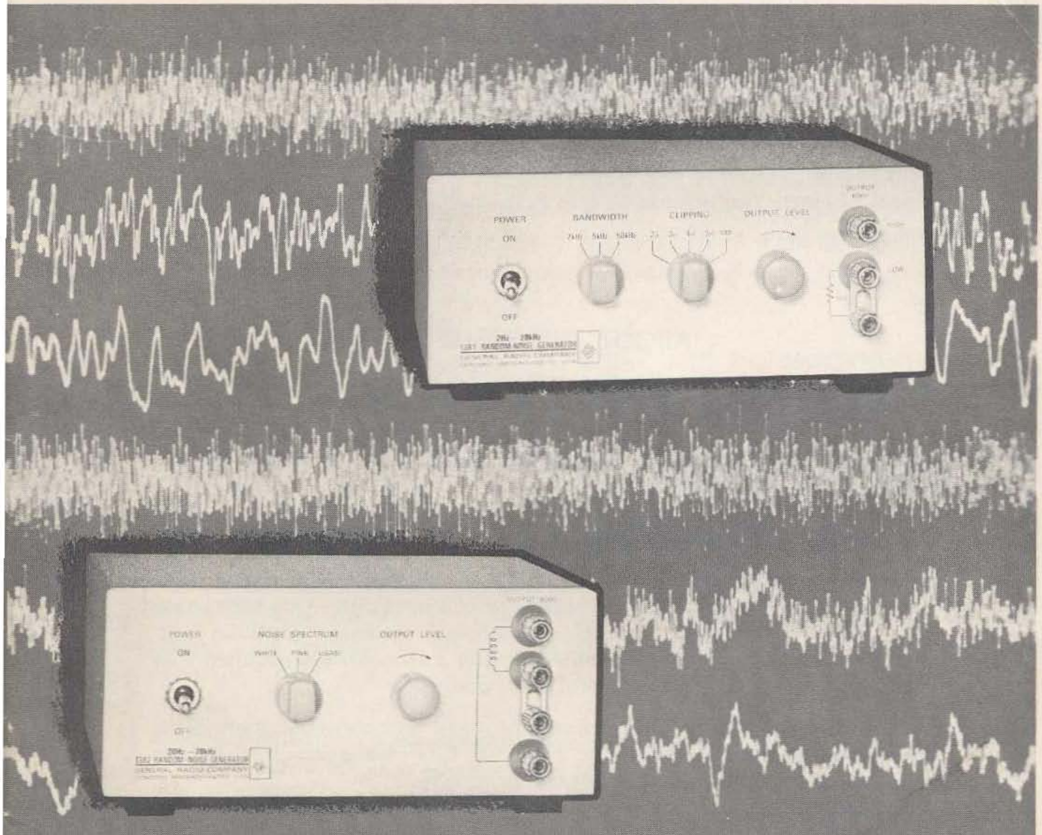
the GENERAL RADIO Experimenter

GR Experimenter Index 1968

Index by Type Number

THE GENERAL RADIO

Experimenter



TWO NEW RANDOM-NOISE GENERATORS

VOLUME 42 · NUMBER 1 / JANUARY 1968

the Experimenter

Volume 42 • No. 1 January 1968

Published monthly by the General Radio Company

CONTENTS

ABOUT THIS ISSUE

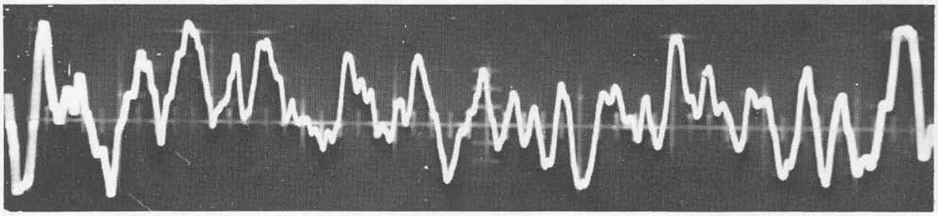
Noise, in the electrical-engineering context, has come a long way toward respectability since it was first defined as an "unwanted signal in an electronic communication system." To a fast-growing body of engineers, noise is regularly wanted and used as a test signal, and the random-noise generator has an established role in many test programs. This month's feature article introduces two new, solid-state generators that deliver random noise in a variety of forms used in audio and subaudio testing.

How does your RC oscillator indicate 20.0 kHz? As 200 $\times$ 100 Hz? As 2.00 $\times$ 10 kHz? As 20000 Hz? Even the best of us can be forgiven an occasional slipped decimal point in converting conventional oscillator readouts into practical terms. GR's new digital oscillator (page 14) solves the problem with an in-line readout that includes decimal point and units.

A one-percent, one-farad capacitor? The mind boggles, but there it is, on page 20. It seems anticlimactic to add that it is also a standard for six other values down to 1 μF, with accuracies to $\pm 1/4\%$.

On the Cover: Oscillograms show various types of Gaussian noise produced by GR's new random-noise generators. From top to bottom: noise in the three bandwidths (50, 5, and 2 kHz) of the 1381 and the white, pink, and USASI noise outputs of the 1382.

The *General Radio Experimenter* is mailed each month without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.



RANDOM-NOISE GENERATORS

The uses of laboratory random-noise generators have multiplied rapidly since the introduction of the first commercially available model some years ago. At first they were principally used in laboratory studies of methods of overcoming the effects of noise in communications systems; the random-noise generator was a controlled source of noise disturbance for such tests. Now, however, random noise has come into its own, and in many applications it has become the test signal. It is useful in electrical measurements because its wide, continuous frequency spectrum and its amplitude distribution simulate the characteristics of many natural phenomena and because it offers the possibility of using a single measurement as an indicator of performance over a wide frequency band. In acoustical measurements, bands of random noise are used to smooth response curves that might otherwise be difficult to interpret. The use of random noise in psychoacoustical experiments has greatly increased our understanding of the process of hearing. Also, because random noise best simulates the vibrations that aircraft and rockets are subjected to in flight, it is commonly used in vibration and fatigue testing of aerospace components, assemblies, and structures.

Random-noise generators can be classified according to the spectrum of their output, and instruments are available to cover frequencies from near dc to microwave. Our consideration here will be limited to audio-frequency noise generators.

WHAT ARE THE IMPORTANT CHARACTERISTICS OF AUDIO-FREQUENCY RANDOM-NOISE GENERATORS?

Just as it is desirable that sine waves have low harmonic distortion and that rectangular pulses have short rise and fall times, there very definitely are desirable characteristics for random noise. However, measuring and specifying them requires the use of different concepts than those needed for sine-wave, pulse, and other periodic waveforms.

Random noise is a signal whose instantaneous amplitude cannot be precisely predicted. It can be described only in terms of certain average properties, of which the most important are the spectrum and the amplitude distribution. For accuracy, measurements of these properties must be averaged over a long period of time, as there is always some fluctuation present in measurements of a random function. For instance, when the spectrum of random noise is measured with an analyzer, there is always some fluttering of the meter pointer. The user expects such fluctuation, as it assures him that he is measuring random noise, and he chooses the averaging time to reduce the fluctuation to an adequately low level.

The Spectrum

The spectrum of a periodic signal is composed of discrete lines, each corresponding to a component in the frequency spectrum. Such a signal is predictable, not random, because of the periodicity of each of its components. The spectrum of random noise, on the other hand, is a continuous function of frequency, containing no line components.

The function used to describe the spectrum of random noise is *spectral intensity*, expressed in units of volts squared per unit frequency. (When divided by the resistance across which that voltage appears, it becomes the power spectrum, expressed in watts per unit frequency.) The spectral intensity is the cosine Fourier transform of the autocorrelation function, and it is a convenient function in theoretical considerations of noise. It is not, however, the most convenient function for practical measurements; spectra are usually measured (as with an analyzer) as voltage in a given bandwidth, and filter responses, used in shaping noise spectra, are usually measured in terms of voltage (not voltage squared or power) as a function of frequency. We therefore

A WORKING GLOSSARY OF NOISE TERMINOLOGY

Amplitude Density Distribution: a function giving the fraction of time that the voltage dwells in a narrow range.

Amplitude Distribution Function: a function giving the fraction of time that the instantaneous voltage lies below a given level.

Gaussian or Normal Distribution: a particular amplitude distribution of great fundamental importance in the theory of probability—the Gaussian Probability Density Distribution is the "bell-shaped curve."

Noise: any unwanted signal, including noise, hum, crosstalk, etc.

Pink Noise: noise whose spectral intensity is inversely proportional to frequency over a specified range, therefore dissipating in a constant resistance equal power in any octave bandwidth in that range.

Random Noise: a signal whose instantaneous amplitude is determined at random and is therefore unpredictable. Truly random noise contains no periodic frequency components and has a continuous spectrum.

Spectral Intensity: a function precisely defining the spectrum and having the units of voltage squared per unit frequency.

Spectrum: the distribution of the components of a signal across the frequency range.

Stationarity: a property that random noise is said to have if its spectral intensity and amplitude distribution do not change with time.

Voltage Spectrum: a function which is the square root of the Spectral Intensity, having the units of voltage in a unit frequency band.

White Noise: noise whose spectral intensity is constant over a specified range, therefore dissipating in a constant resistance equal power in equal bandwidths anywhere in that range.

use, practically, the *voltage spectrum*, whose units are sometimes inelegantly called "volts per root hertz," but which we prefer to express in terms of "voltage in a one-hertz bandwidth." Numerically, it is the square root of the spectrum level, as defined above.

The most generally desirable spectrum for random noise is one that is constant over a wide range of frequencies. Such noise is called "white noise" by analogy with white light, which contains more or less equal intensities of all visible colors.* There are, of course, other spectra more convenient for certain uses; some of these will be discussed later.

In measuring the spectra of modern wide-range noise generators, it is necessary to use analyzers that cover a wide range of frequencies. This generally means using several analyzers covering different ranges, and it also means knowing the effective bandwidth of each so that data can be accurately reduced to the same equivalent bandwidth. The characteristics of the detector in each analyzer must be known, too; peak, average, and true rms detectors respond differently to random noise, and appropriate corrections must be applied.

The Amplitude Distribution

The relationships between the root-mean-square, the rectified average, and the peak value of sine waves, pulses, and other well-defined periodic waveforms are generally easy to determine by mathematical calculation. Knowledge of these relationships is necessary in the use of voltmeters of different types, which may measure the rectified average, the peak, or the rms value of a voltage, but which are generally calibrated to indicate the rms value of a sine wave. To determine the response of a voltmeter to random noise, it is necessary to know the relative occurrence of various amplitudes in the noise voltage, given by the *amplitude density distribution*, $p(v)$ or by the *amplitude distribution function*, $P(v)$. These functions describe the amplitude distribution in terms of probability. The value of $p(v_o)dv$ is the probability, on a scale from 0 to 1, that at any instant in time the amplitude of the noise will lie between v_o and $(v_o + dv)$. $P(v)$ is the integral of $p(v)$; $P(v_o)$ gives the probability that at any instant in time the amplitude of the noise will lie below v_o.

The normal, or Gaussian, amplitude distribution is of fundamental importance in statistical theory and describes many natural phenomena. The central-limit theorem of statistics states, in essence, that the distribution of the sum of a number of independent random variables approaches the Gaussian distribution as the number of such variables is increased, regardless of the distributions of the individual variables. By extension of this reasoning, reducing the bandwidth of a non-Gaussian random noise will generally make it more Gaussian. In that sense, the Gaussian distribution is a stable distribution. It is the distribution of normally occurring random errors in experimental measurements. It is also the amplitude distribution of natural electrical noise (e.g., shot noise in an electron stream and thermal noise in a resistance). The properties of the Gaussian distribution have been studied intensively and are well known. For all these reasons, the

* Although, as Bennett (Ref. 1, p. 14) points out, the analogy has been drawn incorrectly, since spectroscopists were measuring intensity as a function of wavelength, and they found it to be substantially constant per unit *wavelength*, *not* per unit *frequency*.

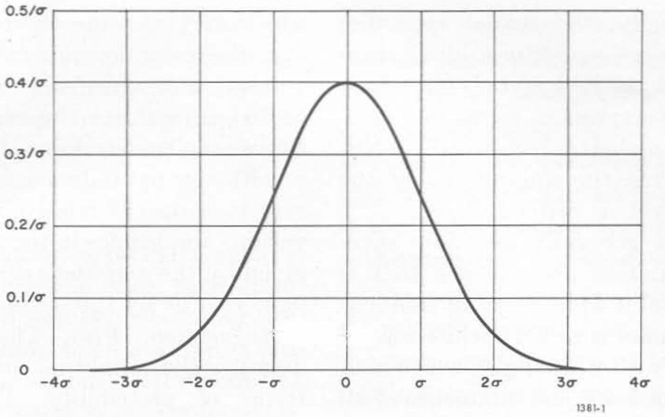


Figure 1. The amplitude density distribution $p(v)$ of Gaussian random noise.

Gaussian distribution is most desirable for a general-purpose random-noise generator.

The Gaussian probability density distribution, $p(v)$, is characterized by the well known bell-shaped curve shown in Figure 1. It is plotted in terms of σ, the rms amplitude of the noise (the standard deviation in statistical theory). The Gaussian probability distribution function, $P(v)$, is plotted in Figure 2, again in terms of σ. It can be seen from Figure 2, for example, that a Gaussian random noise exceeds its positive root-mean-square amplitude only about 16% of the time and twice that value only about 2% of the time.

The rms amplitude of a noise with Gaussian amplitude distribution is σ; the rectified average value is $\sigma\sqrt{(2/\pi)}$. In a measurement of Gaussian noise, a voltmeter that responds to the rectified average and that is calibrated to indicate the rms of a sine wave will indicate a voltage that is low by the factor 0.891 (−1.05 dB).* In general, rectified-average-responding voltmeters are recommended (with the above correction being used) over true-rms-responding meters for measuring noise voltages; the response time is usually faster, and the time constant required for a particular degree of smoothing is slightly less. Although the response of peak-responding voltmeters to random noise is known[2], calibration depends upon accurate knowledge of the constants of the voltmeter circuit; because of this, peak-responding voltmeters are not recommended for noise measurements, except where the application requires their wide frequency range.

The *amplitude distribution function* can be measured with commercially available amplitude-distribution analyzers. These generally comprise level-crossing detector circuits, together with some system for measuring the average fraction of time that the level is exceeded. Direct measurement of the *amplitude density distribution* requires two level-crossing detectors arranged to determine the fraction of time that the voltage lies within a very small range (dv). Other, less accurate, methods are also useful in assessing the amplitude distribution of random noise.

* Much practical information concerning random noise is contained in Reference 13.

For instance, an oscillographic display of the noise can be photographed through an optical wedge,[3] with a long time exposure to give some idea of the distribution of a noise voltage containing frequencies too high for level-crossing detector systems. Sampling techniques can also be useful in determining the amplitude distribution of high-frequency noise.

Other Characteristics

Other desirable attributes of random noise are

1. Stationarity;
2. Freedom from contaminating signals such as hum; and
3. True randomness.

The output of a random-noise generator is said to be stationary* if its average characteristics do not vary with time. The noise voltage is constantly changing, of course, and never repeats the same pattern with time, but, if the noise is stationary, its spectrum and its amplitude distribution remain exactly the same. The property of stationarity is important to the experimenter simply because he wishes to be assured that the characteristics of the noise do not change during the course of his experiment. Rigorous methods of checking for stationarity have been developed[4].

It is important that random noise not be contaminated by hum related to the power-line frequency, by $1/f$ noise, which might arise in amplifiers, or by other disturbances. Periodic signals could possibly be correlated with components of other signals being used, producing erroneous results in sensitive measurements. The presence of periodic signals or low-frequency semiconductor noise ($1/f$ noise) could alter the spectrum of the noise. In a random-noise generator, such contamination should be kept as low as possible.

An important property of random noise is true randomness. Pseudo-random noise, available as a substitute, is periodic, repeating one pattern over and over. Among the limitations of pseudo-random-noise generators are a relatively low crest factor (ratio of peak to rms amplitude) and a significant departure from Gaussian amplitude distribution. For many of the reasons that lead one to select random noise as a test signal in the first place, he is generally better off with noise that is truly random.

*Bennett, W. F., Ref. 1, p 52.

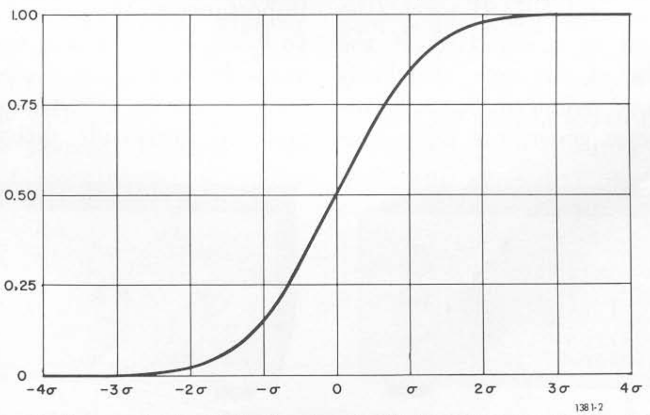


Figure 2. The amplitude distribution function *P(v)* of Gaussian random noise.

TWO NEW NOISE GENERATORS

The new TYPES 1381 and 1382 Random-Noise Generators (Figure 3) embody many improvements over previously available instruments:

1. The amplitude distribution is symmetrical and is accurately Gaussian to beyond 4σ.

2. These generators use a semiconductor noise source and transistor circuits so there is excellent stability and no "warmup" time delay.

3. The method of generating and processing the noise ensures no contamination with $1/f$ noise.

4. A variety of spectra is available, built into the instruments.

5. The output impedance is 600 ohms.

The different characteristics of the two units give the user a convenient choice of noise generator for his specific application. The TYPE 1381 has three different output spectra: white, with upper cutoff frequencies of 2, 5, and 50 kHz; each flat down to 2 Hz; and amplitude clipping, if desired, at 2, 3, 4, or 5 σ. The TYPE 1382 has three spectra, white, "pink," and USASI noise (the latter specified in a standard of the United States of America Standards Institute) [5], and balanced, floating output. Both noise generators have continuous output-level controls that permit the output voltage to be reduced over a range of 60 dB from the maximum of 3 volts open-circuit.

CHARACTERISTICS OF THE NEW NOISE GENERATORS

The two instruments use the same method of generating random noise; they differ mainly in the spectral filters and output circuits. In each instrument the noise is generated by a semiconductor noise diode in a band extending roughly from 80 to 220 kHz. The amplitude in this bandwidth is kept constant by an automatic level circuit, which corrects for the temperature coefficient of the diode. The output from the noise diode is heterodyned down to the audio-frequency band in a balanced, symmetrical modulator, thus making the amplitude distribution symmetrical. It becomes even more accurately Gaussian because of the subsequent further bandwidth reduction by filtering.

It is becoming of increasingly greater importance to the user of random noise to know its amplitude distribution accurately, and an important feature of these new noise generators is a specification on amplitude distribution. This

Figure 3. Types 1381 (left) and 1382 Random-Noise Generators.

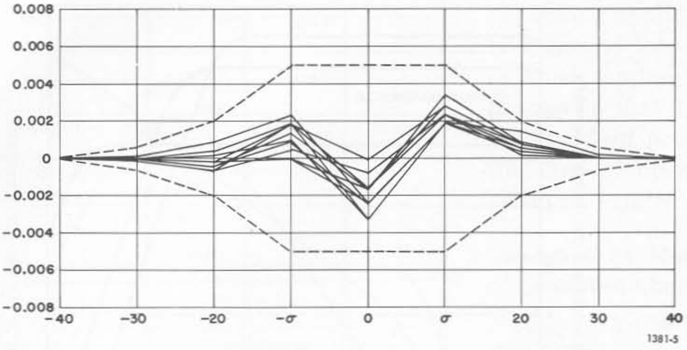


Figure 4. Departure from Gaussian amplitude-density distribution of output of 1381 and 1382 random-noise generators. Measurements were made on 10 different units with a "window" of 0.2σ centered at 0, ±1, ±2, ±3, and ±4 σ. Broken line shows limits of specifications on amplitude distribution of these noise generators.

specification is given in terms of the amplitude density distribution as measured with a window, dv, of finite width, in this case, 0.2σ. The specification is stated at 0, ±1, ±2, ±3, and ±4 σ. Peaks of even higher levels are present in the output noise, but these are usually of little importance. The energy contribution due to peaks above 4σ is entirely negligible. The results of measurements of the amplitude density distribution of a number of these instruments are shown in Figure 4. Measurements were made in a "window" of 0.2σ, and the adherence to the Gaussian distribution out to 4σ is seen to be very good. The maximum crest factor of the noise is limited by the voltage or current swing capability of the output amplifier stage. At full output, clipping in the output stage will not occur below 4σ; reducing the output level to half the maximum will ensure that peaks to 8σ can be present. However, the probability of occurrence of peaks above 5σ is so slight that waiting for one to occur lies somewhere between tedium and hopelessness.

1381 RANDOM-NOISE GENERATOR

The 1381 is intended as a noise source for driving vibration test systems and as a general-purpose noise source in the audio-frequency range. A block diagram of it is shown in Figure 5.

The output spectra of the 1381 are white (flat±1 dB) over three ranges, from 2 Hz to 1, 2.5, or 25 kHz. The upper cutoffs are determined by Butterworth filters having slopes of -12 dB per octave, and upper cutoff frequencies of 2, 5, and 50 kHz, respectively. As the bandwidth is reduced, the gain is increased so that the output power is the same for each range. The voltage spectra of these three outputs of the 1381 are plotted in Figure 6.

The output of the 1381 can be symmetrically clipped, if desired, at ±2,

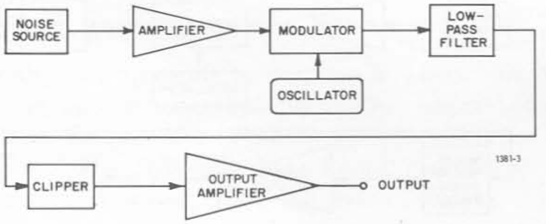


Figure 5. Block diagram of the 1381 Random-Noise Generator.

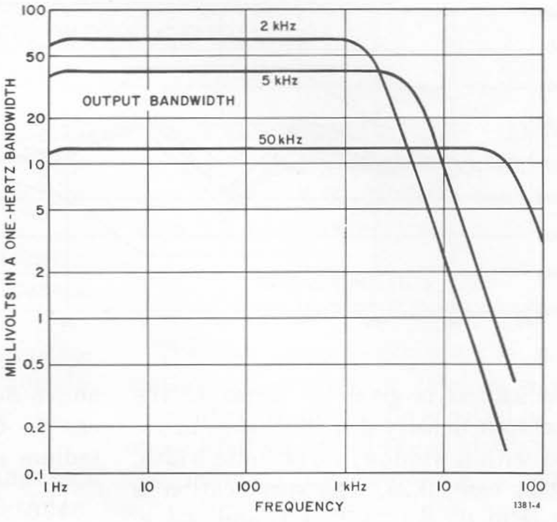


Figure 6. Voltage spectra of the 1381 for the three different output bandwidths at 3 volts rms output level.

± 3, ± 4, or $\pm 5\ \sigma$. Such clipping has almost negligible effect on the total power or on the shape of the spectrum. Clipping is useful in cases where high-power amplifiers are being driven at levels where occasional overloads from noise peaks could be harmful. Such clipping is also used for precautionary reasons in vibration testing.

1382 RANDOM-NOISE GENERATOR

The 1382 is intended specifically for use in the audio-frequency range, as a broadly useful generator of test signals. A block diagram of it is shown in Figure 7. The 1382 Random-Noise Generator also offers three choices of spectrum: white noise, "pink" noise, and "USASI" noise. The white noise is flat (± 1 dB) from 20 Hz to 25 kHz and has an upper cutoff frequency (-3 dB) at 50 kHz, with an upper cutoff slope of -12 dB per octave. Pink noise, by further (and a little more appropriate) analogy with optics, is so called because of its emphasis on lower frequencies, as in reddish light. Pink noise has a spectral intensity that is inversely proportional to frequency, that is, a voltage spectrum that is inversely proportional to the square root of frequency. It has equal energy in each octave band and is therefore useful in measurements made with constant-percentage-bandwidth analyzers.[14] The pink-noise output of the Type 1382 is pink over the range from 20 Hz to 20 kHz. The spectrum of

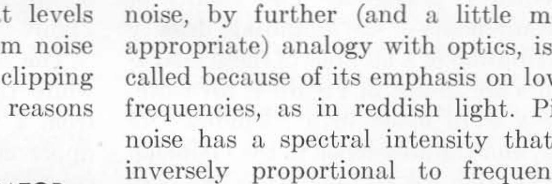


Figure 7. Block diagram of the 1382 Random-Noise Generator.

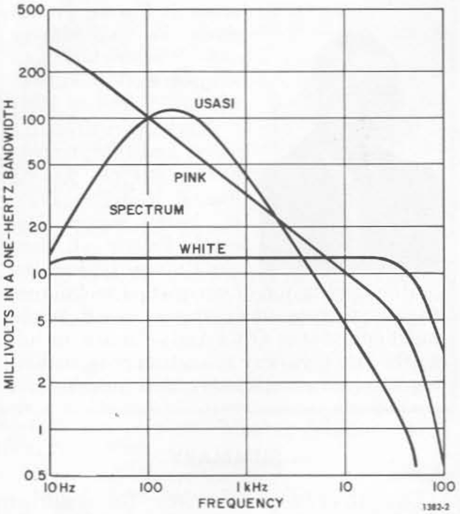


Figure 8. Voltage spectra of the 1382 for the three different output spectra at 3 volts rms output level.

USASI noise results from the passage of white noise through two simple RC filters: a high-pass unit with a cutoff frequency of 100 Hz and a low-pass unit with a cutoff frequency of 320 Hz. The spectrum accurately follows these characteristics from 20 Hz to 20 kHz. USASI noise roughly simulates the distribution of energy with frequency in speech and music, and it has been used for testing amplifiers and loudspeakers. These three outputs of the 1382 are plotted as voltage spectra in Figure 8. In this instrument, the output amplifier includes a transformer, so that the output can be taken floating, single-sided, or balanced.

APPLICATIONS

It is almost impossible to keep track of all the applications that ingenious users are finding for random-noise generators. The following list should demonstrate, however, that the ever-expanding applications of random noise now extend into many fields, for each of which random noise has some unique properties that have by now made its use standard procedure. We can divide the uses of noise generators into five broad categories:

Simulation of Naturally Occurring Noise of Controlled Characteristics

Noise of known amplitude and known spectral characteristics is the most effective for testing various methods of signal detection and recovery in the presence of noise, as in radio, telemetry, radar, and sonar systems. It also can be used for simulation of noise on telephone lines, for noise interference tests on multi-channel systems, and as background noise in comparisons of signal-processing systems.

As a Test Signal for Electrical Measurements

Noise has many uses as the test signal itself in electrical measurements. These include intermodulation-distortion and crosstalk measurements on multi-channel communications systems,[6] the simulation of randomly occurring traffic in communication systems, tests on servo amplifiers, and studies made with analog computers. Wideband noise can be used for determination of the impulse response of networks and systems by cross-correlation of the output with the input. It is commonly used for setting levels on carrier equipment. Its relatively high and symmetrical crest factor makes it suitable for measuring overload characteristics of amplifiers. Broadband noise can also be applied to two supposedly identical networks, their outputs then being compared by Lissajous pattern techniques.

As a Test Signal in Acoustical and Psychoacoustical Measurements

Bands of noise are often used in making frequency-response measurements on microphones, loudspeakers, and rooms, to smooth the resultant curves for easier interpretation.[7] Noise is an excellent test signal for measurement of reverberation time[8] and for reverberant testing of acoustical properties of materials. Similar applications include tests of the sound-transmission properties of walls, panels, and floors[8] and measurements of sound attenuation in ducts. Noise is also used in testing of silencers for air-conditioning systems and sound-proofing of aircraft. In psychoacoustics, noise is used in many hearing tests and masking experiments[9] and in tests of intelligibility of speech in the presence of noise. Random noise is also used in studies of the application of correlation techniques to acoustic receiving systems.[10]

As a Driving Signal for Vibration Testing

Random noise is used to drive shakers for vibration tests of components, assemblies, and structures[11] and is similarly applied to loudspeakers for subjecting the same test objects to high-intensity sound waves. It is also used for fatigue testing of structures subjected to sound or vibration.[15]

In Demonstrations of Statistical Theory and Information Theory

In the classroom, random-noise generators are naturally called upon to familiarize students with properties of random noise, including amplitude distribution.[12] In the laboratory, noise is useful in experiments on signal-detection and signal-processing systems, including correlation detection systems.

James J. Faran, Jr. received his AB degree from Washington and Jefferson College in 1943 and his MA and PhD at Harvard University in 1947 and 1951, respectively. Before joining General Radio in 1952, Dr. Faran was a Research Fellow at Harvard University, working on the application of correlation techniques to acoustic receiving systems. As a development engineer in GR's Audio Group, he has worked on a variety of instruments, including recorders, voltmeters, and analyzers.

SUMMARY

The diversity of uses for random noise led us to offer two compact, inexpensive generators, each of which includes those features most desirable in a certain family of applications. To summarize the characteristics of each instrument, from the applications viewpoint:

The 1381 generates noise that is flat down to 2 Hz and is especially well suited for random-vibration tests and for general-purpose use in the audio and subaudio range. The upper frequency limit can be switched to 2, 5, or 50 kHz, and the output signal can be clipped symmetrically at 2, 3, 4, or 5 times the rms amplitude. Amplitude distribution is Gaussian.

The 1382 generates noise in the 20-Hz to 20-kHz band and is ideal for electrical, acoustical, and psychoacoustical tests. Three spectra are offered: white (flat), pink (−3 dB/octave), and USASI. The output can be taken balanced or unbalanced, floating or grounded.

Both instruments are housed in cabinets only 3½ in. high and 8½ in. wide, with rack-mounting options.

J. J. Faran, Jr.

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SPECIFICATIONS

TYPE 1381

Spectrum: Flat ±1 dB from 2 Hz to 1, 2.5, or 25 kHz; upper cutoff frequency (3-dB point) can be switched to 2, 5, or 50 kHz. Spectral density at 3-V output and for 1-Hz bandwidth is approx 64, 40, and 13 mV, respectively, for 2-, 5-, and 50-kHz upper cutoff. Upper cutoff slope is 12 dB/octave. (See Figure 6.)

Waveform:

Voltage	*Gaussian Probability Density Function*	*Amplitude Density Distribution*
0	0.0796	0.0796 ± 0.005
$\pm\sigma$	0.0484	0.0484 ± 0.005
$\pm 2\sigma$	0.0108	0.0108 ± 0.003
$\pm 3\sigma$	0.000898	0.000898 ± 0.0002
$\pm 4\sigma$	0.0000274	0.0000274 ± 0.00002

These data measured in a window of 0.2 σ, centered on the indicated values; σ is the standard deviation or rms value of the noise voltage. Noise can be clipped at approx ±2, ±3, ±4, or ±5 σ to remove the extremes of amplitude. Such clipping has negligible effect on the spectrum or rms value of output.

Output Voltage: > 3 V rms max, open-circuit, for any bandwidth.

Output Impedance: 600 ohms, unbalanced. Can be shorted without causing distortion.

Amplitude Control: Continuous adjustment from full output to approx 60 dB below that level.

Terminals: Output at front-panel binding posts and rear-panel BNC connector.

Accessories Supplied: Power cord, spare fuses, rack-mounting hardware where appropriate.

Power Required: 100 to 125 or 200 to 250 V, 50 to 400 Hz, 6 W.

Mounting: Convertible-bench cabinet.

Dimensions (width x height x depth): Bench, 8½ x 3⅞ x 9⅞ in. (220 x 99 x 250 mm); rack, 19 x 3½ x 9 in. (485 x 89 x 230 mm).

Weight: Net, 5½ lb (2.5 kg); **shipping,** 10 lb (4.6 kg).

TYPE 1382

Spectrum: Choice of (a) white noise (constant energy per hertz bandwidth) ±1 dB, 20 Hz to 20 kHz, with 3-dB points at approx 10 Hz and 50 kHz; (b) pink noise (constant energy per octave bandwidth) ±1 dB, 20 Hz to 20 kHz; or (c) USASI noise, as specified in USA Standard S1.4-1961. (See Figure 8.)

SPECIFICATIONS (continued)

Waveform: Same as 1381, except clipping is not provided.

Output Voltage: Same as for 1381; see above.

Output Impedance: 600 ohms. Output is floating, can be connected balanced or unbalanced.

Amplitude Control: Same as for 1381; see above.

Terminals: Output at front-panel binding posts and rear-panel jacks for double plugs.

Accessories Supplied:
Power Required:
Mounting:
Dimensions:
Weight:
} Same as for 1381; see above.

Catalog Number	*Description*	*Price in USA*
	Random-Noise Generator	
1381-9700	2 Hz to 20 kHz, Bench Model	$375.00
1381-9701	2 Hz to 20 kHz, Rack Model	398.00
1382-9700	20 Hz to 20 kHz, Bench Model	375.00
1382-9701	20 Hz to 20 kHz, Rack Model	398.00

WIDE-RANGE RC OSCILLATOR WITH IN-LINE DIGITAL FREQUENCY READOUT

Central to the popularity of the general-purpose RC oscillator is its continuously adjustable frequency dial. In fact, the RC oscillator did not become popular until variable air capacitors were used to provide continuous adjustment over 10:1 frequency bands. This method of frequency selection, with dial and vernier, provides a convenient compromise between ease of reading, settability, resolution, and ability to sweep. It will no doubt continue to be most popular for general laboratory use.

As requirements for greater accuracy and resolution and for digital programmability have had their effect on signal sources, noncontinuous or discrete frequency controls have replaced the continuously variable air capacitor. This trend is most apparent in sources using synthesizer techniques, where seven or eight digits of resolution are common, and also in RC oscillators where push-button and rotary switches are used to control three or four digits. Since the use of discrete steps allows closer tracking of the tuning elements, this type of tuning most often appears on high-performance, high-cost instruments, where the objective is to improve performance characteristics of the source, such as stability of output, distortion and noise. In almost all cases the actual method of frequency selection was resorted to as an undesirable necessity,

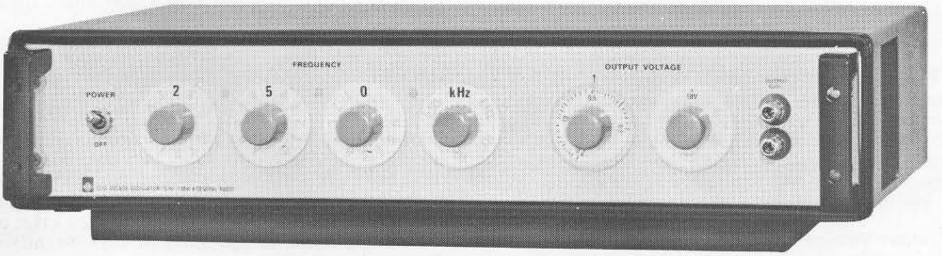


Figure 1. The Type 1312 Decade Oscillator.

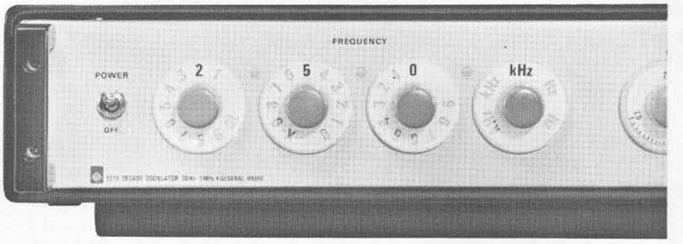


Figure 2. In-line readout of frequency with positioned decimal point and units. Least significant digit is continuously variable with a detent at zero.

and little attempt was made to utilize the discrete steps to provide a more readable or more easily set control. The use of X1, X10, etc range multipliers and confusing decimal-point locations with strange units is still common.

And yet, this type of frequency selection can be quite desirable in its own right. Its greatest advantage is that the same frequency can be reset very accurately, over and over again, limited only by the stability of the oscillator. And it can eliminate perhaps the worst disadvantage of the continuously adjustable dial: difficulty in reading the frequency quickly, accurately, and unambiguously.

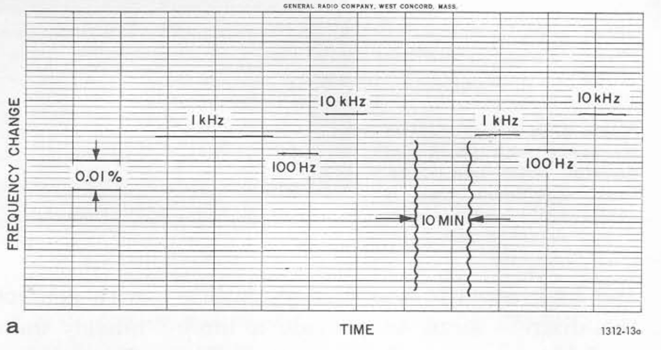
The new General Radio TYPE 1312 Decade Oscillator provides for the first time a general-purpose oscillator with a discrete repeatable type of frequency selection, at a price comparable to that of other general-purpose solid-state oscillators. The 1312 embodies many of the characteristics of the popular TYPE 1310-A[1], with additional features of an 80-dB step attenuator on the output and a rack-width package with front and rear output terminals.

The frequency range is 10 Hz to 1 MHz with an accuracy of $\pm 1\%$ of reading. The frequency is determined by four rotary controls, which provide an in-line three- or four-digit readout with positioned decimal point and frequency units. The decimal-point location and units have been selected so that the frequency appears as it is normally written. Twenty kilohertz appears as 20.0 kHz, not as 20000 Hz, 2.00 X 10 kHz, or 200 Hz X 100. A single exception is that 1 megahertz appears as 1000. kHz, a frequency indicated by some oscillators as 999 kHz.

The first two most-significant digits and the units control are multi-position rotary switches with a light but positive detent mechanism for easy setting. The third digit is a continuously adjustable potentiometer with a detented zero position. This allows the selection of 100 discrete, highly repeatable frequencies within each decade and continuous coverage in between. Rotary controls were selected over push-buttons because of operator preference in high-rate testing. This in-line readout of the rotary switch is generally preferred over the columnar readout of the push-button, just as it is on electronic counters.

The high repeatability of this type of frequency selection is illustrated by the two examples shown in Figure 3. These show the frequency of a typical 1312 after warmup in a normal production environment. Three frequencies, one decade apart, were selected. The same three frequencies 10 minutes later are all within 0.002% of their original

[1] R. E. Owen, "A Modern, Wide-Range RC Oscillator," *General Radio Experimenter*, August 1965.



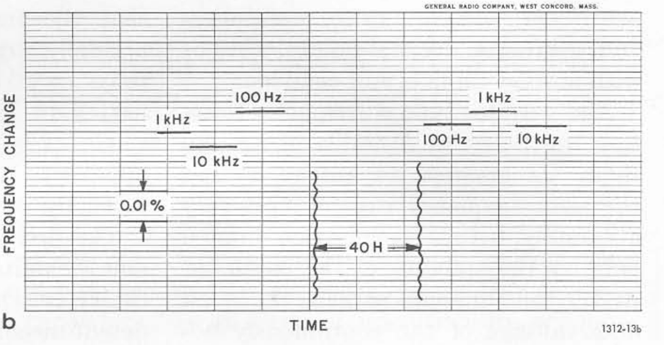


Figure 3. Frequency resettability of the 1312. Difference in frequency at 100 Hz, 1 kHz, and 10 kHz after (a) 10 minutes and (b) 40 hours.

value. Even after 40 hours the change is less than 0.01%. The effects of line-voltage changes on the output frequency are also seen to be small (Figure 4).

The oscillator output is 20 volts open-circuit behind 600 ohms. The voltage is quite constant with changes in frequency (Figure 5). Frequency-response measurements can be made quickly, since controls do not have to be adjusted to keep the output constant. The distortion in the output is low, particularly in the middle of the frequency range, and remains low regardless of the size of the load or attenuator setting. Even a short circuit at full output will not cause clipping of

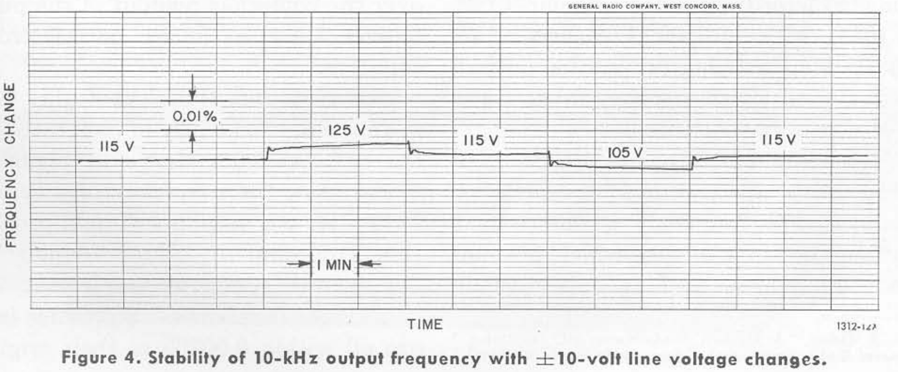


Figure 4. Stability of 10-kHz output frequency with ±10-volt line voltage changes.

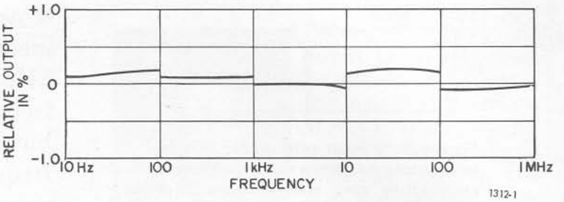


Figure 5. The output-voltage-vs-frequency characteristics of a typical 1312 with 600-ohm load.

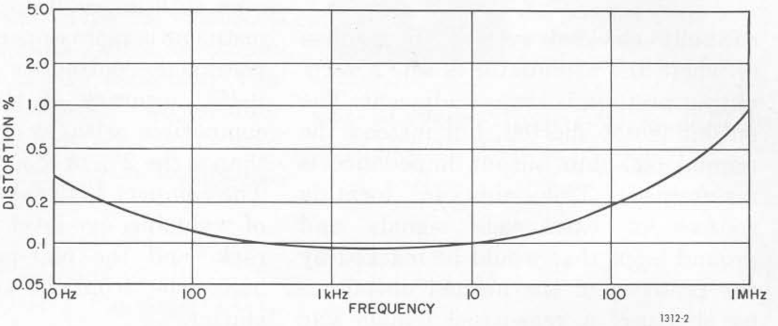


Figure 6. Typical output distortion of the 1312 with 600-ohm load.

the waveform. Noise at frequencies far from a 1-kHz fundamental, measured in a bandwidth of 5 Hz to 500 kHz, is typically less than 0.02%. Noise close to the fundamental is also low, as a close-in spectrum analysis of the 1-kHz fundamental shows (Figure 7).

The output level of the oscillator can be reduced to 200 μV open-circuit by means of a stepped 80-dB attenuator (20 dB/step) and a continuously adjustable attenuator with a range of more than 20 dB. The continuous control is not calibrated but is marked with the approximate open-circuit output voltage. The closely spaced numbered graduations permit convenient return to a given attenuator setting.

Between the 20-dB step positions of the attenuator are so-called "zero-volt-output" positions. This feature, also available on other GR oscillators, provides a convenient transient-free means of reducing the output to zero without disturbing the continuous control setting or shorting or disconnecting a

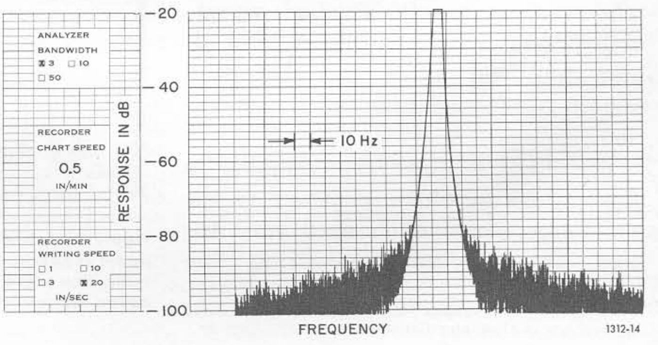


Figure 7. Noise spectrum close to 1-kHz output. Note — 100-dB resolution.

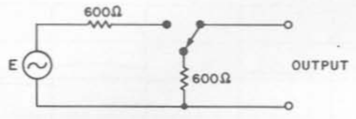


Figure 8. Output attenuator has intermediate positions where voltage is reduced to zero with 600-ohm output impedance maintained.

carefully shielded system. Regardless of where the attenuator is set, a zero-output position is always adjacent. The output is not shorted, but instead the normal 600-ohm output impedance is maintained. This aids in locating sources of extraneous signals and ground loops that would be masked by the presence of the normal output or by shorting. A rear-panel female BNC output is in parallel with the front-panel binding posts.

The 1312 has the external-synchronization feature [2] originally introduced on the General Radio line of RC oscillators. This permits locking the oscillator frequency to an external signal's frequency, and it also provides a constant 0.8-volt output regardless of the attenuator setting. Connection is made by way of a rear-panel female BNC connector.

The 1312 has obvious applications in production or quality-control testing. Repetitive measurements at a variety of widely separated frequencies can be made very accurately. The high repeatability assures uniform testing and allows pretuned filters and distortion meters to be used to speed the operation.

The 1312 is also very valuable in applications in which, contrary to the above, frequency is changed very seldom. In capacitance-measuring systems, for example, almost all measurements are made at 60 Hz, 120 Hz, 1 kHz, 100 kHz, or 1 MHz. Once the frequency is set it is rarely changed, but it must often be verified that the frequency is in fact the desired one. The unambiguous readout makes this easy even with low-skill operators. A misadjustment is more apparent than on conventional continuous dials, and the ±1% accuracy of the 1312 is more compatible with system requirements than is the 2% or 3% of dial oscillators. The compact 1312 takes up a minimum of valuable eye-level space in a test rack, and the rear-panel connections leave the front of the rack free of clutter.

[2] General Radio *Instrument Note* IN-109 "Principles and Applications of Oscillator Synchronization."

How It Works

The 1312 Decade Oscillator uses the modified Wien circuit shown in Figure 10. This circuit oscillates (i.e., its trans-

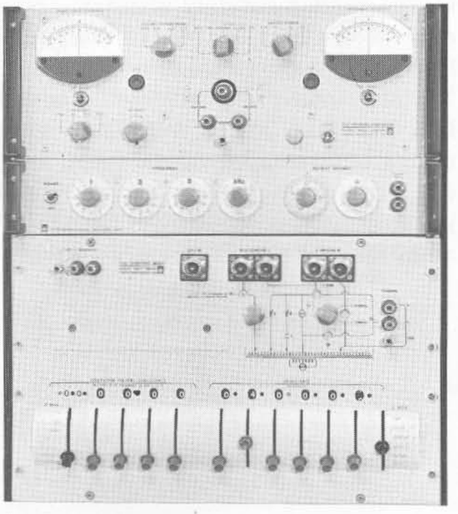


Figure 9. A low 3½-inch rack panel height, rear panel connections, and an easy-to-read frequency indication make the 1312 ideal for systems. Verification of correct operating frequency takes but a glance.

fer function, $\frac{e_o}{e_i}$, is real and at maximum) at a frequency

$$\omega_o = \frac{G}{C}\sqrt{1+\frac{\alpha g}{G}}$$

The values of capacitance C are switched in decade steps by means of the dimensional-units switch to obtain the five ranges of the oscillator. The conductance G comprises two conductance decades that determine the two most significant digits of the frequency. That is

$$G = \overline{G}\,(L+M),$$

where $\overline{G}$ is a normalizing conductance, L varies from 0.1 to 1.0 in steps of 0.1,

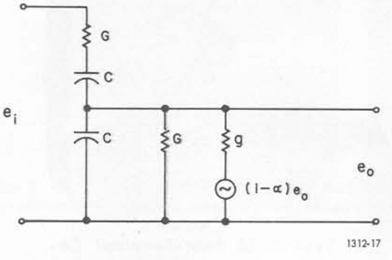


Figure 10. The RC frequency-determining network used in the 1312.

Robert E. Owen received his B.E.E. from Rensselaer Polytechnic Institute in 1961 and his M.S.E.E. from Case Institute of Technology in 1963. He came to GR as a development engineer in 1963 and has for the past few years concentrated on RC oscillator designs.

and M varies from 0.00 to 0.10 in steps of 0.01. Then

$$\omega_o = \frac{\overline{G}}{C}(L+M)\sqrt{1+\frac{\alpha\,\mathrm{n}}{L+M}}$$

$$= \frac{\overline{G}}{C}\left[L+M+\frac{\alpha\mathrm{n}}{2}-\frac{\alpha^2\mathrm{n}^2}{L+M}\cdots\right]$$

where $\mathrm{n} = \frac{\mathrm{g}}{\overline{G}}$. If $\alpha\,\mathrm{n} \ll 1$ then

$$\omega_o \approx \frac{\overline{G}}{C}\left[L+M+\frac{\alpha\,\mathrm{n}}{2}\right].$$

n is made equal to 0.02 so that $\frac{\alpha\mathrm{n}}{2}$ varies continuously from 0.000 to 0.009 as α varies from 0 to 0.9.

R. E. Owen

SPECIFICATIONS

FREQUENCY

Range: 10 Hz to 1 MHz in five decade ranges.

Accuracy: $\pm 1\%$ of setting.

Stability (typical at 1 kHz): Warmup drift, 0.1%. After warmup: 0.001% short term (10 min), 0.005% long term (12 h). Resettable within 0.005%.

Control: Step control of two most significant digits, continuously adjustable third digit with detented zero position. In-line readout with positioned decimal point and frequency units.

Synchronization: Frequency can be locked to external signal. Lock range $\pm 3\%$ per volt rms input up to 10 V. Frequency controls function as phase adjustment.

OUTPUT

Voltage: >20 V open circuit.

Power: >160 mW into 600 Ω.

Impedance: 600 Ω. Isolated from chassis by 10 Ω across 0.1 μF.

Attenuation: Continuously adjustable attenuator with >20-dB range, and 80-dB step attenuator with 20 dB per step. Intermediate steps reduce output to zero while maintaining 600-Ω output impedance.

Distortion: <0.25%, 50 Hz to 50 kHz with any linear load. Oscillator will drive a short circuit without clipping.

Hum: <0.04% of max output or 4 μV, whichever is greater.

Amplitude vs Frequency: ±2%, 10 Hz to 100 kHz with >600-Ω load; ±4%, 100 kHz to 1 MHz with <600-Ω load.

Synchronization: Constant-amplitude (0.8-V) high-impedance (27-kΩ) output to drive counter or oscilloscope.

SPECIFICATIONS (continued)

GENERAL

Power Required: 100 to 125, 200 to 250 V, 50 to 400 Hz, 13 W.

Terminals: Front-panel output, GR 938 Binding Posts; rear-panel output, female BNC connector. Sync, rear-panel, female BNC.

Accessories Supplied: Power cord, spare fuses.

Accessories Available: 776-A Patch Cord (BNC to shielded double plug).

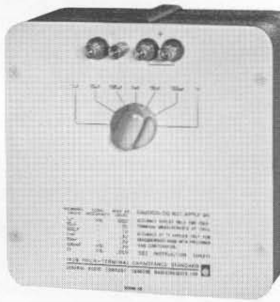
Mounting: Rack-bench cabinet.

Dimensions (width x height x depth): Bench, 19 x 3⅞ x 11 in. (485 x 99 x 330 mm); rack, 19 x 3½ x 8⅞ in. (485 x 89 x 225 mm).

Weight: Net, 13¼ lb (6.5 kg); shipping, 17 lb (8 kg).

Catalog Number	*Description*	*Price in USA*
	1312 Decade Oscillator	
1312-9700	Bench Model	**$415.00**
1312-9701	Rack Model	**415.00**

STANDARDS OF LARGE CAPACITANCE



Type 1426 Four-Terminal Capacitance Standard.

One problem in making wide-range bridges is that of providing adequate standards to check them with. We designed the 1617 Capacitance Bridge[1] to measure capacitors up to 1.1 farads because electrolytic capacitors are available up to this value. However, electrolytic capacitors make poor standards because their value varies so much with temperature, voltage, and time, and because such high values have high loss (high *D*). Paper or plastic capacitors of this value would be extremely expensive. We catalog a 100-μF polystyrene capacitor that sells for $1950. Who could afford a farad at $19,500,000? An alternative approach seemed necessary.

[1] H. P. Hall, "The Measurement of Electrolytic Capacitors," *General Radio Experimenter*, June 1966.

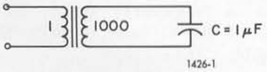


Figure 1. Standard 1-farad capacitor using impedance transformation.

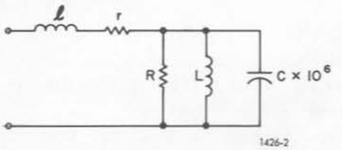


Figure 2. An equivalent circuit for capacitor of Figure 1.

As we all remember, a transformer transforms impedance by the square of the turns ratio. Thus, our standard capacitor could look like Figure 1. We would use a convenient 1-μF capacitor and a 1000:1 turns ratio. This was tried and worked rather well for medium capacitance values. An equivalent circuit for such a "capacitor" is shown in Figure 2. The transformer series leakage inductance, ℓ, increases capacitance, the parallel inductance, L, decreases capacitance, and both r and R cause D to increase. The deterioration caused by

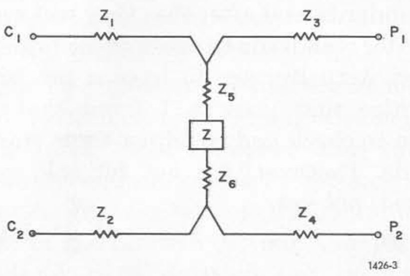


Figure 3. A four-terminal impedance. Only impedances Z_5 and Z_6 affect the measured value.

these series and parallel impedances can be kept reasonably low for rather high values, but at 1 farad the series impedances just can't be kept low enough. At 120 Hz, the standard test frequency, the reactance of 1 farad is only 1.3 milliohms, and the resistance of binding posts, not to mention that of leads, would give excessive D, while an inductance of 0.2 μH would cause a 10% capacitance error.

The bridge that we are trying to calibrate, the 1617, is a "four-terminal" bridge, designed to tolerate substantial series resistance and inductance without error. However, series impedance can be tolerated only in the four-lead system. Once it becomes a two-lead system, such impedances become part of the

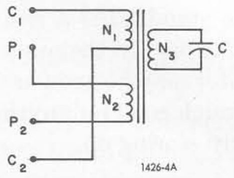


Figure 4. Elementary schematic diagram of Type 1426 Four-Terminal Capacitance Standard.

impedance being measured (see Figure 3).

The standard itself should be four-terminal as long as it is low impedance. On the secondary, high-impedance side of the transformer, a reasonably low lead impedance causes negligible error. The 1426 Four-Terminal Capacitance Standard is designed this way, as shown in Figure 4. The four-terminal capacitance is now ideally

$$C\,\frac{N_3^2}{N_1 N_2}.$$

Actually, there are still errors caused by magnetizing inductance, core loss, and secondary impedance (Figure 5), but these are relatively small and can be taken into account in a calibration at one frequency.

Both primary windings of the 1426 can be switched to provide the various decade capacitance values. Each value has a separate padding capacitor to obtain good accuracy. On the lowest position, the input is directly connected to the capacitor so that in this position

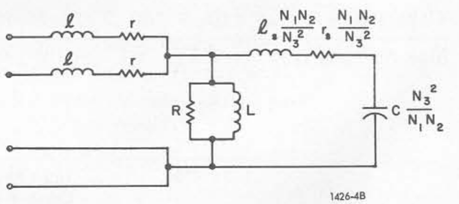


Figure 5. An equivalent circuit for the 1426.

(1 μF) the standard is a real capacitor and acts like one. In the other positions, the capacitor may be good as a standard but isn't much good for anything else — particularly storing dc.

Characteristics

The construction described above explains most of the special characteristics given in the specifications. The unit is frequency-dependent because of inductance, and it is slightly voltage-level-sensitive because the magnetizing inductance depends on the voltage level. While the voltage ratings given seem very small, the currents are rather large. The voltage applied by the 1617 bridge is less than the rated value of the 1426.

The standard may be used as a two-terminal capacitor up to 10 mF (10,000 μF), but the capacitance value will be a few percent high and the *D* will be somewhat high.

