the GENERAL RADIO Experimenter

the GR Experimenter

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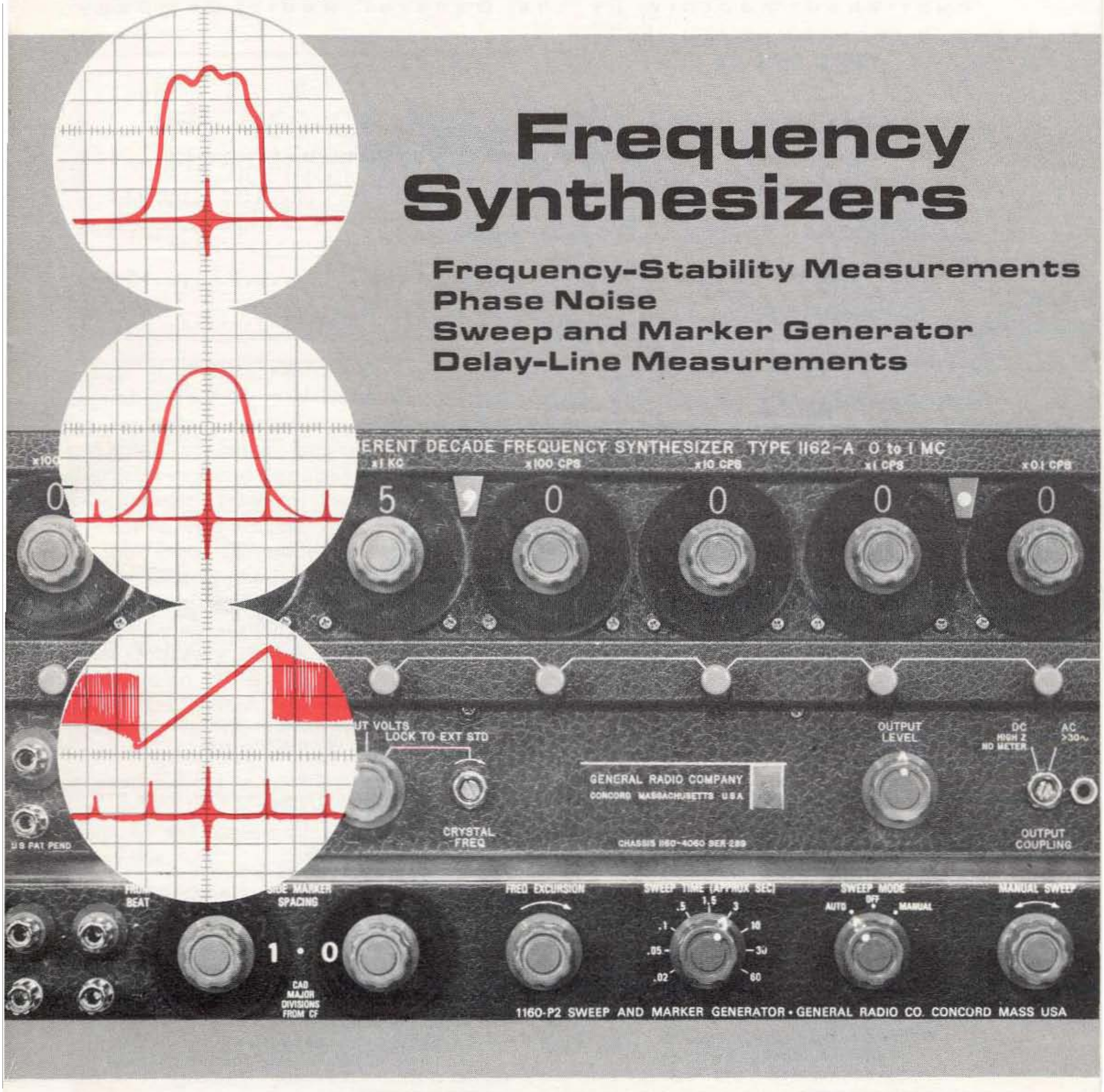

Frequency Synthesizers

Frequency-Stability Measurements
Phase Noise
Sweep and Marker Generator
Delay-Line Measurements

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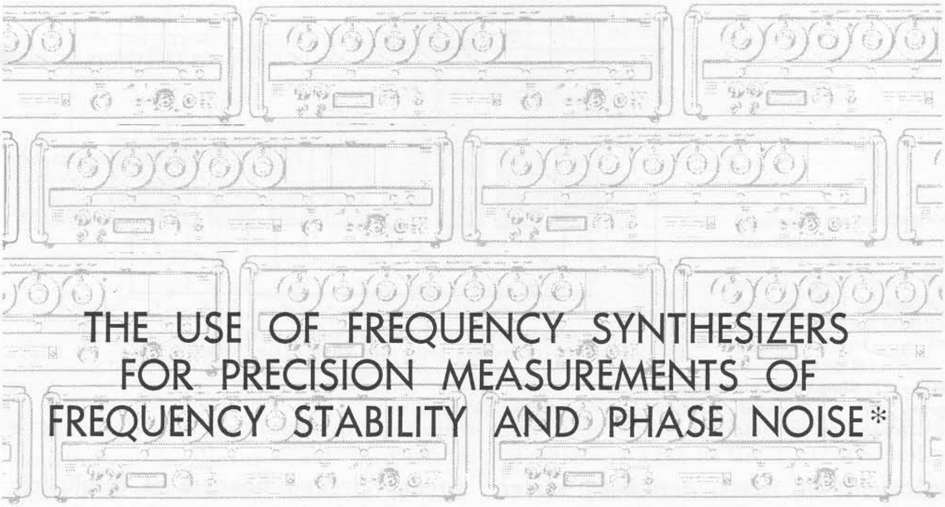

THE USE OF FREQUENCY SYNTHESIZERS FOR PRECISION MEASUREMENTS OF FREQUENCY STABILITY AND PHASE NOISE*

Few new instruments have been accepted as quickly and as widely as has the frequency synthesizer, and few have prompted so many inquiries about their characteristics and capabilities. This month's *Experimenter* responds to the growing interest in synthesizers with four articles: In the first, Dr. Atherton Noyes, who directs GR's synthesizer development, tells how synthesizers can be used to measure the frequency stability and phase noise of other frequency sources. This is followed by a discussion of the phase noise of the synthesizer itself. A new sweep and marker generator for the synthesizer is described in the third article. Finally, the synthesizer is seen solving a problem in production-line testing — the precise, direct measurement of the delay of a solid ultrasonic delay line.

The 1160-series frequency synthesizers, in addition to serving as versatile frequency sources, can be used to track and to record frequency variations in other frequency sources.[1] This is so because the frequency of the continuously adjustable decade (CAD) can be smoothly controlled manually and electrically. By making use of this capability one can maintain exact equality between the frequency under investigation and an adjustable and precisely known standard.

Precision measurements of frequency by establishment of approximate equality between an unknown and a fixed standard and observation of variations in the beat frequency between them have been common practice for many years. For high-precision comparisons, the beat frequency is usually measured in terms of rate of change of phase, over minutes or hours.

The task of measuring phase drift between a *fixed*-frequency standard and an unknown frequency and translating it into frequency error is laborious, and the results often yield only an average over a relatively long time interval. A synthesizer, on the other hand, provides a highly precise standard frequency that can be continuously adjusted to

* Much of the material in this paper was presented at the 21st Annual ISA Conference in New York on October 27, 1966.

[1] Atherton Noyes, Jr., "Coherent Decade Frequency Synthesizers," *General Radio Experimenter*, September 1964, page 11.

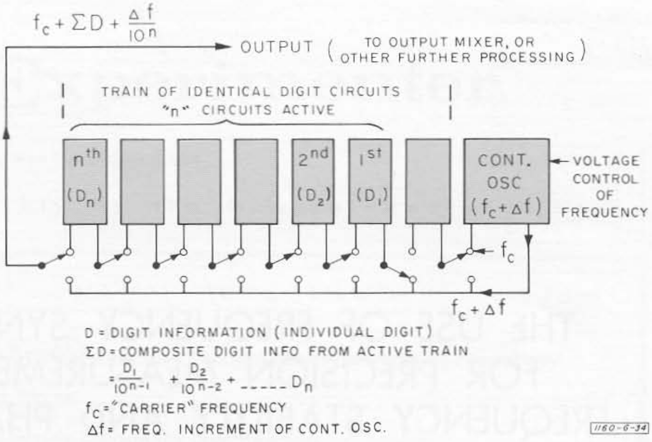

Figure 1. Simplified block diagram of a General Radio frequency synthesizer

maintain an exact, zero-beat match with an unknown. For manual control, a Lissajous figure on an oscilloscope is a convenient indication of exact zero beat (zero rate of change of phase). A frequency difference of only a millihertz, for example, is apparent within a few seconds of observation.

However, it is usually easier to compare the two signals in a phase detector that, for a phase difference of 90°, produces a dc output of zero volts (on which are superimposed the effects of random phase fluctuations — "phase noise" — in both of the compared signals). An unchanging dc component of the phase-detector output indicates that the frequencies are equal; also, any ac components indicate phase noise directly.

By using a voltage-tunable frequency synthesizer one can set up a closed-servo-loop frequency control to maintain zero frequency difference in the presence of drifts and fluctuations of the signal under examination. Most suitable for such applications is a synthesizer that, like the GR 1160-series, uses a series of repetitive circuit modules to synthesize an output frequency to any desired fineness of resolution, and that includes a continuously adjustable voltage-controlled oscillator which can replace a step digit unit at any chosen rank.

Figure 1 is a simplified block diagram of such a synthesizer. Each digit unit processes the signal passing through the train in such a way as to reduce by a factor of 10 any frequency variation occurring in its input signal. So, if the continuous oscillator feeds the first of, say, seven digit units, the variation in output frequency from the train is only $\frac{1}{10^7}$ the variation in the starting signal from the continuous oscillator. A change of 100 Hz in the oscillator thus produces a change of only 100×10^{-7} Hz, or 10 microhertz, at the output.

Conversely, if the synthesizer is constrained to adjust its output frequency smoothly to maintain exact equality with a second, perhaps slightly unstable, frequency by variation of the continuous oscillator, then a drift of only 10 microhertz in the unknown will produce a 100-Hz change in the continuous oscillator.

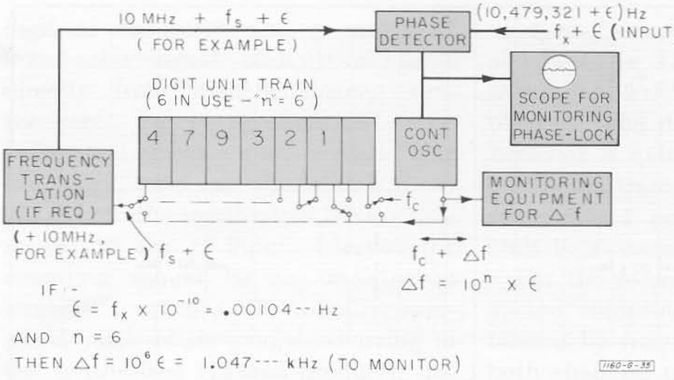

Figure 2. Block diagram illustrating frequency-tracking scheme. ε represents any small frequency drift of the unknown, which is matched and canceled by the servo loop.

Since the total tuning range of the continuous oscillator used in such a system is always the same, no matter which decade the oscillator replaces, simple switching allows selection of the proper magnification for the task at hand. Circuits for monitoring the continuous oscillator are not affected by such switching.

Figure 2 is a diagram illustrating frequency tracking by this technique. The unknown frequency f_x and the synthesized output are here maintained equal in frequency by the phase-lock loop, which includes the voltage-controlled oscillator (*VCO*) at the start of the synthesizer train. Resulting frequency changes in the *VCO* are monitored as shown. Note that a change of 1 part in 10^{10} in f_x (1.04 - - - millihertz) produces a very easily measured change of 1.04 - - - kHz in the *VCO* frequency, in the example illustrated here. In this example, the standard frequency is offset by a fixed frequency of 10 MHz. Such frequency translation is sometimes useful for achieving a frequency match at frequencies outside the band of the synthesizer, or for obtaining higher resolution with the available number of digits in the synthesizer.

There are obviously a great many ways in which Δ*f* of the *VCO* can be measured or recorded. The method we have used at GR is diagrammed in Figure 3. This setup makes use of the fact that, in the 1160 series of GR synthesizers, internal circuits compare the output frequency of the CAD (Continuously Adjustable Decade) with

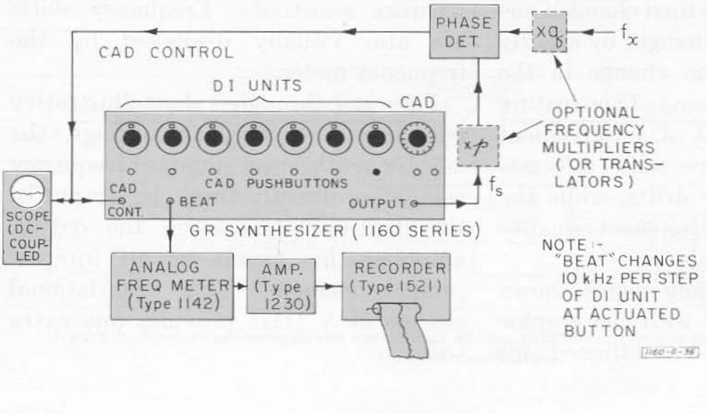

Figure 3. Block diagram illustrating the use of the internal frequency-comparison circuits.

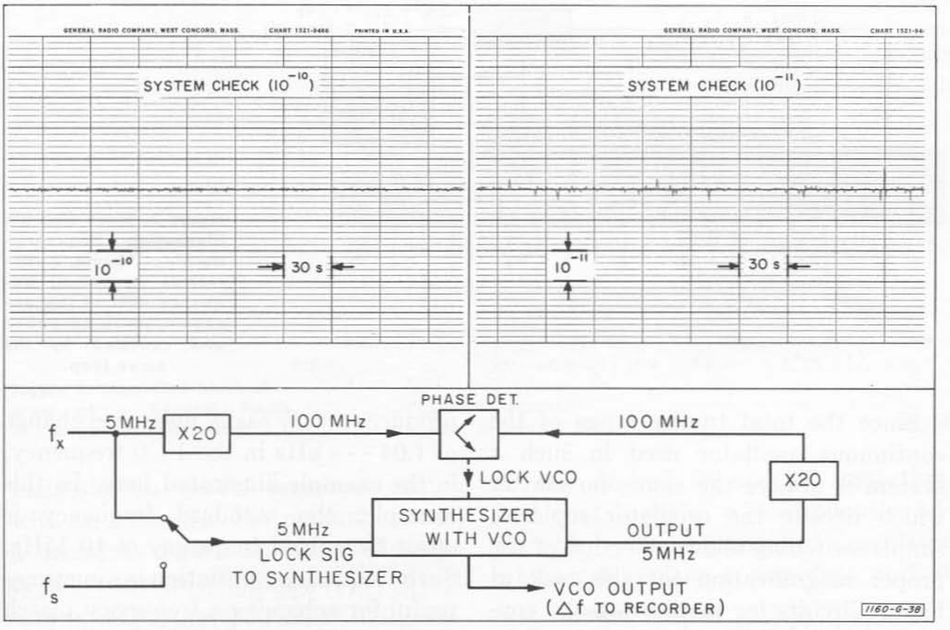

Figure 4. Recordings to check system. Synthesizer lock signal and unknown signal are identical.

the frequency generated by the digit-insertion units that it functionally replaces by push-button selection, and the resulting beat-note is available at the BEAT terminals.[1] The CAD frequency can be tuned manually and can be voltage-controlled about the manual setting. If the frequency of the CAD exactly matches that of the replaced digit units, the BEAT output frequency is zero. If the setting of the digit unit at the actuated button is then changed one step, the beat signal changes by exactly 10 kHz. (There is no change in the main output frequency.) This feature permits establishment of a convenient reference-offset to give sense information about frequency drifts, while the phase detector maintains exact equality of the compared signals.

The analog frequency meter shown in Figure 3, together with the graphic level recorder, can be adjusted for center-of-chart pen positions when the input to the frequency meter is 10 kHz. Hence, positive or negative frequency drifts of f_x are unambiguously traced on the recorder chart, with the chart center line exactly corresponding to the dialed synthesizer frequency. The scale factor of the recording can be changed in decade steps by actuation of the synthesizer pushbuttons and in smaller amounts by adjustments of the recorder sensitivity. Frequency shifts are also visually displayed by the frequency meter.

Figures 4 through 8 show illustrative recordings. For these recordings, the 5-MHz synthesized standard frequency was taken directly from the output of the digit-unit train (via the output amplifier, by means of an internal jumper change). This translational change of 5 MHz provides one extra

[1] *Ibid.*

digit of resolution and a somewhat lower-noise signal than is available directly from higher-frequency synthesizers.

Figure 4 shows a system check. The frequency used as the standard to phase-lock the synthesizer is also connected to the f_x input. Ideally, the recording should be an undisturbed straight line at $\Delta f/f = 0$. Actually, occasional small phase jumps occurring in the synthesized channel produced uncorrelated disturbances recorded as a few parts in 10^{12}.

For each recording of Figure 5, two separate precision oscillators were used, one to phase-lock the synthesizer and the other as f_x. The lock signal f_s was provided by the General Radio Bolton Plant standard, by way of the plant distribution system. For the left-hand trace, a typical 1115-B Oscillator was used as f_x and recorded against this standard. In the right-hand trace, a somewhat less stable oscillator supplied f_x. The difference in short-term behavior is quite apparent. Note that for both traces one major division represents 1 part in 10^{11}. The time scale is 30 seconds per division.

For the recordings of Figures 4 and 5, the recording bandwidth was enhanced by frequency multiplication of both channels to 100 MHz before the phase detector. By this process the phase-lock bandwidth was increased to $\pm\frac{1}{4}$ Hz, at a sensitivity of 10^{-11} per division, and to ten times this, or $\pm 2\frac{1}{2}$ Hz, at 10^{-10}. (To measure lock bandwidth, the manual tuning control of the CAD was moved to the high and low limits of the lock range, as indicated by the phase-detector monitor. This CAD dial range, multiplied by the

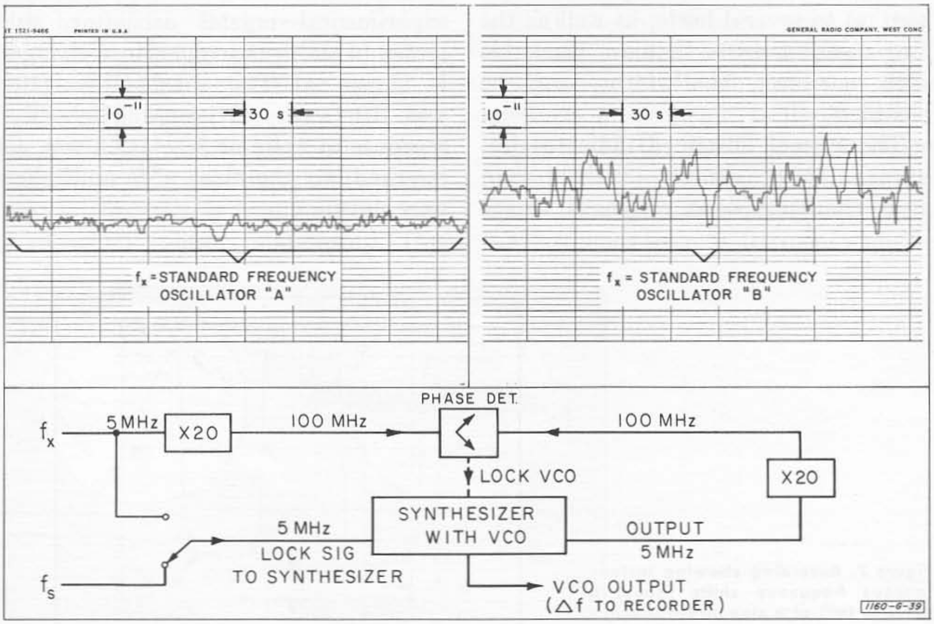

Figure 5. Frequency-comparison recordings of two oscillators relative to a standard used to lock the synthesizer.

