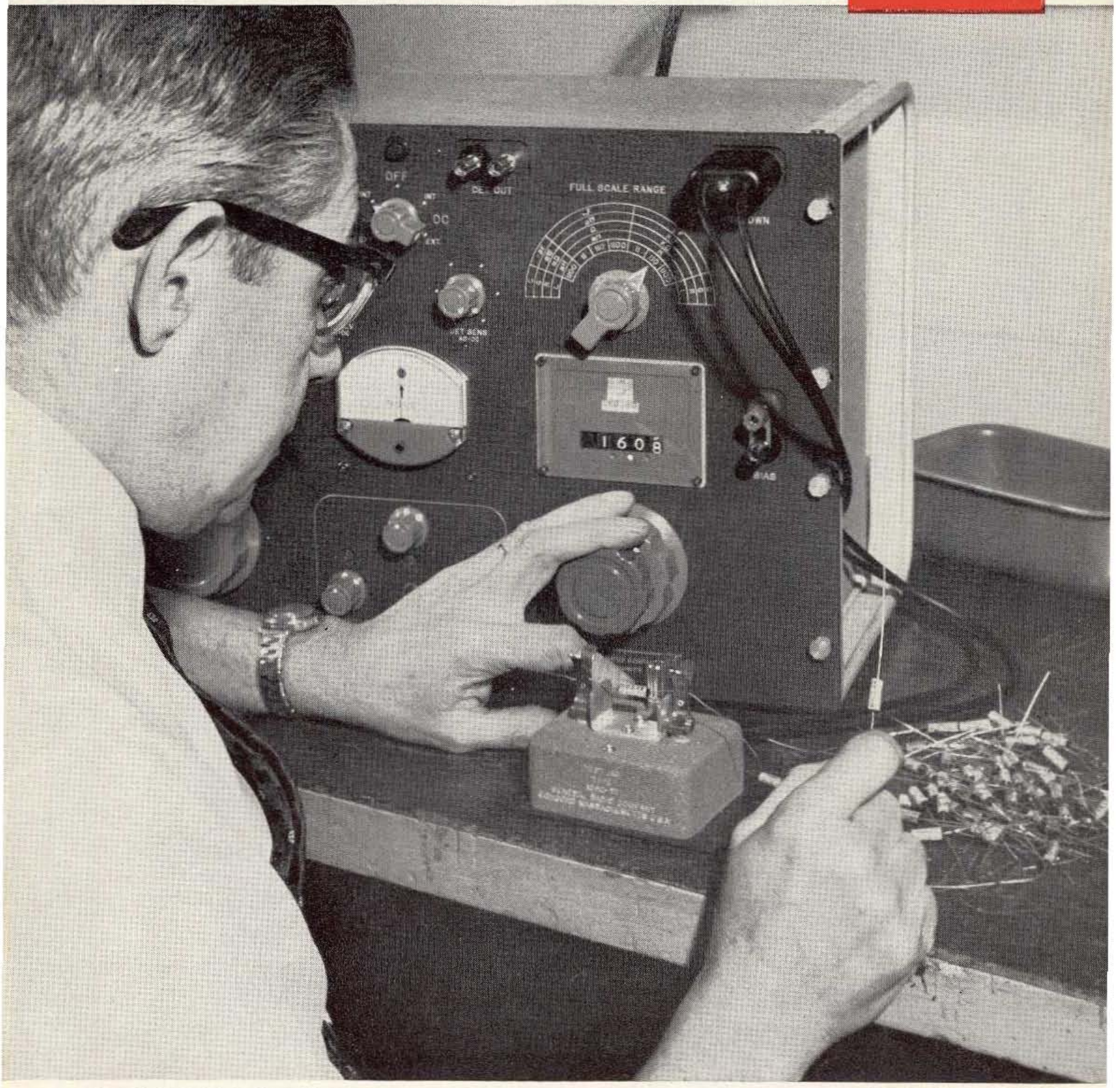

VOLUME 36 No. 3

MARCH, 1962

IN THIS ISSUE

New Precision Impedance Bridge

THE GENERAL RADIO

EXPERIMENTER

Published Monthly by the General Radio Company

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COVER

For the rapid measurement of components to 0.1%, the Type 1650-P1 Test Jig can be used with the new Type 1608-A Impedance Bridge, described in this issue.

A PRECISE, GENERAL-PURPOSE IMPEDANCE BRIDGE

A few years ago, when we were considering the redesign of the old, sloping-panel TYPE 650 Impedance Bridge,[1] we had to decide whether to make a 1% bridge, like the 650, or a bridge of higher accuracy for modern precision components. Generally, a redesigned instrument should do all that its predecessor did, only better, by taking advantage of improved components and techniques. We knew that our precision components could be used with confidence in an 0.1% bridge. A 1% bridge, however, has the advantage that the main component (*C*, *R* or *L*) can be presented on a single logarithmic dial, thus providing a simple balance adjustment for rapid measurements. This was so important that we decided to make two bridges. The TYPE 1650-A Impedance Bridge[2] introduced three years ago has a number of important improvements, but it retained the single CRL dial and 1% basic accuracy. It was an immediate best seller. For those who need greater accuracy, we now introduce the precision impedance bridge, the 0.1% TYPE 1608-A.

[1]Robert F. Field, "The Convenient Measurement of *C*, *R*, and *L*," *General Radio Experimenter*, 7, 11 & 12, April-May, 1933.

[2]Henry P. Hall, "A New Universal Impedance Bridge," *General Radio Experimenter*, 33, 3, March, 1959.

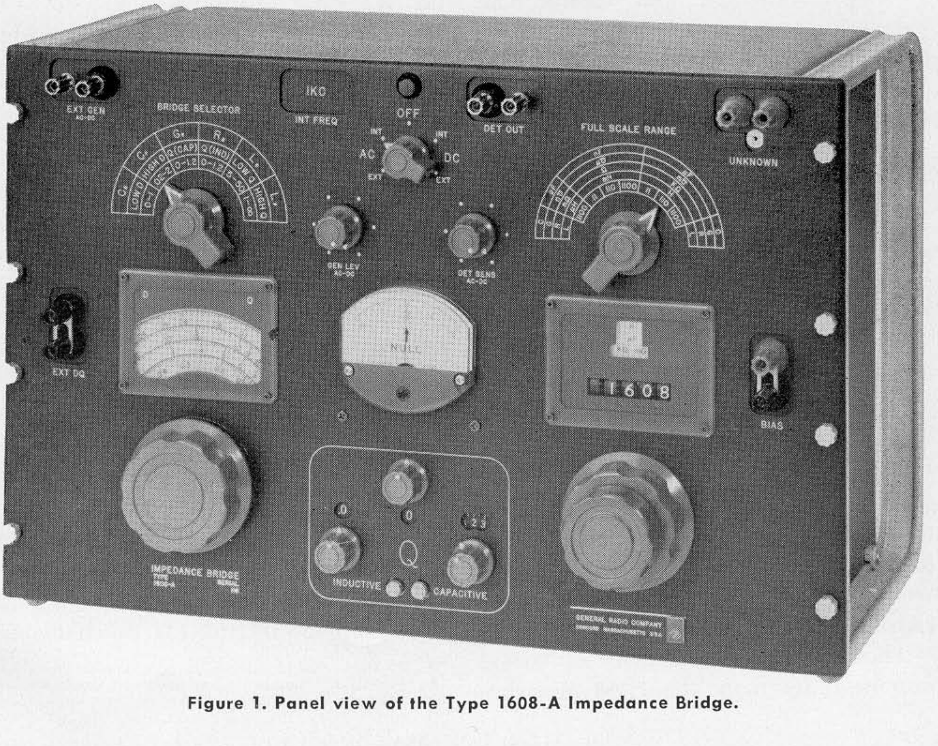


Figure 1. Panel view of the Type 1608-A Impedance Bridge.

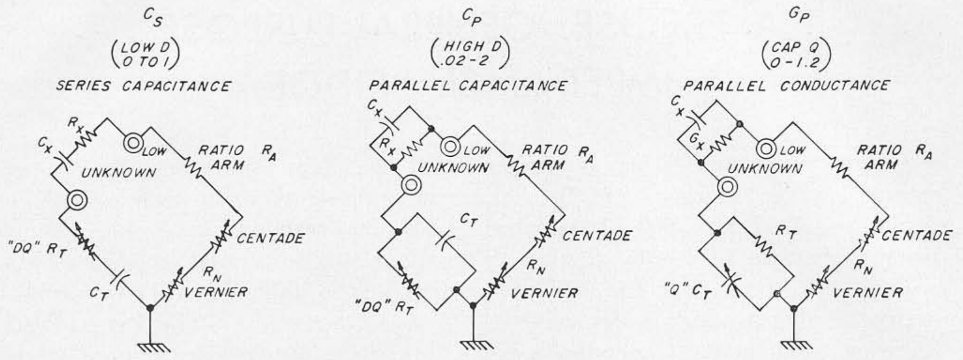


Figure 2a. Elementary schematics of the capacitance and conductance bridges.

This new, self-contained bridge system includes six bridge circuits for complete phase coverage of the passive half of the impedance plane, a 1-kc oscillator and selective detector and three dc power supplies for dc resistance and conductance measurements. The main design objective for this instrument, besides high accuracy, was to get a simply adjusted and easily read balance and readout system. The unique main balance system used is almost as easy to balance as a simple 1% dial and certainly much easier to read.

THE BRIDGE CIRCUITS

The bridge circuits in the TYPE 1608-A, Figure 2, are the familiar series and parallel inductance and capacitance bridges used in the TYPE 1650 and similar instruments. Also included are ac series-resistance and parallel-conductance bridges, both of which have phase (Q) adjustments not found in other bridges of this type. These circuits make possible a precise ac balance on an inductive or capacitive resistor and give a measure of its reactance, which is valuable for predicting the frequency characteristic.

The inclusion of these two bridges is important for more than just measurements on resistors, since they fill out the passive half of the complex plane, as shown in Figure 3. They make possible the measurement of a very lossy inductor or capacitor without a serious "sliding null," since the component can be measured as a resistor and the inductance or capacitance calculated from the measured Q and R (or G). Some bridges have wide D or Q ranges on the appropriate L or C bridge to measure lossy components, but when the Q is below ½ (D above 2), the resulting tedious sliding null makes them useless for practical measurement. The TYPE 1650-A Impedance Bridge uses the patented ORTHONULL®[3] mechanism, which greatly extends the useful range. This device requires a logarithmic CRL adjustment, which would be impractical with the linear digital readout of the new bridge. Although the use of the R and G bridges does require a calculation to get L or C

$$\left(L = \frac{RQ}{\omega}, C = \frac{Q}{\omega R}\right),$$

the high accuracy of the Q reading ($\pm 2\%$ ± 0.0005) results in better L or C accuracy at very low Q's than does even the ORTHONULL mechanism.

[3] H. P. Hall, "Orthonull — A Mechanical Device to Improve Bridge Balance Convergence," *General Radio Experimenter*, 33, 4, April, 1959.

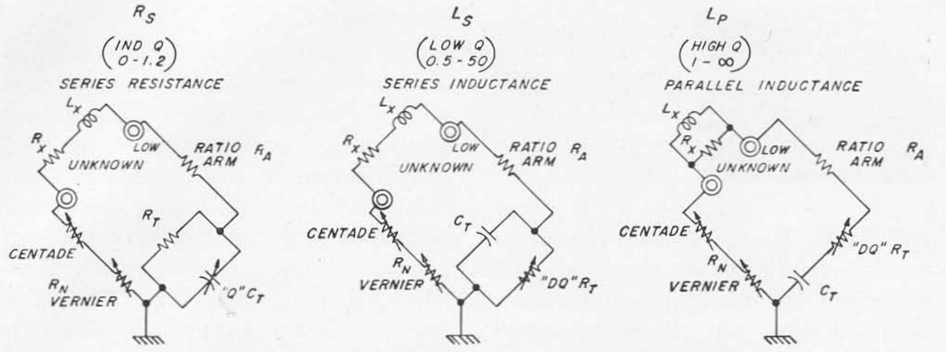


Figure 2b. Elementary schematics of the resistance and inductance bridges.

The bridge ranges extend up to 1100 μf, 1100h, 1.1MΩ and 1.1℧ and down to 0.05pf, 0.05μh, 0.05mΩ and 0.05n℧ (20kMΩ) which is the maximum resolution, corresponding to one-half of the last digit. The *L* and *C* ranges are more than adequate to cover any practical audio-frequency component, and the combination of an *R* and a *G* bridge gives resistance coverage from 0.05mΩ to 20kMΩ with the range from 1Ω to 1MΩ covered on both bridges. These two bridges can be used for both ac and dc measurements.

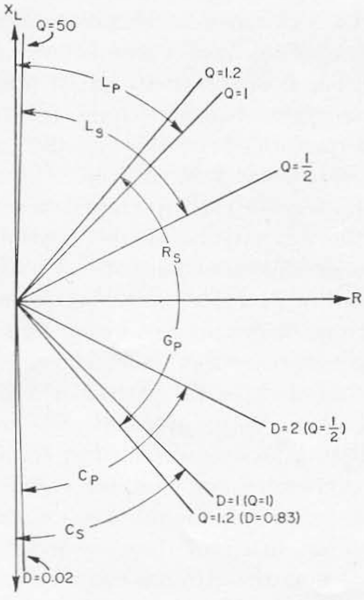


Figure 3. Phase coverage of the Type 1608-A Impedance Bridge.

CONTROLS

The Main Readout (C, G, *R*, or *L*)

An important design objective was a readout system that non-technical personnel could read without error. Even the most experienced engineer occasionally makes a mistake in interpolating a scale or applying a multiplying factor, and in an important project one occasional mistake can be costly. For repetitive production testing, the operator may be skilled at reading scales, but few can keep this up all day without a mistake. We felt, therefore, that our customers would value a simple digital display, very similar to an automobile odometer, but with decimal point and unit indicated, as shown in detail in Figure 4. There are two obvious ways to drive such an indicator, which, of course, must accurately track the precision-rheostat balance adjustment. One is to have a separate adjustment for each digit, as is done on a decade box. This requires four knobs, with the resulting annoyance of adjusting all four to vary the setting between 9999

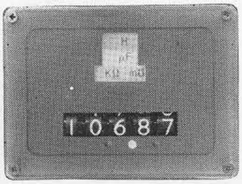


Figure 4. The digital CRL readout includes automatic decimal point location and illuminated indication of the units of measurement.

and 10000. The other method uses one input drive, but, to get from one end to the other, the operator would have to grind through all 10,000 positions, which is tedious and time consuming.

We compromised on two controls, a coarse and a fine, each adjusting two digits (the first actually goes to 114). Each of these controls can be swept through its range in less than one revolution. Therefore, during a balancing procedure when the operator comes to the end of the fine adjustment range, the coarse control is moved one digit, and less than 1 revolution of the fine control will reset it to zero.

To facilitate further this problem of transition between the two adjustments, the vernier scale extends beyond 99 to 106. This overlap of the adjustments is particularly useful for precision components, which are usually in integral values, so that the final adjustment is varied about a reading such as 9999 and 10000. However, the second dial (from the right) could not be labeled 10 since this would give the sequence 9999, 99100, which would be misleading. Also, a little thought will show that a geneva-type transfer to "carry the 1" would cause confusion since the operator would not know where he was on the fine adjustment. Therefore, we have adapted the convention of using X to represent 10 in one digit. The sequence now becomes 9999, 99X0, 99X1, etc., which seems unusual at first, but is easy to master.

The "Centade"

The coarse control of the main readout varies from 0 to 114, and each digit must represent a precise detented step of resistance in the bridge circuit. We have referred to this adjustment as a "centade" since it is similar to a decade resistance box, but has approximately 100 fixed steps controlled by one knob. The obvious (and expensive) way to do this is to put 114 precision resistors in series on a detented switch, which should be a shorting type to avoid discontinuities. The cheap way is to use seven resistors in a binary sequence and code a multiple contact switch to give a decimal scale. This method results in large adjustment discontinuities, which would give large momentary bridge unbalances at those steps where some resistors switched out and others switched in. The worst step is between 63 and 64 where the series combination of $1 + 2 + 4 + 8 + 16 + 32$ is switched out and 64 is switched in. This coded binary scheme would have a momentary value of either 0 or 127 between 63 and 64 unless these two switching functions were performed exactly at the same moment, which is impossible.

The scheme used in the centade has all the advantages of the continuous series arrangement but uses 40 resistors instead of 114. The switching sequence is shown diagrammatically in Figure 5. Here each resistor in the series chain has a value of three times that of a single step, R, in the adjustment. The intermediate values are obtained by shunting each series resistor with one or both of the two resistors mounted on the rotor. As shown, at step 1 the resistance value is $3R$ plus the parallel combination of $3R$, $6R$ and $2R$, which comes out to be a total of $4R$. In step 2 the $2R$ resistor is removed, giving a total of $5R$, and in step 3 both shunting resistors are removed, giving $6R$. Note that when the shunting rotor resistors are moved into position for shunting the next series resistor they are not in circuit, and so there is no coincidence problem. This method could actually be extended to reduce further the number of resistors, but the number saved for each additional rotor contact becomes smaller, and the design becomes more complex.

The switch stator is on an etched board with a rhodium-plated contact pattern similar to that shown diagrammatically in Figure 5, and the precision resistors are mounted directly on this board along with the phase-compensating components (see below). The rotor is a small etched board mounting the two shunting resistors. The contacts are precious metal and the rotor takeoff is a slip ring.

The fine, or vernier, control is a wire-wound rheostat connected in series with the centade and compensated to obtain the desired linearity and phase characteristics.

The *D* and *Q* Adjustments and Readouts

The D and Q adjustments for the L and C bridges are two ganged 40-db exponential rheostats. The whole range of D or Q adjustment for each bridge is on a single scale, so that no multiplier is necessary. The appropriate scale is illuminated, as is the letter D or Q above the dial, so that there is no question of what function is being read.

Separate D scales are used for the series- and parallel-capacitance bridges, and separate Q scales for the series- and parallel-inductance bridges. The ranges have wide overlap, so that inductors with Q values from 1 to 50 (or capacitors with D's from 0.02 to 1) can be measured as either a series or parallel configuration. At Q values above 50 (D below 0.02), the difference between the series or parallel value is 0.04%, at most.

The Q balance for resistance or conductance measurements consists of two decades of capacitance and a variable capacitor, with dials arranged to give in-line readout. An indicating light indicates whether the Q balance is inductive (for R measurements) or capacitive (for G measurements). This readout also uses the X indication to facilitate balances that occur in the awkward region beyond 9 on a decade adjustment.

ACCURACY

The basic bridge accuracy of 0.1% at 1 kc is primarily a function of the

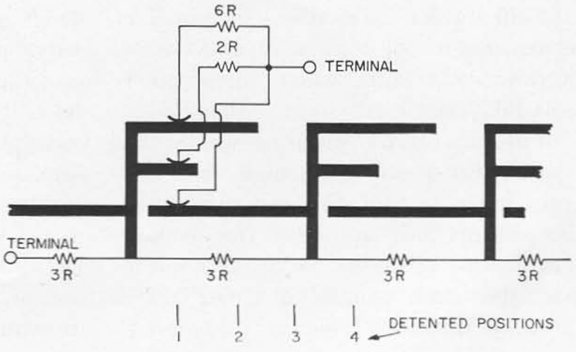


Figure 5. Functional diagram of the centade switching sequence.

calibration and the stability of the standard components used. All the fixed resistors, including those of the centade adjustment, are precision wire-wound units, similar to those used in General Radio decade resistors. The standard capacitor is similar to our precision silver-mica standards, but is shunted by a small, stabilized, polystyrene capacitor to reduce the over-all temperature coefficient. The total capacitance is 0.15 μf, so that in 3-terminal capacitance measurements substantial stray capacitance can be placed across it without causing appreciable error.

On the lowest impedance range of each bridge we have added an additional 0.1% to the accuracy specifications, because this range uses a one-ohm ratio-arm resistor, which is slightly less stable and more affected by possible variations in switch-contact resistance than are those of higher value. The range switches have solid-silver double contacts to keep these variations small.

The accuracy specification also includes a $\pm$0.005%-of-full-scale term, which is $\pm$½ division on the last digit of the indicator. This limitation is imposed by the ability to read the counter, the linearity of the fine adjustment, and backlash in the counter drive. This term reduces the bridge accuracy by a negligible amount at full scale but at 1/10 of full scale increases the possible error to $\pm$0.15% (except on the lowest impedance ranges when it becomes $\pm$0.25%).

The other error terms given in the specifications are important only for very high-D or low-Q measurements or have effect only at higher frequencies. It should be noted that the basic accuracy even at 10 kc is 0.2% for L and C measurements and 0.3% for resistance and conductance.

The residual impedances are the impedances associated with the unknown terminals themselves. An internal four-terminal connection is made to these terminals, so that the resistance is that of the binding post and the inductance is that of the loop completed by the shortest connection between the terminals. The 0.25-pf capacitance of the terminals can be removed by installation of a grounded shield between them.

Substantial effort was put into the design of this instrument to get the fixed phase error term down to $\pm$0.0005 radian (the D of C and L and the Q of R and G) which we felt was necessary for a precise bridge. For example, a 0.1% capacitor to be used in a precision twin-T null circuit requires a dissipation-factor measurement to 0.001, since this D error could cause just as much unbalance in the null circuit as a 0.1% capacitance error. The fixed error term, $\pm$0.0005, is the more important, since it is larger than the 5% term for D values up to 0.01.

With the resistance bridge, this phase precision enables one to check the Q of many bobbin-type, wire-wound resistors, which have appreciable reactance even at audio frequencies.

If the bridge phase shifts were not so closely controlled, a more subtle difficulty would occur. A D error could exist that would put the final balance off the end of the adjustment range for a very low-D component. If this D error were, for example, as great as 0.01, the capacitance balance would be limited to about ½%, since the null-meter deflection is proportional to the square root of the sum of the squares of both adjustment unbalances.

POWER SOURCES AND DETECTORS

An internal, two-transistor *RC* oscillator drives the bridge through a bridge transformer when the instrument is set for 1-kc measurements. It has an adjustable output and applies a maximum of about 1 volt behind 50 ohms. The detector uses six stages to get very high gain and 25-db second-harmonic rejection. This circuit has compression to give added range to the null-meter deflection to reduce the necessity of detector gain adjustments. On the extreme ranges, where more gain is required, an extra 20 db of gain is automatically added by the range switch.

The selective circuits used in the oscillator and detector are mounted on a module that slides in from the rear of the instrument and also provides a panel indication of the internal frequency. A 1-kc module is usually supplied, but other frequencies are available upon request. When modules for other frequencies are used, the proper frequency-dependent multiplication factors for the *D* and *Q* readings are also indicated on the panel. The instrument can also be used with an external oscillator, which is applied through the internal bridge transformer. The internal detector has a flat frequency response for this mode of operation, and an external selective detector, such as the TYPE 1232-A Tuned Amplifier and Null Detector, is useful for measurement at low levels or on nonlinear components.

Three internal dc supplies are included to give good dc sensitivity. These supplies of 350 volts, 35 volts, and 3.5 volts are current limited to avoid damage to the bridge or unknown and are adequate to apply the standard EIA test voltages for various types of resistors over most of the range. The range switch automatically chooses the optimum supply for each range. Moreover, the switching changes the manner in which the source is connected to the bridge to get maximum sensitivity and to prevent excess meter damping for low-resistance measurements.

The dc detector is a sensitive, shaped null indicator. To avoid the necessity of zero adjustment, no dc amplifier is used. With this system, 0.1% balance may be made from 1 ohm to 1 megohm if care is taken in reading the null detector. Provision is made for the use of external dc sources and high-gain dc null detectors.

SOME ADVANTAGES OF AC RESISTANCE MEASUREMENTS

The addition of a *Q* balance for the *R* and *G* bridges makes possible the precise measurement of resistors at 1 kc instead of at dc. This not only overcomes the sensitivity limitation of the dc bridges at the range extremes (see above), but also makes available some general advantages that ac bridges have over dc bridges. If the resistors are going to be used at ac, then an ac measurement is more logical. Further, the *Q* indication is useful in many cases. Ac measurements can be made at a much lower level than dc measurements, since ac detectors are generally more sensitive than dc detectors, because they have no dc drift. This means that, in many cases, the ac measurement is a better measure of the low-level *dc* value of resistors that are voltage or power sensitive.

The last statement is true only if there are no appreciable frequency effects that would make the zero-level ac and dc values different. There are many possible frequency effects to be considered, but the only important ones are distributed capacitance effects which cause difficulties only in the high megohm range at 1 kc. The trick that this bridge uses is that it requires that *series* resistance be measured when the unknown resistor is inductive and that *parallel* conductance be measured when the resistor is capacitive. Series inductance does not affect the value of series resistance, and parallel capacitance does not affect the value of parallel conductance (or parallel resistance). Of course, a resistor will have both series inductance and parallel capacitance, but the resonant frequency of a resistor would have to be 45 kc or lower to get an error of 0.1% at 1 kc. Few self-respecting resistors behave like that. (This does not mean to imply that the ac and dc resistances of a transformer or motor winding are anywhere near equal. Here, iron losses are the main cause of difference.)

APPLICATIONS

R, L, C Components

Bridges of this type are designed primarily to measure resistors, capacitors and inductors, and this one will measure components whose specified tolerances are well below 1%. It is often desirable to measure 1%, or poorer, components to much higher accuracy, for example, in acceptance tests on border-line cases, in quality-control work where a distribution within the tolerance range is desired, or in development work when cumulative tolerance effects are being studied. Although specialized precision bridges are recommended for measurement of reference impedance standards, this bridge is adequate to check most secondary standards used in production testing. Note that it will *compare* two components of decade value to 0.01%, since they can be balanced to the full-scale resolution.

Networks

The wide range and complete phase-angle coverage of this bridge make it useful for measuring the impedance of "black boxes." The bridge will balance for almost all passive impedances at 1 kc, the exceptions being the inductors above 1100h (which would probably be capacitive at 1 kc) and capacities above 1100μf (which should usually be measured at low frequencies). Thus, such things as potted networks, transducer impedances, and amplifier input and output impedances can be measured, and their audio-frequency characteristics plotted if an external oscillator is used.

In-Situ Capacitance Measurements

We've found that the ability to measure small capacitances has been useful for measuring the capacitance between components, wires or mounting structures. The advantage of three-terminal capacitance measurements is useful here, since it is possible to measure such quantities as the capacitance between any two conductors on an etched-board pattern with the others grounded. This ability to make measurements in the presence of large capacitance to ground permits the use of long shielded cables to connect remote or otherwise inaccessible components and to reduce the shunting effect of lead capacitance in the measurement of small capacitors.

Testing

The Type 1650-P1 Test Jig connects conveniently to the bridge (see cover photograph), thus placing quick-connect, spring terminals on the bench directly in front of the operator. If the gain of the instrument is adjusted to give a conveniently read deflection for a given bridge unbalance, this combination provides a versatile and accurate setup for the rapid tolerance testing of components.

— HENRY P. HALL

CREDITS

The TYPE 1608-A Impedance Bridge was developed by H. P. Hall. R. A. Soderman, Administrative Engineer; P. K. Bodge, Design Engineer; C. S. Kennedy, Layout Draftsman; W. H. Higginbotham, Production Engineer; and D. B. Bradshaw, Test Engineer, have all contributed to the final design.

— EDITOR

SPECIFICATIONS

RANGES

Capacitance: 0.05 pf to 1100 μf in seven ranges, series or parallel.

Inductance: 0.05 μh to 1100 h in seven ranges, series or parallel.

Resistance: 0.05 mΩ to 1.1 MΩ ac or dc.

Conductance: 0.05 n℧ to 1.1 ℧ ac or dc (20 kMΩ to 0.9 Ω).

***D* of Series *C*:** 0.0005 to 1.

***D* of Parallel *C*:** 0.02 to 2.

***Q* of Series *L*:** 0.5 to 50.

***Q* of Parallel *L*:** 1 to 2000.

***Q* of Series *R*:** 0.0005 to 1.2 inductive.

***Q* of Parallel *G*:** 0.0005 to 1.2 capacitive.

ACCURACY

C, G, R, L

At 1 kc: $\pm 0.1\% \pm 0.005\%$ of full scale except on lowest R and L ranges and highest C and G ranges where it is $\pm 0.2\% \pm 0.005\%$ of full scale.

Additional % error terms for high frequency and large phase angle:

C and L

$$\left[\pm 0.001\left(\frac{f}{1\text{ kc}}\right)^2 \pm 0.1\text{D}\frac{f}{1\text{ kc}} \pm 0.5\text{D}^2\right]\%$$

of measured quantity.

R and G

$$\left[\pm 0.002\left(\frac{f}{1\text{ kc}}\right)^2 \pm 0.000001\left(\frac{f}{1\text{ kc}}\right)^4 \pm 0.1\text{Q}\right]\%$$

of measured quantity.

Residual Terminal Impedance: $R \simeq 1$ mΩ, $L \simeq$ 0.15 μh, $C \simeq 0.25$ pf.

Dc Resistance and Conductance: Same as for 1-kc measurements, except that accuracy is limited by sensitivity at the range extremes. Balances to 0.1% are possible from 1Ω to 1MΩ with the internal supply and detector.

$D\left(\text{or } \frac{1}{Q}\right)$ of C or L:

$\pm 0.0005 \pm 5\%$ at 1 kc or lower.

$\pm 0.0005 \frac{f}{1\text{ kc}} \pm 5\%$ above 1 kc.

***Q* of *R* or *G*:** $\pm 0.0005 \frac{f}{1\text{ kc}} \pm 2\%$.

GENERATOR AND DETECTOR

Internal Oscillator: 1 kc $\pm 1\%$ normally supplied. Plug-in modules for other frequencies available on request. Level control provided.

Internal Ac Detector: Can be used either flat or selective at frequency of plug-in module (normally 1 kc). Second-harmonic rejection approximately 25 db; sensitivity control provided.

Internal Dc Supplies: 3.5 v, 35 v, 350 v; adjustable, and power limited to less than ⅓ watt.

Internal Dc Detector: Null indicator, 1 μa/mm.

External Oscillator and Detector: TYPE 1210-C Unit RC Oscillator and TYPE 1232-A Tuned Amplifier and Null Detector are recommended.

Dc Bias: Provision is made for biasing capacitors to 600 v with external supplies, and for biasing current in inductors.

GENERAL

Accessories Supplied: TYPE CAP-22 3-Wire Power Cord; spare fuses and indicator lamps.

Accessories Available: TYPE 1650-P1 Test Jig; external generator and detector, if used, as listed above.

Power Input: 105 to 125 (or 210 to 250) volts, 50-60 cps, 10 watts.

Mounting: Either relay-rack or bench, as listed below.

Dimensions: Rack model, panel, 19 by 12¼ inches (485 by 315 mm); bench model, width 19, height 12½, depth 11½ inches (485 by 320 by 295 mm), over-all.

Net Weight: 36¾ pounds (17 kg).

Type		*Code Word*	*Price*
1608-AM	**Impedance Bridge (Bench Mount)**............	ARGON	**$1175.00**
1608-AR	**Impedance Bridge (Rack Mount)**............	ANVIL	**1175.00**

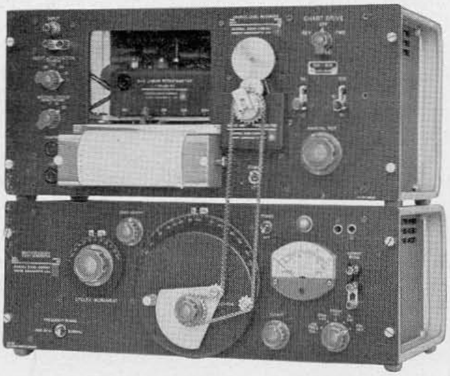
RELAY-RACK MOUNTING FOR THE OUTPUT POWER METER

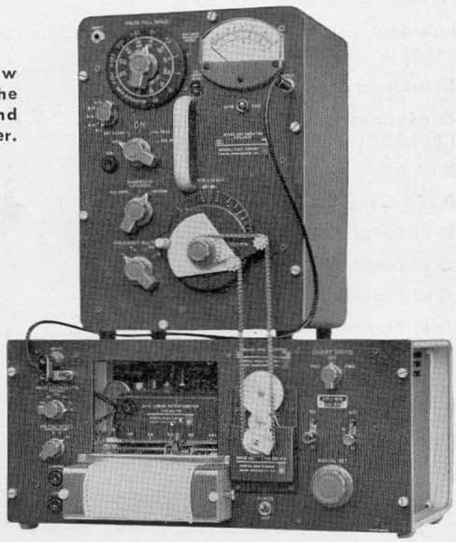
The Type 1840-A Output Power Meter, described in the January-February issue of the *Experimenter*, can be adapted for relay-rack mounting by the addition of panel extensions. Order Type 480-P212 Panel Extensions as listed below.

Type		*Net Weight*	*Code Word*	*Price*
480-P212	**Panel Extensions (pair)**	**4 oz (115 g)**	EXPANELBAT	**$6.00**

NEW LINK UNIT FOR THE GRAPHIC LEVEL RECORDER

The Type 1521-A Graphic Level Recorder with the new Type 1521-P14 Link Unit is shown driving (below) the Type 1304-B Beat-Frequency Audio Generator and (right) the Type 1554-A Sound and Vibration Analyzer.





A new Link Unit, Type 1521-P14, is now available to couple the Type 1521-A Graphic Level Recorder to the Type 1304-B Beat-Frequency Audio Generator or to the Type 1554-A Sound and Vibration Analyzer for the automatic recording of frequency response characteristics. With this link unit, the audio generator can be mounted either above or below the recorder. The analyzer is operated above the recorder.

The Type 1521-P11 Link Unit, with which the generator could be mounted only above the recorder, is discontinued.

Type		*Code Word*	*Price*
1521-P14	**Link Unit**	PANIC	**$18.00**

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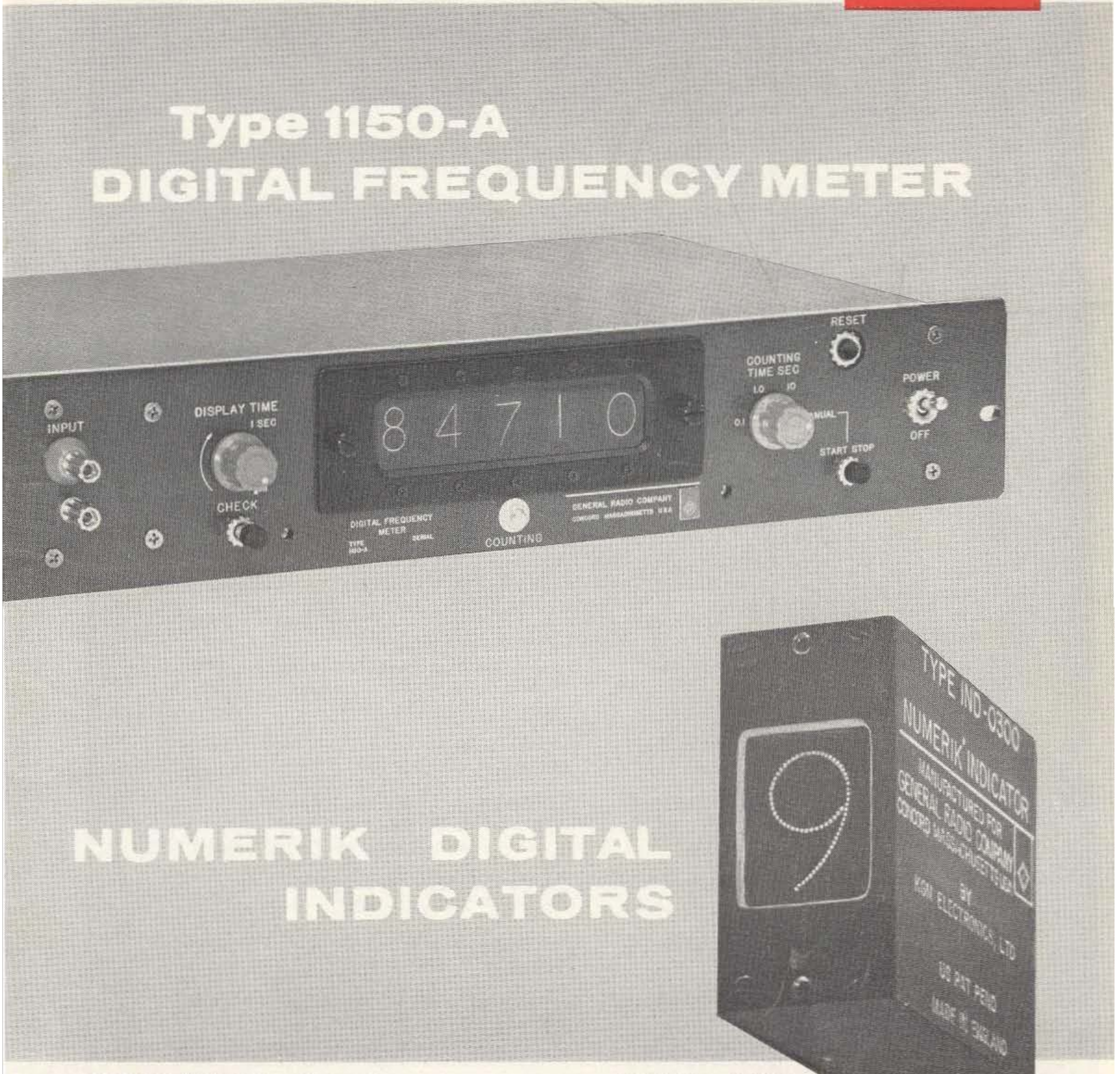
THE GENERAL RADIO

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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GENERAL RADIO COMPANY (OVERSEAS), ZURICH, SWITZERLAND
REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

A FIVE-DIGIT SOLID-STATE COUNTER FOR FREQUENCY MEASUREMENTS TO 220 kc

The TYPE 1150-A Digital Frequency Meter is the first of a line of inexpensive, basic counters for laboratory and industrial applications. Like the TYPE 1130-A Digital Time and Frequency Meter announced a year ago, it is accurate, reliable and easy to operate. Unlike the TYPE 1130-A, which has memory circuits and a neon-tube columnar readout, the TYPE 1150-A uses transistors instead of tubes and an in-line readout with alternate counting and display periods.

The counting circuits used are simple, but unconventional by today's practice. Instead of a scale-of-ten derived from a scale-of-sixteen circuit through appropriate feedback, the TYPE 1150-A is based upon the earlier ring-of-ten counting system.[1] Although the ring circuits require ten bistable elements, instead of four,[2] we have found[3] that modern semiconductors can be used in a simple and economical design. Not only is the overall number of circuit components for a complete decade, including indicators, actually less than that required for the coded scale-of-sixteen circuit, but performance at high counting rates is better, and the circuits can be designed to have tremendous latitude in accommodating variations in transistor characteristics.