Unless the transformer is damaged, the calibration of all values should change by the same percent with time or temperature. Thus, if the 1-μF value is checked on an accurate bridge and shows a variation from its nominal value, all values can be corrected by the same percent. This procedure should make it unnecessary to have calibration backup for this standard once it has been initially checked. However, we're sure that there will be some people who will want a precision bridge that goes to 1 farad to check these standards, and after that they will want better standards to check those bridges, etc. Actually, we do have a precision bridge that goes to 1 farad that we use to check and to adjust these standards. However, it's not for sale — at least, not yet.

H. P. Hall

A brief biography of Mr. Hall appeared in the June 1966 *Experimenter*.

SPECIFICATIONS

Capacitance: 1 μF to 1 F in 7 switch-selected decade values.

Accuracy: $\pm\frac{1}{4}\%$, except $\pm\frac{1}{2}\%$ for 100 mF and $\pm 1\%$ for 1 F; measured at 120 Hz at 23°C at < Max Volts specified below. Measurements must use 4-terminal connections with all but the 1-μF value; at 1 F, lead arrangement must be as prescribed in operating instruction manual.

Dissipation Factor: < 0.0003 for 1 μF at 120 Hz; < 0.1 for larger values.

Max Ac Voltage: Voltage 10:1 larger than specified will not damage standards above 1 μF but will cause an error of approx 1%.

Max Dc Voltage: No dc permissible as values above 1 μF are dc short circuits and could be changed in value by dc current; 100 V max for 1-μF standard only.

Temperature Coefficient: 140 ppm/°C typical.

Frequency Characteristics: 1-μF standard is true capacitor with 170-kHz resonance; other values very frequency dependent. Add $\frac{1}{4}\%$ error at 100 Hz, add 1% from 60 to 150 Hz.

Mounting: Aluminum cabinet.

Dimensions (width x height x depth): 8 x 5⅞ x 8 in. (205 x 150 x 205 mm).

Weight: Net, 7½ lb (3.5 kg); **shipping,** 11 lb (5 kg).

Capacitance	1 μF	10 μF	100 μF	1 mF	10 mF	100 mF	1 F
Max Ac Volts	100 V	2 V	1 V	0.3 V	0.2 V	0.2 V	0.05 V

Catalog Number	*Description*	*Price in USA*
1426-9700	**1426 Four-Terminal Capacitance Standard**	**$275.00**

THIESSEN RETIRES

Arthur E. Thiessen, Chairman of our Board of Directors and for many years the Company's chief sales executive, has retired after a 39-year career with General Radio. A 1926 graduate of Johns Hopkins University, he joined GR in 1928, became Commercial Engineering Manager in 1936, Vice President for Sales in 1944, and Chairman of the Board in 1960.

GR's marketing activities have expanded manyfold under his stewardship and now include a strong domestic sales force, three European sales subsidiaries, and an extensive network of overseas representatives. His important role in the growth of our Company made it particularly fitting that he chronicle this growth in his *A History of the General Radio Company*, published in 1965.

In addition to his duties at General Radio, he has found time to participate in the activities of many professional and industry organizations, including the Institute of Radio Engineers (now the IEEE), the Electronic Industries Association, and the Scientific Apparatus Makers Association, which presented him its 1967 Award in recognition of outstanding contributions to the development of the scientific instrument industry.

His responsibilities at General Radio are now undertaken by three men: D. B. Sinclair, who continues as President and will preside at meetings of the board of directors; W. R. Thurston, Vice President for Marketing, who assumes full responsibility for domestic marketing; and S. W. deBlois, Manager of our International Division.

OF THE FUTURE

Electronics is a big industry, reputedly the third largest in the nation today. To have been a part of it almost from its beginnings has been a rich experience for the Company and an absorbing and challenging one for those who have worked for it. Every day new advances in the art are made in scientific laboratories all over the world. These advances, or discoveries, some in small ways and some in large, continually broaden the already immense field of electronics and make it ever more useful to mankind. And in all of this progress, scientific and material, men strive by means of measurements to assign numbers to those phenomena that they observe and use. Those numbers are essential to the scientific and technological processes. They are the invaluable means of communication for the exchange of knowledge among the practitioners; and they are the links from the scientist to the engineer to the production line to the market.

As the art advances, so does the need for better, more accurate, faster, and more sophisticated measuring instruments. It will be General Radio's job for the future, as it has been in the past, to make available those instruments for progress. In so doing, the Company and the men and women associated with it will know an ever more promising and exciting future.

from *A History of the General Radio Company*, by A. E. Thiessen.

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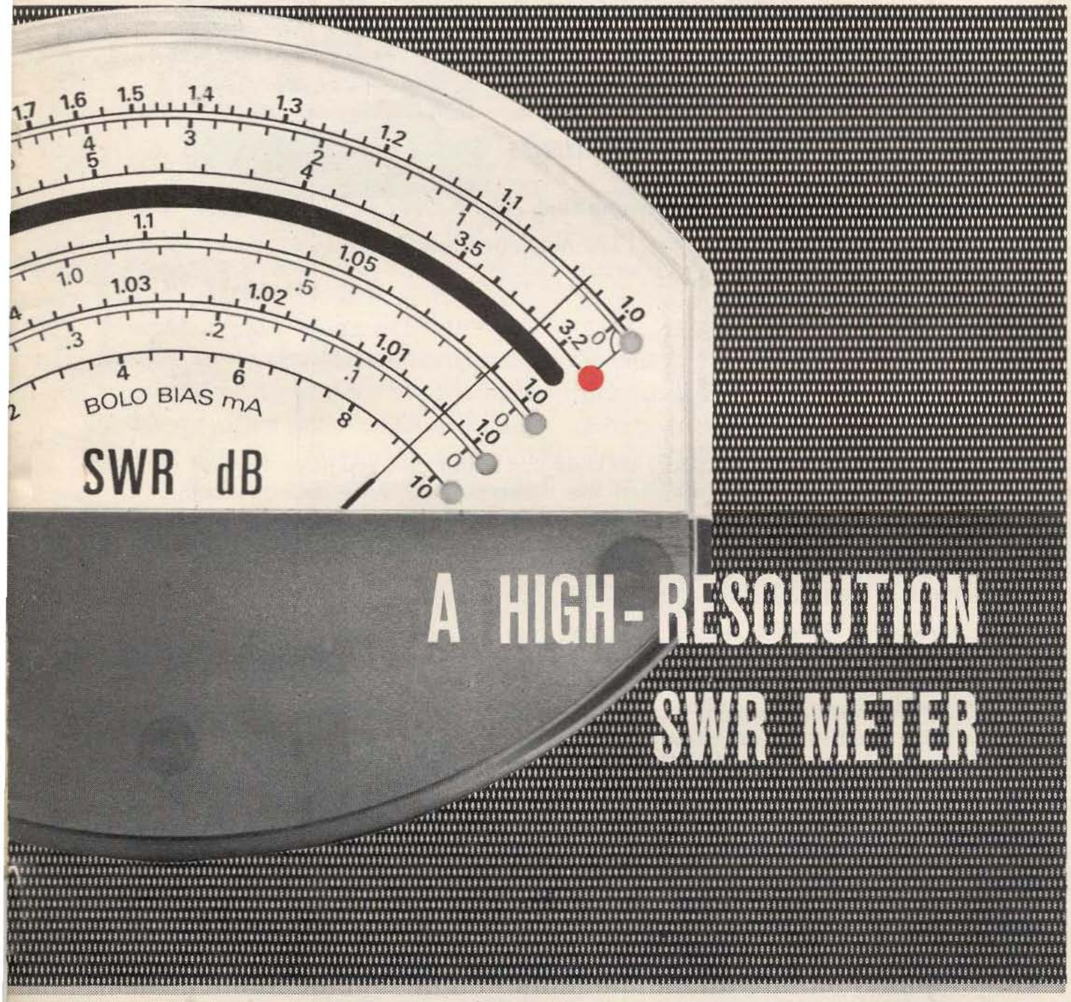


A HIGH-RESOLUTION SWR METER

VOLUME 42 · NUMBER 2 / FEBRUARY 1968

the Experimenter

Volume 42 • No. 2 February 1968

Published monthly by the General Radio Company

CONTENTS

Put a 1.0-to-1.05 SWR scale on a six-inch meter and you have the kind of resolution that an accurate slotted line deserves. Add three step attenuators and an attenuator "memory" dial and you have the means for fast, accurate attenuation measurements. Another feature of the Type 1234 Standing-Wave Meter is in the "why-doesn't-everyone-do-that?" category: a set of meter lights (see front cover) to tell you which scale to read.

We have gone on record as saying that the pulse-output possibilities of our Type 1395 Modular Pulse Generator are virtually limitless. To cement the case we introduce in this issue still another module: an NRZ (Non-Return-to-Zero) Converter/Sampler. In addition to its primary function of preventing voltage in a binary word from falling to zero until a zero is called for, the new unit can be coupled with a random-noise generator to yield random binary sequences.

Users of GR frequency synthesizers will want to note two important new accessories described in this issue: a standard-frequency oscillator and a programmable digit-insertion unit. The line is now fully programmable, right up to 70 MHz.

The *General Radio Experimenter* is mailed each month without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.

AN SWR METER FOR PRECISION MEASUREMENTS

Figure 1. Type 1234 Standing-Wave Meter.

Since the advent of precision coaxial connectors and precision slotted lines, SWR values have been meaningfully expressed in terms of hundredths and thousandths. An obvious need in measurements with a precision slotted line is an SWR meter with enough resolution to translate the precision of the slotted line into precise measurement results. This need is now filled by GR's new TYPE 1234 Standing-Wave Meter, which numbers among its features a large scale that is expandable to a full-scale SWR value of only 1.05.

The new SWR meter is no less useful for larger SWR values, and its three step attenuators and an attenuator "memory" control permit fast, accurate attenuation measurements.

One of the chief objectives in the design of the 1234 was ease of operation and of reading, and the instrument features several interesting innovations in the so-called "human-engineering" area. For example, the usual annoyance and frequent confusion associated with a multiscale meter are neatly dispatched: The scale in use is always unmistakably identified by a light adjacent to it (see front cover photo). Once one is on the right scale, reading it is no problem; the meter is of the familiar and highly popular GR design, with a large 6-inch face, a mirror scale, and a tracking accuracy or linearity of ½ percent.

Description

The 1234 is basically a low-noise tuned ac amplifier, calibrated for use with square-law detectors. The circuit (Figure 2) comprises five stages of audio amplification and four stages of controlled attenuation, staggered to provide optimum signal-to-noise ratio. The total attenuation range is 70 dB, of which 20 dB is controlled by the meter range switch. This feature, which simplifies high (> 4) SWR measure-

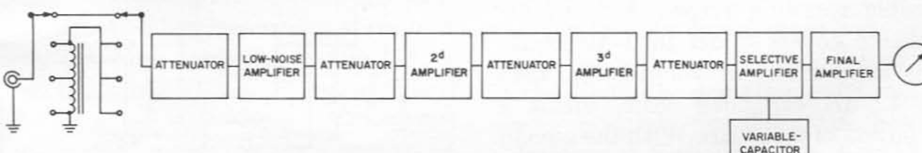


Figure 2. Block diagram of the 1234.

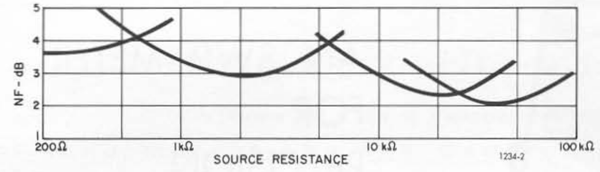


Figure 3. Typical noise figure as a function of source resistance.

ments by eliminating the need for adjustment of the attenuator control, also limits use of the expanded scales to measurements with at least 20 dB in the circuit — in other words, to measurements where the signal-to-noise ratio permits accurate measurements.

At the input, an rf low-pass filter prevents rf signals that leak past the detector bypass capacitor from causing measurement errors. The input stage is a high-input-impedance, low-noise amplifier, whose optimum source resistance is adjustable between 200 ohms and 35 kilohms in four steps. Figure 4 is a plot of noise figure vs source resistance.

The amplifier circuit is designed so that gain is essentially independent of frequency and bandwidth adjustments. The frequency-selective circuit (Figure 3) is a feedback amplifier in which positive and negative feedback are equal at resonance, and the feedback circuit does not affect circuit gain. At other frequencies, however, the positive feedback is less than the negative feedback, reducing gain.

The main attenuator covers a range of 45 dB in nine steps. A "memory" dial permits attenuation measurements by substitution techniques to be made without subtraction of readings and possible resulting errors. A third attenuator covers 5 dB in 1-dB steps. This attenuator, in conjunction with the 1.6-dB expanded scale, yields a resolution of 0.025 dB. With the second expanded scale the resolution is 0.005 dB (i.e., 0.02 dB per small, ⅛-inch-wide division).

Two outputs are available for use with recorders or other auxiliary equipment: a dc output of 1.5 volts behind 1.5 kilohms at meter full scale and a 1-kHz output of 0.1 to 1 volt rms maximum (depending on range switch position) behind 500 ohms.

A 60-ohm impedance between circuit ground and line ground reduces potentially troublesome ground loops. Errors are eliminated when recorders with balanced inputs are used and are usually small even with unbalanced-input recorders.

A highly regulated power supply makes amplifier gain virtually independent of line-voltage changes. Figure 5 shows a barely noticeable gain change on the highest expansion range as the line voltage is changed ±10%. The 1234 can be operated from battery as well as ac voltage. Any battery capable of supplying 90 mA at from 22 to 35 volts can be used. Gain will not be affected noticeably as long as the battery voltage stays above 22 volts. The TYPE 1538-P3 Battery and Charger unit is available as an accessory. The 1234 will operate for 40 hours on one charge of this nickel-cadmium battery.

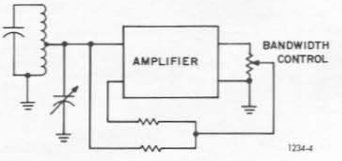


Figure 4. Frequency-selective amplifier circuit.

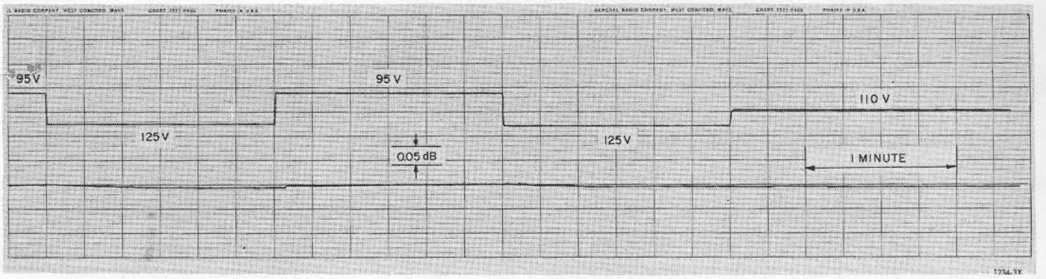


Figure 5. Recording showing gain stability with line-voltage changes, as measured on a production instrument.

Use with Bolometers

The bolometer is often preferred over the crystal detector for accurate measurements. The signal level must be at least five times the residual noise level for an error of 0.1 dB or less. The power level required to produce such a meter deflection has been measured on a typical instrument to be −52 dBm peak, with a 100%-modulated signal. The high limit for a 0.1-dB error is generally about 0 dBm. For errors of less than 0.05 dB the peak power must be limited to the range of −45 to −15 dBm.

Accurate measurements at high signal levels require a high input impedance. In the 1234, the input circuit is designed to provide an optimum signal-to-noise ratio over a wide range of source resistances, while presenting a high input impedance to the source. In the two bolometer-input positions of the SOURCE switch, this optimum source resistance is 200 ohms, while the input impedance is 3.5 kilohms in parallel with 5 kilohms inductive reactance. At high input-signal levels, a low impedance will increase the bolometer's deviation from square law. This error is sometimes called the voltage-transfer error.[1] A high load impedance (i.e., a high input impedance of the amplifier) will reduce this effect to the point where it may become negligible.

Another factor affecting bolometer accuracy is the bias-current supply. If this current is not supplied from a high-impedance source, the source impedance may be considered to appear in parallel with the input impedance of the amplifier, thus aggravating the voltage-transfer error. The 1234 supplies a bolometer dc bias current from a true current source (source resistance about 10^5 ohms). The current is adjustable ±10% of the nominal value by means of a potentiometer accessible through a hole in the rear cover of the instrument. The current source is voltage-limited to protect the bolometer element.

Use with Crystal Detectors

The crystal diode, although its dynamic range is less than that of a bolometer, is widely used as a detector because of its higher sensitivity. The power level required to produce a meter deflection five times the noise level is about −60 dBm. The upper limit, where the deviation from square law becomes significant, is about −30 dBm. This level can be raised by the selection of a proper load impedance for the crystal. At impedances very low compared with the dynamic resistance of the diode, the deviation from square law is positive — that is, the output voltage or current increases more than

[1] G. J. Sorger and B. O. Weinschel, "Comparison of Deviations from Square Law for R. F. Crystal Diodes and Barreters," *IRE Transactions on Instrumentation*, Vol. I-8, No. 3, December 1959, pp 103–111.

the input rf power does. With very high load impedances the deviation from square law is negative, since the detector approaches the linear operating region.

We found that the best results are obtained when the load resistance is about equal to the dynamic resistance of the diode. Measuring the dynamic resistance is a simple matter of connecting a resistor across the top of the detector and calculating R_D from the resulting voltage drop. Dynamic resistances of point-contact diodes vary from 5 to well over 100 kilohms. A shunt resistor to reduce the error at high levels will of course decrease the detector sensitivity, but the ability to change the optimum source resistance helps to reduce this loss in sensitivity, so that this method can be used to increase dynamic range significantly.

An example appears in Figure 6. The diode dynamic resistance is 7 kilohms. The 6.8-kilohm shunt resistor reduces sensitivity by 2 dB. For a maximum error of 0.1 dB on the high-level and 0.1 dB at the low-level end,

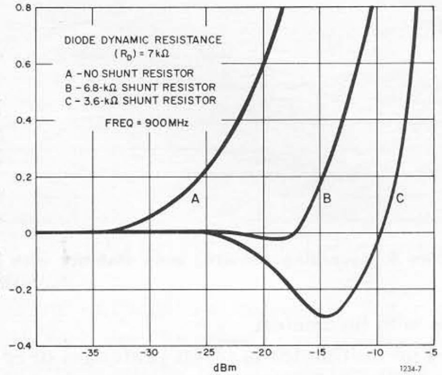


Figure 6. Extension of the square-law range of a diode detector by a shunt resistor across the detector output.

the dynamic range is increased from 35 to 45 dB.

When the meter deflection is at least five times the residual noise deflection, the error due to the noise contribution is less than 0.1 dB.[2] However, the ability to read the meter correctly is limited by the fluctuations of the meter needle. The "slow," or damped, meter response reduces these fluctuations to the point where accurate readings are possible.

— M. Khazam

A brief biography of Mr. Khazam appeared in the July-August 1967 issue of the *Experimenter*.

[2] W. R. Bennett: "Response of a Linear Rectifier to Signal and Noise," *Bell Systems Technical Journal*, Vol. 23, January 1944, pp 97–113.

SPECIFICATIONS

Input:	Crystal				Bolometer
Optimum Source R	35 kΩ	20 kΩ	2 kΩ	200 Ω	200 Ω
Input Impedance	1 MΩ	350 kΩ // 80 H	35 kΩ // 8 H	3.5 kΩ // 0.8 H	3.5 kΩ // 0.8 H
Sensitivity (fs)	1.2 μV	1 μV	0.32 μV	0.1 μV	0.1 μV
Noise*	0.2 μV	0.2 μV	0.06 μV	0.02 μV	0.02 μV

*Equivalent input noise level with source resistance equal to optimum and with minimum bandwidth.

Meter Scales: SWR, 1 to 4, 3.2 to 10, 1 to 1.2, and 1 to 1.05; dB, 0 to 10, 0 to 1.6, and 0 to 0.45; bolometer current, 0 to 10 mA.

Meter Accuracy: 0 to 10-dB scale, ±(0.01 dB + 1.5% of reading); 0 to 1.6-dB scale, ±0.02 dB; 0 to 0.45-dB scale, ±0.007 dB.

Attenuator: Three separate attenuators: 20 dB in 10-dB steps, accuracy ±0.1 dB/10 dB; 45 dB in 5-dB steps, accuracy ±0.05 dB/5 dB; 5 dB in 1-dB steps, accuracy ±0.01 dB/1 dB.

Bandwidth: 10 to 100 Hz, adjustable with constant gain.

Frequency: 1 kHz, adjustable ±30 Hz.

Gain Control: Coarse and fine, 6-dB range.

Bolometer Bias Current: 4.3 and 8.7 mA, adjustable ±10%. Voltage limited for bolometer protection.

Meter Speed: Slow and fast, switch selected.

Outputs: Dc, 1.5 V max behind 1500 Ω. Ac, 0.1 V rms (1 to 4 SWR range), 0.3 V rms (1 to 1.2 range), and 1 V rms (1 to 1.05 range); 500-Ω source impedance. Load resistance > 6000 Ω.

GENERAL

Power Required: 100 to 125 or 200 to 250 V, 50 to 60 Hz. Or 22 to 35 V dc, 90 mA from ext battery, 1538-P3 Battery and Charger suitable.

Accessories Supplied: Spare fuse, battery connector.

Accessories Available: 1538-P3 Battery and Charger.

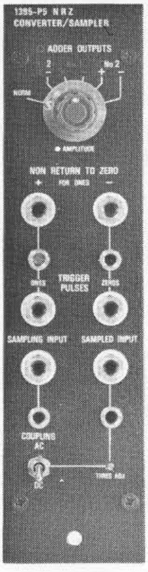
Mounting: Flip-Tilt case.

Dimensions (with x height x depth): 8⅜ x 8¾ x 11¼ in. (215 x 225 x 290 mm).

Weight: Net, 9 lb (4.1 kg); **shipping,** 12½ lb (6.0 kg).

Catalog Number	*Description*	*Price in USA*
1234-9701	**1234 Standing-Wave Meter**	**$495.00**

NRZ AND RANDOM PULSES FROM THE 1395 MODULAR PULSE GENERATOR



NRZ Converter/Sampler.

It often happens in the course of work on digital equipment that binary words of non-return-to-zero (NRZ) pulses are required. The 1395-A Modular Pulse Generator[1] is now available with a new plug-in module, the 1395-P5 NRZ Converter/Sampler, designed to provide NRZ pulses when used in conjunction with the Type 1395-P6 Binary-Word Generator.

The NRZ Converter/Sampler is easy to use. A timing signal, normally derived from the same master clock (a Type 1395-P1 PRF Unit) that drives the Type 1395-P6 Word Generator, commands the NRZ Converter/Sampler to examine the signal at its SAMPLED input. The signal at this input is the output of the Word Generator, stretched with the use of a 1395-P2 Pulse/Delay Unit. The NRZ Converter determines whether the SAMPLED input is in the one or zero state at the moment the SAMPLING pulse arrives. The appropriate output terminal then assumes a high or low voltage and holds that state until the sampling process finds the opposite state at the input.

There are two output terminals for the NRZ signal: One gives high-level voltage for ones; the other gives low-level voltage for ones. Both outputs are available simultaneously. The exact voltage levels are set at the user's pleasure by a gain control on the NRZ Converter/Sampler and by the DC Component control on the 1395-A

[1] G. R. Partridge, "Pulses to Order," *General Radio Experimenter*, May 1965.

Gordon R. Partridge received his BE, ME, and PhD degrees in Electrical Engineering at Yale University. His academic background also includes an associate professorship in EE at Purdue University. He joined GR in 1962 as a Development Engineer, and he has since specialized in the design of pulse generators and amplifiers. He is the author of *Principles of Electronic Instruments*, published by Prentice-Hall in 1958.

Modular Pulse Generator (main frame), which sets the baseline voltage.

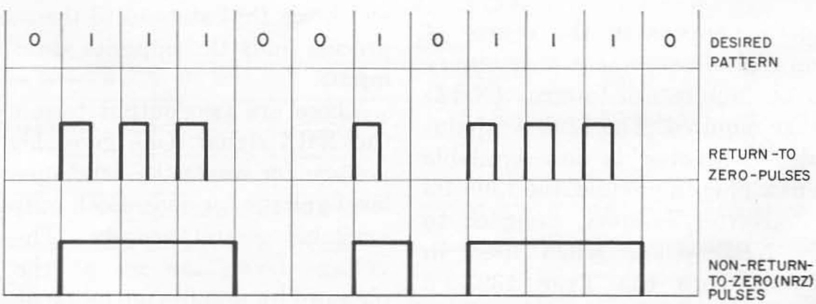
Since trigger pulses as well as NRZ pulses may be desired, the NRZ Converter/Sampler also provides triggers corresponding to its one or zero decisions. The user may select trigger pulses for ones or trigger pulses for zeros; both are available simultaneously at front-panel output jacks.

Communication links, whether in the sense of person-to-person, as in pulse-code modulated telephony, or between business machines involving binary data, normally are subject to interference from random trains of pulses at the same bit rate as the message. Such trains may be generated for test purposes by the NRZ Converter/Sampler operating with a random-noise generator such as the GR 1381 or 1382.[2] Unlike systems of pseudo-random noise generation, this method does not restrict the maximum number of ones or zeros it is possible to obtain in a continuous sequence. The result is a truly random set of ones and zeros.

The noise-sampling technique is also useful for producing low-frequency white noise. The power spectrum of a random sequence of non-return-to-zero ones and zeros is given by Rice.[3]

Specifically, this spectrum follows a $(\sin x/x)^2$ law, where x is the product πfh. The quantity h is the time between SAMPLING pulses, and f is the frequency at which the power density in watts/cycle is being evaluated. As f approaches zero, the value of $\sin x/x$ approaches unity, showing that at low frequencies, the power density is essentially independent of frequency and therefore "white." The table below

[2] J. J. Faran, "Random-Noise Generators," *General Radio Experimenter*, January 1968.
[3] S. O. Rice, "Mathematical Analysis of Random Noise," *Bell Systems Technical Journal*, Vol. 23, pp 282–332, July 1944. In particular, note Rice's equation 2.7-9.



What is an NRZ pulse? In the above diagram, a binary word (top line) is shown as converted to return-to-zero (center) and non-return-to-zero (bottom) pulse sequences. In return-to-zero sequence, voltage drops to zero at end of each "1" pulse, whether or not a "0" follows. With NRZ pulses, voltage returns to zero only at binary "0".

shows the relationship between whiteness and the SAMPLING frequency, which is the reciprocal of Rice's time interval h.

FREQUENCIES AT AND BELOW WHICH NOISE IS "WHITE" WITHIN TOLERANCES LISTED

Sampling Frequency (Hz)	*Frequency in hertz equal to or less than (0.1 dB)*	*(1.0 dB)*	*(3.0 dB)*
1	0.083	0.261	0.440
3	0.249	0.783	1.32
10	.83	2.61	4.40
30	2.49	7.83	13.2
100	8.3	26.1	44.0
300	24.9	78.3	132.
1000	83.	261.	440.
3000	249.	783.	1320.
10000	830.	2610.	4400.
30000	2490.	7830.	13200.
100000	8300.	26100.	44000.

The random binary pulse train is available at the NRZ output terminals of the NRZ Converter/Sampler. No other plug-ins are required except a 1395-P1 PRF Unit to generate a SAMPLING frequency. The total cost of such a system for generating random binary pulses, including the required random-noise generator, is then better than a thousand dollars under that of the most nearly similar commercially available generator of random binary sequences.

Since the NRZ Converter/Sampler gives a trigger pulse output each time a SAMPLING pulse is received, it follows that the basic assembly just described may be enlarged by the addition of other 1395-family plug-ins. Thus, the virtually limitless variety of pulses that can be produced with a 1395 Modular Pulse Generator are now available in random sequence.

— G. R. PARTRIDGE

SPECIFICATIONS

SAMPLING INPUT

Pulses: 10 to 15 V, positive-going, 75- to 150-ns duration, dc to 2.5 MHz.

Sine Wave: At least 17 V rms, 1 to 2.5 MHz. Sine-wave sampling below 1 MHz not recommended.

Input Impedance: Approx 4500 Ω across 40 pF.

SAMPLED INPUT

Sensitivity: 0.2 V pk-pk up to 2.5 MHz, optimized by adjustment of threshold control. Input required increases at other settings.

Coupling: Ac or dc, switch-selected.

Threshold Control: Compensates for dc components between approx ±0.6 V.

Input Impedance: Approx 100 kΩ across 40 pF.

NRZ OUTPUTS Both positive- and negative-going transitions are available simultaneously with the dc component controlled by the 1395.

Amplitude: >20 V, open circuit: >1 V across 50 Ω.

Transition Times: <15 ns with 50-Ω load at max input. Typically 80 ns + 2 ns/pF of load capacitance for high-impedance loads.

Output Impedance: 1 kΩ max.

TRIGGER OUTPUTS Two outputs available simultaneously: pulse generated when SAMPLED terminal is in ONE state, another when terminal is in ZERO state.

Amplitude: > +10 V. **Duration:** Approx 70 ns.

Output Impedance: Approx 160 Ω.

Delays: Trigger and NRZ outputs are delayed approx 190 ns from the SAMPLING input.

Accessories Supplied: Eight patch cords — two each TYPES 274-LMB, 274-LMR, 274-LSB, 274-LSR; two double plugs, four insulated plugs.

Catalog Number	*Description*	*Price in USA*
1395-9605	**1395-P5 NRZ Converter/Sampler**	**$295.00**

A GR900® COMPONENT MOUNT

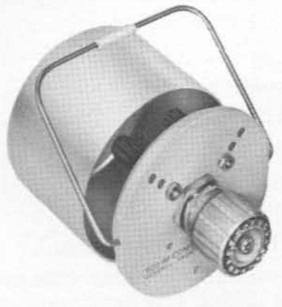


Figure 1.
Type 900-M Component Mount.

The gateway to many precision measurements is the GR900® precision coaxial connector, but if what you are measuring has a simple pair of wire leads, the problem becomes that of finding a gateway to the gateway. Just adapting from wire lead to coaxial is not enough; the transition must be made in such a way that the component is measured with the same lead length and dressing as it will have in the circuit in which it is used, and the leads must be positioned to produce minimum inductance or capacitance.

The solution to the problem is the new TYPE 900-M Component Mount, which permits reproducible measurements of wire-lead components (e.g., resistors, capacitors, inductors, diodes, transistors) at a well-defined reference plane and which also serves as a coaxial packaging device for the permanent mounting of components to be used as standards or terminations.

It is a particularly useful accessory for GR900-equipped instruments, such as the 1606-B RF Bridge, the 900-LB Precision Slotted Line, and the 1609 Immittance Bridge. It is also a very convenient housing for capacitance standards, which can be accurately calibrated on a GR900-equipped 1615-A Capacitance Bridge.

The mount, shown in Figure 1, consists of a GR900 precision coaxial connector, a length of coaxial line, a mounting disk, and a removable cover. The inner conductor of the coaxial line is kept from rotating by a dielectric rod that has been compensated so that it is essentially reflectionless. The line length (i.e., the electrical length of the mount) is exactly 4 cm to correspond to the electrical lengths of the 900-WN4 and -WO4 short and open circuits.

A sample measurement of a 1000-pF disk ceramic capacitor was performed to demonstrate the use of the mount. The lead lengths were $\frac{3}{32}$ inch. The resulting reactance-vs-frequency characteristic is shown in Figure 2.

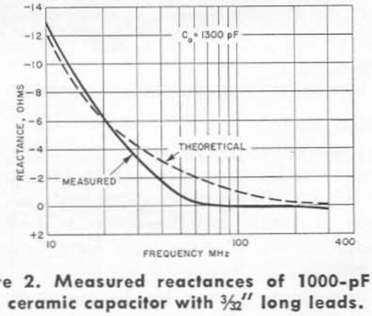


Figure 2. Measured reactances of 1000-pF disk ceramic capacitor with $\frac{3}{32}$" long leads.

SPECIFICATIONS

Electrical Length: 4.0 cm ±0.04 cm to ideal short circuit at terminals.

Residual Capacitance (at low frequencies): 2.93 pF, typical, with screw.

Accessories Required: 900-WN4 Precision Short Circuit, 900-WO4 Precision Open Circuit for establishing reference plane.

Weight: Net, 8 oz (230 g); **shipping,** 11 oz (315 g).

Catalog Number	*Description*	*Price in USA*
0900-9540	**900-M Component Mount**	**$70.00**

GR Product Notes

COMPLETE PROGRAMMABILITY FOR GR SYNTHESIZERS

With the introduction of the TYPE 1164-RDI-3 10-MHz/step Programmable Digit-Insertion Unit, GR's frequency-synthesizer line is completely programmable, from lowest frequency to highest (70 MHz). The 1164-RDI-3 is a plug-in replacement for the highest-frequency (X-10 MHz) module in a TYPE 1164 synthesizer. Conversion from nonprogrammable to programmable is just a few minutes' work. Of course, any synthesizer can be ordered equipped with programmable decades.

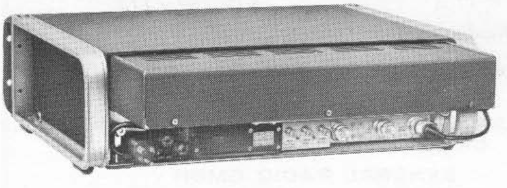
STANDARD-FREQUENCY OSCILLATOR FOR USE WITH SYNTHESIZERS

For many applications, the crystal oscillator built into a GR frequency synthesizer provides adequate stability. The way to higher stability, for those who need it, is a phase lock with an external frequency standard such as GR's TYPE 1115. Now, for those applications requiring less than the ultimate in stability but more than the synthesizer alone can provide, an inexpensive solution is available — an accessory standard-frequency oscillator that mounts inconspicuously on the rear of the synthesizer without adding to over-all height.

The oscillator (TYPE 1160-P3) uses a 5-MHz crystal in a temperature-controlled oven, a buffer amplifier, and its own power supply (so that it can be operated from the power line independently of the synthesizer).

The oscillator holds frequency within 1 part in 10^8 of its room-temperature frequency for any ambient temperature between 0 and 50°C. Aging is less than 3 parts in 10^9 per day. This performance puts the 1160-P3 significantly ahead of a "barefoot" synthesizer but still behind the synthesizer phase-locked to a frequency standard such as the GR TYPE 1115.

Accessory standard-frequency oscillator mounts neatly on rear of synthesizer, requires no additional relay-rack height. Separate power connector allows oscillator to run independently of synthesizer.



Catalog Number	*Description*	*Price in USA*
1164-9489	**1164-RDI-3 Programmable Digit-Insertion Unit,** 10 MHz/step	**$575.00**
1160-9650	**Hookup cable,** 50-ft, 12-conductor, shielded, for above unit.	**15.00**
1160-9603	**1160-P3 Standard-Frequency Oscillator**	**525.00**

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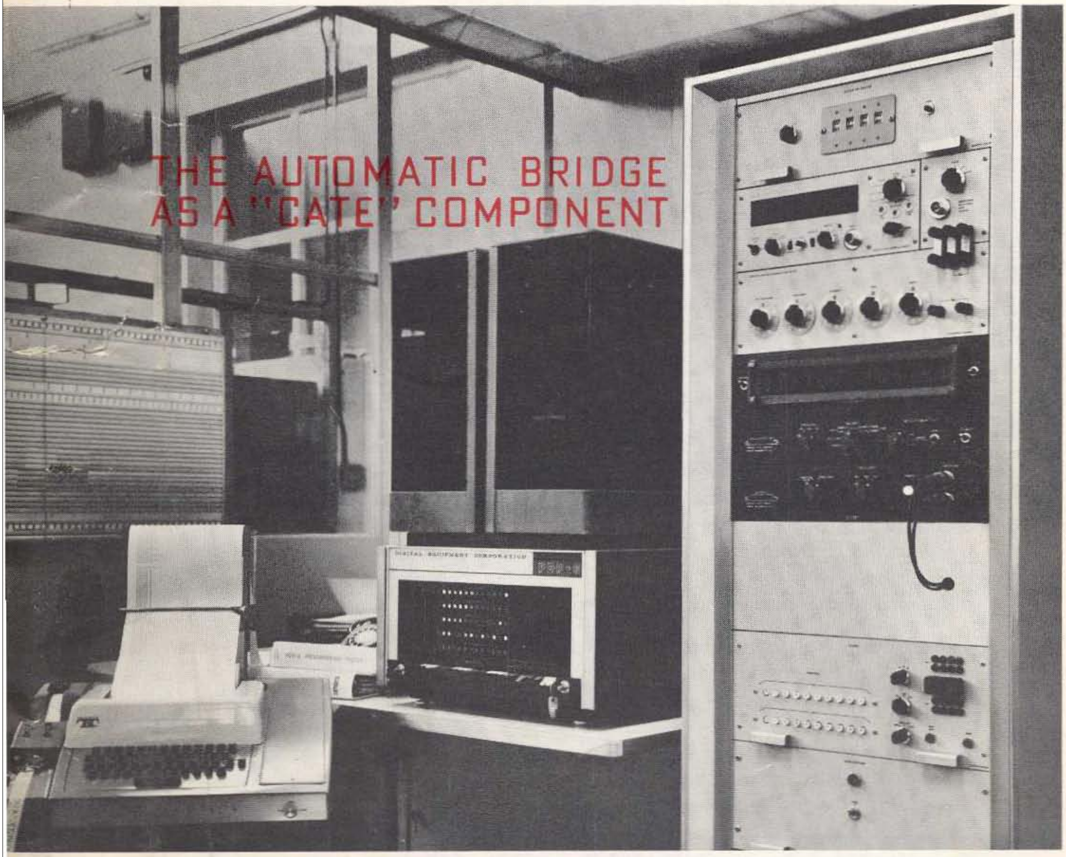


ALSO IN THIS ISSUE

- A RECEIVER FOR PRECISE TIME CALIBRATIONS
- A PARALLEL-STORAGE UNIT FOR THE SYNCRONOMETER
- IMPROVED PERFORMANCE FROM THE 1115 FREQUENCY STANDARD

VOLUME 42 · NUMBER 3 / MARCH 1968

the GR Experimenter

Volume 42 • No. 3 March 1968

Published monthly by the General Radio Company

CONTENTS

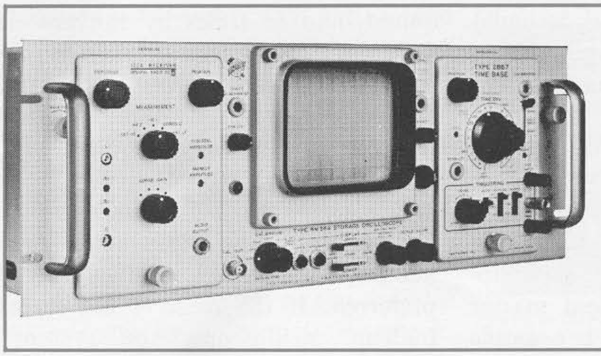
ABOUT THIS ISSUE

General Radio owes its name to the glory that was "radio" back in 1915, when the Company was founded. Now, with "radio" a household synonym for "radio receiver," we sometimes find ourselves explaining to the public that despite our name we do not make radio receivers. Except for the one pictured on the opposite page, that is. Designed for use with the GR Syncronometer® digital time comparator, it covers WWV, CHU, and Loran-C frequencies and includes an oscilloscope for visual comparison of off-the-air time signals with the master tick of the Syncronometer. (It is, in other words, a special-purpose, and not a general, radio.)

Two other important companions to the Syncronometer—a parallel-storage unit and an improved frequency standard—are also introduced in this issue. The mission of the former is to store time information coming rapidly from the Syncronometer until slower data-handling equipment can accept it. The frequency standard is, of course, the key to the accuracy of a local time standard, and the improved crystal-oscillator performance announced in this issue is of obvious importance.

Cover: The Plessey Company's "CATE" (Computer-controlled and Automated Test Equipment) System for component testing includes a now-familiar combination: GR's automatic capacitance bridge and a digital computer. (See page 9.) (Photo courtesy The Plessey Company, Ltd.)

The *General Radio Experimenter* is mailed each month without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.



The Type 1124 Receiver.

A RECEIVER FOR PRECISE TIME CALIBRATIONS

A common method of establishing accurate local time is to drive a local clock with a frequency standard, checking the absolute accuracy regularly against standard time signals broadcast by various agencies throughout the world. An elegant variation of this technique involves the use of primary time standards flown to the local site for periodic calibrations. The "flying-clock" method, though obviously expensive, offers the greater accuracy, since it eliminates the uncertainties of radio transmissions. However, once the clock flies away, there exists the problem of maintaining time until it flies in again. For this purpose, as well as for the many applications that require accurate timekeeping but that do not justify flying clocks, the use of broadcast time signals is the answer.

A time-comparison system using broadcast signals typically takes the form shown in Figure 1. General Radio has for years made two of the components used in such a system: standard-frequency oscillators and Syncronometer® digital time comparators (clocks). We now offer the rest of the system, a receiver designed specifically for time standardization with the GR Syncronometer, with a built-in oscilloscope for automatic visual comparison of the received signals against those from the local clock.

The General Radio time standard included in Figure 1 operates as follows: The 100-kHz output from the 1115-C Standard-Frequency Oscillator is fed to the Syncronometer, which translates the zero crossings into a pulse train and thence into a one-pulse-per-second master tick. These ticks are accumulated in six digital counting circuits, and the totals are presented as digital time-of-day infor-

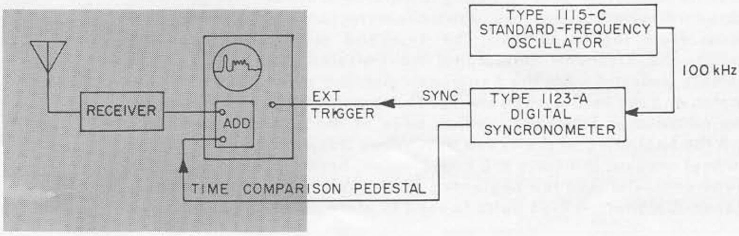


Figure 1. The components of a time-calibration system. Functions in shaded area are performed by the new 1124 receiver. (Antenna is included for Loran-C.)

mation (hours, minutes, and seconds) on the Syncronometer front panel.

To calibrate the system in terms of off-the-air time signals, one need only compare the one-second master tick with a one-second time signal broadcast by one of several agencies offering such services. Then, with both local and broadcast time signals on the oscilloscope, synchronizing the local master tick with the received signal is a simple matter of adjusting a few thumbwheels and pushing a button on the Syncronometer.

This procedure, which sounds simple in the telling, is complicated by the vagaries of radio propagation, by the high degree of precision usually required, and at times by the practical difficulties of integrating and interconnecting several instruments into an efficient system. A stable, sensitive receiver is obviously required. It is unreasonable to expect a general-coverage receiver to have optimum performance at a few selected frequencies, and a special-purpose receiver is usually preferred. If the receiver has its own built-in oscilloscope, the system is greatly simplified.

The new GR receiver (Type 1124) is specially designed for dependable, consistent reception and faithful display of time signals on six fixed frequencies. These are the 2.5-, 5.0-, and 10.0-MHz frequencies of WWV, two CHU fre-

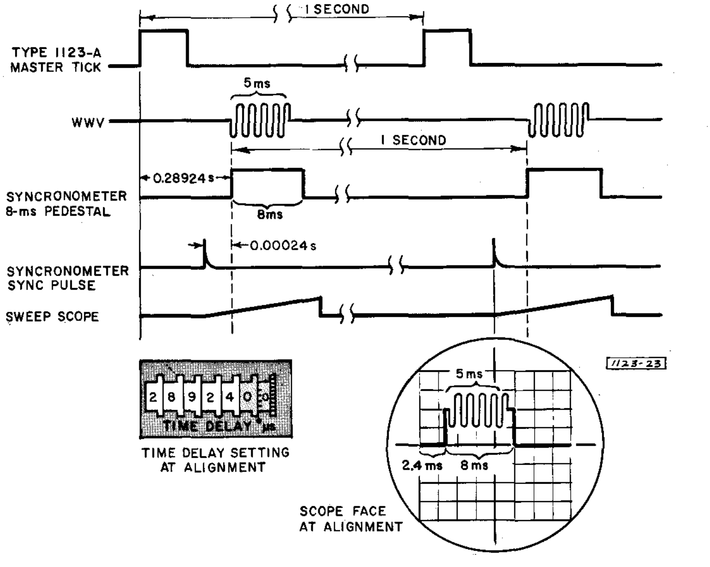
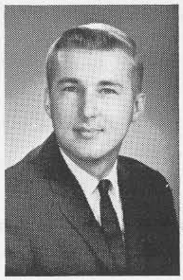


Figure 2. Syncronometrics: Above timing diagrams illustrate the principle of time standardization using the Syncronometer and off-the-air time transmissions. From top to bottom: The 1-second master tick from the Syncronometer; the 1-second time-signal transmissions from WWV; the 8-ms adjustable pedestal from the Syncronometer, the sync pulse from the Syncronometer, and the oscilloscope sweep. The thumbwheels on the Syncronometer are adjusted to bring the leading edge of the pedestal into coincidence with the beginning of the WWV tick. When this adjustment is made, the thumbwheel readout indicates the time interval between the master tick from the Syncronometer and the beginning of the WWV tick. With higher-precision Loran-C signals, a 1-μs pulse is used in place of the 8-ms pedestal.

quencies (3.33 and 7.335 MHz), and 100 kHz for Loran-C transmissions. An "external" mode permits use of signals from other sources (such as, for instance, flying clocks).

The five high-frequency circuits are all fixed-tuned with crystal-controlled local oscillators, and all are mounted on plug-in etched boards. The two used most frequently are left in the receiver and selected by a front-panel switch. Usable input sensitivity for the high frequencies is greater than 3 μV, and agc circuits hold the receiver output within 3 dB over an input-signal range of 10 μV to 100 mV. A 3-MHz crystal i-f filter and three tuned rf stages provide image and i-f rejection of more than 80 dB, with all other spurious responses at least 70 dB down. Low distortion of the modulating waveform, even with 90% modulation, ensures faithful pulse reproduction. There is an audio output monitor, isolated from the display to prevent loading.

The Loran-C receiver is a fixed-tuned amplifier with 100-kHz center frequency and a bandwidth of about 20 kHz (needed to preserve the Loran pulse waveshape). Its 50-ohm input impedance matches the impedance of the loop antenna supplied. Input sensitivity is 3 μV for a signal-to-noise ratio of 2 or greater. An important feature is a pair of notch filters for rejection of unwanted signals near 100 kHz. These filters tune from 80 to 95 kHz and from 105 to 125 kHz and have greater than 40-dB rejection. A gain control with a 60-dB range supplements the oscilloscope gain control for Loran-C presentations.



Dale O. Fisher joined General Radio in 1964, after receiving his BSEE degree from Northeastern University. A development engineer in GR's Frequency and Time Measurement Group, Mr. Fisher has specialized in the design of digital time comparators and related equipment. He is now completing work toward his MS degree from Massachusetts Institute of Technology.

Visual display is by means of a built-in Tektronix RM564 Storage Oscilloscope. The storage mode is especially useful in this application, since it will average out time variations due to unstable propagation characteristics and will increase the signal-to-noise ratio, since the random noise is stored less frequently than is the desired signal. The Tektronix Type 2B67 Time Base provides up to 1-μs/cm display for accurate Loran-C comparisons and allows single-pulse triggering for photographic records.

It is expected that the new receiver will be widely used for its Loran-C capability. Those interested in learning more about the use of Loran-C in precision frequency and time measurements are invited to ask for our recently published monograph on the subject. The 12-page booklet, No. 2 in GR's *Frequency/Time Notebook* series, is available free on request.

— D. O. Fisher

SPECIFICATIONS

HIGH-FREQUENCY RECEIVERS

Rf Frequencies: 2.5, 3.33, 5.0, 7.335, and 10 MHz. Any two are selected by a front-panel switch.

Sensitivity: Better than 3 μV.

Input Impedance: Approx 50 Ω.

Max Input Signal: $>$100 mV.

Bandwidth: I-f 3-dB bandwidth approx 3 kHz; 3.0 MHz center frequency of i-f amplifier and crystal filter.

Automatic Gain Control: Receiver output is within 3 dB for signal change of 10 μV to 100 mV.

Image and I-F Rejection: >80 dB; all other spurious responses at least 70 dB down.

LORAN-C RECEIVER

Center Frequency: 100 kHz; 3-dB bandwidth approx 20 kHz.

Sensitivity: 3 μV for S/N>2.

Input Impedance: Approx 50 Ω.

Max Input Signal: >100 mV.

Gain Control: 4 fixed steps, 60-dB total range.

Notch Filters: Two, front-panel screwdriver-control, 80 to 95 kHz and 105 to 125 kHz (other ranges with internal-capacitor change). Rejection >40 dB; 6-dB bandwidth <3 kHz.

EXTERNAL INPUT Intended for comparing other timing signals with the GR 1123 comparator.

Sensitivity: Approx 0.5 V for full-screen deflection.

GENERAL

Front-Panel Controls: Amplitude (20-dB range), vertical position, input-channel selector, gain; screwdriver controls: notch-filter tuning (2), 1123 pedestal amplitude, and 1123 marker amplitude.

Connections: Front panel: audio output, approx 1 V, for monitoring hf receiver. Rear panel (BNC connectors): Loran antenna, hf antenna, ext-signal input, and pedestal, sync, and marker pulses from 1123.

Power Required: 105 to 125 or 210 to 250 V, 50 to 60 Hz, 240 W.

Accessories Supplied: Storage-oscilloscope accessories, shielded-cable set, 1124-P1 Antenna.

Mounting: 19-inch rack-mount.

Dimensions (width x height x depth): 19 x 7 x 18½ in. (485 x 180 x 470 mm).

Weight: Net, 42 lb (19.5 kg); shipping, c 70 lb (32 kg).

1124-P1 Antenna

Center Frequency: 100 kHz.

Bandwidth: Approx 20 kHz at 3-dB points, with 50-Ω load.

Dimensions (width x height x depth); 58 x 86 x 3¾ in. (1480 x 2200 x 96 mm).

Catalog Number	*Description*	*Price in USA*
1124-9701	**1124 Receiver**	**$3250.00**

GR1125

PARALLEL-STORAGE UNIT FOR THE SYNCRONOMETER

In the *Experimenter* article introducing the Type 1123 Syncronometer® digital time comparator,[1] we said, "No commercial equipment presently available can accept time readings as fast as the comparator can supply them. Required is a parallel-entry storage register with a capacity of 11 four-bit binary words. The register must accept and store the data from the clock in a time well under 5 microseconds." We can now drop the other shoe, by announcing the availability of the Type 1125 Parallel-Storage Unit.

The Syncronometer, it may be recalled, is essentially a precise accumulator of time in 10-microsecond increments. Feeding such fast-changing data to auxiliary data-handling equipment presents an obvious problem: Most such equipment (printers, tape punches, etc)

[1] D. O. Fisher, R. W. Frank, "A New Approach to Precision Time Measurements," *General Radio Experimenter*, February-March 1965.

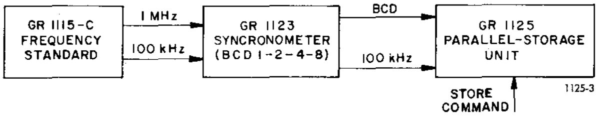


Figure 1. Block diagram of local time standard with parallel-storage unit.

can't keep up with the Syncronometer, and the Syncronometer can't be interrupted for interrogation. Enter the parallel-storage unit, which accepts the time-of-day information — to 10-μs resolution — from the Syncronometer in 2 microseconds. On command from an external source, the unit displays the data on an in-line digital readout and simultaneously presents it in 1-2-4-8 BCD form (thus only the 1-2-4-8 version of the Syncronometer can be used here). Inhibit circuits, controlled either internally or by an external device, are incorporated to prevent storage while clock data are changing or while the storage-unit's output is being used.

The 1125 Parallel-Storage Unit is an all-solid-state instrument containing 11 four-bit storage registers, 11 indicator circuits, and command and inhibit program circuits. In normal operation, the unit receives data input and a 100-kHz inhibit signal from the Syncronometer, by way of cables supplied with the storage unit. These connections, as well as the command connection, the data output connection, and other connections provided for specialized systems applications, are made at the rear panel, leaving only the power switch, indicator brightness control, and indicators on the front of the instrument.

Systems Using the 1125

The simplest system using the storage unit includes an 1123 Syncronometer digital time comparator (driven by a frequency standard such as GR's 1115-C) and the 1125. This system, shown in Figure 1, stores time to a resolution of 10 μs each time a new "store" command is received at its input. Storage of data is automatically timed by means of the 100-kHz inhibit signal to avoid transitions of the counting registers in the Syncronometer. Stored data are displayed on the storage unit's 11 indicators, changing only when new data are stored.

Better use of the system can be made with the addition of a data printer such as GR's TYPE 1137 (Figure 2), which makes a permanent record of time information immediately after it is stored. The only connection required is made by a single cable between the 1125 and the 1137. In this application, the storage unit's internally generated print command and inhibit signals would be used to control the printer and to prevent storage of new data while the printer is operating. If the three-line-

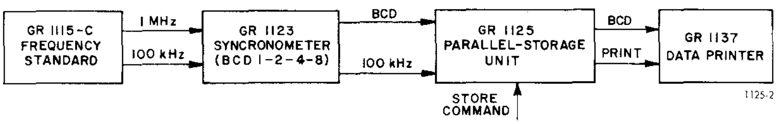


Figure 2. Addition of data printer greatly increases usefulness of system in many applications.

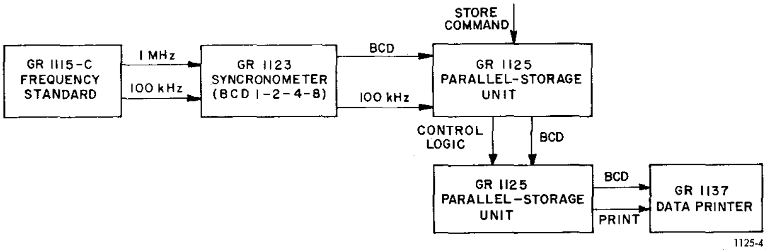


Figure 3. Two or more parallel-storage units can be cascaded to store time data arriving too fast for the printer.

per-second print rate of the standard 1137 printer is not fast enough, models with speeds up to 20 lines per second are available commercially.

If the events to be timed occur too closely together to be recorded by even a high-speed printer, another system is called for, in which two or more storage units are cascaded (Figure 3) to hold the time information until it can be transferred to the printer. The first time data are stored in unit 1, then immediately transferred to unit 2. The first unit will then be ready to accept new data in only 20 μs, rather than having to wait for a 5-ms or longer print cycle.

Numerous variations of serial or parallel combinations of storage units are possible, and inquiries for special systems are invited.

— D. O. FISHER

SPECIFICATIONS

TRANSFER CHARACTERISTICS

Capacity: 11 decimal digits (44 bits) parallel-entry jam-transfer; 10-μs resolution.

Transfer Time: 2 μs approx, for up to 8-ft data cables.

Mode: Data are stored until next store command.

INPUT

Data: 4-line BCD, 1-2-4-8; fully compatible with output of 1-2-4-8 versions of GR 1123 time comparator.

Store Command: Positive or negative transition (switch-selected) between 0 and at least +5 V. Input impedance >100 kΩ, dc coupled.

Inhibit Signal (from 1123): Inhibits transfer while input data are changing.

Inhibit Signal (internal or external): Inhibits transfer while stored data are read by output equipment. Internal inhibit is equal in duration to print command (see below). External inhibit signal can be presence of either 0 or at least +5 V (switch-selected); input impedance >100 kΩ, dc coupled.

OUTPUT

Data: 4-line BCD, 1-2-4-8 with GR 1123.

Logic Levels:
"1" – +10 V
"0" – 0 V

Output Impedance: Approx 6.8 kΩ at logic "1," approx 12 kΩ at logic "0."

Print Command: Pulse of +15 V behind 5 kΩ, duration adjustable 20 ms to 0.5 s; set initially for use with GR 1137 Data Printer.

Data-Ready Signal: Pulse of +15 V behind 5 kΩ, 50-μs duration, occurs 10 μs after transfer is completed.

Readout: 11 in-line digital indicators. Indicates 10's of hours through 10 μs.

GENERAL

Power Required: 90 to 130 or 180 to 260 V, 50 to 60 Hz, 25 W.

Accessories Supplied: Data and inhibit signal cables for connection to GR 1123, output data connector, power cord, spare fuses, mounting hardware with rack models.

Accessories Available: GR 1137 Data Printer and other digital-data acquisition instruments.

Mounting: Rack-bench cabinet.

Dimensions (width x height x depth): Bench, 19 x 3⅞ x 17 in. (485 x 99 x 435 mm); rack, 19 x 3½ x 16¾ in. (485 x 89 x 425 mm).