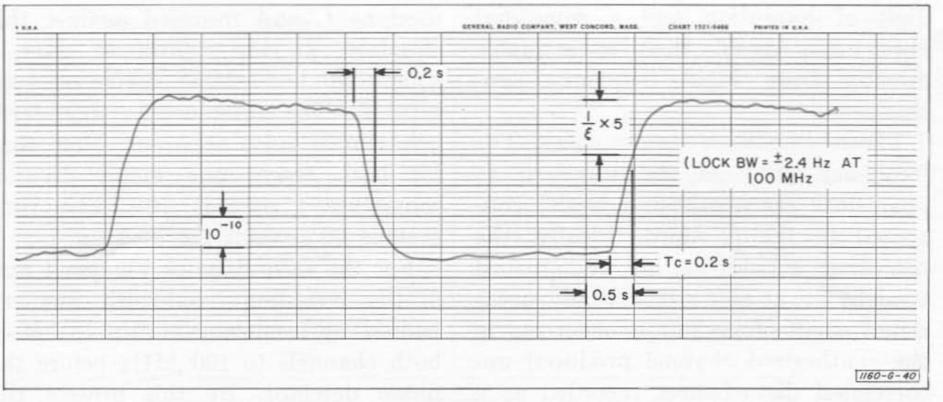

Figure 6. System-bandwidth recording, showing response to abrupt frequency steps of about 5×10^{-10}.

Hz/step at the CAD functional position, gave bandwidth directly.)

Figure 6 shows the response of the system, at 10^{-10} per division sensitivity, to abrupt f_x steps of about 5×10^{-10}. A time constant of about 0.2 second is indicated, so that, at this sensitivity setting, fluctuations at frequencies from zero up to several hertz, as well as the longer-period drifts that are trackable with narrower bandwidths, can be recorded.

The system described indicates frequency shifts *directly,* rather than as a changing slope of phase versus time as is customary in precision measurements. As a further example, Figure 7 shows immediately the frequency shifts that resulted when a precision oscillator was rotated 90° in the earth's gravitational field.

Figure 8 shows a less demanding measurement. The behavior of a simple, experimental crystal oscillator, subjected to ambient temperature changes, is shown at two sensitivities (10^{-8} and 10^{-9} per division). High-speed response at extreme resolution was not required, so the two ×20 multipliers were not used.

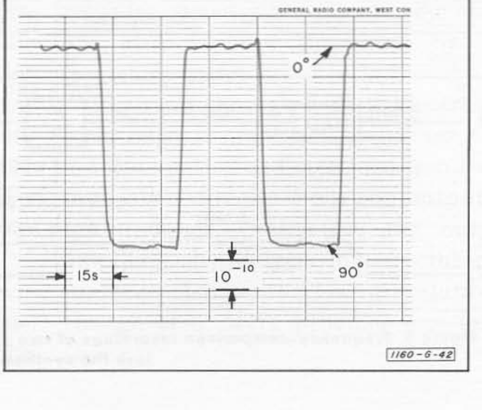

Figure 7. Recording showing instantaneous frequency shifts (about 6 parts in 10^{10}) of a standard-frequency oscillator as it is repeatedly rotated 90° relative to the earth's gravitational field.

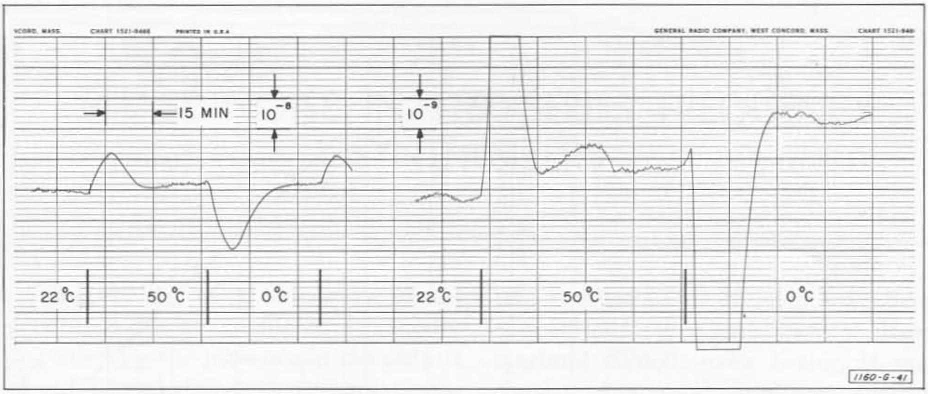

Figure 8. Recordings showing the response of a crystal oscillator as a function of temperature.

Use of the Setup for Phase-Noise Studies

This general setup is also useful for observing short-term fluctuations. The oscilloscope, which in Figure 3 was shown connected as a dc monitor of phase lock at the control-signal output of the phase detector, is also a useful indicator of high-frequency phase noise, at all frequencies outside the phase-lock bandwidth. Such frequencies are not degenerated by the phase lock and appear at the phase-detector output.

The oscilloscope, therefore, can supply quantitative information on phase jitter. The only function of the phase-lock, in this case, is to maintain f_s and f_x at a 90° phase relationship. One should keep the loop bandwidth as low as is consistent with the stability of the oscillator under test, in order to avoid degenerating noise frequencies of interest. It is easy to calibrate the phase deviation, ΔΘ, in terms of the peak value of a beat note created by frequency displacement of one of the inputs, as shown in Figure 9, at the left. If the oscilloscope gain is then increased, under centered phase-lock conditions, the height of the grass measures phase jitter, as illustrated in Figure 9, at the right. Suitable filters placed in the oscilloscope input can limit the noise bandwidth, as desired. For low-noise sources, frequency multi-

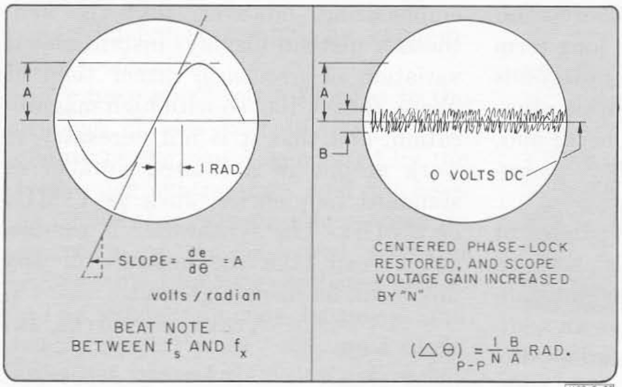

Figure 9. Oscilloscope calibration for phase noise.

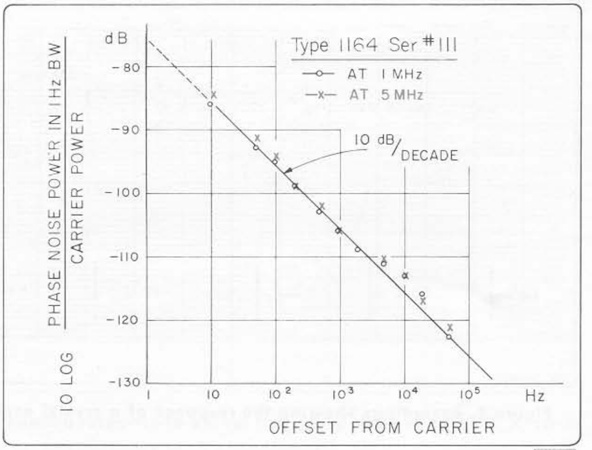

Figure 10. Single-side phase-noise power-density, referred to carrier.

plication before the phase detector (as previously described) is desirable, in order to magnify the noise prior to detection.

A high-gain selective voltmeter connected to the phase-detector output can be used to observe sideband noise distribution. Figure 10 shows a plot of phase-noise density, as a function of offset from signal, taken with a GR 1900 narrow-band wave analyzer. More details on this type of measurement are given in an accompanying article.[2]

CONCLUSION

Systems such as those described above can thus be used to observe and to record not only relatively long-term drifts over periods ranging from seconds to days but also shorter-period fluctuations at rates up to several hertz and, additionally, higher-frequency, noise-sideband power distribution.

The synthesizer, used as an error magnifier, can supply the *total* standard signal, either directly or with multiplication, or it can be inserted as an additive component, as circumstances dictate. For measurements of very high-quality signals, translation by frequency addition is to be preferred, for two reasons: First, the added frequency component can perhaps have somewhat lower noise than the synthesized component, so that the composite signal may be significantly better than one obtained by straight multiplication; second, adding the synthesizer signal to a large fixed component will make available finer fractional resolution than the synthesizer has itself.

The techniques described above have become available with the advent of the synthesizer. Equally good, and perhaps better, results can be achieved by other methods. It is worth re-emphasizing, however, that the synthesizer method displays instantaneous variation of *frequency*, rather than of phase, that it does so with high magnification, and that it is not necessary to work at one of a limited number of standard frequencies, such as 1 MHz or 5 MHz. The synthesizer is nimble, and it can take the measure of any unknown, on its own ground.

— Atherton Noyes, Jr.

[2] "Phase Noise in 1160-Series Frequency Synthesizers," this issue, page 11.

PHASE NOISE IN 1160-SERIES FREQUENCY SYNTHESIZERS

The quality of a sine-wave signal generated by an oscillator is usually degraded by the presence in the output of additional frequency components, which can be classified in the following groups:

(1) Harmonics of the desired frequency.

(2) Sidebands accompanying the desired frequency, created by modulation of the signal by power-line frequency components (60 Hz, 120 Hz, etc).

(3) Other nonharmonically related sideband components. (In frequency synthesizers, these are generally due to the heterodyning of high-order harmonics of signals applied to one or more of the synthesizer frequency mixers. Such harmonics are usually generated in the mixer itself. Spurious sidebands can also be created by signals leaking from one part of the circuit to another, owing to ground loops or imperfect shielding.)

(4) Sidebands created by random-noise modulation.

We have given much thought to the problem of how best to describe the quality of the signals generated by the 1160-series synthesizers, and we have concluded that it is sensible to separate discussion of the undesired sideband components by classes as listed above.

The single-frequency harmonic and nonharmonic "spurs" are most easily measured by means of a sharply selective receiver. This receiver may be of conventional type, such as the Hammarlund SP600; even better, it may consist of a wave analyzer, such as GR's TYPE 1900, used as a tunable i-f amplifier and output indicator, after an external mixer. For stability, the local oscillator of this composite receiver should be a synthesizer, preferably with its 5-MHz internal oscillator phase-locked to the synthesizer being tested. This technique, employing such a highly selective receiver, serves well to make the single-frequency measurements of harmonics, power-line-related sideband components, and nonharmonically related spurs (Classes 1, 2, and 3 above). Specifications on these have already been published in the General Radio Catalog and in the *Experimenter* (September 1964, November-December 1965, and September 1966).

The TYPE 1900 Wave Analyzer can also serve as a high-gain selective volt-

Atherton Noyes received his undergraduate and graduate degrees (AB, AM, SM, ScD) at Harvard University. From 1937 to 1960 he was on the engineering staff of Aircraft Radio Corporation. He has directed GR's frequency-synthesizer development since 1960, and he is currently the Group Leader of the Signal-Generator Group.

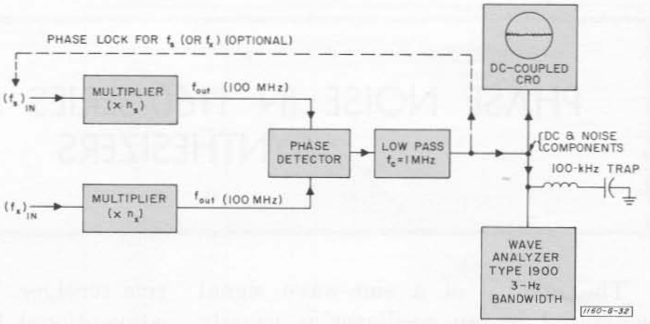

Figure 1. Equipment setup for the measurement of phase-noise density.

meter for measuring random-noise components at any chosen spacing from the carrier. Measurements made on representative groups of 1162, 1163, and 1164 synthesizers by this method are reported here. Figure 1 shows the experimental setup in block-diagram form. This is merely a variation of the selective receiver suggested above, in which the local-oscillator frequency and signal frequency are identical, and the tunable i-f amplifier and output indicator measures *both* sidebands corresponding to any component of frequency modulation of the signal. (The 100-kHz trap at the 1900 input ensures that any 100-kHz sidebands in the measured signal are not able to leak past the input circuits into the 100-kHz fixed-tuned amplifier of the Wave Analyzer in sufficient magnitude to affect the measurement.)

As shown in Figure 1, two precision frequencies are each multiplied to 100 MHz (to increase sensitivity, as explained below) and compared in a phase detector. When the two 100-MHz signals are in phase quadrature, the dc output of the phase detector is zero, and the noise in the phase-detector output is "phase noise" (i.e., phase jitter will produce an output; amplitude jitter will produce no output). The dc-coupled oscilloscope serves to establish the fact that phase quadrature (zero dc) has been achieved and to examine the character of the noise. (Amplitude noise can be measured in the same setup if the 100-MHz signals are brought to equal, instead of quadrature, phase. In general, amplitude noise is 10 dB or more lower than phase noise.)

When the two frequencies are derived from two similar sources, the phase-noise power observed may be assumed to be twice that of a single source; when one frequency is less noisy than the other by 10 dB or more, its contribution to the observed noise is, for practical purposes, negligible.

The GR 1900 Wave Analyzer is used to explore the phase-noise distribution. A plot of phase-noise power in a 1-Hz-wide band, as a function of the spacing (in frequency) of this band from the carrier is given in Figure 2 for the Types 1162, 1163, and 1164 synthesizers. The noise of the reference signal is so low that it can be neglected. We have used a Type 1115-B Standard-Frequency Oscillator as a reference at times, and also the 5/5.1 output of a Type 1161. In the latter case, the CAD can be used for narrow-band phase lock for convenience.[1]

[1] As described more fully in a companion article, this issue, page 3, "The Use of Frequency Synthesizers for Precision Measurements of Frequency Stability and Phase Noise."

The plots of Figure 2 are calculated from experimental data in accordance with the following facts:

(1) When a frequency is multiplied by a factor n, the magnitude of any small-amplitude sideband grows, relative to the carrier amplitude, by the same factor. The "dB down" value measured at 100 MHz must therefore be increased by 20 $\log_{10}$ (n) to refer the noise to the actual synthesizer frequency used. If n = 20, the frequency correction is 26 dB, and for n = 100 the correction is 40 dB.

(2) The noise power in a narrow bandwidth is proportional to the bandwidth (with white noise this is strictly true; with other types of noise, such as the $1/f$ distribution that we observe for synthesizers, it is very close to the truth, provided the bandwidth is small compared with the offset frequency). To check this fact experimentally, the 3-, 10- and 50-Hz bandwidths of the 1900 Wave Analyzer were used at a relatively large displacement from the carrier, and it was confirmed that such calculations from the data of each measurement produced identical values for 1-Hz bandwidth, within experimental accuracy. For the 3-Hz bandwidth of the 1900, then, the dB correction to convert to 1-Hz bandwidth is 10 $\log_{10}$ 3 = 4.77 dB (rounded out to 5 dB). The 3-Hz bandwidth of the 1900 was used to obtain data for Figure 2.

(3) Measurements made with the circuit of Figure 1 include noise from both the upper and lower sidebands (i.e., the intermediate frequency is zero) and corresponding noise components are coherent. That is, we are actually measuring the phase displacement "Modulation Index" β, which at any frequency (at low modulation index) produces an upper and lower sideband, each having an amplitude relative to the carrier of $\beta/2$. Since the associated noise power is proportional to the square of the sideband amplitude, the single-sided noise power is 6 dB lower than the double-sided value measured in the circuit of Figure 1.

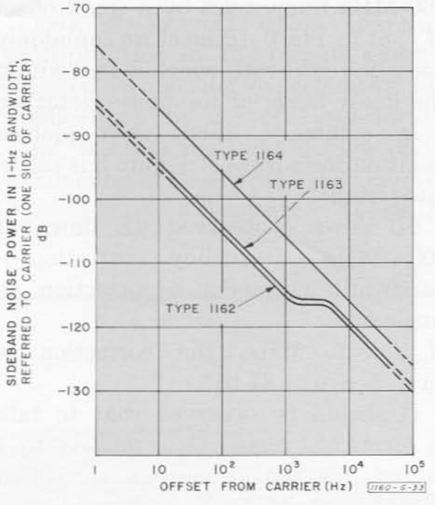

Figure 2. Phase-noise power-density distribution in GR 1160-series synthesizers.

(This fact can be experimentally verified by observation of a single discrete component. For instance, if, in the circuit of Figure 1 with $f_s = f_x$, we measure a 120-Hz component as, say, -66 dB, we can then offset f_s or f_x to produce a beat note in the mixer of, say, 10 kHz and measure the sideband spurs, relative to this 10-kHz beat note, at 10,120 Hz and at 9,880 Hz. These now measure -72 dB.)

It has become popular to show "single-sided" noise curves like those of Figure 2, so we have adopted this method of presentation. The 0-dB reference level used is the TYPE 1900 indication of the level of a beat note deliberately set up between the two

100-MHz frequencies by a small offset of f_x or f_s. The 0-dB level we commonly use is 250 mV, rms, which is well within the linear range of the phase detector.

In summary, then, each plotted point on the curves of Figure 2 is calculated from:

dB down = observed dB down at 100 MHz + frequency correction + bandwidth correction + correction to single side.

If $f_x = 5$ MHz, the correction is $26 + 5 + 6 = 37$ dB.

It should be observed that to talk in terms of "phase noise relative to 1 radian," the 6-dB single-side correction should be omitted.

It should also be noted that the curves of Figure 2 apply at any frequency within the range of the synthesizer. For instance, Figure 10 on page 10 shows a series of points taken at 5 MHz (n = 20) and another series of points at 1 MHz (n = 100) with a typical TYPE 1164 synthesizer. Note that both sets of points fall on the same plot. This condition exists because the synthesizers are beat-frequency oscillators, and what we are really measuring is the noncoherent noise on the inputs to the final output mixer.

The 1164 plot is about 8-dB noisier than the plots for the 1162 and 1163. However, the 1164 goes to much higher frequencies, so that, if the output is going to be multiplied to some still higher frequency (like 1 or 2 GHz), the 1164, with its higher possible starting frequency, will generally come out ahead. If the synthesizer frequency is to be added to some less noisy precision microwave frequency, the 1162 or 1163 will produce a slightly better result, provided the available frequency range is adequate and the combining circuits can be properly worked out.

A striking fact exhibited by Figure 2 is that the noise power density falls off inversely with offset frequency. Note that the curves in Figure 2 have a falling slope of 10 dB per decade, or 3 dB per octave. (In the 1162 and 1163 there are some small anomalies in the region beyond a 2-kHz offset, which are not yet satisfactorily explained.)

Since this is experimentally true, it is convenient to define the noise power density very simply, for comparison purposes, by a single number, namely the "dB down" figure at the intercept of the 10-dB-per-decade line with the 1-Hz offset ordinate. From Figure 2, these figure-of-merit numbers are:

Synthesizer	*Noise Power Density at 1-Hz Offset*
1164	−76 dB
1163	−84 dB
1162	−85 dB

Given this number, one can at any time reconstruct the curve, accurately enough for most purposes, by drawing a line with 10-dB-per-decade negative slope through the tabulated point.

It should be noted that the curves of Figure 2 show the average of measurements on eight or ten production synthesizers of each type. Individual measurements vary about ±2 dB around these averages.

Note Regarding Type 1161

No data on the TYPE 1161 are presented here, because we have not as yet made measurements in this type of setup. Presumably the data, when obtained, will show the same slope, and a 1-Hz intercept 20-dB lower (theoretically) than the TYPE 1162, or about −105 dB. (This is not yet a measured value, however.)

For most purposes, noise data on the 1161 are not as important as data for the others, since such information is chiefly of interest where frequency multiplication of the output signal is planned. We do not expect many such applications for the 1161 (or for the 1162, for that matter), in view of the availability of the TYPES 1163 and 1164.

— ATHERTON NOYES, JR.

Figure 1. Type 1160-P2 Sweep and Marker Generator.