Transistors typically operate at relatively low voltages and high currents and are therefore better suited to lighting incandescent lamps than to operating gas-discharge-type indicators. The ring-of-ten circuit is, as a result, directly adaptable to the ten-lamp, end-fire-illuminated NUMERIK indicator (also described in this issue) without interconnecting circuitry. Since the count proceeds around the ring, one flip-flop at a time, in identical reset-set operations, there is no time lost in feedback

[1]See, for instance, V. H. Regener, *Review of Scientific Instruments*, 17, p. 180, 1946.

[2]R. W. Frank and H. T. McAleer, "A Frequency Counter with a Memory and with Built-in Reliability," *General Radio Experimenter*, 35, 5, May, 1961.

[3]A similar approach, for instance, but using complementary techniques, is described by R. A. Stasior, *G. E. Technical Information Series RG1-EGP16*, May 16, 1961.

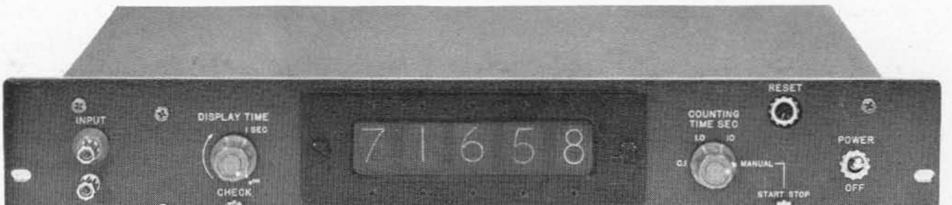


Figure 1. The Type 1150-A Digital Frequency Meter. Encased in General Radio's multi-purpose cabinet,[4] the instrument is readily mounted in a rack or adapted to table-top use with end frames. The panel is 3½ inches high. The controls and fittings are simple and clearly marked. Frequencies between 10 cps and 220 kc can be counted over preselected time intervals from 0.1 to 10 seconds. Display times can be continuously adjusted from 0.5 to 5 seconds and ∞.

[4]H. C. Littlejohn, "The Case of the Well-Designed Instrument," *General Radio Experimenter*, 34, 3, March 1960.

mechanisms, and the effects of delays in the transistors are therefore minimized. There is, furthermore, no need to interlock dc levels between various parts of the circuit to maintain adequate margins for reliability. The ring circuits used in this counter are, in fact, so noncritical that they will work with intermixed transistors, ranging in characteristics from the lowest-cost alloy-junction transistors to MADT types!

The fact that incandescent lamps can be lit directly by the collector currents of the counting transistors themselves leads to the utmost simplicity and economy. At the same time, this very simplicity puts a premium on displaying the count according to the conventional count-display time cycle, rather than the continuous regime made possible by the use of auxiliary memory circuits. For ultimate ease of reading and efficiency of counting, the memory system pioneered by the TYPE 1130-A Digital Time and Frequency Meter remains pre-eminent. The need for memory in this less expensive instrument is less, however, because the eye is less fatigued by the blurring that occurs during the counting interval in in-line presentation than by the running-up-and-down appearance of columnar displays.

Five-digit presentation necessarily means, for a full register, a precision of ± 1 part in 10^5, determined by the ± 1 count uncertainty in the last place. The TYPE 1150-A Digital Frequency Meter, like other GR counters, is designed to yield an accuracy figure that takes full advantage of its inherent precision. Even in this simple, inexpensive instrument, this demands the inclusion of an oscillator with a stability over several minutes of at least ½-part per million. This accuracy is necessary because the counter can count a signal of 220-kc frequency for up to 10 seconds. A temperature-controlled quartz crystal, operating at 100 kc, is therefore used as the fundamental reference for the various timing signals required to produce the accurately known time intervals during which the counts accumulate.

MECHANICAL DESIGN

The mechanical design and component layout are simple and efficient, with all circuits easily accessible for maintenance and open for cooling (Figure 2). Ring counting units, input circuits and time-base dividers are on removable, vertical plug-in cards. These cards are arranged at right angles to the front panel and directly behind a large air filter. A small fan expels air from the box and draws cool air in over the vertical circuit cards. With this cooling system a maximum temperature rise at the hottest spot on any of the circuit boards is less than 10°C, thereby insuring reliable operation at high ambient temperatures.

The NUMERIK indicators (see page 10) have a large and efficient heat sink for their incandescent lamps, which prolongs lamp life. In the event of failure, however, any individual lamp can be easily replaced. The group of five NUMERIK indicators are accessible from the front panel of the instrument. A quarter turn of two mounting thumbscrews permits the entire block of indicators to be pulled forward and withdrawn several inches. Eight replacement 330-type lamps are stored in pockets behind the bezel. In order to replace a burned-out lamp, one simply removes two screws holding the backing plate of the indicator and replaces the lamp in the numbered socket. The entire re-

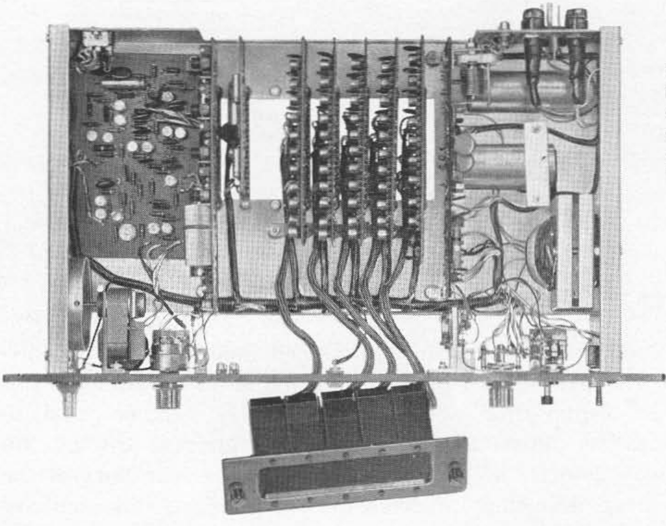


Figure 2. Interior view, showing NUMERIK Indicator assembly detached. The eight spare indicator lamps can be seen behind the panel to the left of the NUMERIK assembly.

placement can be accomplished in about one minute after a defective lamp has been isolated.

When a lamp fails, the entire decade becomes inoperative; no light shows. This fail-safe feature eliminates any possibility of incorrect reading.

A Polaroid filter is used in front of the bank of five indicators to eliminate specular reflection.

GENERAL CIRCUIT DESCRIPTION

The Type 1150-A Digital Frequency Meter counts the number of cycles of the input signal in a time interval established by the stable 100-kc crystal oscillator. The block diagram, Figure 3, shows the functional elements of the system.

The signal from the 100-kc crystal oscillator (or from an external 100-kc frequency-standard) is converted to a pulse train to drive a series of precision frequency dividers utilizing unijunction transistors. The output of this divider chain consists of pulses with periods of 0.1, 1, and 10 seconds. These pulses establish the precision gate interval.

In order to understand the program, let us assume that the time-base gate control flip-flop is in the "1" state and that the time-base gate is therefore open. The main-gate control flip-flop is in the "0" state, and the main gate is closed. The next pulse from the time-base divider circuits passes through the time-base gate, complements the main-gate control flip-flop to the "1" state and opens the main gate. When the main gate is open, pulses of the input-signal frequency are admitted to the five ring counting units and totaled. The next pulse from the time-base divider passing through the time-base gate resets the main-gate control flip-flop and closes the main gate. The main gate is therefore open for exactly one period of the time-base signal.

The NUMERIK indicators now begin to display the resultant count. This display interval is determined by the display-time monostable circuit and is adjustable from 0.5 to 5 seconds. When the main-gate control flip-flop closes the main gate, it also sets the time-base gate control flip-flop to the "0" state, closing

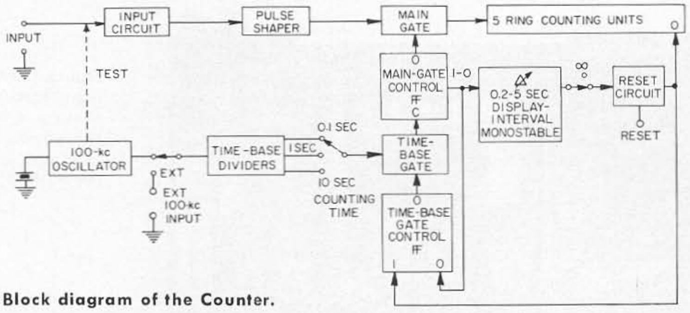


Figure 3. Block diagram of the Counter.

the time-base gate so that no more time-base pulses can pass to the main-gate control flip-flop until the display-time monostable circuit issues its output pulse operating the reset circuit. The reset circuit sets the ring counting circuits to zero, and the end of the zero-setting pulse will set the time-base gate control flip-flop to "1." The process then repeats.

An infinite display time can be obtained if the output of the display-time monostable circuit is opened so that no reset pulse is produced. A switch at one end of the time-interval adjustment range performs this operation and makes possible indefinite display of the results of a single measurement initiated when the manual reset button is pressed.

Provision is also made for manual control of the counting time. When the counting-time switch (Figure 3) is set to a fourth position, the main-gate flip-flop can be controlled from a push-button. The main gate is opened when this button is released and closed when the button is depressed. With manual gating, one can use the counter as a simple totalizing instrument. In the manual position, when the main-gate control flip-flop is set to "0" a normal display and clearing cycle is initiated. A normal display time of 0.5 to 5 seconds can be obtained or the total reading can be retained indefinitely if the display-time switch is set to infinity.

External circuitry can be used to control the gate interval through an auxiliary plug on the rear skirt of the instrument. In addition, this auxiliary plug also provides: (1) the carry pulse from the last of the five decades; (2) means for applying an externally generated reset pulse; (3) a positive pulse corresponding in time to the reset pulse; (4) +20 volts dc; and (5) ground.

RING COUNTING UNITS

The heart of this counter is the ring counting unit. A ring counting system can be considered to be a stepping switch with ten contact positions, as shown in Figure 4. A lamp is connected

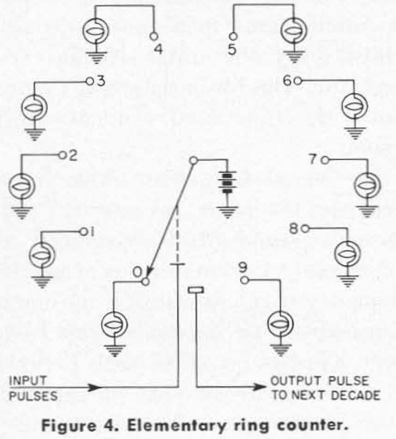


Figure 4. Elementary ring counter.

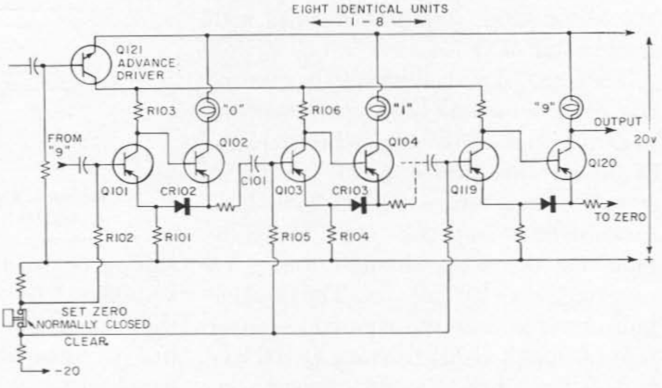


Figure 5. Simplified schematic of one of the ring counting units used in the Type 1150-A Digital Frequency Meter.

to each contact, and the switch rotor is connected to a source of voltage. Each time a pulse is received, the rotor moves from one contact to the next, and the particular lamp that is lighted indicates the number of pulses that have been received. By generating an output pulse each time the rotor moves from "9" to "0," the ring counter functions as a simple decade divider and drives the succeeding unit at one-tenth the rate.

Five ring counting units (RCU's) are used in the TYPE 1150-A Digital Frequency Meter. Twenty-one transistors are required for each decade, two for each of the ten digits, and one more to "move the rotor." Figure 5 is a simplified schematic of one of these five ring counting units. All five are identical in structure. The first unit differs from the other four only in the choice of component values for best high-frequency operation. Each unit consists of a ring of ten bistable circuits, each of which has one "high-current" transistor that drives its associated incandescent lamp in the NUMERIK indicator for that decade.

Referring to Figure 5, let us assume that the decade has been set to its zero state. Q-101 will be OFF and Q-102 ON. Q-102 has its base forward-drive current provided through R-103 and is in saturation, passing 80 milliamperes to light the "0" lamp in the indicator. This 80 milliamperes produces a voltage of -5.5 volts across R-101. The base of Q-101 is returned via R-102, to the set zero buss voltage of about -5.0 volts. The base of Q-101 is, therefore, reverse biased with respect to the emitter, and Q-101 remains OFF. The circuit is stable in this state.

All other pairs in the ring have the opposite stable state. Left-hand transistors, Q-103, etc., are all saturated, and the right-hand transistors, Q-104, etc., are OFF. These are also stable states. Look at Q-103, for example. When it is ON, nearly 1 milliampere of base forward drive flows through R-105, which is connected to the clear buss (at the same potential as the set zero buss). The drop across the 68-ohm resistor R-104 in the common emitter circuit is but 0.07 volt, and practically the full 20-volt collector supply voltage appears across R-106. The very small emitter-to-collector drop of Q-103 is normally below the conduction-knee voltage of Q-104 and keeps it OFF. Complete cutoff of Q-104, even at elevated temperatures and for all possible transistor combinations, is insured by the silicon diode (CR-103 in series with the emitter of Q-104).

The input signal advances the state of the decade by one stage per pulse. A negative pulse is first applied to the base of the advance driver Q-121, turning it OFF. The lamp driver, Q-102, loses base forward drive and goes OFF. The common-emitter voltage changes from −5.5 to zero, and Q-101 goes ON. The positive pulse produced at the common emitter is fed through C-101, turning Q-103 OFF and Q-104, the "1" lamp driver, ON. Each succeeding pulse applied to Q-121 advances the count by one digit. At the count of ten the circuit is switched to the initial conditions, and the negative pulse, as the "9" lamp extinguishes, is fed from the RCU as a carry pulse to the advance driver of the succeeding RCU.

In the simplified schematic of Figure 5 the zero-set system is depicted as a manual switch for simplicity. Opening this switch obviously returns the clear buss to −20 volts, causing all left-hand transistors of the bistable circuits to saturate, and turning the lamp drivers for lamps "1" through "9" OFF. Q-101, on the other hand, loses its forward bias, desaturates, and permits Q-102 to go ON, thereby turning the "0" lamp ON. A fast transistor switch is actually used in the TYPE 1150-A Digital Frequency Meter to accomplish zero setting for all five RCU's.

TIME BASE

The elementary circuit for the 100-kc crystal oscillator is shown in Figure 6. It uses two transistors, an *npn* and a *pnp*, in a modified Pierce circuit. All the open-loop gain (60 db) of this transistor pair is used as negative feedback. Thus, the circuit gain is very stable with respect to variations in temperature, voltage, and transistor parameters, re-

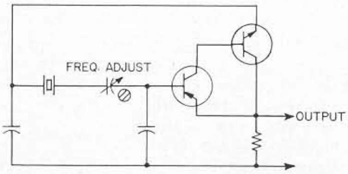


Figure 6. Elementary schematic of crystal time-base oscillator.

sulting in excellent oscillator-frequency stability. The temperature control for the crystal operates as long as the power line is connected, whether or not the panel power switch is on.

APPLICATIONS

The basic counter has many frequency-measurement applications. It can be used to calibrate, and to adjust the calibration of, any signal source lying in the frequency range from a few cycles per second to a maximum counting frequency of 220 kc. A few of its more important applications are: the calibration and setting of oscillators and signal generators, the monitoring of frequencies to 220 kc, and measurements on precision filters and other frequency-selective devices, where high resolution is needed for accurate frequency setting.

Other physical measurements offer a wide range of uses, including the measurement of rotational speed (with a photoelectric or magnetic pickup),[5] and, with appropriate transducers, pressure, temperature, strain, and weight.

With an appropriate transducer, photoelectric or otherwise, the counting of objects on production lines, of particles in liquids, and of similar events that may not be periodic will form an-

[5] A photoelectric pickoff device is now in the final stages of design at General Radio and will soon be available for use with the basic counter and with our stroboscopic equipment line. This unit is a completely self-contained light source and photoelectric cell, which will be powered by the counter or the stroboscope and which will feed its output signal back to the counter or the stroboscope for measurement. A three-connector jack is provided in the rear of the counter to accept this new photoelectric pickoff.

other group of important applications. The simplicity of operation, the accuracy and the reliability of this counter are determining factors in its acceptability in many of these applications.

— R. W. Frank
— J. K. Skilling

CREDITS

The Type 1150-A Digital Frequency Meter was developed by R. W. Frank and J. K. Skilling. H. P. Stratemeyer, Development Engineer, R. A. Mortenson, Design Engineer, H. G. Stirling, Designer, R. A. Chipman, Production Engineer, and William Howard, of Cornell University, have all contributed to the final design.

— Editor

SPECIFICATIONS

Frequency Range: 10 cps to 220 kc.

Input Impedance: AC-coupled approximately 0.5 megohm shunted by less than 100 pf.

Sensitivity: Better than 1 volt, peak-to-peak; for pulse input, duty ratio should be between 0.2 and 0.8.

Display: In-line register, incandescent-lamp-operated.

Display Time: Adjustable from 0.5 to 5 seconds, approximately, or ∞.

Counting Interval: 0.1, 1, 10 seconds, or can be set manually.

Accuracy: ± 1 count ± crystal oscillator stability.

Crystal-Oscillator Stability

Short-Term Stability: Better than ½ part per million.

Cycling: Less than counter resolution.

Temperature: 2½ parts per million for 0 to 50°C ambient.

Warm-up: Within 1 part per million after 15 minutes.

Aging: Less than 1 part per million per week after 4 weeks; decreasing thereafter.

Crystal Frequency Adjustment: The frequency is within 10 parts per million when received. Frequency adjustment provided.

Power Input: 105 to 125 (or 210 to 250) volts, 50 to 60 cps, 45 watts.

Accessories Supplied: Type CAP-22 3-Wire Power Cord; spare fuses; 8 replacement indicator lamps.

Dimensions: Bench model, width 19, height 3⅞, depth 12½ inches (485 by 99 by 320 mm), over-all; rack mount, panel, 19 by 3½ inches (485 by 90 mm); depth, 12¾ inches (325 mm).

Net Weight: 17½ pounds (8 kg), approximately.

Type		*Code Word*	*Price*
1150-AM	**Digital Frequency Meter, Bench Mount**	OFFER	**$915.00**
1150-AR	**Digital Frequency Meter, Rack Mount**	OCCUR	**915.00**

Patent Applied For.

NEW 3-WIRE POWER CORD

The Type CAP-22 Power Cord (3-wire) is now supplied with all General Radio instruments that use a detachable power cord, replacing the Type CAP-15 formerly supplied. This new design has a hammerhead connector at its power-line end, with both male and female connectors, thus permitting two or more cords to be attached in parallel. Two No. 18 conductors are used. The covering is rubber, and the connector bodies are molded integrally with the cord. This power cord is also available separately, as listed below. Its length is 7 feet, and it is rated at 250 volts, 7 amperes.

Type CAP-22 Power Cord.

Type		*Code Word*	*Price*
CAP-22	**Power Cord**	TRUCO	**$2.25**

THE NUMERIK INDICATOR

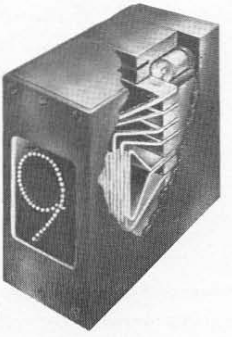


Figure 2. Cutaway view of the Type IND-0300, showing construction.

Selection of the best available readout indicator has been an important part of the development of new in-line-presentation instruments like the one described in this issue.

In the majority of general-purpose applications, and particularly in transistor circuits, the use of incandescent lamp illumination is especially suitable. Transistors operate typically at low voltages, and incandescent lamps adapt easily to these conditions without requiring complicated ancillary circuitry.

After a careful survey of available designs, the products of K.G.M. Electronics, Richmond, England, were judged to have the best combination of these desirable characteristics:

1. Excellent presentation with clear, brilliant readout. The white light is both more pleasing to the eye and actinically more efficient than the orange-red of neon displays.
2. Neat, compact design, with effective means of heat dissipation to insure long lamp life.
3. Use of standard, readily available replacement lamps. Replacement, infrequently required, is easily done.
4. Efficient use of light from low-power lamps.
5. Low parallax. All symbols appear to be nearly in the same plane.
6. Wide viewing angle.
7. Reasonable price.

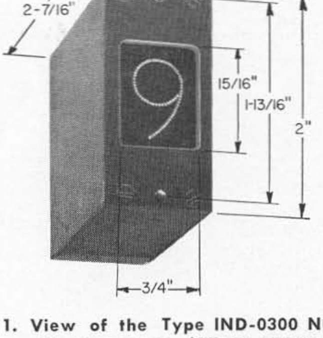


Figure 1. View of the Type IND-0300 NUMERIK Indicator, with dimensions. (Allow additional 1/16" on depth, for common terminal.)

The TYPE IND NUMERIK Indicators combine ten (or twelve) incandescent bulbs which can be illuminated individually and which, in turn, end-fire illuminate clear plastic strips, as illustrated in Figure 2. Light is introduced at one end of a thin, clear acrylic plate and is conducted with little attenuation along reflective ducting to the display surface where it is translated into a bright display by closely spaced dots scribed in the form of the numeral or symbol. A stack of ten plates is just over 5/16 inch deep.

Light transmission through the acrylic is so good that the bottom symbol of the stack appears to have about the same brightness as the top symbol. Thin sheets of reflective opaque material, which separate the paths through which the light to the display surface is conducted, reduce cross illumination to the point where all symbols except the one illuminated are, for practical purposes, not visible.

Because of the thin stack and excellent light conductivity, the NUMERIK Indi-

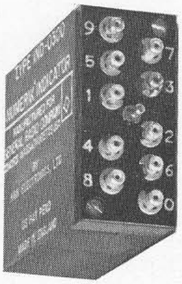


Figure 3. Rear view of the Type IND-0300, showing terminals.

cator has the unusually wide viewing angle of 120°.

The units are conveniently mounted behind the panel with only two screws.

To achieve long lamp life, the lamps are mounted in a drilled, solid aluminum block which serves as an efficient heat sink. Further, the sink is joined to the front panel of the instrument by large-cross-section aluminum side blocks. This configuration leads to cool operation and to lamp life averaging 5,000 hours under switching conditions.

Lamps are readily replaced. The removal of two screws at the back of the Indicator frees the lamp block and terminal plate as a unit.

Typical uses of the NUMERIK Indicator are found in annunciators, computers, counters, digital voltmeters and similar instruments, indicator boards for process control, paging systems, programmers, radar, timing systems, and clock displays.

Two types are available from stock: The TYPE IND-0300, which has ten numerals, zero through nine; and the TYPE IND-1801, which has the ten numerals plus a comma on the right side and a decimal point centered on the left side of the numerals. Additional types with letters and other symbols are available on special order.

The NUMERIK Indicators are manufactured for General Radio by K.G.M. Electronics under an agreement that makes General Radio the exclusive distributor for the United States and Canada.

SPECIFICATIONS

Lamps: 14-volt, 80-milliampere, 0.5 candle-power T-1 3/4 bulb; G.E. #330 or equal. Working life approximately 5000 hours (switching with 10% duty ratio).

Viewing Angle: 120° horizontal; 60° vertical.

Symbol Height: 13/16 inch.

Lamp Holder Block: Solid aluminum heat sink with nylon-filled Bakelite backing block. Nickel-silver contact springs and 11 silver-plated terminals (13 for TYPE IND-1801), one for each lamp and a common ground. The ground connection is to the case of the TYPE IND-0300; it is insulated from the case in the TYPE IND-1801.

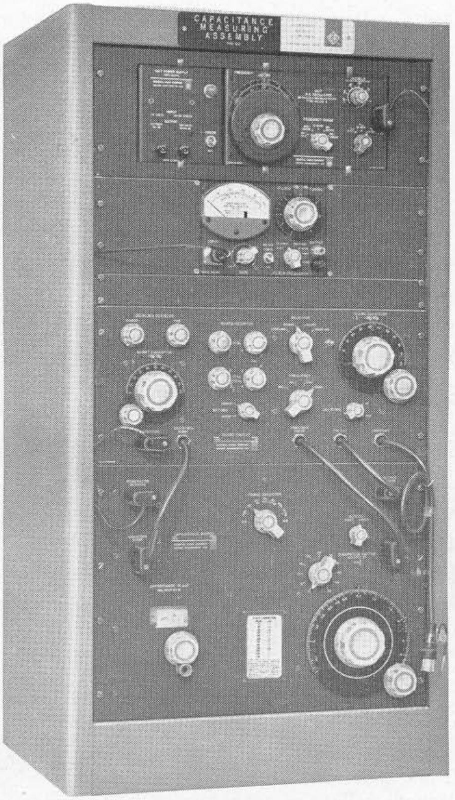
Mounting: Back-of-panel by two 4-40 screws.

		TYPE IND-0300	TYPE IND-1801
Window Size:		3/4 by 15/16 inch	3/4 by 1 1/16 inches
Mounting:		1 13/16 inches, centers	2 3/16 inches, centers
Dimensions:	Width	1 inch	1 inch
	Height	2 inches	2 3/8 inches
	Depth (including terminals)	2 1/2 inches	2 11/16 inches
Net Weight:		4 1/2 ounces	5 ounces
Code Word:		INDAK	INDIG
Prices:	1–19	$32.20	$33.60
	20–49	30.60	32.00
	50–99	28.60	30.00
	100–299	27.20	28.60
	300–999	24.70	26.10
	1,000–4,999	22.00	23.30
	5,000–9,999	18.40	19.60
	10,000–up	16.90	18.10

Patent Applied For.

TYPE 1610-B

CAPACITANCE-MEASURING ASSEMBLIES



View of Type 1610-B Capacitance-Measuring Assembly.

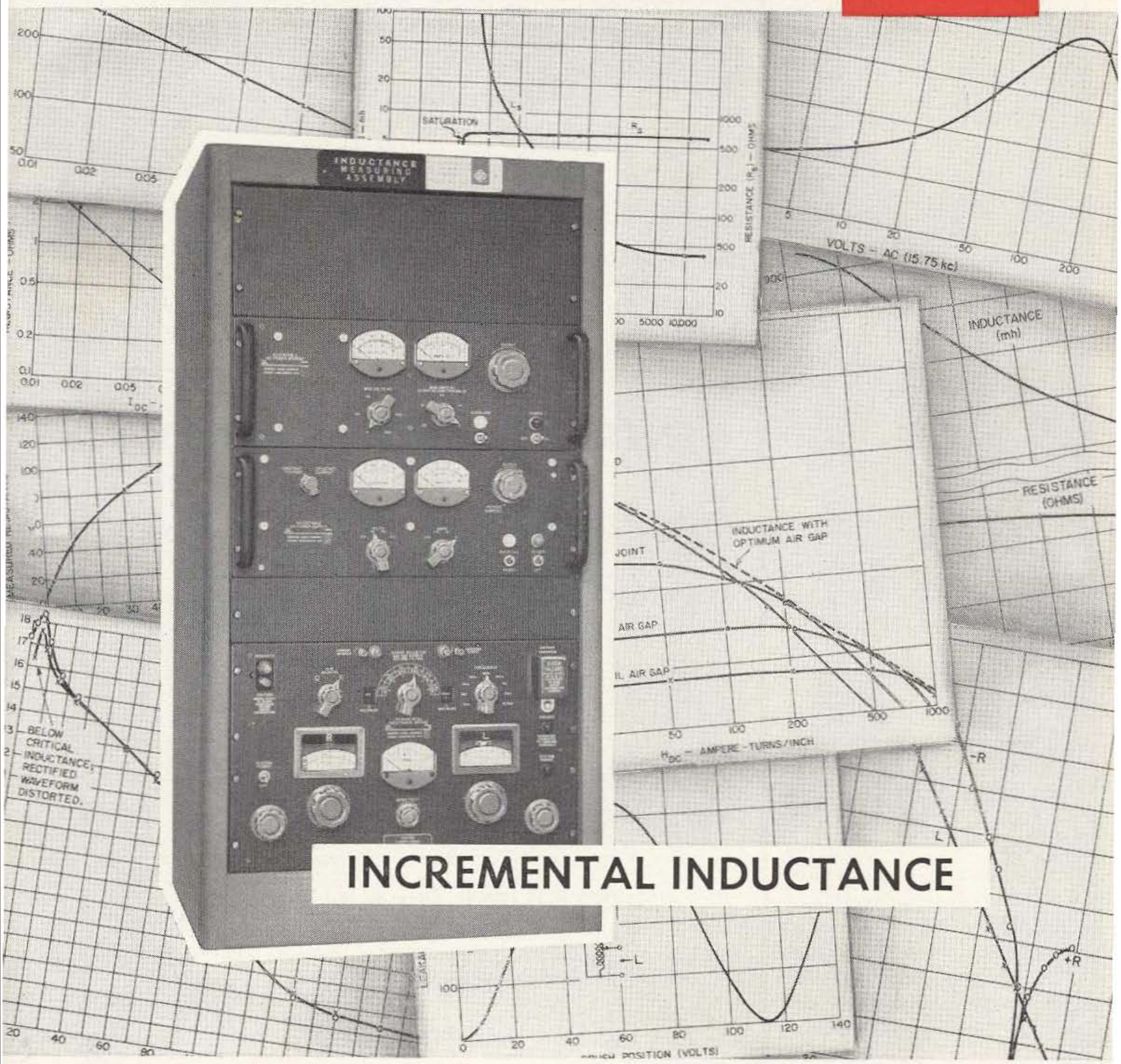
This capacitance-measuring system is now equipped with the Type 1210-C Unit RC Oscillator and the new Type 1232-A Tuned Amplifier and Null Detector, which replace the Type 1302-A Oscillator and the Type 1231-BRFA Amplifier and Null Detector used in the Type 1610-A Assemblies. Sensitivity is increased over that of previous models, and the detector is continuously tunable from 20 cps to 20 kc, with two additional fixed points, 50 kc and 100 kc. The oscillator frequency is continuously adjustable from 20 cps to 500 kc. The capacitance bridge, Type 716-C, remains unchanged.

Prices for the Type 1610-B Assemblies are lower than for the Type 1610-A, owing to the use of the less expensive but more modern oscillator and detector.

Two models are available: (1) the Type 1610-B, which includes the Type 716-P4 Guard Circuit, and is suitable for both three-terminal and two-terminal measurements, and (2) the Type 1610-B2, which is suitable for two-terminal measurements only. Each is complete in cabinet-type relay rack, with connection cables and power cord.

Type		*Code Word*	*Price*
1610-B	**Capacitance-Measuring Assembly**..........	SEDAN	**$1740.00**
1610-B2	**Capacitance-Measuring Assembly**.......... **(less Guard Circuit)**	SABER	**1465.00**

THE GENERAL RADIO
EXPERIMENTER



INCREMENTAL INDUCTANCE

VOLUME 36 No. 5 MAY, 1962

IN THIS ISSUE

New Incremental-Inductance Bridge with Power Supplies

THE GENERAL RADIO

EXPERIMENTER

Published Monthly by the General Radio Company

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GENERAL RADIO COMPANY (OVERSEAS), ZURICH, SWITZERLAND
REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

A NEW SYSTEM FOR MEASURING THE INDUCTANCE OF IRON-CORE COILS

The TYPE 1630-A Inductance Measuring Assembly was designed primarily for measurements of the inductance and loss of transformers, chokes, and similar components at high dc and ac excitation levels. Easy to operate and flexible in application, it can also measure other nonlinear elements such as Zener diodes, rectifiers, thermistors, and lamps.

This article describes the design of the bridge and its associated power supplies and shows a number of examples of measurements.

A system for measuring the inductance and loss of coils with ferromagnetic cores is necessarily more complicated than the simple bridge that suffices for measurements on air-core coils. The iron-core coil is nonlinear, and, for a sinusoidal applied voltage, the current will contain harmonics. Since inductance is usually defined with respect to the fundamental component of current flowing with a sinusoidal applied voltage, for meaningful measurements the ac source driving the measuring instrument must have a very low impedance to harmonics, and any impedances which the measuring system places in series with unknown impedance must be very small with respect to the unknown. To minimize the effects of harmonics, the detector must be sharply tuned to the fundamental frequency. In addition, because the inductance is a function of the applied ac voltage, dc bias current, and previous history of the core, the instrument must be capable of making the measurement with the ac voltage level and dc bias current at which the inductance is desired.

In iron-core coils the term incremental inductance rigorously refers to the ac inductance measured with a small signal superimposed on a relatively large dc bias current,[1] but the term is also frequently used to describe inductance as measured over a wide range of ac and dc excitation. In other important measurements no dc bias is applied. Therefore, the system must be able to handle high voltages and currents, and these must be adjustable.

A new measuring system, the TYPE 1630-AL Inductance Measuring Assembly, has been developed, which is capable of meeting these requirements. This system, shown in Figure 1, operates at the power-line frequency and consists of a bridge, a 200-watt dc source, and a 200-voltampere ac line-frequency supply. The bridge unit includes a null detector.

A second assembly, the TYPE 1630-AV, now under development, includes a 200-va audio oscillator in place of the line-frequency supply and is capable of measurements from 20 cps to 20 kc.

An accuracy of 1% has been found to be adequate for practically all measurements on ferromagnetic components, which makes possible the use of a convenient single-dial readout for inductance and another for series resistance or Q.

[1]F. E. Terman, *Radio Engineering*, 3rd Edition, McGraw-Hill, 330 West 42nd St., New York 36, N. Y.

Many other nonlinear elements present the same measurement problems as iron-core coils, and the TYPE 1630 assemblies are well suited to measurements on such components as Zener diodes, rectifiers, neon lamps, incandescent lamps, and thermistors.

THE BRIDGE CIRCUIT

The TYPE 1633-A Incremental-Inductance Bridge, the main element in the measuring system, uses a new circuit, which includes active elements,[2] in order to achieve important operational features. The active elements are three multistage, transistor, feedback amplifiers, designed to have parameters at least an order of magnitude more stable than is required for the desired bridge accuracy. Two amplifiers are used for isolation and a third for a phase inversion. Figure 2 is the elementary bridge schematic.

[2] H. P. Hall, R. G. Fulks, "The Use of Active Devices in Precision Bridges," *Electrical Engineering*, May 1962.

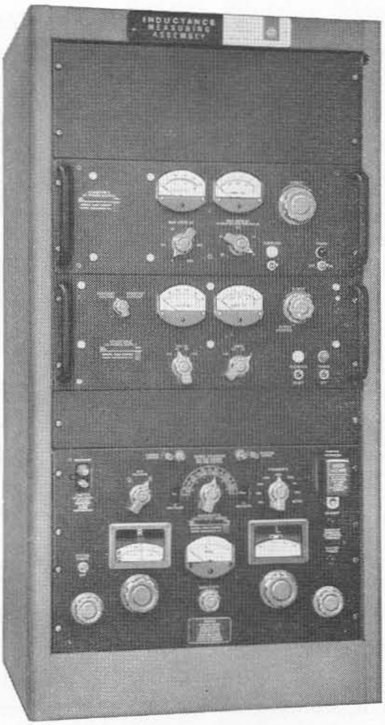

Figure 1. View of the Type 1630-AL Inductance Measuring Assembly. Space is provided at the top of the rack for the addition of an oscilloscope, which permits the current waveform or the hysteresis loop to be viewed during the measurements.

Each isolation amplifier is used with a potentiometer to form a variable-voltage source with a low output impedance. This permits the use of fixed capacitance and conductance standards, C_S and Q_S, because the current through these impedances is adjusted by variation of the voltage applied to them, rather than by variation of their magnitudes. Thus, both adjustments are simple, easily balanced potentiometers rather than multiple decade assemblies. The negative-gain amplifier on the right-hand side of the bridge converts the voltage from the resistive divider into a current of opposite phase, which is required for a bridge balance.

Null Conditions

The equation for null can be easily obtained by setting the sum of the currents into the detector equal to zero (Figure 2). If we let R_B' equal the total resistance to ground from point B, including R_F, then

$$\frac{I_1 + I_2 + I_3 + I_4}{E_{IN}} = \frac{R_B'}{R_B' + R_X + j\omega L_X}\left(\frac{1}{R} + \alpha j\omega C_S + \alpha\beta G_S\right) - \frac{R_D}{R_E(R_C + R_D)} = 0$$

where α and β are the ratios of potentiometer output voltage to input voltage,

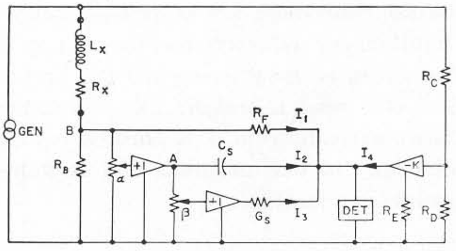

Figure 2.
Simplified schematic of the bridge.

or $\frac{1}{R_F} + \alpha j\omega C_S + \alpha\beta G_S =$

$$\frac{R_B' + R_X + j\omega L_X}{R_B'} \times \frac{R_D}{R_E(R_C + R_D)}$$

If

$$\frac{1}{R_F} = \frac{R_D}{R_E(R_C + R_D)},$$

then

$$L_X = \frac{\alpha C_S R_B' R_E(R_C + R_D)}{R_D}$$

$$R_X = \frac{\alpha\beta G_S R_B' R_E(R_C + R_D)}{R_D}$$

$$Q_X = \frac{\omega L_X}{R_X} = \frac{\omega C_S}{\beta G_S}$$

The unknown inductance is proportional to the value of α, and the dial of the α potentiometer is calibrated to read L. For the connection shown in the schematic, Q_X is inversely proportional to β, so that the dial on the second potentiometer reads Q_X. However, if this potentiometer is connected to point B instead of point A, α disappears from the equation for R_X, making resistance proportional to β, and the dial therefore reads R_X.

Advantages of the New Circuit

This circuit has several advantages over conventional Maxwell or Owen bridges as ordinarily used for this measurement.

1. Both balance controls are single-dial potentiometers permitting rapid balances.

2. The bridge reads either R_X or Q_X directly. A measurement in terms of R_X is desirable for low-Q values where the Q arrangement has a bad "sliding null" (slow balance convergence).

3. The bridge can read Q directly, with no multiplying factor, at several frequencies if the value of G_S is switched as the operating frequency is changed.

4. The voltage and current applied to the bridge and the unknown inductor are constant as the balance adjustments are made. This is a valuable feature in the measurement of nonlinear devices.

5. The generator and detector are both grounded, avoiding the need for bridge transformers that may be susceptible to magnetic pickup.