Weight: Net, 23½ lb (11 kg); shipping, 45 lb (20 kg).

Catalog Number	*Description*	*Price in USA*
	1125 Parallel-Storage Unit	
1125-9801	Bench Model	**$2600.00**
1125-9811	Rack Model	**2600.00**

HIGHER PERFORMANCE FOR THE 1115 FREQUENCY STANDARD

The time/frequency systems described in the preceding articles, like all other time and frequency measuring systems, are only as good as the local frequency standards driving them. The availability of a new, fast-stabilizing crystal has enabled General Radio again to upgrade the performance of its 1115 piezoelectric frequency standard.[1] Aging specifications are now less than 5×10^{-10} per day after three days (formerly 30 days) of operation, less than 1×10^{-10} per day, typically, after six months (formerly a year).

Like its predecessor, the new standard has outputs at 5 and 1 MHz and 100 kHz, excellent spectral purity, and a built-in nickel cadmium battery that automatically takes over for up to 35 hours in the event of power-line failure.

[1] H. P. Stratemeyer, "TYPE 1115-B Standard-Frequency Oscillator," *General Radio Experimenter*, June 1964.

Catalog Number	*Description*	*Price in USA*
1115-9803	**1115-C Standard-Frequency Oscillator** (Bench)	**$1800.00**
1115-9813	**1115-C Standard-Frequency Oscillator** (Rack)	**1800.00**

THE AUTOMATIC BRIDGE AS A CATE COMPONENT

GR's TYPE 1680 Automatic Capacitance Bridge and a digital computer are natural companions, frequently used in combination to perform measurement tasks that would have been unthinkable a few years ago. As an impressive example, take the CATE (Computer-controlled and Automated Test Equipment) system used by The Plessey Co.'s Product Assessment Laboratories at Titchfield, England.

The complete system, shown on the cover, is designed primarily to support reliability trials on thin-film components, resistive and capacitive. The 1680 is used to measure the capacitors. These are aluminum-silicon monoxide elements deposited on a glass substrate, two elements per substrate. Capacitance values range from 100 to 10,000 pF.

Under a long-term-reliability trial, 2000 of these elements are stored at an ambient temperature of 70°C, with maximum rated dc voltage applied. Periodically during the reliability trials the components are removed from the test environment for measurement under standard conditions at 20°C. Capacitors are connected to CATE in groups of 100 at a time, via multiway miniature coaxial connectors, which are in turn linked to a 100-channel scanner at the input to the 1680.

The computer, a Digital Equipment PDP-8, performs several functions: It controls the over-all sequence of operations, it commands the scanner to select the next channel, it recognizes the end of a measurement and transfers measured data to core storage, it calls up subroutines to process input data, and it feeds data to a typewriter and tape punch.

The BCD output from the 1680 is adapted to the PDP-8 by a special interface. The BCD digits are transferred to core in 12-bit words, each word containing three digits. The transfer of all display and range data requires 70 control instructions from the computer. Stored BCD digits can be decoded and converted to equivalent binary words, the capacitance and conductance values being assembled into double-precision binary numbers. The time required for transfer and conversion is less than 130 microseconds.

The simplest program transfers data directly from the 1680 to the teletypewriter, producing a printed output and an equivalent punched tape. CATE can easily handle more complex functions. It can, for instance, check values against preset limits and indicate out-of-tolerance results by, say, a typewritten asterisk. Or it can accumulate sequential measured values and calculate mean and standard deviations of parameter distributions. Or it can compare measured data with earlier data stored on paper tape, calculating percentage change. Since CATE also includes a temperature-measuring digital voltmeter, temperature coefficients can be calculated from changes in capacitance and temperature.

Acknowledgment: We are indebted to Mr. Brian A. Mair of The Plessey Company Ltd. for the information presented above.

MAKING THE 1602 AND 1607 BRIDGES DIRECT-READING BELOW 40 MHz

The susceptance standard supplied with the TYPE 1602-B Immittance Bridge and with the TYPE 1607-A Transfer-Function and Immittance Bridge is calibrated down to 40 MHz; below this frequency a correction is required if the bridge is to be direct-reading.

Either bridge can be made direct-reading below 40 MHz by the simple addition of a tee and a variable capacitor between the susceptance standard and the bridge. The 874-VCL Variable Capacitor (14-70 pF) and the 874-TL Tee are ideal for the purpose; an 874-ML Component Mount fitted with a low-loss variable capacitor or a fixed silver-mica capacitor, plus an 874-TL Tee, can also be used. The extra capacitance needed can also be supplied by an 874-L Air Line inserted between the susceptance standard and the bridge.

After these components are connected, capacitance is adjusted in accordance with the instruction book for the 1602-B (section 3.1) or the 1607-A (sections 3.1.3.2 or 3.1.8.3).

ADDENDUM

In "Precision Capacitance Measurements with a Slotted Line" (*Experimenter*, September 1967), the equation for C_x given on page 11 may not be entirely suitable at the higher frequencies. The following, more exact equation is now recommended:

$$C_x = \frac{41.072 \times 10^8}{f\sqrt{L(\omega)}\tan[1.61799 \cdot 10^{-8} fl\sqrt{L(\omega)}]}$$

where f is frequency in hertz
l is slotted-line position in meters
$L(\omega)$ is slotted-line inductance per unit length, in nH/cm
and the argument of the tangent is in radians.

GR Product Notes

KEEPING COAXIAL CONNECTORS CLEAN

In an increasingly polluted environment, it seems that cleanliness is next to impossible. It is also next to essential in precision microwave measurements, where a little dirt and grime on a connector contact surface can introduce significant error. With a connector as carefully designed and made as the GR900, anti-dirt warfare is especially important. The best weapon is a TYPE 900-TOC Connector Cleaning Kit, which includes a 16-ounce spray can of Freon TF solvent, two nylon brushes, a pipette, cloth patches, and an Allen wrench.

ROTATABLE GR900® CONNECTIONS

The gear teeth on a GR900® precision coaxial connector can occasionally present a problem in alignment. If, for example, GR900 connectors on two large pieces of equipment are to be mated, the teeth on the two connectors may be oriented so as to prevent connection.

Where this problem exists, one answer is the new 900-PKMR Panel Mounting Kit, which adapts standard GR900 connectors for panel mounting and which includes a rotatable gear ring. Another is a rotatable centering ring (Catalog No. 0900-9499) that replaces the ring on a standard GR900 connector.

CHOKE FOR VARIAC® AUTOTRANSFORMERS

The W50-P1 Choke is designed to eliminate circulating currents in parallel-connected Variac® autotransformers. The new choke can carry up to 60 amperes of load current and thus can handle a parallel pair of any size Variac. Additional chokes can be used for parallel connections of more than two autotransformers.

Catalog Number	*Description*	*Price in USA*
0900-9610	**900-TOC Cleaning Kit**	$ 7.50
0900-9500	**900-PKMR Panel Mounting Kit,** Rotatable	20.00
0900-9499	**Centering Ring,** Rotatable	15.00
3150-5016	**W50-P1 Choke**	16.00

Experimenter Index

An index to the *Experimenter* for the year 1967 is now available on request. Write to Editor, *General Radio Experimenter*, General Radio Company, West Concord, Mass. 01781.

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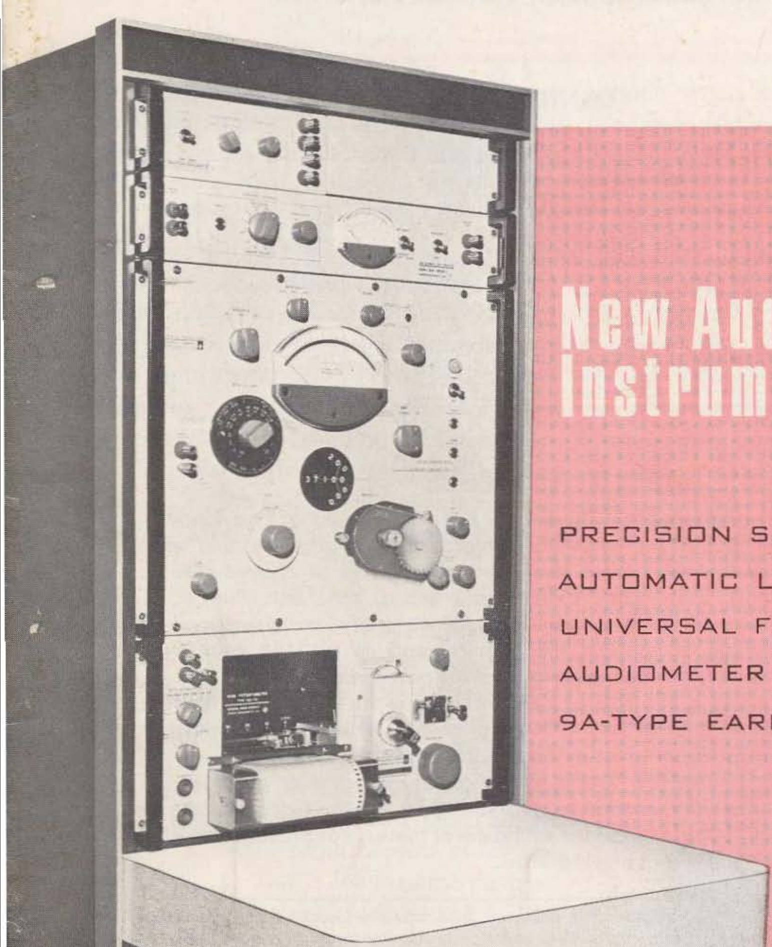


New Audio Instrumentation

PRECISION SOUND-LEVEL METER

AUTOMATIC LEVEL REGULATOR

UNIVERSAL FILTER

AUDIOMETER CALIBRATION SET

9A-TYPE EARPHONE COUPLER

VOLUME 42 · NUMBER 4 / APRIL 1968

the Experimenter

Volume 42 • No. 4 April 1968

Published monthly by the General Radio Company

CONTENTS

ABOUT THIS ISSUE

In the business of measurement, there are, broadly speaking, two kinds of advances: those that extend the accuracy or measurement range beyond what was previously available and those that give new speed or convenience to existing measurements. Both types are represented in this issue. The 1561 sound-level meter (page 3) puts a new order of accuracy and precision into acoustical measurements, exceeding the requirements of all USA and international standards for both precision and general-purpose sound-level meters. Instruments to make life in the lab easier are a high-pass/low-pass tunable filter (page 14) and a regulator that will control an audio test signal to produce a constant sound or vibration level as frequency is swept.

Two new items are directed to the attention of audiometer users: a GR earphone coupler (page 21) similar to the NBS type 9-A and an audiometer calibration set (page 20) that includes sound-level meter, calibrator, and earphone coupler, all in one tidy carrying case.

COVER—One of many systems possibilities using GR's new level regulator (second from top in relay rack). Here the regulator is combined with random-noise generator and recording wave analyzer to produce a narrow band of constant-level white noise.

The *General Radio Experimenter* is mailed each month without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.

A PRECISION SOUND-LEVEL METER

Recognizing that it is unrealistic to expect one sound-level meter to be all things to all people, we have for the past two years offered two models: the 1551-C[1] and the compact, less-expensive 1565[2], both of which meet the USA standard (S1.4-1961) for general-purpose sound-level meters. The variety of demands on these instruments and the trend toward more precise measurements in some research work suggested a need for a third member of the family, with even higher performance. Thus the new TYPE 1561 Precision Sound-Level Meter joins General Radio's acoustical lineup.

The new sound-level meter meets the requirements of the standards for a precision sound-level meter as adopted by the International Electrotechnical Commission (IEC 179, 1965) and as proposed, but not yet adopted, by the United States of America Standards Institute (USASI). The instrument also satisfies the requirements of USASI and IEC standards for general-purpose sound-level meters, USASI S1.4-1961 and IEC 123, 1961.

The chief contributor to the new instrument's precision performance is a new piezoelectric ceramic microphone (TYPE 1560-P7), a descendant of the 1560-P5,[3] which benefits from a series of design improvements and from improved manufacturing techniques and careful testing. Backing up the microphone is an all-solid-state instrument designed throughout for precision work.

Description

The circuit includes a high-gain transistor amplifier with attenuator, A-, B-, and C-weighting networks (plus flat response), an indicating meter, and an electrical calibration system. A low-noise field-effect transistor in the first amplifier stage provides high input impedance and low input noise. The

[1] E. E. Gross, "TYPE 1551-C Sound-Level Meter," *General Radio Experimenter*, August 1961.
[2] W. R. Kundert, "A Compact, Inexpensive Sound-Level Meter," *General Radio Experimenter*, October-November 1964.
[3] B. A. Bonk, "The New General Radio Microphone," *General Radio Experimenter*, May–June 1967.

Figure 1. Precision sound-level meter, portable model.

Ervin E. Gross, Jr. joined General Radio in 1936, after receiving his BSEE degree from Northeastern University. As a development engineer in GR's Audio Group, he has designed many sound- and vibration-measuring instruments. He is a Senior Member of IEEE and a member of the Standards Committee of the IEEE Group on Audio and Electroacoustics.

Carlton A. Woodward, a development engineer in GR's Audio Group, received his B.Sc. degree from Northeastern University in 1935 and joined General Radio the following year. His work includes a number of instrument developments, at low and high frequencies. He is a Senior Member of IEEE.

meter circuit uses the rms response that is standard in GR acoustic instruments. Phone jacks allow insertion of an external filter in place of the standard weighting characteristics. The internal electrical calibration system allows the user to set the gain to match the sensitivity of the microphone and cable combination in use.

The 1561 comes in two versions: a portable model (1561) in a General Radio Flip-Tilt case (Figure 1) and a relay-rack model (1561-R). The expected applications of the two models lead to some differences in connections and accessories, as indicated below.

Portable Model

The portable model is equipped with a 1560-P7 Precision Microphone and a 10-foot microphone extension cable. All the connectors and all controls except two are on the front panel. A gain control and the calibrated microphone-sensitivity reference are accessible through openings in the case. Power is supplied either by standard dry cells or by rechargeable nickel-cadmium batteries. The dry cells will power the instrument for 8 to 16 hours of continuous use or about twice as long at four hours per day. The nickel-cadmium batteries will yield 14 to 22 hours of operation between charges.

When an instrument is purchased with rechargeable batteries, a battery charger is also supplied. This charger will charge one set of batteries, either in the sound-level meter or in the charger, or it will charge two sets simultaneously. Charging time is 14 to 16 hours.

Relay-Rack Model

The relay-rack instrument, TYPE 1561-R (Figure 2), is supplied without a microphone or microphone extension cable; both are, of course, available

Figure 2. Type 1561-R (relay-rack version) precision sound-level meter.

separately. All controls except one gain adjustment are on the front panel. Input and output connections may be made at the rear panel as well as at the front. The jacks for the external filter are at the rear only. Coaxial attenuator controls are designed to prevent overload of the input amplifier stage under all conditions.

The relay-rack instrument is ac-powered, but rechargeable batteries can be installed if desired.

Uses

The primary use for the precision sound-level meter is the measurement of noise to the degree of accuracy specified by precision-measurement standards. This instrument is in fact useful in any application involving current American or international standards for sound-level meters. In this connection it is interesting to note that one of the "general" standards, USASI S1.4-1961, actually has tighter over-all tolerance requirements for the C-weighting characteristic below 1000 Hz than has the precision-sound-level-meter standard, IEC 179, 1965.

The precision sound-level meter is also useful as a low-noise, high-gain amplifier, and the relay-rack model is particularly suitable for such use as a system component.

— C. A. Woodward
— E. E. Gross

SPECIFICATIONS

Sound-Level Range (rms, dB re 20 μN/m^2):

Frequency Characteristic	*With 1560-P7 Microphone and 10-ft cable*	*With 1560-P7 Microphone and 1560-P40 Preamplifier**
Flat	35 to 150 dB	31 to 130 dB
C Weighting	32 to 150 dB	27 to 130 dB
B Weighting	31 to 150 dB	26 to 130 dB
A Weighting	31 to 150 dB	27 to 130 dB

*Min obtained with ×10 preamp gain, max with ×1.

Allowance is made for a peak-to-rms ratio of 14 dB. When a sine-wave signal is applied to the 1560-P40, the range can be extended to 141 dB. The signal-to-noise ratio is at least 5 dB for the lower values given above.

Frequency Characteristics: A, B, and C weighting in accordance with USA Standard S1.4-1961, IEC Publication 123, 1961 and IEC Publication 179, 1965 for precision sound-level meters. Also provided is a flat response from 20 Hz to 20 kHz to permit measurement of sound-pressure level. Jacks are provided for insertion of an external filter.

Microphones: The GR 1560-P7 Precision Microphone is supplied with portable models with a 10-ft cable to permit microphone to be located away from instrument and observer to minimize diffraction effects (1561 gain is set to compensate for cable loss).

Sound-Level Indication: Reading is sum of meter and attenuator setting. Meter calibrated -6 to $+10$ dB; attenuator calibrated 30 to 140 dB in 10-dB steps.

Output (full-scale meter reading): 1.25 V behind 5500 Ω; harmonic distortion $< 0.5\%$.

Input Impedance: > 100 MΩ, across 40 pF in portable model, across 90 pF in rack model.

Meter: Rms response; fast and slow meter speeds in accordance with above USASI and IEC standards.

Calibration: Absolute calibration of the 1561 is set acoustically at 500 Hz and a level of 114 dB re 20 μN/m^2. Microphone response and sensitivity are measured in a free field 20 Hz to 15 kHz by comparison with a WE 640AA Laboratory Standard microphone with calibration traceable to the National Bureau of Standards. Complete electrical frequency-response measurements are made on each instrument. Panel adjustment provided for standardizing gain with internal calibration circuit, which has adjustment to permit calibration in terms of microphone sensitivity (control is internal and accessible through case of portable models, on front panel of rack models). The 1562 Sound-Level Calibrator or 1559 Microphone Reciprocity Calibrator can be used for making periodic over-all acoustic checks.

Temperature and Humidity Effects: The instrument will operate within specifications, for meter indications above 0 dB, over a range of 10 to 50°C and 0 to 90% relative humidity, when standardized by its internal calibration circuit or an external calibrator. No damage to microphone from -30 to $+60$°C and 0 to 100% relative humidity.

Magnetic-Field Effects: In a 60-Hz, 1-oersted (80 A/m) magnetic field and oriented for max reading, the rack model will indicate about 42 dB, the portable model about 53 dB (C weighting).

Accessories Supplied: Portable models include Precision Microphone Type 1560-P7, 10-ft microphone cable, and either one set of dry-cell batteries or two sets of rechargeable batteries and Battery Charger Type 1560-P60. Rack model includes power cord and spare fuses.

Accessories Available: 1952 Universal Filter and 1560-P40 Preamplifier (power supplied by 1561); 1560-P5 or P7 Precision Microphone; microphone extension cables.

Power Required: The rack-mount 1561-R contains ac power supplies for operating the instrument and for recharging the batteries (not supplied) that can be used to power the instrument. This model operates from 100 to 125 or 200 to 250 V, 50 to 60 Hz, 2.5 W max.

The portable 1561 is supplied with either 3 Burgess type PM6 dry-cell batteries (or equivalent), which give about 15-h average operation, or with 2 sets of rechargeable nickel-cadmium batteries and the 1560-P60 Battery Charger. This unit will simultaneously recharge two sets of batteries (one set in the 1561, the other in the charger) from a power line of 105 to 125 or 210 to 250 V, 50 to 60 Hz, 5 W.

The nickel-cadmium batteries will provide about 20 h of operation and recharge in about 15 h; dry-cells about 15 h.

Mounting: The 1561-R is in a rack-mount cabinet, the portable model in a Flip-Tilt case; the charger in an aluminum case.

Dimensions (width x height x depth): Portable, 10¾ x 6⅛ x 5¾ in. (275 x 160 x 150 mm); rack, 19 x 3½ x 15 in. (485 x 89 x 385 mm); Battery Charger, 4¼ x 3¾ x 8 in. (110 x 96 x 205 mm).

Net Weight: Portable, 5½ lb (2.5 kg); rack, 15 lb (7.0 kg).

Shipping Weight (est): Portable, 20 lb (4.6 kg); rack, 23 lb (10.5 kg).

Catalog Number	*Description*	*Price in USA*
	1561 Precision Sound-Level Meter Portable Models, incl precision microphone and 10-ft cable	
1561-9700	with dry-cell batteries	**$675.00**
	with 2 sets rechargeable batteries and recharger	
1561-9701	for 115 volts	775.00
1561-9702	for 230 volts	on request
1561-9703	**1561-R Precision Sound-Level Meter** Rack Model (no battery or microphone)	725.00
1560-9605	**1560-P5 Microphone**	80.00
1560-9607	**1560-P7 Precision Microphone.** Sold only with 1561-R	145.00
1560-9673	**1560-P73 Extension Cable** (25 ft)	15.00
1560-9977	**1560-P72B Extension Cable** (100 ft)	29.00
8410-3000	**Replacement Dry Cell**, 3 req'd	1.20
8410-1040	**Rechargeable Battery**, 2 req'd	12.00

SOUND-LEVEL METERS, GENERAL-PURPOSE AND PRECISION

The two sound-level-meter standards in general use in the United States today are the *USA Standard Specification for General-Purpose Sound Level Meters* (USASI S1.4 1961) and *International Electrotechnical Commission Publication 179, Precision Sound Level Meters* (IEC 179, 1965). Recently there has been an increasing call by specification-writing and code-writing committees for measurements made with precision sound-level meters. For those accustomed to thinking in terms of a single standard, this new emphasis on precision meters may cause some confusion. Whatever confusion there is is not mitigated by the fact that the general-purpose standard is actually tighter than the precision standard at frequencies below 1000 Hz. Perhaps we can clarify the situation somewhat by comparing the pertinent characteristics and tolerances allowed by the respective specifications.

Both standards specify the same three weighting characteristics: A, B, and C. The IEC (Precision) Publication

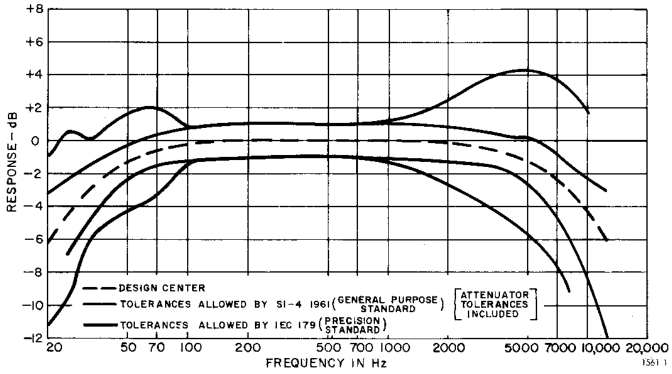


Figure 1. C-weighted response tolerances allowed by general-purpose and precision standards.

requires that an instrument include at least one of the three characteristics; the USASI (General-Purpose) Standard requires all three. As a beginning let us compare the C-weighting characteristics of the two standards.

Figure 1 shows the design-center C-weighted response, along with the tolerances allowed by the two sound-level-meter standards. Note that between 800 and 100 Hz the tolerances are identical, below 100 Hz the general-purpose specification has tighter tolerances, and above 800 Hz the precision specification has tighter tolerances. Note also that the general-purpose specification extends only to 10 kHz vs 12.5 kHz for the precision specification. One further important fact becomes evident as we study the curves and notes of Figure 1: The general-purpose tolerances apply at *any* attenuator setting, while the precision tolerances apply only at the reference setting of the attenuator or at a sound-pressure level of 80 dB. An additional tolerance is permissible at attenuator settings other than 80 dB.

In practice, it is the microphone characteristic that usually determines whether a modern sound-level meter qualifies as a precision meter or a general-purpose meter. If the microphone response uses up most of the tolerance available in the general-purpose specification, the manufacturer must then hold his attenuator tolerances much tighter than those allowed even by the precision specification. If the microphone response does not use up the available tolerances, then the manufacturer can relax tolerances on his attenuator (but no more for the general-purpose than for the precision meter).

The precision-meter specification requires that the error in over-all level introduced by a change in range (i.e., by the attenuator) be less than 0.5 dB. The general-purpose specification requires that the attenuator tolerance be no greater than ±0.5 dB between adjacent steps, nor greater than ±1.0 dB between any two attenuator steps.

As the sound level differs from the reference level or the level at which the response is determined, the indicating meter also contributes errors as its pointer moves over its operating range. Both standards place tolerance limits on the meter indication. These limits are stated differently in each standard, but the actual allowable errors are much the same over the most-used portions of the meter scale. IEC 179 (precision)

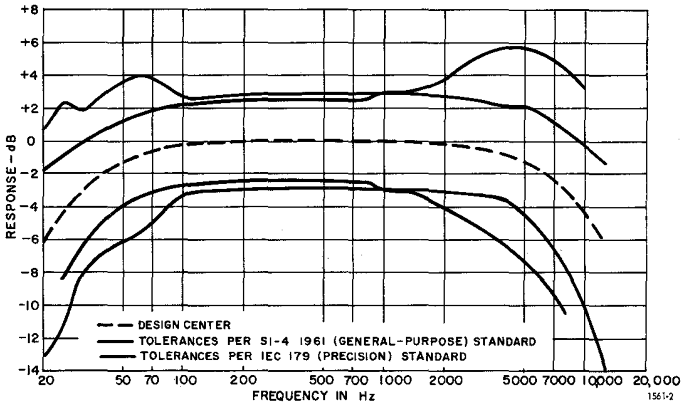


Figure 2. Tolerances imposed on over-all sound-level-meter performance (for C weighting) by general-purpose and precision standards. These tolerances include allowances for attenuators.

allows ±0.2 dB for meter scale graduations and ±0.2 dB for meter scale readability, for a total of ±0.4 dB. S1.4 (general-purpose) states that the indication shall be accurate to ±0.5 dB.

Both standards require that the absolute acoustical calibration of the instruments be ±1.0 dB at a given frequency and sound-pressure level. The precision standard (IEC 179) calls for a frequency between 200 Hz and 1000 Hz (preferably 1000 Hz) and a sound-pressure level of 80 dB. The general-purpose standard (S1.4) calls for calibration at 400 Hz or at some frequency between 320 and 500 Hz but does not specify the sound-pressure level.

Figure 2 shows the C-weighted design-center curve for sound-level meters, along with the maximum deviations (including all the tolerances discussed) from the desired curve allowed by the two sound-level-meter standards. For frequencies below 1000 Hz it can be seen that the general-purpose standard requires greater over-all control of the sound-level-meter performance than does the precision sound-level meter.

A- and B-weighting tolerances are stated differently for the two types of meter. The precision-meter standard requires the same tolerances on all three weightings (A, B, and C). These are the tolerances shown in red on Figures 1 and 2. The general-purpose standard places a close tolerance on the difference between the A or B weighting and the C-weighted response. At frequencies of 800 Hz and higher, for example, the A and B network tolerance with respect to the C weighting is ±0.5 dB. The precision specifications, while requiring closer tolerance to a design center curve, would permit the A- or B-weighted responses to differ from the C-weighted response by as much as 2 dB. The general-purpose B-weighting tolerance remains at ±0.5 dB, with respect to the C-weighting curve, down to 160 Hz and increases to ±1 dB from 125 Hz down to 25 Hz. The general-purpose A-weighting tolerances (with respect to the C response) increase to ±1 dB at 630 Hz, ±1.5 dB at 250 Hz, and to ±2 dB from 80 Hz down to 25 Hz.

General Radio has for many years been manufacturing sound-level meters to meet the requirements of the USA and international standards for general-purpose sound-level meters. Acoustically these meters have always been well within the requirements of the

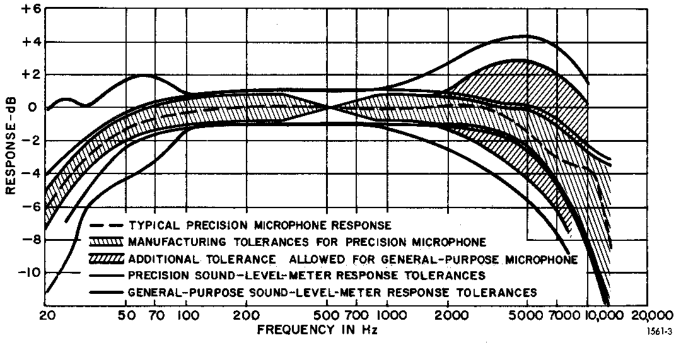


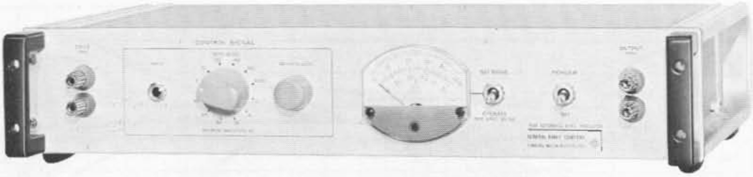
Figure 3. Curves showing data of Figure 1 plus performance of GR precision microphone.

standards, and the electrical characteristics have even met the requirements of the precision-meter standard, IEC 179, 1965. As microphone technology has improved, the over-all characteristics of our meters have come closer and closer to meeting the precision requirements. Careful control of our ceramic microphone response has made this advance possible. Figure 3 illustrates the improvement that has taken place. The broken curve shows a typical C-weighted response of the precision sound-level-meter microphone to sounds of random incidence. The area cross-hatched in red shows the manufacturing tolerances on microphone response. The area cross-hatched in black shows the additional tolerances allowed in the manufacture of microphones for general-purpose sound-level meters. The solid red and black curves duplicate those in Figure 1 and show the response tolerances permitted by the pertinent standards. It can be readily seen that below 2 kHz both kinds of microphones are held to the same close tolerances. Above 2 kHz much tighter control is maintained on the response of the precision microphone than on that of the general-purpose microphone.

While both standards require that the instruments be set to indicate sound-pressure level correctly within ±1 dB at one frequency, all General Radio sound-level meters must indicate sound-pressure level correctly within ±0.5 dB at 500 Hz when compared with a Western Electric type 640AA laboratory standard microphone maintained in calibration by reciprocity techniques and by comparison with other 640AA microphones that have been calibrated at NBS.

One very important fact must be remembered when one is considering whether to choose a precision sound-level meter or a general-purpose unit: For a sound-level meter to meet the stringent high-frequency-response requirements of IEC Publication 179, 1965, the associated microphone must be removed from the instrument and placed at the end of a cable so that the instrument and operator do not disturb the sound field that the microphone is intended to measure. The effects of the operator and of typical instruments on the response of the microphone have been explored and reported by R. W. Young, "Can Accurate Measurements be Made with a Sound-Level Meter Held in the Hand?", *Sound,* Vol. 1, No. 1, January-February 1962.

— E. E. Gross



A VERSATILE LEVEL REGULATOR FOR SWEPT-FREQUENCY SOUND AND VIBRATION TESTING

Swept-frequency sound and vibration tests can be made more quickly and simply if a control device is used to keep the test signal at a constant level as frequency is varied. With the exceptions of some rather expensive special-purpose instruments, no such control unit has been commercially available, and many people have been forced to build their own, often with only limited success.

The new TYPE 1569 Automatic Level Regulator is such a control unit, designed for use in sound or vibration test systems of the type shown in Figure 1. This setup includes an oscillator or a source of swept band-limited noise, a power amplifier and transducer, and a control transducer. As the oscillator is swept, variations in test-chamber sound pressure due to the nonuniformity of the loudspeaker response are sensed by the control microphone and fed to the regulator. This control signal causes the regulator output level to change as necessary to correct for the loudspeaker variations and thus to maintain constant sound pressure.

The control transducer may be a microphone, hydrophone, or vibration pickup. The combination of a high (25-MΩ) input impedance at the regulator's control-signal input and an adjustable high-gain amplifier makes the 1569 compatible with almost any transducer. In effect, the transducer preamplifier is built in, and control signals anywhere in the range from 5 mV to 4 V are acceptable. Sensitivity can be extended to 500 μV with an inexpensive preamplifier (TYPE 1560-P40) powered by the regulator.

The 1569 operates at frequencies from 2 Hz to 100 kHz, satisfying nearly all vibration applications, all airborne sound applications (including high-frequency modeling studies), and most underwater sound applications. The test signal can be a sine wave, a band of random noise, or any signal having components in the 2-Hz to 100-kHz frequency range.

A panel meter indicates the output-signal voltage and the relative level in decibels to show the user where in its 50-dB control range the instrument is

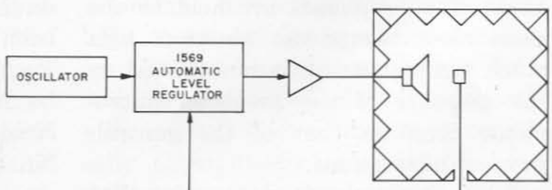


Figure 1. A typical test setup using the level regulator. A microphone in the test chamber serves as a control transducer, sensing variations in sound pressure and sending a control signal to the regulator, which adjusts its output level to correct for loudspeaker variations.

operating. A green band on the meter indicates the operating range that gives greatest signal-to-noise ratio and least distortion. Meter calibration is uniform in decibels.

The control rate (the rate at which the regulator corrects errors in the control loop) is adjustable in 1-3-10 steps from 3 to 1000 dB/second. The rate is adjusted to suit the operating frequency range and the magnitude-phase conditions in the control loop. The regulator can be stabilized for time delays as long as one second and can tolerate a narrow filter or a sharp transducer resonance in its loop.

Because the 1569 will normally be used in a system with other instruments, input and output connectors are on both front and rear panels. The 600-ohm output can be connected to any load without affecting the linear operation of the circuit. The output varies from 10 mV to 3 V in normal operation, a range compatible with most power amplifiers.

Applications

In most frequency tests where the ratio of two quantities is to be measured and plotted automatically, the 1569 is a must. Controlling one signal and measuring the other gives the ratio directly. This approach is applicable to tests on amplifiers, loudspeakers, microphones, and other transducers as well as to mechanical impedance measurements.

Warren R. Kundert received his BSEE and MSEE degrees from Northeastern University in 1958 and 1961, respectively. He came to General Radio as a development engineer in 1959, and he was appointed Audio Group Leader earlier this year. He is a member of the IEEE, Acoustical Society of America, Audio Engineering Society, and Eta Kappa Nu.

Aside from simplifying measurements, the 1569 allows tests to be made at a constant output level on devices with limited dynamic ranges. For example, the output of a loudspeaker or power amplifier can be held constant while the input voltage or power is measured.

The following paragraphs describe some test systems typical of those that can be assembled using General Radio instruments.

Transmission-Loss Tests

In transmission-loss testing it is desirable to maintain constant sound pressure on one side of a partition in order to plot transmission vs frequency directly. When such tests are made in accordance with ASTM Designation E 90-66T (*Laboratory Measurement of Airborne Sound Transmission Loss of Building Partitions*), one-third-octave bands of pink noise are used as the test signal. Figure 2 shows a system for performing this measurement. The signal

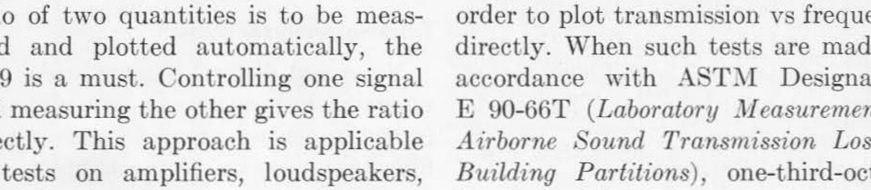


Figure 2. Block diagram of a system for performing transmission-loss tests.

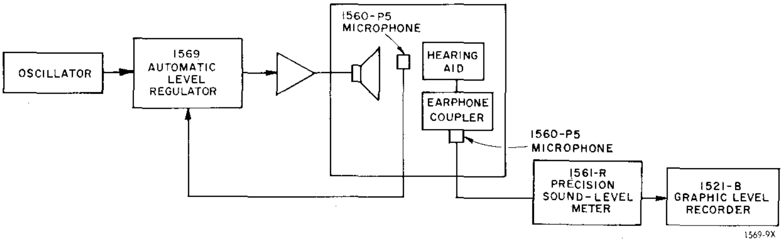


Figure 3. Block diagram of a system for plotting response of hearing aids.

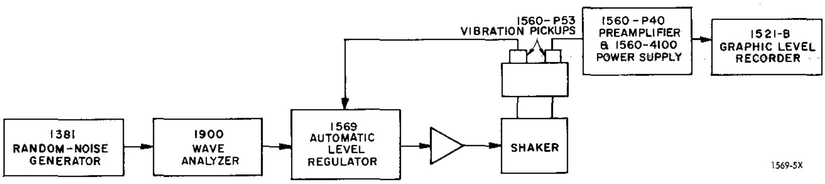


Figure 4. Block diagram of a system for making swept random-vibration tests.

in this setup is derived from the Type 1382 Random-Noise Generator driving the Type 1564-A Sound and Vibration Analyzer.

Hearing-Aid Testing

A setup for automatically plotting the frequency response of hearing aids is shown in Figure 3. The test signal is a sine wave from the Type 1304-B Beat-Frequency Audio Generator. Sound pressure is controlled in a suitable test chamber, and the output from the earphone is coupled to a sound-level meter through a standard coupler. The sound-level meter drives a Type 1521-B Graphic Level Recorder.

Swept Random-Vibration Tests

In the system of Figure 4, a narrow constant band of white noise from the Type 1900-A Wave Analyzer and Type 1382 Random-Noise Generator is fed through the level regulator and power amplifier to an electrodynamic shaker. The control transducer is a Type 1560-P53 Vibration Pickup, which maintains constant acceleration on the table.

Tracking-Analyzer System

Noise and distortion introduced in a vibration test setup can be eliminated from the control signal by means of a filter that "tracks" the test signal. The

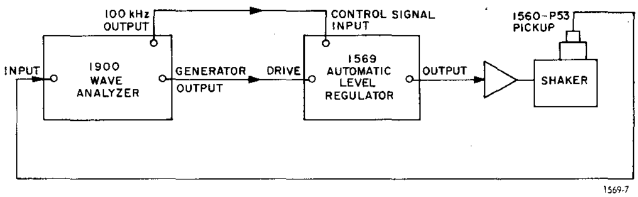


Figure 5. Block diagram showing use of the GR 1900 Wave Analyzer as a tracking filter to eliminate noise and distortion in a vibration test setup.

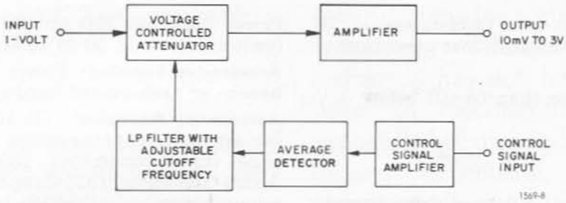


Figure 6. Block diagram of the automatic level regulator.

GR 1900-A Wave Analyzer,[1] used in its tracking mode, provides the necessary sine-wave signal and tracking filter. The setup is shown in Figure 5.

HOW IT WORKS

The 1569 Automatic Level Regulator accepts an ac signal applied to its DRIVE terminals and changes the level of this signal in accordance with a second signal applied to the CONTROL SIGNAL INPUT terminals. The adjusted signal is then fed to the OUTPUT terminals. A small change in the level of the control signal causes a large change in the output, and the result is that a potential level variation of, say, 25 dB in a system is effectively compressed to a variation of less than 1 dB.

The block diagram, Figure 6, shows a voltage-controlled attenuator and an amplifier in the main signal path. The attenuator uses field-effect transistors as variable-resistance elements. The source-to-drain resistances of the FET's are controlled by the dc voltage from the detector and low-pass filter. The filter, whose cutoff frequency is selected by the panel RATE control, determines the speed at which the 1569 can make corrections and limits the lower operating frequency and the amount of delay or phase shift that can be introduced in the control path.

The 1569 automates and simplifies many tests that have been tedious and time consuming. Its great flexibility allows it to be used with a variety of signal sources and control transducers over a wide range of frequencies.

— W. R. Kundert
— C. A. Woodward

A system corresponding to that diagrammed in Figure 5.

[1] A. Peterson, "New Wave Analyzer Has 3 Bandwidths, 80-dB Dynamic Range," *General Radio Experimenter*, April 1964.

SPECIFICATIONS

RANGES

Frequency Range: 2 Hz to 100 kHz.

Control Range: 50 dB.

Compression Ratio: 25 (0.04 dB per dB).

DRIVE (INPUT)

Voltage Required (for normal operation): 1 V.
Impedance: 100 kΩ.

OUTPUT

Voltage: 3 V max to 10 mV min.

Impedance: 600 Ω. Any load impedance can be connected without affecting linear operation of output circuit.

Noise: Typically better than 65 dB below 3 V in 100-kHZ band.

Harmonic Distortion: <1% total for <1-V output level.

Automatic "Shut-Down": A loss of drive (input) voltage from signal source causes the output voltage to drop to zero to protect equipment connected to output.

CONTROL-SIGNAL INPUT

Voltage: 5 mV to 4 V required.

Impedance: 25 MΩ.

Power Required: 100 to 125 or 200 to 250 V (switch selected), 50 to 60 Hz, 4 W.

Accessories Supplied: Power cord, spare fuses, bench- or rack-mount hardware.

Accessories Available: GR 1560-P40 Preamplifier (power for preamplifier available at rear-panel input connector); 1304 Beat-Frequency Audio Generator, 1521 Graphic Level Recorder; microphones and vibration pickups.

Mounting: Rack-Bench Cabinet.

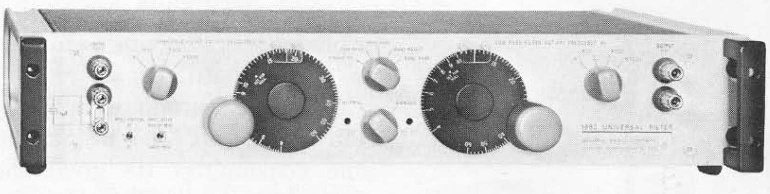
Dimensions (width x height x depth): Bench model, 19 x 3⅞ x 13 in. (485 x 99 x 330 mm); rack model, 19 x 3½ x 10½ in. (485 x 89 x 275 mm).

Weight: Net, 13 lb (6 kg); shipping, 30 lb (14 kg).

Control Rates and Corresponding Min Operating Frequencies:

1000 dB/s	300 dB/s	100 dB/s	30 dB/s	10 dB/s	3 dB/s
600 Hz	200 Hz	60 Hz	20 Hz	6 Hz	2 Hz

Catalog Number	*Description*	*Price in USA*
	1569 Automatic Level Regulator	
1569-9700	Bench Model	$485.00
1569-9701	Rack Model	485.00



A UNIVERSAL FILTER FOR LOW-FREQUENCY WORK

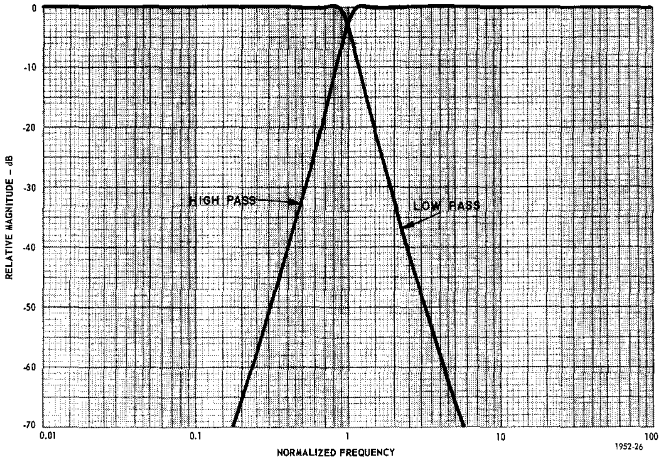
A tunable low-pass/high-pass filter is a very handy device in a variety of laboratory applications. It can, for example, be used to limit bandwidth in a measuring system in order to reduce noise or other interfering signals, to remove one frequency component of a signal in order to measure another, to produce controlled bands of noise, or to function as part of a spectrum analyzer.

The Type 1952 Universal Filter, whose basic magnitude and phase curves are shown in Figure 1, is a tunable high-pass/low-pass filter with some unusual features, notably including the ganged tuning of both filters to provide constant fractional bandwidth, closely controlled and well defined filter characteristics with excellent magnitude response, dc coupling, and the option of line or battery operation.

THE FILTERS

Each filter covers the frequency range from 4 Hz to 60 kHz in four bands, with a cutoff-frequency accuracy of 2% of dial indication. The designs are fourth-order Chebyshev approximations to ideal magnitude response, with negligible (±0.1 dB) pass-band ripple. The Chebyshev design is generally considered to have a magnitude-vs-frequency response superior to that of other designs for a given degree of filter complexity. Though it is an efficient design, it has not heretofore been applied to instruments of this type, probably because it is somewhat more difficult to control, requiring careful circuit design and the use of high-quality components.

The excellent filter characteristics of the 1952 allow narrow band-pass and band-reject characteristics when the filters are combined. Figure 2 shows two band-pass characteristics, of one octave and of 26% (approximately one-



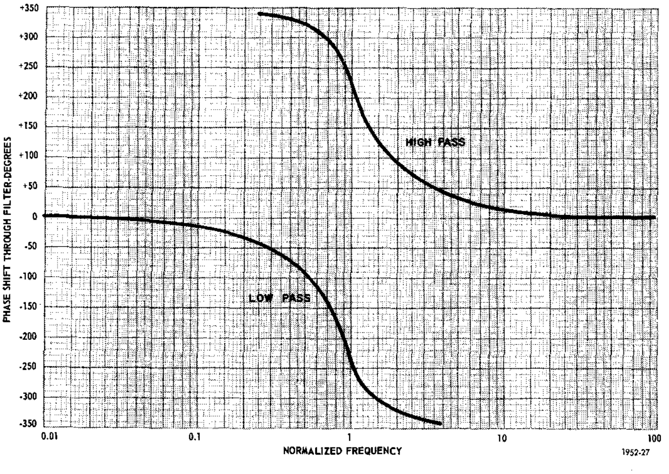


Figure 1. Basic magnitude (top) and phase (bottom) curves for the 1952 Universal Filter.

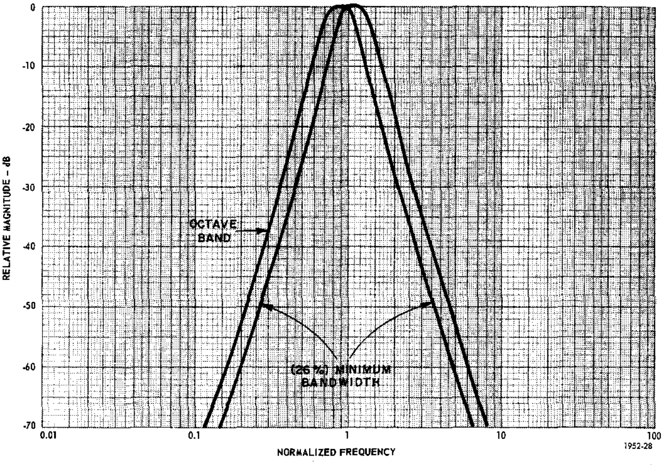


Figure 2. Band-pass characteristics for one-octave and 26% (the minimum) bandwidths.

third octave), the minimum bandwidth. Figure 3 shows a family of band-reject curves, including a narrow "null" response.

It is enlightening to compare the Chebyshev frequency response of the 1952 with the Butterworth response found in some other designs. The fourth-order low-ripple Chebyshev design used in the 1952 is far superior to the fourth-order Butterworth response. The normalized magnitude curves for these two filters, together with the magnitude curve for an eighth-order Butterworth filter, a much more costly design requiring about twice as many circuit components, are shown in Figure 4. In the pass band at 0.8 times the cutoff frequency, the Butterworth filter deviates from the desired uniform frequency response by 0.7 dB, while the error from the Chebyshev filter is only 0.1 dB. In the stop band at, say, 1.2 times the cutoff frequency, the Chebyshev filter has 10.5-dB attenuation while the attenuation of the Butterworth filter is only 7.2 dB.

Ultimate (or asymptotic) attenuation rate is often quoted when stop-band attenuation rate is specified. It should be noted that this rate is the *maximum* rate for the Butterworth design but the *minimum* rate for the Chebyshev design. Figure 4 shows that, at the cutoff frequency, the Chebyshev filter has an attenuation rate of 27 dB/octave (it has already exceeded its asymptotic rate of 24 dB/octave) while the rate for the Butterworth filter is only about 12 dB/octave. In the region near cutoff, where good performance is difficult to achieve, the Chebyshev filter used in the Type 1952 is by far the better choice. Even the eighth-order Butterworth filter, with its much greater circuit complexity, does not perform as well near cutoff as does the fourth-order Chebyshev design.

Ganging

A pulley-and-clutch arrangement operated by a front panel control allows the low-pass and high-pass sections to be ganged. This greatly enhances the usefulness of the filter for band-pass and band-reject applications. Operation is simple. The dials are set to indicate the desired upper and lower cutoff fre-

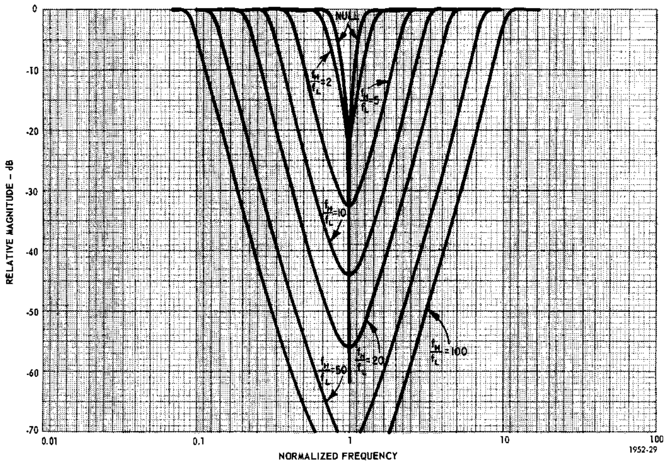


Figure 3. A family of band-reject curves from the 1952.

quencies, the ganging switch is set to GANGED, and the instrument tunes as a constant-fractional-bandwidth band-pass or band-reject filter. (Precise logarithmic dial calibration permits constant fractional bandwidth tuning.) When the dials are set to critical frequencies (indicated by dots) and the ganging clutch is engaged, the filters operate to provide a tunable notch. The narrow response of the notch is shown in Figure 3. The frequency ranges have sufficient overlap to allow tuning through successive ranges without resetting of the dials.

Input/Output Facility

The input circuit includes coupling and attenuation controls as well as a fixed passive filter to prevent frequencies outside the normal operating range of the instrument from overloading the active circuits. The instrument is normally direct-coupled, but a panel switch provides ac coupling with a lower cutoff frequency, about 0.7 Hz.

Output impedance is 600 ohms, and the output can be connected to any load impedance without affecting linear operation of the output circuit.

Input and output connectors are provided on the rear panel as well as on the front for convenience in systems installations.

Battery or Line Operation

Though the instrument is normally operated from a 115- or 230-V line, it can be battery-operated. This feature not only permits operation when line power is unavailable but is also particularly useful in applications requiring total isolation from the power line.

Other Features

The low input terminals are normally connected to the chassis by means of a ground link. In systems installation,

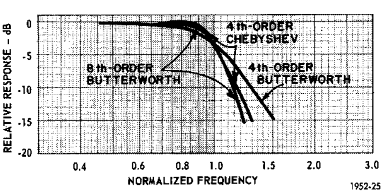
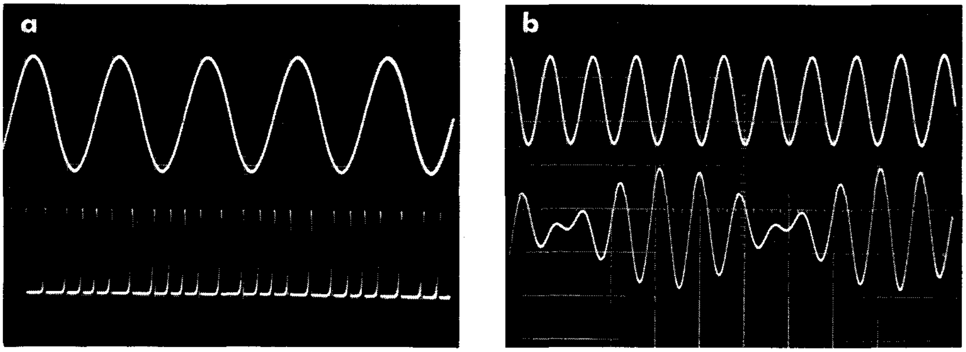


Figure 4. Fourth-order Chebyshev and 4th- and 8th-order Butterworth filter magnitude characteristics around cutoff.



1952-30

Figure 5a. Oscillogram showing results of using the 1952 to recover a 10-kHz modulating signal (upper waveform) from the output of an 1142-A Frequency Meter and Discriminator (lower trace). The signal fed to the 1142 included a 50-kHz carrier modulated at 10 kHz with a deviation of 9 kHz. Figure 5b. Oscillogram showing effect of filter "null" response to separate two equal-amplitude signals at 1 kHz and 1.2 kHz. Upper trace shows 1-kHz signal at 1952 output; lower trace shows the sum of the two sine waves as applied to the filter.

where the chassis of the filter is connected to the chassis of other instruments, it is desirable to have signal currents follow well defined paths. Removal of the ground link connects a circuit consisting of 10 ohms in parallel with 1 μF between the low signal path and the chassis. This impedance is low enough to provide good cabinet shielding with little change in the effects of internal stray capacitance but high enough to prevent low-frequency ground currents in the chassis.

Because low-pass/high-pass filters are often used to remove hum and noise from signals, they should not contribute any themselves. Particular attention has been given to maintaining noise at a minimum level in the 1952.

Applications

The many applications for this versatile filter fall into three basic groups: bandwidth limiting in a measuring system, the generation of controlled bands of noise, and spectrum analysis.

The filter can be combined with the Type 1142-A Frequency Meter and Discriminator, a pulse-count discriminator whose output is a train of "equal area" pulses. Figure 5a shows the output from the 1142-A in the lower trace and the output from the filter in the upper trace. The 1952 filter has been used in its low-pass mode to remove the carrier-related frequency components, leaving the modulating signal. Figure 5b shows the effect of using the "null" response of the filter to separate two equal amplitude signals at 1 kHz and 1.2 kHz.

Controlled bands of noise are generated when the filter is driven by a source of random noise. The new GR Type 1381 and 1382 Random Noise Generators are ideal for this purpose.

In conjunction with a voltmeter or a sound-level meter or vibration meter, the 1952 functions as a spectrum analyzer. The bandwidth is selected with the frequency controls and the controls are ganged. Octave, half-octave, third-octave and other bandwidths used in sound and vibration work are readily produced.

— W. R. Kundert

SPECIFICATIONS

FREQUENCY RANGE

Cutoff Frequencies: Adjustable 4 Hz to 60 kHz in four ranges.

Pass-Band Limits: Low-frequency response to dc (approx 0.7 Hz with ac input coupling) in LOW PASS and BAND REJECT modes. High-frequency response uniform ±0.2 dB to 300 kHz in HIGH PASS and BAND REJECT modes.

Controls: Log frequency-dial calibration; accuracy ±2% of cutoff frequency (at 3-dB points).

FILTERS

Filter Characteristics: Filters are fourth-order (four-pole) Chebyshev approximations to ideal magnitude response. The nominal pass-band ripple is ±0.1 dB (±0.2 dB max); nominal attenuation at the calibrated cut-off frequency is 3 dB; initial attenuation rate is 30 dB per octave. Attenuation at twice or at one-half the selected frequency, as applicable, is at least 30 dB.

Tuning Modes: Switch selected, LOW PASS, HIGH PASS, BAND PASS, and BAND REJECT.

Ganged Tuning: The two frequency controls can be ganged in BAND PASS and BAND REJECT modes so the ratio of upper to lower cutoff frequencies remains constant as controls are adjusted. Range overlap is sufficient to permit tuning through successive ranges without the need to reset frequency controls if ratio of upper to lower cutoff frequencies is 1.5 or less.

Minimum Bandwidth: 23% (approx $\frac{1}{3}$ octave) in BAND PASS mode.

Null Tuning: in BAND REJECT mode, setting the frequency controls for a critical ratio of upper to lower cutoff frequency (indicated on dials) gives a null characteristic (point of infinite attenuation) that can be tuned from 5 Hz to 50 kHz.

INPUT

Gain: 0 or −20 dB, switch selected. Accuracy of gain is ±1 dB, of 20-dB attenuator is ±0.2 dB.

Impedance: 100 kΩ.

Coupling: Ac or dc, switch selected. Lower cutoff frequency (3 dB down) for ac coupling is about 0.7 Hz.