A SWEEPER FOR GR SYNTHESIZERS

The introduction of the 1160-P2 Sweep and Marker Generator (Figure 1) extends the usefulness of the already versatile General Radio Synthesizers by providing a convenient means of varying the synthesizer output frequency at a controlled known rate and through an accurately known range. The instrument also supplies markers for easy visual monitoring of the continuous band of frequencies generated. Several additional features facilitate frequency-response testing of active and passive networks and frequency-selective instruments. These include a wide choice of automatic sweep speeds, a calibrated center-frequency marker, accurate adjustable side markers, and a simple means of expanding the sweep coverage about any stable synthesized center frequency.

To understand the way in which the sweeper operates, remember that the continuously adjustable decade (CAD)[1] in GR synthesizers can be functionally substituted for all step-digit modules below a chosen rank, and that the frequency of the CAD can be varied by an external control voltage of approximately 0.3 volt per CAD major division. In each synthesizer, built-in monitor circuits permit calibration of the CAD frequency against the frequency generated by the replaced digit modules. In addition, the deviation of the CAD frequency from that represented by the

[1] Atherton Noyes, Jr., "Coherent Decade Frequency Synthesizers," *General Radio Experimenter*, September 1964.

replaced digit dials is continuously monitored and is available at the synthesizer BEAT terminals. This continuous monitoring can be used to produce an accurately defined center-frequency marker and side markers indicating known deviations from center frequency.

In use, the Sweep Unit is connected to any GR synthesizer and to a display device. A storage oscilloscope is especially desirable at lower sweep rates. All input and output connections are available in parallel at both the front and rear of the Sweep Unit, to suit the requirements of any setup.

The sweep voltage supplied to the EXT CAD control input to vary the CAD frequency has a triangular waveform. The resulting sweep amplitude is continuously adjustable from ±1 CAD division to ±10 CAD divisions by the FREQ EXCURSION control. A synchronized sweep voltage, of constant amplitude, is available for horizontal deflection of the display device. The triangular, or two-way, sweep control permits the user to observe the effects of the sweep time on the response of the device being tested. Nine one-way sweep times, ranging from 20 milliseconds to sixty seconds, are provided. If narrow-band devices are swept at a faster rate than the response time will accept, a characteristic leaning of the response display will be easily detected visually, since the forward and backward traces will lean in opposite directions and fail to coincide.

When the function switch is in the MANUAL position, the frequency sweep and the synchronized deflection-voltage sweep are manually controlled, and the sweep follows the position of the MANUAL control. This mode is sometimes useful for close examination of a critical portion of a response. With the CAD FREQ EXCURSION control, one can change the swept frequency coverage *without changing the display width and without affecting the selected center frequency*. Markers always indicate actual frequency and move as required when sweep excursion is changed.

An internal oscillator, step-adjustable by two decade controls, provides a reference signal for generating side markers. This marker oscillator is calibrated in terms of the CAD dial divisions departure from the synthesizer center frequency, as chosen by the operator. The synthesizer BEAT output always changes by 10 kHz for each major CAD division swept.[1] The variation in output frequency correspond-

[1] *Ibid*

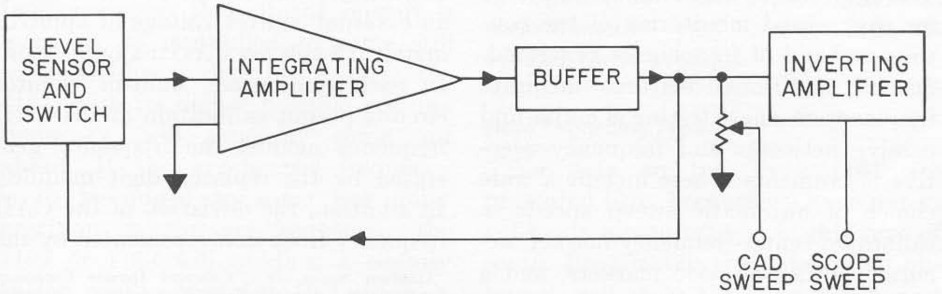

Figure 2. Block diagram of sweep-generator section.

Robert L. Moynihan is a graduate of Northeastern University, where he received the BSEE degree in 1959 and MSEE degree in 1961. Before joining GR, he served for two years in the US Army Signal Corps working on advanced radar R&D, was an instructor of electrical engineering at Northeastern, and a senior engineer at Sylvania Electronic Systems, where he developed subsystems for use in phased-array radar. At General Radio, he is a development engineer in the Decade-Frequency Section of the Signal-Generator Group.

ing to a CAD dial division, on the other hand, can be selected in decade steps with the synthesizer pushbuttons. The marker oscillator generates frequencies from 10 to 59 kHz in 1-kHz steps (each corresponding to one minor division of the CAD dial). The marker-oscillator signal is mixed with the BEAT signal of the synthesizer, and markers occur whenever the BEAT and the marker oscillator frequencies are equal.

Markers appear at symmetrical spacings above and below the center-frequency marker. The center-frequency marker occurs when the BEAT frequency itself is zero and is distinguishable by its bipolar characteristic on the display. It always indicates the point at which the swept frequency is exactly the frequency set by all the synthesizer digit units (including those functionally replaced by the CAD). As the sweep rate is raised, the side markers have less time for buildup and gradually disappear. The center-frequency marker holds position* and always serves as a reassurance that the center frequency is as selected. Side markers can be checked for position by a momentary reduction in sweep speed.

Markers cannot be placed closer to the center frequency than 1 CAD division. Such close-in markers are unnecessary because of the extremely good linearity of the CAD sweep below 1 division. To subdivide this frequency region, one merely positions the 1-division marker at a suitable line on the oscilloscope graticule (by adjustment of the oscilloscope horizontal gain) and uses the graticule divisions as vernier markers.

HOW IT WORKS

Sweep-Generator Section

The sweep-generator section is shown in Figure 2. The sweep-voltage output is obtained by integration of a step of voltage in an operational amplifier with capacitive feedback. The differential-input section is balanced to ground so that either positive or negative inputs are amplified and integrated with polarity retained. The output of the amplifier is monitored by a level sensor which, at preset positive and negative voltage extremes, applies a reversed-polarity step of voltage to the integrator. Sweep rates are adjusted by variation of input step amplitude and of the integration time of the amplifier. The amplitude of the resulting triangular output-voltage waveform is maintained constant by the level sensor. Subsequent buffering and adjustable attenuation result in controlled-rate variable-amplitude sweep voltage. In a second channel, an inverting amplifier is used to obtain a synchronized horizontal-deflection sweep signal of constant amplitude and opposite polarity. (Inversion is needed, because the synthesizer frequency increases with *negative* sweep voltage,

* At high sweep speeds the center-frequency marker changes character but remains symmetrical, so that its center is still well defined and obvious.

whereas oscilloscopes deflect to the right for *positive* voltage.)

Marker Section

The reference oscillator for the frequency markers is a transistorized Wien-bridge oscillator (Figure 3), whose frequency is determined by the relationship $f_o = G/2\pi C$ when the resistance and capacitance values in the series arm equal those in the shunt arm of the Wien bridge. The frequency of the oscillator is controlled by simultaneous variation of the conductances of the shunt and series arms. This is accomplished by the switching of two parallel banks of precision resistors in shunt to obtain multiples of G_o (10 kHz = $G_o/2\pi C$). One switch selects multiples of G_o and the second selects 0.1-G_o increments. The oscillator produces frequencies from 10 to 59 kHz in 1-kHz increments with 1% tolerance. These frequencies correspond to CAD dial divisions ranging from 1 to 5.9 in 0.1-division steps. A negative-temperature-coefficient thermistor in the negative-feedback divider of the bridge keeps the output amplitude constant at all frequencies.

The oscillator output is mixed with the BEAT signal from the synthesizer, and subsequent narrow-band filtering and rectification produces sharp, accurate markers.

APPLICATIONS

Since the synthesizer design is such that one CAD division can correspond to an output frequency range anywhere from 0.001 Hz to 100 kHz as chosen by pushbutton, the system provides an extremely wide choice of sweep widths, sweep rates, and marker spacing about a selectable crystal-controlled center frequency. Slow sweep rates are required with narrow-band devices so that the full amplitude of the response can be reached. A crystal filter, for example, is energized by the source during only that part of the sweep range when the output is within the

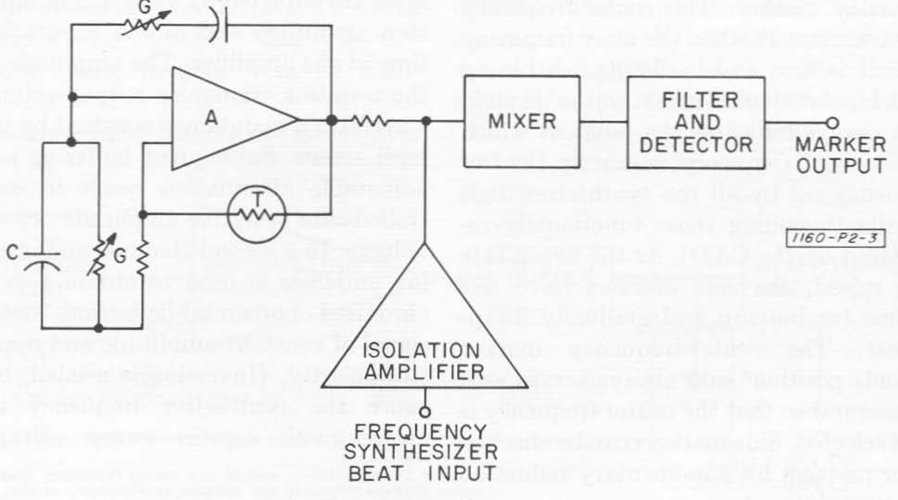

Figure 3. Diagram of marker-generator section.

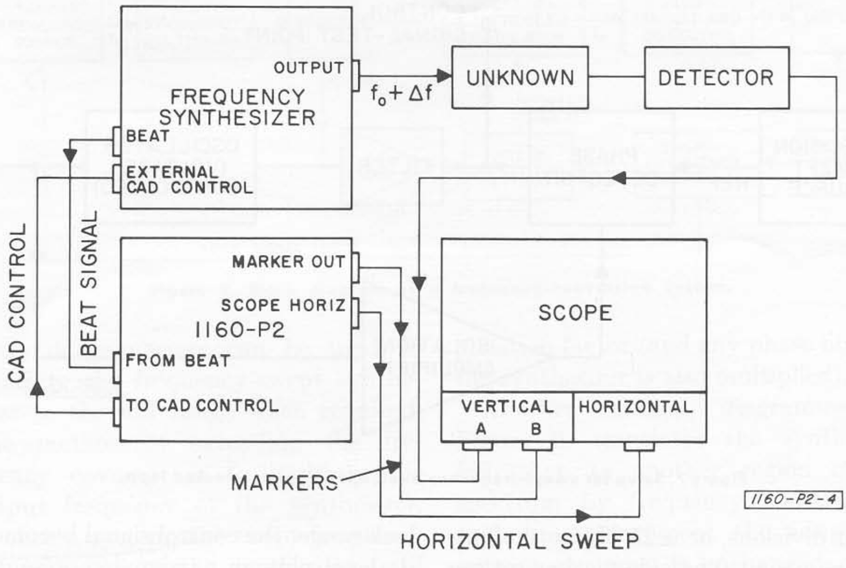

Figure 4. Block diagram of sweep unit and synthesizer used as a precision swept source.

bandpass of the device. Narrow sweep widths are necessary to keep the signal within the bandpass for a reasonable percentage of the sweep time and to yield an expanded display for convenient analysis. The level output and constant source impedance of the synthesizer are also important factors in achieving useful and reliable response testing of crystal and mechanical filters and frequency-selective instruments.

Figure 4 is a block diagram of the sweep unit and synthesizer combined as a precision swept source. The response of a 50-MHz crystal filter is shown in Figure 5. The CAD was in the ×10-kHz position and the side markers were set for ± 1 CAD major division (i.e., ± 10 kHz at the output frequency). The response of the GR 1900 Wave Analyzer is shown in Figure 6. The CAD was in the 10-Hz position and was swept ±2½

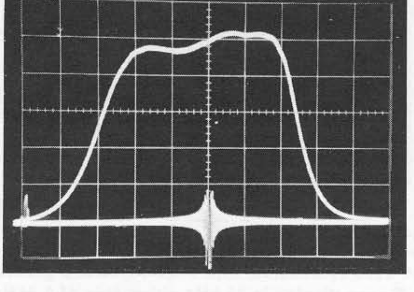

Figure 5. Response of 50-MHz crystal filter. Horizontal scale: 2 kHz/cm. Center frequency marker: 50.00000 MHz.

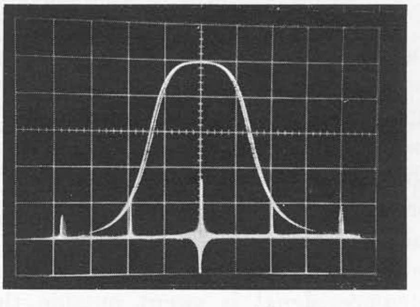

Figure 6. Response of GR 1900 Wave Analyzer tuned to 10 kHz and set for 10-Hz bandwidth. Horizontal scale: 5 Hz/cm. Center frequency marker: 10,000.00 Hz.

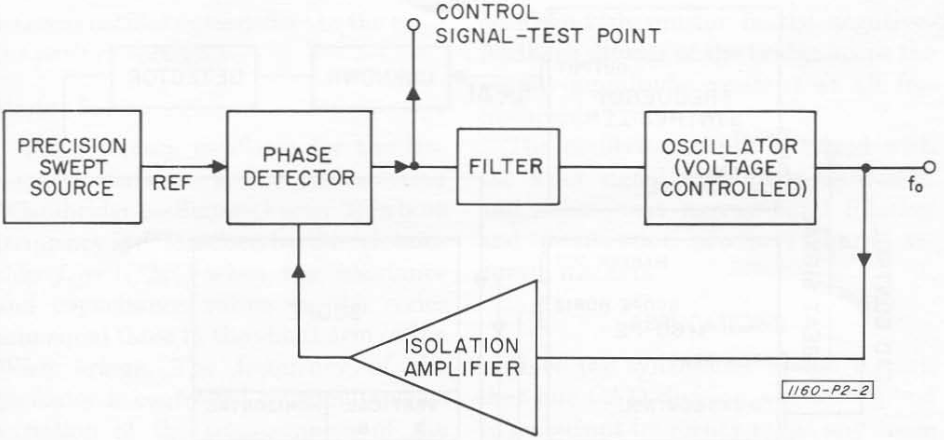

Figure 7. Setup for swept-frequency evaluation of phase-locked loop.

major divisions, or ±25 Hz at the output frequency. The side-marker setting was 1 CAD major division. Markers at ±20 Hz also appear; they are created by the mixing of second harmonics of the BEAT and marker-oscillator signals.

Another application involves the study of a phase-locked oscillator, a device often used as an active bandpass filter in frequency multipliers and phase-locked telemetry receivers. Proper operation of such oscillators is established by measurement of several parameters, including lock range, capture range,[1] and loop stability. Figure 8 is an elementary block diagram of a possible test setup using the precision frequency swept-source. The reference signal is swept at a controlled rate and width, and the control signal of the lock loop is monitored, providing useful information regarding lock-loop performance and stability. The control signal during lock is a ramp. When the reference signal is swept outside the lock range, the control signal becomes a dc level plus an ac signal representing the beat between the reference and the free-running oscillator. The display (Figure 8) is easy to interpret visually, and the results of adjustments can be seen immediately. Loop bandwidth can be measured and center frequency adjusted on the production line.

Measurements need not be confined to the frequency range of present synthesizers. Suitable wide-band multi-

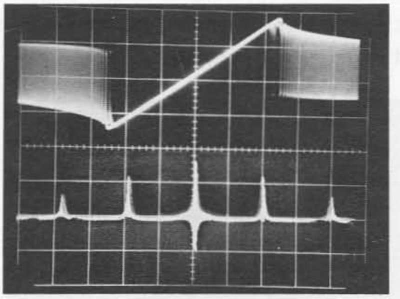

Figure 8. Oscilloscope presentation resulting when 5-MHz phase-locked oscillator was swept with 1163 synthesizer and 1160-P2 sweep generator. Sweep display is 25 kHz either side of 5 MHz; markers are at 10-kHz spacing. Lock range (equals capture range) is approximately ±12 kHz about 5-MHz center frequency.

[1] McAleer, H. T., "A New Look at the Phase-Locked Oscillator," *The Proceedings of the IRE*, Vol 47, June 1959, pp 1137–1143.

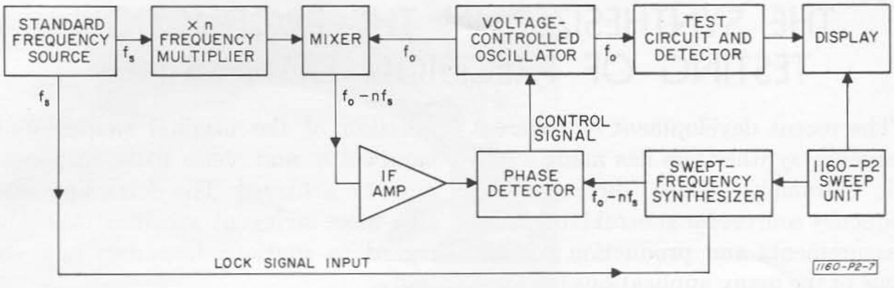

Figure 9. Block diagram of a frequency-conversion system.

pliers or converters can be used to translate the frequency-swept synthesizer to the uhf range when required. One method of extending the frequency coverage is to multiply the output frequency of the synthesizer. In such a system, the swept frequency excursion, Δf, of the synthesizer is of course increased by the chosen multiplication factor (and any phase noise of the synthesizer is also multiplied).

Another method, diagrammed in Figure 9, translates the synthesizer frequency to another region of the spectrum by frequency addition. In such an arrangement the swept frequency excursion is unchanged.

— R. L. Moynihan

SPECIFICATIONS

AUTOMATIC SWEEP

Sweep Voltages: Symmetrical triangular waveforms centered on 0 V dc.

Time for One-Way Sweep: 0.02 to 60 s, automatic, selectable in 9 steps. **Sweep-Time Accuracy:** ± 10%.

Outputs: CAD sweep is continuously adjustable from ±1 to ±10 major CAD divisions (±0.3 to ±3 V approx). The SCOPE HORIZ output is nonadjustable (12 V, p-to-p, behind approx 10 kΩ).

MANUAL SWEEP

Outputs: Sweep excursions are the same as in automatic mode, with continuous manual control.

MARKERS

Location: Center marker occurs at the frequency set on synthesizer digit dials. Side markers are displaced symmetrically from the center marker by the amount set on the MARKER SPACING dials.

Side Marker: Spacing from center marker can be from 1 to 5.9 CAD divisions in 0.1 steps.

Accuracy: ±1% of dial setting.

GENERAL

Power Required: 100 to 125 or 200 to 250 V, 50-400 Hz, 3W.

Ambient Temperature: 0 to 50°C.

Accessories Supplied: Two 2-foot (Type 1160-0320) and two 4-foot (1160-0321) BNC coaxial patch cords, CAP-22 3-wire power cord, spare fuse.

Terminals: Connections to synthesizer and MARKER OUT and SCOPE HORIZ outputs available front and rear.

Cabinet: Rack-bench. End frames for bench mount and fittings for rack mount are included.

Dimensions: Bench model — width 19, height 2, depth 14½ inches (485, 52, 370 mm), over-all; rack model — width 19, height 1¾, depth behind panel 13¼ inches (485, 43, 330 mm).

Weight: Net, 12 lb (5.5 kg); shipping, 16 lb (7.5 kg).

Catalog Number	*Description*	*Price in USA*
1160-9600	**1160-P2 Sweep and Marker Generator**	**$495.00**

THE SYNTHESIZER IN THE PRODUCTION TESTING OF PRECISION DELAY LINES

The recent development of coherent frequency synthesizers has made available extremely precise, direct-reading frequency sources for general laboratory measurements and production testing. Some of the many applications for these versatile instruments have been mentioned in earlier *Experimenter* articles. The following is a brief description of the interesting way in which one manufacturer uses a synthesizer to calibrate ultrasonic delay lines.