The use of active elements, which make possible these features, may bring to the minds of some readers a host of difficulties long associated with vacuum-tube amplifiers, particularly those employing little feedback. However, such problems as gain stability, life, hum, noise, excess heat, and warmup time can be completely eliminated or greatly reduced by the use of multistage, transistor, feedback amplifiers. When one thinks of the high precision and reliability of the operational amplifiers used in modern analog computers, one wonders why they have not been used more extensively in bridge circuits.

Ranges and Accuracy

The arrangement of the panel controls is shown in Figure 3. The six ranges of inductance are more than adequate for practical measurements at any given frequency. Moreover, the standard capacitor, C_S, is switched by the frequency-selector switch so that the inductance range is changed to give reasonable values at the various operating frequencies. For example, the inductance range extends up to 1,000 henrys on the four lower frequency positions, but an inductance of this size would surely be above resonance at 1 kc, so that the maximum range is 100 henrys at 1 kc and is reduced to 10 henrys at 10 kc. This shift in range permits measurements down to 0.1 μh at the higher frequencies.

The bridge reads R or Q as selected by the panel control. The Q scale is direct reading at nine frequencies, which include common power frequencies and their second harmonics, as well as other standard test frequencies. The Q range is ∞ to 1, below which a bothersome sliding null occurs; the R scale should be used for these low-Q measurements. Inductors or inductive resistors of any Q value can be measured when the bridge is set to read L and R. The resistance range extends from 0.01 ohm (first dial division) to one megohm and is independent of frequency.

Power Ratings

The maximum current and voltage that may be applied depend upon the value and power rating of, respectively, the bridge resistor (R_B) in series with the unknown and the other resistor (R_C) adjacent to the unknown. The resistance R_B should be as low as possible for low dissipation, and low compared to the reactance of the unknown inductor for minimum waveform distortion, but is also limited to reasonable values by sensitivity and lead-resistance considerations. In this bridge, R_B has a rating of 100 watts, but the power dissipation is

Figure 3. Panel view of the Incremental-Inductance Bridge.

limited to 50 watts for continuous operation to avoid overheating of the bridge elements. On the lowest three ranges, it has a value of one ohm, so that the rating for continuous use is 7 amperes. The values and ratings for the other ranges are tabulated in the specifications (see page 13).

The highest voltage that can be applied is determined by R_C. On the four highest inductance ranges it has a value of one megohm and permits 1250 volts to be applied to the bridge. On the lower ranges, where the impedance of the unknown inductor is lower, current rating is the more important, and the voltage decreases, becoming 12.5 volts on the lowest range.

For those applications requiring more than 7 amperes, the TYPE 1633-P1 Range-Extension Unit,* which contains a 0.1-ohm resistor, can be externally connected to shunt R_B on the three lowest bridge ranges; the inductance and resistance values are then reduced by a factor of 10. With this resistor, measurements up to 50 amperes, ac or dc, are possible.

The Internal Detector

Measurements on nonlinear impedances require the use of a highly selective detector because the large harmonic signals developed are not nulled at the fundamental balance. When magnetic circuits are highly saturated, it is possible to have harmonics as large as the fundamental itself and therefore approximately 40-db rejection would be necessary for a balance of 1%. Even more rejection is desirable.

The detector in this bridge was designed to meet these requirements. It is selective at the nine fixed frequencies at which the bridge Q scale is direct reading. Two cascaded, active, RC, selective amplifiers are used to obtain a second-harmonic rejection of 60 db. When measurements at other test frequencies are required, an external null detector should be used. The TYPE 1232-A Tuned Amplifier and Null Detector is recommended. This detector has a 35-db second-harmonic rejection, which is adequate for most measurements. However, if a given test frequency is to be used often, it is not difficult to change the detector and the bridge circuit to obtain selectivity and a direct Q reading at the desired frequency.

Due to high detector sensitivity and low noise, measurements can be made at excitation levels below one volt on the high inductance ranges and below 10 millivolts on the lowest range. Hence, on any range, the ac excitation can be varied over a span of more than 1200 to 1.

An additional detector feature is the high degree of amplitude compression, which makes repeated gain-control adjustments unnecessary.

THE COMPLETE SYSTEM

In the complete measuring system the ac supply, a dc supply, and the bridge are connected in series as shown in Figure 4. Descriptions of the two supplies now available are given elsewhere in this issue.

In Figure 4 the output circuit of each power supply is shown in greatly simplified form, to indicate the path of the high dc and ac excitation currents. Each supply is capable of passing the maximum current available from the other. The rms sum of the maximum dc and ac currents (5 amperes each) is equal to the

*Available on special order.

rated bridge current of 7 amperes. It should be noted that the maximum power rating of the bridge is 1250 volts at 7 amperes or 8750 voltamperes, far greater than the output of these 200-watt power supplies. Users who require power of this magnitude will presumably construct their own power supplies.

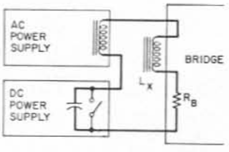

Figure 4. Simplified diagram of the system high-current circuit.

The ac voltage applied to the inductor should be as sinusoidal as possible. If the unknown inductor is nonlinear, the current will be distorted, and the series impedance in the loop must be low to avoid subtracting a large distorted voltage from the generator signal. Therefore, the output impedance of the Type 1266-A Adjustable AC Power Source and the ac impedance of the Type 1265-A Adjustable DC Power Supply have been made very low. Resistor R_B is also in this loop, but in most cases it is much smaller than the unknown impedance and is usually smaller than the winding resistance of the coil being measured. On the other hand, the dc applied should be from a constant current source. Current, rather than voltage, is the variable that should be controlled, because it determines the dc ampere turns (magnetic field intensity, H) applied. When the inductor dissipates a large amount of dc power, it will heat up, and its winding resistance will increase. The dc supply is therefore current regulated to keep the applied current constant.

Protective Devices

Because the bridge is designed to measure inductors in which the stored energy can be very large, a number of features have been incorporated to protect the equipment and to alert the operator to possible danger. The most prominent among these is a hinged cover over the UNKNOWN terminals. A lock on this cover is mechanically connected to a switch that shorts the generator terminals to ground (through a 1-ohm, 50-watt resistor). Thus, any current in the unknown inductor is discharged harmlessly before the operator disconnects the unknown. A panel light under the UNKNOWN terminals indicates when the generator is *not* shorted.

The SYSTEM POWER switch controls the power to two receptacles on the rear panel of the instrument, so that the bridge and its ac and dc generators can be conveniently controlled by the same switch. In this way the generators are also prevented from supplying power when the bridge power and warning light are turned off.

If the generator is disconnected with current flowing in the circuit, the induced voltage transient is applied directly across the bridge. Thyrite varistors have been included to limit this voltage and to help prevent damage to the bridge.

APPLICATIONS

The wide impedance, signal, and frequency ranges over which the Type 1633-A Incremental-Inductance Bridge is useful suggest many applications. The major use is undoubtedly the measurement of iron-core* components, but there are many other important applications in the measurement of nonlinear resistances.

*The term "iron-core" as used here includes all types of ferromagnetic cores.

Iron-Core Components

The ability of this bridge to operate with almost any generator waveform and at high power levels makes possible the measurement of an inductor, transformer, motor, or other electromagnetic device with the generator voltage either supplied by the circuit of the inductor or simulated by the bridge generators.

A dc power-supply filter choke provides a good illustration of this type of measurement. Since the resistors in the bridge circuit are arranged to cause very little extra loading of the generator, the bridge can simply be inserted in the leads of the choke in the power-supply circuit and be measured under the actual source and load conditions. Connections for a typical measurement are shown in Figure 5. Here the generator signal is a full-wave rectified sine wave. Since the sharply tuned detector discriminates against harmonics, the bridge can easily be balanced at the fundamental frequency of this waveform, which is twice the line frequency, or 120 cps. A plot of the measured inductance of a choke as a function of load current is shown in Figure 6. The actual measured points are indicated by crosses.

In order to provide a comparison, the generators of the TYPE 1630-AL Inductance Measuring Assembly were set up to measure the same choke with the same dc current and the same ac flux density (set on the ac supply) as in the above measurements. The rms fundamental component (120 cps) of the full-wave rectified sine wave is $\frac{2\sqrt{2}}{3\pi} E_{peak}$. To get the same flux density at 60 cps requires half this voltage. The points measured with the simulated generator are shown as circles in Figure 6. The difference between the two curves is very small except at low currents where the inductance is less than "critical,"[1] and the waveform in the full-wave case becomes distorted due to the discontinuous current in the choke. This measurement shows that the operating conditions can be duplicated without duplicating exactly the actual waveforms.

[1]Op. cit.

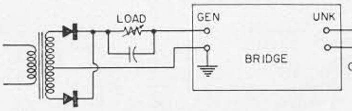

Figure 5. Circuit for measuring filter choke under load conditions.

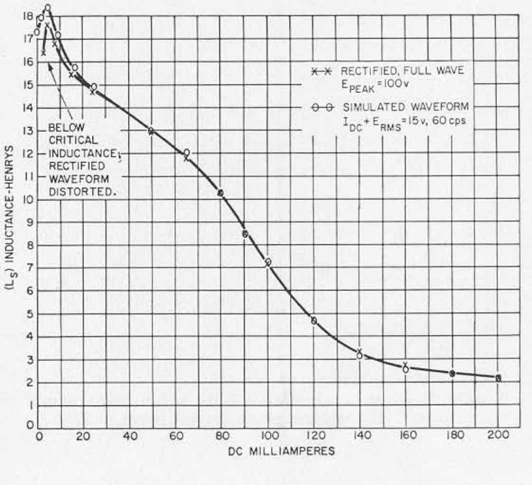

Figure 6. Plot of measured inductance of choke as a function of load current.

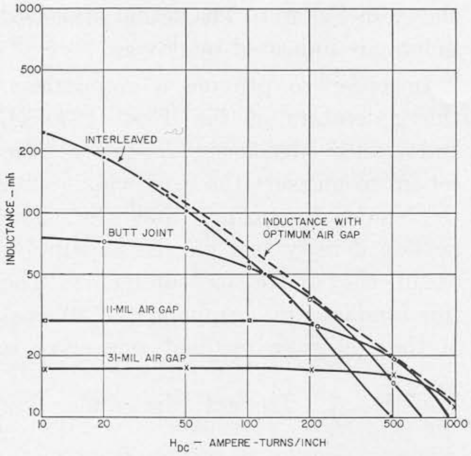

Figure 7. Inductance of choke as a function of dc amperes-turns with various air gaps.

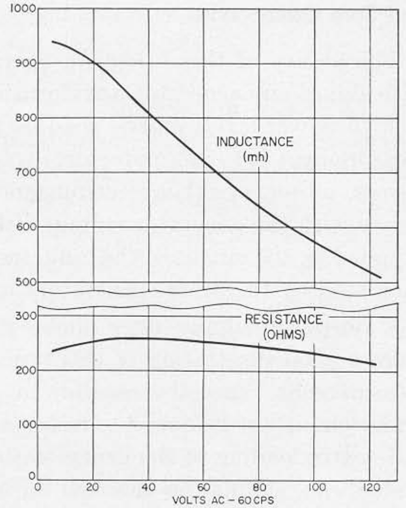

Figure 10. Measured inductance and resistance of the control winding of an ac servomotor.

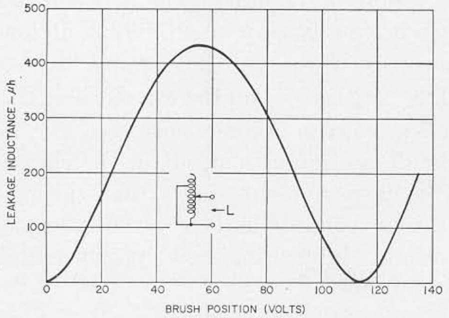

Figure 8. Leakage inductance of a Type W50 Variac Autotransformer at 50 amperes.

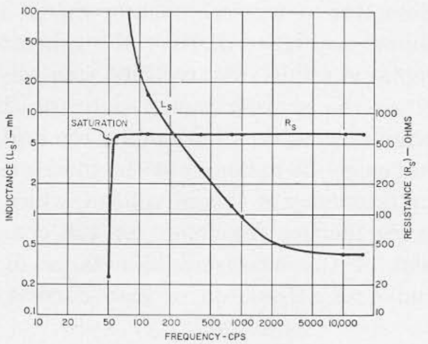

Figure 11. Input impedance of a loaded audio-frequency transformer.

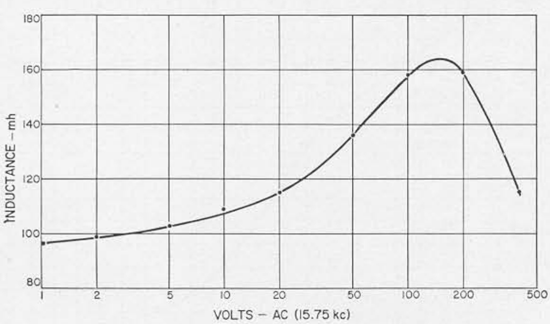

Figure 9. Inductance of a small pulse transformer as a function of ac signal level.

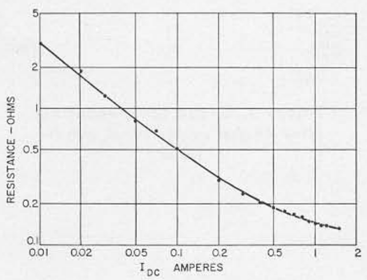

Figure 12. Zener diode: dynamic impedance.

A power-supply choke usually has an air gap in the magnetic circuit to prevent the core from saturating. With a given core, coil, and dc current, there is an optimum air gap for maximum inductance. A systematic procedure for the design of such inductors has been described by C. R. Hanna[3] and is widely used. The measurements required for such a design can easily be made with the TYPE 1630-AL assembly. An example showing the inductance of a coil *vs* direct current with several air-gap sizes is shown in Figure 7. The locus of the optimum-gap inductance values is shown by the dotted line. Figure 8 shows a measurement at high current. Data for this curve of the leakage inductance of a TYPE W50 VARIAC® Autotransformer were taken at 50 amperes, with the TYPE 1633-P1 Range-Extension Unit shunting the internal bridge resistor (R_B).

There are many measurements at low signal levels where the ability to set the signal to a known value independent of the balancing procedure is necessary. Figure 9 shows the inductance of a small pulse transformer measured at 15.75 kc as a function of ac signal level. This frequency position was included in the bridge frequency calibration for measurements on the magnetic components used in the horizontal circuits of television receivers.

Figure 10 illustrates another example showing the measured inductance and resistance of the control winding of a small ac servomotor. From such measurements it is a relatively simple matter to determine the value of series capacitors necessary to give any desired phase relationships at a particular operating point. It also indicates the load on the servoamplifier.

Another useful application of the bridge is in the measurement of loaded transformers of high power rating. Figure 11 shows the input-impedance components of an audio-frequency transformer terminated in its normal load.

AC Resistance Measurements

An important group of applications comprises ac resistance measurements on level-sensitive components. A typical example is the measurement of the dynamic impedance of Zener diodes, in which a small ac signal is superimposed on a dc bias current. For this measurement the bias current is supplied by the TYPE 1265-A DC Power Supply and the ac signal by the TYPE 1266-A AC Power Source. A dc voltmeter across the Zener diode measures the breakdown voltage at the test current. An example of this measurement is shown in Figure 12.

A similar measurement is that of the dynamic forward impedance of rectifiers. This information is useful in the voltage and regulation calculations in the design of power supplies. Figure 13 shows a dynamic resistance of a rectifier tube at various values of dc load current.

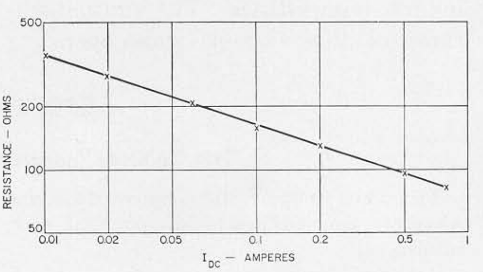

Figure 13.
Dynamic resistance of a vacuum-tube rectifier.

[3] C. R. Hanna, "Design of Reactances and Transformers Which Carry Direct Current," *Journal of AIEE*, February 1927, pp 5-8.

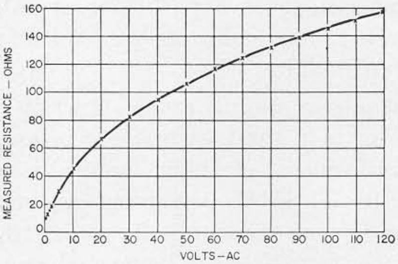

Figure 14. Dynamic resistance of a 100-watt lamp.

Lamp bulbs and thermistors are often used in controlled applications where their resistance is a function of the power applied. The peak current surge when incandescent lamps are turned on can be calculated from the knowledge of the cold resistance of the lamp. A plot of the resistance of a 100-watt, 110-volt lamp bulb is shown in Figure 14.

Figure 15 shows the dynamic inductance and resistance of a neon lamp. The negative resistance was measured by adding sufficient series resistance to make the sum positive. The effective inductance is due to the ionization time of the gas.

From these examples it is evident that the TYPE 1630-AL Inductance Measuring Assembly will satisfy most of the requirements for measurements on nonlinear components having resistive or inductive impedances. The unusual features of this system, wide operating ranges, easily set signal levels, and simplicity of operation, are intended to take the measurement of nonlinear components out of this special-measurements class and put them in the category of general-purpose measurements.

— R. G. FULKS
— H. P. HALL

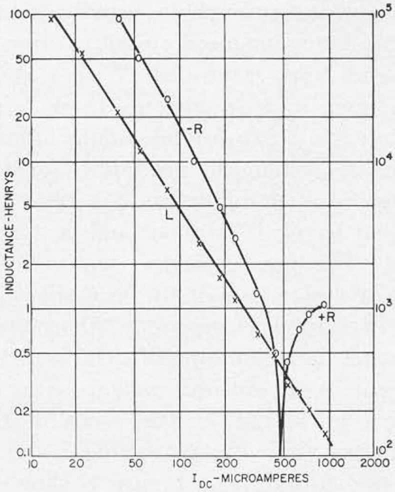

Figure 15. Dynamic inductance and resistance of a neon lamp.

CREDITS

The TYPE 1633-A Incremental-Inductance Bridge was developed by R. G. Fulks and H. P. Hall. R. A. Soderman, Administrative Engineer; H. C. Littlejohn, Design Engineer; J. E. Norton, Layout Draftsman; W. H. Higginbotham, Production Engineer; D. B. Bradshaw, Test Engineer, and S. P. Roberts, Engineer, have all contributed to the final design.

— Editor

SPECIFICATIONS

TYPE 1630-AL Inductance Measuring Assembly

The TYPE 1630-AL Inductance Measuring Assembly, mounted in a bench-type relay rack, consists of:

TYPE 1633-A Incremental-Inductance Bridge
TYPE 1265-A Adjustable DC Power Supply
TYPE 1266-A Adjustable AC Power Source

A connecting cable is supplied.

Type		*Code Word*	*Price*
1630-AL	**Inductance Measuring Assembly**...........	CANON	**$2300.00**

SPECIFICATIONS (Continued)

TYPE 1633-A Incremental-Inductance Bridge

	Frequency	Full-Scale Ranges a	b	c	d	e	f	Smallest Division
L	50c, 60c, 100c, 120c	10 mh	100 mh	1 h	10 h	100 h	1000 h	20 μh
L	400c, 800c, 1 kc	1 mh	10 mh	100 mh	1 h	10 h	100 h	2 μh
L	10 kc, 15.75 kc	100 μh	1 mh	10 mh	100 mh	1 h	10 h	0.2 μh
R	all	10 Ω	100 Ω	1 kΩ	10 kΩ	100 kΩ	1 MΩ	10 mΩ
Q	all	∞ — 1 Direct Reading at Above Frequencies						$Q = 1000$
	Max rms volts	12.5	125	1250	1250	1250	1250	
	Max rms amp*	7	7	7	2	0.7	0.2	

*Maximum rms current $= \sqrt{I_{dc}^2 + I_{ac}^2}$

ACCURACY

Inductance: $\pm 1\%$ of reading or 0.1% of full scale, $\pm \left(\frac{2\pi}{100} \times \frac{f_{kc}}{Q_x}\right) \%$.

Resistance: $\pm 2\%$ of reading or 0.1% of full scale, $\pm \frac{Q_x f_{kc}}{2\pi} \%$.

$\frac{1}{Q}$: $\pm 2\%$ or 0.001.

INTERNAL DETECTOR

Frequency: Selective at any one of nine specific frequencies, accurate to $\pm 1\%$, 50, 60, 100, 120, 400, and 800 cps, and 1, 10, and 15.75 kc.

Response to Second Harmonic: Approximately 60 db below fundamental.

GENERAL

Power Input: 105 to 125 (or 210 to 250) volts, 50 to 60 cps; power consumption, approximately 6 watts.

Accessories Supplied: One TYPE CAP-22 3-wire Power Cord and spare fuses.

Accessories Required: Generator to cover desired ranges of frequency and power, and a source of dc bias current (if desired).

Accessories Available: TYPE 1265-A Adjustable DC Power Supply (200 watts); TYPE 1266-A Adjustable AC Power Source (200 voltamperes).

Mounting: Relay-rack panel in aluminum cabinet. End frames are supplied with bench models.

Dimensions: Bench model, width 19, height 12¾, depth 10¼ inches (485 by 325 by 260 mm), over-all; rack model, panel 19 by 12¼ inches (485 by 315 mm); depth behind panel, 8¾ inches (225 mm).

Net Weight: 31 pounds (14.5 kg).

Type		*Code Word*	*Price*
1633-AR	**Incremental-Inductance Bridge, Rack Mount**	ABYSS	**$925.00**
1633-AM	**Incremental-Inductance Bridge, Bench Mount**	AUGER	**925.00**

THE TYPE 1265-A ADJUSTABLE DC POWER SUPPLY

The characteristics required for dc power supply for use with the incremental-inductance bridge are somewhat specialized and are not met by available units of conventional design. Among them are wide ranges of current and voltage, an output circuit that will pass high alternating currents, and a choice of voltage or current regulation.

The TYPE 1265-A Adjustable DC Power Supply, a completely solid-state design, includes these features. The in-

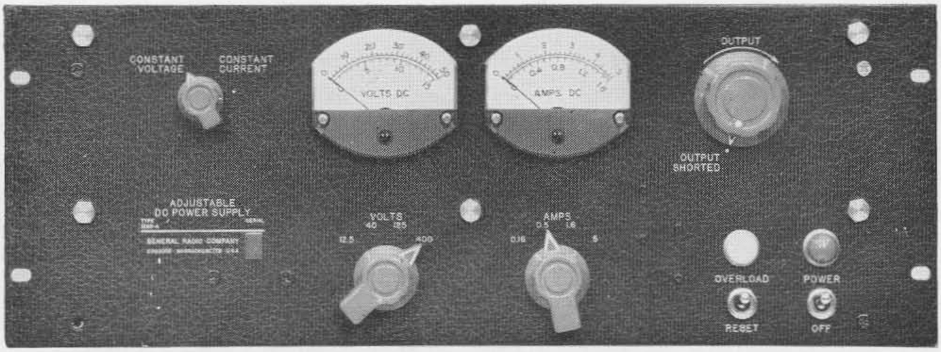

Figure 1. Panel view of the Adjustable DC Power Supply.

strument has four voltage ranges and four current ranges (see specifications below), and it will deliver its maximum rated power of 200 watts to not just one optimum resistance but to 8 ohms, 80 ohms, or 800 ohms.

A conventional regulated supply has a low ac output impedance obtained by large feedback. As a result, this impedance is low over only a limited dynamic range and only a relatively small ac current can be passed through the output. In applications where ac and dc must be combined to form a composite signal, a passive, low ac impedance is much more attractive.

The elementary diagram of the Type 1265-A is shown below. It shows the control loop used for regulation. Either the output voltage or current is sampled, amplified, and then used to control the conduction angle of two power-transistor trigger circuits used as controlled rectifiers. These rectifiers control the current into the output transformer whose several taps provide a choice of output voltages. The selected voltage is rectified and then filtered by components that are switched with both the voltage and current range adjustments in order to keep the output-circuit time constants independent of range. It should be noted that the output is sampled before the filter. Except for losses in the filter, which are compensated for, the dc voltage and current are the same on either side of the filter. Putting the sampling elements before the filter avoids the impossible stability problem that would arise from having the filter and the load (which could have any impedance) in the control loop.

Other important features are voltage and current metering, a mechanical connection between switches that prevents switching to range combination over 200 watts, and a trigger circuit that prevents damage to the instrument from overloads.

— H. P. Hall

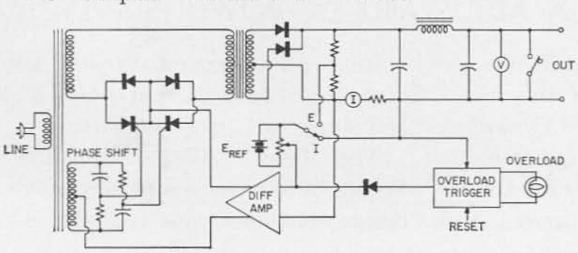

Figure 2. Simplified schematic of the dc power supply.

SPECIFICATIONS

Full-Scale Output Ranges: 12.5, 40, 125, 400 volts, dc; 0.16, 0.5, 1.6, 5 amperes, dc; in any combination up to 200 watts.

Meters: Voltage and current; ranges are switched with output ranges.

Overload Protection: Overload circuit trips at approximately 1½ times full-scale current.

Regulation: (Voltage or current) 0.2% for 20% of line-voltage change; 1% for 100% load change.

Speed of Response: Approximately 0.1 second.

Hum Level (rms): Approximately 70 db below full scale dc output (60 db on 12.5-volt, 5-ampere range).

Accessories Supplied: Type CAP-22 3-Wire Power Cord and spare fuses.

Dimensions: Bench model, width 19, height 7½, depth 17¼ inches (485 by 190 by 440 mm), over-all; rack model, panel 19 by 7 inches (485 by 180 mm), depth behind panel, 15 inches (385 mm).

Net Weight: 70 pounds (32 kg).

Type		*Code Word*	*Price*
1265-AR	Adjustable DC Power Supply, Rack Mount . .	ABASE	$875.00
1265-AM	Adjustable DC Power Supply, Bench Mount .	BAIZE	875.00

THE TYPE 1266-A ADJUSTABLE AC POWER SOURCE

The Type 1633-A Incremental-Inductance Bridge requires relatively large amounts of alternating and direct-current power and the bridge and power sources must tolerate alternating and direct currents, simultaneously.

The Type 1266-A Adjustable AC Power Source has been designed to meet these specific requirements. Six voltage ranges and five current ranges are selected by rotary panel switches, mechanically interlocked to interdict any combination that might exceed the 200-voltampere capacity of the supply. Voltage, in each range, is continuously

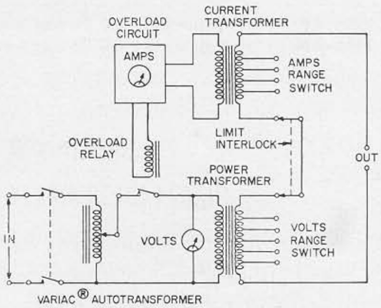

Figure 2. Simplified schematic of the ac power source.

adjustable from zero to the maximum value selected, by means of a VARIAC® adjustable autotransformer. Meters are

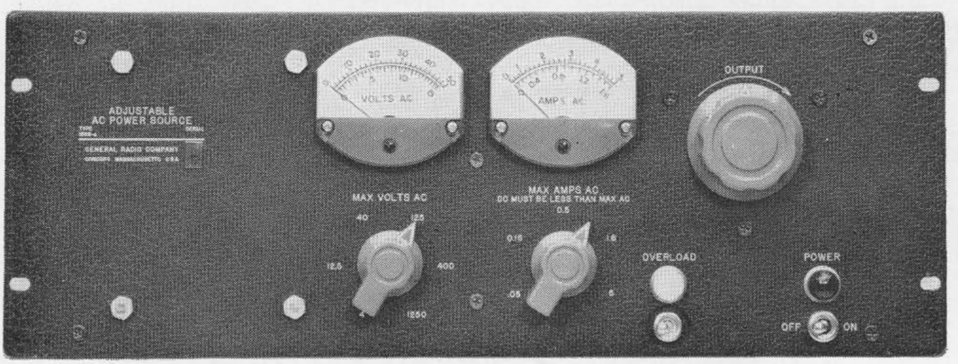

Figure 1. Panel view of the Adjustable AC Power Source.

provided for both voltage and current. An automatic trip circuit, with manual reset, protects the source against unintentional overload. The transformers are designed to tolerate a direct current at least as great as the maximum alternating current for the range selected.

These characteristics will be found useful wherever an adjustable 200-voltampere, 60-cycle power source is required. — GILBERT SMILEY

SPECIFICATIONS

Frequency: Power-line frequency (50 to 60 cps).

Full-Scale Output Ranges: 4, 12.5, 40, 125, 400, 1250 volts, rms; 0.05, 0.16, 0.5, 1.6, 5 amperes; in any combination up to 200 voltamperes.

Dc currents up to the rated ac current may be superimposed on output from external source.

Meters: Voltage and current; ranges are switched with output ranges.

Overload Protection: Overload circuit trips at approximately 1½ times full scale of current meters; can be reset by panel switch.

Accessories Supplied: TYPE CAP-22 3-Wire Power Cord and spare fuses.

Dimensions: Bench model, width 19, height 7½, depth 17¼ inches (485 by 190 by 440 mm), over-all; rack model, panel, 19 by 7 inches (485 by 180 mm), depth behind panel, 15 inches (385 mm).

Net Weight: 46 pounds (21 kg).

Type		*Code Word*	*Price*
1266-AR	Adjustable AC Power Source, Rack Mount...	CANDY	$345.00
1266-AM	Adjustable AC Power Source, Bench Mount.	BAFFY	360.00

SEMINAR ON HIGH-SPEED PHOTOGRAPHY TECHNIQUES AT M.I.T.

The scientific and engineering uses of high-speed photographic measurement techniques will be the subject of a one-week seminar at the Stroboscopic Light Laboratory, Massachusetts Institute of Technology, starting Monday, July 16.

Both theory and laboratory practice will be covered.

Subjects include pulsed stroboscopic lighting, optical high-speed cameras, Kerr cells, Faraday shutters, image converters, etc.

The program is under the direction of Professor Harold E. Edgerton of the Department of Electrical Engineering at M.I.T.

For further information inquire from the Office of the Summer Session, Room 7-103, M.I.T., Cambridge 39, Massachusetts.

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VLF STANDARD-FREQUENCY CALIBRATION

The initiation of standard-frequency broadcasts in the vlf band has made possible an improvement in the precision of calibration of stable oscillators used as local frequency standards which approaches a factor of 100 over that achievable by other means. With high-frequency transmissions, the best precision is about $\pm 1 \times 10^{-9}$. This precision is generally attainable only by the use of standard-time broadcasts and accurate time-comparison devices. If the high-frequency carrier of the standard-frequency station is used instead of the time pulses, the precision deteriorates slightly, but the accuracy suffers from propagation variations, so that the overall performance of any one calibration can seldom be guaranteed to a figure better than $\pm 1 \times 10^{-7}$, and often this figure is optimistic. Repeated calibrations, and cross-checking of time-tick and carrier-frequency calibration methods, suffice to allow some degree of assurance in local frequency-standard calibration, but that laboratory is indeed fortunate which has never been asked to "guarantee" a frequency calibration obtained from the high-frequency broadcasts to better than $\pm 1 \times 10^{-9}$.

The advent of vlf standard-frequency broadcasts, coinciding approximately with the proving-in of atomic frequency-standard devices, has made available both precision and accuracy of calibration which approach $\pm 1 \times 10^{-11}$ under ideal conditions and $\pm 1 \times 10^{-10}$ most of the time. At present, the vlf standard-frequency broadcast stations are maintained within $\pm 1 \times 10^{-10}$ of their nominal frequencies under normal circumstances, and calibration corrections are available from the operating agencies which include data to $\pm 1 \times 10^{-11}$.

It is difficult to justify any attempt to obtain increased precision of calibration beyond $\pm 1 \times 10^{-11}$ under present conditions. The propagation variations and the difficulty in maintaining perfect stability at the transmitting stations impose a practical limit on the performance of a calibration system based on vlf transmission over long ranges. For increased resolution and precision of frequency comparison, therefore, it is still desirable to have both oscillators being compared in the same location, or at least not widely separated. The $\pm 1 \times 10^{-10}$ guarantee on the accuracy of the signals as broadcast is a further

EXPERIMENTAL RESULTS OF FOUR METHODS OF USING VLF TRANSMISSIONS

deterrent to any attempts to claim greater accuracy of calibration. The various possible methods of frequency comparison using vlf transmissions have been described by J. A. Pierce.[1] The methods used by the present author were either taken directly from this publication, or derived from it indirectly. Several methods have been tried, with the results indicated below.

In order to use vlf transmissions for frequency-standard calibration, it is necessary to make use of the *phase* of the carrier wave of the standard-frequency signal. The vlf signal supplies a phase, or time, reference which suffers only a slight variation in propagation time compared with that of an hf signal. Hence, the phase difference between the stable oscillator to be calibrated and the received vlf signal is measured, and the frequency difference determined by conversion of the phase change per unit time, $\frac{d\phi}{dt}$, into the appropriate units. For example, there are 86,400 seconds in a day, or, in round numbers, approximately 10^5, corresponding to 10^{11} microseconds per day. Thus, a relative frequency difference of $\pm 1 \times 10^{-11}$ corresponds to a change of phase of only 1 microsecond in 24 hours, while $\pm 1 \times 10^{-10}$ frequency difference corresponds to $\pm$ 10 microseconds. In simpler terms, if the local phase changes +10 microseconds the first day, and +10 microseconds the second day also, the oscillator is maintaining a constant frequency which differs from that of the vlf signal by approximately $\pm 1 \times 10^{-10}$. If, however, the frequency of the oscillator is drifting by, say, 1×10^{-11} per day, the phase difference will change by 1 microsecond in each 24 hours, and, for example, will be 10 microseconds the first day, 11 the second. Over a longer period, a uniform frequency drift rate will produce an increase in phase proportional to the square of the elapsed time.

The first, and simplest, method tried was an experiment to assess the usefulness of the signal provided by NBA on 18.0 kc. The general arrangement was as shown in Figure 1. In this method, the signal is heterodyned to 2 kc by the injection of 20 kc into the second rf stage of the tunable vlf receiver. The 2-kc beat is then selected by a fixed-tuned filter, amplified, and applied to a circular-sweep oscilloscope driven by the frequency standard being calibrated. In this case, since the frequency of the heterodyning local oscillator is higher than the 18.0 kc signal, an increase in the local-oscillator frequency results in an increase in the beat frequency, resulting in a clockwise rotation of the pattern on the oscilloscope. The time for a one-cycle change in phase gives the frequency difference by simple calculation. The sense of the difference is obtained from the direction of rotation of the pattern. If the local standard is set to reduce the rate of rotation to a very

[1] "Intercontinental Frequency Comparison by VLF Radio Transmission," *Proc. IRE*, June, 1957 (pp. 794-803).

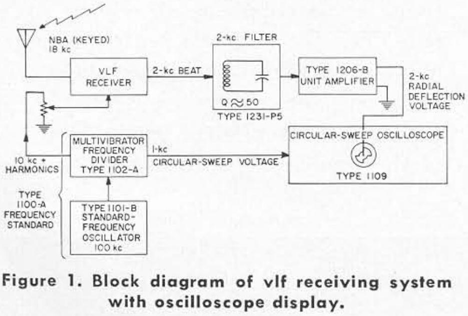
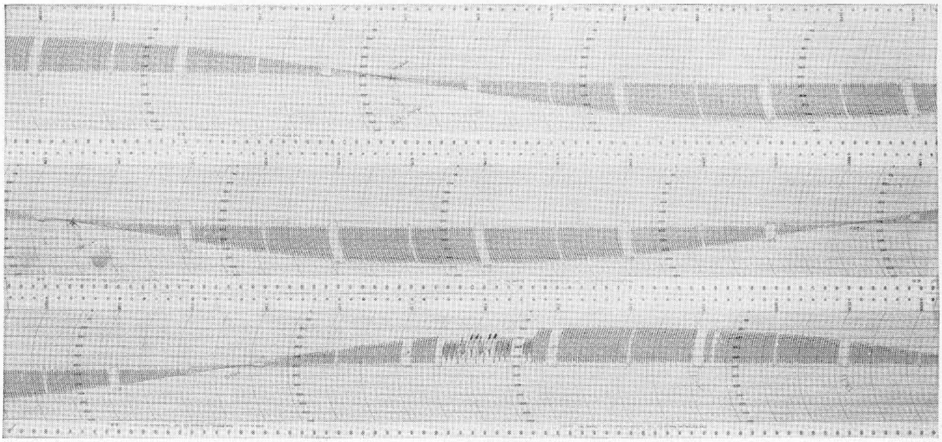


Figure 1. Block diagram of vlf receiving system with oscilloscope display.



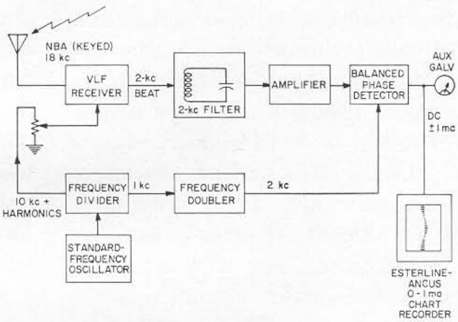


Figure 2. (Right) Block diagram of vlf receiving system using chart recorder on output of phase detector. (Above) A fragment of a recording produced by this system. Frequency difference is measured by the time between the zero crossings on the chart.

slow rate, the phase difference may then be noted and recorded. Unfortunately, for the greatest precision, this method requires rather frequent attention to the phase of the display and is thus somewhat tedious. In addition, it is not practicable to obtain reliable overnight calibrations by this means unless continuous observation of the indicator 'scope is provided, as will be apparent from a study of the typical photographic phase records presented later (Figure 6).

The second method of calibration used experimentally in the GR laboratories is shown in Figure 2. The current from a balanced phase detector was applied to a chart recorder set for zero-center deflection. In view of the small deflection in the vicinity of the zero crossings, the recorder was supplemented with the galvanometer indicator to help estimate the exact moment of zero current in the output circuit of the phase detector. Although somewhat more convenient than the first method described, this still required some manual operations and generated yards of recording tape, which required calibration. The timing pulses radiated by NBA provided an automatic time scale. Alternatively, the frequency of the local standard oscillator could have been set near zero beat with the incoming signal, but the sense of the frequency difference would then not have been immediately apparent; hence a better arrangement was desirable.