Max Voltage: Max sine-wave input is 3 V rms (8.4 V pk-pk) or 30 V rms with input attenuator at 20 dB. Max peak input voltage for dc coupling is ±4.2 V. For ac coupling max peak level of ac component must not exceed ±4.2 V and dc component must not exceed 100 V. Input can tolerate peak voltages of ±100 V without damage. An LC filter at input limits bandwidth to 300 kHz, thus reducing danger of overloading active circuits at frequencies above normal operating range.

GENERAL

Output: 600-Ω impedance. Any load can be connected without affecting linear operation of output circuit. Temperature coefficient of output offset voltage is between 0 and +4 mV/°C.

Noise: <100 μV in an effective bandwidth of 50 kHz.

Distortion: Max harmonic distortion, with all components in the pass band, for a linear load, is less than 0.25% for open-circuit voltages up to 3 V and frequencies up to 50 kHz.

Power Required: 100 to 125 or 200 to 250 V (switch selected), 50 to 60 Hz, 2.5 W. Or 19.2 V, approx 20 mA from rechargeable nickel-cadmium batteries (not supplied), about 10-h operation. Connections for external battery.

Accessories Supplied: Power cord, spare fuses, bench- or rack-mount hardware.

Accessories Available: Rechargeable batteries (two required) and 1560-P60 Battery Charger.

Dimensions (width x height x depth): Bench, 19 x 3⅞ x 15 in. (485 x 99 x 385 mm); rack, 19 x 3½ x 11¾ in. (485 x 89 x 300 mm); charger, 4¼ x 3¾ x 8 in. (110 x 96 x 205 mm).

Weight: Net, 20½ lb (9.5 kg); shipping, 25 lb (11.5 kg).

Catalog Number	*Description*	*Price in USA*
	1952 Universal Filter	
1952-9801	Bench Model	**$950.00**
1952-9811	Rack Model	**950.00**
8410-1040	**Rechargeable Battery (2 req'd)**	**12.00**
	1560-P60 Battery Charger	
1560-9660	115 volts	**95.00**
1560-9661	230 volts	**on request**

NEW AUDIOMETER CALIBRATION SET

Figure 1. The 1565-Z Audiometer Calibration Set.

The TYPES 1560-P81 and -P82 Earphone Couplers[1] were originally introduced to give industrial hygienists concerned with noise and its harmful effects a means of checking audiometers with the sound-level meters they were using and with which they had become familiar. Demand for audiometer calibrators, however, comes not only from the industrial hygienists but from many people who are not otherwise concerned with noise measurement and who thus cannot be expected to own sound-level meters. For these people we now offer a packaged system designed specifically for checking audiometers. The TYPE 1565-Z Audiometer Calibration Set consists of a sound-level meter (TYPE 1565-A), a sound-level calibrator (TYPE 1562-A), an earphone coupler, instruction sheets, a calibration chart, and spare batteries, all mounted in a small foam-lined carrying case (see Figure 1).

The earphone coupler supplied with the set may be either the TYPE 1560-P82 (USASI TYPE 1) or the TYPE 1560-P83 (9A TYPE)[2]. The calibration chart provided with the system, unless otherwise specified, gives the correct sound-level-meter readings for a TDH 39 earphone (with the MX41AR ear cushion if the 9A-type coupler is used) at a hearing level of 60 dB, based on the ISO-1964 audiometer reference threshold.

With this compact, portable calibration set it is entirely feasible to check the accuracy of an audiometer before every use, the surest way to establish confidence in hearing-loss measurements.

[1] E. E. Gross, "A Standard Earphone Coupler for Field Calibration of Audiometers," *General Radio Experimenter*, October 1966.

[2] See page 21, this issue.

SPECIFICATIONS

Components: Sound-level meter (1565-A); earphone coupler (1560-P82 or -P83); sound-level calibrator (1562-A); storage case.

Dimensions (width x height x depth): 11¼ x 4¼ x 10 in. (290 x 110 x 255 mm).

Weight: Net, 5 lb (2.3 kg); **shipping** (est) 12 lb (5.5 kg).

Catalog Number	*Description*	*Price in USA*
	1565-Z Audiometer Calibration Set	
1565-9900	with 1560-P82 Earphone Coupler	**$565.00**
1565-9901	with 1560-P83 Earphone Coupler	**565.00**

A 9A-TYPE EARPHONE COUPLER

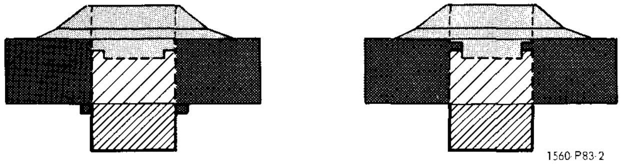


Figure 1. Cross-section drawings of NBS type 9-A (left) and GR 9A-type (right) couplers. Note microphone stop in GR coupler.

Among audiometer specialists, the most popular type of earphone coupler is the type 9-A. We believe, as we stated in an article announcing our type 1 coupler,[1] that the type 1 is easier to use in precision measurements than the 9-A, and this is why our first coupler was a type 1 design. But we also believe in responding to customer demand, and we are therefore adding a 9A-type coupler to our line.

Our interest in ease of use persists, however, and it accounts for a slight difference between the NBS type 9-A coupler and the GR 9A type. As shown in Figure 2, the GR coupler includes a microphone locating stop not found in the NBS 9-A. A series of carefully conducted measurements with both NBS- and GR-type couplers indicates that this minor difference in configuration causes no discernible variation in coupler response. These measurements were made with a Western Electric 640AA microphone and a GR 1560-P5 ceramic microphone.[2] Six earphones were used: two TDH-39's with high electrical impedance, two TDH-49's with high electrical impedance, and two TDH-39's with the standard low electrical impedance. All earphones used the MX41AR Ear Cushion.

The test setup for these measurements was as shown in Figure 2. With the switch in position 2, we applied a constant voltage across the earphone and noted the sound-level-meter reading at each audiometric frequency. Then we set the switch to position 1 and accurately measured the open-circuit

[1] E. E. Gross, "A Standard Earphone Coupler for Field Calibration of Audiometers," *General Radio Experimenter*, October 1966.

[2] B. A. Bonk, "The New General Radio Microphone," *General Radio Experimenter*, May–June 1967.

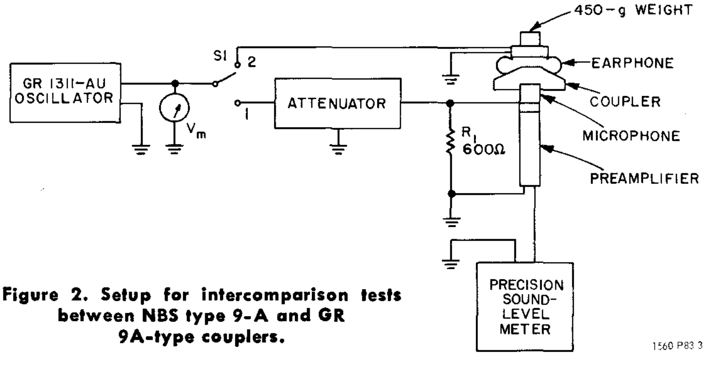


Figure 2. Setup for intercomparison tests between NBS type 9-A and GR 9A-type couplers.

TABLE I
VARIATIONS IN dB FOR 6 EARPHONES

f_{Hz}	*MAXIMUM VARIATIONS BETWEEN MICROPHONES (640AA—GR 1560-P5)* in NBS 9-A Coupler	in GR 9A Coupler	*MAXIMUM VARIATIONS BETWEEN COUPLERS (NBS 9-A—GR 9A)* with 640AA Microphone	with GR 1560-P5 Microphone
125	0.2	0.1	0.4	0.2
250	0.1	0.2	0.2	0.2
400	0.2	0.2	0.1	0.1
500	0	0	0	0
750	0.1	0.1	0.1	0.1
1000	0.1	0.1	0.1	0.1
1500	0	0.1	0.1	0.2
2000	0.2	0.1	0.1	0.1
3000	0.3	0.3	0.1	0.2
4000	0.3	0.4	0.4	0.4
6000	0.4	0.1	0.4	0.5
8000	0.3	0.1	0.3	0.3

microphone voltage required to produce the same sound-level-meter reading noted before. The pressure-response sensitivity of each microphone was carefully measured before the test was performed. Table 1 shows the results.

The hold-down spring is a Hunter Negator spring, designed to exert a force of 450 grams independently of its extension. The spring coil fits in a plastic bobbin so that sharp edges of the spring cannot mar the earphone or the smooth surface of the coupler. This mechanism provides a novel yet simple, dependable hold-down for most earphones and does not require fussy adjustments.

— E. E. Gross

SPECIFICATIONS

Coupler Type: GR 9A type (modified NBS type 9-A).

Volume: 5.642 cm^3, including volume added by microphone (Type 1560-P5 or 1560-2131 cartridge).

Axial Holding Force: 450 grams nominal.

Frequency Range: 100 to 8000 Hz. Duplicates response of NBS 9-A coupler within 1 dB at audiometric test frequencies, when used with TDH-39 or TDH-49 earphones with MX41AR ear cushions.

Dimensions: Coupler diameter 2$\frac{15}{16}$ in., height 1¼ in.; over-all height 3½ in.

Net Weight: 8½ oz (245 g).

Catalog Number	*Description*	*Price in USA*
1560-9683	**Type 1560-P83 Earphone Coupler**	**$35.00**

New GR Catalog

Now available is General Radio's new, 288-page Catalog T, which covers the entire GR product line, from Adaptors to Z-Y bridges. It's free on request, of course.

GR Product Notes

Charting Hearing-Aid Response

In measurements of the response of hearing aids, a recording scale factor of 50 dB per decade is specified by pertinent standards, including USASI Standard S3.8-1967, "Method of Expressing Hearing-Aid Performance." We can now supply 50-dB/decade chart paper for use with our 1304 Beat-Frequency Audio Generator and 1521 Graphic Level Recorder. The catalog number is 1521-9470, and the price is $2.75 per 100-foot roll.

New Coaxial Connector, Adaptor

Two new items further expand GR's coaxial lines: a GR900® precision cable connector for use with RG-58/U and similar cables, and a locking adaptor from GR874® connectors to BNC jacks.

The 900-C58 Precision Coaxial Cable Connector has a SWR much lower than that of even the best-made cables and a frequency range of dc to 8.5 GHz. It can be used with the following cables: RG-58/U, -29/U, -55/U, -141A/U, -142A/U, -159/U, and -223/U.

The 874-QBPAL Adaptor is one of 30-odd adaptors that help make GR874 one of the most versatile lines of coaxial equipment. It should be noted that the "P" in the suffix indicates that the adaptor *contains* a BNC plug and therefore mates with a BNC jack.

Catalog Number	*Description*	*Price in USA*
0900-9431	**900-C58 Precision Coaxial Cable Connector**	**$58.00**
0874-9801	**874-QBPAL Adaptor**	**8.25**

Sixth Conference on Precision Electromagnetic Measurements

The 1968 Conference on Precision Electromagnetic Measurements will be held June 25–28, at the Boulder Laboratories of the National Bureau of Standards, Boulder, Colorado. This meeting will be the sixth in the biennial series begun in 1958. The Conference is sponsored by the NBS Institute for Basic Standards (U. S. Department of Commerce), the Group on Instrumentation and Measurement of the Institute of Electrical and Electronics Engineers, and the U S Commission 1 of the International Scientific Radio Union (URSI).

The aim of the Conference is the advancement of electromagnetic measurements at levels of precision and accuracy appropriate to national standards laboratories. In this year's Conference, emphasis will be placed on the rapidly developing field of automated precision measurements. Methods for precise pulse and waveform measurements will also receive special attention.

Further information concerning the Conference may be obtained from:

George Goulette
Bureau of Continuation Education
328 University Memorial Center
University of Colorado
Boulder, Colorado 803020.

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* AN AC-DC DVM

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ALSO IN THIS ISSUE:

- AN EXERCISE IN VOLTAGE DIVISION
- NEW COAXIAL CAPACITANCE STANDARDS

VOLUME 42 · NUMBER 5 / MAY 1968

the Experimenter

Volume 42 • No. 5 May 1968

Published monthly by the General Radio Company

CONTENTS

ABOUT THIS ISSUE

When you specify voltage ratio to an accuracy of 0.2 part per million, as we do for our precision decade transformers, it is reasonable for people to ask how you can be so sure. In answer, Henry Hall conducts "An Exercise in Voltage Division" in this month's feature article. . . . Is it too much to expect a single digital voltmeter to be a dc multimeter, a uhf (to 1.5 GHz) voltmeter, and an ac/dc millivoltmeter, to offer dB as well as linear readouts, and to have an input impedance of 100,000 megohms on all ranges? No. GR's first DVM does all this and more, with only two plug-ins (page 8). . . . Among the many improvements incorporated in the latest version of our popular 1650 impedance bridge (page 15) is a conductance-measuring capability. Even if you don't have any conductors you want to measure, the things you can do with ac resistance and conductance measurements should whet your imagination. The new 1650 also looks nicer, is priced lower than its predecessor. . . . With the introduction of the 1405 and 1407 series of coaxial capacitance standards (page 19), the benefits of the GR900® precision connector are applied to standards from 1 pF to 0.1 μF. These benefits include accuracy, repeatability, and traceability to NBS.

The *General Radio Experimenter* is mailed each month without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.

AN EXERCISE IN VOLTAGE DIVISION

Voltage ratio, being dimensionless, has no legal unit and no national standard. While resistance, mass, voltage, etc are all defined in terms of standards kept by national laboratories, anyone can make ratio measurements and claim any accuracy he feels his equipment and technique have given him. Whether or not anyone will believe him is, of course, another matter.

The GR 1493 Precision Decade Transformer can easily be read to one part-per-billion* (ppb) of input. At reasonable frequencies and voltages, ac ratio-measuring systems have sensitivities of this order, and comparisons on dividers can be repeated to better than 10 ppb of input, even to 1 ppb on successive measurements. Using two of these dividers and reversing one with respect to the other, one can determine the correction for the 0.5 point repeatedly to within a few ppb.

With resolution and sensitivity available and with confidence that we knew one point very accurately, it took more restraint than we possessed not to try to get the remaining steps of 1/10 input.

CHOICE OF METHOD

The results of any one determination of ratio would not be of too much use. We'd never really know how close we were, and analysis of all possible errors would take years. Therefore, we decided to use three completely different methods and to see how they compared.

Many methods have been used for ratio measurement. We wanted to use those that offered the greatest accuracy as well as those that used the least special equipment. We ruled out all methods using resistors, even though these can be quite accurate under certain conditions. At 1 kHz and 100 volts, where we wanted to work, low-valued resistors dissipate too much power and thus change value, and high-valued resistors exhibit too much phase shift.

The three methods we decided on were (1) the cyclic capacitor method of Cutkosky and Shields,[1] (2) the "bootstrap" method of Sze,[2] and (3) a "straddling" method based on the common reversing technique.

Cyclic Capacitor Method

This method was chosen because Cutkosky and Shields [1] had such good results: 4 ppb at 50% confidence for a 10:1 ratio. While we didn't plan to duplicate their equipment, as a manufacturer of reference standard capacitors we do have a lot of these available, and this method requires good capacitors. The decade transformer is compared with a capacitive divider, or rather to several capacitive dividers, for each capacitor is used in both halves of the divider, and the average of all measurements is taken. The beauty of the method lies in its mathematics. As an example, take the 3:1 divider shown in Figure 1. If each capacitor in turn is connected to the high input and the others to the low, the average of the three ratios is:

$$\frac{1}{3}\left(\frac{C_1}{C_1+C_2+C_3}+\frac{C_2}{C_1+C_2+C_3}+\frac{C_3}{C_1+C_2+C_3}\right)=\frac{1}{3} \qquad (1)$$

All that is required is that the capacitors maintain their values during the cycle of measurement. This means that

* An American billion, or 10^9, not an English billion (10^{12})

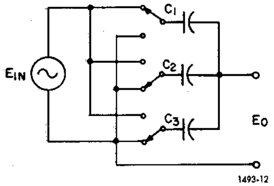


Figure 1. Simplified diagram of capacitive divider using only three capacitors.

they must have low temperature coefficients and that the measurements must be made in a well controlled room.

For our measurements, 10 capacitors were used. For odd values except 0.5, 10 separate measurements had to be made; for even values, five measurements will do (switching two capacitors at a time); at the 0.5 point only two measurements are necessary.

The circuit diagram for these measurements is shown in Figure 2. It should be noted that the capacitors (Type 1404-A, 1000 pF) are three-terminal and have appreciable capacitance from each terminal to case. At first we tried to use a separate guard divider to adjust the case potential as a preliminary balance, but the balance of such a guard was extremely critical. Therefore, we tied this point directly to the output of the divider under test. This causes negligible loading error because (1) the output impedance is low, (2) the stray capacitances form a divider of almost the same ratio, and (3) the cyclic rotation of capacitors rotates these strays also so that this source of error is also averaged out.

The only special equipment required was the special switch box to connect the low side of each capacitor to either the high or low side of the input. This box requires no special internal shielding and is very easy to make. Unfortunately, many laboratories may not have ten 1404-A capacitors.

The Bootstrap Method

This method uses a shielded, 10:1 transformer as a yardstick to measure each step of one-tenth input. While this yardstick may not be completely accurate, we know that the sum of the 10 steps must be unity. This extra condition gives the extra equation necessary to determine all voltages. Putting this into algebra instead of words, we have, from Figure 3:

$$E_n - E_{n-1} = \delta + E_y, \qquad (2)$$

and because the input is unity by definition:

CYCLIC CAPACITOR METHOD

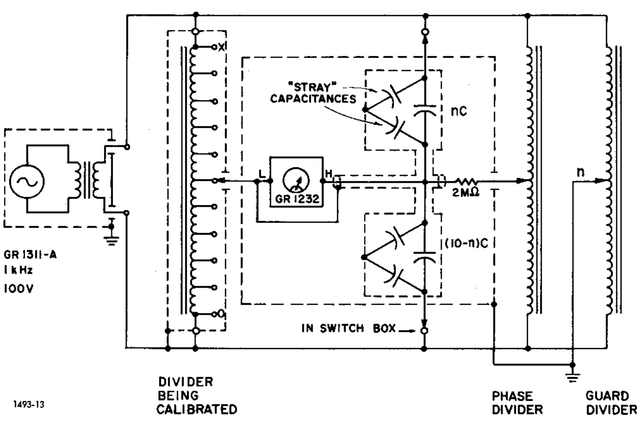


Figure 2. Connection diagram for "cyclic capacitor" method (note: n = number of tenths of input).

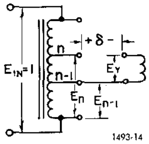


Figure 3. Basic principle of "bootstrap" method, where E_y is the voltage obtained from the shielded 10:1 transformer.

$$10E_y + \sum_{j=1}^{10} \delta_j = 1, \qquad (3)$$

from which

$$E_y = \frac{1}{10}\left(1 - \sum_{j=1}^{10} \delta_j\right). \qquad (4)$$

To find the value of each step, we add up all voltages from zero, or:

$$E_n = \sum_{k=1}^{n} \delta_k + nE_y$$

$$= \frac{n}{10} + \sum_{k=1}^{n} \delta_k - \frac{n}{10} \sum_{j=1}^{10} \delta_j. \qquad (5)$$

In actual dividers, the voltage difference between the 0 and full-scale settings is not equal to the input voltage because of voltage drop in the wiring. Therefore, the input voltage must be measured at the high and low input terminals if expression (3) above is to be valid. The corrections for the 0 and X (10) positions can easily be determined from the small difference between these settings and the input terminals.

While in theory the voltage, E_y, could be any value, it is highly desirable that it be as nearly equal to one-tenth as possible, so that all differences will be small for accurate measurement and so that changes in the E_y error will not be important. This transformer must be doubly shielded; we want E_y to be independent of ratio setting, and this would not be so if there were capacitances to either winding from variable voltages. We first used a simple shielded toroidal transformer, which had an error of about 30 ppm (30,000 ppb) of input. This would not have been too bad if it was very constant, but unfortunately this ratio depended heavily on input voltage. A better transformer was required.

In his paper on this method Sze used a "two-stage transformer" as described by Brooks and Holtz[3] and by Cutkosky[4]. Such a circuit greatly reduces the error due to voltage drop in the primary by sampling the flux and adding additional voltage. The full input voltage in Figure 4 is applied to winding #1. However, because of the voltage drop due to z_1, e_2 and e_3 are slightly low. The difference between e_{in} and e_2 is applied to a second transformer whose output is added to e_3 as a correction.

The transformer we used, shown in cross-section in Figure 5, was constructed quite differently from Sze's. Two of these transformers were made, and both had ratio errors of less than 20 ppb of input (or 200 ppb of output). Note that the outer winding of the two

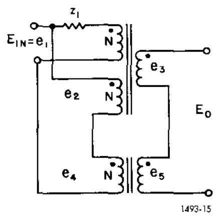


Figure 4. Simplified diagram of "two-stage" transformer, which is actually two transformers interconnected.

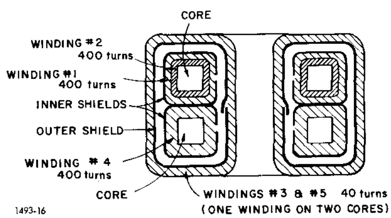


Figure 5. Cross-section diagram of "two-stage" transformer. Note two toroidal cores used.

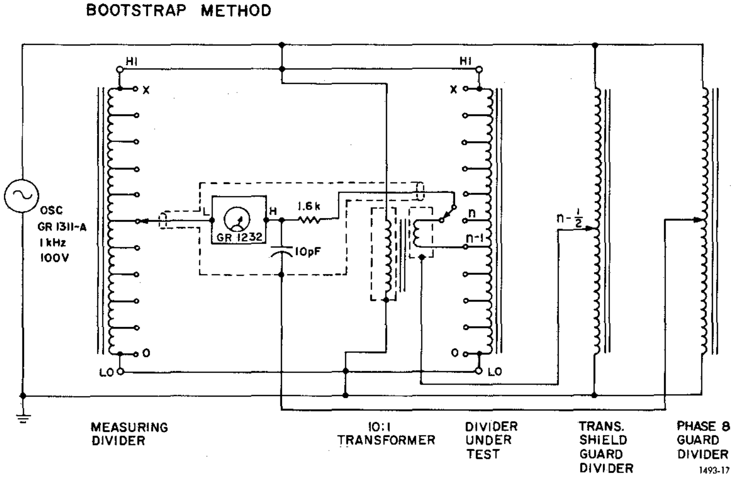


Figure 6. Connection diagram for "bootstrap" method. (Note: n = number of tenths of input. At midpoint n = 5 and n − ½ = 4.5, which is a ratio of 0.45).

windings on the upper core is used as winding #1 because it has more nearly equal coupling to output (#3) and to sampling winding (#2) than the inner winding has.

This bootstrap method was the easiest to use. Measurements and calculations for all 1/10 steps could be made easily in one hour. The complete circuit diagram is shown in Figure 6. The special transformer is not too difficult to make, particularly if a toroidal winding machine is available.

The Straddling Method

As said before, the correction for the 0.5 point can be easily found by reversal of one divider with respect to another. By continuing such division, one can get ¼ points, then ⅛ points, etc (see Zapf[5]). Unfortunately, most dividers are decade rather than binary, so this subdivision is not suitable. However, if one knows the ratio between all pairs of adjacent decade steps, it is possible to calculate the correction for each step. These ratios can be determined by the use of another decade transformer straddling each pair of two 1/10 steps (see Figure 7).

The algebra necessary to determine the expression for any point is tiresome (though not difficult), and so it will not be repeated here. The arithmetic required to evaluate each point after measurement is also dreary, so we programmed a computer to do the job for us.

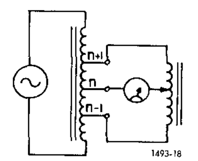
The circuit diagram is shown in Figure 8. Several points are worthy of note:

1. The divider being tested has to have all its 1/10 steps brought out. While it would be possible to bring out the taps of a commercial divider, we made a special divider using the first transformer from a GR 1493.

2. The correction for the straddling divider at the 0.5 point is determined by reversal of its leads.

3. The straddling divider loads the divider under test. In practice, the resulting error is almost negligible. The magnitude of the change caused by

Figure 7. Principle of "straddling" method.



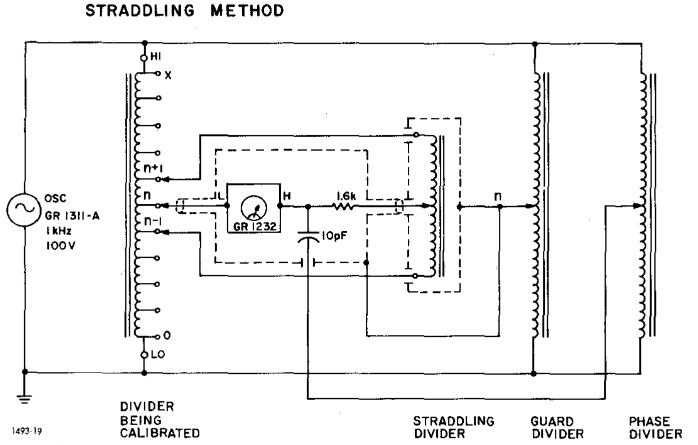


Figure 8. Connection diagram of "straddling" method.

loading was measured by the addition of still another divider in parallel with the straddling divider, and the resulting change of setting was used as a correction.

4. The high and low input terminals should be used for the measurement, instead of the X and 0 points, so that the total voltage is unity by definition.

RESULTS

Figure 9 shows the measured corrections for GR 1493, Serial No. 110, made by two runs using each method. Note that the worst spread of points is less than 30 ppb. Many more runs have been made by the bootstrap method, and they all fell within this spread. Also in this plot is the 0.5 point determined by simple reversal. Its location is reassuring.

We are somewhat hesitant to place a number on our uncertainty because of the clamor this usually raises from statisticians. We prefer to let the results speak for themselves. However, we feel quite confident that the average of these measurements is within 20 ppb of the true ratio. Some readers may disagree with this conclusion and argue that all three methods could be wrong in the same direction, which could be true. Maybe someday we'll try still another method or two to see how they come out.

An independent check is our NBS calibration, even though the Bureau claimed only 200 ppb of input.* The NBS values differ by as much as 50

* NBS Boulder. They no longer make these measurements.

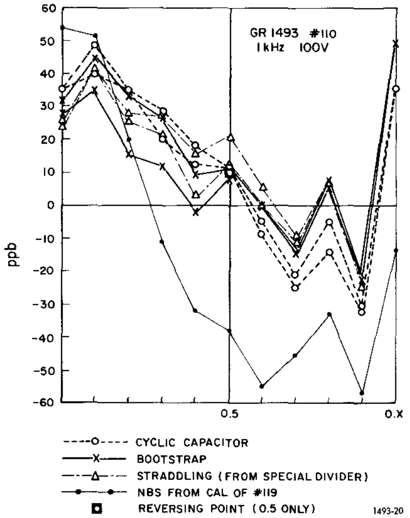


Figure 9. Results of six calibrations, two using each method. Correction in parts-per-billion plotted vs ratio.

ppb from our average value. We have complete confidence that we are within 50 ppb of true value. As a result, we now check 1493's to be within 150 ppb of both NBS and our average value. We thus feel quite comfortable with our specifications for these dividers (±200 ppb of true ratio).

One note of advice for others attempting such measurements. Many of the measurements were made before we realized the importance of demagnetizing the dividers carefully before measurement. In some cases, previous overloads (removed at peak current) caused magnetization that resulted in changes in calibration of up to 10 ppb. Such overloads may occur in a comparison circuit if the two dividers are set to widely differing values and the detector circuit has low impedance. Cyclic demagnetization restores the original calibration. We feel that magnetization may have been our largest source of error.

— H. P. Hall

The author wishes to thank Dr. J. F. Hersh for advice on several details of the measurements

A brief biography of Mr. Hall appeared in the June, 1966 *Experimenter*.

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[1] R. D. Cutkosky and J. Q. Shields, "The Precision Measurement of Transformer Ratios," *IRE Transactions on Instrumentation*, Vol. 1-9 No. 2, September 1960, p. 243.

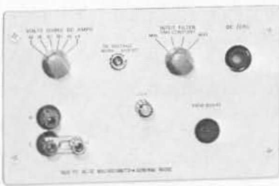
[2] W. C. Sze, "An Injection Method for Self-Calibration of Inductive Voltage Dividers," *NBS Journal of Research, Section C*, January-March 1968.

[3] H. B. Brooks and F. C. Holtz, "The Two-Stage Current Transformer," *AIEE Transactions*, Vol 41, June 1922, pp 382–393.

[4] R. D. Cutkosky, "Active and Passive Direct-Reading Ratio Sets for the Comparison of Audio-Frequency Admittances," *IEEE Transactions on Instrumentation and Measurement*, Vol IM-13, No. 4, December 1964, pp 243–250.

[5] T. L. Zapf, "The Calibration of Inductive Voltage Dividers," *ISA Transactions*, Vol. 2 No. 3, July 1963, p 195.

A VERSATILE DIGITAL MULTIMETER WITH dB READOUT



Type 1820 Digital Voltmeter, with 1820-P1 plug-in in place. Inset shows -P2 plug-in.

Perhaps one of the simplest decisions an instrument user should have to make is which digital voltmeter to buy. This is usually true if one is concerned primarily with dc voltage measurements. For as soon as one specifies the desired dc accuracy, the choice is usually limited to a group of instruments, and

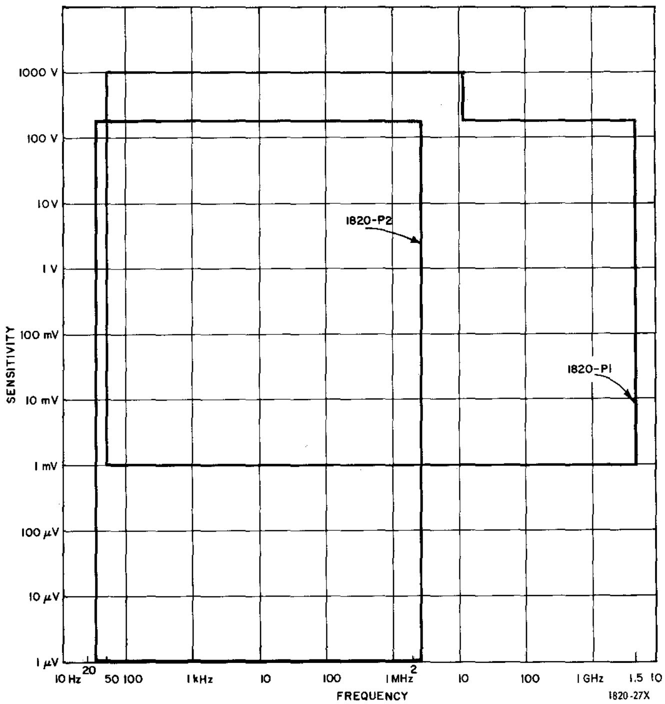


Figure 1. Voltage and frequency range of the 1820 DVM with each of its plug-ins.

one then need only select a unit out of this group to meet his other requirements.

Digital voltmeters are also becoming increasingly useful in the field of ac measurements. Here, however, the choice is more difficult, because many of the features that one would normally seek in an ac meter have, until now, not been available in a digital instrument. General Radio has addressed its first digital voltmeter, the TYPE 1820, to this lack. Designed primarily for ac measurements, the 1820 will also measure dc voltage, resistance, and ac and dc currents.

The 1820 is available with either of two plug-ins, the 1820-P1 and the 1820-P2. Both are ac-dc plug-ins; the differences between them lie in their voltage and frequency ranges and are seen at a glance in Figure 1. Which plug-in to use depends on whether one wants ultra-high sensitivity (1 μV for the last digit) and a frequency range to 2 MHz or ultra-high frequency (to 1.5 GHz) with medium sensitivity (1 mV for the last digit). This combination of plug-ins extends both the sensitivity and frequency ranges far beyond those achieved by other available digital voltmeters.

Direct dB readout, so often desired in ac measurements, is rather common on analog voltmeters, but rare on DVM's. The 1820 offers both linear and dB readouts, switch-selectable at the front panel.

As an ac voltmeter, the 1820 is obviously in a class by itself. As a dc voltmeter, it offers two features not now available in other DVM's. First, no at-

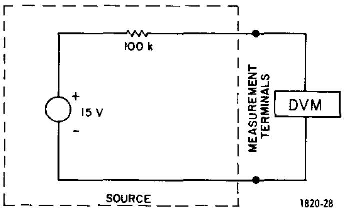


Figure 2. Input impedance of DVM is critical in measurement such as that indicated here (see text).

tenuator is used — on any range. Second, there are as many as six (with the -P2 plug-in) automatically selectable ranges on ac or dc. It is worth taking a few moments to see how these features are translated into increased accuracy of measurement.

One of the commonest errors a user of DVM's encounters, especially with high-impedance circuits, is that due to the finite input impedance of the voltmeter. For instance, suppose that we are about to measure a voltage of 15 volts dc, with a source impedance of 100 kilohms, to 0.1% accuracy (see Figure 2). If we use a popular DVM with a stated accuracy of 0.005% of reading, our measurement will really be accurate only to 1%, because the input impedance of this voltmeter, like that of most others on the market, is only 10 megohms above 10 volts. On the other hand, if we made the same measurement with the 1820 — with a 0.1% basic accuracy specification — we are making a true 0.1% measurement, since the input impedance of the 1820 with the -P1 unit is at least 100,000 megohms on all ranges. (With the -P2 plug-in it is greater than 1000 megohms.)

The availability of six automatically selected ranges with no external preamplifier means, for example, that one can measure +200 volts and, immediately thereafter, −2 millivolts, both with 0.1% accuracy. Again, if we were to use a popular DVM with a nominal accuracy of 0.005% of reading + 0.001% of full scale and with sensitivity of 9.9999 volts full-scale (without external amplifier), our measurement of 2 mV will be good to only 1% accuracy.

The foregoing examples indicate the danger of selecting a voltmeter on the basis of accuracy specifications alone. A moderate-accuracy DVM such as the 1820 may in fact greatly outperform a high-accuracy instrument in many applications.

GENERAL DESCRIPTION

The 1820 is a ramp-type digital voltmeter, shown in generalized block-diagram form in Figure 3. The start pulse initiates the ramp and also starts the time interval. When the ramp voltage equals the input signal, the time interval ends and is indicated digitally.

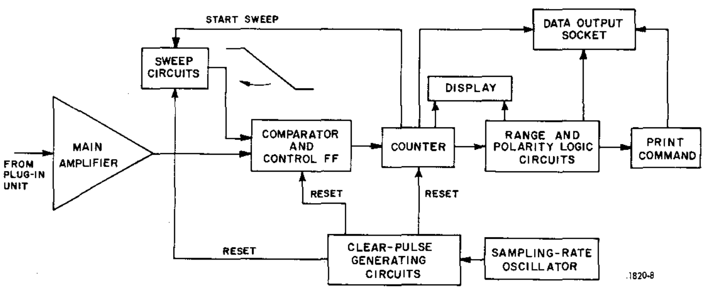


Figure 3. Simplified block diagram of the 1820.

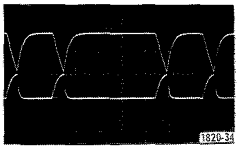


Figure 4. Oscillogram showing the linear (top) and log (bottom) sweeps of the 1820.

If the ramp voltage is linear with time, the time interval is proportional to the input voltage. If the ramp voltage is an exponentially decaying function of time, the time interval will be logarithmically related to the input voltage. Since the current in a series RC circuit is an exponential function of time for a step voltage applied to the series combination, a ramp-type DVM is ideal for logarithmic readout. As a matter of fact, it would be economically prohibitive to incorporate the logarithmic-readout feature of the 1820 in another type of DVM.

Figure 4 shows an oscillogram of the linear and log sweeps of the 1820.

Input Circuits

One of the chief features of the 1820 is the high input impedance on all ranges, achieved through the use of a large negative feedback applied around the input amplifier.

In general:[1]

$$Z_{fb} = Z_{in} \frac{1 - (A\beta)sc}{1 - (A\beta)oc}$$

where $(A\beta)sc$ is the short-circuit loop gain and $(A\beta)oc$ is the open-circuit loop gain. Since the loop gain with input terminals open is zero, the equation above is reduced to:

$$Z_{fb} = Z_{in}\,(1 - A\beta)sc$$

In the 1820, the loop gain is approximately 80 dB on all but the 2-mV range (where it is 60 dB). Thus, if we take the 80-dB figure:

$$Z_{fb} = Z_{in} \times 10^4.$$

The input circuit in either plug-in is a full-wave photochopper operating at about 270 Hz. The impedance of the photochopper is about 500 MΩ in the -P1 and 50 MΩ in the -P2. Thus the typical input impedance is about 5×10^{12} ohms for the -P1 and 5×10^{11} ohms for the -P2 plug-in. Figure 5 shows the ac waveform at the output of the photochopper circuit in the -P1 with the feedback disconnected. It is interesting to note the fast rise and fall times of the chopped waveform at 270 Hz.

Output Amplifier

The output amplifier is capable of ±250 volts swing. This is necessary in order to maintain the high input impedance on all ranges, including the 220-volt range.

[1] T. S. Gray, *Applied Electronics*, Second Edition, John Wiley & Sons Inc., New York,

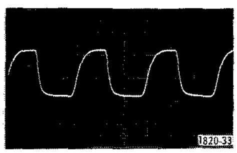


Figure 5. Ac waveform at photochopper output.

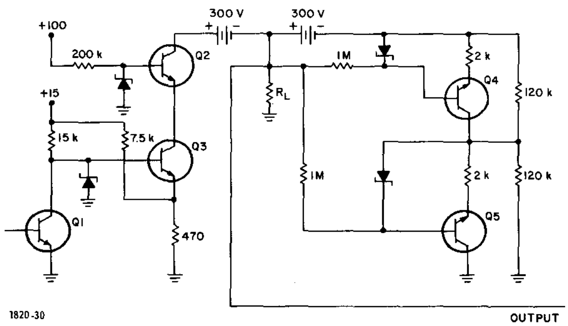
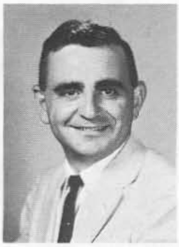


Figure 6. Simplified schematic diagram of the output amplifier.



K. George Balekdjian received his S.B. and M.S. degrees from M.I.T. He joined GR as a development engineer in 1964, and is presently a member of the Industrial Instruments Group. He is a member of IEEE, Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

Figure 6 is a simplified schematic diagram of the output amplifier. Transistors Q4 and Q5 are current sources capable of providing a 250-volt drop across the load resistor, R_L. Transistors Q2 and Q3 are current sinks which, when in the saturated state, will absorb the current supplied by transistors Q4 and Q5.

Transistor Q1 is the low-level control unit for the output amplifier. If this transistor is saturated, Q2 and Q3 are cut off, and all the current supplied by transistors Q4 and Q5 flows into the load resistor. On the other hand, if Q1 is cut off, Q2 and Q3 will be saturated and the voltage across R_L will be -300 volts less a small voltage drop from the collector of Q2 to ground.

UHF Probe

The 1820-P1 uses a uhf probe for measurements up to 1.5 GHz. Voltages up to 200 volts rms can be measured directly with this probe, although at frequencies above 500 MHz the maximum voltage is reduced somewhat. Physically, the probe is similar to that used in the GR 1806-A voltmeter,[2] but the linearizers to correct the response of the probe at different levels are more accurate.

[2] James J. Faran, Jr., "Higher Accuracy, Higher Frequencies with New Electronic Voltmeter," *General Radio Experimenter*, July 1963.

Resistance Measurements

The primary function of a DVM, of course, is to measure voltages. Some instruments, including the 1820, can also measure resistance and current by converting these quantities into voltages.

Currents are usually measured by means of a shunt resistor at the input terminals. The scheme used for resistance measurements is more complex and must involve some kind of current source at the input terminals.

There are two problems associated with high-value ($>10 M\Omega$) resistance measurement with DVM's: (1) The input impedance may not be high enough on all ranges; (2) most transistor current sources become much less than ideal at these impedance levels.

Since the 1820 has at least 10^{11} ohms input impedance on all ranges, it easily meets the first requirement mentioned above. The following technique was used to overcome the limitations of transistor current sources (see Figure 7).

A constant-current source is obtained by means of a floating voltage source in series with a resistance in the feedback path. Since no current will flow in the error amplifier, the current, i, must flow through the unknown resistors, R_x. This will produce across the input terminals a voltage V_x that

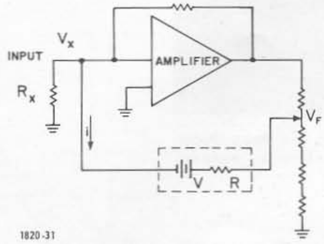


Figure 7. Resistance-measurement technique with the 1820-P1.

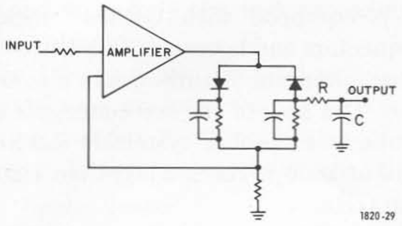


Figure 8. Basic circuit diagram of the operational rectifier of the 1820-P2.

will be equal to V_f by virtue of the high gain of the error amplifier. But, since current must flow through the range-setting resistors at the output of the error amplifier, a correction must be made for each range. One could measure resistors up to 200 megohms with the 1820-P1 if the maximum voltage across the unknown were not limited to 70 volts. Those wishing to measure resistors up to 200 megohms may disconnect the 70-volt limiter, provided that they can tolerate up to 200 volts across the input terminals.

The Average-Responding Detector

The 1820-P2 uses an operational rectifier to convert ac into dc. The detector is calibrated to read in terms of rms volts for sine waves. Figure 8 shows the basic circuit diagram for such a detector.

A simple passive RC filter will filter out the ac signal at the output without introducing long settling times. At low frequencies, the input-filter time constant may be increased by a front-panel control to reduce the ac component further.

Dc Differential Adaptor

It is often desirable to use a DVM in a bridge circuit or in a floating configuration. The common-mode voltages either at dc or power-line frequency usually limit the accuracy of measurement in such situations. The problem is more severe as source impedances greater than 1000 ohms are encountered because of the unbalance between the high and low terminals of the instrument.

The 1820-P3 Dc Differential Adaptor (Figure 9) was designed specifically to improve the common-mode rejection for high-impedance-unbalance circuits. It plugs into either of the two plug-ins to convert the unbalanced input to a closely balanced differential input. This adaptor is actually a low-frequency (20-Hz) sample-and-hold device in which the sampling period is equal to the holding period. The input and output relays are timed so that one pair opens before the other pair closes. Thus the output is never in contact with the input terminal.

Figure 9. The 1820-P3 Dc Differential Adaptor, shown in place on the 1820-P2.

APPLICATIONS

Automatic Dc Leakage-Current Measurements

The 1820 can be used to make leakage-current measurements on capacitors, diodes, transistors, and other components. A typical capacitor leakage-current assembly is shown in the block diagram of Figure 10. In this system, a GR 1770 Scanner connects the components to the voltmeter serially. An external programmer activates a programmable power supply so that proper bias voltages are applied to the com-

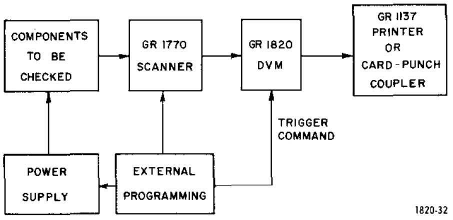


Figure 10. Typical system for measuring capacitor leakage current.

ponents. This programmer also determines the length of time necessary for the component to stabilize before a measurement can be made. After this time interval has elapsed, the programmer supplies a gate signal to the DVM to make the measurement. The 1820 with the -P2 plug-in is quite suitable for low-level leakage-current measurements, since the resolution on the most sensitive range is 1 picoampere. Automatic ranging of the 1820 will display the results within a fraction of a second, on the most appropriate range. A permanent record of results can be obtained with the GR 1137 data printer or a card-punch coupler. The standard data output of the 1820 is 1-2-4-2 BCD, convertible to 1-2-4-8 by a simple modification.

UHF Measurements

The small ceramic diode used in the ac probe has a very low inductance, resulting in a resonance frequency higher than 3000 MHz for the probe assembly. Very precise linearizers for each ac range give excellent linearity even down to 200 mV, one tenth of the most sensitive ac range.

For uhf measurements, the probe should be used in a closed coaxial system to avoid connection errors. The 1806-P1 Tee Connector is available as an accessory to replace the probe cap. It is equipped with GR874® locking connectors and is compensated to introduce minimum disturbance in a smooth line. The SWR of the tee connector and probe in a 50-ohm system is less than 1.10 at 1000 MHz and less than 1.20 at 1.5 GHz.

Precision Ac Measurements

Precision ac measurements at frequencies up to 2 MHz are possible with the 1820-P2. The combination of this DVM and a GR 1310 oscillator is most effective for amplifier response testing. If a lower input capacitance is desired, a Tektronix 10:1 oscilloscope probe can be used directly with the 1820. Also, ac currents can be measured with a Tektronix Type 6020 Current Probe and an adaptor supplied with the -P2 plug-in.

With its 1-μV sensitivity, the 1820 can be used for many measurements that could not before have been handled by a DVM, at least not without external amplifiers. The decibel readout should also prove most useful for many amplifier-response measurements.

Resistance Measurements

As described earlier, the 1820 uses an unusual method of resistance measurement, which allows the user to measure resistances up to 50 megohms without degradation of accuracy.

The 1820 is also useful in resistance measurements on integrated circuits, where the applied voltage must be limited to about 5 volts. The 1820 with either plug-in will measure resistance up to 2 megohms with a maximum of 2 volts applied to the device. A 4-volt Zener diode at the input terminals of the 1820 will prevent any damage to the circuit under test from accidental range changing of the DVM. If a GR 1770

Scanner is added, the test procedure can be automated.

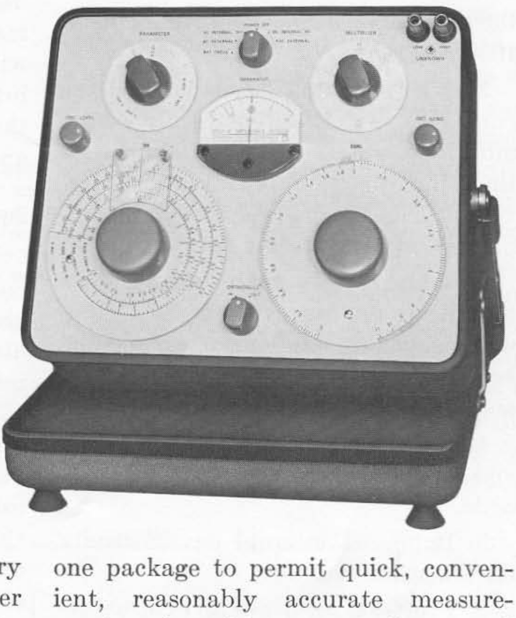
The 1820 can also be used in automatic test setups for resistance sorting. One can measure resistances from 2.000 ohms to 20.00 megohms full-scale with the 1820-P2 automatically and without any external circuitry. With the -P1 plug-in, resistors up to 200 megohms can be measured, with no degradation in accuracy.

— K. G. Balekdjian

Catalog Number	*Description*	*Price in USA*
	1820-A Digital Voltmeter (no plug-in)	
1820-9700	Bench model	**$1985.00**
1820-9701	Rack model	**1985.00**
	Plug-ins and Accessories	
1820-9601	**1820-P1 DC Multimeter/UHF Voltmeter**	**525.00**
1820-9602	**1820-P2 AC/DC Millivoltmeter**	**550.00**
1820-9603	**1820-P3 Differential Adaptor** (for use with 1820-P1 or -P2 plug-in)	**90.00**
1806-9601	**1806-P1 Tee Connector**	**40.00**

Detailed specifications on the 1820 Digital Voltmeter appear in General Radio Catalog T

THE UNIVERSAL IMPEDANCE BRIDGE — NEW FACE, NEW FEATURES



Type 1650-B Impedance Bridge.

One of the truly basic laboratory instruments, along with the voltmeter and the oscilloscope, is the impedance bridge. Although there is a great variety of impedance bridges, by far the most popular is the general-purpose type that combines several bridges in one package to permit quick, convenient, reasonably accurate measurements of resistors, capacitors, and inductors.

The grandfather of this class of instrument was GR's Type 650 Impedance Bridge, succeeded about a decade ago by the 1650. The 1650, with its own generator, detector, and battery power supply, a basic 1% accuracy from 20 Hz to 20 kHz, an Orthonull® balancing mechanism that eliminated sliding balance, and a Flip-Tilt carrying case, soon won an excellent reputation on its own merits, even among a new generation that had never heard of the 650.

The next chapter in the story will be written by the 1650-B, which now replaces the 1650-A. A number of important improvements lie behind the change in letter suffix; despite this fact, it has been possible to reduce the price of the bridge.

The 1650-B offers all the features of the 1650-A plus the following:

1. A conductance bridge, which offers direct micromho readout and extends the range of resistance measurements to 1000 megohms.

2. A slow-motion mechanism on the CGRL dial, which automatically comes into play over a narrow sector each time the direction of rotation is reversed. The effect is a considerable improvement in operating convenience.

3. White sectors on the balance dials to indicate the ranges where the Orthonull balancing mechanism should be switched in for quickest, easiest balance.

4. A battery-check switch position and a corresponding sector on the meter scale.

5. Improved internal dc sensitivity for low-value resistors.

6. Provision for use of an external resistance decade box to extend the DQ range.

7. Access to the bridge arm opposite the unknown arm, so that an external capacitor can be added to obtain a reactance balance of inductive resistors.

8. Relocation of all jacks and terminals except the "unknown" terminals to a side panel, out of the way of the operator.

9. Automatic closure of the BIAS and EXT DQ phone jacks when they are not in use, so that one doesn't have to check the connection of shorting links between binding posts.

10. A redesigned bridge transformer, requiring less drive power at low frequencies.

"Who needs a conductance bridge, anyway?" Because the 1650 is used chiefly to measure components, not including conductors as such, this question will inevitably be asked. Answering it also gives us the chance to discuss the usefulness of ac resistance measurements, possible with the 1650 but not with some other "universal" impedance bridges. Here are just a few of the things you can do with ac resistance and conductance measurements.

Equivalent-Circuit Determinations

When developing an equivalent circuit of an unknown, one usually measures the reactive part on the 1% CRL dial and calculates the resistive part using a DQ measurement, which is only 5% accurate. With ac-resistance capability one can measure the resistive part to 1% and calculate the reactive part using the value of an external capacitor (Figure 1). Use of both methods leads to the most accurate determination of the equivalent circuit.

Impedance Measurements on Batteries

To measure the impedance of a battery, one has only to insert a blocking capacitor in the bridge's bias jack to

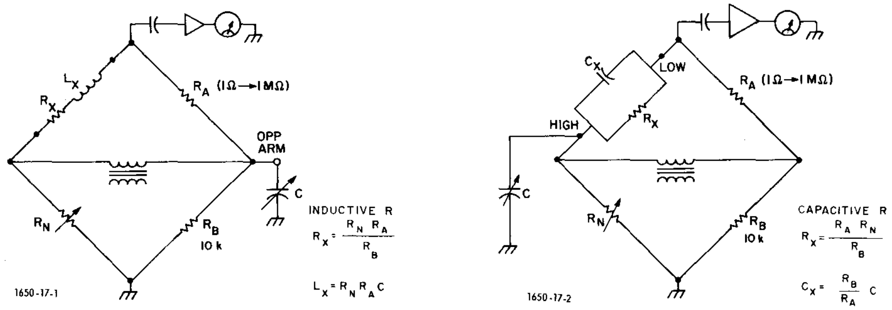


Figure 1. Diagrams showing use of the ac-resistance bridge to make reactive balances.

prevent the flow of large currents through the ratio arm (Figure 2). It is easy to tell sintered-plate (low resistance) and pocket-plate (higher resistance) nickel-cadmium batteries apart by the difference in resistance. The resistance is also a function of the batteries' state of charge. The short-

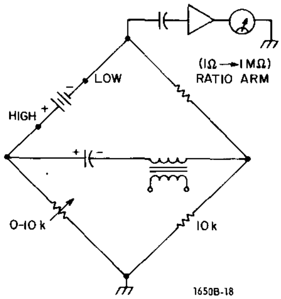
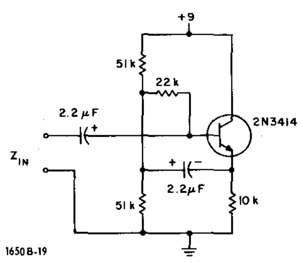


Figure 2. Capacitor inserted in bias jack stops flow of large dc currents through the low-resistance ratio arm.

circuit current of the cell is accurately predicted by the resistance measurement $I_{cell} = \frac{V_{cell}}{R_{ac}}$. This technique is valuable for measuring many low impedances (e.g., simple regulated power supplies and Zener diodes) with dc voltage present.

Input Impedance of Transistor Amplifiers and Other Active Circuits

The 1650-B can be used to determine the input impedance of transistor amplifiers and other active circuits. Consider the bootstrapped emitter-follower circuit of Figure 3a. The ac input resistance is measured and found to be about 760 kilohms, but the null is not sharp, indicating a large reactive component. A capacitor inserted between the OPP ARM jack and the case improves the balance, indicating that the input impedance is inductive. Finally, measurement on the series inductance bridge, with Orthonull, yields an inductance value of 4.92 henrys and a Q of 0.041, which corresponds to a series resistance of 755 kilohms, agreeing with the ac resistance measurement. Knowledge like this is invaluable in explaining the behavior that results when this ampli-



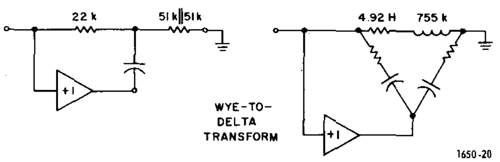
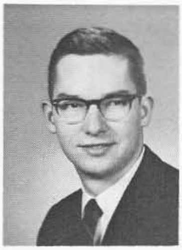


Figure 3a((left). Bridge can be used to determine input impedance of transistor amplifiers such as the bootstrapped emitter follower shown here. Figure 3b (right). A redrawing of the circuit of Figure 3a, showing why the input appears inductive.

fier is incorporated in a large circuit. Figure 3b shows why the input looks inductive.

In general, it is desirable to monitor the output voltage from an active circuit with an oscilloscope to ensure linearity. The oscillator level control can be decreased to prevent overdriving the active circuit.



Charles D. Havener joined GR in 1965, after receiving his B.E.E. and M.E.E. degrees from Cornell. As a development engineer in the Impedance Group, he has worked chiefly on bridge design. He is a member of IEEE and Eta Kappa Nu.

Measurement of Transistor h Parameters

All the complex hybrid parameters of a transistor for the equivalent circuit of Figure 4 can be measured with a 1650-B and an easily constructed test jig. The hybrid π model is in vogue these days, but its parameters are calculated in part from h_{ie}, h_{oe}, h_{fe}, and h_{re}; it is thus still necessary to measure these h parameters. The four-terminal parameters, h_{re} and h_{fe}, must be calculated from impedance values measured with the base ac-shorted to the collector or the emitter. This all may sound a little troublesome, but in practice the measurements go very smoothly. (Further details will appear in a future issue of the *Experimenter*.)

These measurements have proved quite accurate as verified by comparison with measurements made on other testers and by a computer simulation of an amplifier designed to emphasize the measured h parameters. The computed frequency response agreed with that measured within a few dB to 100 kHz when the capacitance in parallel with h_{ie} was included in the model.

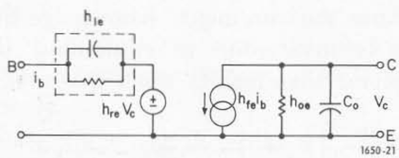


Figure 4. h-parameter equivalent circuit of a transistor.

Ac resistance measurements are also useful for determining the approximate magnitude of a completely unknown impedance, for measuring conductivity of solutions (where dc would polarize the electrodes), and for making incremental resistance measurements on nonlinear components (diodes, thermistors, etc).

Student Use of the Bridge

The universal impedance bridge has long been a mainstay of the student laboratory. In recognition of this fact, the latest in our series of Student-Experiment pamphlets is devoted entirely to the use of the 1650-B by students. Publication STX-107, "The Universal Impedance Bridge," discusses the principle and techniques of bridge measurement and includes some suggested student experiments. It is free on request.