Delay-Line Applications

Delay lines are widely used to provide transient, readily accessible memories, either analog or digital, in computers and in systems for processing radar echoes (e.g., video pulse integrators and moving-target indicators).

In these applications, delay lines must store pulses for long times — several milliseconds — with very little degradation of the original shape. Wide bandwidth and very little dispersion must be achieved. The delay line must also meet stringent specifications with regard to spurious responses and stability.

The Solid Delay Line

These requirements are met by solid ultrasonic delay lines using fused quartz or glass as the propagating medium. Solid media propagate both longitudinal (compression) and transverse (shear) acoustic waves. Because of its lower velocity, the transverse mode is utilized in delay lines. The velocity of transverse waves in fused quartz is 0.148 in/μs or 12,300 ft/s, about 11 times the speed of sound in air. A 30-MHz signal traveling with this velocity has a wavelength of 0.00493 inch.

For short delays, up to 50 μs, a single bar is used with a transducer on each

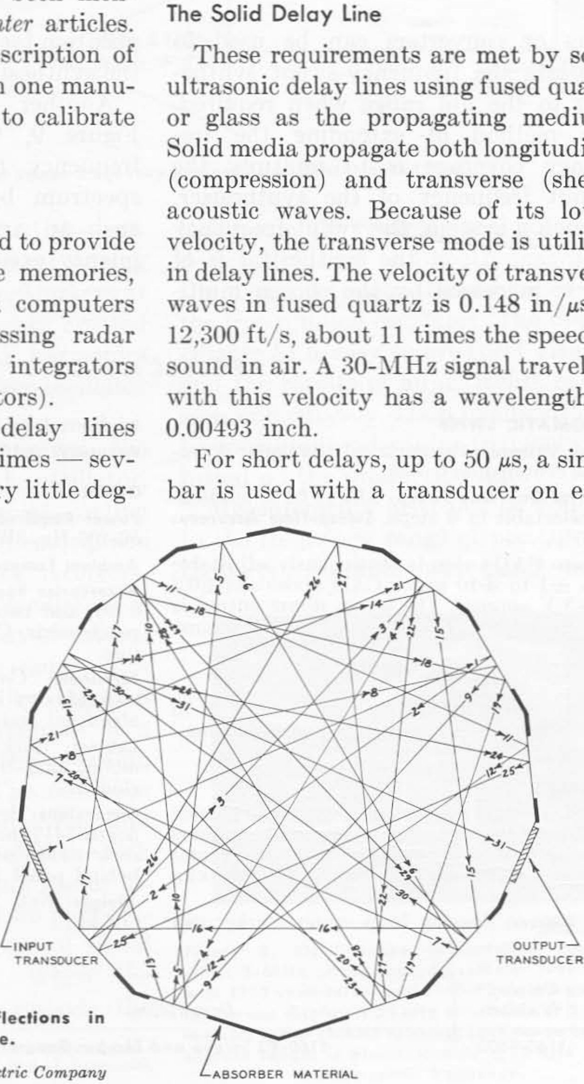

Figure 1. Multiple reflections in 31-pass quartz delay line.

Courtesy Bliley Electric Company

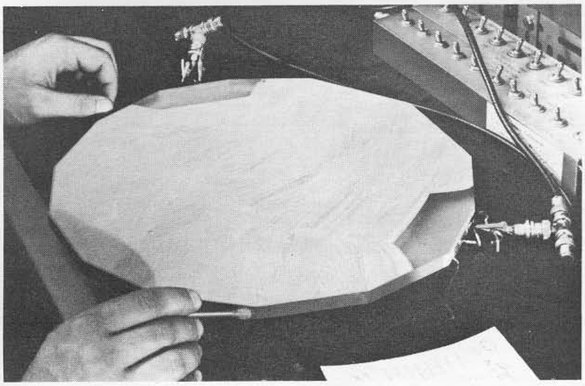

Figure 2. Fused-quartz line during manufacture.

Courtesy Microsonics, Inc.

end, but for longer delays, to save weight and space, the propagation path is folded inside a flat, polygonal slab by means of internal reflections at the sides. Figure 1 shows a complex design in which the wave is made to traverse the slab 31 times in order to achieve a long delay; Figure 2 shows the line itself, unhoused, during production. The input and output transducers are piezoelectric quartz crystal or ceramic wafers cemented to the slab. In order to suppress spurious responses, surfaces not in the path of the main beam are doped with absorbing material.

Delays up to 200 μs can be obtained with a special glass having a zero temperature coefficient of delay. For longer delays, however, the high attenuation of the glass prohibits its use and the line must be made of fused quartz, whose temperature coefficient of delay is −75 parts per million per degree Celsius. In many applications, quartz lines must be very precisely temperature-controlled to provide the necessary stability of delay.

Delay-Line Measurements

Extremely tight tolerances in delay-line specifications require precise measurements of delay time and of other characteristics during manufacture. For example, a line used for digit storage in a computer's memory may have a delay of several milliseconds that must be held to a tolerance of a nanosecond, or to a few parts in 10^7. Microsonics, Inc., a manufacturer of delay lines, has developed a fast, direct-reading method for making precise delay-time measurements with a frequency synthesizer.

Figure 3 shows how it is done. The oscillator frequency is the center fre-

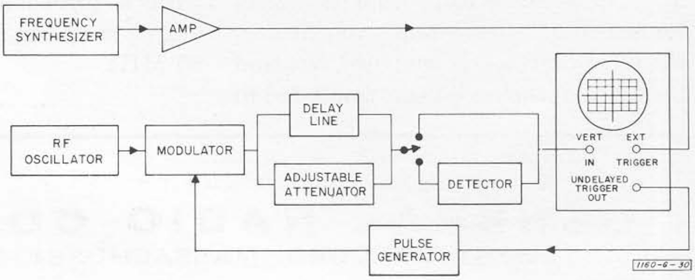

Figure 3. Equipment setup for measurement of delay line.

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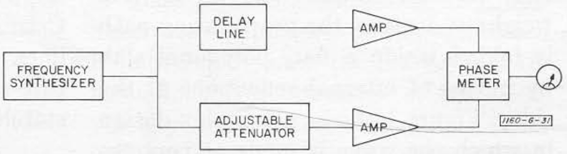


Figure 4. Equipment setup for measurement of stability of delay line.

quency of the line's passband. The oscillator output is pulse-modulated at a repetition rate derived from the synthesizer, which also synchronizes the oscilloscope. The delay line is paralleled by an attenuator, and both delayed and undelayed pulses are displayed on the oscilloscope screen. The two pulses coincide if the pulse rate, $f_{synthesizer}$, is exactly a multiple n of 1/delay time, which we will refer to as the repetition frequency of the line, f_{line}. That is, at coincidence, $f_{synthesizer} = nf_{line}$. If n is made some power of 10 the synthesizer dials read directly the repetition frequency of the delay line. The pulse envelope from the detector is used to achieve rough coincidence; then the detector is switched out and the rf pulses themselves are compared to provide a vernier reading. At a typical frequency of 30 MHz the measurement is accurate to about ± 1 nanosecond. Additional measurements of delay at the lower and upper ends of the passband determine the line's dispersion.

Another measurement requiring the synthesizer's precision is that of the temperature-control oven's ability to maintain stability of the delay time. In this case the synthesizer provides a cw signal at the line's center frequency. As shown in Figure 4, the phases of the delayed and undelayed rf carrier are compared with a phase meter. The method measures drift in the delay time with a precision corresponding to 1 degree of phase, or 1/10 nanosecond at 30 MHz.

— D. A. Gray

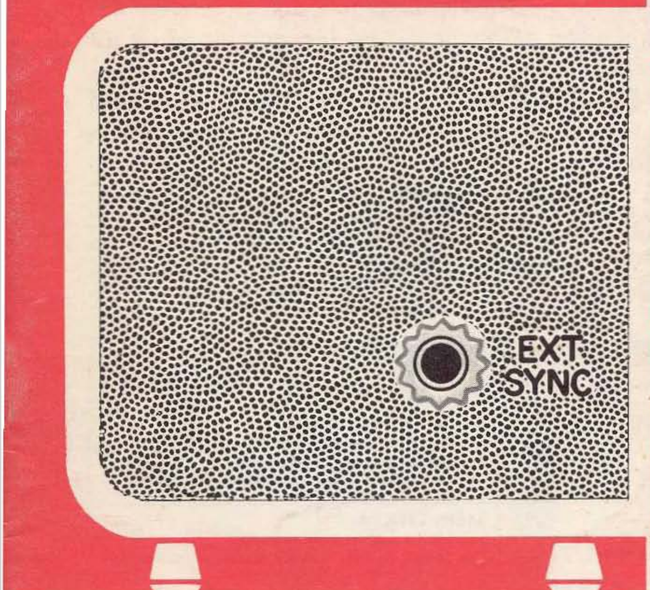
THE GENERAL RADIO

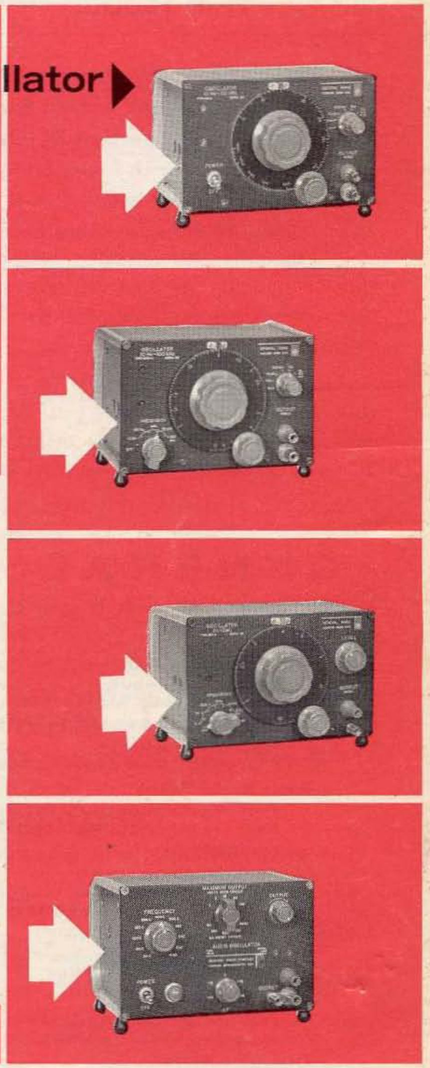
Experimenter

Single-Range RC Audio Oscillator

RC-OSCILLATOR

Synchronization





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RC-OSCILLATOR Synchronization

Four General Radio oscillators — the Types 1309,[1] 1310,[2] 1311,[3] and 1313 (see page 12) — are equipped with an auxiliary connector that serves several very useful purposes. Used as an input connector, it allows the oscillator frequency to be locked to an external signal while simultaneously filtering that signal. Used as an output connector, it provides a constant-amplitude synchronizing signal for an oscilloscope or counter. Inasmuch as this "synchronizing jack" is a fairly recent development in oscillator design (originated by GR in 1962), a summary of synchronization characteristics and of typical applications may help users of these instruments get even more out of them.

[1] R. E. Owen, "All-Solid-State, Low-Distortion Oscillator," *General Radio Experimenter*, March 1966.
[2] R. E. Owen, "A Modern, Wide-Range RC Oscillator," *General Radio Experimenter*, August 1965.
[3] R. G. Fulks, "High-Performance, Low-Cost Audio Oscillator with Solid-State Circuitry," *General Radio Experimenter*, August-September 1962.

CHARACTERISTICS

Frequency-Synchronization Characteristics

When a signal is injected through the auxiliary connector into the active RC-oscillator circuit and the oscillator is tuned within a certain range of this signal, normal oscillations cease, and the oscillator appears to oscillate stably at the injected-signal frequency. The range of frequencies over which this locking takes place is a linear function of the amplitude of the component of the input signal to which the oscillator is locked.

General Radio RC oscillators are designed so that each has a frequency lock range of ±3% for each volt input (see Figure 1). Inputs of up to 10 volts can be used without altering the operation. As Table 1 shows, the 1313-A oscillator is an exception. It is not of

TABLE 1. SYNCHRONIZATION CHARACTERISTICS

	INPUT CHARACTERISTICS			*OUTPUT CHARACTERISTICS*		
Oscillator Type	*Lock Range %/volt*	*Phase between input and output*	*Gain Factor*	*Open-Circuit Output — volts*	*Output Impedance — kΩ*	*Phase with respect to main output*
1309	±3	180 ±90°	0.47 at 5-V output	1.4	12	0°
1310	±3	0 ±90°	0.28 at 20-V output	0.8	27	180°
1311	±3	180 ±90°	0.94 at 100-V output	1.0	4.7	0°
1313	±1 to ±40	180 ±90°	–	0.7	330	0°

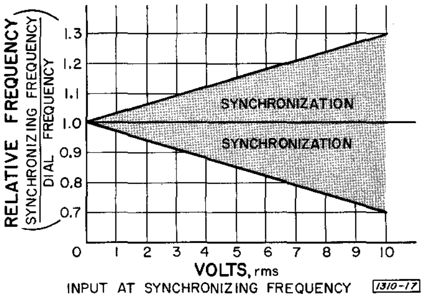

Figure 1. Locking range vs input voltage for Types 1309, 1310, and 1311 Oscillators.

the conventional Wien-bridge type, and its locking-range sensitivity varies appreciably over its frequency range.

The oscillator maintains synchronization if either the oscillator dial frequency or the synchronizing frequency is changed, within the lock range. However, a time constant of about one second is associated with the synchronization mechanism. Thus, if the amplitude or frequency of the synchronization signal or the dial setting of the oscillator is quickly changed, transient changes in amplitude and phase will occur for a few seconds before the oscillator returns to steady-state synchronization.

This time constant is caused by the thermistor amplitude regulator readjusting to the different operating conditions. The thermistor is sensitive to changes in average values of frequency or amplitude only when the averaging time is in the order of seconds. Hence, frequency-modulated and amplitude-modulated synchronizing signals, whose average values of frequency and amplitude are constant over a period of a second or less, are not affected by this time constant. They *are* affected by the equivalent time constant of the filter characteristic discussed in the next section.

For slow changes in frequency or amplitude, the lock range and the capture range are the same; i.e., the frequency or amplitude at which the oscillator goes from the synchronized state to the unsynchronized state is the same as that at which it goes from the unsynchronized state to the synchronized state.

There is a phase difference between the input synchronizing signal and the oscillator output, which depends upon the frequency's relation to the oscillator dial frequency, as Figure 2 shows. Note that the phase shift is a function of amplitude, since the lock range is a function of amplitude. Hence, the constancy of the phase shift at other than 0° depends on the amplitude stability of the input signal as well as on the frequency stability of the oscillator. As a practical matter, the useful range of phase shifts is limited to somewhat less than ±90° because of the steepness of the curve near the limits of the lock range. The data in Figure 2 are displaced by 180° for the 1309, 1311, and 1313 Oscillators because they do not have a phase-inverting output stage.

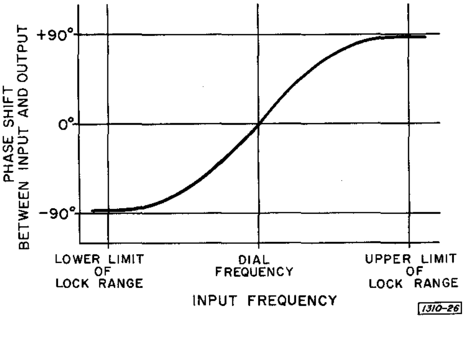

Figure 2. Phase shift relative to input frequency (and amplitude).

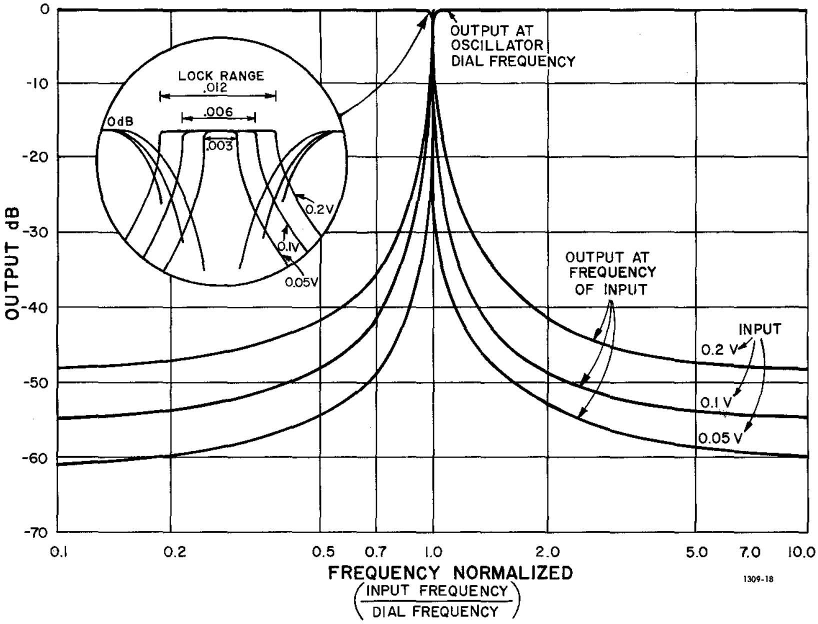

Figure 3. Response of a 1309 Oscillator for three different input-voltage levels.

Frequency-Selective Amplification Characteristics

When the output of the oscillator is locked to the input synchronizing signal, the oscillator is not oscillating in the conventional sense but is, in fact, producing an amplitude-stabilized, frequency-selective regeneration of the input signal. The result is that all the frequency spectrum of the synchronizing signal appears in the output, although most of it is greatly attenuated. Figure 3 shows the response of a 1309-A for three different input-voltage levels and for frequencies up to ten times and down to one tenth of the oscillator dial frequency. The oscillator output at both the input frequency and dial frequency is given, except within the lock range where the dial frequency oscillations stop, as seen in the magnified portion of Figure 3. The apparent increase in the Q of the response as the input level decreases is due to the fact that the output is constant within the lock range (the normal output level of the oscillator) regardless of input, while at all other frequencies it is a direct function of the input (doubling the input voltage increases the output by 6 dB).

Figure 3 is a family of curves for different input voltages, with the output plotted in dB relative to the normal oscillator output, for one particular oscillator. The single curve of Figure 4, together with Table 1, can be used to calculate the response for any input level with any GR oscillator. Figure 4 is a plot of the voltage gain versus frequency for an equal-element Wien-bridge oscillator between its synchronization input and the output. Note that for frequencies distant from the dial frequency the gain asymptotically approaches 2.0. In each oscillator this gain is modified by the resistive input divider and output amplifier. Table 1 gives the appropriate multiplying gain factor for the four GR oscillators set for maximum output voltage.

For example, the voltage amplification between the input synchronization jack on the 1309 and the full output, at twice the dial frequency, is (0.47) (4.5) = 2.1. Thus, if there were a 0.1-volt input at twice the dial frequency, there would be (0.1) (2.1) = 0.21 volt in the unattenuated output, or $\frac{(0.21)\text{ volt }(100)}{5.0\text{ volts}} = 4.2\%$ of the output at the dial frequency, regardless of the amount of output attenuation.

The input impedance of the synchronization connection is the same as the output impedance listed in Table 1, for frequencies outside the lock range. At the synchronizing frequency the input impedance, in general, is complex and can vary over a wide range, including negative values because the connection is also a source at the synchronizing frequency.