The phase-locked oscillator shown in Figure 3 was a logical extension of the

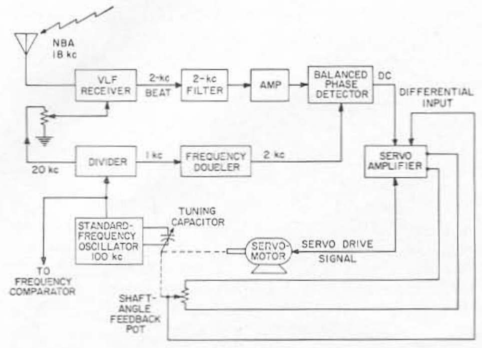


Figure 3. Block diagram of a motor-driven phase-locking system.

arrangement shown in Figure 2. A servo amplifier having a pair of differential input terminals was used to drive a servomotor which changed the frequency of the local standard-frequency oscillator by means of a variable capacitor. For a phase-locked oscillator system, it was necessary to provide one differential input signal to the servo amplifier from the phase-detector output and another from a potentiometer on the capacitor shaft. This system is an electromechanical equivalent of a voltage-variable reactance. The frequency of the local standard oscillator was thus phase-locked to that of NBA. The frequency of this oscillator was then recorded on the frequency comparator used for intercomparison of crystal oscillators in the GR laboratories. In this way, the frequency of NBA was made available as a reference for calibration of other oscillators. The precision attained by this device was approximately $\pm 2 \times 10^{-9}$ at best, and, under conditions of high noise level or sudden ionospheric disturbances (SID), the results were somewhat difficult to interpret. A sample recording is shown in Figure 4. The principal virtues of this system were its completely automatic operation and the direct intercomparison with other locally operated oscillators.

The phase-locking technique illustrated here is not so satisfactory a frequency calibration system as is the servo-driven phase-tracking receiver system originally described by Pierce[1], and now offered in many commercial models. The phase-locked oscillator, however, permitted direct comparison of the frequency of the NBA signal with other local signals without the construction of special phase-tracking equipment. The phase-locked oscillator was put together from already available components.

After several months of experience with the automatic recorder, it became obvious that the daily frequency shifts at dawn and dusk changed in nature, depending on the time of year, and that, even during daylight hours, some large fluctuations seemed to occur. Further investigation seemed indicated if more accuracy was to be obtained from the

[1] Loc. cit.

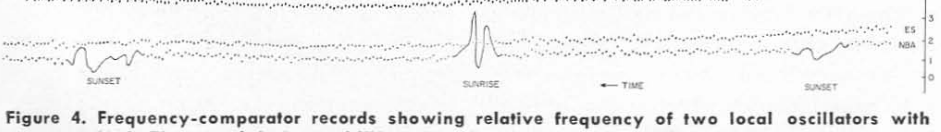
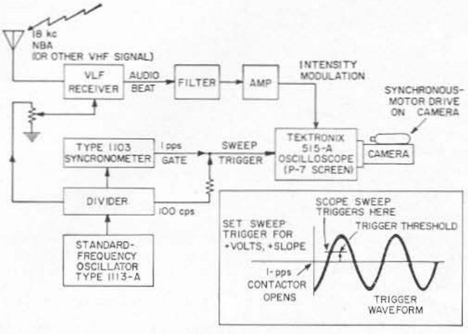


Figure 4. Frequency-comparator records showing relative frequency of two local oscillators with respect to NBA. The record designated WS is that of GR's working standard; ES represents an experimental standard; while NBA is a record of the oscillator phase-locked to NBA. The records show the frequency difference between each of these and a common reference oscillator. The units for the scale at the right are parts in 10^8. The total span of the recorder chart is 5.5 parts in 10^8.



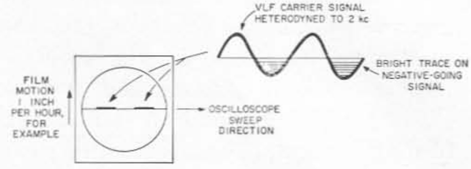


Figure 5. (Left) Block diagram of receiving system using a photographic phase recorder. Inset shows wave form of gated sweep trigger. (Above) Diagram showing mechanism of intensity-modulated oscilloscope display for photographic recording.

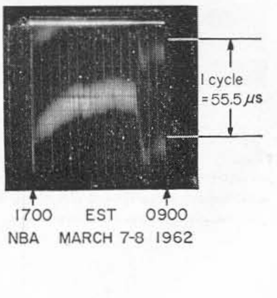
vlf transmissions. Accordingly, a photographic technique was devised, after the manner described by Pierce[1].

The photographic recording system is shown in Figure 5. The oscilloscope sweep is triggered either by means of the gated 100-cycle sweep trigger shown or by the ungated 100-cycle signal. The 1-per-second sweep is used with NBA to avoid sweeping when the signal is not present. The 100-sweeps-per-second arrangement is used to record communications-keyed signals.

The use of a modified oscilloscope camera with a synchronous-motor drive on the film holder provides a number of different sweep rates (in inches per hour of film drive) for the recording film, which is exposed one frame at a time in a Polaroid film-back. The vlf signal intensity modulates the trace on the oscilloscope tube, the sweep speed on the oscilloscope being set to show part of two consecutive carrier-frequency cycles. The sweep-time calibration is thus derived from the spacing between the consecutive cycles rather than from a separate oscilloscope calibration. The system can record signals with relatively poor signal-to-noise ratio, since the use of the P-7 screen and photographic recording seems to produce approximately 20-db improvement in apparent signal-to-noise ratio. It is probable that the photographic method alone accounts for most of this improvement, and that almost any phosphor can be used.

Photographic records made by the method of Figure 5 are reproduced in Figure 6a, b, and c. Minor fluctuations in phase apparently occur quite frequently, even almost continuously. It is not immediately apparent, however, that these fluctuations are anything that can be separated from the effects of impulse noise, such as atmospheric static or "sferics." In other words, this recording system records the noise along with the signal phase. One then makes frequency calibrations by calculating the indicated phase change per unit time and converting this figure to a frequency difference or deviation from the received standard frequency. Diurnal phase shifts are readily observed and calibrated, as shown by the record of Figure 6a. Small instabilities in either the received signal phase or the

Figure 6a. Night-time phase record of NBA showing diurnal phase shifts.



local standard oscillator are easily seen and are free from the effects of the complex time constants in the electromechanical systems described above.

One particularly important advantage of the photographic calibration method is illustrated in Figure 6b which is a record of two signals on one frequency, which were of different phase as received, and which were keyed alternately. Both signal phases are easily discerned in the photographic record. When this same set of signals was applied to the servo-driven phase-locked oscillator, it resulted only in continuous hunting of the servo system.

It is of interest to note the characteristics of some of the signals recorded. The signal from NPM over a west-east path shows 5 stable phases during the sunrise shift (Figure 6c). The corresponding sunset shift is smoothed or blurred by the nature of the recombination process in the ionosphere after sunset, which results in a slower change in effective height than that observed at sunrise. As a simplified explanation, the stepwise variation in phase may be thought of as the result of a ray bouncing between ionosphere and earth, with two relatively stable ionospheric effective heights and a transition zone between them. The ray will not change its path length except at certain critical intervals during which one or more of the reflection points is in the transition region as the rotation of the earth causes the transition region to move along the path.

As is readily apparent, the photographic phase-recording method provides the possibility of phase comparison to within a very few microseconds under most conditions. The over-all performance of this system is such that it can provide, in a 24-hour interval, frequency calibration of a precision better than $\pm 1 \times 10^{-10}$, even approaching a few parts in 10^{-11}. More refined photographic techniques should be slightly better but not by so much as an order of magnitude, in view of the propagation fluctuations apparent even on these relatively unsophisticated photographs.

An inspection of these recordings indicates what the possibilities of the photographic method may provide in the way of calibration data, and serves to illustrate the excellence of simple calibration methods which are based on sound basic principles.

A few remarks about the nature of the vlf signals may be of interest to those contemplating exploration of the vlf spectrum. Most of the transmitters whose carrier frequencies are stabilized are used for communication. In the vlf band, this generally means Morse-code-keyed cw, although other types of

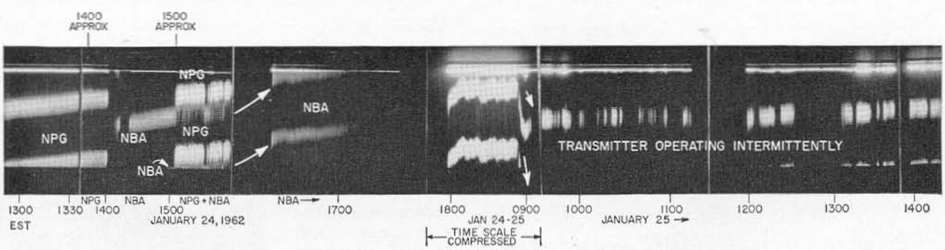


Figure 6b. Photographic record showing phase of signals from NPG and NBA. Note that recording shows phases of both NPG and NBA on January 24, when their transmitters were being keyed alternately. During the nighttime hours the time scale was compressed by slowing the film speed. Note the sudden phase shift at sunrise.

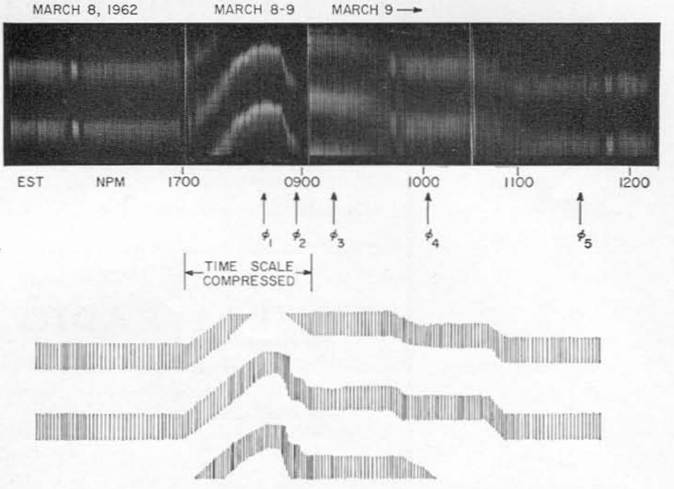


Figure 6c. Record of phase of NPM over a west-to-east transmission path. Note the five relatively stable phases of the signal during the sunrise transition.

modulation are occasionally used. The first vlf standard-frequency station, GBR, Rugby, England, is on 16.0 kc, and occasionally 19.6 kc. All the stations of the United States Naval Radio Service on vlf are used for communications except the NBA transmitter on 18.0 kc (Panama Canal Zone). The other stations are NSS, Annapolis, Maryland, 22.3 kc; NPG/NLK, Jim Creek, Washington, 18.6 kc; NPM, Lualualei, Hawaii, 19.8 kc; and NAA, Cutler, Maine, 14.7 kc. The National Bureau of Standards maintains WWVL on 20 kc and WWVB on 60 kc (Boulder, Colorado) as standard frequency transmitters, transmitting uninterrupted carrier signals with Morse-code-keyed call-letter identification every 20 minutes. The MSF transmitter (England) on 60 kc transmits continuous carrier signals during its scheduled operating times.

The NBA signal has an approximate radiated power of 30 kw, keyed with time signals, also call-letter identification and rated frequency offset from A1* frequency in Morse code. In addition, some silent periods and locked-key periods are observed. The strongest signals are radiated by NAA (approximately 1 megawatt radiated power) and NPG/NLK (250 kw), with NSS (100 kw) close behind, as is NPM (100 kw). The WWVL (20 kc) effective radiated power is estimated at 14 to 15 watts.

Receiving systems that depend on uninterrupted carrier reception for calibration are not generally successful when tried with the keyed-carrier signals. For example, extremely narrow-band filters exhibit keying transients which may obscure the desired phase information. If the signal is not above the noise level, the narrow-band filter has to be very carefully designed to avoid ringing, which may completely mask the desired signal. This is especially true in systems employing voltage-sensitive trigger circuits. In general, the servo-driven phase-tracking receivers and the photographic recording methods have proven more reliable, with the photographic method generally able to handle the widest variety of signals.

— F. D. Lewis

*A1 time is that time scale in which the transition frequency of cesium is measured as 9,192,631,770 cycles per second.

THE GENERAL RADIO

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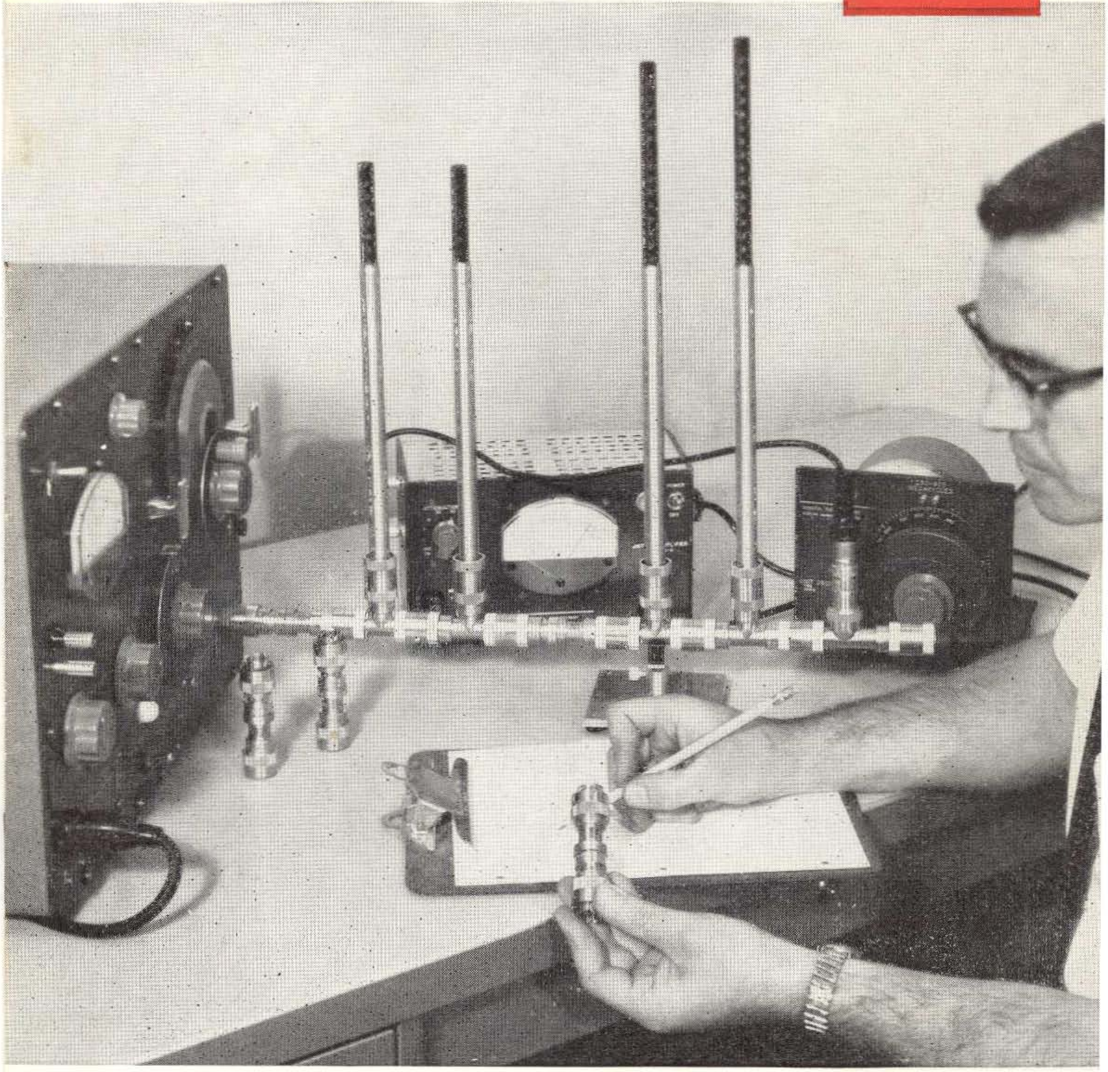

VOLUME 36 No. 7

JULY, 1962

IN THIS ISSUE

New —

- **Coaxial Elements**
- **Pink-Noise Filter**
- **Decade Capacitor**
- **Metered Variac® Autotransformer**

THE GENERAL RADIO

EXPERIMENTER

Published Monthly by the General Radio Company

VOLUME 36 • NUMBER 7 **JULY, 1962**

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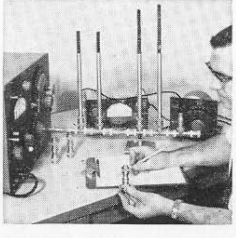

The assembly shown here is used for the calibration of coaxial attenuators. The excellent shielding of the locking Type 874 Connectors is particularly valuable in this measurement.

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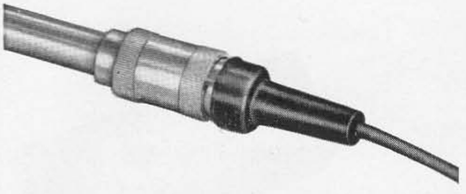
COAXIAL EQUIPMENT WITH LOCKING-TYPE CONNECTORS



The locking version[1] of General Radio's TYPE 874 Coaxial Connector* is now standard equipment on several instruments and coaxial elements. With components fitted with these new connectors, measurement setups can be assembled that possess an exceptionally high degree of mechanical rigidity and electrical stability. At the user's option, however, the locking feature can be disregarded, and the basic quick-connect/disconnect feature of the TYPE 874 Connector will permit rapid changes and substitution of parts. Further, since both the locking and the non-locking types are completely compatible, either can be plugged directly into the other.

Electrically, the voltage-standing-wave ratio is essentially unchanged from the low value that is characteristic of TYPE 874 Connectors. Typical figures for a pair of locking connectors are: <1.02 up to 3 Gc and <1.06 up to 7 Gc.

The shielding qualities of the locking connector produce an outstanding gain for measurement systems. Leakage is down approximately 50 db below that of the non-locking type.

Coaxial instruments now equipped with locking-type connectors are the TYPE 1602-B UHF Admittance Meter and the TYPE 874-LBA Slotted Line. In the admittance meter, only the detector connector, for mechanical reasons, remains a non-locking type.

The following coaxial elements are now available with either type of connector. Those with locking connectors are identified by the letter L added to the type number, as listed below.

[1]"New and Improved Coaxial Connectors," *General Radio Experimenter*, 35, 10, October, 1961.
*U.S. Patent No. 2,548,457.

Type		*Code Word*	*Price*
874-D20L	**20-cm Adjustable Stub**	COAXCOUGAR	**$16.00**
874-D50L	**50-cm Adjustable Stub**	COAXJAGUAR	**19.00**
874-EL-L	**Ell**	COAXYLLAMA	**10.50**
874-GAL	**Adjustable Attenuator**	COAXYHORSE	**67.00**
874-G3L	**Fixed Attenuator, 3 db**	COAXYBISON	**38.00**
874-G6L	**Fixed Attenuator, 6 db**	COAXBADGER	**32.00**
874-G10L	**Fixed Attenuator, 10 db**	COAXBEAVER	**32.00**
874-G20L	**Fixed Attenuator, 20 db**	COAXYCAMEL	**32.00**
874-LAL	**Adjustable Line**	COAXYTAPIR	**27.00**
874-LK10L	**Constant-Impedance Adjustable Line, 10 cm**	COAXYHIPPO	**42.00**
874-LK20L	**Constant-Impedance Adjustable Line, 22 cm**	COAXYRHINO	**42.00**
874-LTL	**Constant-Impedance Trombone Line, 44 cm**	COAXYMOOSE	**97.00**
874-MRL	**Mixer Rectifier**	COAXYOTTER	**34.50**
874-R22L	**Patch Cord**	COAXYFIXER	**10.00**
874-TL	**Tee**	COAXOCELOT	**14.00**
874-VCL	**Variable Capacitor**	COAXMONKEY	**61.00**
874-VQL	**Voltmeter Detector**	COAXYLEMUR	**32.00**
874-VRL	**Voltmeter Rectifier**	COAXAGOUTI	**32.00**

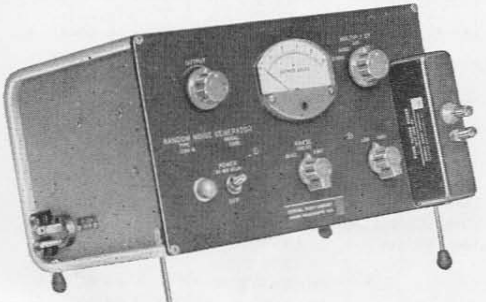
ROSE-COLORED GLASSES FOR WHITE NOISE

Broad-band electrical noise, often referred to as random noise, has proved to be a remarkably useful test signal when supplied by a controlled generator such as the TYPE 1390-B Random-Noise Generator.[1] Such a signal, embracing a wide range of frequencies and having a randomly varying instantaneous amplitude, closely approximates the signals normally encountered in many busy communication systems.[2]

The output of such a generator is characterized by a uniform spectrum over the frequency band to which the instrument is set. When the spectrum is uniform over a broad band, i.e. 20 cps to 20 kc, the noise is frequently referred to as "white" (constant energy per cycle) in that particular band. If this noise output is analyzed with a constant-percentage-bandwidth analyzer, such as the TYPE 1554-A Sound and Vibration Analyzer,[3] the amplitude-frequency characteristic of the white noise appears to slope upward with increasing frequency. The bandwidth of the analyzer increases in direct proportion to the frequency to which the analyzer is tuned. Since noise voltage increases as the square root of the bandwidth, it can be seen that the voltage output of the generator increases by a factor of 1.4 (3 db) when the analyzer frequency is doubled, thus giving the amplitude-frequency characteristic a slope of 3 db per octave.

View of the Type 1390-P2 Pink-Noise Filter.

If the random-noise generator is used as a source for a system under test, and if the output of the system is to be analyzed by a constant-percentage-bandwidth analyzer, it is usually desirable to compensate for this sloping characteristic at the output of the generator. The resulting output is then known as "pink" noise,[4] that is, noise having constant energy per octave.



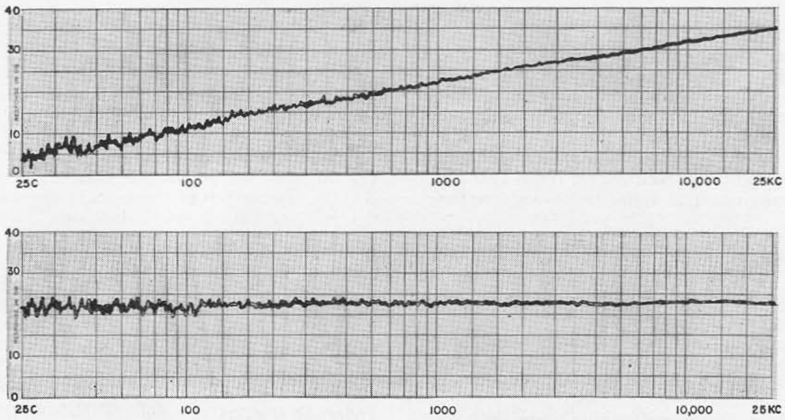
The Filter connects directly into the output terminals of the Random-Noise Generator, as shown.

[1] A. P. G. Peterson, "A New Generator of Random Electrical Noise," *General Radio Experimenter*, 34, 1, January, 1960.

[2] S. S. Stevens, J. P. Egan, and G. A. Miller, "Methods of Measuring Speech Spectra," *Journal of the Acoustical Society of America*, Vol 19, No. 5, September, 1947, pp 771-780.

[3] J. J. Faran, "A New Analyzer for Sound and Vibration," *General Radio Experimenter*, 33, 12, December, 1959.

[4] C. G. Mayo and D. G. Beadle, "Equipment for Acoustic Measurements (Part 4)," *Electronic Engineering*, Vol 23, December, 1951, pp 462-465.



(Upper curve) White noise output of the Type 1390-B Random-Noise Generator as measured by a one-third-octave bandwidth and (lower curve) pink noise output of the filter.

The Type 1390-P2 Pink-Noise Filter converts the electrical output of the random-noise generator from "white" noise to "pink" noise in the audio-frequency range. It is designed to plug into the output binding posts of the Type 1390-B Random-Noise Generator, but can also be used at any point in a system where such a filter characteristic is needed, provided that the source impedance is less than 1 kilohm and the load impedance is at least 20 kilohms. The input terminals are recessed plugs at the rear, and the output terminals are binding posts on the front. For shielding, the case of the filter is grounded to the low input and output terminals.

This RC low-pass filter has an amplitude-frequency characteristic of −3 db per octave from 20 cps to 20 kc. Beyond 20 kc the attenuation has been made 6 db per octave in order to reduce the unwanted frequencies outside the audio-frequency range.

Pink noise has been found to have a wide variety of applications. Some noises that occur in nature, such as the low-frequency noise in semiconductors and certain acoustical background noises, are closer in spectral characteristics to pink noise than to white noise. To simulate electrical signals generated in such cases, it is convenient to use pink noise.

Frequency-response measurements on electroacoustical and electromechanical equipment constitute the most important use. With sine-wave excitation, measurement results are difficult to interpret because of the large amplitude fluctuations that may occur. When the data are averaged over a narrow range of frequencies, the response curve is considerably smoother and much easier to use. "Warble tones" are often used for this purpose. A more convenient method, however, is to use pink noise as the tone source and the Type 1554-A Sound and Vibration Analyzer, with one-third-octave bandwidth, as the measuring system. The frequency-response characteristic can be automatically recorded with the Type 1521-A Graphic Level Recorder.[5]

—J. J. Faran

[5]M. C. Holtje and M. J. Fitzmorris, "A Graphic Level Recorder with High Sensitivity and Wide Ranges," *General Radio Experimenter*, 33, 6, June, 1959.

SPECIFICATIONS

Frequency Response: Sloping at the rate of −3 db per octave from 20 cps to 20 kc. Sloping at the rate of −6 db per octave at all higher frequencies. Output voltage is approximately −5 db with respect to the input voltage at 20 cps and −35 db at 20 kc. It lies within 1 db of the straight line connecting these two points on a graph of output in decibels *vs* log frequency.

Attenuation-frequency characteristic of the Pink-Noise Filter.

Over-all Output Level: When the filter is used with the Random-Noise Generator, set for the 20-kc range, the output voltage of the filter is approximately 30 db below its input, and the voltage level in each one-third-octave band is approximately 17 db below that. Thus, when the output meter of the generator indicates 3 volts, the output of the filter is approximately 0.1 volt, and the level in each one-third-octave band is approximately 15 millivolts.

Input Impedance: The filter should be driven from a source whose impedance is 1 kilohm or less. Input impedance is variable from 6.5 kilohms + load resistance at zero frequency to 6.7 kilohms at high frequencies.

Output Impedance: The filter should not be operated into a load of less than 20 kilohms. Internal output impedance is variable from 6.5 kilohms + source resistance at low frequencies to approximately 200 ohms at high frequencies.

Input Voltage: 15 volts rms, maximum.

Terminals: Input terminals are recessed banana pins on ¾-inch spacing at rear of unit. Output terminals are jack-top binding posts with ¾-inch spacing.

Dimensions: Width 1⅜, height 5, depth 2⅞ inches (35 by 127 by 73 mm), over-all.

Net Weight: 5½ ounces (155 g).

Type		*Code Word*	*Price*
1390-P2	**Pink-Noise Filter**	FATAL	**$45.00**

A 10-MICROFARAD DECADE CAPACITOR

Following the successful introduction of the TYPE 1424-A Standard Polystyrene Decade Capacitor in June 1961,[1] it appeared that many users of a 10-microfarad-total decade might not need its sophisticated performance specifications. Accordingly, another version of this decade capacitor, the TYPE 1424-M, is being introduced, having less rigorous performance specifications and being appreciably lower in price. It is made from twenty ½-microfarad sealed foil-

[1]P. K. McElroy, "A New 10-Microfarad Capacitance Standard," *General Radio Experimenter*, 35, 6, June, 1961.

paper capacitors, of noninductive or extended-foil construction, employing a viscous impregnant to improve stability with time and position as compared to older types of paper units. Despite the difference in form of the capacitors, the effective internal inductance is about the same as that of the TYPE 1424-A, and the natural period of the capacitor with the terminals shorted is also essentially the same.

In appearance the TYPE 1424-M Decade Capacitor resembles the TYPE 1424-A; dimensions are the same except for depth, which is 1¾ inches less.

— P. K. McELROY

SPECIFICATIONS

Nominal Value: 0 to 10 microfarads, in steps of 1 microfarad.

Adjustment Accuracy: ±1% at 1 kc.

Stability: Change is less than ±0.35% per year.

Frequency: Calibrated at 1 kc. Variation with frequency down to 60 cps is typically less than +0.7%. At higher frequencies, terminal capacitance rises as resonant frequency, f_o, is approached (see curves). The increase can be calculated from $\frac{\Delta C}{C} = \left(\frac{f}{f_o}\right)^2$. f_o varies from approximately 570 kc at 1 μf down to 240 kc at 10 μf.

Voltage Recovery: Less than 5%, final, of original charging voltage after a charging period of 1 hour, and a 10-second discharge through a resistance equal to one ohm per volt of charging.

Dissipation Factor: Less than 0.005 at 1 kc. (See curves for variation with frequency.)

Temperature Coefficient: Approximately +180 ppm per degree C.

Maximum Operating Temperature: 90 C.

Insulation Resistance: Greater than ten thousand ohm-farads.

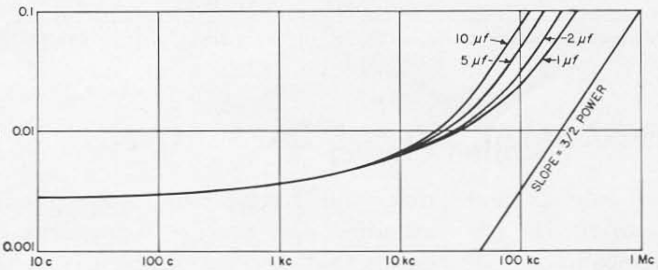
Maximum Voltage: 500 volts peak, up to 2 kc.

Mounting: Aluminum cabinet and panel, finished in gray.

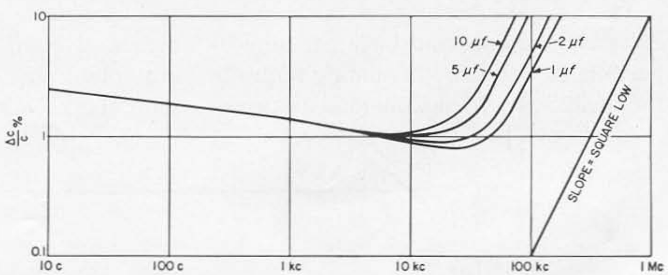
Terminals: A separate ground terminal is provided, permitting 2- or 3-terminal use.

Dimensions: Width 8, height 6, depth 9½ inches (205 by 150 by 240 mm), over-all.

Net Weight: 7¾ pounds (3.5 kg).



(Left) Variation in dissipation factor with frequency.



(Right) Percentage change in capacitance with frequency.

Type		*Code Word*	*Price*
1424-M	**Decade Capacitor**	REFER	**$195.00**

NEW VARIAC® AUTOTRANSFORMER WITH BUILT-IN VOLTMETER

This new VARIAC autotransformer consists of a TYPE W5 VARIAC Autotransformer and a voltmeter, mounted in a metal case, finished in hammertone gray, with switch, cord, plug, and convenient carrying handle. Line cord and output receptacle are 3-wire types. A double-pole on-off switch disconnects both sides of the input-line. The load circuit is protected by a circuit breaker.

SPECIFICATIONS

Input Volts: 120.

Frequency: 50 to 60 cps.

No-Load Loss: 9 watts.

Output Volts: 0 to 140.

Maximum Load Current: 5 amperes.

Voltmeter Range: 0 to 150 volts.

Terminals: Line, 3-wire cord and plug. Load, 3-wire output receptacle.

Dimensions: (TYPE W5 case) Width 4⅛, height 6⅝, depth 5½ inches (125 by 170 by 140 mm), over-all.

Net Weight: 8¼ pounds (3.8 kg).

Type		*Code Word*	*Price*
W5MT3VM	**Metered Variac® Autotransformer**..........	DANDY	**$54.00**

U.S. Patent No. 2,949,592.

Vacation Closing

During the weeks of July 23 and July 30, our Manufacturing Departments will be closed for vacation.

There will be business as usual in the Sales Engineering and Commercial Departments. Inquiries, including requests for technical and commercial information, will receive our usual prompt attention. Our Service Department requests that, because of absences in the manufacturing and repair groups, shipments of equipment to be repaired at our plant be scheduled to reach us after the vacation period.

General Radio Company

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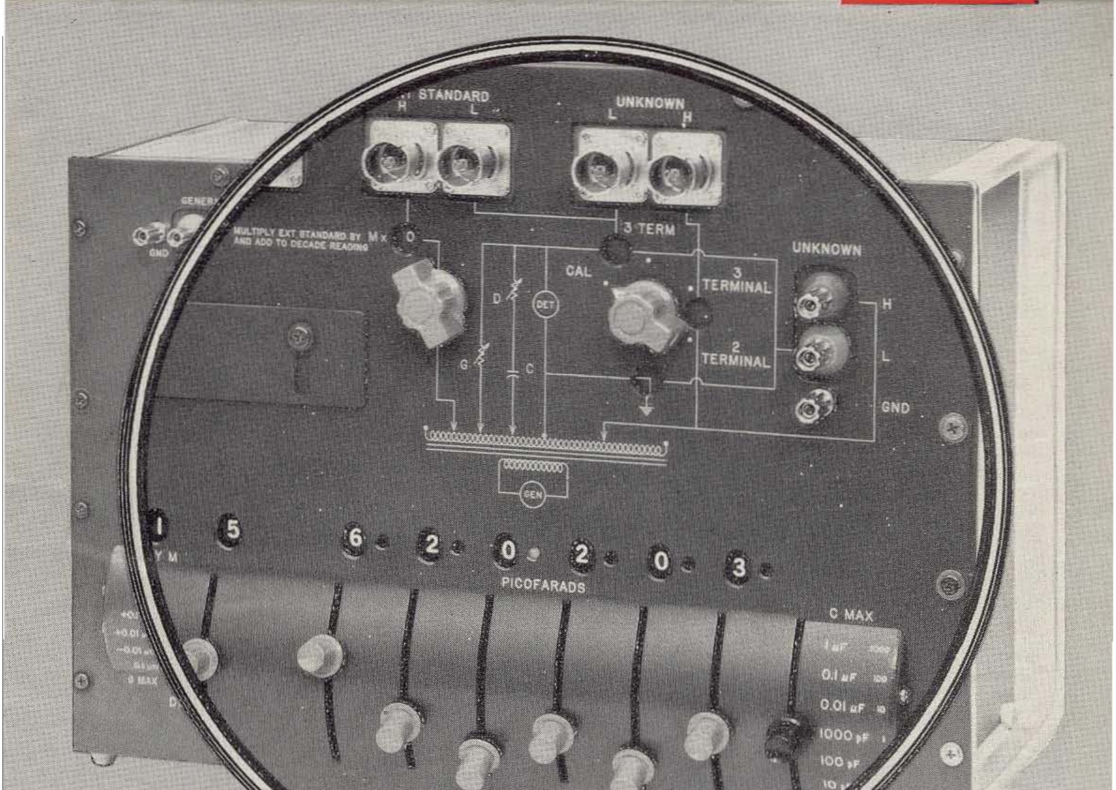
THE GENERAL RADIO

EXPERIMENTER



VOLUME 36 Nos. 8 & 9 AUGUST—SEPTEMBER, 1962

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- 1962 EIME Schedule

THE GENERAL RADIO

EXPERIMENTER

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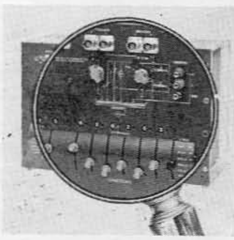

Lever-type switches and digital readout on the Type 1615-A Capacitance Bridge bring a new order of convenience to capacitance measurements. This new bridge has an accuracy of 0.01% and a resolution of one part in a million.

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GENERAL RADIO COMPANY (OVERSEAS), ZURICH, SWITZERLAND
REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

ACCURACY, PRECISION, AND CONVENIENCE FOR CAPACITANCE MEASUREMENTS

The direct-reading accuracy of 0.01% in the new Type 1615-A Capacitance Bridge here being introduced is an improvement of an order of magnitude in accuracy over most preceding bridges. This has been made possible by the improvements introduced in recent years in standards of absolute capacitance, in capacitance bridges, and in reference capacitors at the National Bureau of Standards[1] and other standards laboratories. The computable cross capacitor developed at the National Standards Laboratory of Australia by Thompson and Lampard[2] has made it possible to determine the unit of capacitance with an accuracy that is now a few parts per million and which may be expected to improve. The advantages of ratio transformers and of transformer ratio arms in bridges, exploited initially in Great Britain and then more fully in Australia, have been utilized to increase the precision of measurement and to extend the range to capacitances below 1 micro-picofarad (10^{-18} farad). Furthermore, the reference capacitors used to store and transfer the capacitance unit have

[1] M. C. McGregor, J. F. Hersh, R. D. Cutkosky, F. K. Harris, and F. R. Kotter, "New apparatus at the National Bureau of Standards for absolute capacitance measurement," *IRE Trans. on Instrumentation*, vol. I-7, pp 253–261; December, 1958.

[2] A. M. Thompson, and D. G. Lampard, "A New Theorem in Electrostatics and its Application to Calculable Standards of Capacitance," *Nature*, 177, 888 (1956).

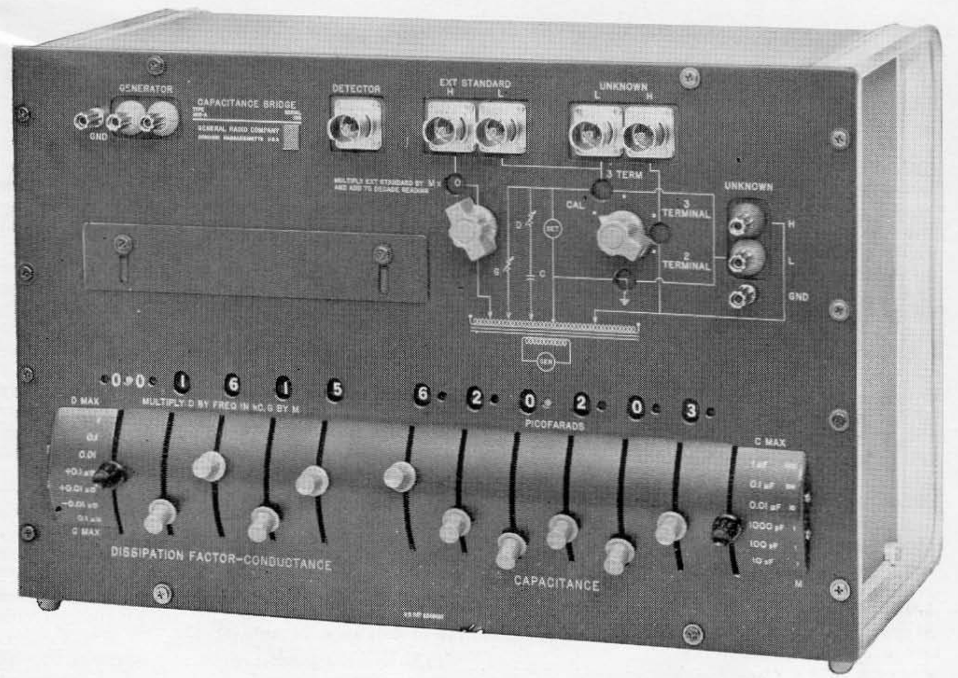


Figure 1. Panel view of the Type 1615-A Capacitance Bridge.

been improved by the use of three-terminal construction to minimize connection errors, which limit accuracy in capacitors of 1000 pf or less,[3] and by the use of new construction methods, low-temperature-coefficient materials, and sealed containers to increase the stability. With this new apparatus, the National Bureau of Standards can now certify capacitors to an accuracy of 50 ppm or better.