— C. Havener

Catalog Number	*Description*	*Price in USA*
	1650-B Impedance Bridge	
1650-9702	Portable model	$450.00
1650-9703	Rack model	450.00

Detailed specifications on the 1650-B Impedance Bridge appear in General Radio Catalog T

MORE COAXIAL CAPACITANCE STANDARDS

The TYPE 1406 two-terminal capacitance standard[1] has been joined by two new series: the lower-valued TYPE 1405 and the higher-valued TYPE 1407. Altogether, the range of values for two-terminal capacitance standards using precision coaxial connectors now extends from 1 pF to 0.1 μF.

In the 1405 and 1407 standards, as in the 1406, the combination of the GR900® precision coaxial connector and careful capacitor design and fabrication results in significantly improved performance up into the rf range. The high repeatability, low inductance, and precisely known reference plane of the connector effectively eliminate one of the chief problems in the calibration of two-terminal capacitance standards — the variation in stray capacitance from one binding-post connection to another. The low, stable and known inductance of these capacitors keeps the capacitance change with frequency at a very low value.

Two-terminal capacitance standards usually are calibrated in terms of the capacitance added to the bridge terminals. When the terminals of the bridge and of the standard are binding posts, the capacitance added by the standard is not clearly defined. Connecting the standard changes the stray capacitance between the bridge binding posts and also adds stray capacitance from the capacitor case to its own binding posts. The fact that no two binding-post pairs have exactly the same configuration adds to the problem as the standard is connected to various instruments.

[1] R. W. Orr," Capacitance Standards with Precision Connectors," *General Radio Experimenter*, September 1967.

The use of a GR900 precision connector for both bridge and standard terminals changes all this. The connector has a precisely known reference plane, which separates bridge and standard capacitances. Bridge capacitance can be defined and measured as the internal capacitance on one side of this reference plane, and the capacitance of the standard is the total capacitance on the other side.

The result of using the coaxial connector is a reduction in error due to stray capacitance of almost two orders of magnitude over open, two-terminal standards at rf frequencies. The new two-terminal capacitors thus approach the reproducibility found before only with three-terminal capacitance standards at the lower frequencies.

1406 Series Standards

The 1406 series of capacitance standards, described in an earlier *Experi-*

Capacitance values available in the 1405, 1406, and 1407 series coaxial capacitance standards

Series	*Type*	*Capacitance*
1405	E	1 pF
	D	2 pF
	C	5 pF
1406	E	50 pF
	D	100 pF
	C	200 pF
	B	500 pF
	A	1000 pF
1407	A	0.001 μF
	B	0.002 μF
	C	0.005 μF
	D	0.01 μF
	E	0.02 μF
	F	0.05 μF
	G	0.1 μF

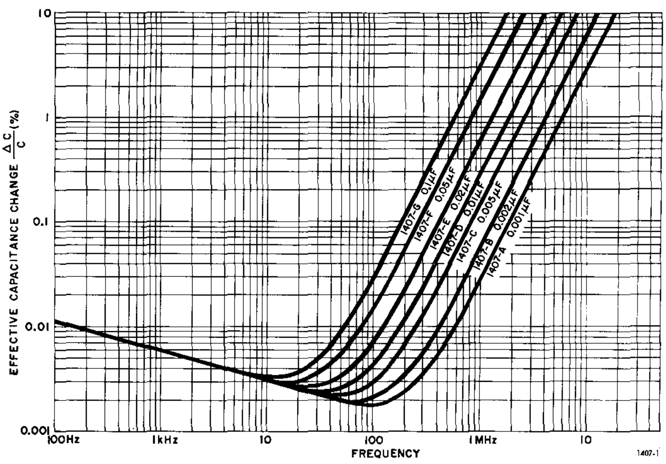


Figure 1. Capacitance-vs-frequency characteristics of the 1407 capacitors.

menter[1], include fine standard capacitors, from 50 to 1000 pF. To these are now added seven TYPE 1407 standards, in values from 0.001 to 0.1 μF, and three 1405 units, in values of 1, 2, and 5 pF.

Type 1407

The TYPE 1407 capacitor consists of a silvered-mica and foil stack which is clamped under heavy spring pressure for mechanical stability. This internal structure is similar to that used in the GR TYPE 1409 standard capacitor, whose stability has been proven over the past decade.

The mica is selected for low dissipation factor. Short, heavy conductors connect the capacitor to the GR900 connector mounted on the top plate of the case, resulting in minimum inductance. The aluminum case is sealed, with a port provided for testing the case under vacuum for leaks. Capsules of silica gel inside the case provide continuous desiccation.

Sealing the mica capacitor in the case with a desiccant makes it possible to reduce the insulating material, other than mica, to a minimum. The only insulation except mica and dry air is a Teflon bead in the GR900 connector and a small amount of low-loss rubber in the gasket sealing the connector. This, together with the high quality of the mica, makes it possible to keep the capacitance changes caused by temperature and interfacial polarization to a minimum.

In the 1407 series, interfacial polarization in the mica dielectric is a source of capacitance change with frequency at the lower frequencies (see Figure 1). Careful choice of and drying of the mica and the elimination of practically all other insulation keep this effect at a

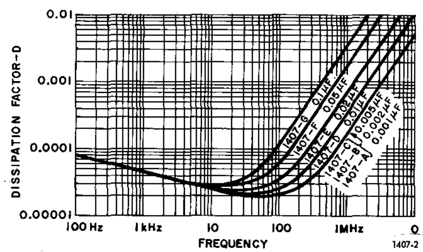


Figure 2. Dissipation-factor-vs-frequency characteristics of the 1407 capacitors.

[1] *Ibid.*

minimum, but there is still a slight increase in effective capacitance below 100 kHz.

At low frequencies (see Figure 2), the dissipation factor also is largely determined by interfacial polarization in the mica dielectric. At higher frequencies, the dominant losses result from the resistance of the metallic conductors. Above 100 kHz, the dissipation factor is approximately proportional to the 3/2 power of the frequency, because of skin effect.

Type 1405

Each of the TYPE 1405 capacitors comprises a section of coaxial line (with inner conductor built up in diameter for the 2-pF and 5-pF units), cantilevered from a PPO (polyphenylene oxide) support rod. The PPO has excellent physical stability and a silicone coating greatly reduces the effects of changes in humidity.

The PPO support contains a threaded PPO cylinder used at the factory for trimming the capacitance to the exact value. A simple residual inductance value is not given for the 1405 because of the distributed nature of the capacitance in this construction.

The capacitance in this 1405 series maintains its dc value to very high frequencies (see Figure 3) and changes thereafter so little that these capacitors are useful in the calibration of measuring instruments well into the uhf region. A 1405 also can serve as a low-capacitance terminating or support means for any two-port device.

Calibration Certificate

A calibration certificate supplied with each capacitor gives the measured value of capacitance at 1 kHz and the temperature and relative humidity at the time of measurement. The calculated effective capacitance of the unit at a higher frequency, such as 1 MHz, also is supplied with some units. The calibrations apply at the well-defined reference plane of the GR900 connector. The 1-kHz capacitance is obtained by comparison with working standards whose absolute values are known to an accuracy of $\pm 0.01\%$.

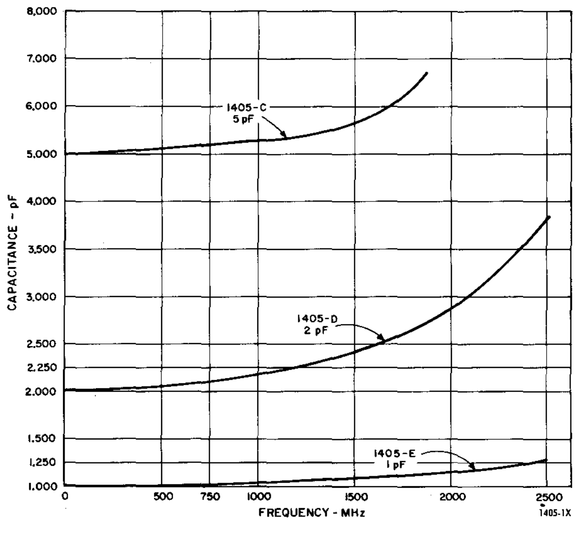


Figure 3. Capacitance-vs-frequency characteristics of 1405 capacitors.

The stability of these capacitance standards permits them to be monitored quite conveniently at 1 kHz to make certain that changes have not occurred over any of the frequency range. The very small capacitance changes that may occur due to normal aging will be the same at all frequencies. Changes in inductance are more serious at the high frequencies. The internal construction is very rugged, however, and any treatment likely to affect the inductance will show up as low-frequency capacitance changes or as obvious damage to the capacitor.

High-precision measurements can be made readily at 1 kHz using any of a number of bridges, such as the TYPE 1620-A, while high-precision measurements at 1 MHz can be made at only a few places outside of NBS. Repeated measurements at 1 MHz are not necessary, owing to the low inductance and the rugged construction of the standard. The fact that any change in its 1-MHz capacitance value will be accompanied by a proportional change in its 1-kHz capacitance value permits a standard to be NBS-calibrated at 1 MHz, and this high-frequency value thereafter can be monitored by means of periodic 1-kHz measurements.

Application Notes

Any open connector, including the GR900, has a fringing capacitance consisting of the total stray capacitance between its terminals. When a capacitor with a GR900 connector is coupled to another GR900 on an instrument, all fringing capacitance is eliminated. The joined connectors act as a straight section of coaxial line.

The value of capacitance calibrated for the standard is the value of the capacitor up to the precise reference plane provided by its connector.

The use of a GR900 precision connector for the instrument terminals provides a precisely known reference plane for the instrument also. Its internal capacitance can be defined as the internal capacitance up to this reference plane. However, an open GR900 connector on an instrument has a fringing capacitance beyond its reference plane. When a capacitor with a GR900 connector is added to the open instrument connector, the net increase in capacitance equals the value of the capacitor, as measured at its reference plane, minus the fringing capacitance of the instrument connector. This fringing capacitance can be anticipated approximately,* but even better accuracy can be obtained if an initial balance is made with a small capacitor whose value is known precisely at its reference plane.

The small capacitance of the 1405-E 1-pF unit, for example, provides the means for eliminating, accurately, this fringing capacitance in the initial setting of a bridge or for setting up an accurate reference plane. This same unit also provides a low-capacitance termination for supporting the inner conductor of the 900-LZ Reference Air Lines. These air lines also are terminated with a GR900 precision connector and can serve as accurate capacitance standards up to 20 pF.

Because of their wide frequency range and acceptance by the National Bureau of Standards for calibration above 30 kHz, these capacitance standards are expected to be used chiefly as standards for the calibration of

* The fringing capacitance of an open GR900 connector is 0.155 ± 0.008 pF in the usual environment on a bridge, with no conductors within several inches of the open connector, and 0.172 ± 0.008 pF with a 900-WO Open-Circuit Termination.

two-terminal bridges and other impedance-measuring instruments. Of course, the most convenient arrangement and the most accurate measurements will result when a bridge is equipped with a GR900 connector.

The next best thing is a precision adaptor, and two of these have been designed specifically for use with bridges. One, the TYPE 1615-P2, is used with the 0.01%, 1-kHz TYPE 1615-A Capacitance Bridge. A trimmer capacitor is included so that the terminal capacitance can be effectively eliminated from the measurement.

The TYPE 900-Q9 Adaptor mates with binding posts on ¾-inch spacing, such as those used on the GR716 Capacitance Bridges, with other posts with a ¼-28 thread, or with tapped holes on ¾-to-1-inch spacing. Among the many instruments accommodated are the Boonton Radio Type 260A Q Meter and the Boonton Electronics Model 75 Capacitance Bridge.

— R. ORR
J. ZORZY

Brief biographies of Mr. Orr and Mr. Zorzy appeared in the August, 1966 *Experimenter*.

SPECIFICATIONS FOR TYPE 1407

Calibration: A certificate of calibration is supplied with each unit giving the measured capacitance at 1 kHz and at a specified temperature. The measured value is the capacitance at the reference plane of the GR900 connector. This value is obtained by comparison to a precision better than ±0.005% with working standards whose absolute values are known to an accuracy typically ±0.01%, determined and maintained in terms of reference standards periodically calibrated by the National Bureau of Standards.

Stability: The capacitance change is less than 0.01% per year.

Accuracy: Within ±0.05%, at 1 kHz, of the nominal capacitance value marked on the case.

Temperature Coefficient of Capacitance: +20 ± 10 ppm/°C, between 10 and 70° C.

Dissipation Factor: 50×10^{-6} typical at 1 kHz and 23° C. Max values given in table below; see Figure 2 for D vs frequency. Measured D at 1 kHz is stated in certificate to an accuracy of ±0.00005.

Series Inductance: 7 nH typical.

Frequency Characteristics: See Figures 1 and 2.

Insulation Resistance: Minimum of 5000 ohm-farads or 100 GΩ, whichever is the lesser, when measured at 500 V dc after two minutes electrification.

Max Voltage: 500 V pk.

Accessories Available: Adaptors 1615-P2 for convenience in calibrating with the 1615-A Capacitance Bridge and 900-Q9 for connecting 1407 to ½-in. x 28 threaded stud (GR 938 Binding Post) or tapped hole.

Terminal: GR900 precision coaxial connector.

Mounting: Aluminum panel and cylindrical case.

Dimensions (dia x ht): 3 x 4¾ in. (77 x 125 mm).

Weight: Net, 1¼ lb (0.6 kg); shipping, 4 lb (1.9 kg).

Catalog Number	*Type*	*Nominal Capacitance*	*Max D at 1kHz and 23°C*	*Price in USA*
1407-9700	**1407-A**	0.001 μF	0.00030	**$ 85.00**
1407-9701	**1407-B**	0.002 μF	0.00025	**95.00**
1407-9702	**1407-C**	0.005 μF	0.0002	**95.00**
1407-9703	**1407-D**	0.01 μF	0.0002	**95.00**
1407-9704	**1407-E**	0.02 μF	0.0002	**100.00**
1407-9705	**1407-F**	0.05 μF	0.0002	**115.00**
1407-9706	**1407-G**	0.1 μF	0.0002	**115.00**

Detailed Specifications on the Type 1405 Coaxial Capacitance Standards appear in General Radio Catalog T

Catalog Number	*Type*	*Nominal Capacitance*	*Accuracy*	*Peak Volts*	*Frequency for 10% C Increase*	*Price in USA*
1405-9702	**1405-C**	5 pF	±0.010 pF	1 kV	0.75 GHz	**$55.00**
1405-9701	**1405-D**	2 pF	±0.005 pF	1 kV	1.0 GHz	**55.00**
1405-9700	**1405-E**	1 pF	±0.005-pF	3 kV	1.7 GHz	**55.00**

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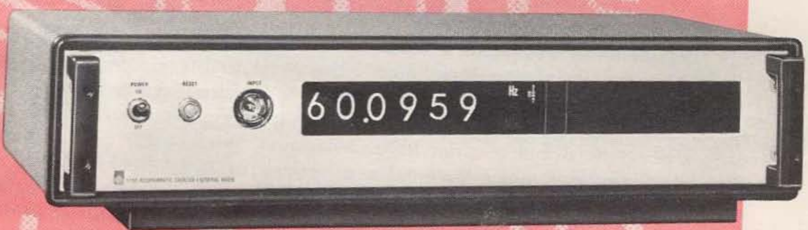
THE RECIPROMATIC COUNTER



Also In This Issue

New Digital Impedance Comparator

New RF Oscillators and Power Supplies

VOLUME 42 · NUMBERS 6,7 / JUNE – JULY 1968

the Experimenter

Volume 42 • Nos. 6, 7 June-July 1968

Published monthly by the General Radio Company

CONTENTS

ABOUT THIS ISSUE

The theme of this month's *Experimenter,* if we were given to thematizing, might be "The Resolution Revolution." The lead article describes a counter that can give six significant digits of readout for a frequency as low as 0.6 Hz. At that point the right-hand digit is actually displaying *microhertz,* a unit probably making its debut in print here. Moreover, this counter can make such ultra-low-frequency measurements in a second or less, thanks to a built-in computer that translates a multiple-period measurement into a frequency readout. . . . Then consider the resolution of the new digital impedance comparator (page 10). Here the percent magnitude difference between two impedances is spread over five digits, with five more digits indicating phase difference. Since the full-scale range can be as little as 10 percent, the comparator detects as little as 10 ppm difference between unknown and standard C, R, or L. . . . Tolerances on inner and outer conductors of our GR900® reference air lines (page 26) are 100 and 50 microinches, respectively, electrical lengths are controlled to ± 0.002 cm, and both conductors are overlaid with pure silver. The same attention to detail has produced a precision attenuator that represents a fivefold improvement in SWR over previously available units.

The *General Radio Experimenter* is mailed each month without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.

THE RECIPROMATIC COUNTER

Automatic ranging, fast measurement of low frequencies, full use of six-digit resolution, and no-hands operation make the 1159 a most extraordinary counter.

Added convenience and speed are the key ideas behind much of today's efforts in instrument design. In the area of frequency measurement, the digital frequency counter has eliminated many of the tedious steps associated with older methods. Still, several nagging problems have persisted, especially in the area of low-frequency measurement.

One way of making a precise low-frequency measurement is to count the number of cycles of the unknown frequency until the desired resolution is achieved. But this takes time — too much time in most cases. For example, a six-digit measurement of 60 Hz would take over 2½ hours by this method.

A much faster way of getting high resolution is to measure period rather than frequency — that is, to count not the cycles of the unknown signal but the higher-frequency pulses from the counter's time base, using the unknown signal to start and to stop the count. If the counter time base is a 10-MHz oscillator, there is no problem in producing a six-digit readout in a very short time (for our 60-Hz measurement, about 0.1 second).

A period readout, however, is the reciprocal of the frequency data usually desired and often required by specifications. The operator can, of course, call on a calculator to perform the simple conversion from period to frequency, but this is a time-consuming and potentially error-producing approach.

A better solution is to build a special-purpose computer into the counter, so that a period measurement is displayed in terms of frequency. Our approach to the design problem of adding a computer has led to an exceptional instrument with automatic ranging as well as reciprocal computation.

THE RECIPROMATIC COUNTER

The foregoing is by way of introducing GR's new Recipromatic Counter, which has a frequency range of from 0.6 Hz to 20 MHz, six-digit resolution, and an average measurement time of 100 milliseconds above 6 Hz, 1 second down to 0.6 Hz. All six digits are always used; a measurement at the counter's low-frequency limit is actually presented with *microhertz* resolution.

Automatic ranging is one of the major features of the new counter. As we have seen above, this implies automatic selection of the number of periods to be measured. Many of the earlier, nonautomatic counters with

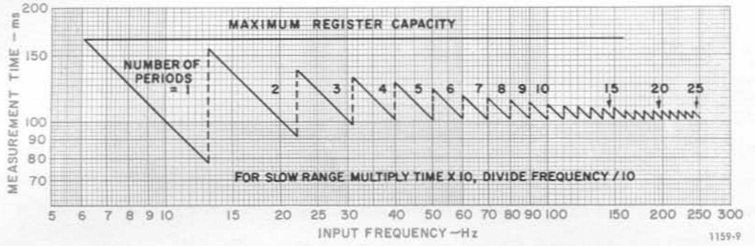


Figure 1. Measurement time as a function of frequency.

which we are familiar have decade period ranges (1, 10, 100 . . . periods). Automatic systems for selecting the decade range are conceivable, but further reflection convinces us that this is the wrong approach. Suppose we program a counter to make a 1-period measurement from 10 to 100 Hz, a 10-period measurement from 100 to 1000 Hz, a 100-period measurement from 1000 Hz to 1 kHz, etc. The measurement time for such a system will vary from 100 ms at the low-frequency end of each range to 10 ms at the high-frequency end. Since the resolution in a multiple-period measurement is proportional to measurement time, a system designed for a specified resolution at the high end of a decade range will require 10 times the measurement time and an extra decade in the counting register for the low end, without gaining any usable increase in resolution. Decade ranges are necessary in standard, general-purpose counters because their only built-in computation facility is that of shifting the decimal point. If, however, we provide an instrument with computational ability sufficient to calculate frequency from period data, then any convenient number of periods can be measured, and it is necessary only to count this number* and to use it as one input to the computation. The quotient of periods/measurement time equals the frequency.

With the foregoing approach our instrument can be programmed for an approximately constant measurement time. The 1159 was designed to give a full six-digit resolution and to use a 10-MHz clock. To obtain this resolution, a measurement time of about 100 ms is required. The lowest frequency that can be measured is that whose period is equal to the maximum capacity of the counting register that counts clock cycles. We must make a single-period measurement from this frequency up to twice this lowest frequency, where we may then change to a two-period measurement. Measurement time will decrease from the maximum value to about half as much at the changeover point. At three times the lowest frequency, where the two-period measurement time has decreased to two-thirds of maximum, we may switch to a three-period measurement and so on, with the lower limit of measurement time increasing with each step. This program makes maximum use of register capacity to give the best possible resolution, but it is rather complicated because of the large number of steps.

The simplest system would be to allow the measurement to be terminated by the first signal pulse occurring after some predetermined measurement time

*Patent applied for.

that is slightly less than half of register capacity. This system makes inefficient use of register capacity, however, since it will be only about half full for most frequencies. A compromise system is to make the lower limit for measurement time a function of the number of periods counted for the first few periods and equal to the nominal time for higher numbers of periods. The 1159 was designed with a register capacity of 167 ms, which corresponds to a frequency of 5.99 Hz. The lower limits of measurement time are programmed as follows, with the results illustrated in Figure 1:

Number of Periods	*Minimum Measurement Time*
1	78.6 ms
2	91.8 ms
3	98.3 ms
4 and higher	101.6 ms

It should be borne in mind that the steps referred to in the above program do not represent range changes in the usual sense; that is, there is no interruption of the measurement, but only a decision as to when to terminate. The range (i.e., decimal-point position and measured units) is determined entirely in the subsequent computation.

In order to measure lower frequencies, the measurement time must be increased. This is accomplished by the lowering of the clock frequency. In the 1159, when the frequency drops below 5.99 Hz, the clock frequency is automatically lowered to 1 MHz, permitting measurements down to 0.599 Hz. When the input frequency rises to about 7.88 Hz, the 1159 automatically switches back to the 10-MHz clock. A three-position toggle switch on the control panel permits operator override of this automatic clock-frequency selection in favor of either the 1-MHz or 10-MHz clock.

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Computation in a small instrument initially seemed prohibitively complicated. Although the availability of integrated circuits helped to bring this approach nearer to economic practicality, the key idea was the use of dual-purpose registers, which function in both the measurement and computation parts of the program. Our first efforts along these lines involved schemes to employ standard counting registers in the computation. Several such methods exist, but they are comparatively slow and, in some cases, less accurate than desired. The 1159 employs two registers that can function either as counting registers or as shift registers by application of proper programming signals. These two registers count signal pulses and clock pulses during the measurement; during computation they become, respectively, the dividend and divisor registers in a fairly standard serial computer.

The method of computation is to subtract the divisor from the dividend and to test for a positive remainder. If the remainder is positive, a pulse is applied to the first decade of the quotient register and another subtraction is performed. When the remainder is negative, the divisor is added to restore a positive remainder, completing the calculation of the first digit. The remainder is then multiplied by ten and the process repeated to calculate the second digit. Usually the dividend is initially smaller than the divisor (except where the signal frequency exceeds the clock frequency), so that the first digit is usually zero. The quotient register is designed to ignore such non-significant zeros, however, and will wait until a non-zero digit is recorded in the first decade before transferring its input to the second decade. The steps required to perform this normalization are counted to determine decimal-point position and units. Computation is continued until seven digits have been computed, the last of which is not displayed but is used to generate a round-off in the quotient register. The time required for computation depends upon the frequency, varying from about 150 μs for 10.0000+MHz to about 675 μs for 9.99999 Hz. Because of this short computation time, the result is displayed directly from the quotient register without the need for buffer storage.

While the program of the TYPE 1159 ensures a constant six-digit resolution, accuracy depends upon the accuracy of the internal crystal oscillator and upon noise. With a resolution of 1×10^{-6} the accuracy requirements of the crystal oscillator are not severe in the light of presently available components. The crystal oscillator for the TYPE 1159 was nevertheless designed for good long-term aging characteristics in order to make the time between recalibrations as long as possible. In addition, facility is included for phase-locking the internal oscillator to an external 100-kHz or 1-MHz reference standard.

The input circuit is of vital importance to the performance of any period-measuring device. As discussed in a previous article,[1] the accuracy of a period measurement is controlled by the over-all signal-to-noise ratio of the source and the input circuit. The noise of the input circuit is of two kinds: random noise generated by the semiconductors in the circuit, and spurious signals harmonically related to the clock and computational frequencies. In order to obtain the lowest noise level in the 1159, the input circuit has a linear input amplifier followed by a Schmitt trigger of comparatively large hysteresis. The effect of any noise in the Schmitt trigger is thereby reduced by the gain of the amplifier so that performance depends primarily upon the noise level of the amplifier. Random noise is reduced by the circuit design of the amplifier, while the spurious frequencies are reduced by careful shielding and decoupling.

The noise level specified for the TYPE 1159 is 50 μV. The error in measurement caused by additive noise is given by the formula $E = \dfrac{1}{n\pi \, S/N}$ where S/N is the over-all signal-to-noise ratio and n is the number of periods measured. For single-period measurements, this implies that the root-mean-squared error due to internal noise is reduced to 1×10^{-6} for a signal of 16 volts rms

[1] R. W. Frank, "Input Noise," *General Radio Experimenter*, February 1966.

(sine wave), for two periods for a signal of 8 volts, and so on. At 1600 periods the signal level needed to make internal noise negligible would be 10 mV rms, which is the specified instrument sensitivity. The formula no longer holds at this point because triggering no longer occurs at the axis crossing. For signal levels of 20 mV, however, internal noise will not affect accuracy at frequencies over 8 kHz for the fast range or 800 Hz for the slow range. From the same formula we may infer that noise effects from the signal source will be negligible for $S/N > 110$ dB — 20 log n. At 1600 periods this requires $S/N > 46$ dB.

In the design of the input circuit, consideration must be given to the types of signal that will be measured. Probably the most common type of signal in frequency measurement is a noisy sine wave with a small amount of distortion. In order to minimize the effects of noise with this type of signal, triggering should take place at the steepest part of the waveform, which is the axis crossing. For this reason the 1159 was made ac-coupled. In order to preserve this relationship under large signal conditions, the signal is clipped symmetrically before each stage of the amplifier.

The 1159 input amplifier also has a programmable bandwidth, which may be changed in decade steps from 10 MHz down to 1 kHz by a rotary switch on the control panel. By reducing the bandwidth to the minimum required by the signal, one may often obtain a significant improvement in signal-to-noise ratio.

For non-sine-wave signals, two additional controls help the user to obtain best results from the 1159. First, a slope switch permits selection of either axis-crossing polarity for triggering. This allows the operator to choose the steeper side of a nonsymmetrical waveform or the side with less noise and jitter (for example, the triggered edge of a one-shot). Second, a trigger-level control helps to ensure triggering on low-level pulses with very low duty ratio.

Although the elimination of range selection removes most of the programmability requirements, all the remaining functions mentioned in the preceding paragraphs are programmable. Programming inputs require DTL micrologic input levels except for the display time (variable resistance) and trigger-level (variable voltage) inputs. Programming connections are made through a multiterminal connector on the rear panel of the counter.

BCD data output is available at a multi-terminal connector on the rear panel. The six digits of readout together with range information are supplied in 1-2-4-8 BCD format at DTL logic levels, together with control signals for operating the 1137 Printer, the 1136 Digital-to-Analog Converter and similar devices.

By the use of a prescaler such as the Type 1156 Decade Scaler or Type 1157 100:1 Scaler, the range of the Type 1159 can be extended to 100 MHz or 500 MHz with the same six-digit resolution and automatic ranging features. Because of its period-type measurement, the Type 1159 does not lose a factor of ten in resolution as is the case with conventional frequency counters. A three-position switch on the rear panel of the instrument permits multiplying the readout by a factor of 10 or 100 to give direct readout when a prescaler is used. This last function, like all the other controls, is programmable.

APPLICATIONS

The 1159 has applications in all areas where general-purpose counters are used to measure frequency. Automatic range switching and constant six-digit resolution make it superior to conventional counters for many high frequency measurements. Specific areas of application include frequency measurements on crystal oscillators, voltage-to-frequency converters, and communication receivers and transmitters. The TYPE 1159 is outstandingly superior to conventional counters in the area of low-frequency measurements, because of the short measurement time needed to produce high resolution. This characteristic is especially useful in measurements on transducers (such as flow meters and radiosondes) and on rotating devices such as gyros and motors, as well as in acoustics, sonar, and vibration measurements.

Complete programmability and data output make the 1159 a natural component of fully automatic test setups for signal-source monitoring, filter and component inspection, etc.

The 1159 is very useful in the routine calibration and setting of oscillators, signal generators, and other signal sources, since it allows the operator to calibrate each point on a dial without having to select the range on the counter. Calibrations are possible over the entire wide frequency range of the 1159 without resort to period measurement at low frequencies. The standard measurement rate of the 1159 of 10 measurements per second is fast enough to allow the operator to tune the oscillator continuously with the counter continuously following his actions.

Many low-frequency oscillators, such as the GR TYPE 1310-A, are stable enough to be set to ±0.001% with the aid of an 1159, even though the settability and accuracy of the oscillator dial calibration allow only ±2%. More stable signal generators, like the GR TYPE 1003 Standard-Signal Generator, can be set to ±0.0002% with the help of the 1159 and the fine-tuning adjustment of the generator. Using the generator dial alone, one can achieve only ±0.1%. For signal sources from 20 to 500 MHz, the 100:1 TYPE 1157 Scaler or the 10:1 TYPE 1156 Decade Scaler can be used ahead of the 1159. The resolution remains a full six digits, and the decimal point as well as the units display remains correct.

When used in combination with a stable signal source, the 1159 lends itself to filter testing. For low-frequency filters, the 1159 gives fast six-digit resolution without requiring that test specifications be written and measurements made in terms of period.

The high sensitivity of the 1159 (10 mV rms over most of its frequency range), coupled with symmetrical limiting of the input signal, helps the 1159 to read the signal frequency even in the presence of large amplitude modulation. The 1159 will read correctly as long as the amplitude of the carrier wave at the negative modulation peak exceeds the maximum sensitivity of the counter. For the TYPE 1003 Standard-Signal Generator, the reading of the 1159 remains correct with amplitude modulations up to more than 90% when the counter is connected to the counter output of the generator.

Other useful input-circuit characteristics of the 1159 are high input resistance and low input capacitance, 1 MΩ in parallel with 27 pF (20 pF when the rear-panel input terminal is disconnected). These make it possible to use an 1158–9600 Input Probe, giving a total input impedance of 10 MΩ shunted by 7 pF and a maximum sensitivity of 100 mV (200 mV at 20 MHz) or better. The counter can then be connected to oscillators and circuits without loading them any more than a good oscilloscope would.

The 1159 will be very useful in production-line work. Its automatic range switching capability and excellent low-frequency performance permit an unskilled worker to obtain a six-digit reading of any frequency between 0.6 Hz and 20 MHz without need for manipulation of controls. The few controls connected with the input circuit and display time are covered by a hinged door in order to prevent misadjustment by curious production help. The use of these controls is unnecessary under normal use of the 1159.

— N. L. Westlake
S. Bentzen

SPECIFICATIONS

Frequency-Measurement Range: 0.6 Hz to 20 MHz. In the fast mode, 6 Hz to 20 MHz; in the slow mode, 0.6 Hz to 9.99999 MHz. Extend range to 100 or 500 MHz without loss of accuracy with GR 1156 (10:1) or 1157 (100:1) Scaler.

Frequency-Measurement Accuracy: $\pm 1 \times 10^{-6}$ $\pm$ clock accuracy $\pm$ noise (see note).

Measurement Rate: Sum of adjustable display time, 0.02 to 10 s and ∞, and measurement time of about 100 ms in fast mode or 1 s in slow mode.

INPUT

Sensitivity: 20 mV rms at 20 MHz; 10 mV rms from 1 Hz to 10 MHz.

Bandwidth: Ac-coupled input, −3 dB at approx 1 Hz. Bandwidth switch sets −3 dB points at approx 10 or 1 MHz, 100, 10, or 1 kHz.

Impedance: 1 MΩ//27 pF for up to 5-V pk-pk input; 0.67 MΩ//30 pF for up to 200 V pk-pk. Input capacitance can be reduced by disconnecting either unused front- or rear-panel input connector. Front only, 20 pF; rear only, 17 pF.

Trigger Threshold: $\pm$20 mV, adjustable.

Slope: Positive- or negative-going, switch-selected.

CLOCK

Internal: 10-MHz, third-overtone quartz-crystal oscillator in proportional-controlled oven.

Temperature Effects: $<1 \times 10^{-6}$ from 0 to 50°C ambient.

Warmup: Within 1×10^{-6} in 10 minutes at 25°C ambient.

Stability: Better than 3×10^{-9} per day after 1 month of operation; better than 1×10^{-6} per year.

External Control: Internal clock oscillator can be phase-locked to external 100-kHz or 1-MHz signal of at least 1 V rms.

GENERAL

Programmability: All control functions can be programmed by contact closures to ground (2- to 4-mA sink current required) except display time, which requires an external resistance of 0 to 100 kΩ, and trigger level, which requires 0 to $\pm$5 V dc.

Note — Noise affects precision of frequency measurement. For additive noise on signal measured, the error in measurement will be $\epsilon = \frac{N}{\pi S n}$ where N is the noise level and S the signal level in the same units; n is the number of periods averaged. Internally produced noise in the counter will determine the limiting error level. For the 1159, this internal noise is approx 50 μV rms.

Data Output: 1-2-4-8 BCD-DTL output for 6 digits of data, decimal point, and measurement units. Data zero is 0.5 V max (12-mA current sinking capability); data one is approx 5 V behind 6 kΩ.

Display: Six neon readout tubes, automatically positioned decimal point, and measurement units. Dimensions can be multiplied by 1, 10, or 100 with rear-panel switch for use with 10:1 or 100:1 prescaler.

Power Required: 100 to 125 or 200 to 250 V, 50 to 400 Hz, 60 W.

Accessories Supplied: Power cord, spare fuses, mounting hardware with rack models.

Accessories Available: GR 1156 (10:1) and 1157 (100:1) Scalers, 1137 Data Printer, 1136 Digital-to-Analog Converter, 1158-9600 input probe (available only with counter).

Mounting: Rack-bench cabinet.

Dimensions (width x height x depth): Bench, 19½ x 4⅞ x 15 in. (495 x 125 x 385 mm); rack, 19 x 3½ x 13⅛ in. (485 x 89 x 335 mm).

Net Weight: Bench, 26 lb (12 kg); rack, 19 lb (9 kg).

Shipping Weight: Bench, 35 lb (16 kg); rack, 28 lb (13 kg).

Catalog Number	*Description*	*Price in USA*
	1159 Recipromatic Counter	
1159-9700	Bench Model	**$2235.00**
1159-9701	Rack Model	**2235.00**
1158-9600	**P6006 Probe,** Tektronix Catalog No. 010-0127-00 (not sold separately)	**22.00**

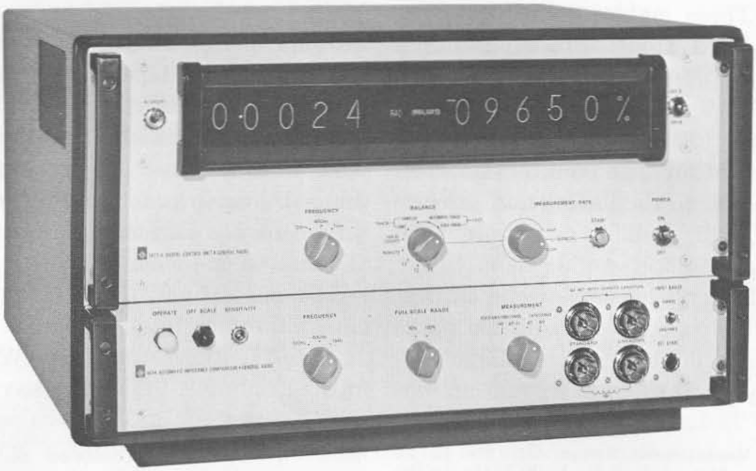


Figure 1. Type 1681 Automatic Impedance Comparator System.

THE AUTOMATIC IMPEDANCE COMPARATOR

The high-performance specifications of today's electronic equipment require increasing use of precision components and force manufacturers and users of components to check large quantities of resistors, capacitors and inductors to close tolerances. The 1680 Automatic Capacitance Bridge,[1] introduced in 1964, represented a major assault on the problems of high-volume measurements. This bridge can make two capacitance measurements a second and presents data in machine-readable as well as visual form. The success of this system was immediate, and the 1680 and associated equipment can now be found in production, quality-control, and inspection installations throughout the world.

The new 1681 Automatic Impedance Comparator System (Figure 1) is similar in appearance to the 1680, and it boasts most of the 1680's advantages, plus a few important ones of its own. Like the 1680, it can make measurements at high speed (60 to 100 per minute), and it presents data in both in-line digital readout and in binary coded decimal form for use with tape punches, card punches, computers, and other recording and handling equipment. The chief differences are (1) that the capacitance bridge is designed primarily for capacitors, while the comparator is equally at home with resistors, inductors, and capacitors, (2) the comparator measures, not absolute value, but percent difference between the unknown component and a standard, and (3) the comparator, spreading a 10- or 100-percent magnitude difference over five digits, is capable of much greater accuracy than is the 1680.

A convenient, if oversimplified, way of describing the new impedance comparator is to consider it as a digital version of GR's long-popular 1605-A analog impedance comparator.[2] Both the 1681 and the 1605-A express differences in impedance magnitude and in phase angle. The great advantages of

[1] R. G. Fulks, "The Automatic Capacitance Bridge," *General Radio Experimenter*, April 1965.
[2] M. C. Holtje and H. P. Hall, "A High-Precision Impedance Comparator," *General Radio Experimenter* April 1956.

the 1681 over its older relation are the high resolution of the digital readout and its compatibility with the automatic component-handling and data-handling equipment so vital in today's technology.

OPERATING FEATURES

The 1681 is a fully automatic comparison bridge providing direct reading in both magnitude and phase-angle difference over a wide range of impedance. The high speed and self-balancing capability of this comparator are obtained without sacrifice of accuracy or impedance range. Basic comparison accuracy is 50 ppm, and impedance range is 2 ohms to 20 megohms. The comparator can detect impedance difference and phase-angle difference to 10 ppm.

The measurement decisions are made by a 1672-A Digital Control Unit, whose circuitry has been described in an earlier article.[3] The operating modes of the digital control unit serve to illustrate the versatility of this instrument.

1. The HOLD RANGE provides speed and maximum resolution of impedance values for components that vary widely in value from the standard. In this mode the bridge balance sequence starts at the most-significant digit.

2. The TRACK CONTINUOUS mode provides continuous tracking of the unknown from the standard. Balance starts from the previous measured value in the least-significant digit. This mode is useful for temperature-coefficient measurements, where small changes in the value of the unknown are being measured.

3. The TRACK SAMPLED mode is used for measurements of components where the difference between the unknown and the standard is small. This is similar to the TRACK CONTINUOUS mode except that balance is achieved here only on command of the operator.

4. The REMOTE position disconnects the front-panel balance control and allows the operating mode to be selected externally by contact closures.

HOW IT WORKS

Figure 2 is a block diagram of the 1681 Automatic Impedance Comparator System. Basically, the 1681 is a transformer ratio-arm comparison bridge with the unknown and standard

[3] Fulks, op. cit.

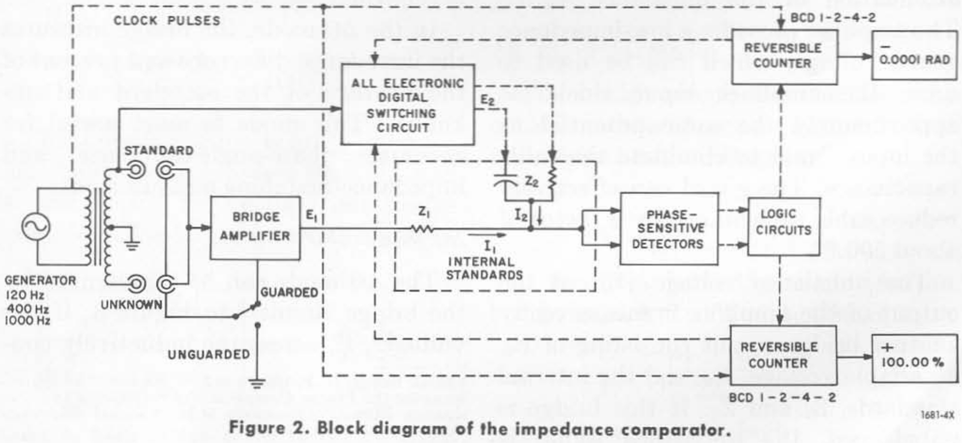


Figure 2. Block diagram of the impedance comparator.

impedances serving as the remaining two bridge arms. The output of the comparison bridge is an unbalance voltage, which is measured by a second bridge, using conventional bridge-balancing techniques as described below.

When the unknown and the standard are unequal, an unbalance voltage results whose phase and magnitude relative to the test voltage are a measure of the impedance difference. The unbalance voltage is fed into a high-input-resistance, low-input-capacitance amplifier so that no loading will occur at the output terminals of the comparison bridge when high-impedance components are measured.

The amplifier's high input impedance is achieved through use of a field-effect transistor in a source-follower configuration. Input capacitance is reduced to a minimum through a guarding technique. Altogether this provides an input impedance of about 1000 megohms and an input capacitance of less than 1 picofarad.

In the case of high-impedance measurement where shielded cable is used to prevent pickup, the cable capacitance to ground will cause phase shift and attenuation of the unbalance signal. The amplifier provides a low-impedance guard voltage, which can be used to drive the amplifier input shield at approximately the same potential as the input signal to eliminate the cable capacitance. The guard can effectively reduce cable capacitance by a factor of about 500.

The unbalance voltage (E_1) at the output of the amplifier is measured by another bridge circuit consisting of E_1, a variable voltage, E_2, and the internal standards, Z_1 and Z_2. If this bridge is unbalanced, the unbalance signal is amplified and fed to phase detectors, which separate the error signal into real and imaginary components proportional to the phase-angle difference and the magnitude difference. These error voltages are fed to the digital logic circuitry, which controls electronic switches to change the reference voltage, E_2, until balance is achieved.

At balance, the counter displays the magnitude and phase-angle difference between the unknown and the standard (in the comparison bridge) on an in-line readout with positioned decimal point and appropriate measurement units.

MEASUREMENT MODES

Two measurement modes are provided, to ensure the highest accuracy under a variety of conditions. Figures 3 and 4 are simplified diagrams of the circuitry used in these two modes.

In the ΔR, ΔL, or ΔC mode, the bridge measures the impedance difference as a percent of the standard impedance. This mode provides the most accurate magnitude-difference measurements over a wide deviation range when the phase-angle difference between the standard and the unknown is very small.

In the $\Delta\Theta$ mode, the bridge measures the impedance difference as a percent of the average of the standard and unknown. This mode is most useful for accurate phase-angle-difference and impedance-matching measurements.

$\Delta\Theta$ Mode

The $\Delta\Theta$ mode can be represented by the bridge circuit[4,5] in Figure 3. If the voltages, E, across the inductively cou-

[4] M. C. Holtje, H. P. Hall, and I. G. Easton, "An Instrument for the Precise Comparison of Impedance and Dissipation Factor," *Proceedings of the National Electronics Conference*, Vol. 10, 1955.
[5] Holtje and Hall, op. cit.

pled ratio arms are equal, the complex output voltage, E_o, is:

$$\frac{E_o}{E} = \frac{Z_x - Z_s}{Z_x + Z_s}.$$

The real part of this expression is:

$$R_e\left(\frac{E_o}{E}\right) = \frac{\dfrac{|Z_x| - |Z_s|}{|Z_x| + |Z_s|}}{1 + \dfrac{\cos(\Theta_x - \Theta_s) - 1}{1 + \dfrac{|Z_x|}{2|Z_s|} + \dfrac{|Z_s|}{2|Z_x|}}}.$$

If $(\Theta_x - \Theta_s)$ is small, say less than 0.1 radian, $\cos(\Theta_x - \Theta_s) \approx$ unity and this equation reduces to:

$$R_e\left(\frac{E_o}{E}\right) \approx \frac{|Z_x| - |Z_s|}{|Z_x| + |Z_s|}.$$

Another approximation is necessary to have the bridge measure the impedance difference as a percent of the standard. If $|Z_x| - |Z_s|$ is very small:

$$\frac{|Z_x| - |Z_s|}{|Z_x| + |Z_s|} \approx \frac{|Z_x| - |Z_s|}{2|Z_s|}.$$

In measurements of resistance, capacitance, and inductance, the in-phase component of the bridge output voltage is a measure of the percent difference of the relative components:

$$\left(\frac{R_x - R_s}{R_s} \times 100\%\right), \left(\frac{C_x - C_s}{C_s} \times 100\%\right),$$

$$\left(\frac{L_x - L_s}{L_s} \times 100\%\right).$$

This approximation is quite good for impedance differences less than 5% and is the basis for comparison-bridge operation. For larger impedance differences, the bridge output becomes quite nonlinear and correction is necessary. (The ΔR, ΔL, ΔC mode, designed into the 1681 comparator to overcome the necessity for correction due to this nonlinearity, is discussed below.) The imaginary part of the bridge output voltage in the $\Delta\Theta$ mode is:

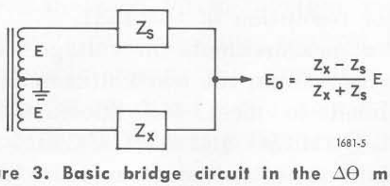


Figure 3. Basic bridge circuit in the $\Delta\Theta$ mode.

$$I_m\left(\frac{E_o}{E}\right) = \frac{\sin(\Theta_x - \Theta_s)}{\cos(\Theta_x - \Theta_s) + \dfrac{|Z_x|}{2|Z_s|} + \dfrac{|Z_s|}{2|Z_x|}}.$$

If Z_x and Z_s are unequal but the difference is small and $\Theta_x - \Theta_s$ is less than 0.1 radian, this expression reduces to:

$$I_m\left(\frac{E_o}{E}\right) = \frac{1}{2}(\Theta_x - \Theta_s).$$

If relatively pure elements are used (C, with D less than 0.1; R, with Q less than 0.1; L, with Q greater than 10), then $(\Theta_x - \Theta_s) \approx \Delta D$ of C and L, or ΔQ of R.

ΔR, ΔL, or ΔC Mode

The bridge circuit for the ΔR, ΔL, or ΔC mode is shown in Figure 4. The unbalance bridge voltage is:

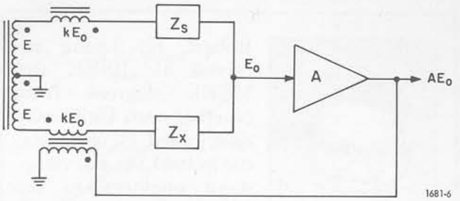


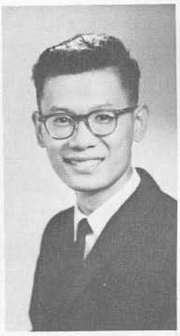
Figure 4. Basic bridge circuit in the ΔR, ΔL, or ΔC mode.

$$\frac{E_o}{E + kE_o} = \frac{Z_x - Z_s}{Z_x + Z_s}.$$

If feedback from the output to the bridge is such that $k = 1$, then:

$$\frac{E_o}{E} = \frac{Z_x - Z_s}{2Z_s}.$$

This is what is desired.



Robert K. Leong received his BSEE and MSEE degrees from Northeastern University and joined General Radio in 1964. As a development engineer in the Low-Frequency Impedance Group, he specializes in the design of bridges and comparators. He is a member of Tau Beta Pi and Eta Kappa Nu.

The real part of the formula is:

$$R_e\left(\frac{E_o}{E}\right) = \frac{|Z_x| \cos(\Theta_x - \Theta_s) - |Z_s|}{2|Z_s|}.$$

For $(\Theta_x - \Theta_s)$ less than 0.01 radian:

$$R_e\left(\frac{E_o}{E}\right) \approx \frac{|Z_x| - |Z_s|}{2|Z_s|}.$$

This approximation is extremely good; it produces a maximum difference error of 0.005% for $(\Theta_x - \Theta_s) = 0.01$ radian.

The imaginary part of the bridge unbalance is approximately equal to:

$$I_m\left(\frac{E_o}{E}\right) \approx \frac{|Z_x|}{|Z_s|}(\Theta_x - \Theta_s),$$

if $(\Theta_x - \Theta_s) \leq 0.01$ radian.

Note that the phase-angle-difference error is directly proportional to the ratio of the unknown impedance over the standard impedance. This error is negligible in comparisons of pure elements (R, L, C) of approximately the same value.

In effect, conventional analog comparison bridges have provided an indication of the percentage difference between the standard impedance and the unknown impedance with respect to the *average* of the two impedances. Ideally, the reading should represent the difference between the unknown and the standard impedances, expressed as a percent of the *standard*. These two approaches produce different readings, but this difference is small when the unknown and standard impedances are close in value. For large differences between the unknown and the standard the readings become widely different, as indicated in Figure 5. For example, an accurate percent difference of 50% would read +40% on one side and −66.7% on the other.

Some analog comparators provide separate meter scales and individual measurement ranges to compensate for this nonlinearity. In the 1681, however, the new ΔR, ΔL, or ΔC mode provides the desired indication over the total measurement range without correction. The other mode is also available and is labeled ΔΘ since it is especially useful in the comparison of phase angles.

APPLICATIONS

Component Measurement

The 1681 can be used to make fast, accurate comparison measurements on almost any type of capacitor. With capacitors of very low value — under 10 pF — the comparator can be made direct-reading in capacitance if a suitable shunting capacitor is connected across the detector. In loss measurements on low-loss capacitors, the user can call on the 10-ppm dissipation-factor resolution of the 1681.

For measurements on voltage-sensitive capacitors, the test voltage can be modified to meet Mil Specifications MIL-C-11015C and MIL-C-39014 for deviation measurements up to ±10%.

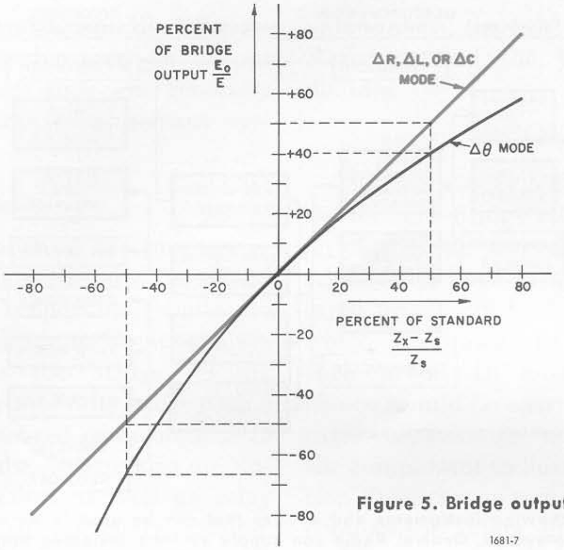


Figure 5. Bridge output characteristics.

1681-7

The comparator will measure inductors, in terms of inductance difference in percent. Thus the instrument can be helpful in the balancing of transformers and in the adjustment of inductors to precise tolerances.

Resistance measurement — again as a percent difference from a standard — is possible with the 1681. Here also the high resolution of the comparator can be put to good use — in measurements of resistance drift, for example, as small as 10 ppm.

Sorting, Inspection, Quality Control

As the number of components to be measured and sorted increases, speed becomes more and more important. The 1681, like its sister instrument, the 1680 Automatic Capacitance Bridge, was designed for installation and use in automatic systems. General Radio manufactures or can provide a wide array of input and output accessories to handle and to sort components and to record measurement data (see Figure 6).

Dielectric and Temperature-Coefficient Measurements

The enterprising engineer or scientist will quickly find many ways to take advantage of the speed and resolution of the 1681. Samples of dielectric materials can be compared, with precise readout of impedance-magnitude and phase-angle differences. Temperature-coefficient measurements on components in environmental chambers usually involve deviation measurements as a function of temperature, and here the comparator, with its 10-ppm resolution and provision for automatic data collection, is of obvious value.

TEST SYSTEMS

Input Devices

Because so many different types of components can be measured over such a wide impedance range, it is impossible to provide a single, all-purpose terminal arrangement. In many cases, the connection of the unknown component to the comparator can be most convenient-

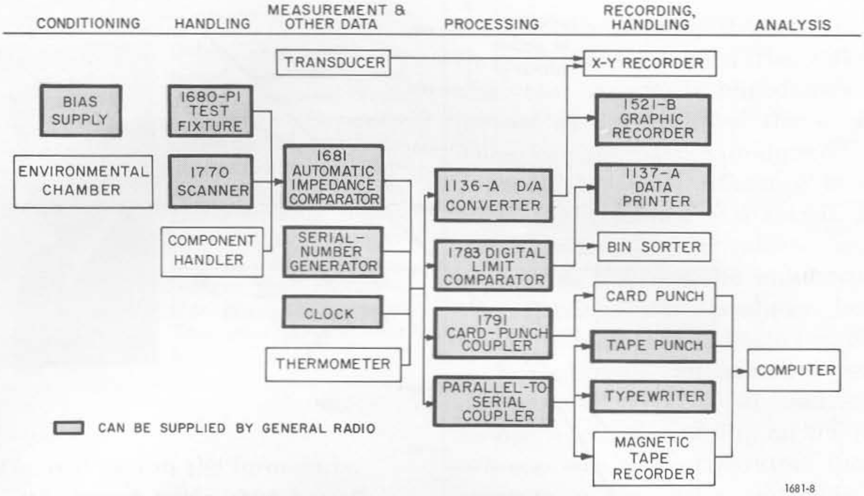


Figure 6. Chart showing instruments and devices that can be used in an automatic component-measuring system. General Radio can supply systems including those components indicated by tinted blocks.

ly made through the use of the 1680-P1 Test Fixture,[6] in which the components are manually inserted.

Automatic input devices, such as the 1770 Scanner System, are available for applications in which components must be connected to the comparator in a prescribed automatic sequence.

Output Devices

Many output devices are available to record the measured data. Among these are printers, analog recorders, card and tape punches, typewriters, and magnetic tape recorders.[7]

The TYPE 1137-A Data Printer (designed for use with GR digital equipment) is probably the least expensive and simplest way of automatically obtaining a permanent record.

In applications where the output data are desired in the form of an analog plot, such as in temperature-coefficient measurements, a TYPE 1136-A Digital-to-Analog Converter can be used to translate the data output into a dc voltage or current for analog recording by the TYPE 1521-B Graphic Level Recorder.

For those who wish to record data in machine-readable form, punched tape is the least expensive solution. A parallel-to-serial converter must be inserted between the bridge and the tape punch because the BCD data from the bridge are punched serially on the tape and the bridge output is presented in parallel form.

A card punch, such as the IBM 526, is the most common output device for obtaining machine-readable records. The chief advantage of the punched card is that, after all the data on a tested component have been punched on a single card, the card can move with the component. The TYPE 1791 Card-Punch Coupler can be used as the parallel-to-serial converter between the bridge and the card punch.

[6] Fulks, op. cit.
[7] H. T. McAleer and R. F. Sette, "Automatic Capacitor-Testing Systems," *General Radio Experimenter*, Nov-Dec 1966.

Typewriters and magnetic tape recorders can also be used as output recording devices; they are generally seen only in large measurement systems.

Processing Equipment

A useful member of the processing-equipment family is the new Type 1783 Digital Limit Comparator (similar to the 1781 Digital Limit Comparator)[8], which compares the bridge reading against manually set limits to determine whether the measured component is in or out of tolerance. It provides GO/NO-GO visual indication as well as relay contact closures for automatic sorting. Further information on this instrument is available on request.

Standards

An external standard is required for the 1681. This standard may be a component identical to the unknown and selected for a nominal value, or it may be a calibrated, fixed standard or, if more convenient, a variable standard. A wide selection of standards is available: the GR 1422 Precision Capacitors (for small values of C), the 1423, 1424, and 1425 Decade Capacitors (for large values of C), the 1433 and 1434 Decade Resistors, and the 1491 Decade Inductors.

BRIDGE OR COMPARATOR?

The availability of both an automatic capacitance bridge (the GR 1680) and an automatic impedance comparator (the 1681), both at approximately the same price, will lead many to wonder which instrument is better suited to their needs. In most instances, the choice should be easy. The bridge indicates capacitance and loss directly, the comparator indicates percent deviation from an external standard. The comparator offers greater resolution, and it measures resistors and inductors just as easily as it does capacitors.