Output Characteristics

Since the injection-synchronization input connects to a resistive divider across the output of the oscillator, it is also an output. This output can be valuable because it is of constant amplitude regardless of the main-output amplitude, which may be reduced by the attenuator. The open-circuit output voltage and output impedance are given for each of the oscillators in Table 1. In each case, the amplitude is sufficient to trigger an oscilloscope or a counter. However, note that the

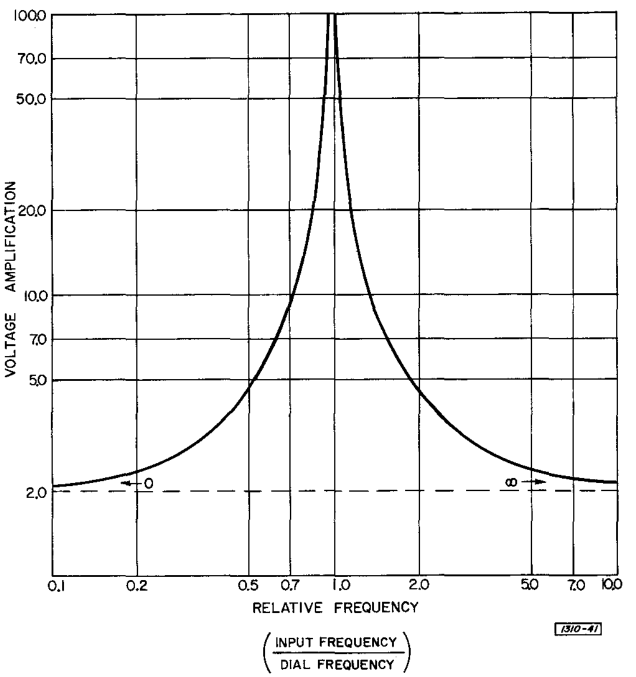

Figure 4. Voltage gain between unattenuated sync input and unattenuated output for a Wien-bridge oscillator.

output impedances are higher than are usually expected from a source. At high frequencies the output may be reduced by the capacitive loading of connecting cables.

Table 1 also gives the phase relation between the synchronization jack and the main output. Because the 1310 Oscillator output is 180° out of phase, an output balanced with respect to ground can be obtained.

This output is always a sine wave, so that on the 1309 and 1313 Oscillators simultaneous sine- and square-wave outputs are available.

APPLICATIONS

The various functions of the synchronizing jack are distinct but do exist simultaneously and can be used in complementary ways. The following applications are among the more obvious and show, with circuits and sample calculations, how the above data can be used.

Locking to a Stable Source

An oscillator with injection-synchronization capability can obviously be locked to a more accurate frequency reference to increase its long-term frequency stability. The advantages of this are many. The frequency selectivity of the oscillator can appreciably reduce the hum, noise, and distortion in the source. It will provide amplification, since less than one volt into a high impedance is necessary for locking, and yet up to 100 volts at low impedances is available in the output. The long-term amplitude stability will be the same as that of the normal oscillator, regardless of the long-term fluctuations in the input. Input-amplitude changes of 20 dB are easily suppressed in the output.

The oscillator isolates the reference source from changes in load and from the addition of spurious signals. Also, with the 1310 and 1311 Oscillators, it is possible to short-circuit the output without increasing distortion.

If the oscillator is locked to one of the harmonics of the source, it functions as a precision frequency multiplier. The accuracy and the long-term stability of the submultiple source are maintained, and the output is sinusoidal.

As an example, Figure 5(a) is the frequency spectrum of the output of a sinusoidal 1-kHz standard frequency derived by division from a crystal frequency standard. Note the 120-Hz hum, the noise close to the fundamental, and the large amount of harmonic distortion. Figure 5(b) is the output of a 1310 Oscillator locked to the same source. The distortion is reduced to almost the normal level of the oscillator, the hum is more than 80 dB below the signal, the noise is noticeably reduced, and yet the long-term frequency stability is the same as that of the reference source. The short-term stability, like the distortion, cannot be made better than that normally existing in the oscillator.

Whenever the synchronized oscillator is used for filtering, as above, the input voltage can be adjusted to an optimum level. The voltage should be high to provide a locked frequency range wide enough so that the oscillator will not drift out of lock, and yet low enough to reject the unwanted signals. Suppose that in the example it is desired to minimize the second harmonic in the oscillator output. The typical long-term stability of the 1310 at 1 kHz is 0.03% after warm-up;

therefore, a lock range of ±0.12% should provide a sufficient margin to ensure that the oscillator will always remain locked. This would require an input at 1 kHz of $\frac{0.12\%}{3.0\%/\text{volt}} = 0.04$ volt. The second harmonic in this signal is 26 dB below the desired 1-kHz fundamental (5.0%). From Table 1 and Figure 4, it is found that the 1310 has a voltage gain at the second harmonic of (4.5) (0.28) = 1.25. Therefore, with a 0.04-volt input there would be (1.25) (0.04) (0.05) = 0.0025 volt of the second harmonic in the 20-volt oscillator output, or $\frac{0.0025 \text{ volt}}{20 \text{ volts}} = 0.0125\%$, regardless of the output attenuator setting. This is below the amount of second-harmonic distortion normally present in the oscillator, as Figure 5(b) shows, so it is certain that the largest possible reduction of the second harmonic has been made.

Frequency-Jitter Reduction

Although the short-term frequency stability, or jitter, of the synchronized oscillator cannot be better than when it is unsynchronized, it can be better than the source to which it is locked. This is, again, because it behaves as a tracking narrow-band filter.

In Figure 6, the output frequency of a drifting, jittery 10-Hz source is

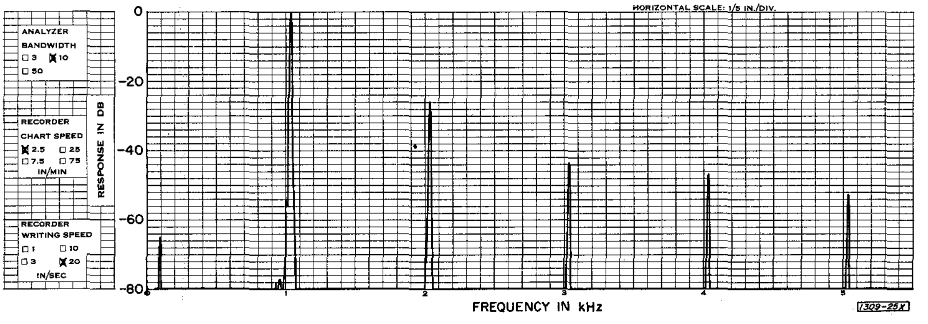

Figure 5(a). Spectrum of a typical sinusoidal 1-kHz standard frequency, derived by division from a crystal frequency standard.

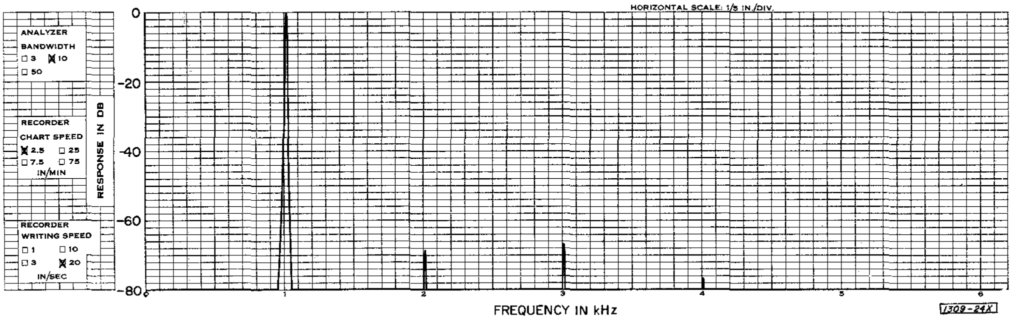

Figure 5(b). Spectrum of the output of a 1310 Oscillator synchronized with the 1-kHz standard of Figure 5(a). Note the reductions in hum, noise, and distortion.

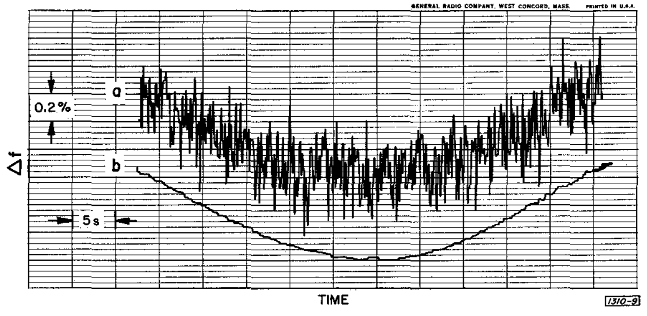

Figure 6. (a) Output frequency of a drifting, jittery 10-Hz source. (b) Output frequency of oscillator synchronized with source (a). Note that jitter is reduced while drift is tracked.

recorded along with the frequency of the output of an oscillator synchronized to that source. The filter selectivity has considerably reduced the short-term jitter, while the oscillator has remained locked onto the long-term drifting average. The low frequency of this example was used for convenience in making the graphic recordings. A reduction in jitter can be made at any frequency where the filter characteristic is sufficiently selective. The ability to track longer-term drift, however, is always limited by the approximately one-second time constant of the locking mechanism.

Harmonic Waveform Synthesis

One of the most popular uses of the synchronized oscillator is as a sinusoidal frequency multiplier for Fourier synthesis of various waveforms. The oscillators are simply locked onto a harmonically rich waveform with the input level adjusted for sufficient suppression of the other harmonics.

Tone bursts can supply a harmonically rich signal for synchronizing. For example, if it is desired to synthesize

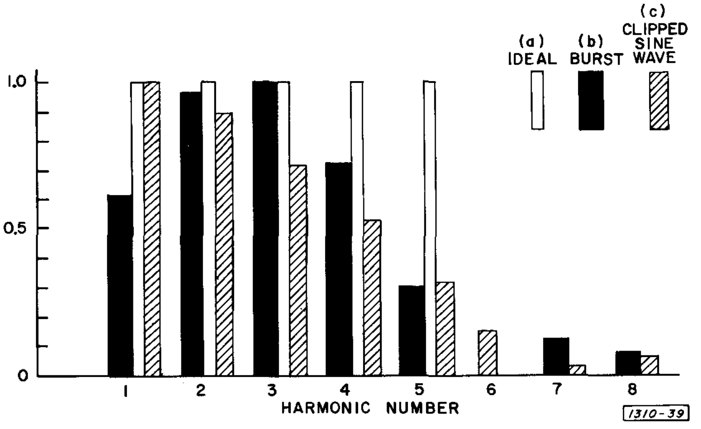

Figure 7. Spectrum of locking signal for generating waveforms with five harmonics. (a) ideal, (b) tone-burst approximation, (c) clipped sine-wave approximation.

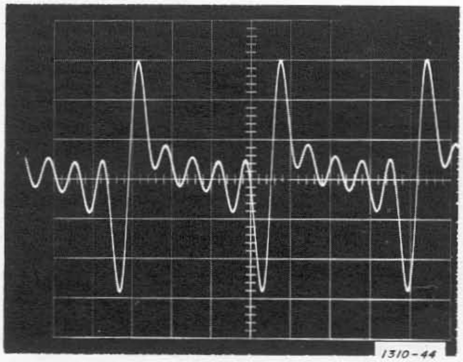

Figure 8. Ideal signal for generating waveforms with five harmonics. It is composed of equal amplitudes of the five in-phase harmonics.

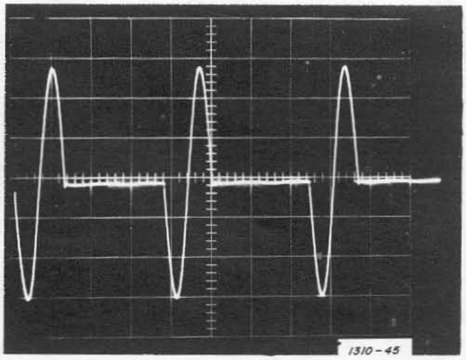

Figure 9. Tone-burst approximation to Figure 8.

pure waveforms from the first five harmonics, a waveform with the spectrum of (a) in Figure 7 would be best. This has the waveshape of Figure 8, which can be approximated by a single cycle of the third harmonic with a repetition rate of the fundamental frequency, as in Figure 9. Its spectrum, (b) in Figure 7, is quite close to ideal. This waveform is easily generated with the GR 1396 Tone-Burst Generator (see Figure 10). The spectrum of a tone burst is very good for synchronizing because it can produce a relatively flat spectrum for large harmonic numbers and because it is not frequency-sensitive.

Conventional nonlinear waveshaping methods can be used to generate a signal with a desired harmonic spectrum. If shaping techniques are used, it is helpful to recall that, for repetitive

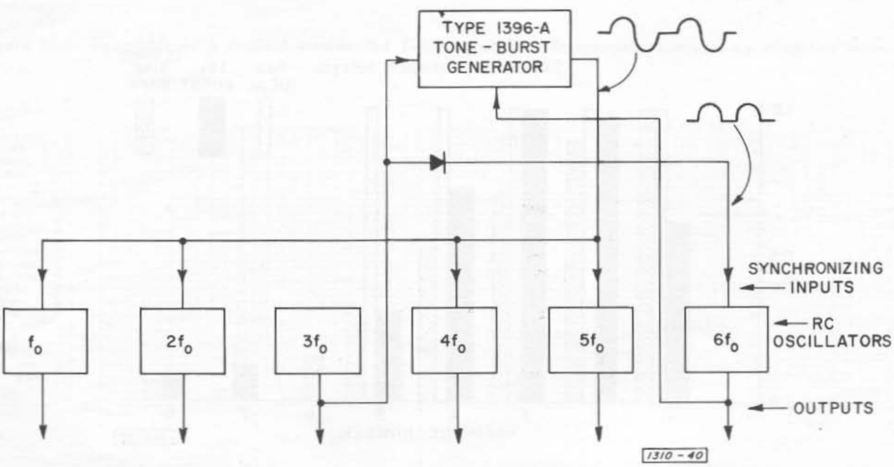

Figure 10. One method of obtaining synchronized oscillators for first six harmonics. Waveform of Figure 9 is used to lock oscillators.

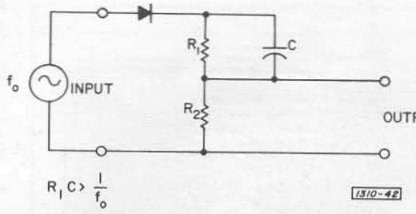

Figure 11. Sine-wave clipping circuit for approximating equal-amplitude harmonics pulses.

equal-amplitude harmonic signals, the waveform goes progressively from a smooth sine wave with one harmonic to an impulse with all harmonics. Hence, the ideal synchronizing signal is an appropriately bandwidth-limited impulse. For low harmonics this can be approximated with a clipped sine wave by means of a circuit such as that of Figure 11. For this five-harmonic example, values for R_1 and R_2 of 10 kΩ and 200 Ω, respectively, produced the waveform of Figure 12 and the spectrum of (c) in Figure 7.

As in the first application, there is an optimum input-synchronizing voltage, which provides the best combination of purity of output and lock range. In this case it usually is desirable to make the lock range large so that the phase of each harmonic can be adjusted. The phase-coherent signal of Figure 8 was generated with the equipment shown in Figure 10, and all undesired higher harmonics were more than 60 dB below the five equal-amplitude ones.

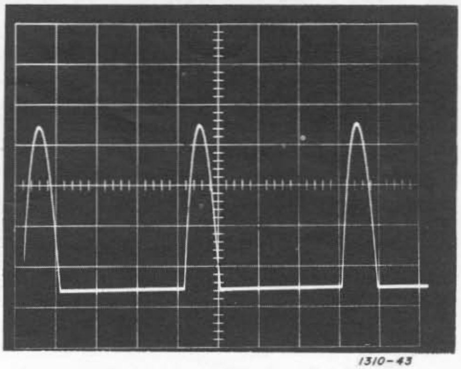

Figure 12. Clipped sine wave with the spectrum of Figure 7(c).

Phase Shifting

The synchronized oscillator can be used as a convenient, single-frequency phase shifter or time delay. Table 1, in conjunction with Figure 2, shows the range of phase shift available for each oscillator. This is particularly useful with the 1309, where the Schmitt trigger in the square-wave circuit permits generation of variable-delay pulses.

— R. E. Owen

GENERAL RADIO RC OSCILLATORS WITH SYNCHRONIZATION

Condensed Specifications

Oscillator Type	*Frequency Range*	*Output Waveform*	*Output Voltage*	*Output Power*	*Distortion*
1309	10 Hz - 100 kHz	∿ ⊓	500 μV - 5 V	10 mW	0.05%
1310	2 Hz - 2 MHz	∿	0.1 - 20 V	160 mW	0.25%
1311	50 Hz - 10 kHz in 11 steps	∿	0-1, 3, 10, 30, 100 V Transformer output	1 W	0.5%
1313	10 Hz - 50 kHz	∿ ⊓	500 μV - 5 V	10 mW	0.5%

Figure 1. Type 1313-A Oscillator.

10 Hz TO 50 kHz WITHOUT RANGE CHANGING

For general laboratory use, the conventional decade frequency range on an RC oscillator is a good compromise between accuracy, resolution, and ease-of-setting. On the production line, however, where measurements are made in rapid succession over a wide frequency range, the necessary range switching and large return sweeps of the dial become an important disadvantage. To eliminate this problem, General Radio has developed a low-cost RC oscillator with the entire audio-frequency range covered in a single range.

The TYPE 1313-A Oscillator (Figure 1) provides sine and square waves from 10 Hz to 50 kHz. The frequency is quickly and easily set and unambiguously indicated on a single-turn dial. There are no multipliers to use or decimal points to slip; the dial is marked the way you would say the frequency: ten kilohertz is 10 kHz, not 10,000 Hz, for example. Also, since there is no range switch, there are no range-changing transients, no fast, high-amplitude pops to rupture a voice coil or mechanical transducer. And there is no necessity for routine replacement of the range switch, as there often is with other oscillators used on production lines.

The TYPE 1313-A is in many respects similar to the popular TYPE 1309-A 10

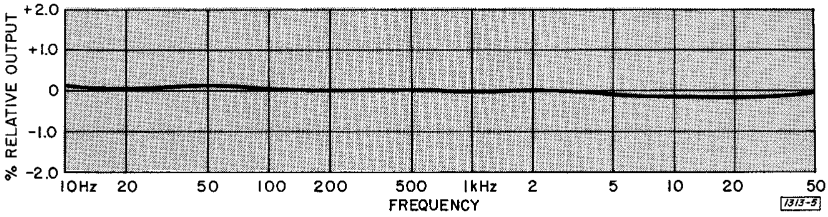

Figure 2. Typical output-vs-frequency characteristic.

Hz-100 kHz Oscillator[1]. It uses the same all-silicon, all-solid-state design, except that a modified Wien Bridge expands the frequency range. The technique used is similar to that employed by Anderson in a seven-decade oscillator[2], but it has been refined with the aid of a digital computer.

The sine-wave output is continuously adjustable over a range of from less than 500 μV to 5.0 volts open-circuit by means of a 60-dB step attenuator and a continuous control. The steady-state output voltage is held within $\pm 2\%$ of its 1-kHz value over the whole dial span; it is typically even better than this, as Figure 2 indicates. Thus frequency-response measurements are not interrupted by periodic readjustments of the output level. Distortion (see Figure 3) is held below 0.5% from 100 Hz to 10 kHz.

The square-wave output has a very fast transition time, typically 40 nanoseconds into 50 ohms. This corresponds to the rise time of a device with a bandwidth of greater than 10 MHz; hence it is adequate for most transient-response testing. The maximum output is greater than + 5 volts peak-to-peak

[1] R. E. Owen, "All-Solid-State Low Distortion Oscillator," *General Radio Experimenter*, March 1966.
[2] F. B. Anderson, "Seven-League Oscillator," *Proceedings of the IRE*, 39, August 1951, pp 881–890.