Many of these improvements have also been incorporated into the new TYPE 1615-A Capacitance Bridge. This bridge has transformer ratio arms for accuracy and stability. Its internal capacitance standards are three-terminal, sealed capacitors having low temperature coefficients. The bridge has six-figure resolution for capacitance from 1 μf to 1 pf and a direct-reading accuracy of 0.01% over this capacitance range and over most of the frequency range from 100 cycles to 10 kc. The impedance of the transformer ratio arms has been kept very low, so that accurate three-terminal measurements can be made even in the presence of large capacitances to ground. The bridge also has the necessary internal shielding to permit one terminal of the unknown to be grounded, so that both two-terminal and three-terminal measurements can be made over the whole capacitance range.

The balance controls are lever-type switches, the readout is digital, and the decimal point is automatically positioned.

These features, and others described below, result in a capacitance bridge that brings to the measurement of capacitance, to the intercomparison of standards, and to the measurement of dielectric properties an unusual degree of accuracy, precision, range, and convenience.

TRANSFORMER RATIO ARMS

Many of the advantages of inductively coupled, or transformer, ratio arms have been known since about 1888, and they are covered in detail in the 1928 British patent of A. D. Blumlein. Little use was made of them, however, until about the time of World War II, when new applications were found in the measurement of very small capacitance. Since that time, transformer ratio arms have become increasingly popular in commercial bridges as well as in the apparatus of the national standards laboratories.

The advantages of such ratio arms are that accuracies within a few parts per million are not difficult to obtain over a wide range of integral values, even for ratios as high as 1000 to 1, and that these ratios are almost unaffected by age, temperature, or voltage. The low impedance of the transformer ratio arm also makes it easy to measure direct impedances and to exclude the ground impedances in a three-terminal measurement without the use of guard circuits and auxiliary balances.

To illustrate these characteristics, a simple capacitance bridge with transformer ratio arms is shown in Figure 2. On the toroidal core, a primary winding, connected to the generator, serves only to excite the core; the number of primary turns, N_P, determines the load on the generator but does not influence the bridge network. If all the magnetic flux is confined to the core—as it is to a high degree in a symmetrically wound toroid with a high-permeability core—the ratio of the open-circuit voltages induced in the two secondary windings must be exactly equal to the ratio of the number of turns. The ratio can be changed by the use of taps along the two secondaries, but, when the number of turns is fixed, the voltage is highly invariant. Changes in the core permeability with time and temperature have only second-order effects on the ratio, because they modify only the very small amount of leakage flux that is not confined to the core in a practical transformer. The ratio is, therefore, both highly accurate and highly stable.

In Figure 2, the two transformer secondary windings are used as the ratio arms of the capacitance bridge with the standard capacitor, C_N, and the unknown, C_X, as the other two arms in a conventional four-arm bridge network. The condition for balance or zero detector current is easily shown to be that $V_N C_N = V_X C_X$ or $C_X/C_N = V_N/V_X = N_N/N_X$. This balance condition is not affected by the capacitances shown from the H and L terminals of C_N and C_X to the terminal G connected to

[3]J. F. Hersh, "A Close Look at Connection Errors in Capacitance Measurements," *General Radio Experimenter*, 33, 7, July, 1959.

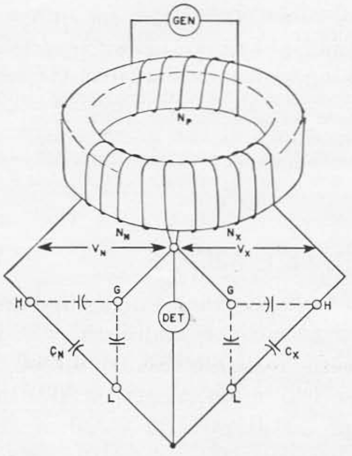


Figure 2. A capacitance bridge with transformer ratio arms.

the junction of the ratio arms. The capacitances between L and G shunt the detector, so that they affect only the bridge sensitivity. The capacitances between H and G are across the transformer windings. To the extent that the transformer can be assumed ideal, i.e., with no resistance in the secondary windings and with no flux that does not link equally both secondaries, the current drawn by the H-G capacitances does not change the voltages V_N and V_X or the balance conditions. In practice, the transformer resistances and leakage inductances can be kept so small that quite low impedances or large capacitances can be connected from H to G before there is appreciable error in the bridge.

The junction of the ratio arms, G, is therefore a guard point or guard potential in the bridge. All capacitances to G from the H or L corners of the bridge are excluded from the measurement. In the three-terminal capacitors represented by the H, L, G terminals in Figure 2, the bridge measures only the direct capacitance, C_X, of the unknown in terms of the direct capacitance, C_N, of a standard, without additional guard circuits or balances.

One can take advantage of the accurate and stable ratios of the transformer by the use in the bridge of a standard arm which is fixed and a ratio which can be varied to balance the bridge.

Figure 3 shows three of the possible ways of balancing a simple transformer-ratio capacitance bridge. For simplicity, the generator and primary are not shown, but it is assumed that the two secondaries have 100 turns each and are excited so that there is 1 volt per turn. The capacitor in the unknown arm is assumed to be 72 picofarads.

In Figure 3a, the two ratio arms are equal and the bridge is balanced in the conventional way with a variable standard capacitor which is adjusted to 72 pf.

The detector current can equally well be adjusted by a variation in the voltage applied to a fixed standard capacitor. In Figure 3b, the standard capacitor is fixed at 100 pf, and this is balanced against the 72-pf unknown connected to the 100-volt end of the transformer by connection of the standard to 72 volts of the opposite phase, obtained from suitable taps on the transformer windings. The inductive divider shown has a winding of 100 turns with taps every 10 turns and, on the same core, another winding of 10 turns tapped every turn. If, as shown, the second winding is connected to the 70-volt tap on the first winding and the capacitor to the 2-volt tap on the second winding, the required 72 volts is applied to the capacitor. Six or more decades for high precision can be obtained in a similar fashion with more turns on one core and the use of additional cores driven from the first. Such inductive dividers have very accurate and stable ratios, but the errors increase with the number of decades because of loading effects.

Another method of balance by voltage variation is shown in Figure 3c, where a single decade divider is used in combination with multiple fixed capacitors. The 100-turn secondary is tapped every 10 turns to provide 10-volt increments. If, then, a 100-pf capacitor is connected to the 70-volt tap and a 10-pf capacitor to the 20-volt tap, the resulting detector current balances that of the 72-pf unknown connected to 100 volts. This bridge can be given six-figure resolution, for example, through the use of six fixed capacitors in decade steps from 100 pf to 0.001 pf, each of which can be connected to any one of the taps on the transformer.

In any of these bridges, the bridge ratio can also be varied by use of taps on the unknown side of the transformer to vary the voltage applied to the unknown capacitor. For example, if the unknown capacitor were connected to a 10-turn or 10-volt tap on the upper half of the transformer, then a capacitance of 720 pf instead of 72 would be balanced by the standard capacitors shown. The range of the bridge can thus be extended to measure capacitors which are much larger than the standards in the bridge.

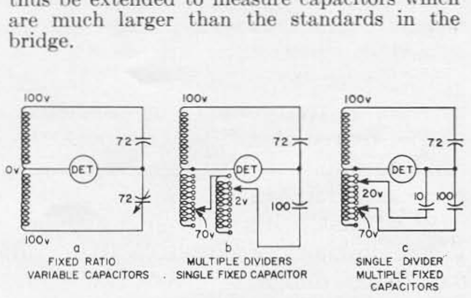


Figure 3. Methods of balancing capacitance in a transformer-ratio bridge.

These advantages of transformer ratio arms and dividers make possible a bridge of very wide range and high accuracy, since not only are the ratios stable and accurate but, when only a few fixed capacitors are required as standards, the standards can be constructed to have high stability and accuracy. This bridge can also have a wide range of frequencies. At low frequencies, a limit is imposed on sensitivity by the maximum voltage obtainable from the transformer, since, for a given core, the voltage at saturation is proportional to frequency. At high frequencies there is a decrease in accuracy resulting from the decrease in core permeability with frequency, from the increased loading of the transformer by its self-capacitance as well as the bridge capacitances and, of course, from the usual residual capacitances and inductances in the bridge wiring and components.

THE TYPE 1615-A CAPACITANCE BRIDGE

CAPACITANCE

The new Type 1615-A Capacitance Bridge is a transformer-ratio bridge of the type that uses a single decade of transformer voltage division and multiple, fixed, standard capacitors to provide six decades of resolution in capacitance. As shown in the elementary diagram of Figure 4, one side of the secondary of the ratio transformer is tapped at intervals of one-tenth, and to these taps can be connected six standard capacitors in any combination required to balance the bridge. If, for example, the standards connected to the six-decades switch are 1000, 100, 10, 1, 0.1, and 0.01 pf, the range of unknown that can be balanced is from 1000 pf to 0.001 pf when the unknown is connected to the full voltage of the other secondary of the transformer. This unknown side of the transformer has, however, a tap at one-tenth of the full voltage, so that when the unknown is driven from this lower voltage, the range is multiplied by ten, and an unknown up to 10,000 pf or 0.01 μf can be balanced by the same internal standards. The range is extended still further by further division of voltage on the unknown side through a second transformer or inductive divider driven from the 0.1 tap on the ratio transformer. This second divider provides additional ratios of 0.1 and 0.01, so that, with the voltage applied to the unknown reduced to 0.01 and 0.001, the bridge is given two more ranges of 0.1-μf and 1-μf maximum capacitance.

To extend the range to smaller capacitances, two additional standards are used, of 0.001 and 0.0001 pf. This yields two more ranges, 0.0001 pf to 100 pf and 0.00001 pf to 10 pf. There are, therefore, eight standard capacitors, only six of which are used for any one range. The connections of these capacitors are made by the same range switch that selects the transformer taps.

With this combination of eight internal standard capacitors and four voltage ratios to which the unknown can be connected, the capacitance range of the bridge extends from a maximum of 1.111,110 μf to a minimum step of 0.00001 pf or 10^{-11} μf. The capacitors and ratios used for each range are indicated in Figure 5.

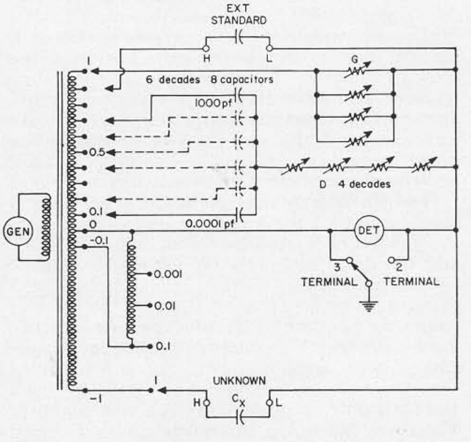


Figure 4. Elementary schematic diagram of the capacitance bridge.

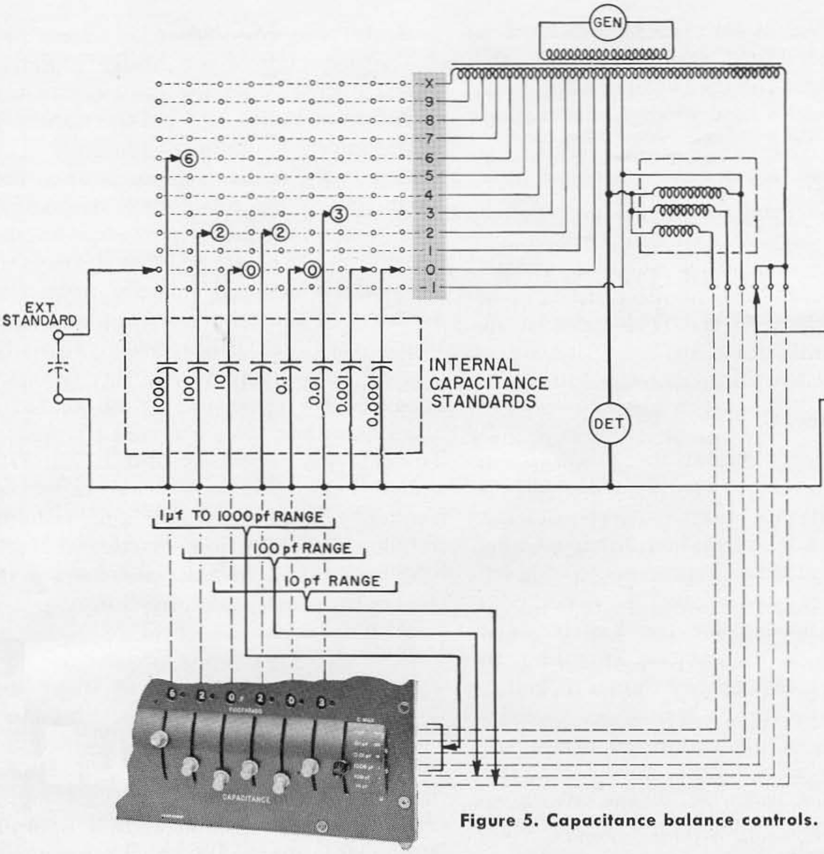


Figure 5. Capacitance balance controls.

LOSS

To obtain a precision of six figures in the capacitance balance, the loss balance must be made equally precise. As shown in Figure 4, the loss balance in this bridge can be made in terms of either the dissipation factor, *D*, or the shunt conductance, *G*, of the unknown. For most purposes, dissipation factor offers the greater range and convenience. Conductance is useful in some measurements of dielectric materials and is necessary when external standards are added to the bridge and when the loss in the bridge standards exceeds that of the capacitor being measured.

Dissipation Factor

The dissipation-factor balance is made by means of four resistance decades connected in series with the common side of all the internal capacitance standards as shown in Figure 4. Since $D = \omega RC_T$, where C_T is the total capacitance connected to the junction of the capacitors and resistors, the resistance decades can be calibrated to read *D* directly at a particular frequency, in this case at 1000 cps. With four decades of 100, 10, 1, and 0.1 ohms per step and with the total capacitance adjusted to 0.001592 μf, the range of *D* at 1000 cps is from 0.01 to 0.000001. At other frequencies, the indicated *D* must be multiplied by the frequency in kilocycles. To extend the range to higher *D*, additional capacitors are added by a range switch to make $C_T = 0.01592$ μf for a maximum *D* of 0.1 and to make

$C_T = 0.1592$ μf for a maximum D of 1. This capacitance added between the resistors and the transformer end of the detector does not change the capacitance balance.

Although the bridge has only four-figure resolution in D, this precision is adequate for the six-figure capacitance balance of capacitors whose D is 0.01 or less, since the smallest division of the 0.01 range of the D decades is one part per million.

Conductance, G

Balance of loss in the unknown in terms of shunt conductance, G, is provided in this bridge by the equivalent of four decades of conductance in parallel with the internal capacitors, as shown figuratively in Figure 4. The conductance needed for the loss in most capacitors is small, corresponding to resistance much greater than a megohm, so that ordinary resistance decades cannot be simply connected across the capacitors. It is simple, however, to use resistance decades in a T network to obtain a variable conductance. With 100-kilohm resistors as the series arms and the same four resistance decades used for D as the shunt arm, the range of G is from 0.1 μmho to 0.00001 μmho. The conductance is reduced by a factor of ten when the network is switched to the 0.1 tap on the transformer instead of to the full winding, and the range is then from 0.01 to 0.000001 μmho. When the loss in the external or internal capacitors exceeds that of the unknown, the bridge must be able to add loss to the unknown. With the conductance balance of loss, the T network can be readily switched to the taps at full or tenth voltage on the unknown side of the bridge to provide the same two ranges of conductance across the unknown $(-G)$ as there are for conductance across the internal and external standards $(+G)$.

ACCURACY

The accuracy of the bridge is determined primarily by the accuracy of the transformer ratios and by the accuracy of the internal standard capacitors. The accuracy of the ratios depends upon the magnitude of the ratio, upon frequency, and upon the load connected to the transformer. The accuracy of the capacitors, which depends initially upon the accuracy of the reference standard with which they are calibrated, is usually limited subsequently by the changes produced by aging and by fluctuations in temperature, pressure, and humidity. To achieve an accuracy of 0.01% in the bridge reading over a wide range of frequency and capacitance and without frequent recalibration, particular care has been taken in the construction of the transformers and capacitors.

Transformers

Relatively low numbers of turns are used in the transformers to keep the leakage inductance, stray capacitance, and resistances of the windings so small that the ratio accuracy remains high, even with loads greater than 1 μf and frequencies above 10 kc. These small residual impedances make it possible, for example, when a 1000-pf capacitor is being measured at 1000 cps with unity ratio, to load the transformer with as much as 1 μf of ground or cable capacitance before the error in the measured direct capacitance exceeds 0.01%. The small bridge inductances are not insignificant, however, when high capacitance is measured at high frequency, and the bridge error is then of the order of $+0.002\%\ C_{\mu f}\left(\frac{f}{1000}\right)^2$, if no correction for the inductance is used.

The accuracy of the ratios when the transformer is lightly loaded is better than 0.1 part per million for the unity ratio and is better than 2 ppm for the 0.1 ratio at 1000 cps or lower frequencies.

The winding self-capacitances act as a load as frequency increases, so that the error in the 0.1 ratio increases to about 20 ppm at 10 kc and to 0.2% at 100 kc. When the auxiliary transformer is connected for ratios of 0.01 and 0.001, the ratio errors are increased by the loading effects of the input impedance of the auxiliary transformer. These errors can, however, to a large extent be eliminated by compensating impedances, and the 0.01 and 0.001 ratios in the bridge are adjusted to within ±20 ppm in the frequency range below 10 kc. The phase errors are, in general, somewhat larger than the magnitude errors of the ratios. At 1000 cps, the phase error is probably within ±10 μradians, but the error increases in approximate proportion to ratio and to the square of frequency.

Capacitors

The internal standard capacitors are constructed to have such small changes with time, temperature, and environment that the initial calibration to ±0.01% may be expected to change less than 0.01% per year in normal use. The temperature coefficients of the 1000-, 100-, and 10-pf units, which are Invar multiple-plate capacitors, are less than 5 ppm/°C; the coefficients of the Invar Zichner-type 1-, 0.1-, and 0.01-pf units and of the cylindrical 0.001- and 0.0001-pf units are less than 20 ppm/°C.

For almost zero changes of capacitance with atmospheric pressure and humidity, all but the two smallest capacitors are hermetically sealed in an atmosphere of dry nitrogen. This sealing is necessary where stability of better than 0.01% is expected, because in an unsealed capacitor the capacitance changes about 2 ppm for each 1% change in relative humidity; hence a 50% change in humidity produces a 0.01% change in capacitance. And the pressure change, for example, resulting from moving the capacitor from the near-sea-level altitude of Washington, D.C., to the more than 5000-ft altitude of Boulder, Colorado, produces a capacitance decrease of about 0.01%.

To minimize long-term drift, all metal parts of the capacitors are Invar to avoid differential stresses, and they are annealed and temperature-cycled to relieve strains and to accelerate the initial aging.

The bridge can be calibrated quickly and accurately by the measurement of a single calibrated external standard capacitor of almost any size within the range of the bridge. Since the six-figure resolution of the bridge permits comparison with a precision better than 0.01% down to 1 pf, the accuracy of calibration is usually determined by the accuracy of the standard. Only one external standard, most conveniently a three-terminal 1000-pf standard,* is required because the accurate, internal 0.1 transformer ratio can be used to insure an accurate ratio of the internal capacitance standards. A −1 position on each capacitance lever switch connects the corresponding internal capacitor to the 0.1 tap on the unknown side of the transformer. This capacitor can be compared with the next decade capacitor, which is connected to the maximum voltage on the standard side when the adjacent lever is set on the x position, and any adjustments required can be made with trimmers accessible beneath a sliding cover on the bridge panel.

Such checks or recalibrations of the bridge need not be made often.

Loss

Although the accuracy of the measurement of loss is not important in the measurement of many capacitors, the Type 1615-A Capacitance Bridge makes possible measurements of dissipation factor to an accuracy which exceeds

*The Type 1404 Reference Standard Capacitors are recommended. These will be described in a subsequent issue.

that of most capacitance bridges. This accuracy of ($\pm 0.1\% + 10$ ppm) of the measured value is applicable over the whole D range and over nearly all the capacitance and frequency ranges. At low frequencies and small capacitance the accuracy will be limited by the reduced sensitivity of the bridge. At high frequencies and at ratios other than unity, the phase errors of the transformers will reduce the accuracy. Within these extremes, the accuracy of the D reading is determined by the resistance decades, which are adjusted within $\pm 0.05\%$, and by the total capacitance connected to the decades, which is trimmed to adjust the D reading to within $\pm 0.1\%$ when a standard of known D is measured.

The loss measurement in terms of shunt conductance, G, is limited to an accuracy of $\pm(1\% + 0.00001\ \mu\text{mho})$ by the accuracy of the 100-kilohm resistors used in the T network. Higher accuracy is seldom needed. It would not only add to the cost but would also require corrections to the bridge G reading. These corrections, amounting to a maximum of 2%, are due to the non-linear relation between the decade resistance and the equivalent conductance of the network.

The loss measured by the bridge as either D or G is the loss of the unknown capacitor relative to the loss of the internal standards. Since the bridge capacitors are carefully cleaned and sealed in dry nitrogen, it is estimated that their dissipation factor does not exceed a few parts per million. The accuracy of absolute loss measured by the bridge is, therefore, the same as that of the loss relative to the bridge capacitors.

CONVENIENCE

Readout and Balance (Refer to Figure 1.)

Past experience leads many of us to picture a bridge of very high precision and accuracy as a massive but delicate laboratory instrument which, when handled with considerable care, coddling and some cunning, may yield an accurate value for capacitance only after the application of numerous corrections. The TYPE 1615-A Capacitance Bridge in no way fits this picture. The moderate size and weight of this bridge permit it to be moved about the laboratory with ease, and the bridge is sufficiently rugged to be transported into the field should its accuracy be required there. It is easy to balance, easy to read, and the reading is accurate without corrections.

A feature which contributes much to the ease of balance and of reading is the use of lever or linear rather than rotary switches for the decades. The small panel space occupied by these switches makes it possible to position the six decades and range switch for capacitance and the four decades and range switch for loss within the span of the operator's right and left hands, respectively. The throw of the switches is about three inches, so the 12-position range of any decade can be covered with only a slight motion of hand or finger.

The position of each decade is indicated by a number appearing in the window above each lever. The bridge capacitance readout thus appears in the convenient form of six closely-spaced digits in a horizontal line and the D or G readout as a similar line of four digits. As the lever at the right is moved to change capacitance range, the decimal point is automatically positioned in the six-figure readout to indicate without multipliers the capacitance in picofarads from a maximum of 1,111,110 pf to a minimum of 0.00001 pf. The lever on the left similarly moves the decimal point when the D range is changed to indicate directly the dissipation factor. The decimal point is also positioned automatically to read conductance in micromhos, but since G

must be multiplied by the factor M, this factor is indicated in orange engraving adjacent to each position of the C MAX range switch lever. This multiplier is required only for *G* and for external standards, and the orange color is used on the panel to indicate all quantities to which M must be applied.

Bridges of high precision are often reputed to be bridges which are not easily balanced. In spite of its wide range and high precision, the TYPE 1615-A Capacitance Bridge can often be balanced with more ease and speed than bridges of lower range and accuracy. For example, when even the approximate magnitude of a capacitor is not known, a rough balance can be made quickly on this bridge by the use of the maximum capacitance range, so that the six decades cover the range from 1 μf to 1 pf and the six levers can be tried in quick succession to determine the balance point without a change in range. The —1 position on each of the capacitance decades, which was mentioned above as useful in the self-calibration of the bridge, also facilitates balance in the region near any zero by permitting a trial reduction of bridge capacitance by one step in a decade without the necessity of moving the adjacent lever.

Connection of Unknown

The convenience of the balance controls is matched by the convenience with which various types of capacitors can be connected to the bridge for measurement. Two types of connector for the unknown capacitors are provided at the upper right corner of the bridge panel: a pair of TYPE 874 Coaxial Connectors and a set of three TYPE 938 Binding Posts with standard ¾-inch spacing. For three-terminal measurements with complete shielding, as is required particularly for very small capacitance, three-terminal capacitors, such as the TYPE 1403 Standard Air Capacitors and TYPE 1422-CD Precision Capacitor, can be connected with coaxial cables to the coaxial bridge terminals. Capacitors having other common types of coaxial connectors can also be connected to the bridge terminals by the use of the appropriate TYPE 874-Q Adaptor. Capacitors, such as the TYPE 1401 and TYPE 1409 Standard Capacitors, which have TYPE 274 Plugs as terminals, can be plugged into the jack-top binding posts. The binding posts can also be used for the connection of patch cords and leads of many types.

The appropriate set of unknown terminals is connected to the bridge (and the unused terminals disconnected) by means of a four-position terminal switch located next to these terminals. As this switch is moved to change terminals, it also shows the corresponding changes of connections and grounds in the simple circuit which is engraved on the panel. This simple circuit diagram does not replace the operating instruction manual, but it does serve even the constant user as a useful and ever-present reminder of the circuit which is in use and of the possible sources of measurement or connection error.

When the terminal switch is set in the position marked CAL, the L or detector side of all the terminals is disconnected. This permits a check or self-calibration of the bridge capacitors at any time without the need for disconnecting the unknown.

Three-Terminal

In the next position, marked 3 TERM, the coaxial TYPE 874 unknown terminals are connected to the bridge, with the L terminal connected to the detector and the H terminal to the transformer. The shields of the connectors and all ground points on the bridge are connected to the guard point, so that all capacitances to the shields or to ground are excluded from the direct capacitance between H and L measured by the bridge.

The third position of the switch, marked 3 TERMINAL, connects to the bridge the H, L, and GND binding posts instead of the coaxial terminals. The H post is connected to the transformer, the L post to the detector, and the GND post to the transformer midpoint and bridge ground. As in the coaxial three-terminal measurement, the bridge measures only the direct capacitance between the H and L posts and excludes capacitances from H or L to any GND or guard point. The open binding posts have a direct capacitance of about 0.2 pf, which must usually be measured and subtracted from the value measured when a capacitor is connected. The bridge can, of course, measure this small terminal capacitance, as well as that of any leads connected between terminals and capacitor.

Two-Terminal

The fourth position of the switch, marked 2 TERMINAL, deserves special attention because of the important changes it makes in bridge connections and bridge measurements. The bridge is again connected to the binding-post terminals with the H post connected to the transformer, but the L and GND posts are now connected together and to the bridge case and panel and to any external ground used. The bridge now measures *all* capacitances between the H terminal and L or GND, including stray capacitances from post and leads to the panel and other environment. These are the capacitances measured by the common two-terminal capacitance bridge, so that it is possible to duplicate with the new TYPE 1615-A Capacitance Bridge the measurements of two-terminal capacitors obtained with older bridges, such as the TYPE 716-C Capacitance Bridge.

In principle, this change of the inherently three-terminal transformer bridge to two-terminal operation is made as shown in Figure 4; the ground point is simply switched from the center of the transformer arms to the junction of the standard and unknown capacitors, thereby grounding one side of the unknown. In practice, this change is complicated by the fact that the center of the transformer, which is the guard point to which the bridge shields are connected, is then connected to the high-impedance side of the detector instead of to ground. To prevent error voltages from entering the detector, all the wires and bridge shields connected to the high side of the detector must be enclosed by a grounded shield. To provide this extra shielding for two-terminal measurements, the bridge components are enclosed in an inner shield box which is enclosed by but insulated from the outer box and panel, and the primary of the main ratio transformer is also enclosed in two separate shields.

External Standards

Range Extension

The usefulness of the bridge is further increased by the provision on the bridge panel of a pair of terminals to permit the connection of an external standard capacitor or resistor to supplement or replace the standards in the bridge. This pair of coaxial TYPE 874 Connectors, located to the left of the coaxial pair for the unknown, has the L terminal connected to the L terminal of the unknown and the H terminal connected to the standard side of the transformer through a rotary switch, by means of which any of the ten steps of voltage from the transformer can be applied to the external standard. This rotary switch, with its digital readout through a window, provides a seventh decade of capacitance or a fifth of conductance whose magnitude is determined by the external standard chosen. For example, the capacitance range can be extended to 11 μf by the connection of an external standard of 0.01 μf. With the C MAX range lever set at the 1 μf maximum, the rotary decade then provides a balance control of 1 μf per step and the lever switches extend the balance range six more decades from 0.1 μf through 1 pf per step.

Accuracy Extension

Since both the unknown and external standard capacitors can be connected to a wide range of accurate transformer ratios, a comparison of external capacitors can be made with an accuracy even higher than that of the direct bridge reading; and the ratios can be chosen so that the magnitudes of the external capacitors do not have to be decade multiples. For example, suppose a standard capacitor of 1000 pf is available with a calibration accuracy higher than 0.01%. This accuracy can be transferred to a capacitor of, say, 5000 pf by connecting that capacitor to the appropriate unknown terminals and the 1000-pf standard to the external standard terminals. When the rotary decade switch for the external standard is set to 0.5 and the C MAX lever to the 0.01-μf position (where M = 10), the external standard is effectively multiplied by 5 to balance the unknown. Small differences between the external capacitors can, of course, be balanced with the bridge capacitance and conductance decades, and any small errors in the bridge reading of the difference are insignificant in the comparison measurement as long as the difference is a small percentage of the total capacitance.

Resolution Extension

The resolution, as well as accuracy, of the bridge can be extended by the use of an external standard capacitor. It has already been noted above that the external standard and its decade switch add a seventh decade, which can have increments either larger or smaller than those of the six lever decades. Even higher resolution is possible when, for example, two 1000-pf external capacitors are compared, because the bridge decades can be used to measure a difference as small as 0.00001 pf or 1 part in 10^8 in this example. Usable resolution of 0.1 ppm is not hard to obtain with the recommended TYPE 1232-A Null Detector, but higher resolution usually requires special detectors.

GENERATOR AND DETECTOR

The fact that the instrument contains neither generator nor detector may not seem a convenience to the occasional user of the TYPE 1615-A Capacitance Bridge, but it is often an engineering and economic advantage. A generator and a detector in separate packages can be better selected or modified to fit the many uses of the bridge over its wide range of capacitance and frequency. For most of the uses and most of the range, the recommended generator is the new TYPE 1311-A Audio Oscillator[4] and the recommended detector is the TYPE 1232-A Tuned Amplifier and Null Detector.[5] A complete system for capacitance measurement, consisting of the bridge and the recommended generator and detector, is available as the TYPE 1620-A Capacitance-Measuring Assembly, illustrated on page 14.

— J. F. Hersh

CREDITS

The TYPE 1615-A Capacitance Bridge was developed by John F. Hersh. Others contributing to the final design are R. A. Soderman, Administrative Engineer; G. A. Clemow, Design Engineer; G. C. Oliver, Design Draftsman; W. H. Higginbotham, Production Engineer, and W. G. Cooper, Assistant Test Engineer.

— EDITOR

[4]Described elsewhere in this issue.

[5]A. E. Sanderson, "A Tuned Amplifier and Null Detector with One-Microvolt Sensitivity," *General Radio Experimenter*, 35, 7, July, 1961.

SPECIFICATIONS

Capacitance Range (6 ranges): 10^{-17} to 10^{-6} farads (10 μpf to 1μf), direct reading; 6-figure resolution, smallest division 10^{-17} farads.

Dissipation-Factor Range (3 ranges): 0.000001 to 1 at 1 kc, direct reading. Directly proportional to frequency at other frequencies. Four-figure resolution; smallest division, 0.000001.

Conductance Range (2 ranges +; 2 ranges −): 10^{-6} μmho to 100 μmho; 4-figure resolution, smallest division 10^{-6} μmho; independent of frequency; varies with C range.

Accuracy:

Capacitance—direct reading, internal standard, ±0.01%, except at the extremes of the range. At high capacitance and high frequency, error is $+ 0.002\% \, C_{\mu f} \left(\frac{f}{1000}\right)^2$. At low capacitance and low frequency, accuracy may be limited by bridge sensitivity.

Capacitance—comparison with external standard, approximately 1 ppm.

Dissipation factor, ±(0.1% + 10 ppm) of measured value.

Conductance, ±(1% + 0.00001 μmho).

Frequency Range: Approximately 100 cycles to 10 kc.

Temperature Coefficients of Internal Standards: Less than 5 ppm/°C for the 1000-, 100-, and 10-pf units; slightly greater for the smaller capacitance units.

Maximum Voltage: 20 volts at 1 kc. Proportional to frequency.

Accessories Required: Generator and detector; the Type 1311-A Audio Oscillator and the Type 1232-A Tuned Amplifier and Null Detector are recommended.

Accessories Supplied: Type 874-WO Open-Circuit Termination, Type 874-R22 Patch Cord, and Type 274-NL Patch Cord.

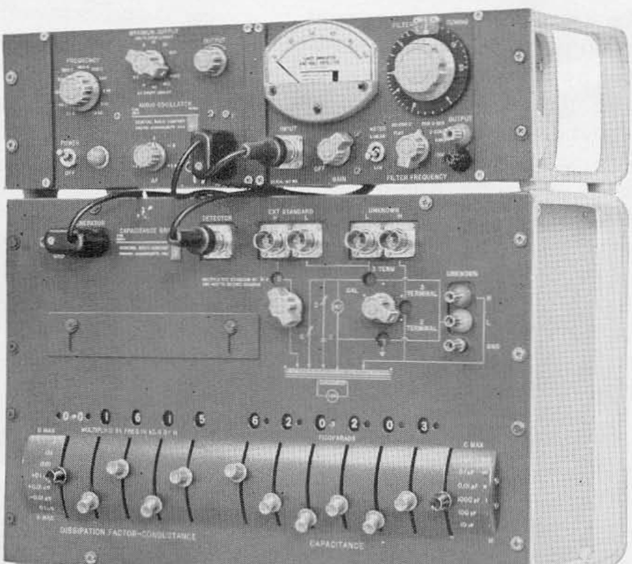
Dimensions: Width 19, height 10½, depth 12¾ inches (485 by 270 by 325 mm), over-all.

Net Weight: 38½ pounds (17.5 kg).

Type		*Code Word*	*Price*
1615-AM	**Capacitance Bridge, Bench Model........**	ATTIC	**$1475.00**
1615-AR	**Capacitance Bridge, Cabinet Model......**	BALMY	**1475.00**

U.S. Patent No. 2,548,457.

TYPE 1620-A CAPACITANCE-MEASURING ASSEMBLY



The Type 1620-A Capacitance-Measuring Assembly consists of the Type 1615-AM Capacitance Bridge with the Type 1311-A Audio Oscillator and the Type 1232-A Tuned Amplifier and Null Detector, thus providing a complete system for the precise measurement of capacitance over the range of 10 μpf to 1 μf (10^{-17} to 10^{-6} farads). Frequency range is approximately 50 cps to 10 kc. The system has sufficient sensitivity to realize the full six-place resolution of the bridge for all measurements except for very small capacitances at the lower frequencies.

Oscillator and detector are mounted side by side as shown in the photograph. The end frames are bolted together to make a rigid assembly without the use of a relay rack. Connection cables are supplied.

The oscillator operates from the power line, the detector from internal batteries.

Type		*Code Word*	*Price*
1620-A	**Capacitance-Measuring Assembly**	ORBIT	**$2080.00**

HIGH PERFORMANCE, LOW-COST AUDIO OSCILLATOR WITH SOLID-STATE CIRCUITRY

Modern solid-state circuitry is used in the new TYPE 1311-A Audio Oscillator to produce a self-contained, compact, inexpensive instrument with many desirable features. Among these are high-power output into a wide range of load impedances, low-distortion even when the load impedance is short-circuited, excellent stability, low noise, and very small size.

The TYPE 1311-A Audio Oscillator supplies power at eleven commonly used

Figure 1. Panel View of the Type 1311-A Audio Oscillator.

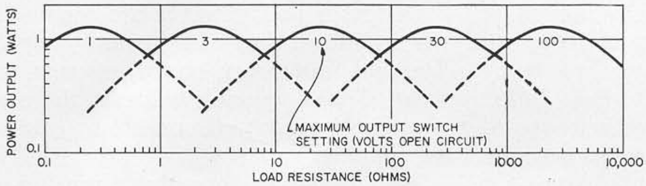


Figure 2. The output transformer allows matching a wide range of load impedances. Data were taken at 1 kc, but are representative of performance at all frequencies.

fixed frequencies: 50, 60, 100, 120, 200, 400 and 500 cps and 1, 2, 5 and 10 kc as selected by a rotary switch. A continuously adjustable incremental-frequency control provides a range of $\pm 2\%$ about the nominal frequency. One additional frequency can be provided by the user at a twelfth switch position by the addition of two resistors.

The output transformer has a tapped secondary winding, so that an output power of at least one watt can be delivered to five different load impedances, and at least one-half watt to any resistive load between 80 milliohms and 8 kilohms, as shown in Figure 2.

In most oscillators, overloading and waveform clipping occur when the load impedance is very low compared to its optimum value. In contrast, the TYPE 1311-A Audio Oscillator can supply a low-distortion signal to any load impedance from an open circuit to a short circuit, independent of the setting of the tap on the output transformer. The over-all distortion is always low, less than 0.5% at a 1-watt output level and typically less than 0.1% over much of the frequency range, as shown in Figure 3. Hum and noise components are less than 0.003% of the maximum output.

CIRCUIT

The oscillator makes use of the familiar Wien bridge network and a multistage, Class-B, transistor amplifier to provide the necessary power output without additional buffer amplifiers. A simplified schematic diagram is shown in Figure 4. The frequency of oscillation is determined by the capacitors and one of eleven pairs of resistors in the positive feedback path. A thermistor is part of the negative feedback path and assures a very stable output signal, as shown in Figures 5 and 6, without the distortion associated with many amplitude-limiting systems.