Bridge or comparator, the result is speed and convenience impossible with manually operated instruments. The higher initial price of the automatic instrument, examined in terms of cost per component tested, is seen as a true saving. Indeed, for the high volume measurer of components, the automatic bridge or comparator is now an economic necessity.

— R. Leong

[8] McAleer and Sette, op. cit.

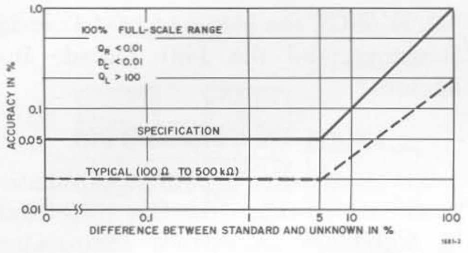
SPECIFICATIONS

	Total Useful Range	*Ranges for Measurement with Stated Accuracy*			
		Full Scale	120 Hz	400 Hz	1000 Hz
Resistance	2 Ω–20 MΩ	100% 10%	10 Ω–2 MΩ 500 Ω–2 MΩ	10 Ω–2 MΩ 500 Ω–2 MΩ	10 Ω–2 MΩ 500 Ω–2 MΩ
Capacitance	20 pF–800 μF	100% 10%	1 nF–100 μF 1 nF– 5 μF	400 pF–50 μF 400 pF– 2 μF	200 pF–20 μF 200 pF– 1 μF
Inductance	400 μH–1000 H	100% 10%	5 mH–1000 H 200 mH–1000 H	2 mH–200 H 50 mH–200 H	600 μH–100 H 30 mH–100 H

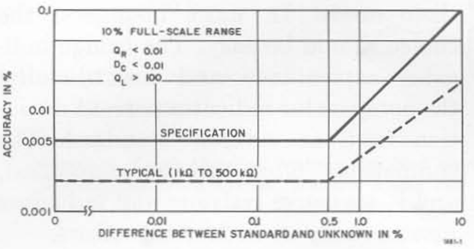
FULL-SCALE RANGES

Magnitude Difference: ±10%, and +100% −40%, full scale.

Phase-Angle Difference: ±0.1 and ±1 radian, full scale. The phase-angle difference is very nearly equal to the D difference (C and L) and the Q difference (R) when the D or Q is less than 0.1.



Accuracy of R, L, or C Difference as Percent of Standard.



ACCURACY

ΔR, ΔL, ΔC Measurement Mode

Magnitude Difference (as % of standard): ±[1% of reading + 0.001 ΔΘ (in counts) + 5 counts].

Phase-Angle Difference: ±[1% of reading + 0.005 ΔZ (in counts) + 5 counts] + additional error when large magnitude differences are measured. Correction chart supplied.

ΔΘ Measurement Mode

Magnitude Difference (as % of average of unknown and standard): ±[1% of reading + 0.001 ΔΘ (in counts) + 5 counts]. Reading in this mode differs from %-of-standard when deviation ≥1%. A correction chart is supplied.

Phase-Angle Difference: ±[1% of reading + 0.005 ΔZ (in counts) + 5 counts].

Max Resolution: 0.001%, 0.00001 radian.

Effects of Leads: For high-impedance measurements with input shield guarded, shielded cables up to 3 feet long can be used without significant error from cable capacitance.

Voltage Across Standard and Unknown: 0.3 V for 100%-full-scale range; 3 V for 10% range. Test voltage can be modified on request to meet MIL Specifications MIL-C-11015C and MIL-C-39014 on ceramic capacitors.

Dc Bias: Can be introduced from external source.

Display: Two 5-digit banks of bright-light, numerical indicators with decimal point and units of measurement. Lamp burnout does not affect instrument operation or coded output. Lamps can be replaced from front panel.

Remote Control: Start and balance controls can be activated remotely by contact closures.

OUTPUT

Numerical Data: 10 digits BCD 1-2-4-2 code.

Print Command (at completion of balance): Change from "1" level to "0" level.

Signal Levels: "1" level, 0 V; "0" level, −12 V; both with respect to reference line at +6V above chassis ground. Impedance of lines 12 kΩ.

Measurement Rate: Panel control allows adjustment of measurement rate so that display time between measurements is between approx 0.1 and 5 s. The rate can be set manually or remotely at any rate compatible with balance time.

Other Measurement Frequencies: With internal modification, the measurement frequencies can be changed to any value between 100 Hz and 2 kHz.

GENERAL

Power Required: 105 to 125, 195 to 235, or 210 to 250 V, 50 to 60 Hz, 100 W. Internal 120-Hz oscillator is locked to power line for 60-Hz operation.

Auxiliary Controls: Sensitivity control on front panel can be used to minimize balance time with a resulting decrease in accuracy. Self start (when component is connected) or ext start (by contact closure) can be selected with a rear-panel switch.

Accessories Supplied: Rack-mounting hardware with rack models; power cord and spare fuses with all models.

Accessories Available: 1680-P1 Test Fixture; R, L, and C standards and decade boxes; various GR digital-data-acquisition instruments and system components.

Mounting: Supplied with hardware for rack mounting or assembled in cabinet for bench use.

Dimensions (width × height × depth): Bench, 19½ × 12 × 19 in. (495 × 305 × 485 mm); rack, 19 × 10½ × 18 in. (485 × 270 × 460 mm).

Net Weight: Bench, 76 lb (35 kg); rack, 71 lb (33 kg).

Shipping Weight: Bench, 160 lb (74 kg); rack, 145 lb (67 kg).

Catalog Number	*Description*	*Price in USA*
	1681 Automatic Impedance Comparator System	
1681-9700	115 V, 60 Hz, Bench	**$4975.00**
1681-9701	115 V, 60 Hz, Rack	**4975.00**
1681-9702	115 V, 50 Hz, Bench	**on request**
1681-9703	115 V, 50 Hz, Rack	**on request**
1681-9704	220 V, 50 Hz, Bench	**on request**
1681-9705	220 V, 50 Hz, Rack	**on request**
1681-9706	230 V, 50 Hz, Bench	**on request**
1681-9707	230 V, 50 Hz, Rack	**on request**
1680-9601	**1680-P1 Test Fixture**	**95.00**

U.S. Patent Applied For.

NEW WIDE-RANGE RF SOURCES

Several recent additions to the General Radio line of wide-range, general-purpose laboratory rf power sources should enhance an already excellent reputation for performance, versatility and dependability at a reasonable price. Two new vhf and uhf oscillators provide increased frequency coverage and improved modulation capability. Three new models of the power supplies offer regulated dc heater voltage for improved oscillator stability. The new oscillators and power supplies are packaged for quick, easy installation and use together, whether on the bench or in a relay rack.

The new oscillators have the low-noise sideband level essential in the local oscillator of a simple superheterodyne receiver using a wide-band single-sideband mixer. In the Type 1241 Heterodyne Detector (see page 24), the 1236 I-F Amplifier and the 874-MRAL Mixer are used with these oscillators to create a precision calibrated receiver. Typical sensitivity is -100 dBm for a 3-dB meter deflection over residual noise with a 0.5-MHz bandwidth. The oscillators achieve both low noise and complete freedom from nonharmonic discrete spurious frequencies in their outputs through the use of high-Q tank circuits operated at high level in a fundamental-frequency mode.

56 to 500 MHz in One Band

The 1363 oscillator delivers power typically in excess of 150 mW from 56 to 500 MHz (see Figure 1a) and replaces the popular Type 1208-C. As a local oscillator in the 1241 Heterodyne Detector, it provides fundamental mixing

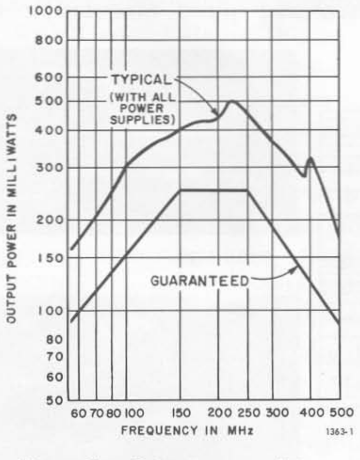


Figure 1a. Output power into a 50-ohm load for Type 1363 Oscillator.

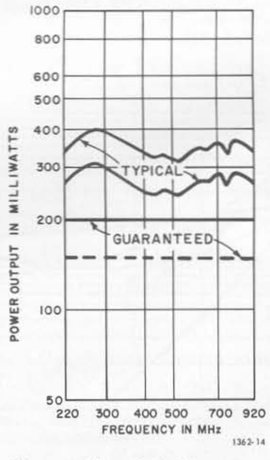
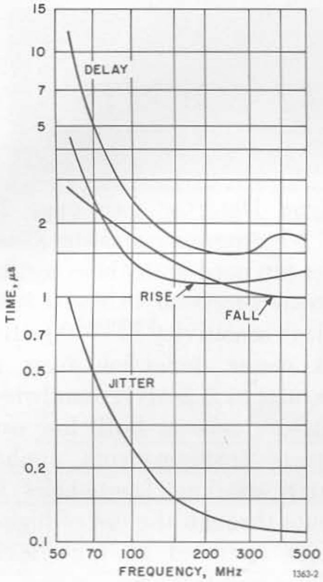
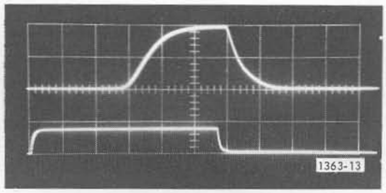
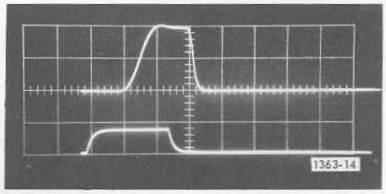


Figure 1b. Output power into a 50-ohm load for Type 1362 Oscillator.





60 MHz 2 μs/div



500 MHz 1 μs/div

Figure 2a. Typical rise time, starting delay, and jitter when the 1363 Oscillator is pulsed by the 1264 Modulating Power Supply, driven by a 1217-C Pulse Generator. Oscillograms show modulating and modulated pulse shapes at frequency extremes of the 1363.

for signal frequencies from 40 to 530 MHz. The basic wide-range tuner[1] consists of a variable inductor and a variable tuning capacitor, constructed as an integral unit. This fundamental-frequency LC oscillator circuit is inherently more stable than RC or beat-frequency circuits. In the new oscillator, we have increased the tuning range while reducing the number of wiping contacts from two to one by using a fixed network to suppress the unwanted resonance in the unused portion of the tuning inductor.

Other important circuit changes ensure compatibility with the 1264 power supply, making possible both square-wave and pulse modulation (Figure

[1] E. Karplus, "VHF and UHF Unit Oscillators," *General Radio Experimenter*, May 1950.

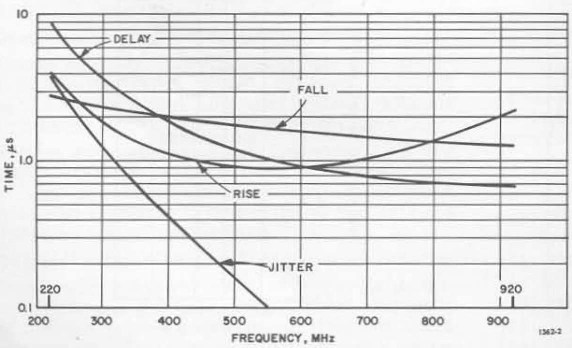
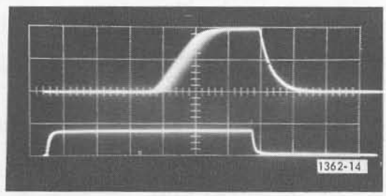
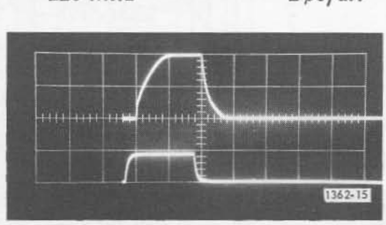


Figure 2b. Pulse characteristics of the 1362 Oscillator. Oscillograms showing modulating and modulated signals are at frequency extremes of the 1362.



220 MHz 2 μs/div



900 MHz 2 μs/div

2a). A front-panel output control is provided, and the rf output connector can be installed either on the front or at the rear of the instrument; the user can change the location in a few moments without any special tools. The GR874® output connector can be easily converted by means of GR874 adaptors to any popular coaxial connector series (BNC, C, N, TNC, OSM/BRM, Microdot, etc).

220 to 920 MHz in One Band

The 1362 Oscillator, with an output power typically in excess of 250 mW from 220 to 920 MHz (Figure 1b), supersedes two widely used oscillators, the 1209-CL and the TYPE 1209-C. The frequency range of the new oscillator includes the entire uhf aircraft communications band (220–406 MHz) and the uhf TV band (470–890 MHz), with margin to spare at the top end.* The tuner is a noncontacting butterfly similar to that used in the earlier oscillators.[1,2]

The oscillator tube is the new planar triode Type Y-1266 developed by General Electric Company in close collaboration with General Radio (Figure 3). This tiny ceramic tube has both the low interelectrode capacitances required for wide tuning range and the stable cathode of high emission capability required for high power output. The cathode operates at a moderate temperature, ensuring long, trouble-free life. This tube has demonstrated its excellence in hundreds of recent production 1209 oscillators and in the high-performance TYPE 1026 Standard-Signal Generator.[3]

The output system is a waveguide-below-cutoff piston, calibrated over a range of 80 dB and adjustable from the front panel. As in the 1361 Oscillator,[4] it is keyed against rotation and can readily be reset to a previously determined position. Relocation of the output coupling loop relative to the butterfly and the use of aperiodic damping to suppress an interdigital rotor resonance result in minimum harmonic content and a very smooth output-versus-frequency characteristic at any setting of the output attenuator.

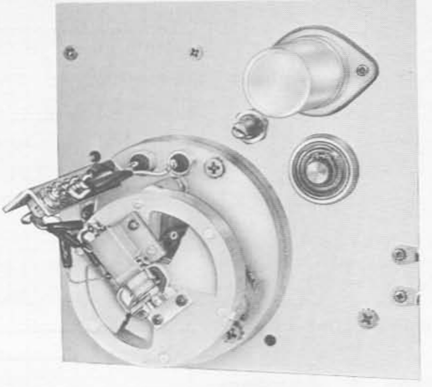


Figure 3. Interior view of the 1362 Oscillator showing new GE Y-1266 planar triode.

Leveled operation over the entire oscillator tuning range with a single setting of the output attenuator can be achieved by means of the 1263-C Amplitude Regulating Power Supply. This combination delivers 20 mW into 50 ohms (+13 dBm), either peak, with 1-kHz square-wave modulation, or cw. The level can be reduced as much as 20 dB if desired. Leveled performance is shown in Figure 4. Alternatively, new circuitry permits direct connection to

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[1] *Ibid.*
[2] E. Karplus, "The Butterfly Circuit," *General Radio Experimenter*, October 1944.
[3] G. P. McCouch, "A New 500-MHz Standard-Signal Generator," *General Radio Experimenter*, March 1967.
[4] G. P. McCouch, "A New UHF Signal Source," *General Radio Experimenter*, March 1961.
* The region below 220 MHz is covered by both the TYPE 1215 Oscillator (50-250 MHz, noncontacting tuner) and the new 1363. The region above 920 MHz is covered by the 1361 (450-1050 MHz) and the 1218 (900-2000 MHz).

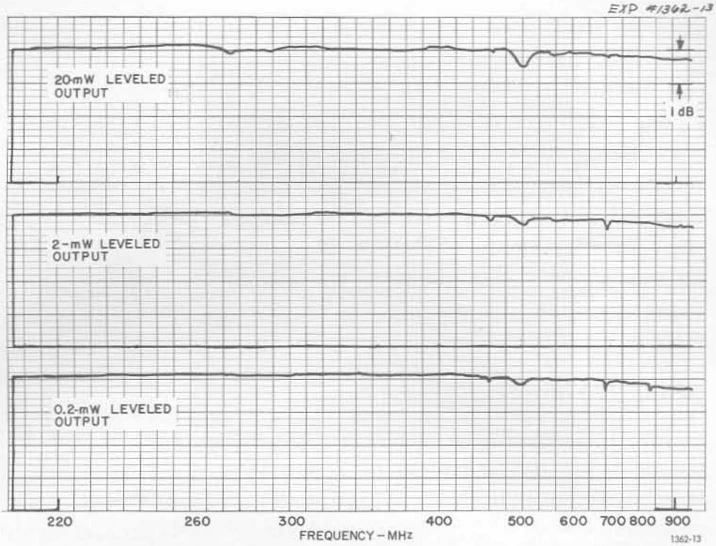


Figure 4. Recordings showing leveled performance of a 1362 Oscillator powered by a 1263-C Amplitude-Regulating Power Supply.

the 1264-B Modulating Power Supply, making possible both square-wave and pulse modulation (Figure 2b). Full power output is delivered during the "on" period and the oscillator is completely cut off during the "off" period.

Improved Stability from New Power Supplies

The new power supplies offer a major improvement in frequency and amplitude stability as well as increased tube life, obtained by close regulation of heater as well as plate supply voltage. The advantages afforded by regulation of both supplies have been clearly established by some years' experience with the 1267-A Power Supply. Well regulated dc heater supplies have now been incorporated in the 1263-C Amplitude Regulating Power Supply and in the 1264-B Modulating Power Supply.

Recent redesign of the 1267 to a "B" model permitted us to maintain the excellent specifications of its predecessor while simplifying the regulators and introducing a dual primary power transformer so that a single model now operates on either 115- or 230-V lines. In all three power supplies, the heater regulators are set to deliver 6.5 volts, thereby allowing 0.2 volt for the drop that occurs in the heater rf filters in the oscillators.

An important feature of the TYPE 1264-B Modulating Power Supply is the internal 1-kHz square-wave generator. A sample of the 1-kHz signal has been brought out to the modulation terminals for use in synchronizing oscilloscope sweeps; conversely, a synchronizing signal from an external oscillator may be injected here.

In the latest model of the 1264, it is much easier to set the 1-kHz frequency to the exact center of the narrow passband of a highly selective detector amplifier, and there is an order-of-magnitude improvement in the stability of the frequency once it is set.

The improved settability of the 1-kHz frequency has been achieved by means of a dual potentiometer with controlled backlash, operated from a single knob. The procedure is to tune with slight overshoot, then, as the con-

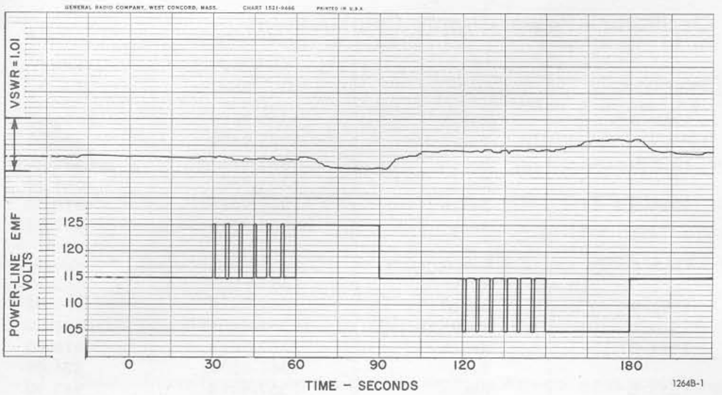


Figure 5. Stability of a 1218-B Oscillator — 1264-B Power Supply Combination used in a slotted-line recording system.

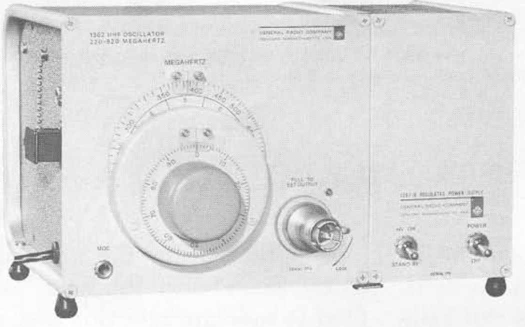
trol is backed up, only the vernier potentiometer comes into play over an arc of 40 degrees. This single control is far easier to use than is the conventional dual concentric knob coarse/fine combination.

The stability against line voltage of a 1264-B used to power a 1218-B Oscillator in an expanded-scale SWR-measuring system[5] is shown in Figure 5. The 1264-B is also ideal for use with other General Radio high-frequency oscillators as a source for conventional slotted-line measurements using the new high-stability 1234 Standing-Wave Meter.[6]

— G. P. McCouch

A brief biography of Mr. McCouch appeared in the March 1967 issue of the Experimenter.

[5] A. E. Sanderson, "A Slotted Line Recorder System," *General Radio Experimenter*, January 1965.
[6] M. Khazam, "A High-Resolution SWR Meter", *General Radio Experimenter*, February 1968.



Type 1362 UHF Oscillator with Type 1267-B Regulated Power Supply.

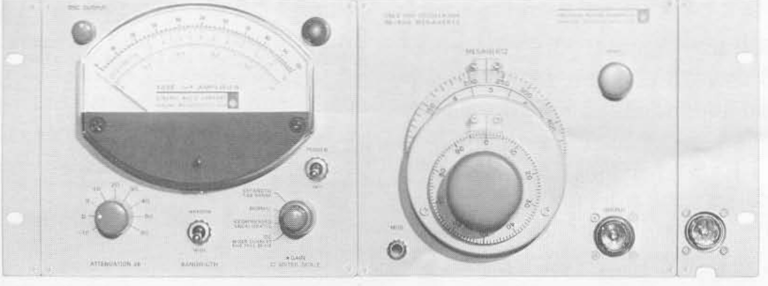
Complete specifications for the instruments described in this article are given in General Radio Catalog T

Catalog Number	*Description*	*Price in USA*
1362-9701	**1362 UHF Oscillator**	**$395.00**
1363-9701	**1363 VHF Oscillator**	**395.00**
1263-9703	**1263-C Amplitude-Regulating Power Supply**	**550.00**
1264-9702	**1264-B Modulating Power Supply** (115 V)	**415.00**
1264-9703	**1264-B Modulating Power Supply** (230 V)	**415.00**
1267-9702	**1267-B Regulated Power Supply**	**195.00**

Oscillator—Power-Supply Combinations

Oscillator	*Power Supply*	*Mounting*	*Catalog Number*	*Price in USA*
1363	**1269-A***	Bench	1363-9419	**$490.00**
		Rack	1363-9509	**515.00**
(56-500 MHz)	**1267-B**	Bench	1363-9417	**590.00**
		Rack	1363-9507	**615.00**
	1264-B	Bench	1363-9414	**810.00**
		Rack	1363-9504	**836.00**
1362	**1269-A***	Bench	1362-9419	**490.00**
		Rack	1362-9509	**515.00**
(220-920 MHz)	**1267-B**	Bench	1362-9417	**590.00**
		Rack	1362-9507	**615.00**
	1264-B	Bench	1362-9414	**810.00**
		Rack	1362-9504	**836.00**
	1263-C	Bench	1362-9413	**945.00**
		Rack	1362-9503	**971.00**

* See August 1963 *Experimenter.*



Type 1363 UHF Oscillator with Type 1236 I-F Amplifier, principal components of a Type 1241 Heterodyne Detector.

NEW HETERODYNE DETECTOR

GR's Type 1236 I-F Amplifier,[1] Type 874-MRAL Mixer,[2] and the oscillators introduced elsewhere in this issue constitute the main elements of a highly sensitive high-frequency heterodyne detector for relative-signal-level measurements and for use as a null detector. We now offer the entire package, including, in addition to the above, a 10-dB pad, a 90° ell, and an appropriate low-pass filter. The assembly is available as the Type 1241 Heterodyne Detector.

Applications for the heterodyne detector are almost limitless. It can be used to measure insertion loss, attenuation, crosstalk, antenna gain, and radiation patterns. It is, of course, a sensitive high-frequency receiver. When calibrated at one signal level and frequency, it can be used at that frequency as a selective voltmeter in a 50-ohm

[1] M. Khazam, "A New 30-MHz Amplifier with Two Bandwidths," *General Radio Experimenter*, July–August 1967.

[2] *General Radio Experimenter*, July–August 1967, page 19.

system. It is now the recommended null detector for the 1602-B UHF Admittance Meter, the 1609 Precision UHF Bridge, and the 1607-A Transfer-Function and Immittance Bridge.

As an SWR indicator with a slotted line, it is especially useful for measurements on nonlinear elements, when a high degree of harmonic rejection and a small applied signal level are required.

The price table indicates the fundamental-frequency coverage of the three basic assemblies. These ranges can be extended through the use of oscillator harmonics, but with reduced sensitivity and dynamic range. To cover a very wide frequency range, one might order one complete detector plus the necessary oscillators and filters for the additional ranges desired.

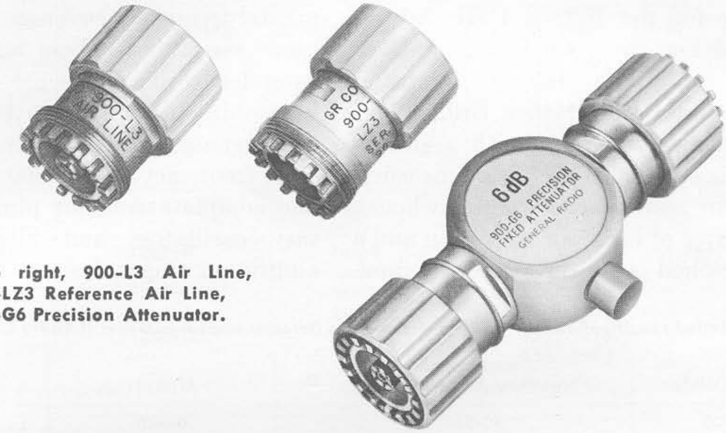
Detailed specifications on the 1241 Heterodyne Detector appear in General Radio Catalog T.

Catalog Number	*Fundamental Frequency Range — MHz*	*Mounting*	*Price in USA*
1241-9700	40-530	Bench	**$1270.00**
1241-9701	40-530	Rack	**1295.00**
1241-9702	190-950	Bench	**1265.00**
1241-9703	190-950	Rack	**1295.00**
1241-9704	870-2030	Bench	**1565.00**
1241-9705	870-2030	Rack	**1610.00**

IMPROVED NBS CALIBRATION ACCURACY FOR COAXIAL IMPEDANCE (1-8 GHz)

Accuracies of impedance measurements in coaxial-line systems have been improved significantly in recent years by the Radio Standards Laboratory (Boulder, Colorado), of the NBS Institute for Basic Standards (U.S. Department of Commerce), and others. This improvement came primarily from the development of precision coaxial-line standards and precision coaxial connectors, such as the GR900® series. According to an NBS release, "these developments have in turn contributed toward improving measurement capabilities of coaxial slotted-line systems to the extent that very accurate measurements are now possible. Errors originally introduced by structural defects of slotted lines have been minimized by the use of precision made, coaxial slotted lines. Refinements in measurement technique have helped, in part, to reduce some systematic errors.

"These improvements, along with other good practices, make the measurement possible of VSWR (Voltage Standing Wave Ratio) up to 8 GHz with an uncertainty in the range of 0.1 to 1 percent. The phase of the reflection-coefficient magnitude can be measured with an uncertainty ranging from 0.1 degree to approximately 1 degree. These uncertainties apply for coaxial impedance standards equipped with the 14-mm precision coaxial connector, where $1 \leq \text{VSWR} \leq 2$, referred to 50 ohms."



Left to right, 900-L3 Air Line, 900-LZ3 Reference Air Line, 900-G6 Precision Attenuator.

NEW GR900® ATTENUATOR, AIR LINES

The usefulness of any connector type depends to a large extent on the number of different things it can connect to, either directly or through adaptors. In this respect, the GR900® precision coaxial connector measures up very well. General Radio alone catalogs over fifty GR900 components, and the basic GR900 connector is used on many devices made and sold by other manufacturers. A full line of GR900 adaptors provides access to all other popular coaxial connectors.

The latest additions to the fast-growing GR900 line are a precision-fixed attenuator and two 3-cm air-line sections.

Attenuator

The accuracy of many microwave measurements (e.g., impedance, attenuation, phase, and power measurements) depends on the impedance match of the generator and detector. Attenuators and pads are commonly used for matching purposes, but the SWR of these devices is usually high enough to limit the accuracy improvement obtainable. In precision applications, the only recourse has been the use of matching tuners and slow, point-by-point measurements. The new GR900 attenuator, with an SWR of less than $1.005 + 0.005\ f_{GHz}$, eliminates the need for tuners in most applications and makes possible accurate swept-frequency measurements.

Attenuation of the 900-G6 is within 0.2 dB of its nominal 6-dB value to 5 GHz and within 0.3 dB of nominal to 8.5 GHz. Since it is equipped with GR900 connectors, it can be accurately calibrated for use as a secondary standard of attenuation.

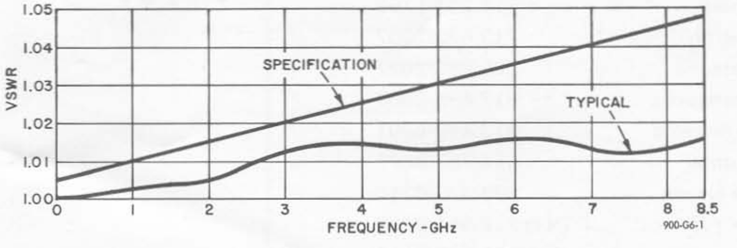
Air Lines

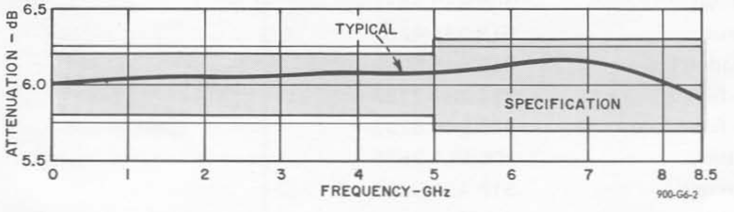
The 900-L3 Air Line and the 900-LZ3 Reference Air Line have been added to the existing series of air lines primarily to extend the usefulness of the series in precise capacitance calibrations. Both units are 2-pF two-port elements.

The chief distinction between the 900-L air lines and the 900-LZ reference air lines is one of precision. The reference air lines contain no dielectric material to support the inner conductor, which must therefore be supported from the connectors to which the line is joined. This no-dielectric-support design affords the ultimate in accuracy but obviously requires more time and care in installation. The regular, 900-L air lines include dielectric supports, which make them easier to use but which also introduce uncertainties. (Simple, absolute calibration of capacitance purely on the basis of dimension is not possible, for instance.)

The characteristics of the two types of air lines can be used to complement one another. Where two reference air lines are to be used in series, some means of supporting the inner conductors must be inserted between the two lines. The short 900-L3, with its own dielectric-supported inner conductor, will serve ideally as a minimal-capacitance coupling between two reference air lines.

SPECIFICATIONS FOR TYPE 900-G6





Frequency Range: 0 to 8.5 GHz.
Attenuation: 6.00 ± 0.2 dB, 0 to 5 GHz; ±0.3 dB, 5 to 8.5 GHz.
SWR: $<1.005 + 0.005\, f_{GHZ}$.
Characteristic Impedance: 50.0 Ω.
Insertion-Loss Repeatability: ±0.001 dB to 30 MHz, ±0.002 dB to 1 GHz; ±0.0025 dB to 8.5 GHz per connector.
Dc Resistance: 50.0 Ω ± 0.3% when terminated in 50.0 Ω.
Max Power: 1.0 W continuous; peak, 500 W with 1-W avg.
Temperature Coefficient: <0.0001 dB/°C/dB.
Dimensions: 3¾ × 1¾ × 1¹⁄₁₆ in. (95 × 45 × 27 mm).
Net Weight: 11 oz (310 g).

Detailed Specifications on the 900-L3 and -LZ3 Air Lines appear in General Radio Catalog T

Catalog Number	*Description*	*Price in USA*
0900-9850	**900-G6 Precision Fixed Attenuator (6 dB)**	**$175.00**
0900-9608	**900-L3 Precision Air Line**	**63.00**
0900-9603	**900-LZ3 Reference Air Line**	**76.00**

GENERAL RADIO COMPANY

WEST CONCORD, MASSACHUSETTS 01781

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the Experimenter

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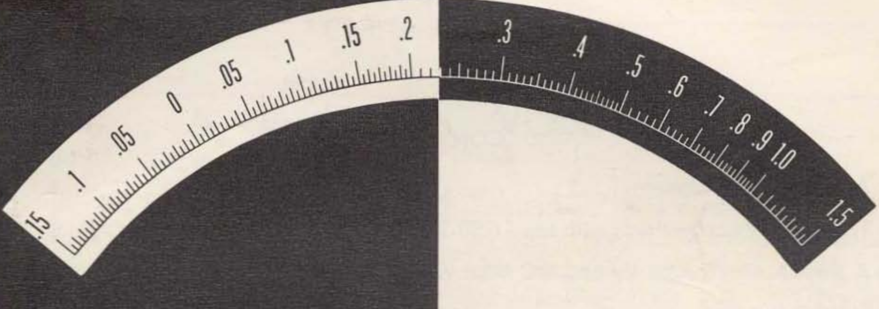


■ MEASUREMENT OF TRANSISTOR h-PARAMETERS WITH UNIVERSAL IMPEDANCE BRIDGE

■ NEW MICROVOLTER

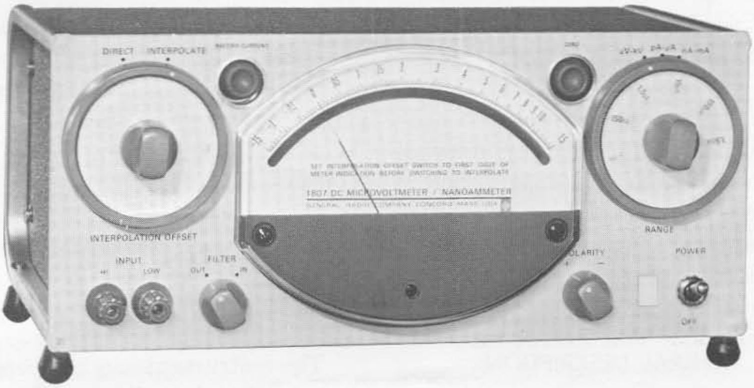
VOLUME 42 · NUMBERS 8, 9 / AUG – SEPT 1968 the GR Experimenter

Volume 42 • Nos. 8, 9 August-September 1968

Published monthly by the General Radio Company

CONTENTS

The *General Radio Experimenter* is mailed each month without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.



A UNIQUE DC VOLTMETER

Among the instruments that can be classified as basic or essential to any electrical or electronics laboratory or production test facility is the dc voltmeter. Yet, with the state of the art in the electronic industry demanding increasingly specialized instruments, it is difficult to find one "basic" voltmeter to meet the needs of even several projects in a given research and development center.

General Radio's new Type 1807 DC Microvoltmeter/Nanoammeter was designed to fill this "versatility gap." It combines the features of microvoltmeter, nanoammeter, null detector, and differential voltmeter, all with 0.2% accuracy. As a microvoltmeter it offers nine decade ranges, from 15 μV (with resolution of 0.05 μV) to 1500 volts full-scale. An input filter is also provided for noise suppression.

As a nanoammeter it can read currents from 15 pA (0.05-pA resolution) to 10 mA full-scale.

As a null detector it has a common-mode rejection ratio of greater than 160 dB and a three-second recovery time for a 10^6 overload.

As a differential voltmeter it offers accuracy 10 times better than that of conventional voltmeters.

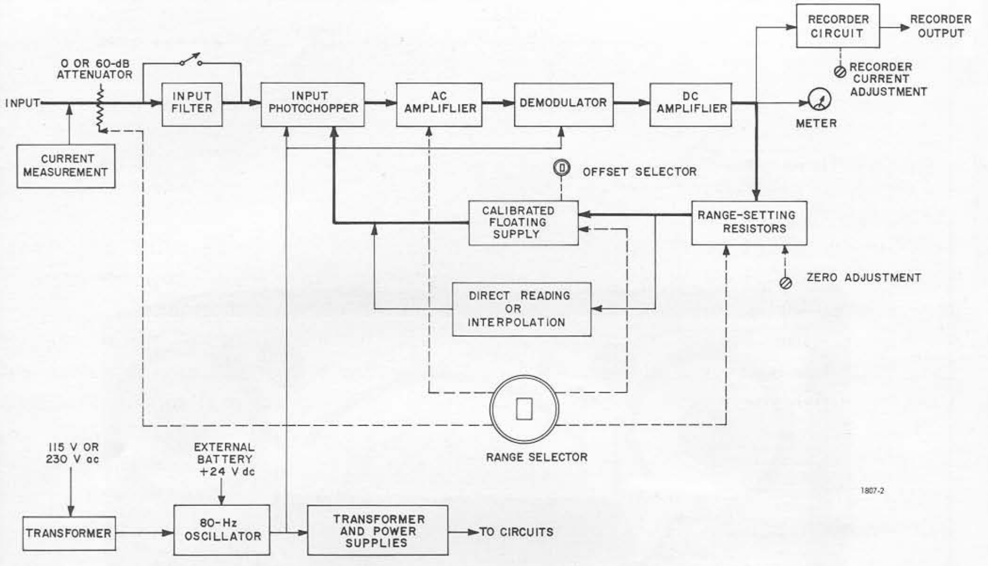


Figure 1. Block diagram of Type 1807 DC Microvoltmeter/Nanoammeter.

GENERAL DESCRIPTION

Figure 1 shows the block diagram of the 1807. What may not be obvious from this diagram is the method of achieving such a high common-mode rejection ratio. The key is Teflon* insulation to ground at all connections to the high and low terminals. For example, the meter is completely isolated by Teflon from the front panel. Even in the power transformer, Teflon is used as the insulation between windings.

The instrument can be operated from a 115-220-volt ac line or from a 24-volt dc supply.

Meter and Output Circuits

An unusual meter scale (see Figure 2) helps to improve the accuracy of reading and to provide higher resolution for null detection. Note that the meter is logarithmic above 10% of full scale and linear below 10% of full scale. The zero

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* Registered trademark of E. I. du Pont de Nemours and Company.

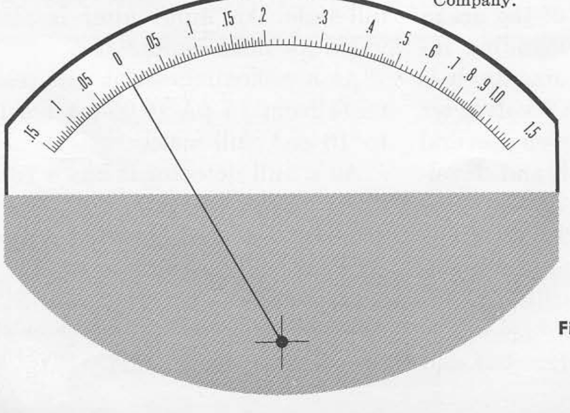


Figure 2. Meter scale of the 1807.

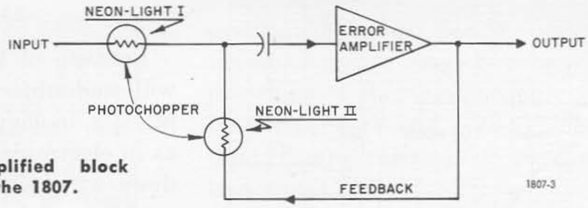


Figure 3. Simplified block diagram of the 1807.

is offset by about 20 degrees to allow the user to take readings about zero and still reserve most of the meter movement for higher resolution. A front-panel polarity switch enables the user to utilize the proper region of the meter.

The dc output amplifier can supply ±2.5 volts or ±1 mA, enough to drive most recorders. This voltage is adjustable at the front panel. Any load, even a short circuit at the recorder terminals, will have no effect on the operation of the instrument.

Interpolation-Offset Feature

The interpolation-offset feature allows the user to read the difference between two signals to within 0.1% of reading plus 0.1% of full-scale accuracy. The user, in setting the interpolation-offset switch, subtracts from the input a *calibrated* voltage equal to the most significant figure of the unknown; he then reads the difference in the conventional manner (see Figure 1). With this feature, the 1807 achieves accuracies usually associated with digital techniques, while preserving the versatility of an analog instrument.

Input Circuitry and Connectors

The high input impedance of the 1807, even down to the microvolt level, eliminates almost all loading errors. This impedance is achieved by means of the high loop gain in the error amplifier (see Figure 3) and the high impedance of the series-shunt cadmium-selenide (CdSe) photochoppers used at the input of the instrument.

The selection of a photochopper modulator was made in order to minimize noise, drift, and offset. Figure 4 shows a recording of the output noise of the 1807 in the picoampere range with no signal applied to the input. (There is a 1-megohm resistor across the input terminals on this range.)

One very annoying problem in the measurement of low-level dc is the presence of thermoelectric voltages generated when junctions of dissimilar metals are at different temperatures. Since it is difficult to keep all parts of the instrument at the same temperature, care was taken in the design of the 1807 to use copper-to-copper junctions at all points of the input circuitry. Thus, the input binding posts are gold-plated copper and the photocell leads

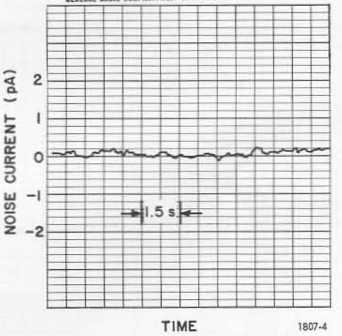


Figure 4. Recording of the output noise current referred to the input with 1 MΩ across the input terminals.

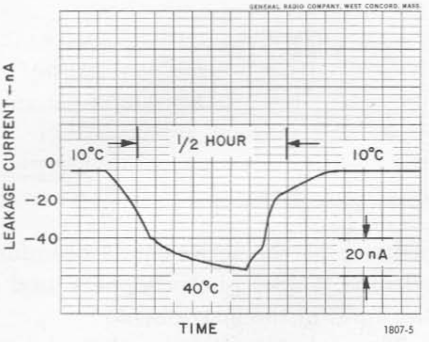


Figure 5. Recording of the leakage current for a Type 1N3604 diode cycled through a temperature change of 10°C to 40°C.

are Dumet, which has a thermoelectric voltage coefficient similar to that of copper.

The high input impedance of the instrument makes it feasible to use an RC low-pass filter with a cutoff frequency of 1.5 Hz at the input. This will filter out any ac noise that may be superposed on the dc signal. For faster response time, the filter can be switched out by means of a front-panel switch.

As an aid for routine voltage and current measurements, a Tektronix 1:1 probe (no attenuation) is available as an accessory. Note that this probe should not be used at extremely low voltage and current levels.

APPLICATIONS

Because of its versatility, the 1807 will undoubtedly find applications in physics, biology, and chemistry, as well as in electronics. Thus the applications discussed below should not be interpreted as being the prime uses of this instrument.

Diode or Transistor Matching

Quite often it is desirable to match a pair of diodes for leakage current or forward voltage drop (emitter-to-base voltage in case of transistors). The 1807 can be used to measure the forward voltage drop with a high degree of accuracy (using the interpolation feature) and the reverse leakage current with high sensitivity. Figure 5 shows a recording of leakage current for a type 1N3604 diode cycled through 10°C to 40°C.

Low-Level Differential Measurements

For comparison of various standard cells (e.g., saturated versus unsaturated), one is interested in making dc differential measurements of a few hundred microvolts with accuracies of about 1 or 2 μV. Figure 6 shows a recording of differential measurements

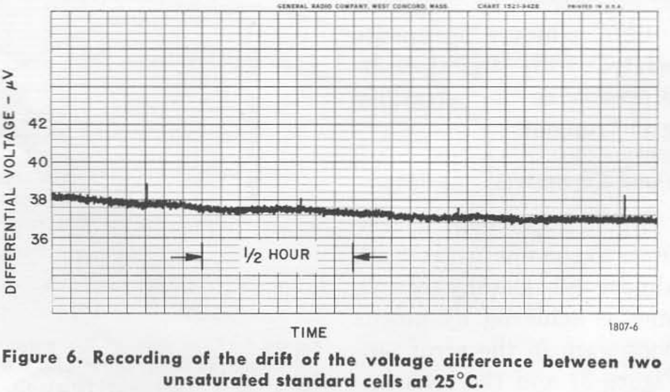


Figure 6. Recording of the drift of the voltage difference between two unsaturated standard cells at 25°C.

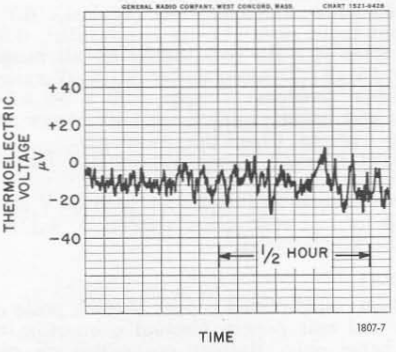


Figure 7. Recording of the thermoelectric voltage generated by a thermocouple.

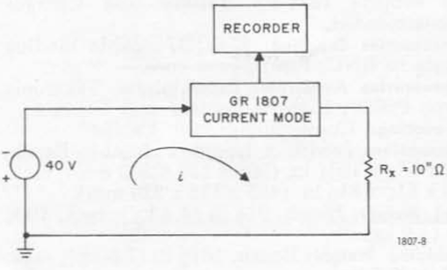


Figure 8. Test set-up for measuring unknown resistor R_x.

for two unsaturated standard cells. This recording, taken over a two-hour period, shows that the cells differed by 38.1 μV at the beginning of the test and by 36.8 μV at the end of two hours. The test temperature was 25°C.

Temperature Measurements

Figure 7 is a recording of the thermoelectric voltages generated by temperature difference between two points in a non-air-conditioned room. A copper-constantan thermocouple with a temperature coefficient of 39 μV/°C was used in this test. The recording, taken over a one-hour period, shows the thermoelectric voltage between the two junctions of the thermocouple to vary between $+8\ \mu$V and $-28\ \mu$V. Thus, the temperature variation is concluded to be about 0.9°C.

High-Value Resistance Measurements

The low-current sensitivity and the high accuracy of the 1807 make it suitable for very-high-resistance measurements. Figure 8 is a typical setup for such measurements. A nominal current i of 410 pA at 50°C was measured with the 1807, using the interpolation feature (Figure 9). Thus, the value of R_x at 50°C is calculated to be $9.75 \times 10^{10}\ \Omega$. As the temperature was dropped to 0°C, the current measured by the 1807 decreased to 376 pA. The resistance of R_x at 0°C, therefore, is $1.06 \times 10^{11}\ \Omega$.

— K. G. Balekdjian

A brief biography of Mr. Balekdjian appeared in the May 1968 issue of the *Experimenter*.

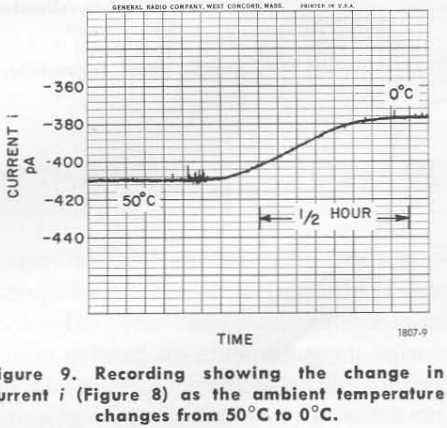


Figure 9. Recording showing the change in current *i* (Figure 8) as the ambient temperature changes from 50°C to 0°C.

SPECIFICATIONS

RANGE

Voltage (either polarity): 15 μV to 1500 V full scale in 9 decade ranges; 0.05-μV/div resolution near zero on most sensitive range.

Current (either polarity): 15 pA to 1.5 mA full scale in 9 decade ranges; 0.05-pA/div resolution near zero on most sensitive range.

ACCURACY

Record-Current Linearity: ±(0.1% of reading + 0.5 μV).

Interpolate: ±[0.1% of full scale (range) + 0.1% of reading + 0.5 μV].

Direct: ±(1.5% of reading + 0.5 μV) above 10% of full scale. ±(0.15% of full scale + 0.5 μV) below 10% of full scale.

Temperature Coefficients (typical)

Record-Current Zero Drift: ±(0.001% of full scale + 0.15 μV) per degree C.

Interpolate: ±(0.001% of reading + 0.001% of full scale + 0.15 μV) per degree C.

Direct: ±(0.02% of reading + 0.001% of full scale + 0.15 μV) per degree C.

INPUT IMPEDANCE

Voltage: 150-μV to 1.5-V ranges, > 500 MΩ on direct and typically 5,000 MΩ on interpolate; 15-μV range, > 50 MΩ; 15-V to 1500-V ranges, 10.5 MΩ.

Current: Internal Shunts, 1 MΩ in pA-μA ranges, 1 kΩ in nA-mA ranges.

Meter: Single scale from −1.5 to 15. Logarithmic (20 dB) above 10% of full scale.

Input Current: Less than 5 pA.

Noise: Typically 0.5 μV for 3σ with 1 kΩ across input.

Common-Mode Rejection: > 160 dB for dc with up to 600 V dc max above ground; > 120 dB for 60-Hz common-mode signal of <8 V pk with input filter.

Record-Current Response Time (typical): 0.1 s without input filter (1.5-Hz bandwidth), 0.3 s with filter (0.5-Hz bandwidth) on all ranges above 15 μV; 10 times slower on 15-μV range.

Maximum Overload: Voltage: 150 V on 1.5-V range and below, 1500 V on 15-V range and above. Current: 10 mA max all ranges.

Overload Recovery Time: Approx 3 s for 10^6 overload.

Recorder Output: Adjustable up to ± 2.5 V open circuit for full scale meter deflection; ± 1 mA into 1.5 kΩ max load.

GENERAL

Terminals: Gold-plated copper binding posts on front and rear panels. Ground connection on rear panel only. Battery connection on rear panel.

Power Required: 105 to 125, 205 to 250 V, 50 to 60 Hz, 5 W. Also operates from external 24-V dc supply; 1538-P3 Battery and Charger recommended.

Accessories Supplied: 274-QBJ adapts binding posts to BNC; fuse; power cord.

Accessories Available: Input probe, Tektronix type P6028; 1538-P3 Battery and Charger.

Mounting: Convertible-Bench Cabinet.

Dimensions (width x height x depth): Bench, 12 x 5⅞ x 10¼ in. (305 x 150 x 260 mm); rack, 19 x 5¼ x 8½ in. (485 x 135 x 220 mm).

Net Weight: Bench, 9½ lb (4.4 kg); rack, 10¾ lb (4.9 kg).

Shipping Weight: Bench, 16½ lb (7.5 kg); rack, 18 lb (8.5 kg).

Catalog Number	*Description*	*Price in USA*
	1807 DC Microvoltmeter/Nanoammeter	
1807-9700	Bench Model	**$665.00**
1807-9701	Rack Model	**665.00**
1807-9601	Input probe, Tektronix type P6028, Cat. No. 010-0074-00	**14.00**

TRANSISTOR MEASUREMENTS WITH THE 1650-B

As mentioned in the May *Experimenter*, the new TYPE 1650-B Impedance Bridge is useful not only for routine measurements on passive components but also for measurements of transistor *h*-parameters. The special test jig required for such measurements, as well as the procedure, is described below.

The *h*-parameters are defined in Figure 1, and the calculations that indicate how to obtain the four-terminal parameters h_{fe} and h_{re} from two-terminal measurements appear in Figure 2. Some thought about biasing the

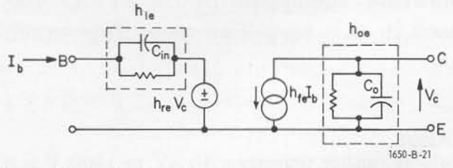
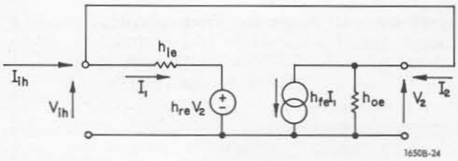
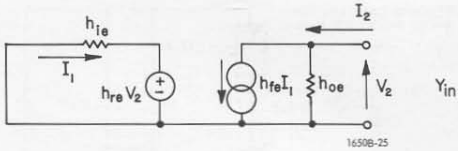


Figure 1. Equivalent circuit of a transistor, showing how the common-emitter h-parameters are defined.





$$I_{in} = I_1 + I_2 \qquad Z_{in} = \frac{V_{in}}{I_{in}}$$

$$V_2 = V_{in}$$

$$V_{in} = h_{ie} I_i + h_{re} V_{in}$$

$$I_2 = h_{fe} I_1 + h_{oe} V_{in}$$

$$\therefore I_1 = \frac{V_{in}\,(1 - h_{re})}{h_{ie}}$$

and $I_2 = h_{fe}\,\dfrac{V_{in}\,(1 - h_{re})}{h_{ie}} + h_{oe}\,V_{in}$

substitute and combine to get

$$Z_{in} = \frac{h_{ie}}{\underbrace{(1 - h_{re})}_{\approx 1}\,(1 + h_{fe}) + \underbrace{h_{oe}\,h_{ie}}_{\text{very small}}}$$

$$\boxed{\therefore\ h_{fe} \approx \frac{h_{ie}}{R_{in}}}$$

$$Y_{in} = \frac{I_2}{V_2}$$

$$V_1 = h_{ie} I_1 + h_{re} V_2 = 0$$

$$\therefore\ I_1 = \frac{-h_{re}}{h_{ie}}\ V_2$$

$$I_2 = h_{fe} I_1 + h_{oe} V_2$$

$$\therefore\ I_2 = \left(h_{oe} - h_{fe}\,\frac{h_{re}}{h_{ie}}\right) V_2$$

$$Y_{in} = \frac{h_{ie}\,h_{oe} - h_{fe}\,h_{re}}{h_{ie}}$$

$$h_{re} = \frac{h_{ie}\,h_{oe} - h_{ie}\ Y_{in}}{h_{fe}}$$

$$\text{but}\quad \frac{h_{ie}}{h_{fe}} \approx R_{in}$$

$$\boxed{\therefore\ h_{re} \approx R_{in}\,(h_{oe} - Y_{in})}$$

Figure 2. Derivations of the formulas for h_{fe} and h_{re}.

transistor so as to isolate the particular impedance we want to measure yields the test circuits of Figure 3. These may be built into a convenient test jig using rotary switches or toggle switches with shielded leads going to the 1650-B's unknown terminals. The shields should be connected to the 1650-B case to guard out the stray capacitance between the center conductors. It is helpful to have the 1650-B plastic-coated condensed instruction sheet handy during measurements so that R_A, the ratio-arm resistance, can be determined.

The procedure is as follows:

a. Use the R_{ac} bridge and the h_{ie} test jig (Figure 3a). Turn the OSC LEVEL control way down to ensure a small-signal measurement of the forward-biased base-to-emitter junction. The diffusion capacitance of the junction is balanced using an external capacitance

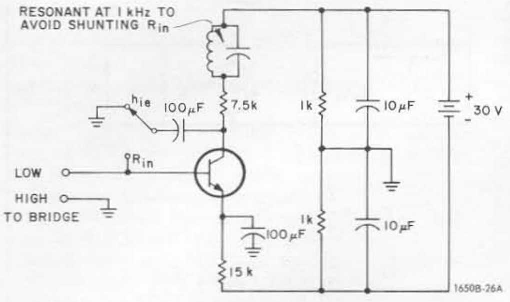


Figure 3a. The h_{ie} test jig.

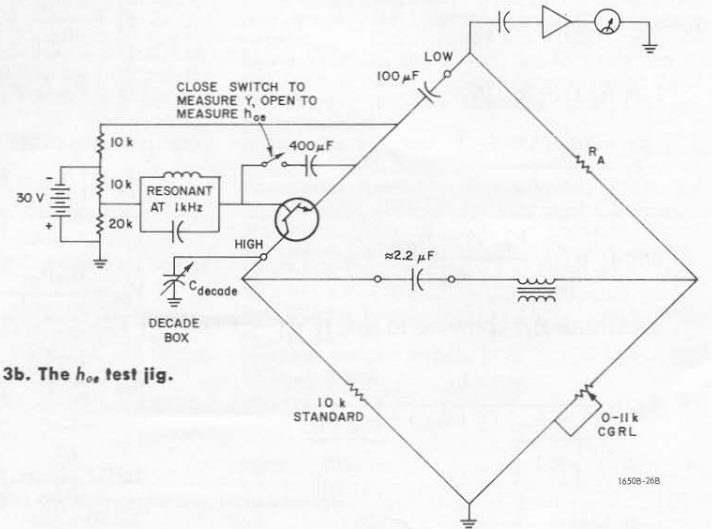


Figure 3b. The h_{oe} test jig.

decade box connected between the HIGH UNK terminal and the 1650-B ground. h_{ie} is usually a few thousand ohms and

$$C_{in} = \frac{10\ \text{k}\Omega}{R_A} C_{\text{decade}}\ ,$$

where R_A is the value of the ratio arm in kilohms (10 kΩ on the 1-kΩ multiplier range).

b. Throw a switch that connects the base to the collector with a large capacitor and measure R_{in}, which will be about 26 ohms for an emitter current of 1 mA. Calculate: $h_{fe} = \dfrac{h_{ie}}{R_{in}}$.