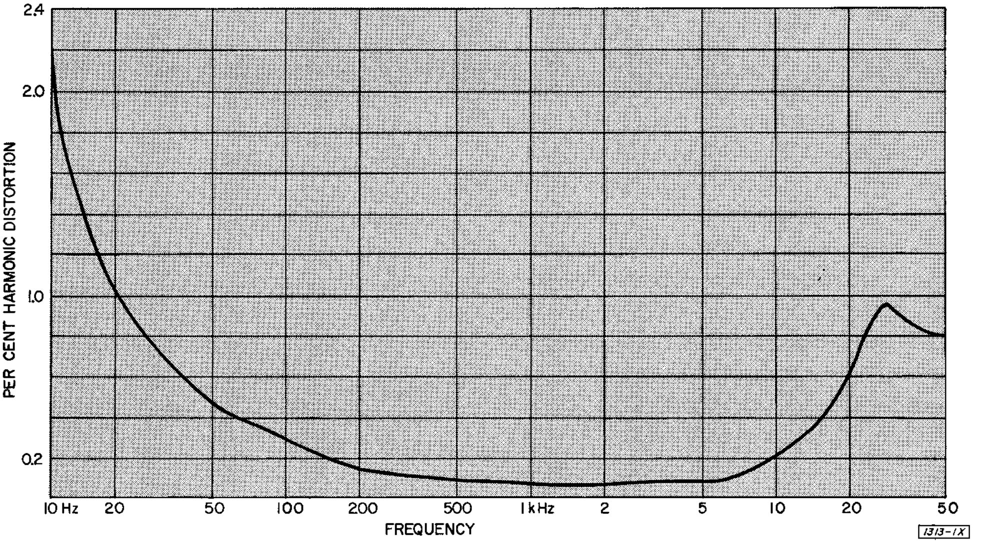

Figure 3. Typical distortion-vs-frequency characteristic.

Figure 4. In a production test, the 1313 drives a loudspeaker over the entire audio range in one turn of the dial.

(Courtesy of KLH Research and Development Co., Cambridge, Massachusetts)

(open-circuit), and it is direct-coupled, so there is no low-frequency tilt, even at 10 Hz. Symmetry is specified at ±2% (48%–52% duty ratio) over the whole frequency range, but typically there is no asymmetry discernible on an oscilloscope. The square-wave output can be continuously adjusted by the 20-dB attenuator.

There are no provisions for mechanically sweeping this oscillator. Sweeping an inexpensive RC oscillator of this type is not recommended because of transient amplitude changes as the frequency varies and because the dial calibration is nonlogarithmic. For sweep applications in this frequency range, we recommend the TYPE 1304-B Beat-Frequency Audio Generator, which provides logarithmic calibration and which maintains constant amplitude when mechanically swept.

The TYPE 1313-A, in common with other General Radio RC oscillators, has a frequency-synchronization capability (see page 3). As a result of the single-range circuit, the frequency locking range varies from less than 1% per volt input at 10 Hz to greater than 40% per volt at 50 kHz. This synchronizing feature permits each station on a production line to have, in essence, a tuned isolation amplifier with independent amplitude and waveform control, operating from one standard-frequency source.

— R. E. OWEN

A biographical sketch of Mr. Owen appeared in the March 1966 *Experimenter*.

SPECIFICATIONS

FREQUENCY

Range: 10 Hz to 50 kHz in one range.
Accuracy: ±4% or ±1Hz, whichever is greater.
Synchronization: An external reference signal can be introduced through phone jack to phase-lock oscillator. 1-V input provides locking range of ±1% to ±40%, depending on frequency.

OUTPUT

Sine Wave

Power: 10 mW into 600-Ω load.
Voltage: 5.0 V ± 5% open-circuit.
Impedance: 600 Ω. One terminal grounded.
Control: Minimum of 20 dB continuously adjustable and 60 dB step attenuator (20 ± 0.2 dB per step). Also, a 0-V output position with 600-Ω output impedance maintained.
Distortion: Less than 0.5% from 100 Hz to 10 kHz.
60-Hz Hum: Less than 0.05% at 1 kHz.
Frequency Characteristic: ±2% over whole frequency range for loads of 600 Ω or greater.

Square Wave

Voltage: Greater than +5 V p-p, open-circuit. Dc-coupled output.
Impedance: 600 Ω.
Rise Time: Less than 100 ns into 50 Ω. Typically 40 ns at full output.
Control: Minimum of 20 dB, continuously adjustable attenuator only.
Symmetry: ±2% over whole frequency range.

GENERAL

Accessories Supplied: CAP-22 Power Cord, spare fuses.
Accessories Available: 1560-P95 Adaptor Cable (phone plug to 274-MB Double Plug) for connection to synchronizing jack; relay-rack adaptor set.
Power Required: 100 to 125 V, 200 to 250 V, 50 to 400 Hz, 6 W.
Mounting: Convertible-bench cabinet.
Dimensions (width-height-depth): 8³⁄₁₆ by 5⅞ by 8⅛ in (210 by 150 by 210 mm).
Weight: Net, 7 lb (3.2 kg); shipping, 9¼ lb (4.2 kg).

Catalog Number	*Description*	*Price in USA*
1313-9701	**Type 1313-A Oscillator, 10 Hz–50 kHz**	**$325.00**
1560-9695	**Type 1560-P95 Adaptor Cable**	**3.00**
0480-9638	**Type 480-P308 Rack-Adaptor Set**	**7.00**

HARMONIC BRIDGE USES RC-OSCILLATOR SYNCHRONIZATION

Ingenious use is made of the synchronizing capability of a GR 1310 Oscillator by Dr. Homer Fay of the Speedway Laboratories of Union Carbide Corporation, Electronics Division. Writing in *The Review of Scientific Instruments*[1], Dr. Fay describes a system used to measure linear and nonlinear electric coefficients and quadratic electrooptic coefficients in high-dielectric perovskite crystals. These coefficients can be derived from a capacitance-bridge measurement if the harmonic content of the driving voltage is known, and this is where the synchronizing oscillator enters the system.

The 1310, along with several other GR oscillators (see page 3, this issue), can be phase-locked to a signal whose frequency is within a certain range of the oscillator dial setting; moreover, once lock is established, the oscillator

[1] Dr. Homer Fay, "Harmonic Bridge for Measurement of Nonlinear Electric and Electrooptic Properties of Crystals," *The Review of Scientific Instruments*, February 1967.

Figure 1.
Type 1310 Oscillator.

frequency control can be used as a phase shifter, over a range of ±75° or so.

In Dr. Fay's setup (Figure 2), the oscillator frequency dial of each of several 1310 oscillators is set in the vicinity of a harmonic of the driving signal, thereby establishing lock. The oscillators then assume control of both the phase (by means of the frequency control) and the amplitude (by means of the output level control) of each harmonic covered. The phase and amplitude of these harmonic components are adjusted to cancel the harmonics created by the nonlinearity of the crystal under test. Finally, measurements of the amount of such compensation at each harmonic are used to calculate the electric coefficients of the crystal.

Such measurements are important because many effects in crystals are directly related to electric displacement. The electrooptic effect is an example. The harmonic bridge, says Dr. Fay, "permits display of the electrically induced phase retardation intensity pattern as a function of electric displacement, from which the electrooptic coefficients may be readily obtained."

— R. E. Anderson

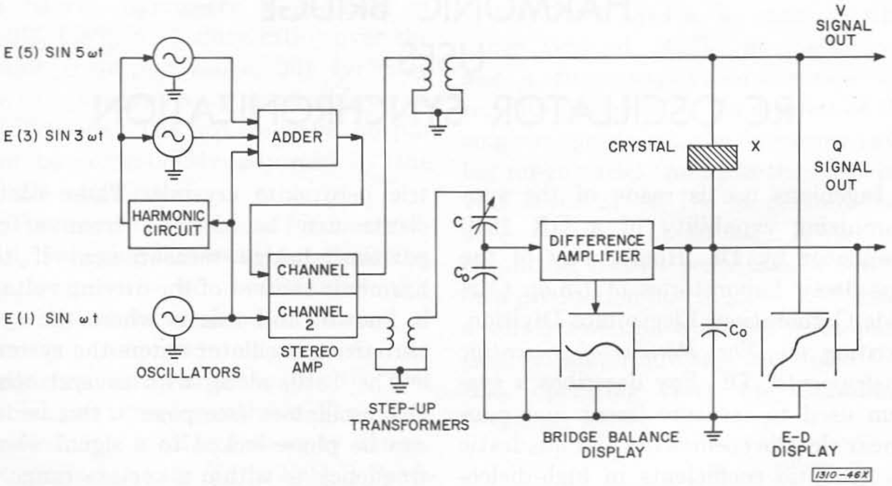

Figure 2. Block diagram of harmonic bridge, showing use of GR 1310 Oscillators to provide harmonics of driving signal.

Figure 1. Type 1232-P2 Preamplifier attached to the 1232-A Tuned Amplifier and Null Detector

A PREAMP FOR USE WITH BRIDGE DETECTORS

In the quest for low noise in an amplifier, one must accept the fact that no one amplifying device is optimum for signal sources of widely differing impedance levels. Some compromise is therefore inevitable in the design of the input stage of a sensitive null detector that is to be used in a variety of applications. The low-noise transistor used in the TYPE 1232-A Tuned Amplifier and Null Detector[1] is suitable for use with most impedance-bridge systems. However, some measurements requiring extremely high sensitivity present a very high impedance to the detector, and in such cases the detector could benefit from a preamplifier with a very high optimum-source resistance. The new 1232-P2 FET Preamplifier (Figure 1), designed to fill this need, can increase sensitivity by a factor of 10 or more in some measurements.

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

Plots of typical equivalent input noise vs resistance for the 1232-A alone and with the 1232-P2 are shown in Figure 2. The input noise can be char-

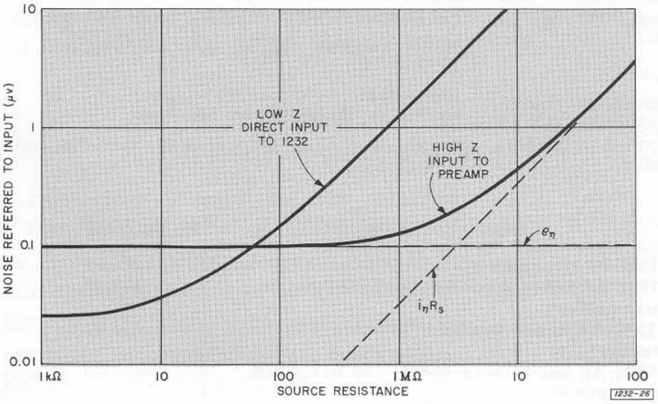

Figure 2. Equivalent input noise vs source resistance for 1232-A alone and for combination of 1232-A and 1232-P2.

acterized by an equivalent voltage noise generator, e_n, and a current noise generator, i_n. On the plot of equivalent input noise *voltage* vs resistance, e_n is a horizontal line and $i_n \times R_s$ is a diagonal line. Note that the two curves cross at 60 kilohms and that below this value the 1232-A is better without the preamplifier.

One application where the addition of the preamplifier is a distinct advantage is the measurement of low-loss dielectric samples on the GR 1615-A Capacitance Bridge. Here the unknown capacitance is usually less than 1000 pF, the lowest D range is usually used (and not the G ranges), and the frequency is usually under 500 Hz. On the lowest D range, the output capacitance of the 1615-A is approximately 1600 pF + C_x + cable capacitance. If C_x and the cable capacitance are small, the output impedance will be about 1 MΩ at 100 Hz, and the preamplifier will improve sensitivity by a factor of about 10, as shown by Figure 2. As the frequency increases, the impedance decreases, and eventually the 70-pF capacitance of the preamplifier's input cable negates the use of the preamplifier, even with a source of infinite impedance.

The preamplifier is of no advantage on the higher D range, where bridge capacitance is 10 times as great as on the lowest D range, or on the G ranges, where the output impedance is shunted by 100 kilohms.

The circuit of the preamplifier consists of a single source-follower stage, using a field-effect transistor. A switch allows the user to bypass the preamplifier in applications where the 1232-A is better off alone. The preamplifier is housed in a thin "pancake" box that is easily added to the side of the 1232-A or between the 1311-A Oscillator and the 1232-A in assemblies. The resulting combinations are available as the TYPES 1232-AP (1232-A plus preamplifier) and 1240-AP (1232-A plus preamplifier plus 1311-A). The entire TYPE 1620 Capacitance Measuring Assembly, when supplied with the preamplifier, is designated TYPE 1620-AP.

— H. P. HALL

SPECIFICATIONS

Input Impedance: Greater than 100 MΩ in parallel with 70 pF.
Output Impedance: 10 kΩ.
Voltage Gain: Approx 0.7.
Noise (referred to input): Open-circuit equivalent, 0.1 pA; short-circuit equivalent, 0.3 μV (when used with Type 1232-A tuned to 100 Hz).
Optimum Source Impedance: 3 MΩ.

Connectors: GR874 on cables, input and output.
Power Required: 12 V, 200 μA, supplied by 1232-A.
Dimensions (width-height-depth): ¾ by 6 by 7½ in (20 by 150 by 190 mm).
Weight: Net, 15 oz (425 grams); **shipping,** (est) 3 lb (1.4 kg).

Catalog Number	*Description*	*Price in USA*
1232-9602	**Type 1232-P2 Preamplifier**	**$ 95.00**
1232-9829	**Type 1232-AP Tuned Amplifier and Null Detector, with preamplifier**	**485.00**
1240-9829	**Type 1240-AP Bridge Oscillator-Detector, with preamplifier**	**725.00**
1620-9829	**Type 1620-AP Capacitance-Measuring Assembly, with preamplifier**	**2325.00**

Type 1123 Digital Syncronometer.

SYNCRONOMETER WITH 1-2-4-8 CODE

A new version of the SYNCRONOMETER® digital time comparator[1] is now available. The output impedance of the new model has been lowered by a factor of 10, and the output coding has been changed to 1-2-4-8 BCD. All other features of the instrument — standby battery power, synchronization capability, fail-safe operation, etc — remain unchanged.

In the new instrument, a buffer transistor is added to each of the 44 data-output lines, reducing the output impedance and permitting the change in coding. Rise times of the data output are thus reduced, simplifying the transfer of precise time data to a parallel-storage unit, printer, or computer.

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

SPECIFICATIONS

Same as 1123-A[1], except as follows:
Time-of-day Data Output:
From all decades, parallel 1-2-4-8 BCD
Logic 0: approx 0.8 V, impedance 1 kΩ.
Logic 1: approx 15 V, impedance 11 kΩ.

Catalog Number	*Description*	*Price in USA*
1123-9760	**Type 1123 Digital Syncronometer (1-2-4-8 BCD Code), 115 V, Bench Model**	**$3450.00**
1123-9763	**Type 1123 Digital Syncronometer (1-2-4-8 BCD Code), 115 V, Rack Model**	**3450.00**
1123-9762	**Type 1123 Digital Syncronometer (1-2-4-8 BCD Code), 230 V, Bench Model**	**3450.00**
1123-9765	**Type 1123 Digital Syncronometer (1-2-4-8 BCD Code), 230 V, Rack Model**	**3450.00**

QUANTITY PRICES FOR ENLARGED SMITH CHARTS

Since some users of the new enlarged Smith Charts ($22\frac{1}{2}'' \times 35''$) announced in September want to order more than one pad at a time, and we are happy to handle the larger orders, the following quantity price schedule has been established.

Number of Pads	*Price per Pad of 75 Sheets*
1	$6.00
2–3	5.75
4–9	5.50
10–19	5.00
20 and up	4.75

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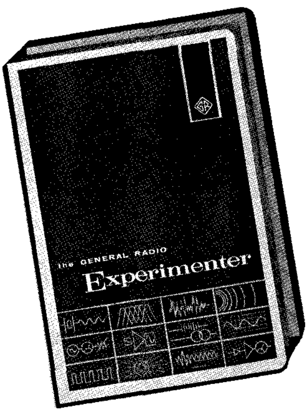

Experimenter INDEX

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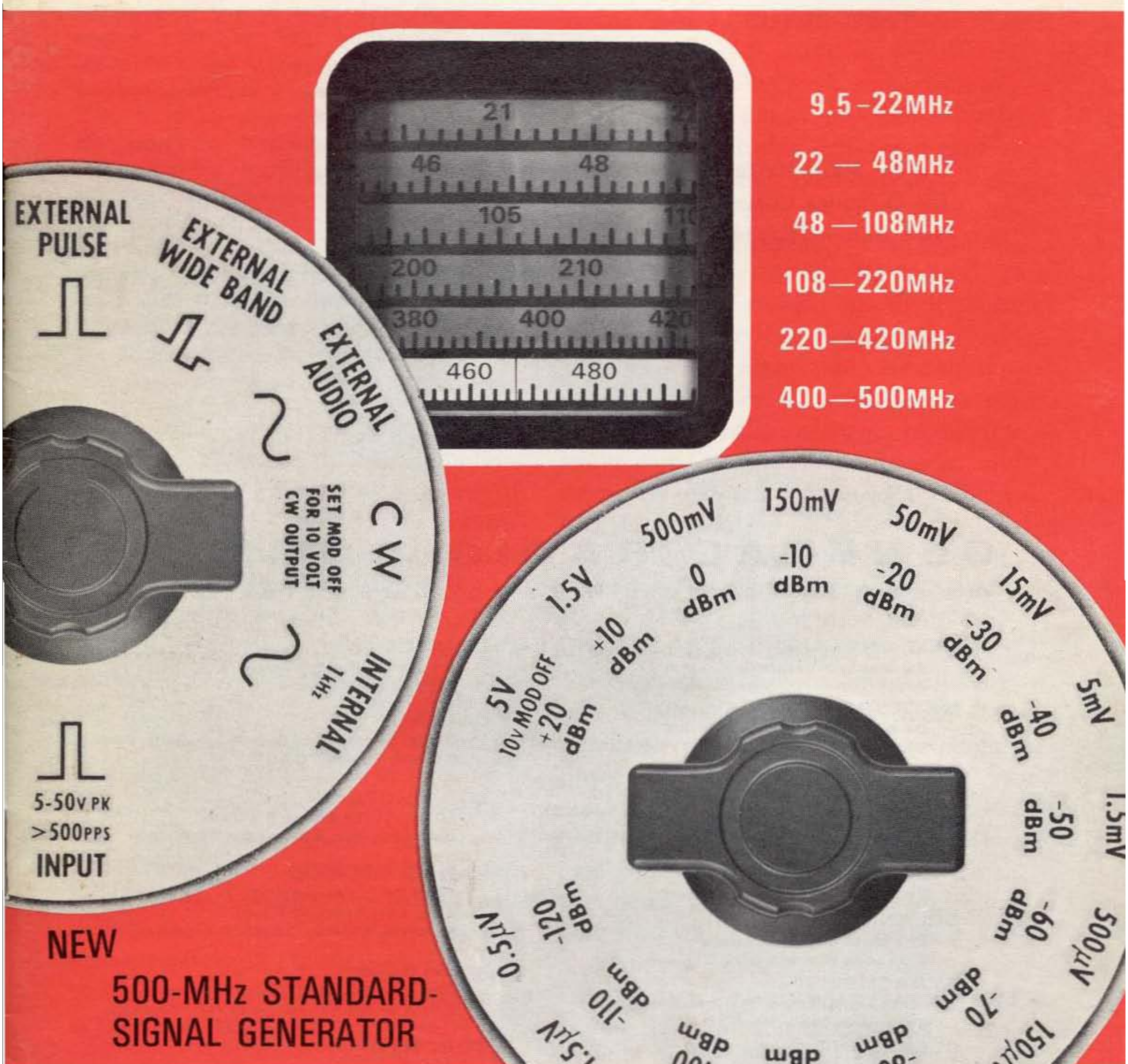
Experimenter



NEW 500-MHz STANDARD-SIGNAL GENERATOR

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A NEW 500-MHz STANDARD-SIGNAL GENERATOR

A great many interacting problems surround the development of a standard-signal generator, and the final instrument usually combines a few successes with more than a few compromises. If the signal generator is to have an output-level range of 160 dB, accurate frequency control without trimmers, and leveling in all modulation modes, the challenges are compounded. Yet these were among the design objectives of GR's latest signal-generator development. A major engineering effort turned the problems into successes and produced the high-performance standard-signal generator introduced in this month's feature.

The basic tool for the alignment and testing of receivers, filters, amplifiers, and attenuators is the a-m standard-signal generator. The term "standard-signal" means that all primary characteristics are calibrated, and it implies signal quality beyond that expected of the ordinary signal source. Thus, it is generally accepted that an a-m standard-signal generator shall have calibrated output frequency, calibrated output level adjustable over a wide range, and calibrated depth of modulation. As technology has evolved, users have demanded increasingly wide ranges of these characteristics, along with major reductions of incidental and spurious effects; at the same time, speed and convenience of use have become prime concerns.