Six transistors are incorporated in a single direct-coupled feedback loop. The high loop gain results in an oscillator which is substantially independent of transistor characteristics, with low distortion and long-term reliability. Noise and short-term amplitude and frequency variations are minimized by the use of low-noise circuitry for the input am-

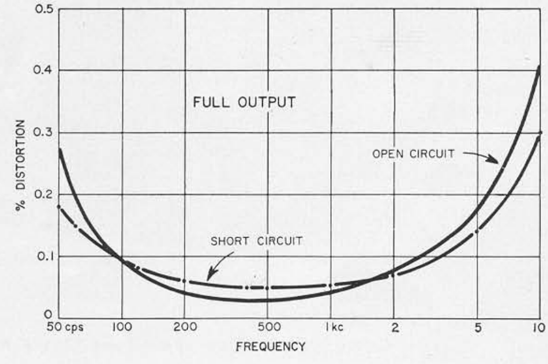


Figure 3. The Type 1311-A Audio Oscillator will drive any impedance with low distortion.

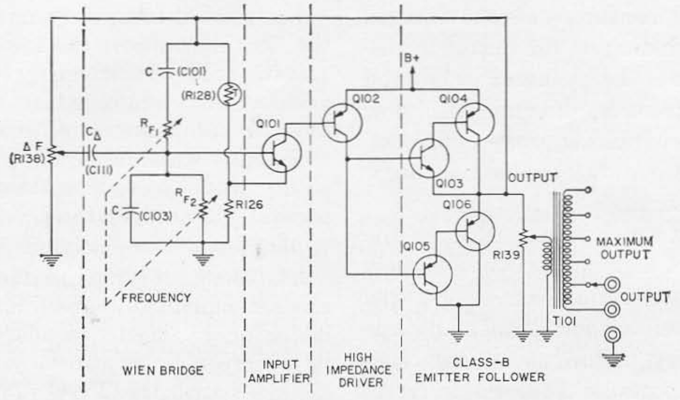


Figure 4. Elementary schematic diagram of the oscillator.

plifier, Q101. The transistor, Q102, provides a high-impedance drive circuit for operation of the Class-B output stage with a minimum of crossover distortion, without the use of complicated temperature-sensitive bias networks.[1]

Since the RC-network capacitors are too large (0.1 μf) to be made adjustable, the incremental-frequency adjustment is produced by a variation in the voltage across part of one of the capacitors by means of a potentiometer. This has the same effect on the circuit as a variation in capacitance, and, since the potentiometer impedance is low compared to that of the capacitor, the control can be calibrated in percentage frequency change.

[1]J. J. Faran and R. G. Fulks, "High-Impedance Drive for the Elimination of Crossover Distortion," *The Solid-State Journal*, August, 1961.

APPLICATIONS

Although the Type 1311-A Audio Oscillator was designed primarily for use as a generator for bridge measurements, its superior performance and many features make it well suited to almost any application where a high-quality audio oscillator is needed.

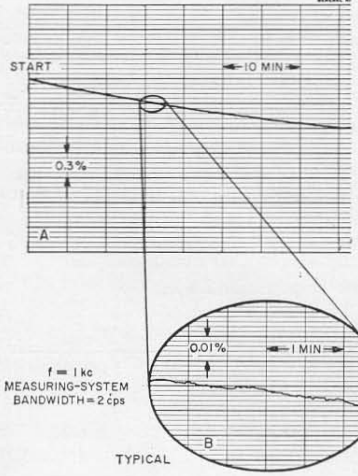
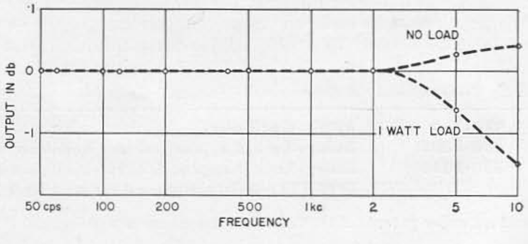


Figure 5. Typical output amplitude stability of the oscillator, showing warmup drift (A) and short-term variation (B).

Figure 6. The flat frequency response of the oscillator eliminates the need for rechecking the level in frequency-response measurements.



For bridge measurements the shielded secondary winding on the output transformer permits the oscillator to be used as a floating source, thus minimizing or eliminating circulating ground currents. This feature is also important in other low-level systems.

For many applications, such as the calibration of high-speed level recorders and analog-to-digital converters, the very low level of short-term amplitude and frequency variations in this oscillator are important. Appreciable errors can be caused by the cycle-to-cycle variation found in most oscillators.

[2]A. E. Sanderson, "A Tuned Amplifier and Null Detector with One-Microvolt Sensitivity," *General Radio Experimenter*, 35, 7, July, 1961.

For general laboratory measurements the floating output, the low distortion, and the ability to drive any load impedance without clipping are among the most useful features of this oscillator, while the small size, simplicity, reliability and excellent stability are important advantages for production-test applications.

The oscillator is mounted in a compact cabinet which can be used either on the bench or, by means of adaptor panels, in a relay rack. It can be conveniently mounted with the TYPE 1232-A Tuned Amplifier and Null Detector[2] as a complete oscillator-detector combination for relay-rack mounting. Relay-rack adaptor sets for this purpose are listed below.

— R. G. FULKS

SPECIFICATIONS

FREQUENCY

Range: 11 fixed frequencies from 50 to 10,000 cps.

Control: 50, 60, 100, 120, 200, 400, 500, 1000, 2000, 5000, 10,000 cps selected by rotary switch. A vernier provides a ±2% adjustment about nominal.

Accuracy: ±1% when Δf control is at zero.

OUTPUT

Power: One watt into matched load. (Taps provide at least one-half watt output into any resistive load between 80 milliohms and 8 kilohms.)

Voltage: Continuously adjustable from 0 to 1, 3, 10, 30, or 100 volts, open circuit.

Current: Continuously adjustable from 0 to 40, 130, 400, 1300, 4000 milliamperes, short circuit (approx).

Impedance: Between one and two times matched load, depending on control setting. Output circuit is isolated from ground and, hence, can be used to drive balanced circuits.

DISTORTION AND NOISE LEVEL

Distortion: Less than 0.5% under any load condition. Typically less than 0.1% over much of range. Oscillator will drive a short circuit without waveform clipping.

AC Hum: Typically less than 0.003% of output voltage.

GENERAL

Terminals: Jack-top TYPE 938 Binding Posts with standard ¾-inch spacing. Separate ground terminal holds TYPE 938-L Shorting Link which can be used to ground adjacent OUTPUT binding posts.

Power Input: 105 to 125 (or 210 to 250) volts, 50 to 400 cps. Total power input varies between 7 and 22 watts, depending on load.

Mounting: Aluminum panel and cabinet, in gray-crackle finish, for bench use. Panel adaptor sets are available to permit mounting in standard 19-inch relay rack.

Accessories Supplied: TYPE CAP-22 Power Cord, spare fuses.

Dimensions: Width 8, height 6, depth 7¾ inches (205 by 155 by 200 mm), over-all.

Net Weight: 6 pounds (2.8 kg).

Type		*Code Word*	*Price*
1311-A	**Audio Oscillator**	TIPSY	**$175.00**
480-P308	**Relay-Rack Adaptor Set (for oscillator only)**	EXPANELDOG	**7.00**
480-P316	**Relay-Rack Adaptor Set (for oscillator and TYPE 1232-A Tuned Amplifier and Null Detector)**	EXPANELHUM	**6.00**

Licensed under patents of the American Telephone and Telegraph Company.

NEW COAXIAL CABLE CONNECTORS

Have Lower VSWR, Are Easier to Install

The continuous development program for GR TYPE 874 Coaxial Connectors,* among whose recent achievements was the locking version of the connector,[1] has now produced a greatly improved design of cable connectors. Both mechanically and electrically, the performance of these new cable (and panel) connectors is commensurate with that of the rigid-line connectors. The new series, which is identified by the letter "A" in the type number and by a gray rubber guard on the cable end, instead of the black previously used, will replace the older series on September 1, 1962.

DESIGN CHANGES

The design objectives were (a) lower VSWR, (b) minimum change in VSWR as a result of assembly variations, and (c) simplified assembly. These have been achieved by (a) redesign of the transition between the basic TYPE 874 Connector and the cable, (b) reduction in the distortion of cable dielectric due to melting during the soldering operation on the inner conductor, (c) an improved method for attaching the cable braid and jacket, and (d) improvements in the rubber guard used on patch-cord types.

In addition, the inner conductor soldering operation for small-diameter cables has been simplified, and a tendency for the TYPE 874-PB Panel Connector to become slightly loose in its panel flange with hard usage has been corrected.

The transition between the 9/16″ air-dielectric connector and smaller solid-dielectric cables, shown in Figure 1, is an important part of the connector. The older, tapered, design has been replaced by a step transition, which is easier to control in machining, easier to inspect, and easier to solder. Crimping of the inner transition to the cable wire is also possible with this design, for non-critical applications.

The most critical part of any cable connector is the actual connection between the cable conductors and the corresponding parts of the connector. The center-conductor connection is usually the most difficult to make with consistently low reflections. One of the causes of high VSWR is flow and distortion of the dielectric material during soldering. Another related cause is the variation in location of the end of the cable dielectric with respect to the end of the connector inner transition section. Furthermore, the plastic dielectric can actually flow into the soldered joint and produce an inductive discontinuity. These conditions are illustrated in Figure 2.

These problems have been eliminated by the addition of a Teflon disk, as shown in Figure 3, which blocks any

*U. S. Patent No. 2,548,497.

[1]"New and Improved Coaxial Connectors," *General Radio Experimenter*, 35, 10, October, 1961.

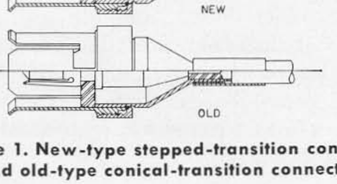
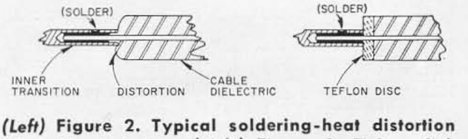


Figure 1. New-type stepped-transition connection and old-type conical-transition connection.



***(Left)* Figure 2. Typical soldering-heat distortion of cable dielectric. *(Right)* Figure 3. Teflon disk eliminates distortion.**

Figure 4. New, perforated ferrule.

flow of the cable dielectric into the soldered joint, and provides heat insulation. It provides also a definite surface for the end of the inner transition to rest against during the soldering operation. Still further control of dielectric distortion is provided by an improved assembly procedure.

The connection between the cable braid and connector also has been improved. A taper has been added at the end of the knurled section of the outer transition, as shown in Figure 5, to reduce the inductive discontinuity due to the step-up in diameter, and a new perforated ferrule, shown in Figures 4 and 5, has been provided to hold both the jacket and braid securely in place.

The cable jacket flows out into the perforations, producing an effective holding force, and preventing the jacket from drawing away from the braid-to-connector joint. In addition, crimping is necessary over only a short length of ferrule, as shown in Figure 5. As a result, the dielectric is compressed very much less than with a solid ferrule, and the effect on the VSWR is relatively small even at frequencies up to 7 Gc. The perforated ferrule is employed in all connectors of the new series.

The knurled, cylindrical transition with a crimped-ferrule method of braid holding was retained in the new design

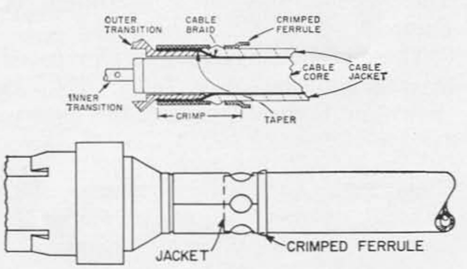


Figure 5. Improved method of jacket and braid retention.

because it offers several advantages over the butt-retention systems commonly employed. The principal advantage is that it provides the least discontinuous transition consistent with the necessary requirement for TYPE 874 Connectors that the inner transition must be pushed slightly forward during assembly to install the insulator in the basic connector.

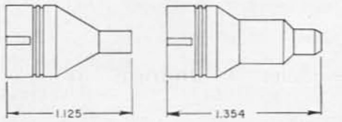


Figure 6. Increased length facilitates soldering.

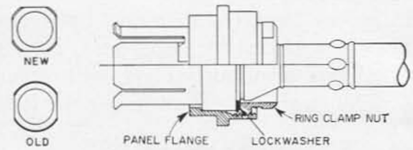


Figure 7. *(Right)* New ring clamp nut has additional flats. *(Left)* Positive locking of connector in "PB" series of panel flanges, with lockwasher.

With this type of assembly, characteristic of all TYPE 874 Cable Connectors, the inner transition and inner connector are accurately positioned, and the tendency for movement caused by flexing, expansion, or contraction of the cable with temperature changes is practically eliminated. Other advantages include rapid assembly and low cost.

The rearward protrusion of the inner transition is now used on all connectors, so that the "58A" and "62A" series of connectors can be soldered as easily as the larger cable types. For this purpose, the new transition pieces in these series have been lengthened (Figure 6).

A lockwasher has been added to prevent possible loosening of the connector in the panel flange, and an additional set of flats has been provided on the nut that clamps the basic cable connector in the panel flange to facilitate tightening behind a panel where accessibility is poor. See Figure 7.

ELECTRICAL PERFORMANCE

The new stepped transitions yield a very low VSWR up to 7 Gc, and the reflections introduced by the connector as a whole are now lower than the reflections inherent in even the best flexible cables.

In the development of the new cable connectors, it was not possible to employ standard flexible cables to test the connectors, because these cables are not made to sufficiently close tolerances and are not sufficiently uniform. It was necessary, therefore, to build sections of dielectric-filled line, accurately constructed to be as close as possible to the desired 50-ohm characteristic impedance. Each stepped section of the transition was designed and tested individually in order to isolate each individual discontinuity and minimize its reflection. This is important in achieving a low-reflection design above 4 Gc. The results of VSWR measurements made on development units are shown in Figures 8, 9 and 10. The test configuration, shown in

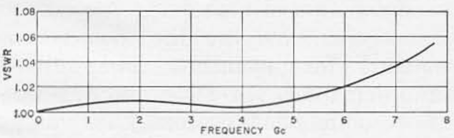


Figure 8. VSWR of a pair of "CA" cable transitions on ideal cable section.

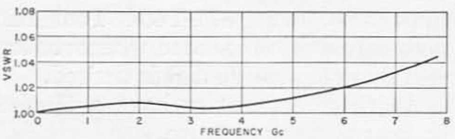


Figure 9. VSWR of a pair of "C8A" cable transitions on ideal cable section.

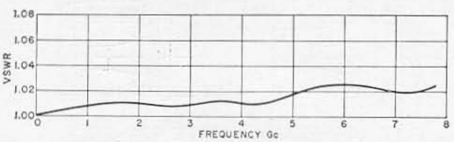


Figure 10. VSWR of a pair of "C58A" cable transitions on ideal cable section.

Figure 11. Test configuration for Figures 8, 9, and 10.

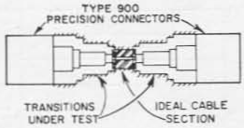


Figure 11, comprises two transitions, less the Type 874-B Basic Connectors, connected back-to-back through a short section of 50-ohm polyethylene line with a Teflon disk placed at each end, simulating the connection to a cable.

The over-all VSWR of the complete connectors installed on actual cables is excellent, as shown in Figures 12 to 17, for two basic attachments. In the first, a connector is installed on an extremely long length of cable, simulating an infinite cable. Measurement of this configuration represents the VSWR of a

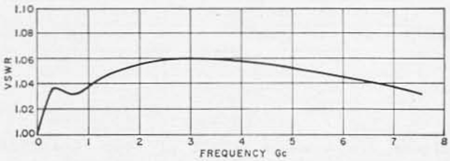


Figure 12. Average VSWR of single Type 874-CA Connector on an infinite length of Type 874-A2 Cable. (Also applies to the Types 874-CLA, -PBA, -PLA, and PRLA.) Peak at 300 Mc is due to cable characteristic-impedance error.

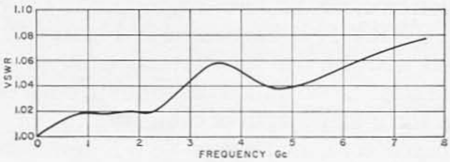


Figure 13. Average VSWR of single Type 874-C8A Connector on infinite length of RG-214/U Cable. (Applies to all the "8A" series.)

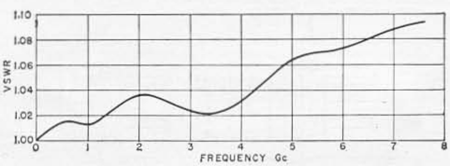


Figure 14. Average VSWR of single Type 874-C58A Connector on infinite length of Type 874-A3 Cable. (Applies to all the "58A" series.)

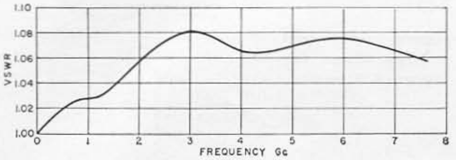


Figure 15. Average VSWR of Type 874-R20A Patch Cord consisting of two Type 874-CA Connectors mounted on Type 874-A2 Cable.

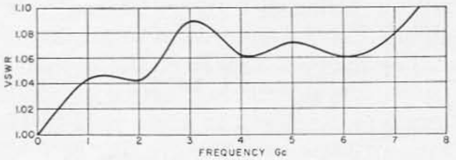


Figure 16. Average VSWR of patch cord consisting of two Type 874-C8A Connectors mounted on RG-214/U Cable (similar to RG-9/U).

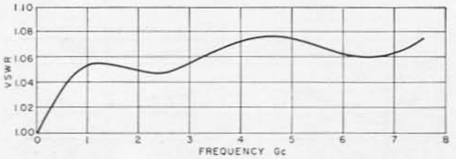


Figure 17. Average VSWR of Type 874-R22A Patch Cord consisting of two Type 874-C58A Connectors mounted on Type 874-A3 Cable.

single cable connector on a typical section of coaxial cable, sometimes referred to as the "rigid-line-to-cable VSWR." It is illustrated in Figures 12, 13 and 14. In the second, cable connectors are installed on opposite ends of a three-foot length of cable, typical for a patch cord, and a low-VSWR termination is plugged into one end. This is referred to as "cable-to-cable VSWR" when the cable length is a multiple of a half wavelength.[2] It is illustrated in Figures 15, 16 and 17.

A note of explanation is required for the "infinite" cable measurements. Most flexible cables exhibit a resonance phenomenon whereby periodic variations in characteristic impedance inherent in the manufacturing process become synchronous with the measurement frequency. These appear as periodic VSWR spikes in the measurements and are due to an accumulation of many reflections. These have been positively identified as occurring in the cable and, for this reason, have been omitted from the connector VSWR graphs.

AVAILABLE TYPES

The 874-series cable connectors are available with a variety of fittings that make them adaptable to both patch-cord and panel-mounting use. These are identified by the letter series C, CL, PB, PL, PRL. The suffix, L, identifies the locking type. The following table lists available catalog items. Also shown below the table are the tools recommended for assembly. The ferrule-crimping tools, 874-TO58 and 874-TO8, are recommended, especially for volume assembly, although suitable ferrule crimping can be achieved with ordinary pliers where appearance is not important. The Type 874-TOK Tool Kit, however, is recommended whenever low and reproducible VSWR is desired.

All these connectors have a 50-ohm characteristic impedance.

—John Zorzy

[2]A. E. Sanderson, "An Accurate Substitution Method of Measuring the VSWR of Coaxial Connectors," *The Microwave Journal*, January, 1962.

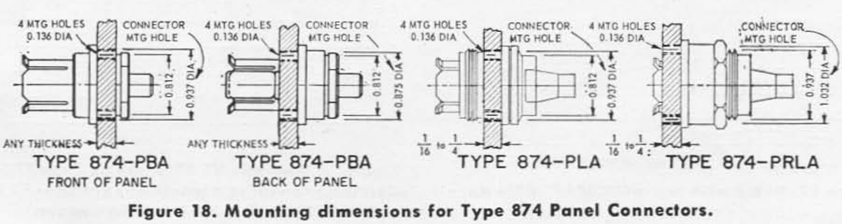
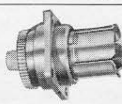
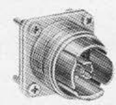
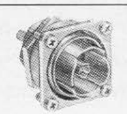


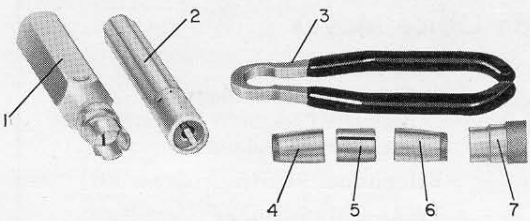
Figure 18. Mounting dimensions for Type 874 Panel Connectors.

CONNECTORS

	Type	Fits	Code Word	Price*	
50-ohm Cable Connectors	**874-CA** **874-C8A** **874-C58A** **874-C62A**	(50-ohm) 874-A2 Cable (50-ohm) RG-8A/U, -9B/U, -10A/U, -87A/U, -116/U, -156/U, -165/U, -166/U, -213/U, -214/U, -215/U, -225/U, -227/U; (non-constant impedance) RG-11A/U, -12A/U, -13A/U, -63B/U, -79B/U, -89/U, -144/U, -146/U, -149/U, -216/U Cables (50-ohm) 874-A3, RG-29/U, -55/U (series), -58/U(series), -141A/U, -142A/U, -159/U, -223/U Cables RG-59/U, -62/U(series), -71B/U, -140/U, -210/U Cables (non-constant impedance)	COAXCABLER COAXCORDER COAXCALLER COAXCANDOR	$2.50 2.50 2.50 2.50	
50-ohm Cable Connectors —Locking	**874-CLA** **874-CL8A** **874-CL58A** **874-CL62A**	Same as Type 874-CA. Same as Type 874-C8A. Same as Type 874-C58A. Same as Type 874-C62A.	COAXYROBIN COAXPARROT COAXYSNIPE COAXYSWIFT	3.50 3.50 3.50 3.50	
50-ohm Panel Connectors —Flanged	**874-PBA** **874-PB8A** **874-PB58A** **874-PB62A**	Same as Type 874-CA. Same as Type 874-C8A. Same as Type 874-C58A. Same as Type 874-C62A.	COAXAPPLER COAXBATHER COAXABATER COAXBARKER	3.40 3.40 3.40 3.40	
50-ohm Panel Connectors —Locking	**874-PLA** **874-PL8A** **874-PL58A** **874-PL62A** **874-PLT**	Same as Type 874-CA. Same as Type 874-C8A. Same as Type 874-C58A. Same as Type 874-C62A. Wire Lead.	COAXYFINCH COAXYVIREO COAXTHRUSH COAXTOUCAN COAXWILLET	3.75 3.75 3.75 3.75 3.75	
50-ohm Panel Connectors —Locking, Recessed	**874-PRLA** **874-PRL8A** **874-PRL58A** **874-PRL62A** **874-PRLT**	Same as Type 874-CA. Same as Type 874-C8A. Same as Type 874-C58A. Same as Type 874-C62A. Wire Lead.	COAXYGOOSE COAXCONDOR COAXCURLEW COAXAVOCET COAXMERLIN	4.00 4.00 4.00 4.00 4.00	

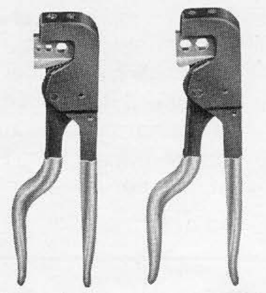
*For quantities of 1 to 99; prices for larger quantities on request.

TYPE 874-TOK TOOL KIT



1. Outer-conductor wrench . . . . (0874-2610)
2. Inner-conductor wrench . . . . . (0874-2611)
3. Coupling-nut wrench . . . . (0874-6801)
4. Front-ring expander (red) . . . (0874-6820)
5. Keeper for ring expanders . . (0874-6840)
6. Back-ring expander (green) . (0874-6800)
7. Ring pusher . . . . (0874-6830)

CRIMPING TOOLS



TYPE 874-TO58 TYPE 874-TO8

Type		Code Word	Price
874-TOK	**Tool Kit . . . . .**	COAXKITTEN	$20.00
874-TO8	**Crimping Tool**	COAXCRIMBA	75.00
874-TO58	**Crimping Tool**	COAXCRIMPO	85.00

Coming — in September

The 3rd Annual ELECTRONIC INSTRUMENT MANUFACTURERS' EXHIBIT

Boston to Washington, D. C.

This popular exhibit (EIME) has now been expanded to include eight leading instrument manufacturers exhibiting together in nine convenient locations. Operating displays of the latest instruments will again be featured, and factory engineers will be on hand to discuss your measurement problems in an unhurried atmosphere.

The sponsors are: General Radio Company, Ampex Corporation, FXR, Lambda Electronics Corporation, Non-Linear Systems, Inc., Panoramic Electronics, Inc., Sensitive Research Instrument Corporation, and Trio Laboratories, Inc.

General Radio will exhibit three new bridges: the TYPE 1608-A Impedance Bridge, the TYPE 1633-A Incremental Inductance Bridge and the TYPE 1615-A Capacitance Bridge. Other displays will include the TYPE 1150-A Digital Frequency Meter, TYPE 1360-A Microwave Oscillator, TYPE 1217-B Unit Pulse Generator, and TYPE 1558-A Octave Band Noise Analyzer.

Waltham, Massachusetts	Monday, Sept. 10	Charterhouse Motel, Route 128
Syracuse, New York	Tuesday, Sept. 11	Sheraton Inn, Carrier Circle
Jericho, Long Island, New York	Thursday, Sept. 13	Meadowbrook Motel
Cedar Grove, New Jersey	Monday, Sept. 17	The Towers
Camden, New Jersey	Wednesday, Sept. 19	Cherry Hill Inn
Philadelphia, Pennsylvania	Thursday, Sept. 20	Marriott Motor Hotel
Watchung, New Jersey	Monday, Sept. 24	Wally's Tavern-on-the-Hill
Red Bank, New Jersey	Wednesday, Sept. 26	Molly Pitcher Hotel
Washington, D. C.	Thursday, Sept. 27	Marriott Motor Hotel

Exhibit hours: Noon to 7:30 p.m.

Washington Office Moves

Effective September 1, 1962 the General Radio Office serving Washington, D.C. and the Southeast is moving from Silver Spring, Maryland to new and larger quarters.

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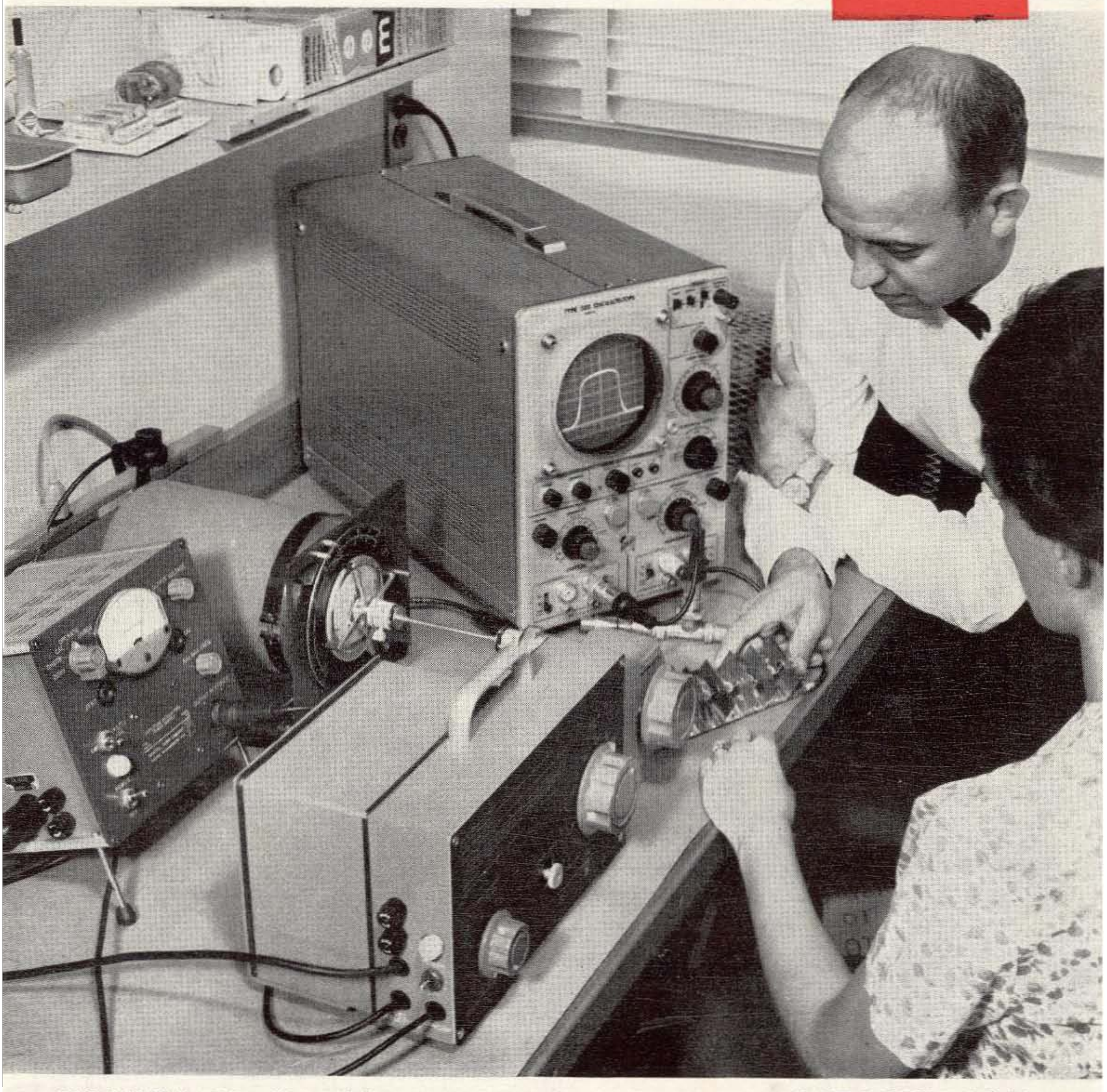

VOLUME 36 No. 10

OCTOBER, 1962

IN THIS ISSUE

New

Octave-Band Analyzer

Unit Oscillator

Photoelectric Pickoff

THE GENERAL RADIO

EXPERIMENTER

Published Monthly by the General Radio Company

VOLUME 36 · NUMBER 10 **OCTOBER, 1962**

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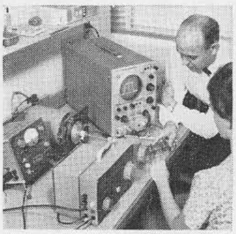
COVER



Measurement of the response of a band-pass filter, using the Type 1211-C Unit Oscillator, described in this issue. The Type 1263-B Amplitude-Regulating Power Supply holds the oscillator output voltage at a constant level. The test frequency is swept by means of the Type 1750-A Sweep Drive to produce the response curve on an oscilloscope screen.

NEW, COMPACT, OCTAVE-BAND ANALYZER

OPERATES DIRECTLY FROM PIEZOELECTRIC MICROPHONE

An octave-band analysis has become the most widely used method of determining the frequency distribution of acoustical noise, because it yields, with minimum effort, sufficient information to solve most noise problems.

Most octave-band filter sets are fairly cumbersome and operate from the output of a sound-level meter, which serves as an acoustic pickup and high-impedance preamplifier. In field applications this arrangement is often inconvenient, owing to the necessity of handling two instruments. The new TYPE 1558 Octave-Band Noise Analyzer, weighing less than 9 pounds, includes the amplification and high input impedance needed for direct use with piezoelectric microphones. With its accessory TYPE 1560-P4 PZT Microphone Assembly, it indicates directly octave-band sound-pressure levels from 44 to 150 db *re* 2×10^{-4} μbar, a range that is adequate for the majority of uses. When the analyzer is operated from the output of a sound-level meter, however, lower levels can be measured.

The new analyzer is available in two models. The TYPE 1558-A is designed to meet the requirements of the current

Figure 1. View of the Type 1558-A Octave-Band Noise Analyzer with the Type 1560-P4 PZT Microphone Assembly.

ASA Specification for octave-band filter sets (Z24.10 1953) in all respects. The TYPE 1558-AP has bands centered at the ASA preferred frequencies for acoustical measurements (S1.6-1960).[1] Ten one-octave bands are included in each model together with an all-pass, or flat, characteristic. In addition, the A-model has a low-pass filter at 75 cps.

The electrical circuits are designed to reduce extraneous signals. Microphonics are held at a minimum through the use of transistors, rather than vacuum tubes, and pickup from external magnetic fields is avoided by RC-active filters, which contain no inductors.

The TYPE 1558 Octave-Band Noise Analyzer has a built-in, feedback-type calibration system for a simple check of the over-all electrical system. A dial setting renders the instrument direct reading in db *re* 2×10^{-4} μbar for piezoelectric microphones ranging in sensitivity from -52 to -62 db *re* one volt/μbar. Power is supplied by rechargeable nickel-cadmium batteries.

The analyzer joins the growing family of GR instruments packaged in the flip-tilt cabinet.

USES

Speech-Interference Level

The reactions of individuals to noise depend on many factors. Among these is the amplitude-frequency characteristic of the noise. It is well known, for example, that high sound-pressure levels in the octave bands 600–1200, 1200–2400, and 2400–4800 cps interfere with speech communication. The average of the band levels in db for these three bands is called speech-interference level, and for satisfactory intelligibility of difficult speech material this level should not exceed the values given in Table 1. Where speech cannot be heard, not only does efficiency suffer, but the danger of accidents is increased because shouted warnings may not be understood.

[1]Also specified by ISO Recommendation 402 and German Standard DIN45-401.

TABLE 1.

Speech-Interference Levels (db)

Distance (Feet)	Voice Level			
	Normal	Raised	Very Loud	Shouting
0.5	71	77	83	89
1	65	71	77	83
2	59	65	71	77
3	55	61	67	73
4	53	59	65	71
5	51	57	63	69
6	49	55	61	67
12	43	49	55	61
24	37	43	49	55

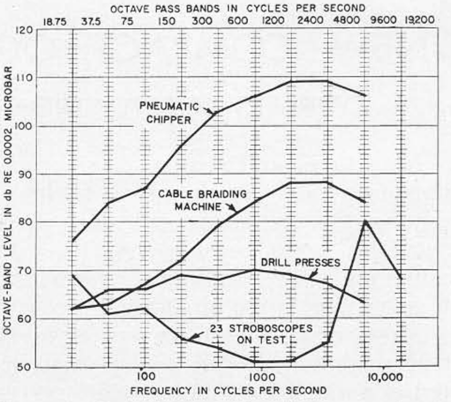


Figure 2. Typical octave-band pressure levels encountered in industry.

Hearing-Damage Risk

The TYPE 1558 Octave-Band Noise Analyzer is particularly suitable for determining the probability of hearing loss due to noise exposure. The noise spectra produced by several industrial activities are shown in Figure 2. While there are no standard methods for assessing the possibility of hearing damage in terms of octave-band analyses, it is generally accepted that the octave-band pressure level in any of the bands from 300 to 4800 cps should not exceed 85 db for daily exposure over a period of years. Thus, ear protection was prescribed for operators of the pneumatic chipper and cable-braiding machine of Figure 2.

Many organizations conduct periodic hearing tests and maintain records of noise exposure of their employees. Such a program is recommended wherever employees are exposed to high-level noise. A guide to recommended procedures is available.[2]

Office Noise

Difficulty in hearing speech can result in poor office efficiency. An octave-band analysis taken in an office allows the noise to be rated by use of a noise-criterion (NC) rating. Like speech-interference level, an NC level attempts to evaluate the noise environment by a single number. Figure 3 and Table 2

[2] "A Guide for Conservation of Hearing in Industry," Subcommittee on Noise of the Committee on Conservation of Hearing, American Academy of Ophthalmology and Otolaryngology. Available from the Research Center, Subcommittee on Noise of the American Academy of Ophthalmology and Otolaryngology, 327 S. Alvarado Street, Los Angeles 57, California.

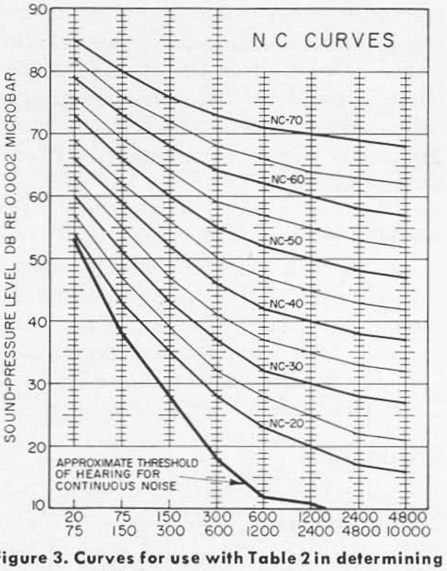


Figure 3. Curves for use with Table 2 in determining the permissible sound-pressure levels in eight octave bands.

TABLE 2. RECOMMENDED NOISE CRITERIA FOR OFFICES

Noise measurements made for the purpose of judging the satisfactoriness of the noise in an office by comparison with these criteria should be performed with the office in normal operation, but with no one talking at the particular desk or conference table where speech communication is desired (i.e., where the measurement is being made). Background noise with the office unoccupied should be lower, say by 5 to 10 units.

NC Curve of Figure 3	Communication Environment	Typical Applications
NC-20 to NC-30	Very quiet office — telephone use satisfactory — suitable for large conferences.	Executive offices and conference rooms for 50 people.
NC-30 to NC-35	"Quiet" office; satisfactory for conferences at a 15-ft table; normal voice 10 to 30 ft; telephone use satisfactory.	Private or semi-private offices, reception rooms, and small conference rooms for 20 people.
NC-35 to NC-40	Satisfactory for conferences at a 6- to 8-ft table; telephone use satisfactory; normal voice 6 to 12 ft.	Medium-sized offices and industrial business offices.
NC-40 to NC-50	Satisfactory for conferences at a 4- to 5-ft table; telephone use occasionally slightly difficult; normal voice 3 to 6 ft; raised voice 6 to 12 ft.	Large engineering and drafting rooms, etc.
NC-50 to NC-55	Unsatisfactory for conferences of more than two or three people; telephone use slightly difficult; normal voice 1 to 2 ft; raised voice 3 to 6 ft.	Secretarial areas (typing), accounting areas (business machines), blueprint rooms, etc.
Above NC-55	"Very noisy"; office environment unsatisfactory; telephone use difficult.	Not recommended for any type of office.

are based on the work of Beranek[3] and his associates. Measured octave-band levels are plotted on the graph, and the noise is rated in terms of the highest NC level reached in any octave band. The chart gives recommended criteria for various types of offices.