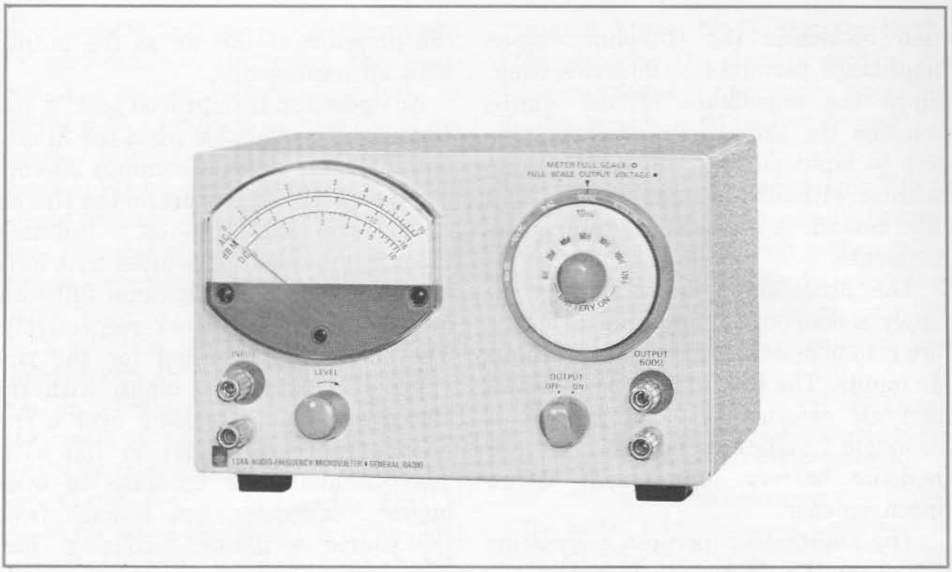
c. Use the G_{ac} bridge and the h_{oe} test jig (Figure 3b) and turn up the oscillator level for increased sensitivity Measure h_{oe} directly in micromhos and balance the output capacitance, C_o, with the external capacitance decade box. Calculate: $C_o = \dfrac{R_N}{R_A} C_{\text{decade}}$,

where R_N is the reading on the CGRL dial (0-11 k) in kilohms and R_A is the value of the ratio arm in kilohms (100 kΩ on the ×10-micromhos range).

d. Throw a switch that connects the base to the emitter with a large capacitor and measure Y in micromhos. Calculate: $h_{re} = R_{in}\,(h_{oe} - Y)$.

— C. HAVENER

A brief biography of Mr. Havener appeared in the May 1968 issue of the *Experimenter*.



A NEW AUDIO-FREQUENCY AND DC MICROVOLTER*

Just as in 1933 when the first General Radio Microvolter was designed, there is still a need for an independent metered attenuator to obtain dc and audio-frequency voltages at the microvolt level. Many modern sources of ac signals, particularly of nonsinusoidal ones, are not metered and do not have an attenuator with the range or the shielding required for working at microvolt potentials. Yet the need for such a low-voltage source is clear, especially with today's wide use of very high-gain operational-amplifier techniques.

A Microvolter is simply a self-contained step attenuator preceded by a meter and a continuously adjustable potentiometer. The meter and potentiometer establish the input voltage of the attenuator over a one-decade range, and this is then reduced to the desired level in decade steps by the attenuator. The range of the instrument is determined by the full-scale sensitivity of the meter and by the amount of attenuation.

The new 1346 Audio-Frequency Microvolter, the third generation to follow the original 546, has many new features, of which the two most important are a dc meter scale and an internal battery, making it a self-contained "floatable" dc source. Output is adjustable from 1 μV to 10 V with either polarity. Gold-plated copper binding posts and careful construction keep thermal emfs negligible even at the lowest microvolt levels. An on-off switch, which in the off posi-

* Trademark registered in USA.

tion maintains the 600-ohm output impedance, permits easy dc zero setting. Since the impedance of the source remains the same, any offset voltage due to input current will be the same with or without the signal applied and will not affect incremental gain measurements.

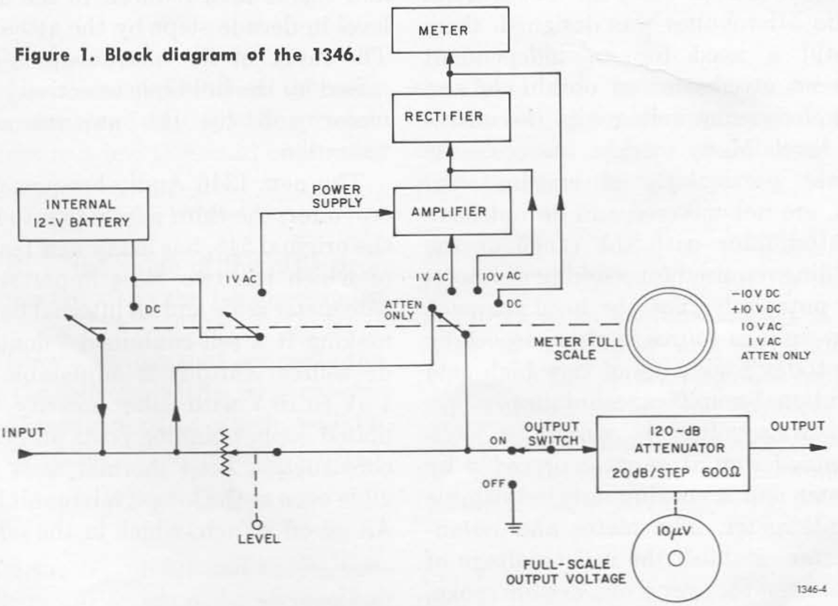
The floatability makes it easy to apply a common-mode signal to measure common-mode rejection for various dc inputs. The battery can be removed and an external dc supply used if complete isolation is not needed or if periodic battery replacement is an inconvenience.

The ability to superpose a small ac signal on the dc greatly increases the usefulness of the Microvolter. If one wants to know when a dc amplifier is operating in its linear region and when it is in saturation, he can simply superpose a small ac signal on the dc through the input binding posts and check for the presence of this ac at the output with an oscilloscope.

Ac operation is improved also. A full 10 volts ac is available when the Microvolter is used with the common 20-volt, 600-ohm audio oscillators on the 10-volt ac full-scale range. A 1-volt ac full-scale range is provided for sources that have lower outputs. The minimum full-scale outputs are 1 and 10 μV respectively. The meter is calibrated for the rms value of a sinusoidal input, with the stated accuracy extending over a frequency range of 10 Hz to 100 kHz. The attenuator is accurate to even higher frequencies, but leakage from the source oscillator, radiation from connectors, and ground-loop effects often exceed microvolt levels at frequencies above 100 kHz.

Distortion introduced in the signal of the Microvolter is reduced to a level consistent with that of typical sources with which it may be used. The output

Figure 1. Block diagram of the 1346.



on-off switch that can reduce the output to zero while maintaining the same 600-ohm output impedance is fully as useful with ac signals as with dc. Noise sources due to ac pickup and ground loops are much easier to measure and to locate with the signal removed and with the impedance level, shielding, and circuit configuration remaining the same.

In addition to positions for the one dc and two ac sensitivities, the meter has an ATTEN ONLY position in which the meter is out of the circuit, leaving just the step attenuator between the input and output terminals. This is desirable when the signal is of unusual waveshape, as for example, a tone burst.

Figure 1 is a block diagram of the 1346, showing the circuits used in the different modes of operation. Note that the battery is used in the dc and 1-V ac positions only; the meter is self-powered on the 10-V ac range. This is the range most often used, because of the profusion of 20-volt audio oscillators available, and the battery thus should normally have a very long life. The battery is an inexpensive and readily available type.

— R. E. OWEN

A brief biography of Mr. Owen appeared in the January 1968 issue of the *Experimenter*.

ACKNOWLEDGEMENT

The author acknowledges the assistance of Peter Young on the attenuator and Alan Ombrello on the meter designs.

SPECIFICATIONS

Function	*10 V ac*	*1 V ac*	*+10 V dc −10 V dc*	*Atten Only*
Open-Circuit Output Voltage	1.0 μV to 10 V ac	0.1 μV to 1.0 V ac	1.0 μV to 10 V dc	0 to −120 dB 20 dB/step
Accuracy at 23°C (above 10% of dc full scale)	±(4%+0.2μV) 10 Hz to 100 kHz	±(4%+0.02μV) 10 Hz to 100 kHz	±(3%+0.2 μV)	±(0.04 dB/step +154 dB below input level) dc to 100 kHz
Source	External ac required 10.0 V into 595 Ω	External ac required 1.0 V into 550 Ω	Internal battery or ext dc source 10 V max	Ext ac or dc source 10 V max input
Input Impedance (approx)*	595 Ω to 25 kΩ	550 Ω to 25 kΩ	610 Ω to 25 kΩ with int battery removed	550 Ω to 5 kΩ

* Input impedance varies as shown in table with setting of input level control. Can be adjusted to remain constant when varying the step attenuator for load impedance of ≥50 Ω.

Distortion (at 1 kHz): <0.01% in 1-V-ac mode, <0.05% in 10-V-ac mode, with level control at max setting.
Output Impedance: 600Ω ±0.5%.
Power Required: None required for 10-V-ac range. In other modes, 12-V dry battery: Eveready 228, RCA VS329, or Burgess PM8. Approx life, 33 hours at 2h/day in either dc mode, 316 hours at 2h/day in 1-V-ac mode.
Mounting: Convertible-Bench Cabinet.

Accessories Supplied: Battery; mounting hardware with rack model.
Accessories Available: GR 1309-A and 1310-A Oscillators, 1396-B Tone-Burst Generator, 1381 and 1382 Random-Noise Generators.
Dimensions (w x h x d): Bench, 8½ x 7½ x 7½ in. (220 x 190 x 190 mm); rack, 19 x 6 x 7⅝ in. (485 x 155 x 195 mm).
Weight: Net, 5¼ lb (2.4 kg); shipping, 9 lb (4.1 kg).

Catalog Number	*Description*	*Price in USA*
	1346 Audio-Frequency Microvolter	
1346-9700	**Bench Model**	**$250.00**
1346-9701	**Rack Model**	**272.00**

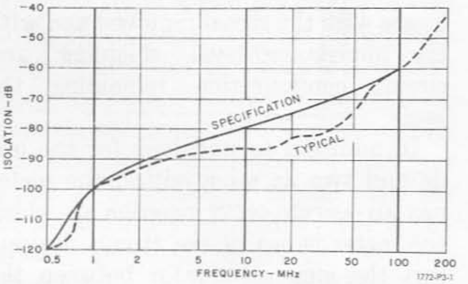
GR Product Notes

NEW SHIELDED SWITCH MODULE

The 1772-P3 module extends the capabilities of the 1770 Scanner System to the automatic scanning of signal lines by providing high isolation of signals up to 100 MHz. Isolation between lines is 120 dB at 500 kHz, 100 dB at 1 MHz and 60 dB at 100 MHz. Pulse reflections are typically less than 15%.

Other switch modules, the 1772-P1 and 1772-P2, accept input lines in a single shielded bundle; the new 1772-P3 accepts BNC-terminated cables. All three will switch 10 channels per module, thus providing up to 100 channels in a single 1770 Scanner System. The system has six scanning modes, with front-panel mode selection.

The 1772-P3 switch module contains ten dry-reed relay switches. Each relay requires 15 V at 30 mA to operate. The module is a special-order item, and the customer can specify the program boards, switching methods, and shielding that suit his application.



Isolation between channels of the 1772-P3 (determined by energizing one channel, applying a signal through it, and measuring the signal on any unenergized channel with a 50-Ω detector).

Typical of the many uses for the 1772-P3 is that of making automatic frequency-stability tests on many oscillators or signal generators. While the shielded modules connect the oscillator outputs in sequence to a single counter, the output data from the counter could also be switched, in order, to several recorders or to other logging devices through other modules.

PRECISION CAPACITOR

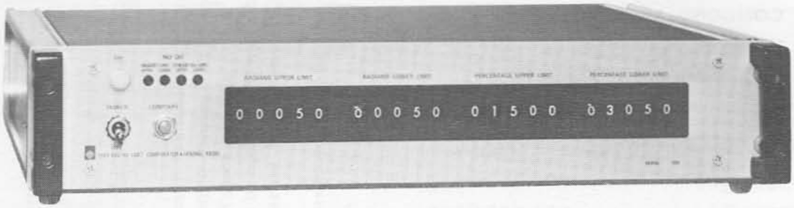
The 1422-CE is a stable and precise variable air capacitor intended for use as a continuously adjustable standard of capacitance. It is a dual-range, three-terminal capacitor with shielded coaxial terminals for use in three-terminal measurements. The calibrated direct capacitance is independent of terminal capacitance to ground and losses are very low.

This low-capacitance member of the 1422 family has ranges of 0.005 to 0.11 pF and 0.05 to 1.1 pF, resolution of 0.00002 and 0.0002 pF per division, respectively, and will satisfy the special requirements for use in the Harris ultralow-frequency capacitance bridge* built at the National Bureau of Standards.

Specifications are comparable to other models of the GR 1422 and are given in full in Catalog T.

* W. P. Harris, "A New Ultralow-Frequency Bridge for Dielectric Measurements," *1966 Annual Report, Conference on Electrical Insulation and Dielectric Phenomena*, page 72.

NEW DIGITAL LIMIT COMPARATOR FOR 1681 AUTOMATIC IMPEDANCE-COMPARATOR SYSTEMS



The 1783 Digital Limit Comparator is similar to the 1781 (an accessory to the 1680 Capacitance Bridge[1]), but it has expanded capabilities for use with the 1681 Automatic Impedance Comparator.[2]

When used with the 1681 comparator, the 1783 automatically makes comparisons between the measured digits and the limit digits and presents the results as a panel-lamp display and as relay-contact closures at a rear-panel connector. The 1783 has independent upper and lower 5-digit limits for both impedance magnitude and phase angle. The limits are set manually on rows of thumb-wheel switches.

A 1681 impedance comparator and one or more 1783 limit comparators form a complete testing system that allows an operator to sort components manually into precise categories. These instruments can also be used with automatic handling equipment.

The 1783 is assembled on special order. It is available either alone or in a system custom designed by General Radio for your application.

[1] *General Radio Experimenter*, December 1966.
[2] *General Radio Experimenter*, June-July 1968.

DC-LEVEL CONTROL FOR THE 1398 PULSE GENERATOR

A new accessory can be used to control the dc level of the output pulse of a 1398-A Pulse Generator. The controlled parameter may be average, positive peak, or negative peak, and control is independent of duty ratio and prf. Time constant of the regulating circuit can be selected according to output-pulse duration.

The 1398-P1 DC Component Control attaches to the side of, and derives power from, the pulse generator.

Catalog Number	*Description*	*Price in USA*
	Precision Capacitor	
1422-9833	1422-CE, with standard calibration	**$510.00**
1422-9580	1422-CEP, with precision calibration	**620.00**
	1783 Digital Limit Comparator	
1783-9801	Bench Model	**2500.00**
1783-9811	Rack Model	**2500.00**
1398-9601	**1398-P1 DC Component Control**	**185.00**
0480-9337	**480-P317 Rack-Adaptor Set** for combination of 1398-A and 1398-P1	**6.00**

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THE GENERAL RADIO

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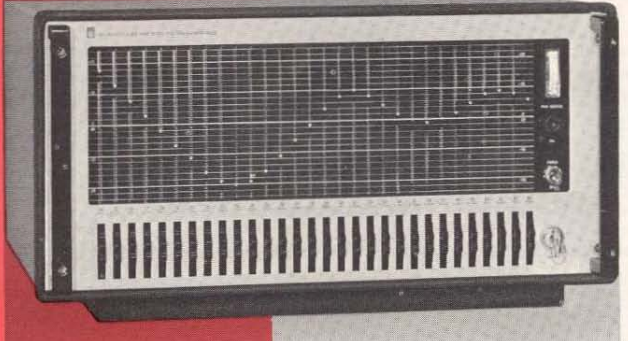
AN ELECTRONIC

MULTIFILTER



ALSO IN THIS ISSUE:

- HIGH DEGREE OF FREQUENCY RESOLUTION WITH A SYNCABLE OSCILLATOR
- 100:1 SCALER
- REDESIGNED TONE-BURST GENERATOR

VOLUME 42 · NUMBER 10 / OCTOBER 1968

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the GR Experimenter

Volume 42 • No. 10 October 1968

Published monthly by the General Radio Company

The *General Radio Experimenter* is mailed each month without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.

HIGH DEGREE OF FREQUENCY RESOLUTION ACHIEVED THROUGH USE OF SYNCABLE OSCILLATOR IN CLOSED LOOP

When Dr. P. E. Armstrong of the Los Alamos Scientific Laboratory wanted to explore the high-Q mechanical resonances of clamped samples of magnetic alloys, he achieved the needed frequency stability and resolution not with a frequency synthesizer but with one of GR's inexpensive "syncable" oscillators.

The experimental setup is shown in Figure 1. Vibrations are excited in the rod by a nearby coil driven through an amplifier by the Type 1310 Oscillator. The rod is slightly magnetized by an

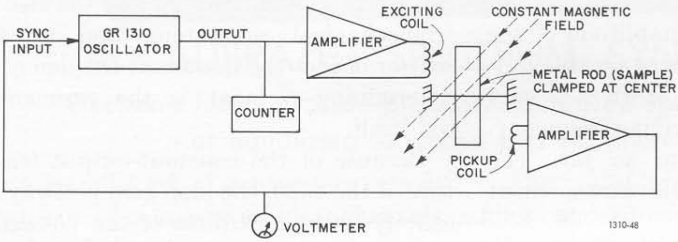
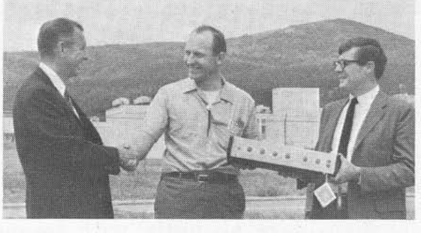


Figure 1. Block diagram of the experimental setup used by Dr. Armstrong to investigate the mechanical resonance of the clamped-magnetic rod.

WINNER OF GR'S "THINK SYNC" CONTEST



GR District Manager Frank Thoma (left) and Sales Engineer Dave McGreenery (right) present 1312 Decade Oscillator to Dr. Armstrong.

At the New York IEEE Show this year, General Radio announced a technical contest to discover novel applications of the synchronizing capability of GR's low-frequency oscillator line. The prize was choice of any of the five "syncable" oscillators.

We were impressed by the ingenuity with which many of the contestants exploited some not-altogether elementary aspects of the sync capability, and from our point of view the contest was an edifying success. The criteria for selection of a winner were originality and usefulness of the application. It was the final decision of our judges to award the prize to Dr. P. E. Armstrong, Los Alamos Scientific Laboratory, Los Alamos, New Mexico, whose application is described in the accompanying article. Our congratulations to Dr. Armstrong, who chose the Type 1312 Decade Oscillator as his prize.

Peter A. Previte received his BSEE degree in 1966 from Northeastern University. After graduation he joined General Radio's Sales-Promotion Department, where he is a product-line specialist for low-frequency oscillators. He is a member of Eta Kappa Nu.

externally applied constant magnetic field and therefore induces in the pick-up coil a signal that, in frequency, phase, and amplitude, is a faithful replica of the mechanical oscillation of the rod. The relative amplitude of the rod's vibration is indicated by the voltmeter, while the counter gives a very precise reading of the driving frequency.

Quite straightforward so far. The remarkable feature of the arrangement is immediately apparent from a glance at some typical resonance data shown in Figure 2. The bandwidth of the resonance is a few hertz at about 15 kHz. One does not expect to be able to achieve this kind of resolution with a general-purpose laboratory oscillator! The trick is closing the loop through the oscillator's synchronizing jack.

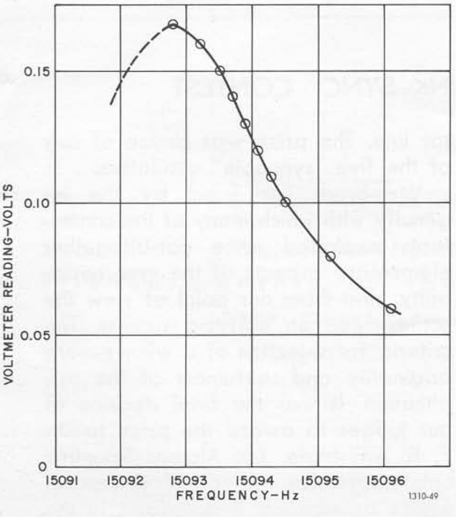


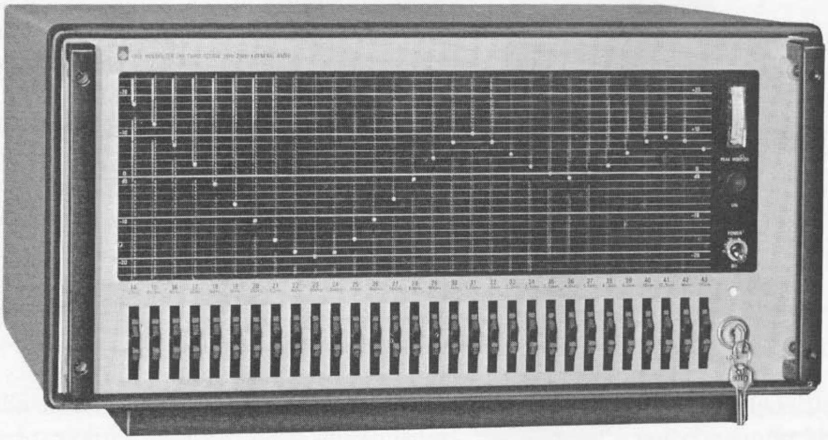
Figure 2. Plot of resonance data for a typical sample. The system turned out to be unstable below resonance, but data for this region could have been obtained, if necessary, by reversing connections somewhere in the loop.

Of course, in the absence of synchronization, the frequency at which the rod is driven would be determined simply by the setting of the 1310's frequency dial. But the oscillation frequency of the closed-loop system depends upon both the 1310 and the electromechanical subsystem, and as a matter of fact the dominant frequency-determining element is the resonant rod itself.

Because of the constant-output feature of the 1310, the loop gain is always exactly unity regardless of the voltage at the synchronizing jack. The system therefore oscillates at a frequency for which the total phase shift around the loop is zero. Now, the locked 1310 is effectively a variable phase shifter. With a sufficient voltage at the synchronizing jack, the phase shift through the 1310 changes only gradually with frequency and with the setting of the frequency dial. On the other hand the phase shift through the high-Q electromechanical part of the system changes very abruptly with frequency in the vicinity of the mechanical resonance. Thus the frequency of oscillation, which is determined by the zero-phase condition, is predominantly under the control of the clamped rod; the 1310's frequency dial serves as a frequency vernier.

— P. A. Previte

The new Type 1925 Multifilter. The model shown here covers the frequency range from 25 Hz to 20 kHz in ⅓-octave bands.



A CALIBRATED SPECTRUM SYNTHESIZER

Electronic filters provide precisely calibrated spectrum shaping or equalizing for sound and vibration work

A parallel set of contiguous octave-band or ⅓-octave-band filter channels, each including an adjustable attenuator — such a system, the basis for a number of spectrum-synthesis and spectrum-analysis techniques, is embodied in GR's new 1925 Multifilter.

The multifilter normally includes 30 filters and 30 calibrated, 1-dB-per-step attenuators, although it can be supplied with more or fewer channels, and the attenuators are optional. The octave- and ⅓-octave-filter characteristics (Figure 1) conform to both American

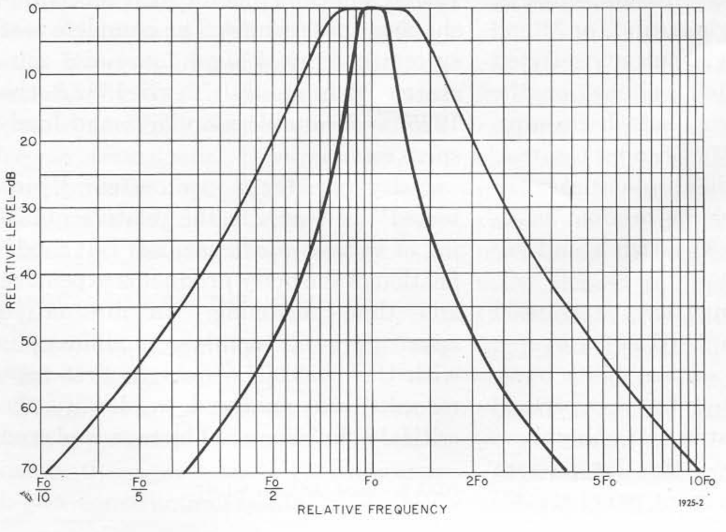


Figure 1. Octave and ⅓-octave filter characteristics.

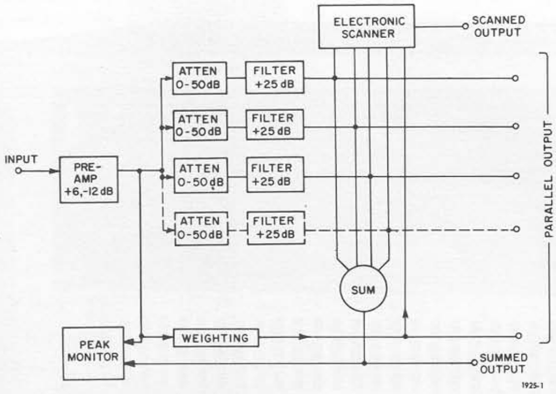


Figure 2. Block diagram of the multifilter.

and international standards. Both meet the IEC Recommendation Publication 225-1966; the third-octave characteristic meets the USA Standard S1.11-1966 Class III (high attenuation); the octave characteristic meets the same standard, Class II (moderate rate — highest for octave-band filters). Channel frequencies from 3.15 Hz to 80 kHz are available in standard models of the instrument.

The outputs from the filter channels are accessible in three different ways: 1) simultaneously at separate outputs, 2) summed at a single output, or 3) one at a time at a single output, selected either by push-button switches on the rear panel, by external switch closures, or by a GR Type 1771 Scanner Control. The 1771 allows either sequential scanning of the channels or random selection by BCD (1-2-4-8) coded standard[1] band numbers.

The channel attenuators are adjusted by thumbwheels in 1-dB steps over a 50-dB range. The setting of each attenuator is indicated by the vertical position of a dot that is part of a unique front-panel display — the dot pattern that appears on the front panel creates a graph of the instrument's over-all transmission characteristic. This graphic display has a vertical scale factor of 10 dB per inch and a horizontal frequency scale factor of 5 inches per decade, matching recorder chart paper commonly used in sound and vibration work.

APPLICATIONS

High-level-sound tests on such structures as airframes are carried out with especially shaped noise spectra. The 1925 is a flexible tool for such spectrum-shaping applications. The complete test system comprises a random-noise generator such as the Type 1382,[2] the 1925, and suitable amplifiers and loudspeakers.

Noisy products are often "jury tested" to compare the relative quieting of various modifications. But modification of the noisy product is expensive and time-consuming. An alternative approach is to simulate modifications with the 1925. The noise is first tape recorded and analyzed to identify the critical noise sources. The tape is played

[1] USA Standard (ASA) S1.6-1967.
[2] *GR Experimenter*, January, 1968.

back through the multifilter, which permits selective control over the amplitudes of individual noise components, and a listening jury is asked to rate the modified noise.

The 1925 can be used to simulate the transmission characteristics of walls, partitions, and other acoustic systems, thus saving the cost of constructing the barrier in order to test its effectiveness.

In broadcast- and recording-industry applications, the 1925 permits unusually flexible and precise control over recording and playback equalization, program-line equalization, and pre- and de-emphasis. As a filter, the 1925 not only offers greater versatility than passive filters or even tunable electronic filters, it also provides a calibrated display of its frequency characteristic.

With accessory equipment, the multifilter becomes a parallel or serial spectrum analyzer. Channels can be selected in sequence with a TYPE 1771 Scanner Control, or the parallel outputs of the 1925 can be used to drive a set of detectors and recorders. The adjustable attenuators are in principle unnecessary in the analyzer, and a model without them might be suitable in this application. However, inclusion of the attenuators makes it possible to compensate for frequency-response errors due to transducer, tape-recorder, or other system components and to extend the system's dynamic range by "prewhitening" the incoming signal.

COMPONENTS OF THE 1925

The block diagram of the multifilter is shown in Figure 2.

The filters are active, six-pole Butterworth types, mounted on plug-in etched boards, three per board. This modular plug-in construction results in a great deal of flexibility, allowing both easy conversion of a standard model of the instrument from one frequency range to another and straightforward assembly of special versions. Models with non-standard frequency ranges, with mixtures of octave and ⅓-octave channels, or even with special-bandwidth channels can be assembled on special order.

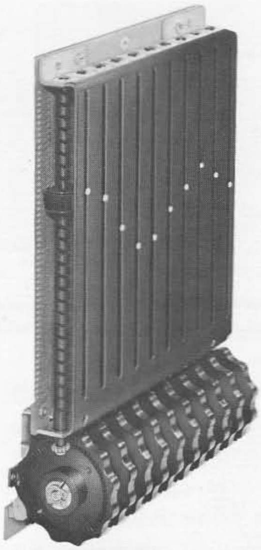


Figure 3. An assembly of ten of the adjustable attenuators. The setting of each attenuator is indicated by the vertical position of a dot in the front-panel display.

The channel attenuators are etched-circuit switches and integrated resistor circuits. An attenuator assembly, shown in Figure 3, has a single input and ten outputs to drive a block of ten filter channels. Three of these assemblies are used in a standard ⅓-octave-band version of the multifilter, and one is used in an octave-band version. The main chassis is built to house up to three attenuator assemblies, thirty filters, and input, output, and power-supply circuitry. Reed relays switch the output of each filter channel to an output connector.

Four additional channels, one each with standard A-, B-, and C-weighting

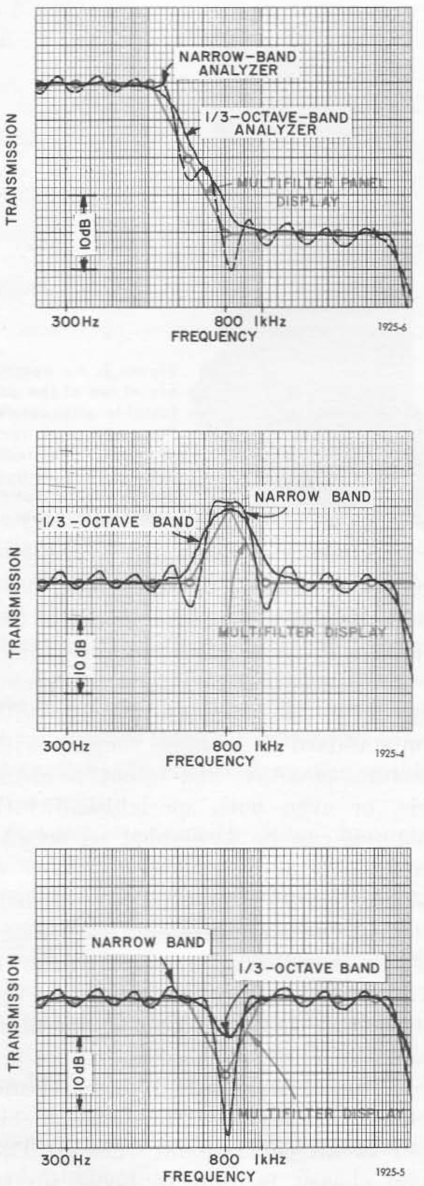


Figure 4. The multifilter's display is generally a better indication of the synthesized spectrum than one can get from a ⅓-octave analysis. Shown here are three awkward spectra. The red circles represent the dots on the front-panel display; the dashed black line is the "true" (narrow-band) spectrum. Notice that the fine-structure in the true spectrum does not appear in the ⅓-octave analysis (solid black line).

networks and one with flat response, are accessible at their own output connectors and can be scanned along with the filter channels. Peak detectors, before and after the filter channels, monitor for overload, and their outputs are displayed by a panel meter calibrated in dB referred to the overload level. Current proportional to the meter deflection is available at a rear-panel connector.

A WORKING MEASURE OF THE SYNTHESIZED SPECTRUM SHAPE

The remarkable versatility of the 1925 has been achieved by dividing up the frequency spectrum into contiguous bands. When one is using the 1925 for spectrum shaping, it is important to bear in mind the particular characteristics and limitations inherent in this method of spectrum shaping.

First, we cannot synthesize detail that is finer than the bandpass characteristic of the individual channels. Suppose we try to generate a narrow band of noise by attenuating all but one ⅓-octave multifilter channel. The spectrum we shall generate is a band of energy ⅓ octave wide — a narrower peak is unattainable with ⅓-octave filters. If we try to generate a dip in the spectrum by attenuating only one channel, we encounter the effect of finite stop-band-attenuation rates. Even though the filters have extremely high attenuation rates, the adjacent channels will partially fill-in for the attenuated channel and thereby limit the attenuation in the notch.

Second, because of phase cancellations and reinforcements when the channel outputs are summed, the spectrum synthesized by the multifilter has a fine-structure of narrow peaks and

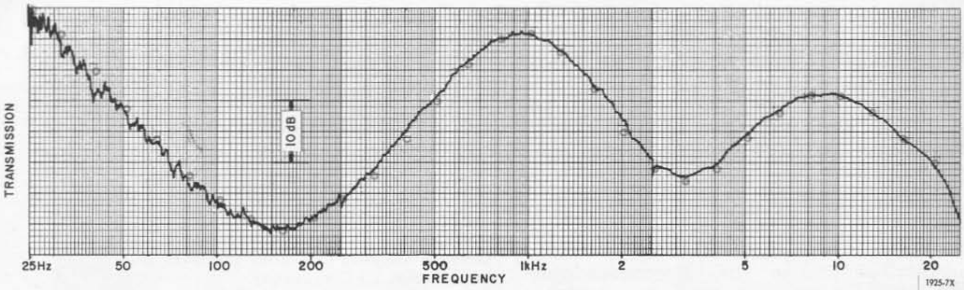


Figure 5. A more likely example. When there are no abrupt discontinuities or steep slopes, the ⅓-octave analysis (solid black line) departs from the multifilter display by less than the measurement error.

dips. The widths of individual features in the fine-structure are always much less than the channel bandwidth, and the spectrum, viewed through a "window" that is as wide as or wider than the channel bandwidth, is a smooth curve that conforms to the dot pattern on the front-panel display.

A parallel can be drawn between the accuracy of a spectrum synthesizer in shaping a spectrum and the accuracy of an analyzer in resolving one. A discrete frequency component, swept by a ⅓-octave-band analyzer, looks like a band of energy ⅓-octave wide. When we use the ⅓-octave analyzer to look at a spectrum with an unusually narrow dip, the ⅓-octave bandwidth and finite stop-band-attenuation rate cause the dip to fill in. When we analyze a spectrum having peaks and valleys narrower than the bandwidth of the analyzer, we get an averaged, smooth curve in which the peaks and valleys do not appear.

One might feel that an analyzer with narrower bandwidth should be used for these measurements; if it is necessary to see fine detail in the spectrum, then, indeed, a narrow-band analyzer must be used. But for most purposes a ⅓-octave analysis is entirely satisfactory. Users of sound analyzers understand and accept the limitations that bandwidth imposes on resolution, and the effects of finite resolution are readily interpreted by those experienced in the art.

Now, these same considerations apply to the spectrum synthesizer. Just as we must use an analyzer with narrower bandwidth to look at finer detail in a spectrum, so must we use a set of narrower filters if the resolution of our synthesizer is insufficient for our purpose. Although the spectrum synthesized by a ⅓-octave (say) multifilter may be, strictly speaking, somewhat in error with respect to the instrument's front-panel display, a ⅓-octave analysis of the spectrum will also be in error, and usually to a greater degree. The situation is illustrated in Figures 4 and 5.

Conclusions: 1) It is helpful when using a spectrum synthesizer to bear in mind the analogy between synthesizers and analyzers. 2) The multifilter's ⅓-octave calibrated panel display gives a better measure of averaged spectrum shape than can be obtained from a ⅓-octave analysis. 3) If we are going to look at a spectrum with ⅓-octave resolution, there is no need to use greater than ⅓-octave resolution to synthesize it.

— W. R. Kundert

A brief biography of Mr. Kundert appeared in the April 1968 issue of the *Experimenter*.

SPECIFICATIONS

FILTERS

Characteristics: (Figure 1) Both octave-band and one-third octave-band filters are six-pole Butterworth designs. Specified bandwidths are effective bandwidths, i.e., bandwidths for noise. Filters meet all current American and international standards: ⅓-octave conforms to USAS 1.11-1966 Class III (high attenuation), the octave filters to USAS 1.11-1966 Class II (moderate rate but highest for octave-band filters). Both octave and third-octave characteristics conform to IEC Recommendation Publication 225-1966.

Accuracy of Center Frequency: ±2%.

Passband Ripple: 0.5 dB max peak to peak.

Uniformity of Levels: At center frequencies (attenuator at +25 dB) ±0.25 dB at 25°C; ±0.5 dB, 0 to 50°C.

Noise: <15μV equivalent input noise.

Distortion: For bands centered at 25 Hz and above, harmonic distortion at band center is <0.1% at 1V out. For bands with center frequency below 25 Hz, distortion at band center is <0.25% at 1V out.

ATTENUATORS

Range: Gain in each channel adjustable in 1-dB steps from +25 dB to −25 dB relative to nominal 0-dB gain by means of panel control.

Accuracy: ±0.25 dB relative to +25-dB attenuation setting.

Readout: Panel display indicates attenuation in each channel and represents transmission between input and summed output. Display has standard 50-dB-per-decade scale factor; 10 dB per inch vertical, 5 inches per decade horizontal. Lock on panel prevents accidental changes in attenuator settings.

CHASSIS (accepts up to 30 filters)

Over-all Gain: 0 dB nominal.

Gain Adjustment: +6 to −12 dB, common to all channels.

Input Impedance: 100 kΩ.

Input Voltage: Ac component, ±17 V pk max referred to dc component of input. Dc component, ±35V max.

Scanner: Any single filter output is selected by a rear-panel pushbutton, external switch closure, or by use of a Type 1771 Scanner Control (available on special order), which displays and outputs (BCD) the channel number.

Peak Monitor: A peak detector senses levels at two circuit points and drives a panel meter calibrated in dB referred to overload level. A signal proportional to meter indication is available at an output jack for driving a dc recorder; 1 mA corresponds to full-scale reading.

OUTPUTS

Channel Outputs (Parallel Output)

Impedance: 20 Ω nominal.

Voltage: ±4.2 V max (3 V rms sine wave).

Load Impedance: 3 kΩ minimum for max output voltage.

Scanned Output:

Impedance: 20 Ω nominal.

Voltage: ±4.2 V max (3 rms sine wave).

Load Impedance: 3 kΩ minimum for max output voltage. Two chassis can be wired in parallel for up to 60 scanned channels.

Summed Output (For equalizing and shaping applications)

Impedance: 600 Ω.

Voltage: ±4.2 V max, open circuit.

Load Impedance: Any. Will not affect linear operation of output.

Weighted and Unfiltered Outputs

Impedance: 20 Ω nominal.

Load Impedance: 3 kΩ minimum for max output voltage.

Gain: 0 dB nominal at 1 kHz.

Weighting: A, B, and C characteristics conform to requirements of current American and international standards including USAS 1.4, IEC R123, and IEC R179.

GENERAL

Accessories Supplied: Power cord, 36-terminal plugs (2), spare fuses, *Handbook of Noise Measurement.*

Power Required: 100-125 V or 200-250 V, 50-60 Hz, 17 W.

Mounting: Rack-bench mount.

Dimensions (width X height X depth): Bench, 19¾ x 9⅛ x 14 in. (500 x 235 x 355 mm); rack, 19 x 8¾ x 12¼ in. (485 x 225 x 315 mm).

Weight: Net, 48 lb (22 kg), approx.

Catalog Number					*Price in USA*	
With Attenuator		*Without Attenuator*		*Description*	*With Attenuator*	*Without Attenuator*
Bench	*Rack*	*Bench*	*Rack*			
				1925 Multifilter		
				One-Third-Octave Bands		
1925-9700	1925-9701	1925-9712	1925-9713	25 Hz to 20 kHz	$3500.00	$3150.00
1925-9702	1925-9703	1925-9714	1925-9715	12.5 Hz to 10 kHz	3580.00	3230.00
1925-9704	1925-9705	1925-9716	1925-9717	3.15 Hz to 2.5 kHz	3680.00	3330.00
1925-9706	1925-9707	1925-9718	1925-9719	100 Hz to 80 kHz	3450.00	3100.00
				Octave Bands		
1925-9708	1925-9709	1925-9720	1925-9721	31.5 Hz to 16 kHz	2110.00	1990.00
1925-9710	1925-9711	1925-9722	1925-9723	4 Hz to 2 kHz	2210.00	2090.00

Note: Rack and bench versions of any model are priced the same.

Figure 1. The 1191-Z 500-MHz Counter.

A 100:1 SCALER FOR FREQUENCY MEASUREMENTS TO 500 MHz

The instrument shown in Figure 1 is the 1191-Z 500-MHz Counter, a combination of the already-popular 1191 Counter and the new 1157 Scaler. The present article discusses the scaler, an instrument that performs a relatively simple task: it accepts an input signal with a frequency up to 500 MHz and provides an output square wave whose frequency is exactly 100 times lower. The 1157, like its companion, the 1156-A Decade Scaler[1], is usable alone for any application where precise frequency division is required, or in combination with the 1191 Counter[2] or the recently-announced 1159 Recipromatic Counter[3].

APPLICATIONS

Counter Range Extension

Scalers (or prescalers) provide a simple and economical means for extending the frequency ranges of counters. The one drawback is a loss of resolution when they are used with low-frequency counters. With high-frequency counters, however, such as the 1191 or the 1159, this drawback disappears. With an 1191-Z combination one can measure 500 MHz with a resolution of 2 parts in 10^7 for a 1-second measurement and 2 parts in 10^8 for a 10-second measurement.

When the Type 1157 Scaler is used to extend the frequency range of a counter, the frequency to be measured is applied to the input of the Scaler, and the output of the Scaler is applied to the input of the counter. The Scaler output signal is sufficient to drive any known counter over the Scaler's entire range. The frequency is read from the counter by moving the decimal point in the counter display two places to the right (multiplication by 100). The accuracy of the measurement is not affected by the Scaler. Accuracy is strictly a function of the counter and is usually specified as ± 1 count (at counter input) $\pm$ crystal-oscillator stability.

[1] *GR Experimenter*, September 1965.
[2] *GR Experimenter*, December 1967.
[3] *GR Experimenter*, June 1968.

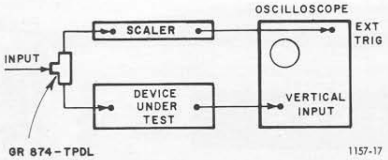


Figure 2. Use of the 1157 Scaler to trigger an Oscilloscope.

FM Measurements

The scaler is a wide-band frequency divider using digital pulse-count techniques and is not restricted to narrow-band sinusoidal signals. Frequency-modulation characteristics of the input signal are therefore preserved (divided by 100) and the scaler can be used to extend the range of frequency discriminators. For example, the 1157 Scaler can be used with the 1142-A Frequency Meter and Discriminator to extend its range to 150 MHz. An 1156-A Decade Scaler connected between the 1157 and 1142-A can extend the range further to 500 MHz.

Oscilloscope Trigger

The TYPE 1157 Scaler is especially valuable in the trigger path of an oscilloscope whose trigger capabilities are inadequate to lock a signal within its vertical passband. This technique is equally useful in testing scalers or frequency dividers even at lower frequencies. If the trigger is taken from the output of the device under test and the output waveform changes or the device fails, the trigger is lost. But if the trigger is taken ahead of the device (see Figure 2), the trigger is independent of the device. This is also an advantage when making time-relationship measurements comparing several points in the device, since the output from the 1157 Scaler can be used as a time reference.

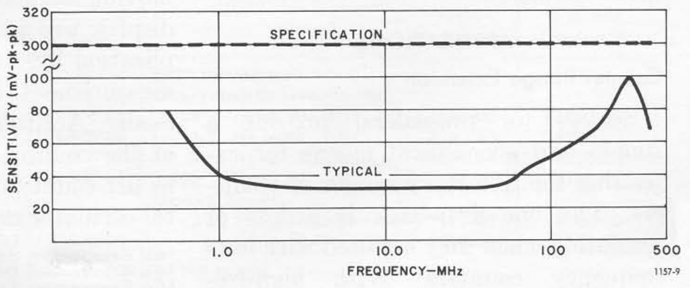
DESCRIPTION

The 1157 Scaler is housed in a 3½-inch-high relay-rack or bench cabinet. Input and output connectors are GR874® locking connectors, which can be conveniently moved to the rear of the instrument for racked-systems applications.

An input-attenuator control and level meter are provided on the panel for convenient adjustment of the input-signal level. The input impedance is 50 ohms for all positions of the attenuator control. Thus the scaler can be used as a 50-ohm load or cable termination. Alternately 50-ohm oscilloscope probes such as the Tektronix P6026, P6034, and P6035 can be used to raise the input impedance.

The input sensitivity is specified as 0.1 volt rms to 500 MHz. Figure 3 shows the sensitivity of a typical instrument as a function of frequency for a sine-wave input.

Figure 3. Typical and specified peak-to-peak sine-wave input voltage required to drive the 1157 Scaler.



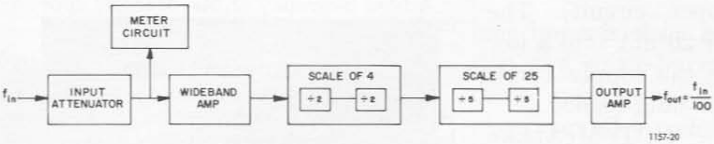


Figure 4. Simplified block diagram of the 1157 Scaler.

HOW IT WORKS

Block Diagram

Figure 4 shows a simplified block diagram of the 1157 Scaler. The instrument contains a 50-ohm attenuator, an input metering circuit, a wide-band amplifier, a scale-of-4 divider (2 cascaded flip flops), a scale-of-25 divider (2 cascaded divide-by-5 circuits), and an output amplifier. Interesting features of the circuits are discussed below.

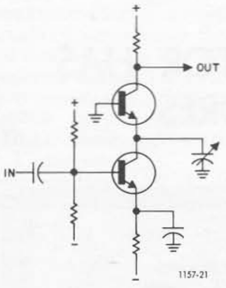


Figure 5. Simplified schematic diagram of the emitter-peaked cascode amplifier.

Input Amplifier

In this amplifier we reinvented the cascode circuit, but with a new twist.[4] Figure 5 shows a simplified schematic diagram of one stage of the amplifier. The small trimmer capacitor connected from the collector of the lower (common-emitter) transistor to ground combines with the inductive input impedance of the upper (common-base) transistor to form a low-Q resonant circuit. This trick extends the frequency response of an ordinary cascode circuit from 200 MHz, or so, to over 500 MHz. Two such stages interconnected by a shunt peaking network are used in the amplifier.

Scale-of-4 Divider

Figure 6 shows a simplified schematic diagram of one of the two flip-flops in the divider. The circuit uses a high-speed tunnel diode in combination with a transistor differentiating circuit to produce a nanosecond pulse which operates the following tunnel-diode circuit as a complemented flip-flop. This simple flip-flop circuit[5] has proved very useful in both low- and high-frequency divider applications.

Scale-of-25 Divider

The scale-of-25 divider uses two cascaded "Englemann Rings"[6], a unique combination of 5 bi-stable Schmitt circuits serially connected to form a ring counter.

Output Amplifier

The output amplifier includes a Schmitt circuit and provides sharp square waves of about 7 volts peak-to-

[4] H. T. McAleer, "Emitter Peaking Pushes Bandwidth to 500 MHz," *Electronics*, Sept 4, 1967.

[5] W. F. Chow, "Tunnel-Diode Digital Circuitry," *IRE Transactions on Electronic Computers*, EC-9(3), 295–301, Sept 1960.

[6] Rudolph Englemann, "Bi-quinary Scaling: Accuracy and Simplicity at 500 Mc," *Electronics*, p 34, November 15, 1963.

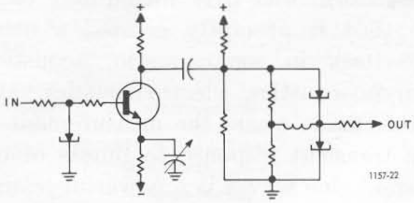


Figure 6. High-speed flip-flop — simplified schematic diagram.

peak amplitude (open circuit). The amplifier can deliver 20 mA into a low impedance and thus can supply 1 volt peak-to-peak to a 50-ohm load.

— H. T. McAleer

A brief biography of Mr. McAleer appeared in the December 1966 issue of the *Experimenter*

ACKNOWLEDGMENTS

Early design of the 1157 Scaler was done by H. T. McAleer. J. K. Skilling and R. L. Moynihan assisted in the latter stages of development.

SPECIFICATIONS

INPUT

Frequency: 1 to 500 MHz.

Sensitivity: Better than 0.3 V pk-pk (0.1 V rms) over entire range.

Impedance: 50Ω.

Attenuator: 1-2-5 sequence for signals up to 5 V rms.

Max Input: 5 V rms.

OUTPUT

Frequency: 0.01 to 5 MHz. Approx square-wave output, 20 mA pk-pk; > 5 V pk-pk open-circuit, 1 V pk-pk into 50Ω.

Impedance: 250Ω.

GENERAL

Power Required: 105 to 125 or 210 to 250 V, 50 to 60 Hz, 25 W.

Terminals: GR874® locking connectors; can be attached to either front or rear panel. Adaptors to other connector types available.

Catalog Number	*Description*	*Price in USA*
	1157 Scaler 100:1	
1157-9801	Bench Model	**$650.00**
1157-9811	Rack Model	**650.00**
	1191-Z Counter (100 MHz)	
1191-9900	Bench Model	**2040.00**
1191-9901	Rack Model	**2040.00**
	1191-Z Counter (500 MHz)	
1191-9902	Bench Model	**2190.00**
1191-9903	Rack Model	**2190.00**

REDESIGNED TONE-BURST GENERATOR HAS CUSTOMER-SUGGESTED FEATURES

Wider frequency range

Increased suppression of signal during off period

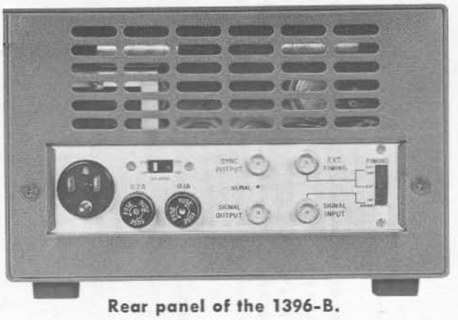
More output

Timed or counted intervals may be selected independently for both on and off periods

When the Type 1396-A Tone-Burst Generator was first introduced early in 1964 it promptly created a name for itself in sonar, audio, acoustics, psychoacoustics, electroacoustics, and other fields where the measurement of ac transient response to bursts of coherent sine waves is a powerful testing technique.[1] The considerable number of unforeseen uses to which the 1396-A was put prompted a redesign, which incorporates features that enhance the instrument's usefulness in many applications.

Among the new features that have been built into the 1396-B is the provision for independent choices of either

[1] J. K. Skilling, "A Generator of AC Transients," *GR Experimenter*, May, 1964; "Testing with Tone-burst Signals," *The Electronic Engineer*, December, 1966 (GR Reprint A130). ———, "The Frequency Spectrum of a Tone Burst," GR *Instrument Notes*, IN-105.



Rear panel of the 1396-B.

counted or timed intervals for both the on and off periods. This flexibility will be found useful in sonar and psychoacoustic applications, where a long burst is often required. The addition of a single-burst button facilitates those applications in physiology and experimental psychology that require the generation of a single burst on the experimenter's command.

Here are some of the applications to which the tone-burst technique is uniquely suited: transducer testing in the presence of reflections, self-reciprocity transducer calibration, measurement of room acoustics, measurement of ac meter ballistics, recovery-from-overload tests, music-power measurements, generation of power-line transients.

SPECIFICATIONS

SIGNAL INPUT (signal to be switched)
Amplitude: Proper operation results from input signals of not greater than 10 V pk (7 V rms) and not less than 1 V pk-pk.
Frequency Range: Dc to 2 MHz.
Input Impedance: 50 kΩ, approx.
TIMING INPUT (signal that controls switch timing). Same specifications as SIGNAL INPUT except:
Input Impedance: 20 kΩ, approx.
SIGNAL OUTPUT
Output On: Replica of SIGNAL INPUT at approx same voltage level; dc coupled; down 3 dB at >1 MHz. Output current limits at >25 mA pk, decreasing to >15 mA at 2 MHz. Output source impedance typically 25 Ω increasing above 0.2 MHz. Total distortion contribution <0.3% at 1 kHz and 10 kHz.
Output Off: Input-to-output transfer (feed-through), <−60 dB, dc to 1 MHz, increasing above 1 MHz.
Spurious Outputs: Dc component and change in dc component due to on-off switching (pedestal) can be nulled with front-panel control. Output switching transients are typically 0.2 V pk-pk and 0.2 μs in duration (120-pF load).
ON-OFF TIMING: Timing is phase-coherent with, and controlled by, either the signal at the SIGNAL INPUT connector or a different signal applied to the EXT TIMING connector. The on interval (duration of burst) and the off interval (between bursts) can be determined by cycle counting, timing, or direct external control.
Cycle-Count Mode: On and off intervals can be set independently, to be of 1, 2, 4, 8, 16, 32, 64, or 128 cycles (i.e. periods) duration or to be 2, 3, 5, 9, 17, 33, 65, or 129 cycles with +1 switch operated.
Timed Mode: On and off intervals can be set, independently, for durations of 10 μs to 10 s. On and off times occur at first proper phase point of controlling signal occurring after time interval set on controls; one interval can be timed while other is counted.
Switching Phase: In above modes, input controls determine phase of timing signal at which on and off switching occurs. SLOPE control selects either positive or negative slope of timing signal: TRIGGER LEVEL control sets voltage level at which both on and off switching occur.
Direct External Control: A 10-V pulse applied to rear-panel connection will directly control switching.
SYNCHRONIZING PULSE: A dc-coupled pulse that alternates between approx +8 V for output on, and −8 V when off. Source resistance approx 0.8 kΩ for positive output and 2 kΩ for negative.
GENERAL
Power Required: 100 to 125 or 200 to 250 V, 50 to 400 Hz, 16 W.
Accessories Supplied: Power cord.
Mounting: Convertible-Bench Cabinet.
Dimensions (width × height × depth): Bench, 8½ × 5⅝ × 10¼ in. (220 × 145 × 260 mm); rack, 8½ × 5¼ × 8⅝ in. (220 × 135 × 255 mm).
Weight: Net, 8 lb (3.7 kg); shipping, 12 lb (5.5 kg).