A NEW STANDARD OF PERFORMANCE

General Radio's new Type 1026 Standard-Signal Generator meets this growing demand for higher performance and simple operation. Over a frequency range from 9.5 to 500 MHz, the 1026 delivers one-half watt of leveled cw power to a 50-ohm load, approximately 100 times the power previously available from a well shielded signal generator. Modulation characteristics are equally impressive. At all carrier frequencies, 95% modulation is available for outputs up to 5 volts behind 50 ohms (compared with up to 10 volts behind 50 ohms for cw operation). At 50% 1-kHz amplitude modulation, incidental fm is less than 1 part per million, an order of magnitude improvement over previous designs, and envelope distortion for the 1-kHz internal modulation is less than 1%. Other unique a-m features include provision for wide-band modulation up to 1.5 MHz and for pulse modulation with leveled and metered peak output.

Electrical fine frequency control permits frequency modulation and, with auxiliary synchronizing equipment, phase-locked operation.

The three front-panel elements used for this month's cover illustration symbolize the extreme ease with which the signal generator can be controlled over its wide operating ranges. The user selects frequency by setting the band switch to the proper range and tuning the frequency control. The large, back-lit drum dial provides an unambiguous readout accurate to 0.5%. There are no secondary frequency controls, no trimmers to peak. Output level is set by a precision step attenuator supplemented by a continuous carrier level control; the meter and large attenuator dial provide a highly legible readout. A flick of the wrist takes the user from 0.1 microvolt to 10 volts behind 50 ohms, without external amplifiers and their attendant tuning, shielding, and level-monitoring problems.

RELATION TO OTHER SIGNAL SOURCES

The characteristics of the standard-signal generator, as embodied in the 1026, set it distinctly apart from such relatively simple signal sources as GR's general-purpose oscillators on the one hand and from our highly sophisticated decade-frequency-synthesizer line on the other. The oscillators are ideal for antenna and bridge measurements and as local oscillators for heterodyne detectors, but they lack the output-level calibration, ultra-high shielding, isolation of frequency from load pulling effects, and calibrated modulation characteristics required for receiver tests and for precision insertion-loss measurements.

Frequency synthesizers are quite complex instruments in which the primary concern is to generate precisely known output frequencies, usually with relatively less attention devoted to achieving a wide range of accurately calibrated output levels and to modulation capabilities. Furthermore, the spectral impurities found in the output of frequency synthesizers are generally of a different character from those in the output of the free-running, continuously tunable LC oscillator used in the 1026. The unwanted outputs of the latter are almost entirely harmonics or hum sidebands close to the carrier; other discrete spurious outputs are not present, and broadband noise sidebands are of extremely low amplitude.

Thus a good conventional a-m standard-signal generator is preferable to a synthesizer for certain important classes of measurements, including receiver spurious-response tests, receiver sensitivity checks, which require accurately known low-level signal amplitudes and known modulation characteristics, and measurements of receiver selectivity, filter cutoff, and attenuator insertion loss, all of which require a wide range of accurately calibrated output levels.

APPLICATIONS

The 9.5-to-500-MHz frequency range of the 1026 includes the important vhf and uhf aircraft communications bands, which use double-sideband a-m, and most of the common high i-f bands. Obvious applications thus lie in the alignment and testing of receivers, filters, amplifiers, attenuators, and other devices used in such service.

The largest single application area for standard-signal generators is receiver testing. The 1026 is ideal for receiver testing because of its high, leveled output, single-dial tuning, and low modulation distortion.

AGC and Squelch Testing

The receiver manufacturer is usually called upon to specify the range of rf input voltage over which the AGC will maintain a relatively constant audio output level. For such tests, the 1026 offers a wide range of output levels, extremely low incidental fm, and a highly accurate output attenuator.

The wide-band- and pulse-modulation modes are also useful for precise checks of squelch and AGC recovery time.

Distortion Tests

The maximum nonlinear distortion in a receiver is generally specified at about 5 to 10%. Any envelope distortion in the signal generator will, of course, introduce error in the measurement. Negative envelope feedback and adequate buffering (which virtually eliminates incidental fm) reduce distortion in the 1026 to 1% at 50% a-m and to less than 3% at the critical 80% a-m level. Not only is the 1026 output leveled, but there is no peaking by the operator to obtain specified modulation performance.

Sensitivity Tests

In measurements of receiver sensitivity, a signal generator must produce an rf signal at a low, accurately known level, with known modulation percentage. It is important to keep these characteristics, as well as carrier frequency, constant during the test procedure. In the 1026, the output attenuator is accurate to ±0.1 dB per step, with a maximum accumulated error of ±0.5 dB. Rf leakage is negligible even when the output is in tenths of a microvolt. After warmup, the output level will remain constant to ±0.01 dB over any 15-minute period, even with ±10% line-voltage fluctuations.

Signal-to-Noise Ratio

In measurements of signal-to-noise ratio, the usual procedure is to compare the receiver audio output when a modulated carrier is applied with the output without modulation. Any spurious modulation due to hum or noise will appear as additional receiver noise. The very low residual a-m hum and noise of the 1026 (at least 70 dB below carrier) permit measurements of signal-to-noise ratios up to 60 dB with confidence.

Other Receiver Tests

Measurements of adjacent-channel rejection, image-frequency rejection, and responses due to local-oscillator harmonics can all be made with greater confidence and convenience, because of the purity and stability of the 1026 output.

Noise-limiter effectiveness is commonly measured by tests on the receiver's ability to limit the noise spikes to some specified modulation percentage, such as 80%. Thus the ability of the 1026 to provide accurate high-percentage modulation makes it particularly attractive for noise-limiter testing.

Amplifier, Attenuator, Filter Tests

In tests of amplifier gain, frequency response, and distortion, the leveled output of the 1026 eliminates repeaking of controls as frequency is changed. Also, the high output levels available from the 1026 will drive i-f amplifiers and low-impedance stages directly, without the addition of power amplifiers. Pulse response of i-f amplifiers in the 30-to-200-MHz range is of great interest because of the widespread use of such amplifiers in radar systems. The clean leveled pulse performance of the 1026 speeds up measurements on this class of equipment considerably.

In filter-response tests, the high output capability of the 1026 is especially useful when the rejection outside the pass band is 70 dB or more, since

Figure 1. The 1026 Standard-Signal Generator.

the filter output signal will still be easily detectable by an untuned high-frequency voltmeter. The high-level output, automatic leveling, and low leakage are valuable in attenuator testing. Phase-lock stabilization of carrier frequency (with auxiliary synchronizing equipment) also permits the use of highly selective detectors to improve signal-to-noise ratios in measurements of insertion loss in excess of 100 dB.

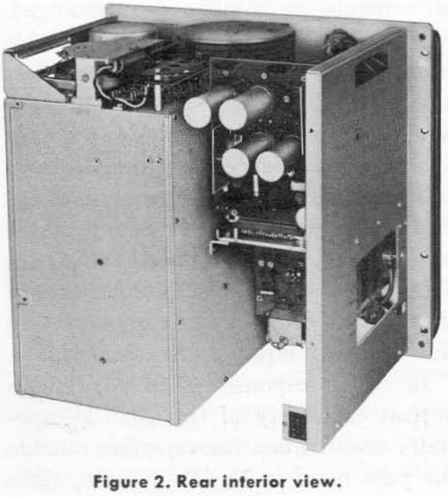


Figure 2. Rear interior view.

GENERAL DESCRIPTION

The 1026 Standard-Signal Generator (Figure 1) is basically a continuously tunable master oscillator-power amplifier chain, with appropriate power supply and modulator circuits to permit automatic level control under a wide range of modulation conditions.

The instrument is built in three main subassemblies: rf assembly, modulator assembly, and power-supply assembly (see Figure 2). Each subassembly is extensively pretested before being secured to the panel assembly, which supports all controls and which provides interunit cabling.

Vacuum tubes are used in the rf stages because of their superior performance and reproducibility in the upper part of the frequency range, but all components in the modulator and power supply are solid-state.

The 9.5-to-500-MHz frequency range is covered in six bands, five of which cover approximately an octave each. The bands were selected so that important allocations, such as the 88-to-108-MHz fm band, the 108-to-156-MHz vhf

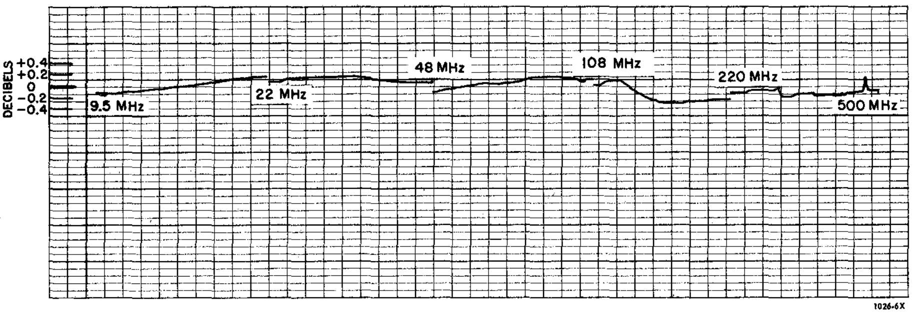


Figure 3. Output level vs carrier frequency. The small discontinuities in level (typically less than 0.2 dB) occur at frequencies where the range is changed.

aircraft band, and the 225-to-406-MHz uhf aircraft communications band, each lie entirely within a single range.

Once the appropriate band has been selected, frequency is controlled by a single dial, the amplifier trimmer control thus passing into history. The output is adjustable from 0.1 microvolt to 10 volts behind 50 ohms, and the shielding challenge implied by such a range of levels has been met.

Highly effective leveling keeps the output constant in the face of changes in frequency or load impedance (Figure 3). The leveling loop is used for envelope feedback for internal or external audio modulation, giving very low modulation distortion. Leveling is also in effect with wide-band externally applied modulation frequencies up to as high as 1.5 MHz, depending on carrier frequency, and with pulse modulation. The use of two buffer stages between oscillator and modulated power amplifier results in unusually low incidental fm in the presence of high-level amplitude modulation. The resulting excellent sideband symmetry is shown in Figure 4.

For added flexibility, the signal generator includes an internal crystal calibrator, a high-level auxiliary output, and provision for electrical fine frequency control for fm or phase-lock operation. The auxiliary output is unusually versatile: It can be used to drive an external counter for monitoring the signal-generator frequency at the same time that a low-level output from the main rf output connection is delivered to a receiver under test; it can be disabled with better than 100-dB isolation by means of an internal coaxial switch, thereby eliminating possible leakage from connected apparatus or cables; it can serve as an input, permitting the signal generator to be used

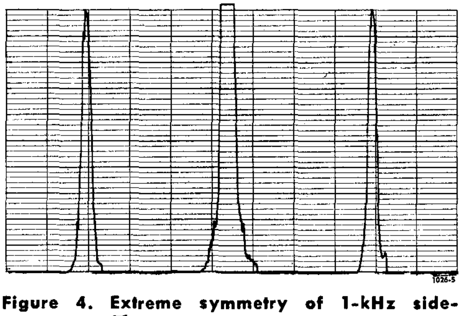


Figure 4. Extreme symmetry of 1-kHz sidebands at 50% modulation is evidence of very low incidental fm. Carrier amplitude (center spike) is offscale because of expansion to show sidebands.

as a heterodyne frequency meter; and it can drive an external phase detector for phase-locked operation.

PRINCIPLES OF OPERATION

The elementary block diagram (Figure 5) shows the rf power-generating stages and those parts of the modulator that serve to level the carrier amplitude and to insert audio modulation. The ancillary circuits employed in wideband and pulse modes of modulation have been omitted from this simplified diagram.

The output stage of the rf power generator and the modulator stages form a negative-feedback loop. Since the loop encloses a signal detector, it is the carrier amplitude or envelope that is fed back rather than the radio-frequency wave itself. The reference against which this loop stabilizes itself is a dc voltage supplied by the CARRIER LEVEL control potentiometer. Modulation voltages can be superposed on this reference, thus forcing the loop to follow an audio-frequency input. Special provisions are made to accommodate wide-band- and pulse-modulating signals that are too fast for the loop to follow.

Radio-Frequency Generation

The output frequency is generated by planar ceramic triodes driven by a Colpitts oscillator and two buffer amplifiers. The Colpitts circuit is conventional except for the guillotine tuning capacitor.[1] The six coils required to cover the 9.5-to-500-MHz frequency range are mounted on a turret that rotates with band changes.

The three lowest frequency ranges cover tuning ratios of 2⅓ to 1, with some overlap between ranges. The plates of the guillotine tuning capacitor are shaped so that frequency varies linearly with dial rotation on these ranges, simplifying interpolation in bandwidth measurements on high-frequency i-f amplifiers and filters. The upper three ranges become progressively narrower in coverage in order to maintain satisfactory interstage tracking and adequate drive levels.

A small amount of electronic tuning is possible through control of the bias on a varactor diode. Owing to the

[1] See "The Guillotine Capacitor," p 17, this issue.

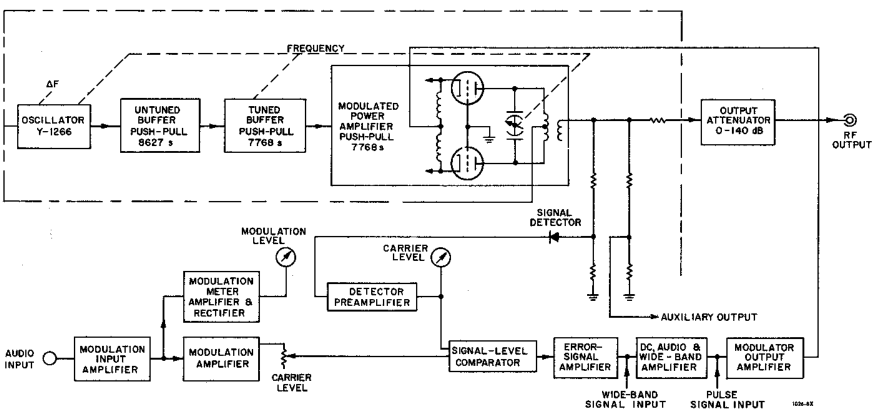


Figure 5. Elementary block diagram.

large signal level at which this diode operates, current is being drawn, and the consequent self-biasing action produces a nearly linear frequency-vs-applied-control-voltage characteristic for small frequency deviations.

Both plate and heater voltages of the oscillator are electronically regulated for minimum residual fm and for maximum stability against line-voltage changes.

The output of the oscillator drives a pair of nuvistor power triodes in a push-pull grounded-grid broadband amplifier stage, which operates at approximately unity gain. This stage effectively isolates the oscillator from reaction due to modulation of the output stage. The untuned buffer in turn drives a push-pull tuned buffer consisting of a pair of high-performance ceramic planar triodes. The tuning capacitor, a second guillotine, and the coil turret are almost identical to those used in the oscillator. This tuned buffer gives the power gain needed to drive the modulated output amplifier.

The modulated output amplifier is almost identical to the tuned buffer. The same tube types are employed, the guillotine tuning capacitor is identical, and the coil turret is similar. The most significant difference is that bias is controlled by the modulator, which is in series with the cathode ground return.

Careful control of the tuned-circuit elements ensures good tracking between stages without front-panel trimmers. Minor detunings, resulting from the inevitable differences in configuration between oscillator and amplifier stages, are controlled to produce similar effects. For example, coil trimming capacitors used on the top two frequency ranges are nominally identical, but they are nevertheless reset automatically by the band-change mechanism to values established during factory alignment of the instrument. High-quality precision gears and tight control of runout in the common drive shaft are also important factors in achieving a tracking accuracy that is typically 0.1% in frequency.

Instrument structure has a first-order effect on the stability, reproducibility, and shielding integrity of a signal generator. The radio-frequency portion of the 1026 is built in a single large casting, with separate pockets for the individual stages (Figure 6). Power and modulation leads are brought in through appropriate low-pass filters, shafts are brought out insulated through wave-guide-below-cutoff pipes, and

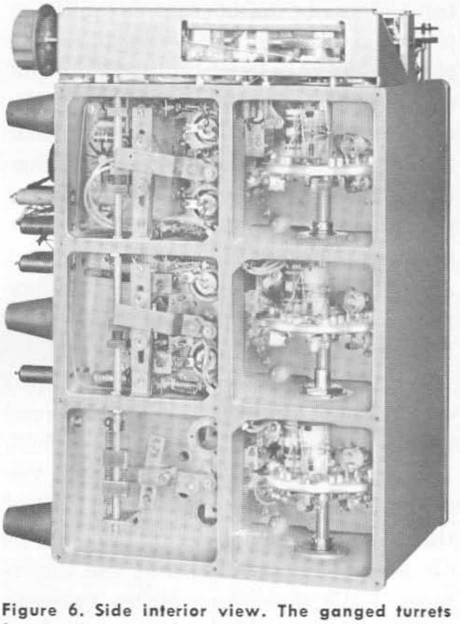


Figure 6. Side interior view. The ganged turrets for the three tuned stages are in the right-hand compartments; the tube sockets for the amplifier and tuned buffer stages are visible in the upper two left-hand compartments.

shield covers have multiple fingers for good contact, with double rows of fingers in selected "hot" locations. The casting supports the individual guillotine tuning capacitors and the precision gear drive that couples them to the front-panel FREQUENCY control. All adjustments and tubes are accessible without major disassembly in case of need for servicing.

Output System

The output of the modulated power amplifier is detected by a crystal diode, whose dc output is amplified and used to drive the CARRIER LEVEL meter. The detected output also serves as the input signal to the feedback-leveling amplifier. The diode is fed from a compensated voltage divider, which reduces the rf level to match the diode ratings yet still keeps it high enough for the diode to operate as a linear peak detector. The divider also isolates the diode somewhat from the rf output and thus reduces the distortion of the carrier by the diode. The automatic leveling makes a zero-impedance Thévenin generator at the driving point for the divider. A 50-ohm resistor between this point and the output attenuator establishes the source impedance in the maximum-output position of the OUTPUT RANGE selector.

Adjustment of the leveled output over a range of up to 20 dB is achieved by the CARRIER LEVEL control, which varies the reference voltage fed to the feedback loop. The uppermost 6 dB of this range is used only in the 10-volt cw mode of operation, and the lowermost 4 dB provides overlap between the 10-dB steps of the attenuator and permits reaching the low-level limit of 0.1 microvolt.

The resistive 10-dB-per-step attenuator includes the input switching required to provide proper impedance characteristics and straight-through operation with zero insertion loss. After an impedance-matching 10-dB input section, the attenuator is a continuous 100-ohm ladder. This provides a 50-ohm output impedance since the output always sees the source and load segments of the ladder connected in parallel. The extremely low VSWR of the output impedance (less than 1.02 over the whole frequency range) is due to close control of resistor values and of physical dimensions, as well as of the size and shape of the shield pockets in which the resistors are located. At the output tap, special rotating shield members are used to prevent pickup of stray leakage from the input, which could reduce the accuracy of the attenuator at the extremely high maximum insertion loss of 140 dB. Resistors are aged and checked for stability to ensure long-term accuracy.

The high-level auxiliary output is taken from the monitoring point by way of a 10:1 divider and through a coaxial switch, which isolates the FREQUENCY METER output connector from the input by over 100 dB when the auxiliary output is not required.

Frequency Monitoring

The 0.5-percent direct-calibration frequency accuracy of the 1026 is adequate for most applications; yet some measurements (e.g., those of steep-sided filter characteristics) do require even greater accuracy. The built-in crystal calibrator improves calibration by at least an order of magnitude. At exact multiples of 1 MHz, the frequency can readily be determined to within 0.01% (the actual crystal frequency is accurate to 0.001%), but practical measurement must include allowances for failure to set to zero beat and for short-term drift. In interpolation between beat points, accuracy is typically better than 0.05%.

Selectivity measurements of narrow-band receivers must sometimes be made directly at the signal frequency rather than at the intermediate frequency. Since in this case the percentage bandwidth is small, it may be necessary to determine individual frequencies to the very high precision possible with a counter. The special switching and shielding provisions associated with the high-level auxiliary FREQUENCY METER output have already been described. As an additional feature, for the user who is annoyed by the noise radiated by some counters, auxiliary contacts on the FREQUENCY CALIBRATOR switch permit automatic quieting of the counter when it is not actually required to count.