Loudness Level

Loudness level, a measure of the loudness of sounds, can be determined from the results of an octave-band analysis. It is a convenient single number, which agrees with subjective estimates of loudness.

Aircraft Noise

Recently a new method of rating aircraft noise has come into use.[4] This method rates the annoyance value rather than the loudness of a sound. The octave-band levels are weighted in a manner to give good correlation with listener judgment of the "noisiness" of both reciprocating-engine and jet aircraft as they pass overhead.

Vehicle Noise

Vehicle noise, especially truck noise, has become a serious problem on city streets and in residential areas. Some work has been done by truck manufacturers to ameliorate this situation, and there is now in effect an Automobile Manufacturers Association Specification which requires that noise levels be measured with an octave-band noise analyzer.

Vibration

Although the trend in vibration analysis is toward the use of an analyzer with narrower bands, such as the General Radio Type 1554-A Sound and Vibration Analyzer,[5] the Type 1558 Octave-Band Noise Analyzer with its low-frequency octave bands may also be found useful. The high input impedance of the octave-band noise analyzer permits direct connection of a piezoelectric vibration pickup.

Acoustical Characteristics of Structures

The sound transmission loss of walls, partitions, and floors can be determined with the octave-band noise analyzer and a wide-band sound source such as the General Radio Type 1390-B Random-Noise Generator.[6]

[3]Leo L. Beranek, "Criteria for Noise in Buildings," *Noise Control*, Vol 3, No. 1, pp 19-27, January, 1957.

[4]K. D. Kryter, "Scaling Human Reactions to the Sound from Aircraft," *Journal of the Acoustical Society of America*, Vol 31, Number 11, pp 1415 to 1429, November, 1959.

[5]J. J. Faran, "A New Analyzer for Sound and Vibration," *General Radio Experimenter*, 33, 12, December, 1959.

[6]A. P. G. Peterson, "A New Generator of Random Electrical Noise," *General Radio Experimenter*, 34, 1, January, 1960.

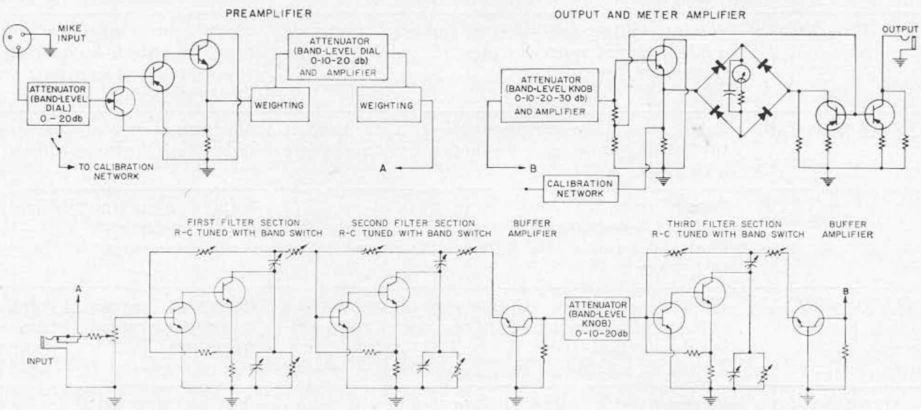


Figure 4. Schematic diagram of the Octave-Band Noise Analyzer.

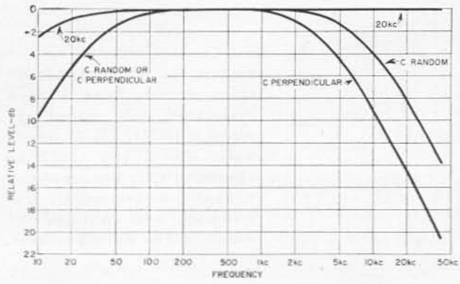


Figure 5. Frequency response characteristics of the preamplifier.

Non-Acoustical Uses

The Type 1558 Octave-Band Noise Analyzer can be used for purposes other than acoustical noise analysis. It is useful for measuring the electrical noise spectra generated by amplifiers, tape recorders, and other electronic devices. It can also function as a tuned voltmeter or a selective amplifier.

CIRCUIT

The Type 1558 Octave-Band Noise Analyzer consists of a high-impedance microphone preamplifier, a tunable filter having a noise bandwidth of one octave, an output amplifier, and a level indicator. An elementary schematic diagram is shown in Figure 4.

Preamplifier Section

The preamplifier section includes an input attenuator, a high-input-impedance, unity-gain amplifier, a weighting network, and a second attenuator and amplifier. It has a maximum mid-frequency voltage gain of 20 db, and its amplitude-frequency characteristic can be set by means of an internal switch to be either essentially flat from 20 cps to 20 kc or C-weighted (see Figure 5). Both attenuators in this section are controlled by the large outer dial of the coaxial BAND LEVEL control. This dial is used to adjust the gain of the preamplifier in 10-db steps to suit the over-all amplitude of the signal being analyzed.

Filter Section

The filter is synthesized as an isolated cascade of three resonant sections. Between the second and third sections is a 20-db step attenuator. The resonant frequencies of the sections are staggered around the center frequency of the selected band to give a Butterworth, or "maximally flat," characteristic. Each filter section uses a highly stabilized current amplifier and an RC feedback network. Both resistors and capacitors in the feedback network are switched in a manner which allows each capacitor set to be used for two bands. Figure 6 is a functional diagram of a single section.

Figure 6. Functional diagram of a single section of the filter.

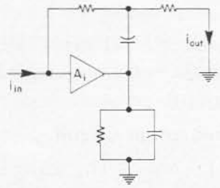


Figure 7 shows the band-pass characteristic of the filters in the -A model. The -AP model has identical characteristics, but different center frequencies.

Output Amplifier and Meter Section

The output amplifier section consists of a 30-db step attenuator, an amplifier, a detector, and a meter. An isolating stage for the output terminals prevents any load from affecting the meter indication. The detector characteristic is quasi-rms[7] so that the meter indication is very closely rms for most types of signals.

The meter is identical to the one used in the Type 1551-C Sound-Level Meter[8] and therefore has the dynamic characteristics specified by ASA Specification for General Purpose Sound Level Meters

[7] E. E. Gross, "Improved Performance Plus a New Look for the Sound-Level Meter," *General Radio Experimenter*, 32, 10, October, 1958.

[8] E. E. Gross, "Type 1551-C Sound-Level Meter," *General Radio Experimenter*, 35, 8, August, 1961.

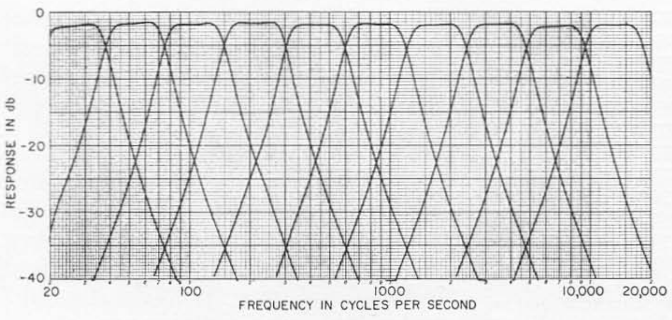


Figure 7a. Filter characteristics of the Type 1558-A measured with signal applied at INPUT (SLM) terminals. The Type 1558-AP characteristics are identical, except that the center frequencies are changed, as noted in the Specifications.

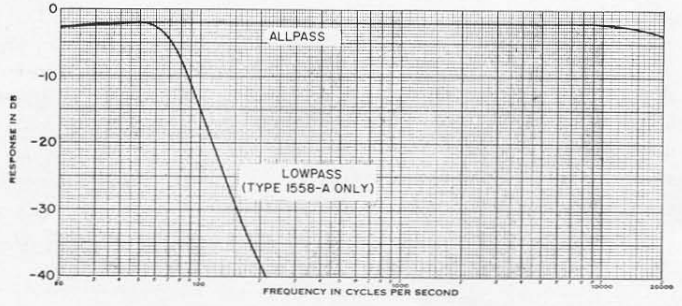


Figure 7b. Lowpass and allpass characteristics of the Analyzer.

(ASA S1.4-1961). Fast or slow meter speeds can be selected by a panel control.

Calibration Circuit

To check the gain of the analyzer, its output is switched to its input through a filter limiter and calibrated attenuator. When the gain is adjusted to equal the known attenuation of this feedback network, the system oscillates. The attenuation of the feedback network is adjustable by means of an internal control, which is calibrated in terms of microphone sensitivity.

Charging Circuit

The battery, a sealed nickel-cadmium unit, is charged through a simple half-wave rectifier and series resistor, which connect directly to the power line. During the charge period, the battery floats on the line. Neither side of the line is connected to the case nor to any other part of the instrument except the charging circuit.

[8] Ibid.

TYPE 1560-P4 PZT MICROPHONE ASSEMBLY

The Type 1560-P4 PZT Microphone Assembly was designed for use with the Type 1558 Octave-Band Noise Analyzer. The microphone is a PZT piezoelectric ceramic type, identical to that supplied with the Type 1551-C Sound-Level Meter and described in detail in a previous article.[8]

As shown in Figure 8, the frequency response of this microphone to sounds of random incidence is virtually flat to 8 kc. Its temperature coefficient of sensitivity is very low (−0.01 db/°C)

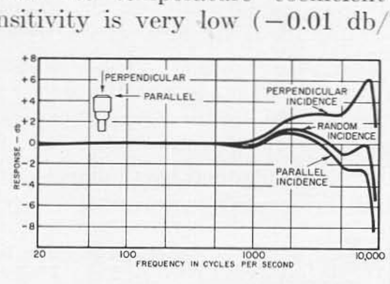


Figure 8. Response of the Type 1560-P4 PZT Microphone Assembly.

and its impedance is nearly independent of temperature. Designed to be durable and dependable, this microphone will withstand, without damage, temperatures of −30 to +95C and relative humidity up to 100%.

The microphone is mounted on one end of a flexible conduit. A detented swivel connector on the other end of the conduit plugs into a receptacle on the panel of the instrument.

— W. R. KUNDERT

SPECIFICATIONS

Bands: Type 1558-A

Lower Cutoff Frequency - cps	*Upper Cutoff Frequency - cps*	*Center Frequency*- cps*
18.75	37.5	26.5
37.5	75.0	53.0
75.0	150	106
150	300	212
300	600	424
600	1200	848
1200	2400	1696
2400	4800	3392
4800	9600	6784
9600	19,200	13,570
LP	75	
ALL PASS		

*Geometric Mean

For Type 1558-AP center frequencies are 31.5, 63, 125, 250, 500, 1000, 2000, 4000, 8000, 16,000.

Filter Characteristics*: Level at center frequency in bands from 37.5 to 9600 cps is uniform within 1 db. Maximum deviation from ALL PASS level at center frequency in any band is 1 db. For bands from 37.5 to 9600 response at nominal cutoff frequency is (3.5 ± 1) db below response at center frequency. Attenuation is at least 30 db at one-half the lower nominal cutoff frequency and twice the upper nominal cutoff frequency for all octave bands. Attenuation is at least 50 db at one-fourth the lower nominal cutoff frequency and four times the upper nominal cutoff frequency for all octave bands. The 75-cycle low-pass filter has at least 30-db attenuation at 200 cps and at least 50-db attenuation at 400 cps.

Sound-Pressure Level Range: 44 to 150 db above 0.0002 μbar in any band when the TYPE 1560-P4 PZT Microphone Assembly is used.

Inputs: Impedance at MIKE terminals is approximately 50 pf in parallel with 50 MΩ. It is intended for use with high-impedance transducers.

*Measured with signal applied at INPUT (SLM) terminals.

Impedance at INPUT (SLM) terminals is approximately 100 kΩ. Maximum input is 3 volts. Low-input terminal is connected to case. This input is intended for connection to the output of a sound-level meter.

Preamplifier Frequency Characteristics: Two frequency characteristics are available. These are C-weighting, which is specified by the American Standards Association (ASA S1.4-1961 SLM), and 20 kc, an essentially flat response.

Outputs: Open-circuit output is at least 1 volt for full-scale meter indication. Output impedance is 6000 ohms. Any load can be connected across the OUTPUT terminals.

Meter Response: FAST or SLOW meter response is selected by panel control. These characteristics are as specified in the American Standard Specification for General Purpose Sound Level Meters, ASA S1.4-1961. Meter indication is closely rms for most waveforms.

Internal Calibration: A built-in reference allows the gain of the analyzer to be calibrated for use with piezoelectric microphones having sensitivities from −52 to −62 db *re* 1 v/μbar. The absolute accuracy for ALL PASS is then ensured within 1 db over a wide range of atmospheric conditions.

Batteries: Two 9.6-volt rechargeable nickel-cadmium batteries (Gould Type 9.6V/450B) give 30 hours operation. They are recharged by connection to a 115-v (or 230-v) 25- to 60-cycle power line. Full charge takes about 14 hours.

Accessories Supplied: Carrying strap, power cord, shielded cable assembly.

Accessories Available: TYPE 1560-P4 PZT Microphone Assembly.

Dimensions: Flip-tilt case: length 10¼, height 9¼, depth 7¼ inches (260 by 235 by 185 mm), over-all, including handle.

Net Weight: 8¾ pounds (4 kg).

Type		*Code Word*	*Price*
1558-A	**Octave-Band Noise Analyzer**	ABATE	**$725.00**
1558-AP	**Octave-Band Noise Analyzer**	ALARM	**725.00**
1560-P4	**PZT Microphone Assembly**	NAVAL	**80.00**

U.S. Patent Nos. 3,012,197; 2,966,257; and D187,740.

THE TYPE 1211-C, AN IMPROVED UNIT OSCILLATOR

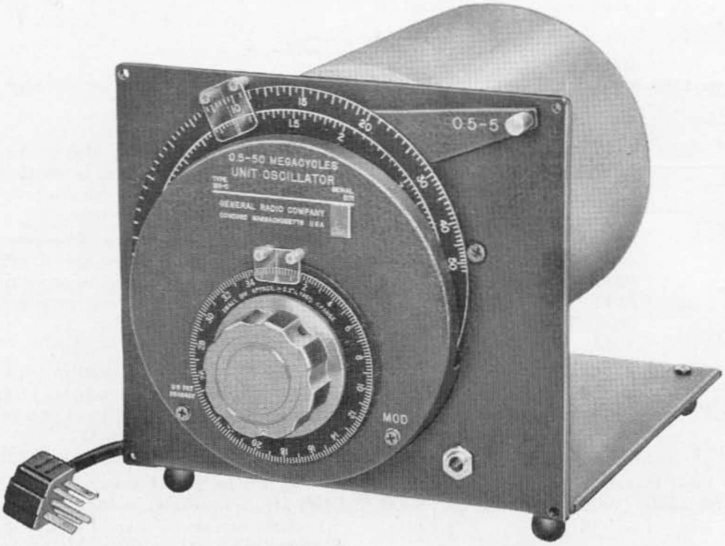


Figure 1. View of the Type 1211-C Unit Oscillator.

The TYPE 1211 Unit Oscillator,[1] 0.5 to 50 Mc, which was first announced in September, 1953, has now been modified as a first step in a program of Unit-Oscillator redesign which should make these popular rf power sources even more useful in the future. The characteristic L-shaped mounting panel of the Unit Oscillators and their open construction, which give excellent shielding and heat dissipation at a minimum price, have been retained, but the panel of the new oscillator, TYPE 1211-C, has been reduced to 7-inch height (four standard relay-rack units). Panel width is 8 inches, corresponding to one half relay-rack width, for mounting side by side in a relay rack with other GR half-rack instruments. Mounting in a relay rack is by means of simple, inexpensive adaptor panels (see Figure 2).

Frequency range, input power requirements, and output power remain es-

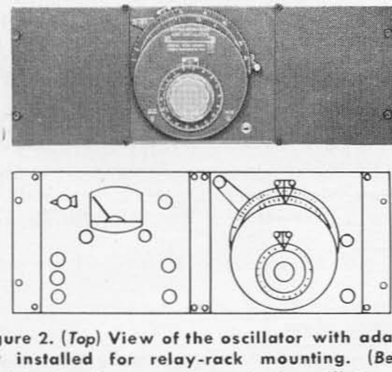


Figure 2. (*Top*) View of the oscillator with adaptor set installed for relay-rack mounting. (*Below*) Sketch showing arrangement of oscillator and amplitude-regulating power supply for relay-rack mounting.

[1]A. G. Bousquet, "A Unit Oscillator for the 0.5- to 50-Mc Range," *General Radio Experimenter*, 28, 4, September, 1953.

sentially unchanged, but distortion has been reduced considerably. The 100-to-1 frequency range of this oscillator is covered in two 10-to-1 ranges, 0.5 to 5 Mc and 5 to 50 Mc. The required 100-to-1 variation in the LC product on each range is obtained by changing the inductance as well as the capacitance of the tuned circuit. The capacitance is varied from 20 to 800 pf (40-to-1), and simultaneously the inductance value is altered by sickle-shaped cores, mounted on the capacitor shaft (see Figure 3). One core is made of aluminum, the other of iron dust. As the frequency dial is

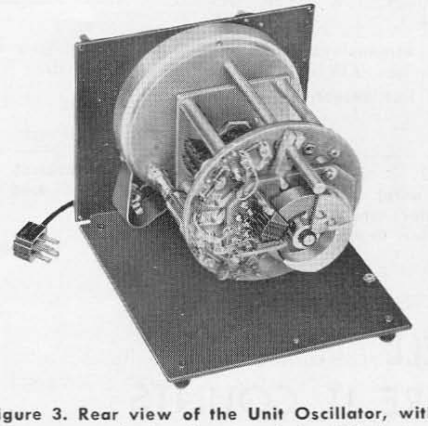
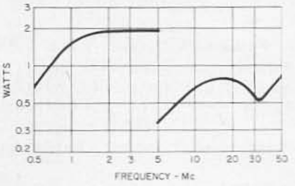


Figure 3. Rear view of the Unit Oscillator, with cover removed, showing tuned circuit inductors and sickle-shaped cores.

rotated, the active core material within the inductors varies smoothly from dust core for maximum inductance to a full aluminum core for minimum inductance. A 2.5-to-1 change in inductance is realized, from 125 to 50 μh for the 0.5- to 5-Mc range and from 1.25 to 0.5 μh for the 5- to 50-Mc range. The cores and the capacitor plates are shaped for logarithmic frequency change with angular rotation.

Figure 4. Typical output curves of the Type 1211-C Unit Oscillator.



Power output over the frequency range varies approximately as shown in Figure 4, when the inexpensive Type 1203-B Unit Power Supply is used. With the plate-regulated Type 1201-B Unit Regulated Power Supply, the output frequency as well as the output power is stabilized against line-voltage change, but output is reduced to about three quarters of that shown.

The frequency dial of the Type 1211-C Unit Oscillator can be swept back and forth mechanically by the Type 1750-A Sweep Drive,[2] the Type 908-P Synchronous Dial Drives, or the Type 908-R96 Dial Drive, and constant output over the frequency ranges can be obtained with the Type 1263-B Amplitude-Regulating Power Supply.[3] The combination of Unit Oscillator, Sweep or Dial Drive, and Amplitude-Regulating Power Supply is used for recording or for oscillographic display of frequency characteristics (see cover photograph). The Type 1263-B Amplitude-Regulating Power Supply will also be found useful for manual operation of the Unit Oscillator.

— E. Karplus

[2] W. F. Byers, "A New System for Automatic Data Display," *General Radio Experimenter*, 24, 11, April, 1955.

[3] W. F. Byers, "Type 1263-B Amplitude-Regulating Power Supply," *General Radio Experimenter*, 35, 9, September, 1961.

SPECIFICATIONS

FREQUENCY

Range: 0.5 to 50 Mc in two ranges.

Calibration Accuracy: ±2 percent at no load.

Warmup Drift: 0.4% ± 0.2%, largest at the high-frequency end of each range.

Controls: A two-position range switch, a six-inch dial with approximately logarithmic calibration, and a slow-motion dial to indicate frequency increments of 0.2 percent per dial division.

SPECIFICATIONS (Continued)

OUTPUT

System: Output available at a TYPE 874 Coaxial Connector (locking) at rear of instrument. Adjacent ground terminal also permits connection by TYPE 274-M Double Plug. Output is controlled by a 250-ohm resistive voltage divider. The dial is calibrated in 100 arbitrary units.

Power: With the TYPE 1203-B Unit Power Supply, at least 200 milliwatts into 50-ohm load at any frequency. Over the 0.5- to 5-Mc range, average output is approximately 1 watt; over the 5- to 50-Mc range, 0.4 watt. See Figure 4 for typical output characteristics.

GENERAL

Circuit: Hartley oscillator coupled directly to output. Capacitance and inductance are simultaneously changed for frequency variation.

Modulation: Plate modulation of 30% at audio frequencies can be produced by external source of 50 volts. Input impedance is about 8000 ohms. For amplitude modulation free from incidental fm, a TYPE 1000-P6 Crystal Diode Modulator can be used at carrier frequencies above 10 Mc.

Power Supply Requirements: 320 volts, 50 milliamperes, dc; 6.0 volts, 0.75 ampere, ac or dc. TYPE 1203-B Unit Power Supply, TYPE 1201-B Unit Regulated Power Supply, or TYPE 1263-B Amplitude-Regulating Power Supply is recommended.

Mounting: Oscillator on aluminum casting is shielded with a spun aluminum cover; assembly is mounted on an L-shaped panel and chassis. Adaptor panels for relay-rack mounting are available.

Accessories Supplied: TYPE 874-R22 Patch Cord, TYPE 874-Q2 Adaptor, telephone plug.

Other Accessories Available: TYPE 1750-A Sweep Drive, TYPE 908 Dial Drives, TYPE 874 Coaxial Elements, TYPE 1000-P6 Crystal Diode Modulator, TYPE 480 Relay-Rack Adaptor Sets.

Dimensions: Width 8, height 7½, depth 12 inches (205 by 192 by 305 mm), over-all.

Net Weight: 11½ pounds (5.5 kg).

Type		*Code Word*	*Price*
1211-C	**Unit Oscillator**	ATLAS	**$305.00**
480-P408	**Relay-rack Adaptor Set (for oscillator only)**	EXPANELJAG	**8.00**
480-P416	**Relay-rack Adaptor Set (for oscillator and Type 1263-B Amplitude-Regulating Power Supply)**	EXPANELNIT	**6.00**

U.S. Patent No. 2,548,457.

USING A PHOTOCELL WHERE IT COUNTS

For those wishing to measure the speed of rotating objects and to present the results as a continuous digital display, we recommend the combination of the new TYPE 1536-A Photoelectric Pickoff and the TYPE 1150-A Digital Frequency Meter.[1]

The pickoff consists of a light source, an optical system, a photocell, an output cable, and a flexible linkage system. Light from the source is reflected, either by the rotating object or by reflective tape attached to it, back to the photocell, which sends electrical pulses to the

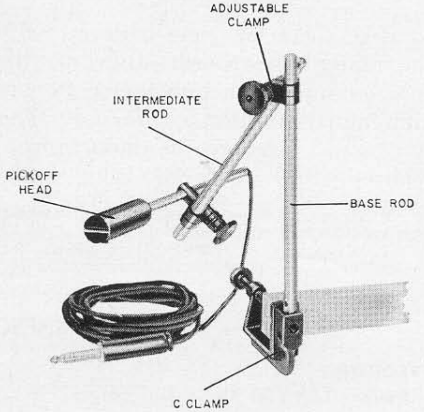


Figure 1. View of the Photoelectric Pickoff with component parts identified.

[1]R. W. Frank, J. K. Skilling, "A Five-Digit Solid-State Counter for Frequency Measurements to 220 kc," *General Radio Experimenter*, 36, 4, April, 1962.

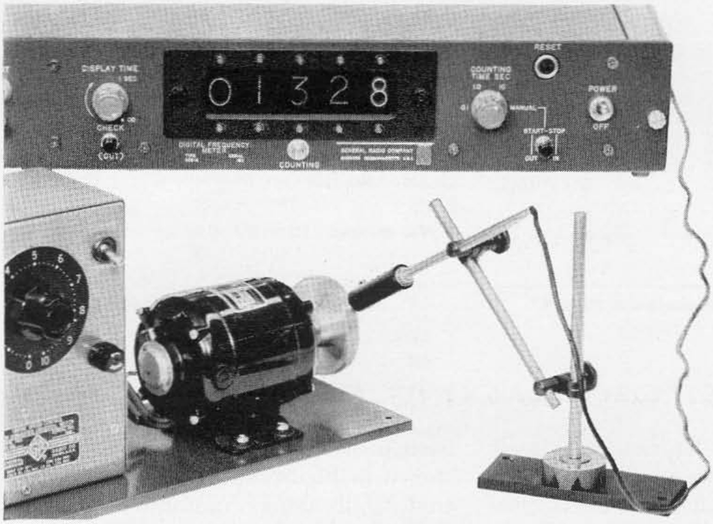


Figure 2. The Photoelectric Pickoff shown with the Type 1150-A Digital Frequency Meter (*top*), as arranged to measure the speed of an electric motor.

frequency meter. This instrument counts the number of pulses arriving per second (or 0.1 second or 10 seconds) and displays that number on an in-line digital readout.

The cylinder containing the photocell and light source must be placed fairly close to the object being observed. The maximum distance depends on the contrast between the reflective and nonreflective parts of the rotating object. The small size of the pickoff head and the double-jointed linkage assembly, mounted on either a C-clamp or a magnet (both supplied), permit the pickoff to be maneuvered close enough to out-of-the-way rotating parts.

With the counter set for a one-second gate (i.e., counting) period, the digital display will be in revolutions per second. For greater accuracy, the counting period can be set to 10 seconds, and the digital readout divided by 10. By obtaining more than one pulse per revolution (as, for instance, by attaching more than one reflecting strip to the rotating surface), one can increase the display possibilities: With six reflective strips and a 10-second counting period, the counter indicates rpm. If 60 strips can be attached, one can obtain a direct rpm statement once a second. As more strips are used, the pickoff must be placed nearer to the object.

Most machine speeds are well within the range of the pickoff-counter combination. The high-frequency limit of the counter is over 13 million rpm, so there is no problem from that quarter. The limiting factor is usually the capacitance of the cable connecting the pickoff to the counter. Under favorable conditions, speeds up to 100,000 rpm can be measured.

Two rolls of pressure-sensitive tape are supplied, one reflecting and one nonreflecting. The latter can be used with objects that are themselves highly reflective.

Uses of the Type 1536-A Photoelectric Pickoff cover almost all rotating machinery, but it is especially desirable for low-torque devices, to which mechanical contactors cannot be attached.

SPECIFICATIONS

Light Source: GE Type 327 bulb, 28 volts, 40 milliamperes.

Power Supply: Power is supplied for both lamp and photocell by the TYPE 1150-A Digital Frequency Meter.

Accessories Supplied: 10-ft roll of $\frac{3}{8}$-inch black tape; 10-ft roll of $\frac{3}{8}$-inch silver tape; carrying case.

Mounting: C-clamp (capacity $1\frac{5}{16}$ inches, flat or round) or $1\frac{1}{2}$-inch magnet, both supplied.

Dimensions: Pickoff head, $\frac{11}{16}$-inch diameter, 2 inches long. Linkage consists of two $\frac{5}{16}$-inch-diameter stainless-steel rods, 6 and $6\frac{1}{4}$ inches long, connected by an adjustable clamp. Second clamp attaches pickoff assembly. Cable (pickoff to counter) is 8 feet long, terminated in phone plug.

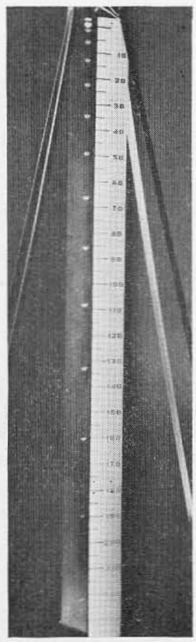
Net Weight: 18 ounces (0.6 kg).

Type		*Code Word*	*Price*
1536-A	**Photoelectric Pickoff**..................	FOTOF	**$65.00**

STROBO-TRACKING GRAVITY

The constant acceleration of a freely falling object is nowhere more graphically demonstrated than at Los Angeles' Occidental College, where a TYPE 1531-A Strobotac® Electronic Stroboscope[1] is used in a classroom demonstration. As shown in Figure 1, a ball is dropped so that it falls along a uniformly calibrated scale. During its fall the ball is photographed by the light of the stroboscope, flashing at a constant rate. The result is the multiple-image photograph of Figure 2, in which the ball is shown at fixed

[1] M. J. Fitzmorris, C. J. Lahanas, and W. R. Thurston, "New Eyes for Modern Industry," *General Radio Experimenter*, 34, 9, September, 1960.



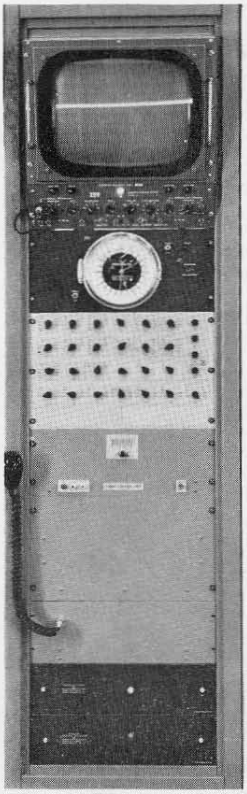
(*Left*) Figure 1. View of the classroom demonstration, with ball about to be dropped.
(*Right*) Figure 2. Photographic record of the travel of the ball.

time intervals on the way down. The widening gaps between images as the ball drops proves the rule of constant acceleration to even the most skeptical sophomore.

The setup at Occidental uses simple equipment, and is easily duplicated. The Polaroid camera uses Polaroid Type 46-L transparency film, and the transparency is projected for class analysis within two minutes after exposure.

Our thanks to Rex R. Nelson, Assistant Professor of Physics at Occidental, for telling us about his interesting demonstration.

GENERAL RADIO AT SEATTLE WORLD'S FAIR



(*Right*) Group of fairgoers listening attentively to the satellite-tracking demonstration. (*Left*) Close-up view of the center rack. The Syncronometer with its 24-hour clock face is near the top; the standard-frequency oscillator and frequency divider are at the base.

The accompanying photographs show the Transit Satellite Demonstration Tracking Station used in the U. S. Science Exhibit at the Seattle World's Fair. The frequency standard for the station is a General Radio Type 1113-A Standard-Frequency Oscillator, shown at the bottom of the center rack with its companion, the GR Type 1114-A Frequency Divider. A third member of

GR's frequency-measuring lineup, the TYPE 1103-B Syncronometer, is seen higher up in the same rack.

The 5-Mc output of the TYPE 1113-A is multiplied up into the UHF region for use as a local oscillator for precision doppler receivers. A multichannel tape recorder records satellite signals along with a 50-kc reference signal from a generator driven by the TYPE 1114-A Frequency Divider. The Syncronometer provides an accurate time display for audience and demonstrators, and serves as a time reference for recorded doppler data.

The satellite tracking demonstration is one of the most popular displays at the fair. The right-hand photograph shows a group of fairgoers paying earnest attention to a tracking demonstration. This demonstration was designed and built by the Applied Physics Laboratory at Johns Hopkins University.

General Radio Company

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THE GENERAL RADIO
EXPERIMENTER

VOLUME 36 No. 11

NOVEMBER, 1962

IN THIS ISSUE

New —

Vibration Pickup System

Rack-Mounted Impedance Bridge

Dallas Sales Engineering Office

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EXPERIMENTER

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CONTENTS

COVER

The light pattern of a row of Strobotac® electronic stroboscopes under test in the GR lab was too much for our photographer (and editor) to pass up. Harry Chisholm, Electrical Inspection Supervisor, is shown adjusting flashing rate.

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GENERAL RADIO COMPANY (OVERSEAS), ZURICH, SWITZERLAND
REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

NEW PZT CERAMIC VIBRATION PICKUP AND CONTROL BOX FOR VIBRATION MEASUREMENTS

A new lead-zirconate-titanate accelerometer replaces the barium-titanate ceramic accelerometer as the general-purpose vibration pickup supplied with the General Radio Company's vibration-measuring instruments. This new transducer increases the upper frequency limit of the TYPE 1553-A[1] Vibration Meter from 1200 cps to 2000 cps. A newly designed control box provides a like increase in the frequency response plus an increase in over-all measuring sensitivity when the pickup is used with a TYPE 1551-C or -B Sound-Level Meter. This new control box is so designed that it can readily be adapted for use with other piezoelectric accelerometers. One such adaptation is for use with the Endevco Model 2217 Accelerometer to meet the frequency-response requirements of Mil-Std-740 (SHIPS).

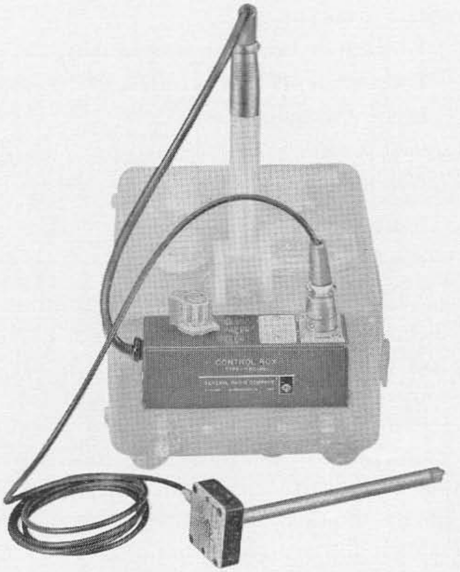


Figure 1. View of the Type 1560-P11B Vibration Pickup System, shown with the Type 1551-C Sound-Level Meter. The control box attaches to the sound-level-meter case.

The Pickup

The new TYPE 1560-P52 Vibration Pickup replaces the TYPE 1560-P51 model.[2] Table 1 compares the characteristics of the new and old units. Compared with its predecessor the new pickup has, in addition to increased sensitivity and increased frequency response, a lower impedance, a wider operating temperature range, and better stability.

Its high sensitivity and low impedance make it an outstanding unit for low-frequency vibration measurements. Frequency response is flat to below 2 cps without special high-impedance preamplifiers. High-gain amplifiers are not required for many common vibration measurements.

The Control Box

The TYPE 1560-P21B Control Box illustrated in Figure 1 has been designed to match the TYPE 1560-P52 Vibration Pickup to the input of the TYPE 1551-C or -B Sound-Level Meter. The combination of pickup and control box is listed

[1]E. E. Gross, "TYPE 1553-A Vibration Meter," *General Radio Experimenter*, 35, 11, November, 1961.

[2]E. E. Gross, "TYPE 1560-P11 Vibration Pickup System," *General Radio Experimenter*, 34, 11 & 12, November-December, 1960.

TABLE 1

	Old	*New*
Type No.:	1560-P51	1560-P52
Material:	Barium Titanate	Lead Zirconate (PZT)
Sensitivity (mv/g):	40	75
Resonant Frequency (cps):	2300	3200
Capacitance (pf):	7000	10,000
Max Acceleration (g):	100	100
Temperature Coefficient of Sensitivity (db/°F):	0.03	0.03
Temperature Range (°F):	0 to +180	−30 to +200
Relative Humidity Range (%):	0 to 100	0 to 100
Cable Length:	5 ft (1.55 m)	5 ft (1.55 m)
Weight:	1.6 oz (45 g)	1.6 oz (45 g)
Pickup Dimensions:	1⅝ x 1 7/16 x 9/16 in. (42 x 37 x 15 mm)	1⅝ x 1 7/16 x 9/16 in. (42 x 37 x 15 mm)

as the TYPE 1560-P11B Vibration Pickup System and replaces the TYPE 1560-P11 Vibration Pickup System. The control box is designed to operate with the sound-level-meter weighting switch at 20 kc and the calibration control set for a −60 db (*re* 1 volt/μbar) microphone. Because the control box is designed for a fixed sound-level-meter sensitivity, the calibration adjustment (variable attenuation) required in the control box is reduced. This reduced attenuation requirement and the increased pickup sensitivity have made it possible to increase the over-all measurement sensitivity by 10 db, as illustrated in Table 2.

TABLE 2

db Conversion Factors for Old and New Vibration Pickup Systems

Vibration Quantity	*Sound-Level-Meter reading in db with:* Type 1560-P11	*Sound-Level-Meter reading in db with:* Type 1560-P11B
Acceleration 1 in/sec²	40	50
Velocity 1 in/sec	80	90
Displacement 1 in	110	120

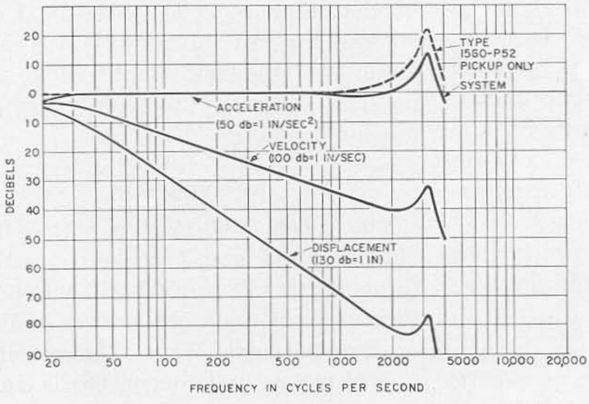


Figure 2. Frequency-response characteristics of the Type 1560-P11B Vibration Pickup System for constant applied acceleration.

The frequency-response characteristics of the TYPE 1560-P11B Vibration Pickup System in combination with the TYPE 1551-C or -B Sound-Level Meter and the response of the TYPE 1560-P52 Vibration Pickup are shown in Figure 2.

Calibration

The electrical response of the circuits in the control box is measured over the frequency range of 20-5000 cps and the response of each pickup is measured over the frequency range of 10 to 5000 cps. Absolute sensitivity of the pickup is then determined at a low frequency by accurate measurement of the displacement of a calibrating shaker by means of a microscope and a Strobotac® electronic stroboscope. Finally, an over-all operating test of the combination is performed at 100 cps with a standardized TYPE 1557-A Vibration Calibrator.[3]

[3]E. E. Gross, "Little Dithers," *General Radio Experimenter*, 34, 11 & 12, November-December, 1960.

SPECIFICATIONS

(*See Tables 1 and 2*)

Type		*Code Word*	*Price*
1560-P11B	**Vibration Pickup System**	PIKUP	**$140.00**

MIL-STD-740 (SHIPS)

While the TYPE 1560-P11B combination of pickup and control box meets the requirements of most vibration-measurement problems, there are some specialized measurements that require a flat response characteristic to higher frequencies. For these, the TYPE 1560-P11S2 combination is recommended, consisting of the Endevco Model 2217 Accelerometer and the TYPE 1560-P21S1 Control Box. A small holding magnet, TYPE 1560-4020, is included. (See Figure 3.)

Figure 3. View of the Type 1560-P11S2 Vibration Pickup System with the Type 1551-C Sound-Level Meter.