Catalog Number	*Description*	*Price in USA*
	1396-B Tone-Burst Generator	
1396-9702	Bench Model	**$550.00**
1396-9703	Rack Model	**572.00**

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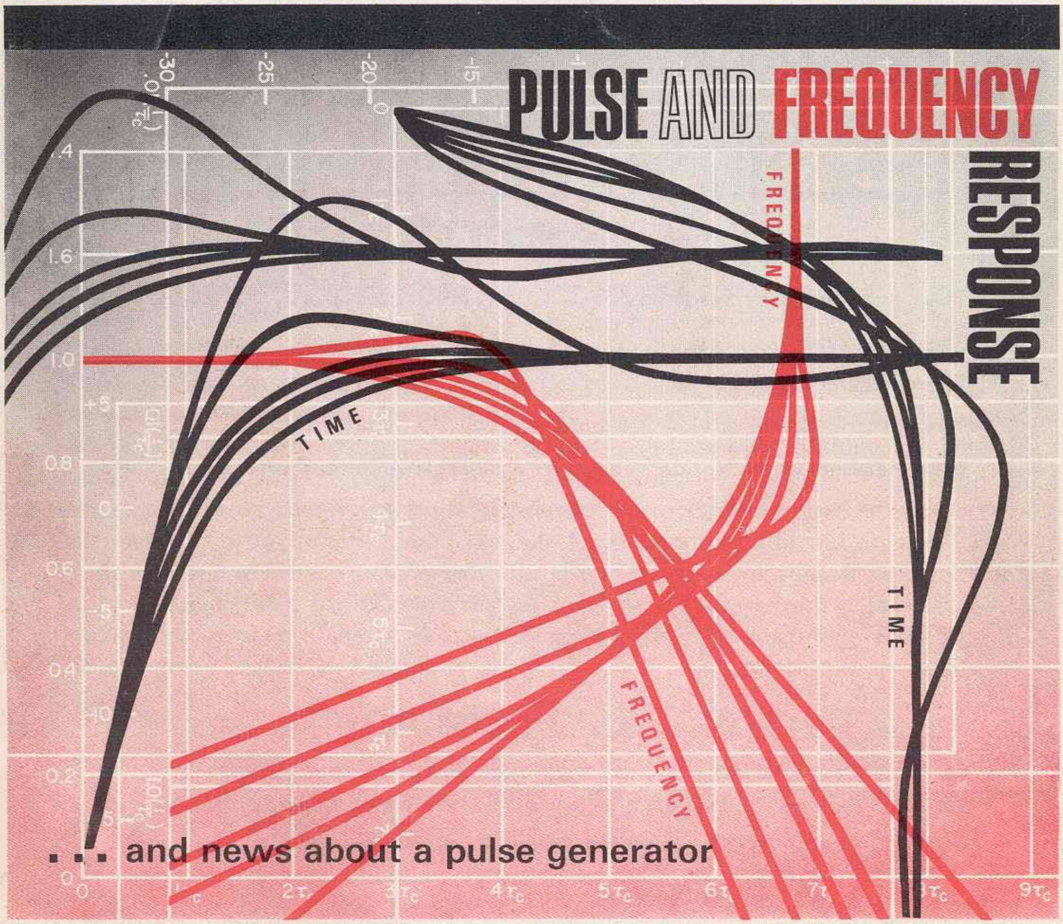
Experimenter



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Volume 42 • Nos. 11, 12 November-December 1968

Published monthly by the General Radio Company

The *General Radio Experimenter* is mailed each month without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.

PULSE AND FREQUENCY RESPONSE

Some Useful Relations

by J. K. Skilling

If a fast pulse is applied to a wideband circuit, and the output has a rise time of 50 nanoseconds and no overshoot, the upper cutoff frequency is approximately 7 MHz. If the same circuit exhibits 25 percent droop in its response to a 1-second pulse, its low-frequency cutoff is 0.05 Hz.

It is often useful to be able to infer the frequency response of a device or circuit from measurements of pulse response. For one thing, instrumentation for pulse work is generally simpler than that for sine-wave measurements. To measure directly the two cutoff frequencies in the above example, one would probably need two sine-wave signal generators to cover the range from 0.01 Hz to 10 MHz. Measurements at several frequencies would have to be taken. A sweeper would be more convenient — although it would give no phase information — but would entail an even larger commitment to instrumentation. On the other hand an economical, general-purpose pulse generator and oscilloscope are the only instruments needed for the rapid pulse testing of wideband circuits and devices.

And there are other advantages to the use of pulse techniques. Pulse excitation allows independent control over signal level and average power in the device under test. Also, the use of an oscilloscope to view the output immediately uncovers gross distortion, clipping, oscillation, etc that might otherwise escape notice.

On the debit side, the present state of instrumentation does not allow the observation of waveforms with the accuracy that characterizes sine-wave testing. As a result, pulse tests give less accurate information about frequency response than do direct measurements.

MODELS

Correlation of pulse and sine-wave responses is generally done, as most active-circuit calculations are done, with models. One calculates both pulse-response waveforms and frequency-response curves for a given model circuit with given component values. The assumption is that if the observed waveform from a real device is the same as a waveform that one has calculated for the model, then the device's frequency response will be the same as that calculated for the model. This is born out, at least to a good approximation, by most common wideband, linear devices and circuits that have reasonably smooth rolloffs at both low and high frequencies.

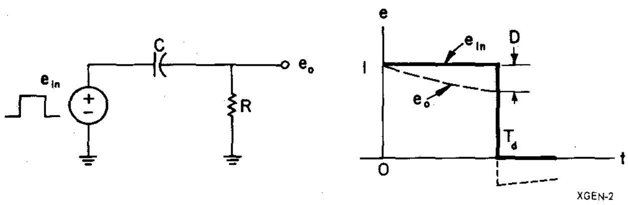


Figure 1. Low-frequency model and input and output waveforms.

LOW-FREQUENCY RESPONSE

The low-frequency response of the circuit governs the amount of "droop" in the output pulse. It is usually adequate to choose a simple RC high-pass circuit as a model for low-frequency behavior, as shown in Figure 1. If the amount of droop is less than 30 percent, the low-frequency cutoff frequency f_1 is given in terms of the droop D (percent) and the pulse-duration time T_d by the approximate formula

$$f_1 \doteq \frac{0.159}{T_d} \cdot \frac{D}{100}$$

In practice, one might adjust the pulse duration until a droop of 25 percent is observed. When $D = 25\%$, the low-frequency cutoff can be calculated with a formula that is a little more accurate than the one given above:

$$f_1 \doteq \frac{0.0456}{T_d} \qquad (\text{droop} = 25\%)$$

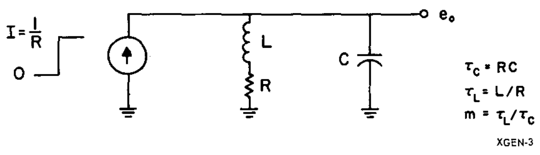
HIGH-FREQUENCY RESPONSE

The high-frequency behavior of the circuit governs the way in which the output waveform responds to an input step. A model suitable for high frequencies is somewhat more complicated than the low-frequency model we have just discussed. The one usually chosen is the series-parallel circuit shown in Figure 2.[1] The source is a current step whose magnitude I is such that $IR = 1$. The ratio m measures the amount of peaking or, inversely, the amount of damping. (It is equal to the square of the circuit's Q at the resonant frequency of L and C.)

Figure 3 shows calculated output waveforms for various values of m. Note that the time is measured in units of τ_C. Also shown in Figure 3 are the definitions of two parameters having to do with the circuit's step response: the overshoot r and rise time t_r.

[1] Valley and Wallman, *Vacuum Tube Amplifiers*, M.I.T. Radiation Laboratory Series, No. 18, Boston Technical Publishers, Inc., Lexington, Mass., 1964.

Figure 2. High-frequency model. The source is a current step of magnitude *I*, and the circuit elements are in a series-parallel configuration.



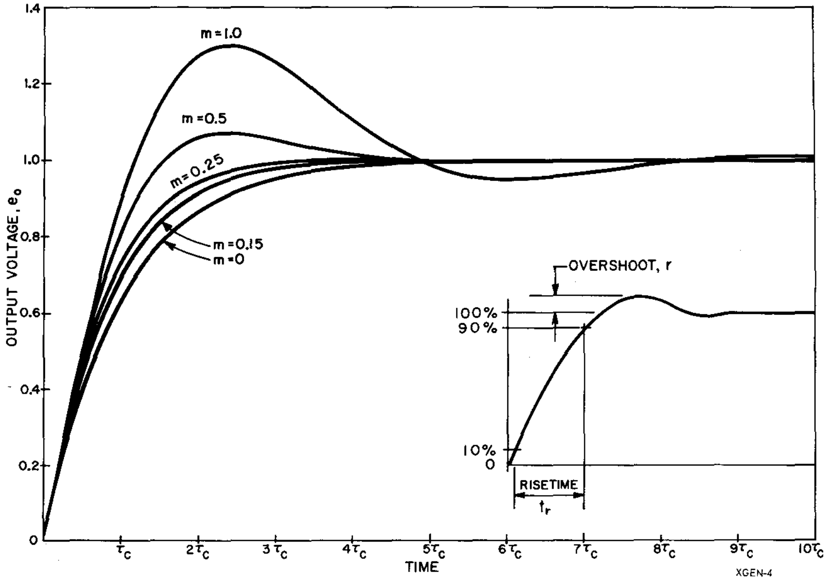


Figure 3. Step response of the series-parallel high-frequency model of Figure 2. The time axis is set off in units of τ_c and the waveforms are functions only of *m*. The illustration in the lower right corner shows how overshoot and risetime are defined.

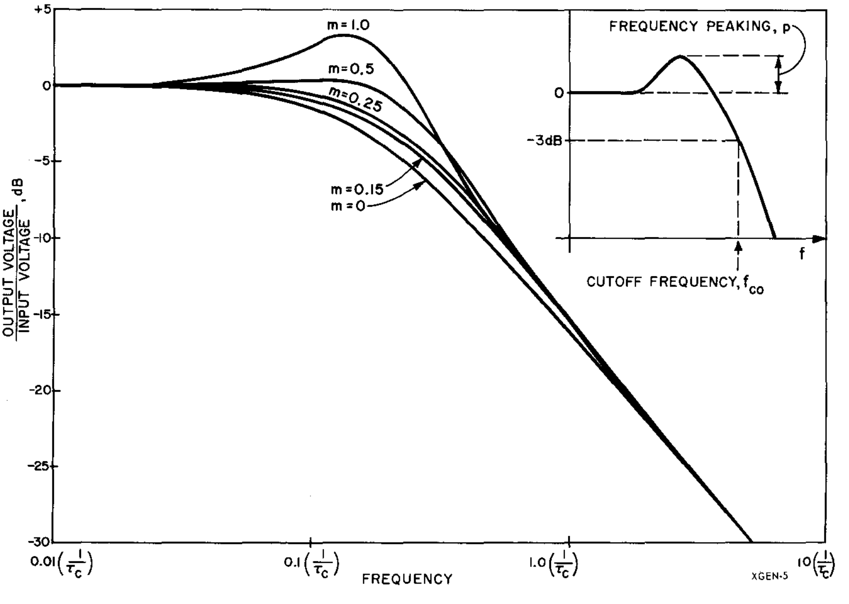


Figure 4. High-frequency response of the series-parallel model of Figure 2. These curves are the companions of those in Figure 3. The frequency axis is marked off in units of $1/\tau_c$, and the shapes of the curves, like the shapes of those in Figure 3, depend only on *m*. The amount of peaking *p* (dB) and the cutoff frequency f_{co} are defined by the sketch in the upper right corner.

The corresponding frequency-response curves are shown in Figure 4. Here the frequency is measured in units of $1/\tau_C$ and the shape of the curve is again a function only of m.

Let us look at a few simple relations for representative values of m.

$\underline{m = 0}$

There is no inductance, and the model reduces to the simple RC low-pass network.

$$t_r = 2.2\,\tau_C, \qquad f_{co} = \frac{0.159}{\tau_C}$$

$$\boxed{t_r f_{co} = 0.35 \text{ (golden rule)}}$$

$\underline{0 < m < 0.25}$

Now there is some inductance present. As m increases, the rise time decreases and the high-frequency cutoff increases.

$\underline{m = 0.25 \text{ (critical damping)}}$

This is the condition for the shortest rise time without overshoot (although one can do better with a more complicated "peaking circuit"[2]).

$$t_r = 1.53\tau_C, \qquad f_{co} = \frac{0.225}{\tau_C}$$

$$\boxed{t_r f_{co} = 0.344}$$

The golden rule survives.

$\underline{0.25 < m \leq 0.50}$

The rise time continues to decrease and the cutoff frequency continues to increase, but overshoot starts to build up and so does frequency peaking. At $m = 0.50$,

$$t_r = 1.12\tau_C, \qquad f_{co} = \frac{0.286}{\tau_C}$$

$$\boxed{t_r f_{co} = 0.32}$$

The golden rule is still barely tarnished.

$\underline{m > 0.50}$

Rise time continues to decrease, but overshoot becomes very large and reproduction of the pulse waveform is consequently poor. The cutoff frequency con-

[2] See Valley and Wallman, *op. cit.*, p 75.

Table 1
Characteristic quantities for the series-parallel high-frequency model

m	normalized rise time t_r/τ_c	overshoot r	normalized cutoff frequency $f_{co} \cdot \tau_c$	rise-time-cutoff-frequency product $t_r \cdot f_{co}$	frequency peaking p
0	2.20	0%	0.159	0.350	0 dB
0.07	2.03	0	0.172	0.349	0
0.15	1.82	0	0.192	0.349	0
0.20	1.68	0	0.208	0.349	0
0.25	1.53	0	0.225	0.344	0
0.35	1.31	1.0	0.258	0.335	0
0.50	1.12	6.7	0.286	0.320	0.25
0.70	1.01	16.	0.294	0.297	1.5
1.0	0.940	30.	0.289	0.276	3.3
2.0	0.865	68.	0.268	0.232	7.8
5.0	0.825	148.	0.245	0.202	15.

tinues to increase until $m = 0.7$, where the bandwidth of the circuit is maximum. The frequency peaking at maximum bandwidth is 1.5 dB. For larger values of m, the model becomes a relatively high-Q resonant circuit with frequency peaking that is excessive for a wide-band device.

The relations discussed above are tabulated as a function of m in Table 1.

AN ALTERNATIVE HIGH-FREQUENCY MODEL

We can gain some confidence in the model-building method if we compare the above results with those for a different high-frequency model, the series circuit shown in Figure 5. The behavior of the series model is shown in Figures 6 and 7 and Table 2.

The response is similar to that of the series-parallel model. The improvement in rise time and cutoff frequency with increasing m is not as marked — the improvement is 10 percent less at $m = 0.25$, for example. The cutoff frequency maximizes at a lower value of m than with the series-parallel model. But the most

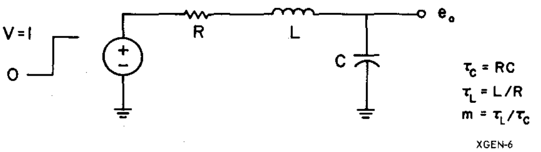


Figure 5. Another high-frequency model. The source in this circuit is a voltage step of unit magnitude and the circuit elements are in series.

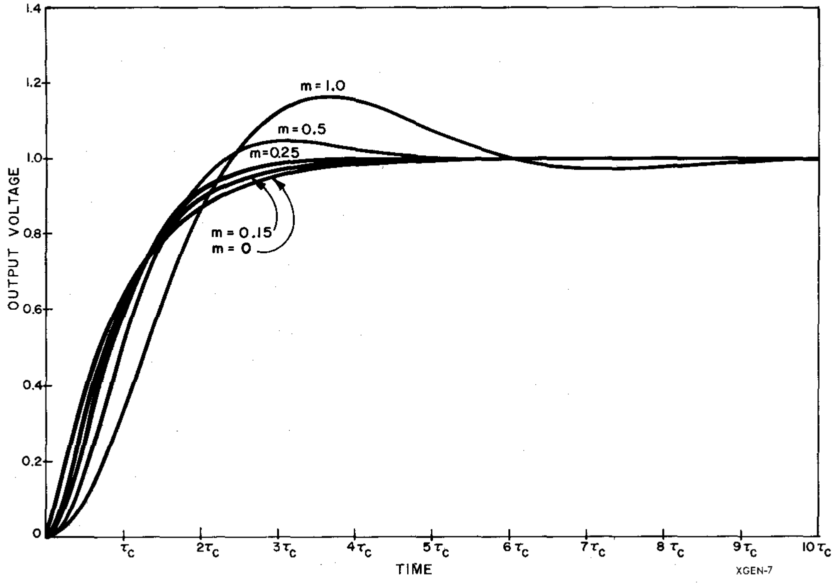


Figure 6. Step response of the series high-frequency model. The improvement in rise time with increasing *m* is a little less than in the series-parallel model, but otherwise these curves are very similar to those of Figure 3.

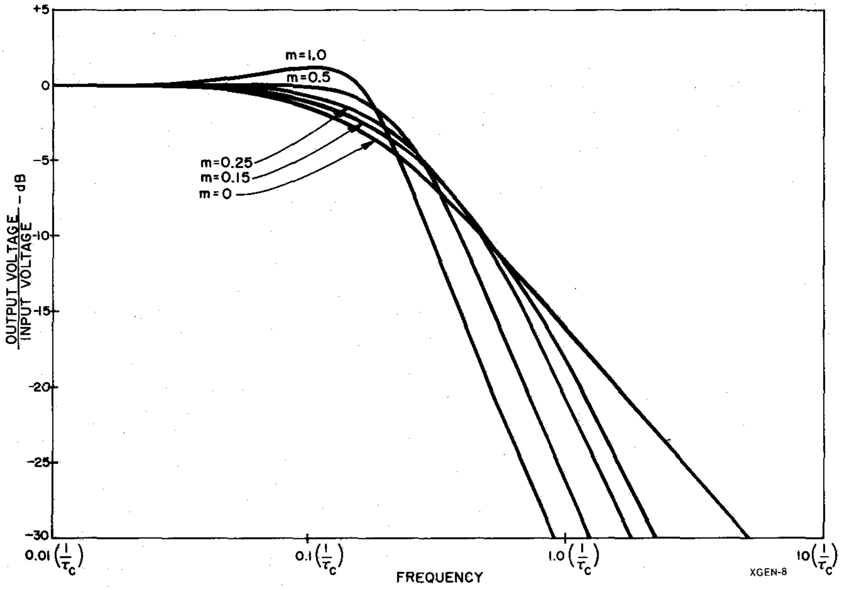


Figure 7. High-frequency response of the series model.

Table 2
Characteristic quantities for the series high-frequency model

m	normalized rise time t_r/τ_c	overshoot r	normalized cutoff frequency $f_{co} \cdot \tau_c$	rise-time-cutoff-frequency product $t_r \cdot f_{co}$	frequency peaking p
0	2.20	0%	0.159	0.350	0 dB
0.07	2.04	0	0.171	0.349	0
0.15	1.87	0	0.186	0.348	0
0.20	1.77	0	0.196	0.347	0
0.25	1.68	0	0.205	0.344	0
0.35	1.57	0.6	0.218	0.342	0
0.50	1.52	4.3	0.225	0.342	0
0.70	1.55	9.6	0.219	0.339	0.74
1.0	1.64	16.3	0.202	0.331	2.49
2.0	1.97	30.5	0.159	0.313	7.18
5.0	2.75	48.6	0.107	0.293	14.42

important thing to notice is that this circuit, too, obeys the golden rule: the rise-time-bandwidth product is very close to 0.35 for small m.

CASCADED STAGES

Cascaded stages with no overshoot have a combined rise time that is approximately the root-sum-square of the individual rise times:

$$t_{r\ combined} = \sqrt{t_{r1}^2 + t_{r2}^2 + t_{r3}^2 + \ldots}$$

This useful relation also allows one to calculate the effect of the measuring system on the rise time. For example, if a pulse generator and oscilloscope by themselves show a rise time of 20 nanoseconds, and if when they are used to measure the rise time of a device the result is 80 nanoseconds, the above relation gives a rise time of 77.5 nanoseconds for the device itself.

Bear in mind that the formula given above applies only in cases of no overshoot. Valley and Wallman[3] give the following rules for cascaded stages with overshoot: "For stages having very small overshoot (1 or 2%) the overshoot grows extremely slowly or not at all as the number of stages increases. . . . For stages having overshoots of about 5 to 10% the overshoot increases approximately as the square root of the number of stages, and the rise time increases substantially less rapidly than as the square root."

[3] *Op. cit.*, p 78.

CONCLUSIONS

The low-frequency cutoff can be determined from the measured amount of droop in the response to a long pulse. The prediction of high-frequency performance is a little more complicated because two cases have to be distinguished. In the overdamped zero-overshoot case ($0 \leq m \leq 0.25$) it is difficult to determine m from an observed pulse waveform. But it is surprising to note that the cutoff-frequency-rise-time product changes by less than 2 percent over the range $m = 0$ to $m = 0.25$. If the modeling procedure is reasonable, the prediction of cutoff frequency from rise time or vice versa anywhere in the overdamped region is accurate enough for engineering purposes.

In the underdamped case ($m > 0.25$), the cutoff-frequency-rise-time product depends upon m. But in this case the amount of overshoot, which one can measure with fair accuracy, gives a good estimate of m.

Acknowledgement

Computation of the data contained in the figures and tables was programmed by Bruce J. Jay.

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James K. Skilling received his BSEE degree from the University of California at Berkeley in 1953 and his MSEE degree from Johns Hopkins University in 1963.

Before joining General Radio in 1959, he held an engineering position at Douglas Aircraft and served as an electronics instructor at the United States Naval Academy.

At GR he has been primarily involved with the development of pulse equipment and digital system techniques. He is a member of IEEE, A.A.A.S. and SIGMA XI.

THE 1340 PULSER

Versatility and price make this the "best buy" of the general-purpose pulse generators.

When a new instrument improves on the performance of products already on the market, it represents progress; when it offers conveniences that the others don't, it is attractive; when at the same time its price meets or beats the competition, it makes news. So the GR 1340 Pulse Generator is news. It is a general-use pulse generator that offers a total combination of performance and features that is not surpassed even by high-priced generators.

0.2 Hz to 20 MHz

Repetition periods from 50 nanoseconds to 5 seconds will interest users in fields from high-speed circuit testing to seismological research. A single pulse for one-shot tests can be triggered manually from a front-panel pushbutton.

The 1340 also offers eight decades of pulse-duration times, from 25 nanoseconds to 2.5 seconds. The maximum duty ratio that can be attained varies from 70 to over 95 percent, depending on the combination of repetition-rate range and duration-time range that is in use. In case of an inadvertent setting that exceeds the allowable duty ratio, a panel lamp flashes to alert the operator that the output waveform may not be trustworthy. In any event there is no risk of a circuit overload if the duty-ratio limit is exceeded.

The repetition and duration controls each combine range switching and continuous adjustment in single easy-to-read dials. The repetition dial reads both prf and period, along with the correct units.

A convenience feature of the 1340 is the "squarewave" position of the pulse-duration control. In this setting the output consists of 50%-duty-ratio pulses, regardless of the prf. Square-wave testing with the 1340 is thus fast and simple, since there is no need to reset the duration when the prf is changed.

10-Volt Output, ± 1-Volt Offset

Both positive and negative pulses are available simultaneously at two low-reflection GR874® coaxial connectors. Each output is a high-impedance (about 1 kΩ) current generator with a maximum source current of 200 mA and a saturation voltage of better than 10 volts. Fifty-ohm resistors are switched across the outputs when reflectionless sources are required. With the resistors switched in, the 1340 can deliver 5-volt pulses to a 50-ohm load.

Another 1340 convenience feature is the output offset. Independently adjustable biases can be applied to the two outputs to offset the pulses up to ± 1 volt (or ± 20 mA). The offset feature eliminates the need for auxiliary equipment, for example, to ensure positive triggering in the presence of noise or small biases. It is a feature that will be of particular interest to those concerned with integrated-circuit testing.

The 1340's synchronizing trigger is a square wave, rather than the narrow pulse that is available at most pulse-generator sync outputs. The square wave has three principal advantages. 1) Even when the 1340 is being used to generate long pulses, the sync pulse is not too short to provide an effective trigger. 2) The square wave provides a stable trigger as the prf is changed, since its dc level does not vary. 3) By changing the scope's trigger level, one can synchronize the scope one-half period ahead of the pulse.

PPM, PDM, PAM

The period, duration, negative amplitude, and positive amplitude can each independently be linearly modulated by externally applied modulating voltages. Voltages between −0.5 and −5 volts sweep the period and duration over the decade ranges selected on the range switches. Voltages between 0 and +5 volts and 0 and −5 volts, respectively, vary the amplitudes of the positive- and negative-going pulses over the entire 10-volt range.

The 1340 has other modes of electronic control. A gating input permits an external switch closure or positive-going pulse to inhibit the 1340's period generator. Gating the oscillator means that a pulse that is already underway is not chopped by the gating signal. Whether one talks about modulation, sweeping, or remote programming, the 1340's electronic-control features mean greater usefulness.

A description of some of the 1340's novel circuitry follows the specifications.

SPECIFICATIONS

PULSE PERIOD (PRF)

Internally Generated: 50 ns to 5 s (20 MHz to 0.2 Hz) in 8 decade ranges. Single-pulse push button on panel.

Externally Controlled: 1 Hz to 20 MHz; triggers on any waveform of >3 V pk-pk. Input resistance approx 100 kΩ. Output pulse is started by negative-going transition. Period control acts as input trigger-level control in external mode.

OUTPUT-PULSE CHARACTERISTICS

Duration: 25 ns to 2.5 s in 8 decade ranges, or square wave.

Rise and Fall Times: 5 ns ±2 ns at 5 V, 50-Ω load, and 50-Ω source resistance.

Amplitude: Positive and negative ground-based pulses available simultaneously with independent amplitude and offset control. Source current continuously adjustable to at least 0.2 A (i.e., across 50-Ω load, 10 V from high source resistance or 5 V from 50-Ω source).

Offset: Continuously adjustable source current from −20 to +20 mA.

Source Resistance: 50 Ω, or high (approx 1 kΩ) for 50-Ω loads.

Distortion: Preshoot, overshoot, ringing, etc, <0.5 V (5% of max output).

Duty Ratio: Duty ratios of over 70% can be obtained on all ranges except decreasing to approx 50% at 50-ns period in 50-to-500-ns range.

SYNCHRONIZING PULSE

Waveform: Square wave. Negative transition precedes start of output pulse by approx 35 ns; positive transition can be used for half-period pretriggering.

Amplitude: 2.5-V pk-pk positive square wave behind 500-Ω source impedance.

MODULATION AND GATING

Modulation: Period and duration are linearly controllable by an external voltage between −0.5 and −5.0 V. Amplitude of the positive-pulse output is linearly controllable by an external voltage of 0 to +5 V, the negative-pulse output by 0 to −5 V. Period and duration are modulatable over the decade range set by range switches; amplitude can be modulated over its full range. Amplitude modulation can be used for noncoherent gating of output pulse.

Gating: An impedance of ≤600 Ω to ground inhibits output; +4 to +8 V allows normal output; 1340's can be used to gate 1340's.

GENERAL

Power Required: 100 to 125 or 200 to 250 V, 50 to 400 Hz, 30 W.

Accessories Supplied: Spare fuses, power cord.

Accessories Available: GR874® coaxial components, attenuators, terminations, tees, etc.

Mounting: Convertible-Bench Cabinet.

Dimensions (width × height × depth): Bench 8½ × 5⅝ × 13 in. (220 × 145 × 330 mm); rack, 19 × 5¼ × 11¼ in. (485 × 135 × 290 mm).

Weight: Net, 9¼ lb (4.2 kg); shipping, 13 lb (6.0 kg).

Catalog Number	*Description*	*Price in USA*
	1340 Pulse Generator	
1340-9700	Bench Mount	$395.00
1340-9701	Rack Mount	417.00

NOVEL VOLTAGE-VARIABLE SCHMITT CIRCUIT

The 1340's period generator makes use of the voltage-variable Schmitt circuit, shown in Figure 1.

Q1 and *Q2* are regeneratively connected to form a Schmitt-type multivibrator. The common emitter current supplied by R_E flows entirely in either *Q1* or *Q2* when the circuit is in one of its two stable states. Consider the case where *Q2* is conducting (on) and *Q1* and *Q3* are off. Under these conditions the voltage divider *R1*, *R2*, and *R3* sets a voltage, E_U, at the base of *Q2*. In order to keep *Q1* and *Q3* off, their base voltages must be more negative than E_U.

Transfer of conduction to *Q1* occurs when the input voltage E_{IN} on the base of *Q1* is raised to a value close to E_U, at which point *Q1* comes on enough to start the turn-off of *Q2*. Regenerative transfer of conduction to *Q1* results,

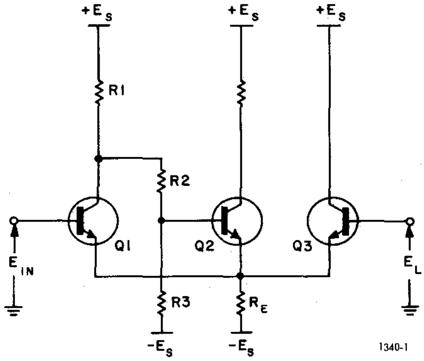


Figure 1. Simplified schematic diagram of the voltage-variable Schmitt circuit, part of the 1340's period generator.

and, with *Q1* on, the base of *Q2* is driven more negative than the base of *Q3*.

In order to transfer the conduction back to *Q2*, E_{IN} is now lowered to a value near E_L. At this point *Q3* begins to draw current, and this reduces the current in *Q1*, causing *Q2* to begin to turn on. The regenerative interconnection of *Q1* and *Q2* again causes the emitter current to shift, this time back to *Q2*.

Conduction is therefore shifted back and forth between *Q1* and *Q2* by cycling E_{IN} first up to an upper trigger level E_U and then down to a lower trigger evel near E_L. The significant advantage of this circuit over other multivibrators is that the lower trigger level is controlled in a very simple manner by an externally supplied voltage E_L. The lower trigger level is, in fact, almost exactly equal to E_L.

Two more transistors, *Q4* and *Q5*, turn the voltage-variable Schmitt circuit into the complete voltage-variable period generator, Figure 2. *Q5* is a constant-current source, supplying a current I to the timing capacitor C. *Q4* is a switched constant-current source, supplying no current when off and $2I$ when on.

When *Q1* is on, both *Q2* and *Q4* are off and the current into C is $-I$; the

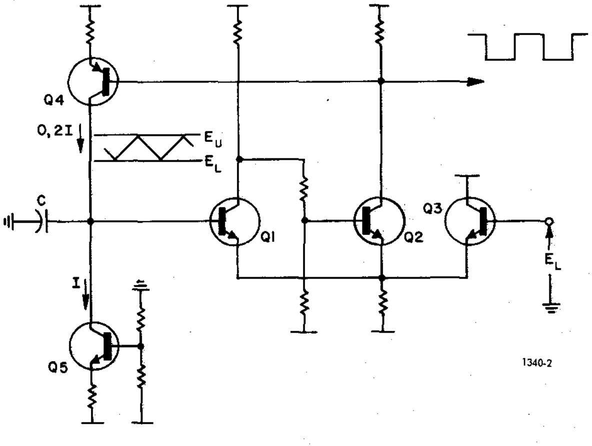


Figure 2. Simplified schematic of the complete period generator. Voltage control of the period is achieved by varying E_L, which governs the capacitor-voltage swing.

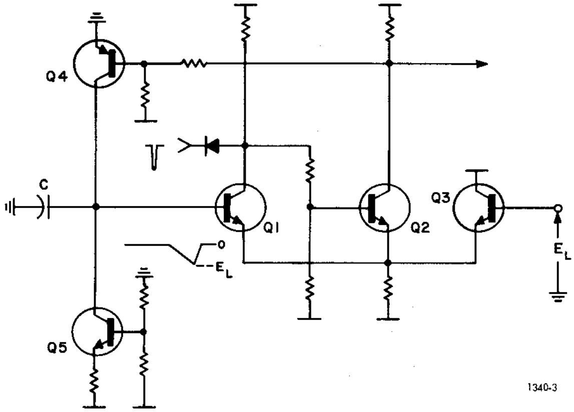


Figure 3. The pulse-duration generator, simplified. The control voltage E_L determines how long the circuit stays in its unstable state.

voltage across C drops linearly with time. When this voltage reaches E_L, the voltage-variable Schmitt shifts to the *Q1*-off, *Q2*-on state, and *Q4* comes on. The net capacitor current is now $+I$, and the capacitor voltage increases linearly. When the capacitor voltage reaches E_U, the Schmitt shifts states back to *Q1*-on, *Q2*-off. In this fashion the capacitor alternately charges and discharges so that the capacitor voltage is a triangular wave, and a square wave is developed at the collector of *Q2*.

Voltage control of the period is effected by varying E_L, which controls the peak-to-peak amplitude of the capacitor voltage. Since the charging currents are constant, the period of the square wave is directly proportional to the peak-to-peak swing of the capacitor voltage.

The duration generator uses the voltage-variable Schmitt circuit in the monostable configuration shown in Figure 3. In this circuit *Q5* is again a constant-current generator. In the stable state, *Q2* is on; *Q2* holds *Q4* on, and the capacitor voltage is clamped to ground. In monostable operation, the base of *Q2* is slightly above ground when it is on; thus the capacitor voltage, which cannot go positive, does not turn *Q2* off. A negative trigger, applied to the collector of *Q1*, turns off *Q2*, and *Q4* turns off with *Q2*. The constant-current source now charges C, and the capacitor voltage starts a negative climb. When it reaches E_L, the

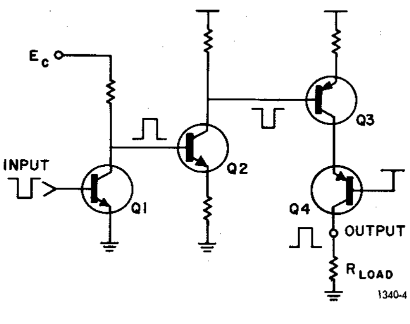


Figure 4. Simplified diagram of the positive-pulse output circuit. The two stages following Q1 are nonsaturating, so that E_c controls the amplitude of the output pulse.

circuit reverts to the stable *Q1*-off, *Q2*-on state. The duration of the output pulse is determined by the amplitude of the voltage change on the timing capacitor and therefore is linearly proportional to the control voltage E_L.

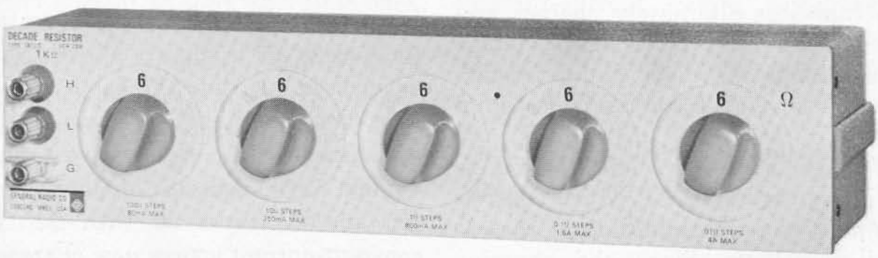
The output circuit for the positive pulse is shown in Figure 4.

The first transistor *Q1* is normally saturated and is turned off by the input pulse. The amplitude of the pulse at the collector of *Q1* is changed by adjusting the supply voltage E_c of this stage. *Q2* is a nonsaturating amplifier that provides current gain to drive the output stage. The cascode (grounded-emitter driving grounded-base) output stage is also nonsaturating and has a very fast rise time. In order to handle the high output current and yet maintain high switching speed, *Q3* and *Q4* are each a pair of parallel transistors. All the amplifier stages following *Q1* are nonsaturating so that the output amplitude is proportional to the control voltage E_c. This allows voltage control of the pulse amplitude.

The negative-pulse amplifier circuit is the same as the positive amplifier except that negative supply voltages and complementary transistor types are used.

— J. K. Skilling

IMPROVED DECADE RESISTORS



The GR line of precision decade resistors has a new look, better performance "under the hood," and a new type number.

General Radio decade resistors have been constantly improved since they were first introduced in 1917. Often these changes occurred as new techniques permitted, unheralded by outward appearance changes or new type-number designations. A review of published specifications, however, would reveal the steady improvements that have taken place. Most recently, both appearance and performance have gotten better, in recognition of which the precision 1432 line has been renamed the 1433. Improvements represented by this new number include:

In-line digital readout; reading errors are far less likely.

Models with from four to seven dials give required resolution.

All models rack mountable.

Better accuracy, $\pm 0.02\%$ two-year accuracy for high values.

"Over-all" as well as "incremental" accuracy specifications now given.

Better ac performance at high resistances, owing to reduced stray capacitance.

Rear-panel terminals for system installation.

"Over-all" accuracy versus "incremental" accuracy

Historically, GR has specified the accuracy of each separate resistor in a decade box rather than that of the total resistance. This practice has reflected the fact that each resistor is individually adjusted. We called this specification "accuracy of resistance increment" because the difference in value between any two positions of the same switch has the same accuracy as a resistor on that switch. Knowledge of the incremental accuracy is needed in many uses, and we are not abandoning it; but more often the absolute accuracy of any given setting of the box — the "over-all" accuracy — is desired.

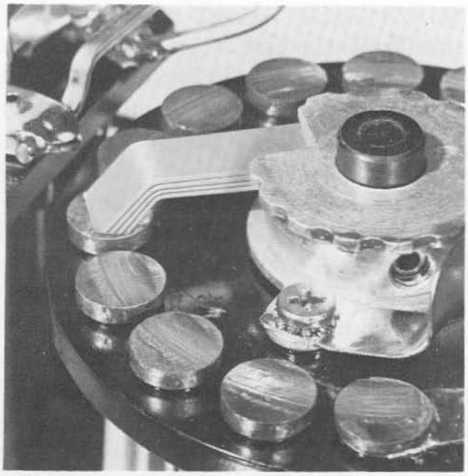
If all the resistors in a box had the same accuracy specification, any combination of them would have this same specification. Lower-value resistors, however, cannot be adjusted as easily and generally are not as stable as higher-value resistors. So our over-all specification has two terms: $\pm$ (0.02% + 2 mΩ).

A single over-all specification is intended to apply to any and all possible settings of the decade box, but there are obviously just too many to permit testing every one. Instead, we adjust the lower-value units to much tighter tolerances than the published $\pm$ 2 mΩ, and test the assembled boxes at carefully selected values with limits substantially closer than the catalog specifications. To substantiate the validity of this method, 300 boxes were "made" on a computer under the assumption of a very poor tolerance distribution. These were first computer "tested" against our laboratory specifications and then computer "measured" at *every* setting. No box that passed the simulated laboratory test was in error by more than 9/10 of the catalog specification at *any* setting.

Such testing, whether real or simulated, checks only the *initial* tolerance of the decade box. Confidence in the two-year accuracy of GR 1433's is assured by stability records of the resistors and by the use of exceptionally good switches.

Switches

Decade-box accuracy depends on the resistors and on the switches. In the over-all specification, switch resistance plays a particularly important part. Commercial rotary switches with solid-silver contacts, such as we use in many instruments, can vary 1 milliohm or more in contact resistance. Obviously, a five-decade box can display a variation of 5 milliohms or more, which would have to be accounted for in the specification. The GR-designed and manufactured switches in our Type 1433 decade boxes have a low and very repeatable contact resistance. Critical contacts (on lower decades and zero settings) use multiple-leaf wipers



The use of multi-leaf wipers on silver-overlaid studs assures a contact resistance that varies by less than 100 microhms.

against silver-overlaid studs (see photo). As a result, contact resistance changes by no more than 100 microhms and hence has but slight effect upon the specification. Our less expensive line, the 1434, uses commercial switches with double contacts to give a $\pm$ 0.05% $\pm$ 5 mΩ over-all specification.

Resistors

For added dimensional stability, the 1-ohm through 10-kilohm units are now wound on ceramic rather than fiber-glass or mica cards as before. While the 100-kilohm units are unchanged, the 1-megohm units are now wound on bobbins.

For years, we've held that flat, single-layer card resistors are far superior to multi-layer bobbin types, as the latter have much higher inductance and shunt capacitance that degrade their performance in ac operation. Indeed, we still feel this is true, but only for single resistors. When 1-megohm cards are wired together in a decade box, the capacitances between cards and from the cards to the box become the main cause of reduced ac performance. These capacitances are lower when bobbins are used, and the ac performance is better in spite of the higher individual shunt capacitance. We have now swallowed our pride and learned to wind precision bobbin resistors, for, besides being better for decade boxes, they're a lot less costly to make and actually permit lower prices on some decades.

Separate Decades

The improved accuracy of the 1433's is available in the individual Type 510 Decades, which are sold separately. These units are ideal for assembly into production-test instruments, bridges, and other experimental or permanent equipment where only one or two decades are needed. The 510's are completely shielded, both mechanically and electrically, by an aluminum housing, and a knob and dial plate are supplied with each unit.

Decade resistors are difficult to specify adequately, but it is our intention to describe the 1433's performance in the specifications below as completely and truthfully as possible. A discussion of some of the terms we use and why we use them is given in the October, 1965 *Experimenter*, and further information on ac performance is given in the instruction sheet.

— H. P. Hall

A brief biography of Mr. Hall appeared in the June, 1966 *Experimenter*.

Acknowledgement

Mr. Walter J. Bastanier is responsible for improvements in the resistors that have made possible the new, tighter specifications.

SPECIFICATIONS

Long-Term Accuracy: Our two-year warranty applies to the tolerances given below unless the resistor is damaged by excessive current. These tolerances apply for low-current measurement at dc or low-frequency ac (see below).

Over-all Accuracy: The resistance difference between that at any setting and at the zero setting is equal to the indicated value $\pm(0.02\% + 2\ m\Omega)$.

Incremental Accuracy: See table. This is the accuracy of the change in resistance between any two settings on the same dial.

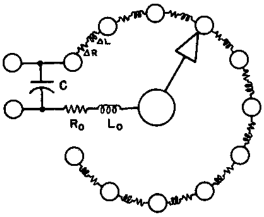
Max Current: The max current for each decade is given in the table below and also appears on the panel of each decade box and on the dial plate of each decade resistance unit.

Frequency Characteristic: The accompanying plot shows the max percentage change in effective series resistance, as a function of frequency for the individual decade units. For low-resistance decades the error is due almost entirely to skin effect and is independent of switch setting, while for the high-resistance units the error is due almost entirely to the shunt capacitance and its losses and is approx proportional to the square of the resistance setting.

The high-resistance decades (510-E, -F, -G, and -H) are very commonly used as parallel resistance elements in resonant circuits, in which the shunt capacitance of the decades becomes part of the tuning capacitance. The parallel resistance changes by only a fraction (between a tenth and a hundredth) of the series-resistance change, depending on frequency and the insulating material in the switch.

Characteristics of the 1433's are similar to those of the individual 510's modified by the increased series inductance, L_o, and shunt capacitance, C, due to the wiring and the presence of more than one decade in the assembly. At total resistance settings of approx 1000 ohms or less, the frequency characteristics of any of these decade resistors are substantially the same as those shown for the 510's. At higher settings, shunt capacitance becomes the controlling factor, and the effective value of this capacitance depends upon the settings of the individual decades.

Typical Values of R_o, L_o, and C for the Decade Resistors:

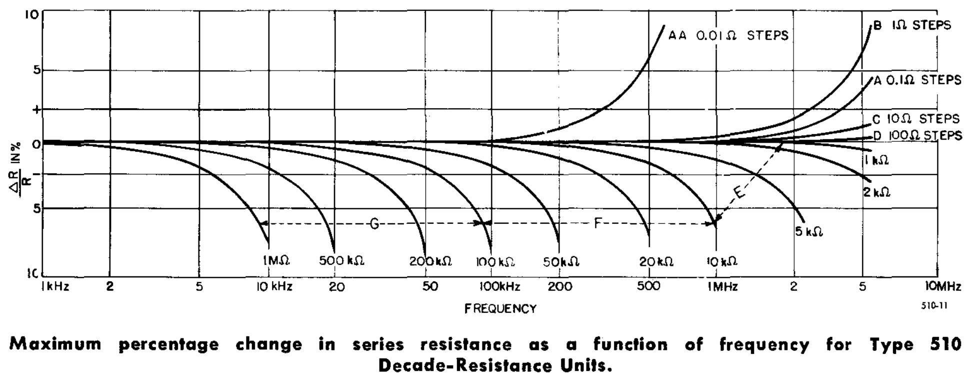


Equivalent circuit of a resistance decade, showing location and nature of residual impedances.

Zero Resistance (R_o): 0.001 Ω per dial at dc; 0.04 Ω per dial at 1 MHz; proportional to square root of frequency at all frequencies above 100 kHz.

Zero Inductance (L_o): 0.1 μH per dial + 0.2 μH.

Effective Shunt Capacitance (C): This value is determined largely by the highest decade in use. With the low terminal connected to the shield, a value of 15 to 10 pF per decade may be assumed, counting decades down from the highest. Thus, if the third decade from the top is the highest resistance decade in circuit (i.e., not set at zero), the shunting terminal capacitance is 45 to 30 pF. If the highest decade in the assembly is in use, the effective capacitance is 15 to 10 pF, regardless of the settings of the lower-resistance decades.



Maximum percentage change in series resistance as a function of frequency for Type 510 Decade-Resistance Units.

SPECIFICATIONS (continued)

Temperature Coefficient of Resistance: Less than ±10 ppm per degree C for values above 100 Ω and < ±20 ppm per degree C for 100 Ω and below, at room temperatures. For the 1433's the box wiring will increase the over-all temperature coefficient of the 0.1- and 0.01-Ω decades.

Switches: Quadruple-leaf brushes bear on lubricated contact studs of ⅜-in. diameter in such a manner as to avoid cutting but yet give a good wiping action. A ball-on-cam detent is provided. There are eleven contact points (0 to 10 inclusive). The switch resistance is less than 0.0005 Ω. The effective capacitance is of the order of 5 pF, with a dissipation factor of 0.06 at 1 kHz for the standard cellulose-filled molded phenolic switch form and 0.01 on the mica-filled phenolic form used in the 510-G and 510-H units.

Max Voltage to Case: 2000 V pk.

Type 1433 Decade Resistors

Catalog Number							*Price in USA*	
Bench	*Rack*	*Type*	*Total Ohms*	*Ohms per Step*	*No. of Dials*	*Type 510 Decades Used*	*Bench*	*Rack*
1433-9700	1433-9701	**1433-U**	111.1	0.01	4	AA, A, B, C	$120.00	$128.00
1433-9702	1433-9703	**1433-K**	1111	0.1	4	A, B, C, D	122.00	130.00
1433-9704	1433-9705	**1433-J**	11,110	1	4	B, C, D, E	125.00	133.00
1433-9706	1433-9707	**1433-L**	111,100	10	4	C, D, E, F	120.00	128.00
1433-9708	1433-9709	**1433-Q**	1,111,000	100	4	D, E, F, G	154.00	162.00
1433-9710	1433-9711	**1433-T**	1111.1	0.01	5	AA, A, B, C, D	146.50	154.50
1433-9712	1433-9713	**1433-N**	11,111	0.1	5	A, B, C, D, E	144.00	152.00
1433-9714	1433-9715	**1433-M**	111,110	1	5	B, C, D, E, F	147.50	155.50
1433-9716	1433-9717	**1433-P**	1,111,100	10	5	C, D, E, F, G	182.50	190.50
1433-9718	1433-9719	**1433-Y**	11,111,000	100	5	D, E, F, G, H	247.50	255.50
1433-9720	1433-9721	**1433-W**	11,111.1	0.01	6	AA, A, B, C, D, E	168.50	176.50
1433-9722	1433-9723	**1433-X**	111,111	0.1	6	A, B, C, D, E, F	166.50	174.50
1433-9724	1433-9725	**1433-B**	1,111,110	1	6	B, C, D, E, F, G	210.00	218.00
1433-9726	1433-9728	**1433-Z**	11,111,100	10	6	C, D, E, F, G, H	276.00	284.00
1433-9729	1433-9730	**1433-F**	111,111.1	0.01	7	AA, A, B, C, D, E, F	191.00	199.00
1433-9731	1433-9732	**1433-G**	1,111,111	0.1	7	A, B, C, D, E, F, G	229.00	237.00
1433-9733	1433-9734	**1433-H**	11,111,110	1	7	B, C, D, E, F, G, H	303.50	311.50

Type 510 Decade-Resistance Units

Catalog Number	*Type*	*Total Resistance Ohms*	*Resistance Per Step (ΔR) Ohms*	*Accuracy of Resistance Increments*	*Max Current 40° C Rise*	*Power Per Step Watts*	*ΔL μH*	*C** pF*	*L_o μH*	*Price in USA*
0510-9806	**510-AA**	0.1	0.01	±2%	4 A	0.16	0.01	7.7-4.5	0.023	$23.00
0510-9701	**510-A**	1	0.1	±0.4%	1.6 A	0.25	0.014	7.7-4.5	0.023	17.50
0510-9702	**510-B**	10	1	±0.1%	800 mA	0.6	0.056	7.7-4.5	0.023	26.00
0510-9703	**510-C**	100	10	±0.04%	250 mA	0.6	0.11	7.7-4.5	0.023	27.00
0510-9704	**510-D**	1000	100	±0.02%	80 mA	0.6	0.29	7.7-4.5	0.023	25.00
0510-9705	**510-E**	10,000	1000	±0.02%	23 mA	0.5	13	7.7-4.5	0.023	20.50
0510-9706	**510-F**	100,000	10,000	±0.02%	7 mA	0.5	70	7.7-4.5	0.023	21.00
0510-9707	**510-G**	1,000,000	100,000	±0.02%	2.3 mA	0.5	—	7.7-4.5	0.023	61.00
0510-9708	**510-H**	10,000,000	1,000,000	±0.02%	0.7*mA	0.5	—	7.5-4.5	0.023	92.00
0510-9604	**510-P4**	Switch only	(Black Phenolic Frame)							11.00
0510-9511	**510-P4L**	Switch only	(Low-Loss Phenolic Frame)							12.00

* Or a max of 4000 V, pk.

** The larger capacitance occurs at the highest setting of the decade. The values given are for units without the shield cans place. With the shield cans in place, the shunt capacitance is from 0 to 20 pF greater than indicated here, depending on whether the shield is tied to the switch or to the zero end of the decade.

The 1218-BV lockable microwave oscillator.

POPULAR UHF OSCILLATOR NOW LOCKABLE

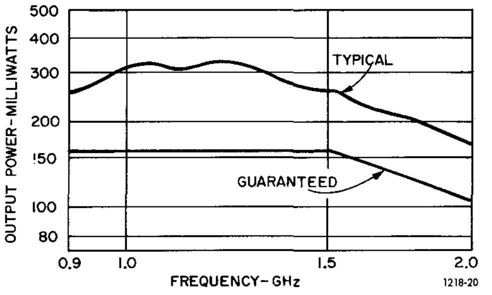
900 to 2000 MHz

The GR 1218 Oscillator has long been valued in many laboratories for its high output power and its spectral purity. Now, the 1218-BV lockable model means that frequency stability too can be achieved by use of a feedback control loop and reference-frequency source.

The oscillator can be phase locked to a stable reference of the same frequency or to a harmonic of the reference signal. Absolute stability of the 1218-BV may be less important than the stability of a difference frequency. In heterodyne systems the 1218-BV, used as the local oscillator, can track small changes in the test frequency. In this way the intermediate frequency is held within the detector passband.

A dc control voltage of ±25 volts will result in a frequency change typically greater than ∓2 MHz. Step-response time is typically under 1 microsecond. As in the 1218-A, output power into 50 ohms is at least 160 mW up to 1.5 GHz, dropping linearly to at least 110 mW at 2 GHz, and, in all other respects as well, performance of the 1218-BV is the same as its predecessor.

The 1218-BV can be used with any of three GR power supplies: 1264-B Modulating Power Supply, 1263-C Amplitude-Regulating Power Supply, or 1267-B Regulated Power Supply. The oscillator is available separately or in combination with any of the above power supplies for bench or relay-rack mounting.



1218-BV power output into 50 ohms.

SPECIFICATIONS

Frequency Range: 900 to 2000 MHz.

Frequency Calibration Accuracy: ±1%.

Warmup Frequency Drift: 0.1% total drift, typical.

Frequency Control: A 4-in. dial calibrated in MHz over 290° (10½-in. scale length); also 800-division logging scale. Slow motion drive of about 8 turns.

ΔF Control (Internal): > ±2 MHz by ⅞ turn of front-panel knob.

Power-Level Pulling (by ΔF control): < ±0.5 dB for ± 2-MHz ΔF.

ΔF Control (Remote): By dc voltage applied at front or rear jacks.

Frequency: > 4 MHz total range for 50-V change.

Voltage: Typical useful range ± 25 V; ΔF control sets center value from +10 to −20 V. Positive-going voltage causes frequency decrease. Applied voltage ±50 V max.

Interface Characteristic: Equivalent to 10 kΩ, 150 pF, and −1.3 mA current source in parallel across terminals, one of which is grounded. Ext source should have < 1000 Ω int impedance; may be ac coupled.

Step-Response Time: < 1 μs, typical.

Output Power (into 50 Ω): >160 mW, 0.9 to 1.5 GHz; drops linearly to >110 mW at 2.0 GHz.

Output Connector: Locking GR874® connector at rear panel. Adaptors available to other connector types.

Level Control: Full output to at least 20-dB attenuation set by uncalibrated front-panel control.

Modulation: AM INPUT jack at front panel for external audio-frequency plate modulation: approx 30 V rms into 6 kΩ required for 30% amplitude modulation. GR 1311 Audio Oscillator recommended as modulator.

Power Supply: Choice of three regulated power supplies. Oscillator is available separately or in combination with any power supply for rack or bench mounting.
The GR 1267-B Regulated Power Supply is suitable for cw operation.
The GR 1263-C Amplitude-Regulating Power Supply automatically holds the output at set level up to 2 V behind 50 Ω, cw or 1-kHz-square-wave modulated.
The GR 1264-B Modulating Power Supply provides full power cw or modulated operation from internal 1-kHz square-wave or external pulse up to 100 kHz.

Mounting: Unit cabinet can be mounted with a power supply in single bench assembly or can be rack mounted alone or with power supply by use of appropriate rack adaptor.

Accessories Supplied: 2- and 3-contact phone plugs for modulation and frequency-control inputs.

Accessories Available: GR874® coaxial elements and adaptors.

Dimensions (width × height × depth): Bench, 12 x 7⅝ x 9 in. (305 x 195 x 230 mm); rack (with 1267 power supply), 19 x 7¼ in. (485 x 180 x 185 mm), 1263 or 1264 power supply adds 7 in. to rack height.

Weight (less power supply): Net, 14 lb (6.5 kg); shipping, 25 lb (11.5 kg).

Catalog Number		*Description*	*Price in USA*	
1218-9724		**1218-BV Lockable Oscillator**	**$825.00**	
Bench	**Rack**	**Oscillator/Power-Supply Combinations**	**Bench**	**Rack**
1218-9901	1218-9902	1218-BV with 1263-C Amplitude-Regulating Power Supply	$1375.00	$1404.00
1218-9903	1218-9904	1218-BV with 1264-B Modulating Power Supply	1240.00	1269.00
1218-9905	1218-9906	1218-BV with 1267-B Regulated Power Supply	1020.00	1046.00

W8MT3VM

W8MT3

NEW MEMBERS OF THE VARIAC®
FAMILY OF ADJUSTABLE AUTOTRANSFORMERS

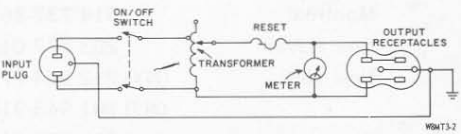
The W8MT3 and its metered companion, the W8MT3VM, provide the most volt-amperes per dollar of any Variac® model, many convenience features, plus the new General Radio "light look."

The W8MT3 and W8MT3VM are cased units that include a power cord and plug, an on-off switch that switches both sides of the input line, a dual receptacle for connection of a multiple load, a manual-reset overload protector, and a carrying handle. The W8MT3VM has a 0-to-150-volt meter that indicates output voltage of the Variac.

A three-wire power cord ties the ground (third) prong on the power plug to the ground-prong receptacle on the load outlet and to the case.

These portable W8-type autotransformers are designed for use at a minimum line frequency of 50 Hz for the stated 120-volt input. They can be operated at up to 400 Hz.

Because of the Duratrak® precious-metal contact surface, these Variac autotransformers can tolerate large momentary overloads without damage and have a comfortable margin-of-safety in normal operation at rated loads.



W8MT3VM circuit. The W8MT3 is the same except that it has no meter.

SPECIFICATIONS

INPUT
Voltage: 120 V.
Frequency: 50 to 60 Hz.
OUTPUT
Voltage: 0 to 140 V.
Rated Current: 10 A; equivalent to 1400 W at max output voltage.
Meter (in W8MT3VM only): 0 to 150 V.
No-Load Loss: 12 W at 60 Hz.
Driving Torque: 10 to 20 ounce-inches.
Replacement Brush: Type VB 2.
Weight: Net, 10 lb (4.6 kg); shipping, 20 lb (9.5 kg).

Catalog Number	*Description*	*Price in USA*
3038-5119	**W8MT3 Variac® autotransformer**	**$38.00**
3038-5015	**W8MT3VM Variac® autotransformer (metered)**	**68.50**
3200-5900	**VB2 Replacement Brush**	**1.00**

GENERAL RADIO COMPANY

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617 369-4400

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