This system with the TYPE 1551-C or -B Sound-Level Meter provides the flat frequency response and low-noise operation required by Mil-Std-740 (SHIPS) for vibration measurement.

The response characteristics of this special control box, an Endevco pickup, and a TYPE 1551-C or -B Sound-Level Meter are shown in Figure 4. The curves show that the electrical system does not modify the pickup response for acceleration and provides true integration for velocity measurements to well beyond 10 kc. The two integrators for displacement measurements perform to 3000 cps. The limit here is determined by the internal noise level of the system. The decibel conversion factors for this system are:

Acceleration — 1 in/sec^2 = 40 db
Velocity — 1 in/sec = 90 db
Displacement — 1 in = 120 db

These are the same numbers, often

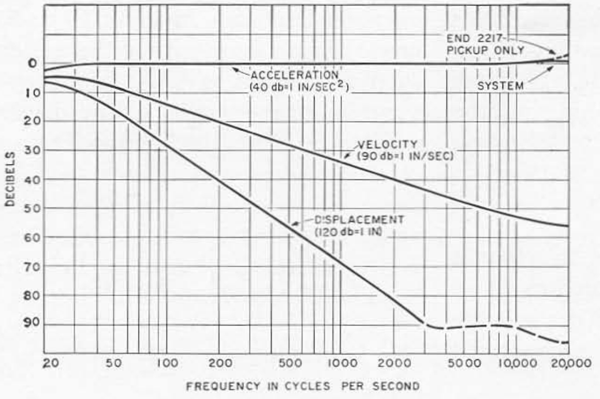


Figure 4. Frequency-response characteristics of the Type 1560-P11S2 Vibration Pickup System for constant applied acceleration. Electrical responses of the control box were measured with a source impedance equivalent to the Endevco Model 2217 plus its connecting cable. The Endevco 2217 response is as specified by the manufacturer.

called Chapman Numbers,[4] that applied to an earlier vibration pickup system using a Rochelle-salt-crystal pickup and used by the Navy for shipboard noise-control measurements.

The addition of the Type 1554-A Sound and Vibration Analyzer[5] to the vibration-measuring system permits measurements of very low level accelerations. The analyzer and the Type 1521-A Graphic Level Recorder[6] make it possible to record the vibration spectrum, either by a continuous narrow-band analysis or a continuous one-third-octave-band analysis.

— E. E. Gross

[4] Robert York Chapman, "Electronic Instrumentation for Submarine Auxiliary Machinery Noise and Vibration Control," *Technical Report No. 210-61*, Acoustical Research and Development Division, Code 380, Engineer and Repair Department, U. S. Naval Submarine Base, New London, Groton, Connecticut.

[5] J. J. Faran, "A New Analyzer for Sound and Vibration Measurement," *General Radio Experimenter*, 33, 12, December, 1959.

[6] M. C. Holtje, M. J. Fitzmorris, "A Graphic Level Recorder with High Sensitivity and Wide Ranges," *General Radio Experimenter*, 33, 6, June, 1959.

SPECIFICATIONS

Type No.:	Endevco 2217
Material:	Ceramic
Sensitivity (mv/g):	72
Resonant Frequency (cps):	35,000
Capacitance (pf):	350
Max Acceleration (g):	1000
Operating Temperature Range (°F):	−65 to +250
Temperature Coefficient of Sensitivity (db/°F):	< 0.01
Relative Humidity Range (%):	0 to 100 (Hermetically sealed)
Cable Length:	8 ft, 2 in. (2.5 m)
Weight:	1.1 oz (31 g)
Pickup Dimensions:	⅝ hex x 0.70 in. (15.5 dia x 18 mm)

Type		*Code Word*	*Price*
1560-P11S2	**Vibration Pickup System***	PILOT	**$355.00**

*Available on special order.

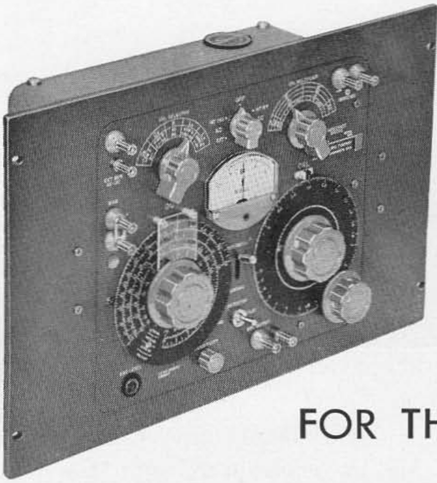


Figure 1. The new relay-rack model of the Type 1650-A Impedance Bridge.

RELAY-RACK MOUNT FOR THE IMPEDANCE BRIDGE

The General Radio Flip-Tilt Case is used on instruments where portability is an important requirement. Prominent among these is the Type 1650-A Impedance Bridge.[1] It is easily carried, is amply protected during transport, and opens easily for operation with the instrument panel held at any desired angle. However, for some applications, as, for instance, production-line test assemblies, relay-rack mounting is desirable, and we are now prepared to supply this versatile bridge on a 19-inch relay-rack panel, as shown in Figure 1.

An adaptor panel provides the conversion to relay-rack mounting. The cover and handle are removed and a heavy, charcoal-gray, crackle-finish, aluminum panel, with suitable cutout for the instrument, is attached to the instrument by two angle brackets. These brackets make use of the same hardware as does the carrying handle for fastening to the instrument case.

[1]Henry P. Hall, "A Universal Impedance Bridge," *General Radio Experimenter*, 33, 3, March, 1959.

SPECIFICATIONS

Dimensions: Panel, 19 by 12¼ inches (485 by 315 mm); depth behind panel, 5 inches (130 mm).

Net Weight: 17¾ pounds (8.5 kg).

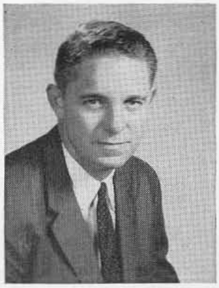
Electrical specifications are identical with those for the portable model, Type 1650-A.

Type		*Code Word*	*Price*
1650-AR	**Impedance Bridge**	BEFOG	**$460.00**

U.S. Patent No. 2,872,639 and 2,966,257.

Figure 2. Flip-tilt case combines ruggedness with convenience.

GENERAL RADIO COMES TO TEXAS



Ed Sutherland

Shirley Redfield

With appropriate pride, we announce the opening of a new sales engineering office in Dallas, Texas. Territory covered by the new office, at 2501-A West Mockingbird Lane, includes all of Texas (except El Paso), Louisiana, Mississippi, Oklahoma, Arkansas, and Colorado.

Manager of the Dallas Office is Edward F. Sutherland, who for the past four years has been a member of our New York sales engineering staff. Ed is eminently qualified for his new assignment: As an electronics engineer (Cornell) with six years' experience at General Radio, he offers expert advice on measurement problems to our Texas-area customers; as one born in Texas (Houston), he speaks as a native son; and, at six-foot-seven, Ed can not only talk to Texans, but can look most of them in the eye.

Our Chicago office reluctantly gives up its senior secretary, Miss Shirley Redfield, to Dallas. Shirley, with almost 20 years at GR, is well known in the Chicago area as a member of the Executive Committee of the Chicago IRE Section.

Dallas is the latest extension of GR's traditional policy of serving customers directly. Our first district sales office was opened in 1934 in New York, and there are now 11 such offices spread from Toronto to Los Angeles. These offices are manned by engineers whose considerable knowledge of electrical measurements is yours for the telephoning. And of course that goes double in Big D, where the number is FLeetwood 7-4031 (code 214).

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THE GENERAL RADIO

EXPERIMENTER

VOLUME 36 No. 12

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IN THIS ISSUE

New

500-Mc Frequency Converter

Washington Service

Counter Errors

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CONTENTS

COVER

Development Engineer H. T. McAleer measures the frequency stability of a Type 1215-B Unit Oscillator at 250 Mc with the Type 1130-A Digital Time and Frequency Meter and the Type 1133-A Frequency Converter.

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GENERAL RADIO COMPANY (OVERSEAS), ZURICH, SWITZERLAND
REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

A NEW CONVERTER FOR FREQUENCY MEASUREMENTS TO 500 Mc

The introduction of our 10-Mc counter, the TYPE 1130-A Digital Time and Frequency Meter,[1] has stimulated requests that we produce a companion instrument to extend the frequency-measurement range. Several such instruments already exist, but we felt that we could contribute significant improvements in sensitivity, selectivity, and ease of use. Figure 1 shows the results of our efforts, the TYPE 1133-A Frequency Converter.

The converter heterodynes an unknown input frequency between 10 and 500 Mc against a 10-Mc multiple of a standard frequency, derived from the 5-Mc time-base oscillator of the counter, and applies the less-than-10-Mc difference frequency to the counter. The instrument can also amplify weak signals between 100 kc and 10 Mc to operate the counter.

Operation of the converter is simple and straightforward. The heterodyne reference frequency to be added to the counter reading is indicated directly by large in-line numerals. The range of output amplitude acceptable for the counter is clearly indicated on the panel meter, and adjustment to this level is made by an output control. An input sensitivity control is also provided.

Among the unique features of the instrument are the use of linear mixers; a tuned amplifier, which can be used or not, as needed; signal lights to indicate proper control settings; and a novel dial readout to reduce reading errors.

Principles of Operation

Figure 2 is an over-all block diagram. For input frequencies above 200 Mc a dual-conversion system is used; single conversion is used below 200 Mc. Input signals are first passed through an attenuator (controlled by the SENSITIVITY control) and through a 100-Mc-wide band-pass filter selected by the hundreds-reference-FREQUENCY control.

Input signals below 200 Mc are routed to the 2nd mixer, either directly or through the tuned amplifier. In the 2nd mixer, the input signal is heterodyned against the tens-reference frequency (from 10 to 190 Mc depending on the setting of the FREQUENCY controls), and the beat-frequency output of the mixer is passed to the video amplifier where it

[1] R. W. Frank, H. T. McAleer, "A Frequency Counter With a Memory and With Built-In Reliability," *General Radio Experimenter*, 35, 5, May 1961.

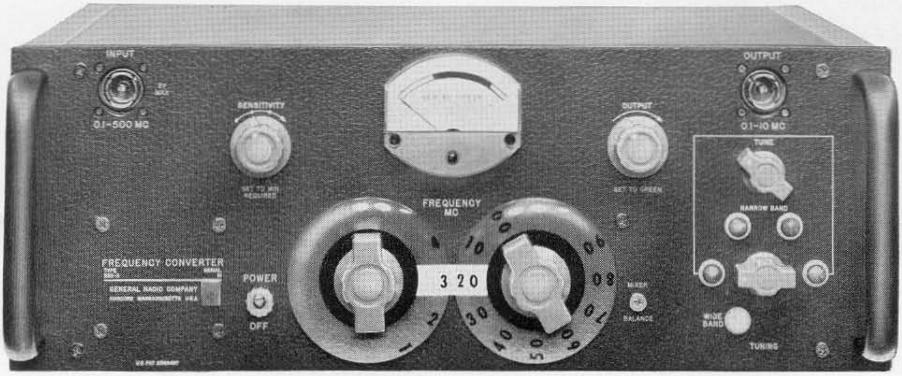


Figure 1. Panel view of the Type 1133-A Frequency Converter.

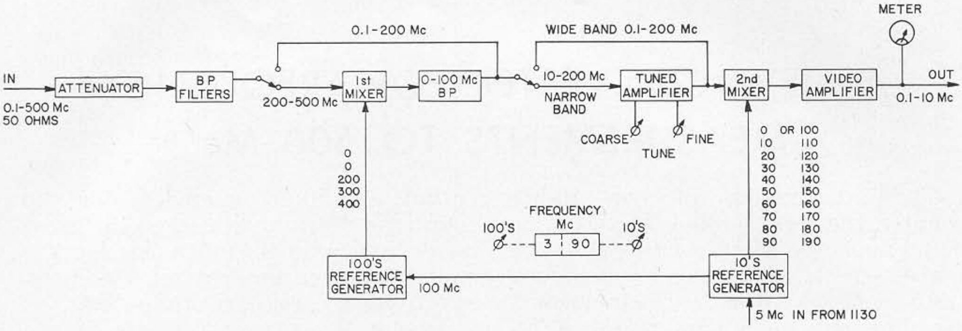


Figure 2. Over-all block diagram. A single conversion system is used below 200 Mc, a double conversion system from 200 to 500 Mc.

is amplified, metered, and applied to the output. The OUTPUT control varies the gain of the video amplifier to set the output level.

Input signals above 200 Mc are heterodyned against the hundreds-reference frequency (200, 300 or 400 Mc) in the 1st mixer. The 0- to 100-Mc beat-frequency output of the 1st mixer is filtered and then processed exactly like an input signal below 200 Mc. If the signal applied to the 2nd mixer is below 10 Mc (input frequencies of 0–10, 200–210, 300–310, 400–410 Mc), the 2nd mixer is converted into an amplifier, and the gain of the video amplifier is increased. In such cases the tuned amplifier is not used (since it doesn't operate below 10 Mc).

Linear Mixing

A mixer circuit usually operates with two signals applied, a reference (local-oscillator) signal and an input signal. The amplitude of the beat-frequency output depends more on the amplitude of the lower-level signal than on that of the higher-level signal. If the reference signal has the higher amplitude, the circuit will function as a linear mixer for the input signal. That is, the amplitude of the beat-frequency output will be proportional to the amplitude of the input signal, and, in particular, the signal-to-noise ratio of the input signal will be preserved in the output. If the input signal has the higher amplitude, however, the circuit will not operate as a linear mixer for that signal. The signal-to-noise ratio of the reference will be preserved, but the signal-to-noise ratio of the input signal will be degraded.

In a heterodyne frequency converter either mixing method can be used. Several existing converters use the nonlinear method, since this eases greatly the requirements on the purity or "cleanliness" of the heterodyne-reference-frequency signals. For successful measurements, therefore, the input signal must be relatively clean; and, consequently, the measurement of low-level signals in the presence of noise may be impossible.

Since we feel that the burden of signal purity should be on the measuring instrument rather than on the unknown signal, clean, high-level reference signals and linear mixing are used throughout the TYPE 1133-A Frequency Converter. As a result, the instrument can be used under an unusually wide variety of measurement conditions.

Tuned Amplifier

To increase the sensitivity of the instrument and to reduce further the effects of noise and extraneous signals, a tuned amplifier can be switched into the measuring system. With input frequencies from 10 to 500 Mc, the amplifier covers the range from 10 to 200 Mc. The amplifier is operated by two controls (see Figure 3). The TUNING control switches the amplifier (1) out of the system (WIDE BAND) to simplify operation when measuring pure, high-level signals, or (2) into the system

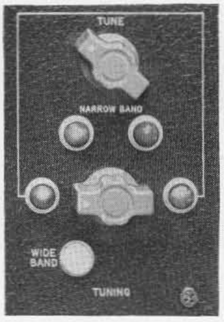


Figure 3. Tuning controls. Signal lights indicate proper setting.

(NARROW BAND) to enable measurement of noisy, low-level signals. It also selects one of the 4 coils necessary to cover the 20-to-1 tuning range. The TUNE control adjusts the tuning capacitor of the amplifier. The positions of the TUNING control corresponding to the different amplifier tuning coils are indicated by lights. With the TUNING control in the WIDE BAND position, the WIDE BAND indicator light will glow. With the TUNING control in any NARROW BAND position, one indicator light will glow, showing the position to which the control should be set. The indicator lights are controlled by the settings of the FREQUENCY controls. For example, with the FREQUENCY controls set to 330 Mc, the second NARROW BAND indicator light will glow, showing that the 20- to 40-Mc range of the tuned amplifier should be used. For certain positions of the FREQUENCY controls (00, 200, 300, 400) the WIDE BAND indicator light will remain on when the TUNING control is moved, indicating that the tuned amplifier cannot be used. For these positions, as mentioned above, the gain of the output amplifier is increased so that no loss in sensitivity occurs.

In addition to increasing the sensitivity of the instrument because of its gain, the tuned amplifier also provides selectivity to guard against noise and extraneous signals which may exist within the ±10-Mc conversion band. Figure 4 shows a typical plot of the bandwidth versus frequency of the tuned amplifier.

The sensitivity figures listed in the instrument specifications are broad, all-inclusive figures. Figure 5 shows a plot of the sensitivity versus frequency of a typical instrument for both WIDE BAND and NARROW BAND operation. This plot was taken for a beat-frequency output of 10.1 Mc, where the sensitivity of the associated counter is poorest. Therefore, the plot is labeled "worst-case sensitivity." For operation at output beat frequencies lower than 10.1 Mc, the sensitivity is significantly better.

Readout

Figure 6 shows a close-up of the new type of dial readout used on the converter. The readout uses transparent plastic dials with characters silk-screened on the rear. A white area on the panel behind the dials effectively "illuminates" the desired characters, and a uniform in-line indication is presented.* The readout has two main advantages:

1) The desired information always

*This readout method was devised some time ago by General Radio engineer Warren R. Kundert and is currently being incorporated in several new instruments.

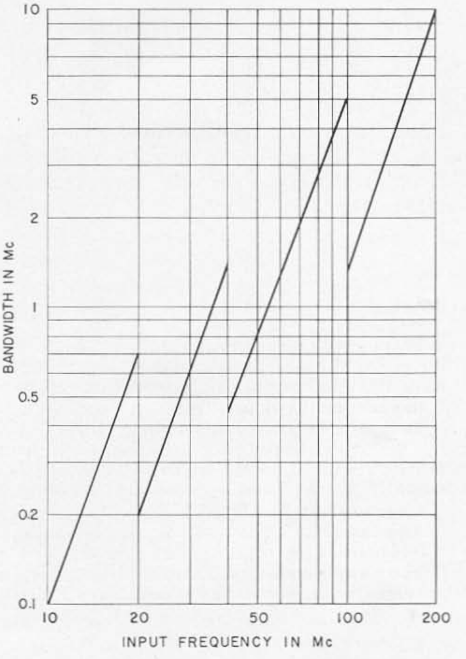


Figure 4. Tuned-amplifier bandwidth versus frequency. For input frequencies above 200 Mc, amplifier operates from 10 to 100 Mc.

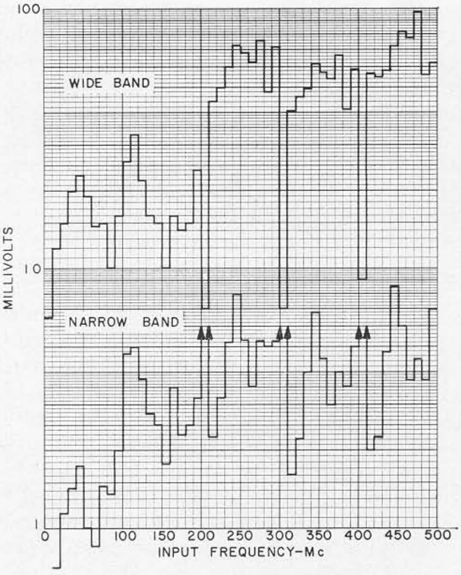


Figure 5. Typical over-all sensitivity of converter and counter for 10.1-Mc counter indication. This is worst case. Sensitivity is better for lower converter output frequencies.

occurs in the same region on the panel, lessening the chance for operator error, and

2) Since the dials are not hidden behind the panel, the operator can easily determine which way to rotate the controls to increase or decrease the frequency setting.

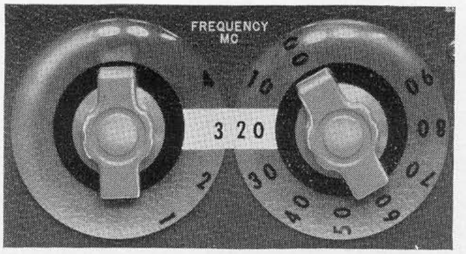


Figure 6. New in-line digital readout.

Over-all Versatility

In this all-in-one-package, wide-range converter, rugged construction and long-term reliability are combined with a number of important operating features — high sensitivity, the ability to make measurements in the presence of noise, positive identification of frequency, and simple controls — to produce an instrument of maximum usefulness in precise frequency measurement.

— H. T. McAleer

CREDITS

The design and development of the Type 1133-A Frequency Converter was carried out by W. F. Byers and H. T. McAleer, assisted by S. Samour. Progress from conception to completion was supported by G. Neagle, mechanical design; W. Montague, design drafting; R. H. Chipman, production engineering; and W. P. Buuck, test engineering.

— Editor

SPECIFICATIONS

INPUT

Frequency Range: 100 kc to 500 Mc.

Sensitivity (with the Type 1130-A counter): Better than 10 millivolts on narrow band; better than 100 millivolts on wide band. See Figure 5.

Impedance: 50 ohms.

Reference Frequency Required: 5 Mc, 0.1 volt, rms, into 50 ohms (normally supplied from 5-Mc output connector on Type 1130-A).

OUTPUT

Frequency: 100 kc to 10 Mc.

Amplitude: 0.25 volt to 1 volt, approximately.

Impedance: 100 ohms, approximately.

Noise and Harmonics: Narrow-band operation provides filtering to reduce noise and extraneous signals. Linear mixer preserves signal-to-noise ratio during conversion process.

GENERAL

Power Input: 105 to 125 (or 210 to 250) volts, 50 to 60 cps, 60 watts.

Accessories Supplied: Two coaxial patch cords for connection to counter, one coaxial cable connector, one 3-wire power cord, and spare fuses.

Dimensions: Bench model, width 19, height 7½, depth 17¾ inches (485 by 190 by 450 mm), over-all; rack model, panel, 19 by 7 inches (485 by 180 mm); depth behind panel, 14 inches (355 mm).

Net Weight: 34 pounds (15.5 kg).

Type		*Code Word*	*Price*
1133-AM	**Frequency Converter, Bench Model**	NOVEL	**$1250.00**
1133-AR	**Frequency Converter, Rack Model**	NEWEL	**1250.00**

U.S. Patent No. 2,548,457.

DIGITS CAN LIE

A Discussion of Error Sources in Counter Measurements

Since its introduction some 15 or so years ago, the counter, or digital time and frequency meter, has become a widely used electronic instrument, almost as ubiquitous as the vacuum-tube voltmeter or the oscilloscope. Wonderful as this instrument seems, however, it is still capable of producing wrong answers, even though its circuits may be functioning perfectly. Recognition and understanding of the sources of error will enable one to minimize their effects and to improve the usefulness and reliability of counter measurements.

When using a counter, one should keep in mind the distinction between precision and accuracy. Precision describes the degree of fineness, the least significant figures, of a measurement; accuracy indicates the possible extent of error. For example, it is quite possible (and fairly common) to measure a time interval with a counter to a precision of 0.1 μsec, but with an accuracy of only ±1 msec.

The simplified block diagram of Figure 1 shows the five basic circuit blocks of a digital time and frequency meter: input circuits, time base, main gate, program control, and decimal counting units. The input circuits generate trigger pulses from the input signal. For frequency measurement these trigger pulses are counted by the decimal counting units during a time interval derived from the time base; for time measurement the trigger pulses cause the main gate to start and stop the flow of time-base clock pulses into the decimal counting units.

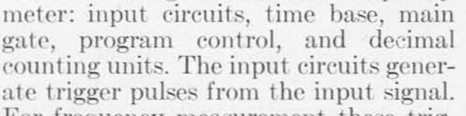


Figure 1. Simplified block diagram of a digital time and frequency meter.

The program control routes pulses to open and close the main gate, selects the proper pulses to be counted, controls the display, and handles the resetting operations.

Errors can, of course, occur if any of these circuit blocks malfunction, but we are concerned here with errors due to other causes. Such errors occur mainly in the time-base, gate, and input circuits. Some are inherent in the counting system; others depend upon the nature of the signal to be measured.

INHERENT ERRORS

Time-Base Error

The time-base reference for most counters is a quartz-crystal oscillator. Such oscillators are exceptionally accurate and stable but will still drift in frequency with time and should be reset occasionally. *Satisfactory self-check operation of a counter does not indicate the accuracy of the time-base reference frequency.*

One-Count Gating Error

Because the rate of the trigger pulses that are counted is not usually synchronous with the rate of the pulses that are opening and closing the main gate, it is possible for a trigger pulse to occur simultaneously with a gating pulse and not to be counted (see Figure 2). This leads to the so-called one-count gating error — the possibility that any particular measurement may be in error by one count. The percentage error versus frequency, caused by the gating error, is plotted in Figure 3 for various methods of measurement.

This one-count gating error applies only to a single measurement. In a series of measurements of the same quantity (a typical use of a counter) *the answer is*

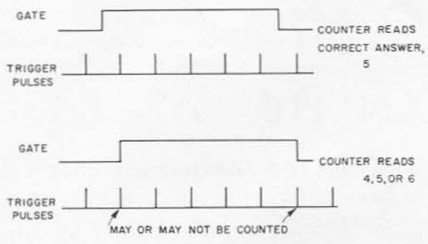


Figure 2. One-count gating error. Pulses occurring simultaneously with the gate may or may not be counted.

averaged, and the one-count error disappears. For example, if the true value of the last digit is between 4 and 5, say 4.5, the reading will jump back and forth between 4 and 5. By observing the relative rate of occurrence of the two digits, the operator can estimate a digit beyond the last one displayed. (Incidentally, it's easier to do this with the thermometer type of readout than with the more popular in-line readout.)

ERRORS CAUSED BY NOISE

There is an additional group of errors, which are caused by noise. Noise, in this instance, refers to anything that causes the input signal processed by the counter to be other than a perfect sine wave of infinite signal-to-noise ratio. Modulated signals, for example, may be considered to be noisy signals.

The effects of noise depend directly upon the operation of the input circuits, which function both as an amplitude

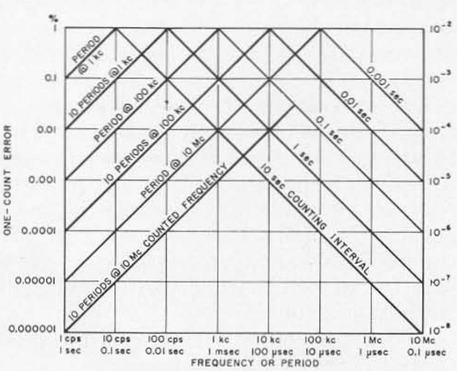


Figure 3. Percentage error versus frequency caused by one-count gating error.

limiter and a trigger pulse generator. In most counters, a Schmitt (or similar-type) circuit generates a trigger pulse when the input-signal voltage increases above a certain reference-voltage level and resets itself when the input voltage falls below another level, as illustrated in Figure 4. The voltage difference between the triggering and resetting levels is called the "hysteresis" voltage of the counter and determines the minimum input voltage necessary to operate the input circuits. A trigger-level control to adjust the absolute voltage of the trig-

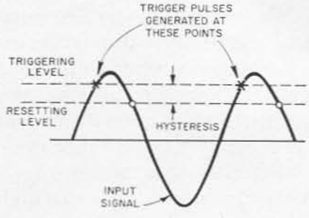


Figure 4. Operation of input circuits.

gering and resetting levels is very useful in combating the effects of noise and obtaining maximum input sensitivity.

Frequency-Measurement Errors

Errors in frequency measurement can be caused by modulation, either amplitude or frequency, or by the existence of spurious signals along with the desired signal.

Amplitude Modulation

If the input signal is amplitude-modulated, an error may occur if the triggering level is offset as shown in Figure 5. With the triggering and resetting levels adjusted symmetrically about 0 volts, however, the correct frequency will be measured as long as the minimum peak-to-peak excursion of the input signal is greater than the hysteresis voltage. Even with optimum adjustment of the triggering level, however, pulses will be lost if the degree of modulation reduces the voltage excursions to less than the hysteresis voltage.

Frequency Modulation

The counter is often used to measure various properties of a frequency-modu-

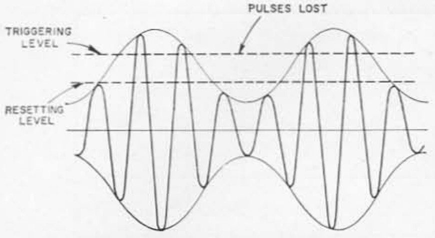


Figure 5. Error caused by amplitude modulation.

lated signal.[1] It can measure the average value of an input frequency during the chosen counting interval, provided that the frequency remains within the resolution capability of the instrument. If the frequency rises above the maximum counting rate of the instrument or falls below the minimum rate, pulses will be lost and the measurement will be in error.

If a heterodyne frequency converter for high-frequency measurements is used, care must be taken that frequency modulation does not drive the input frequency down through zero beat with the heterodyne reference frequency and "out the other side," or the counter reading will be incorrect.

Noise

If the desired signal is accompanied by noise sufficient to cause extra transitions of the hysteresis region, as shown in Figure 6, extra counts will be registered. This error can often be combated by adjustment of the triggering level to the region of steepest signal slope or by attenuation of both signal and noise. Note that it is the absolute value of the noise voltage that is important, rather than the signal-to-noise ratio. In a counter with a 0.2-volt hysteresis region, for example, a 10-volt signal accompanied by 1 volt of noise will not be measured correctly. Attenuating the input 10:1, however, leaves a signal of 1 volt with 0.1-volt noise, which will be measured with no difficulty.

The situation described above is slightly oversimplified; it is correct for signal and noise frequencies well below the maximum triggering rate. The hysteresis voltage of most counter input circuits is not constant for all input frequencies; it decreases at frequencies approaching the maximum triggering rate, even though the over-all sensitivity of the counter may decrease. A 10-Mc counter, for example, will be more sensitive to 10-Mc noise than to 1-Mc noise.

Interfering Signals

Occasionally it is necessary to measure the frequency of a signal in the presence of another signal of nearly equal amplitude. Counters have been severely, and somewhat unjustly, criticized for being unable to make this type of measurement. The actual situation, however, is not as bad as may be believed.

A capture effect occurs in the input circuits of a counter, similar to that encountered in FM receivers. Figure 7 shows an experimental curve describing the effect on a counter measurement of two signals together. If the frequency of the interfering signal is much higher than that of the desired signal, the interference behaves like the type of noise described above. That is, if the peak-to-peak interference voltage is less than the hysteresis voltage of the counter, it will cause no difficulty. If the interference voltage is greater than the hysteresis voltage, however, it may cause errors — depending on the ratio of the frequencies of the two signals and their amplitudes. As the interfering frequency approaches the desired frequency, the counter can tolerate more interference amplitude before making an error.

Let us consider the case of an interfering frequency close to the desired frequency as the interference amplitude

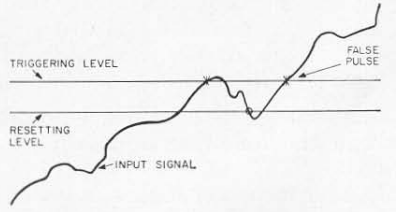


Figure 6. Effects of additive noise.

[1] I. Godier and P. S. Christensen, "New Method of Measuring FM Deviation Uses Electronic Counter," *Canadian Electronics Engineering*, July 1962, p. 38.

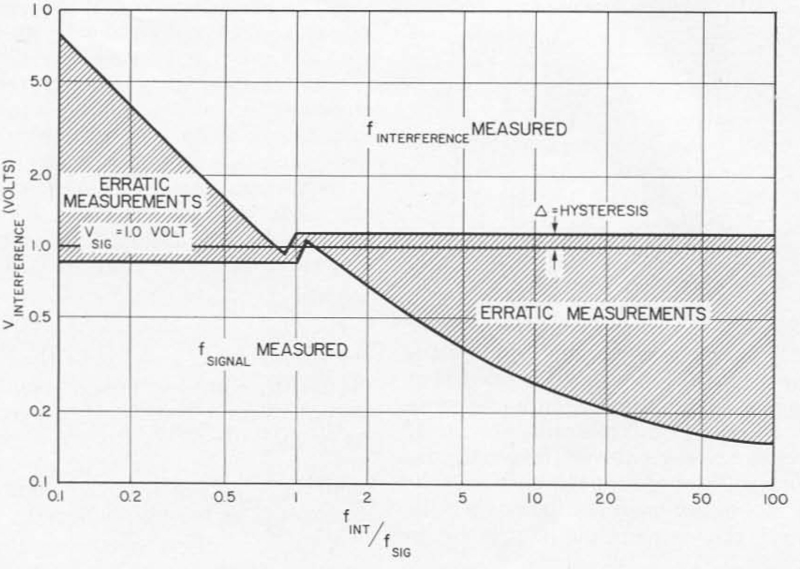


Figure 7. Capture behavior of a counter. Counter ignores an interfering signal with amplitude below the shaded area. If interference amplitude exceeds shaded area, counter will measure interfering signal only.

increases from zero. The composite voltage begins to wax and wane at a rate equal to the difference frequency and with an envelope amplitude equal to that of the interference. The voltage waveform looks like an amplitude-modulated wave, except that the phase of the "carrier" is also varying. The average frequency within the modulation envelope, *i.e.*, the average rate of positive zero crossings, is equal to that of the larger-amplitude, lower-frequency signal. The counter will continue to measure the lower frequency until the interference becomes great enough to compress the composite voltage amplitude within the hysteresis voltage limits. As the interference voltage increases still further, the counter will give erratic readings until the amplitude of the interfering signal exceeds that of the desired signal by an amount equal to the hysteresis voltage. *At that point and beyond, the counter will read the higher frequency.*

This two-signal behavior is summarized in the following approximate relations:

Counter measures higher frequency if

$$V_h > V_l + \Delta;$$

counter measures lower frequency if

$$V_h < \frac{f_l}{f_h} V_l + \mathrm{K}\Delta,$$

where:

V_h = peak-to-peak amplitude of higher frequency signal,
V_l = peak-to-peak amplitude of lower frequency signal,
f_h = higher frequency,
f_l = lower frequency,
Δ = hysteresis voltage of counter,
K = a factor varying between 1 and 2.

If neither condition is satisfied, the counter will give erroneous readings.

A counter, therefore, has a degree of inherent immunity to interference. If the interfering frequency is lower than the desired signal frequency, it will be completely ignored by the counter if its amplitude is less than the signal amplitude by at least the hysteresis voltage. As the interfering frequency exceeds the signal frequency, the counter can tolerate less and less interference amplitude but, in the limit, will still ignore interference amplitudes less than the hysteresis voltage.

Errors in Period Measurement

For period measurements, trigger pulses generated from the input signal open and close the main gate and control the flow of time-base pulses into the decade counters. The accuracy of such a measurement depends on the time accuracy with which the triggering-level crossing of the signal can be determined. Errors in this determination may occur for several reasons: the triggering-level crossing of the signal may vary because of drift, hum pickup, noise, etc., or the triggering level itself may vary for similar reasons. Figure 8 illustrates the effect of uncertainty in either signal or triggering level. These uncertainties are additive and can be combined into a single "noise" voltage. For sine waves triggering at zero crossings, the following relationship applies:

$$\text{Max error in } \% = \pm \frac{1}{\pi} \frac{V_n}{V_s} \times 100$$

where V_n = peak noise voltage,
V_s = peak signal voltage.

Stated in other terms, the fractional error caused by noise is about one-third the noise-to-signal ratio. For example, a noise voltage of 3% can produce an error of about 1%.

The effective noise includes both noise present in the signal and internal noise generated by the counter. The internally generated noise depends on the impedance of the signal source and the positions of the controls. For example, measuring the period of a clean signal of 1-volt rms amplitude from a 600-ohm source will yield an accuracy of about 0.05% on a General Radio Type 1130-A counter, indicating an effective internal noise of about 2 mv. A 10-periods measurement is at least 10 times as accurate since the time error is compared with a time interval 10 times as long.

The above remarks apply to noise of a random nature. If the noise is periodic, the measurement will display a cyclic variation, and it is possible to estimate the mean value of the measurement with greater accuracy than that indicated by the signal-to-noise ratio.

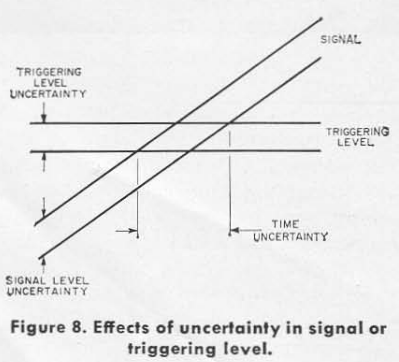


Figure 8. Effects of uncertainty in signal or triggering level.

Errors in Time-Interval Measurements

The sources of error described above apply also to time-interval measurements. Time errors caused by noise on the start and stop signals or triggering levels can be expressed as follows:

$$T = \pm \frac{V_n}{s}$$

where: T = error in seconds,
V_n = peak noise voltage,
s = slope of signal in volts/second.

As the slope of the signal increases, the time error caused by noise decreases, so that, for brief pulses or voltage steps with rise time comparable to one period of the counted frequency, the measurement error is reduced to ±1 period of the counted frequency ± the error of the time-base reference.

Two Rules-of-Thumb for Counter Measurements

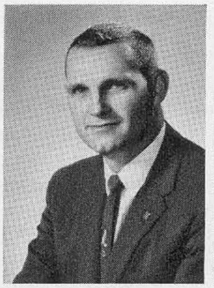
Two good rules-of-thumb to follow are:

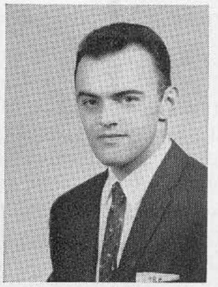
1. A steady reading is usually correct; an erratic reading incorrect.
2. When in doubt, look at the input signal with an oscilloscope.

The purpose of this discussion has been to acquaint the reader with some of the causes of error in counter measurements, not to weaken his confidence in such measurements but, rather, to bolster that confidence through better understanding of the principles involved.

— H. T. McAleer

CAPITOL SERVICE





Donald Brown — Raymond Jones

The establishment of a service laboratory at our Washington, D. C., office brings to six the number of such operations in the U. S. and Canada. Donald W. Brown, formerly Service Supervisor at our New York Office, heads the new facility, located at Rockville Pike at Wall Lane, Rockville, Maryland (Telephone 946-1600). At New York, Raymond J. Jones becomes Service Supervisor.

All six service offices are fully staffed and equipped for the repair, reconditioning, and recalibration of General Radio instruments and for the certification of General Radio standards. All work performed by our service department is guaranteed for one year. Customers preferring to make their own repairs will find most replacement parts stocked at all six service centers.

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"Bridges and Techniques for Impedance Measurement" was the title of a seminar recently conducted by General Radio engineers at the plant of RCA Victor Company, Ltd., in Montreal. Morning lectures by Dr. John F. Hersh were followed by afternoon workshop sessions. Participating in the seminar were engineers from RCA Victor, Canadian Aviation Electronics, Northern Electric, and Canadian Marconi. The photo shows a workshop on coaxial-line measurements.

General Radio Company

extends to all *Experimenter* readers its best wishes

for a Happy and Prosperous 1963.

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