The combination of the crystal calibrator and the high-level FREQUENCY METER output makes it possible to use the signal generator as a heterodyne frequency meter. An external signal applied to the FREQUENCY METER connector is mixed with the signal-generator output in the same detector used for the crystal calibrator function, and resulting beats are amplified and delivered to the BEAT OUTPUT jack. The crystal calibrator provides the accuracy required for many measurements, and the audible monitoring of signal quality often yields information that is completely absent in a counter measurement.

Leveling and Modulation

Leveling of the output rf amplifier is now generally accepted as an essential convenience feature in any modern signal generator. It speeds up the measurement process by eliminating an operator adjustment, and it facilitates scanning of a band to observe frequency response characteristics. Substantial improvement in level stability at any given frequency can be achieved as a useful byproduct of leveling. This amplitude stability, extremely important in precision measurements of insertion loss, is shown in Figure 7. The

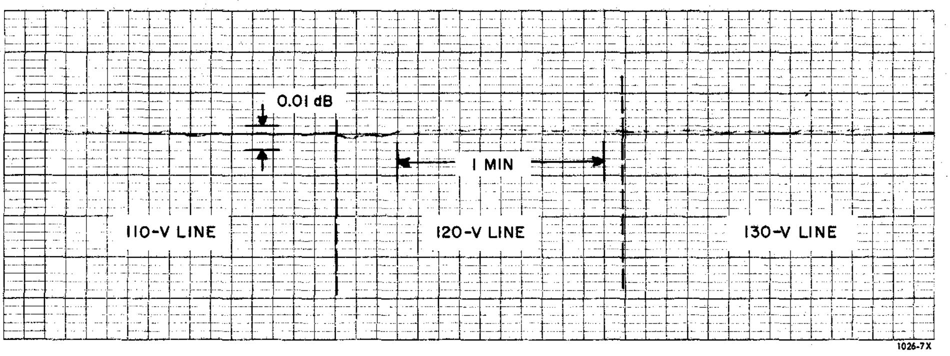
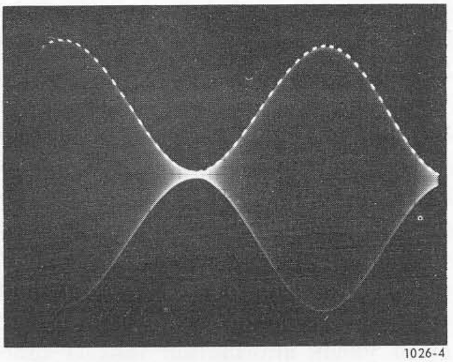


Figure 7. Short-term amplitude stability. The peak amplitude of the envelope of a 400-MHz carrier modulated 95% at 1 kHz, recorded against time with line-voltage steps.



1026-4

Figure 8. Modulation linearity: The envelope of a 400-MHz carrier modulated 95% at 1 kHz is shown with superposed modulation signal. The oscilloscope trace of the 1-kHz modulation signal has been intensity-modulated at a synchronized 30-kHz rate to avoid masking of the modulation envelope.

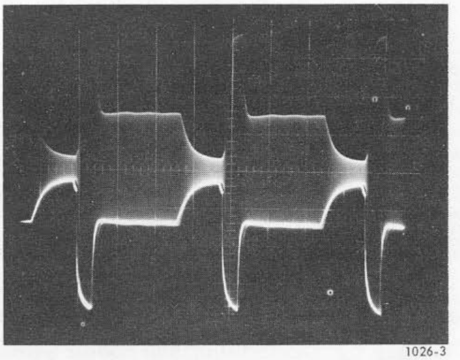
record of the detected modulation envelope vs time and line-voltage steps illustrates the stability of both carrier level and modulation depth.

The accuracy of leveling depends on the frequency characteristics of the detector used to sense the level changes, on the harmonic content of the signal to be leveled and its effect on the detector response, on the magnitude of level changes that require compensation, on the gain of the compensating loop, and on the stability of the dc reference voltage.

Signal harmonics are often ignored in leveling specifications. Moreover, in any circuit with adequate loop gain, their presence is not betrayed by the carrier level meter. In the 1026, appropriate filters reduce harmonics to at least 30 dB below the carrier level. Leveling performance of the 1026 is shown in Figure 3.

Loop gain cannot be increased indefinitely without loss of stable operation. The relation of gain and phase to frequency must be carefully controlled when many stages are enclosed in the feedback loop. For cw and audio-modulation operating modes, envelope feedback is used to reduce residual hum modulation and to provide low envelope distortion at high percentages of modulation. The corner frequency of the loop-gain roll-off is 600 Hz, permitting maximum gain for carrier-level stabilization, while still providing substantial negative feedback to reduce modulation envelope distortion in the normal audio range. In Figure 8, the input audio waveform is superposed on the rf envelope of a 400-MHz carrier modulated 95% at 1 kHz.

Particular attention has been paid to stability of the reference voltage and of the frequency and amplitude of the internal 1-kHz oscillator, in order to maximize the usefulness of the generator in precision loss measurements. A portion of the 1-kHz signal is available for synchronization of oscilloscopes or synchronous detectors.



1026-3

Figure 9. Wide-band modulation: The envelope of a 400-MHz carrier modulated by a complex waveform at a 50-kHz repetition rate is shown with superposed input modulation signal. The 2-μs-duration modulation peak is produced by the negative-going portion of the input signal.

The over-all system amplitude stability shown in Figure 7 includes the effects of the 1-kHz internal oscillator.

The modulator consists of a signal-level comparator followed by a multistage error amplifier to provide the necessary gain and power to control the cathode current of the modulated power-amplifier tubes. One input to the comparator is provided by the signal detector after preamplification. The CARRIER LEVEL meter is connected to this input. The other input from the CARRIER LEVEL control consists of dc only for cw operation or dc with audio superposed for internal 1-kHz or external audio modulation. The modulation level is monitored in terms of the audio voltage superposed on the carrier control voltage.

In the WIDE-BAND mode of operation, long-time-constant networks are added to the feedback loop so that it cannot follow modulation-frequency voltages but can still stabilize the carrier operating point on a dc basis. The modulation voltages are then inserted at a subsequent stage in the loop. A complex wide-band modulation signal and the associated modulation envelope are seen in Figure 9.

In the PULSE mode, the pulses detected in the signal detector are stretched and converted to a dc value

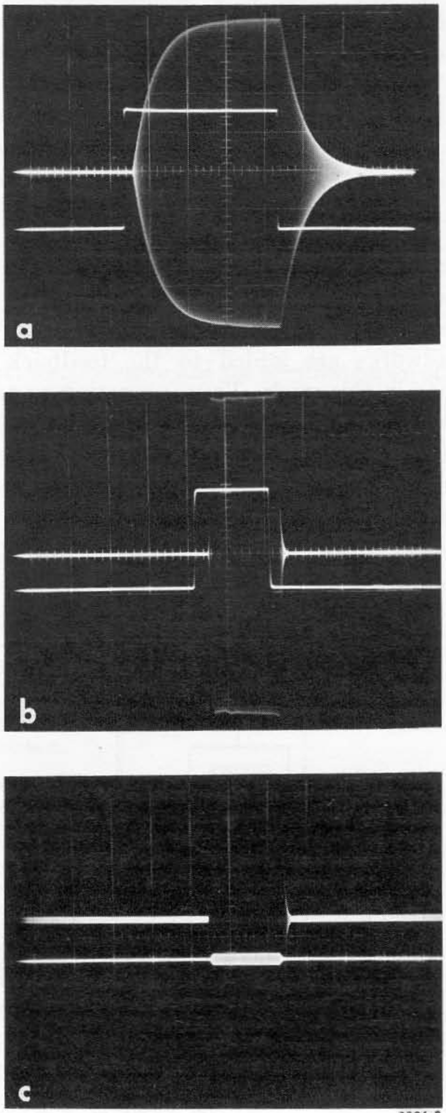


Figure 10. Pulse-modulation characteristics: In (a), the envelope of a 10-MHz carrier modulated by an 8-μs pulse is shown together with the input pulse. Rise and fall times of 3 μs and delays of 0.5 μs are evident from the 2 μs/cm graticule. In (b), the envelope of a 500-MHz carrier modulated by a 2-μs pulse is shown together with the input pulse. Rise and fall times of 0.2 μs and delays of 0.4 μs are evident from the 1 μs/cm graticule. In (c), a composite picture shows an on-off ratio in excess of 50 dB for a 500-MHz carrier modulated by a 2-μs pulse. The upper trace shows the residual signal as a thickening of the base line; the large off-scale deflection during the 2-μs pulse period causes apparent blanking in the center of the trace. The lower trace shows the same pulse with the output attenuator set for a 50-dB reduction in level. The peak amplitude of the lower trace is greater than the interpulse amplitude of the upper trace, demonstrating an on-off ratio in excess of 50 dB.

Gordon McCouch graduated from Harvard University in 1941 and obtained his MA in 1948. After a year with the Radio Research Laboratory he went to England for the office of Scientific Research and Development, later serving as a Technical Observer in continental Europe. From 1945 to 1957 he was with Aircraft Radio Corporation. Since then he has been a member of the GR Engineering Department; he is now Section Leader responsible for the design of signal generators.

He is a Senior Member of IEEE and is currently a member of the Sections Committee of IEEE.

corresponding to their peak amplitude before they are delivered to the signal-level comparator. Long-time-constant networks are added to the feedback loop in this mode. Thus, operating as a dc amplifier, the loop establishes a clamping level to which input pulses can drive the output amplifier. These input pulses are inserted into the loop just before the final power stage of the modulator. A typical pulse-modulation envelope is shown in Figure 10.

Power Supply

In the 1026, power-supply dissipation present in conventional regulated supplies is minimized by a new preregulator that controls the input to the primary of the power transformer (see Figure 11). This preregulation stabilizes the heater voltages of all tubes, thereby contributing to long-life performance, and reduces the swings that the electronic postregulators have to accommodate. The power dissipated in the instrument is only about 90 watts and is almost independent of input line voltage. The power supply operates satisfactorily over the 50-to-60-Hz line-frequency range and can be connected for either 115- or 230-volt operation.

The preregulator, like a conventional electronic regulator, compares an output dc voltage with a reference and amplifies the resultant error voltage.

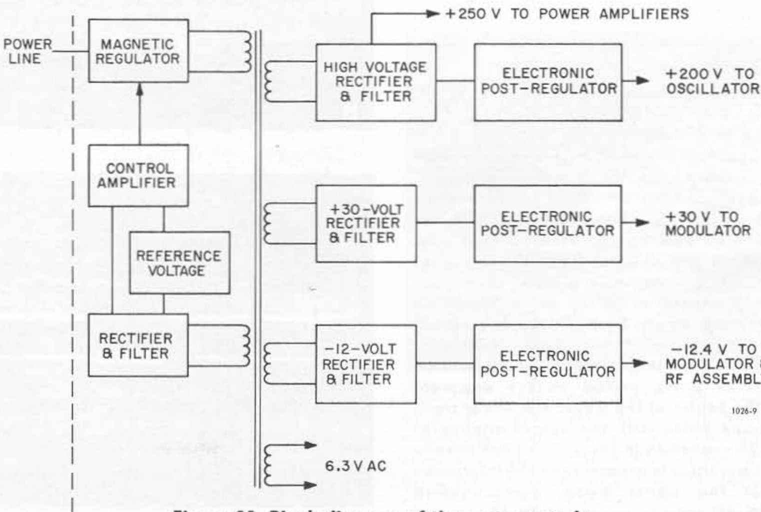


Figure 11. Block diagram of the power supply.

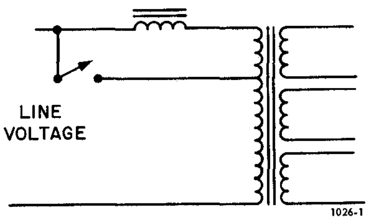


Figure 12. Elementary schematic showing pre-regulator tap-switching scheme.

But, instead of controlling a series losser element, the control voltage operates a magnetic amplifier that switches the input line between high- and low-voltage taps on the power-transformer primary, in such a way that the average output voltage is maintained at the desired value (see Figure 12). Actually, the high-voltage tap is always connected through a series inductor, which saturates and thus has a low drop during the part of the cycle when it carries the line current to the high-voltage tap. During the remainder of the cycle, it floats across the portion of the transformer between primary taps and carries only magnetizing current. At this time, the line current is transferred to the low-voltage tap through the magnetic-amplifier inductor.

Since only a portion of the transformer primary is switched by this regulator, waveform distortion is not severe, and the peak-to-rms ratio is not grossly altered. Thus good regulation of both ac heater power and dc output is realized.

— G. P. McCouch

ACKNOWLEDGMENTS

The closest collaboration between the author and S. Brown Pulliam has characterized the entire development of the 1026 Signal Generator; the attenuator and the electrical design of the guillotine tuning capacitors were contributed by Andrew P. Lagon.

TENTATIVE SPECIFICATIONS

FREQUENCY

Range: 9.5 to 500 MHz in 6 ranges: 9.5 to 22, 22 to 48, 48 to 108, 108 to 220, 220 to 420, and 400 to 500 MHz.

Manual Control: Main frequency control, spinner knob with 100-division vernier dial (25 turns per range) drives main drum-type dial. Illuminated scale indicates selected range. Parallax-free fiducial mark is adjustable for fine calibration. Scales to 108 MHz are linear. An uncalibrated Δf control spans typically ±0.003% at low end of range to ±0.015% at high end (actual spans may vary 2:1 depending on frequency range).

Scale Characteristics:

Frequency Range (MHz)	*Main Scale Interval*	*kHz per Vernier Division*	*Scale Length (in)*
9.5-22	100 kHz	5	14¼
21.2-49.6	200 kHz	11	14¼
47.4-111	500 kHz	25	14¼
100-220	1.0 MHz	45-60	13
216-430	2.0 MHz	80-150	10½
400-500	2.0 MHz	150	4

External Electrical Fine Frequency Control: Applied voltage of ±20 V dc varies frequency typically ±0.04% at low end of range to ±0.2% at high end (actual variation may differ by 2:1 depending on frequency range).

Calibration Accuracy: ±0.5% direct reading, after initial adjustment of fiducial. With internal crystal calibrator, ±0.01% at 1.0-MHz intervals, typically ±0.05% by interpolation.

Calibration Provisions: Internal crystal frequency, accurate to ±0.001%, provides calibration at intervals of 1 and 5 MHz over entire frequency range. Calibration by external counter provided for by output of about 0.1 to 1 V behind 50 Ω. When not needed, this output can be disabled with >100-dB isolation; external counter can be simultaneously disabled by a contact closure provided to eliminate interference from the counter's internal signals.

Harmonic Output: At least 30 dB below carrier.

RF OUTPUT

Range: CW, 0.1 μV to 10 V behind 50 Ω, ½ W into 50 Ω (−133 to + 27 dBm); modulated, 0.1 μV to 5 V behind 50 Ω (−133 to + 21 dBm). Load VSWR > 2.0 may restrict the max output available at some frequencies.

Control: Step attenuator, 140 dB in 10-dB steps, voltage and dBm calibration. Continuous interpolation with metered level control.

Meter Scales: 0.3 to 1.5 V, 1.0 to 5.0 V, and −13 to +1 dBm. Scale extensions (in red), for cw use only, to 10 V and to +7 dBm.

Accuracy: Metering, ±5% to 108 MHz; above 108 MHz, harmonics can add ±3% and rectifier characteristic can add ±2%. Attenuator, ±1% (±0.1 dB) per step; max accumulated error ±0.5 dB.

RF Interference: Leakage has negligible effect on measurements of receiver sensitivity down to 0.1 μV.

Leveling: CW output is held at preset level to within ±2% (0.2 dB) up to 108 MHz and to within ±5% (0.5 dB) to 500 MHz as frequency is varied, including effects due to range switching. Effectiveness of leveling under modulated operation is a function of modulation mode and frequency.

Stability: At any given frequency, in cw operation or internal 1-kHz modulation mode, and after 2-hour warmup, output will typically remain constant within ±0.0025 dB per minute, or ±0.01 dB over any 15-min period. Also under these conditions, variation due to ±10% line-voltage fluctuation is < ±0.005 dB.

Effective Generator Impedance (at panel jack): 50 Ω resistive; VSWR is <1.02 with output attenuator set for 0 dBm or less. At higher outputs, source impedance viewed as Thévenin generator has a VSWR < 1.2.

MODULATION

Modes:

Amplitude Modulation is provided in four modes:

1. Internal 1 kHz. Modulation level adjustable 0 to > 95% and metered to within ±3% of reading ±2% of full scale. Envelope feedback provides leveling and holds distortion to < 1% at 50% modulation and < 3% at 80% modulation. Modulating frequency, 1 kHz ±0.5%; after 2-hour warmup stable to better than 0.1% over 8-hour period or for line-voltage variations of ±10%. 1 kHz signal available at MOD binding posts, about 2.5 V behind 100 kΩ.

2. External Audio. Response flat to dc, down < 3 dB at 20 kHz. Square-wave response 0 to 10 kHz; rise and fall times < 10 μs; overshoot < 10%; rampoff negligible. Modulation level is adjustable 0 to > 95% for dc to 5-kHz input, to > 50% at 20 kHz, and is metered to within ±5% of reading ±5% of full scale for sine-wave inputs from 20 Hz to 20 kHz. For 95% modulation < 3 V, peak required into 3 kΩ. Envelope feedback provides leveling and holds distortion at 50% modulation to < 1% up to 1 kHz, < 5% up to 10 kHz.

3. External Wide Band. Modulation level adjustable 0 to > 80%. Response flat to ± 3 dB for 50-Hz to 1.5-MHz inputs at carrier frequencies above 108 MHz. Average carrier is leveled and metered, but modulation depth and linearity should be monitored externally. For full modulation, about 0.6 to 3.5 V (depending on carrier frequency) is required into 3 kΩ.

4. External Pulse. Required input pulses, at least 10 V peak, positive going (max 30 V); repetition rate 500 Hz to 150 kHz; duration 1 to 300 μs (min 3 μs on 9.5- to 22-MHz range); max 50% duty ratio. Input impedance 3 kΩ. Output pulse, duration within ±0.5 μs of input; rise and fall times < 1 μs each on all ranges but 9.5 to 22 MHz (up to 3 μs); rampoff < 5%. On-off ratio > 30 dB and at max output setting of attenuator is typically > 40 dB. Peak amplitude of pulses is leveled and metered to within ±1 dB added to accuracy specified for cw leveling.

Incidental FM (accompanying a-m): < 1 ppm, peak, at 1 kHz, 50% a-m.

Residual FM: < 0.05 ppm, peak.

Residual A-M: At least 70 dB below carrier level in cw, internal 1 kHz and external audio modes.

GENERAL

Power Required: 105 to 125 or 200 to 250 V, 50 to 60 Hz, 90 W.

Terminals: RF and counter outputs are GR874 Coaxial Connectors, recessed and locking; for rapid conversion to other common types, use locking GR874 adaptors. Modulation connection is to front-panel binding posts and rear-panel multiterminal connector. Audio (BEAT) output from front-panel telephone jack. Electrical frequency control is through rear-mounted 12-pin connector.

Accessories Supplied: Type 874-R22LA Patch Cord (GR874 to GR874), phone plug, 12-pin connector plug, CAP-22 power cord, spare fuses, hardware for bench and rack mounting.

Mounting: Rack-bench cabinet.

Dimensions: (width x height x depth): Bench, 19 x 17¾ x 15¼ in (485 x 450 x 390 mm); rack, 19 x 17½ x 13 in (485 x 445 x 330 mm).

Weight: Net, 96 lb (44 kg); shipping (est), 180 lb (80 kg).

Catalog Number	*Description*	*Price in USA*
1026-9701	**Type 1026 Standard-Signal Generator**	**$6500.00**

THE GUILLOTINE CAPACITOR

The frequency range between 200 and 1000 MHz presents special challenges in the design of wide-range tuned circuits. Cavities become extremely bulky, and their ranges are difficult to extend by bandswitching techniques; operation in both $\lambda/4$ and $3\lambda/4$ modes can extend the tuning range, especially to higher frequencies, as in the GR 1360 signal source, which tunes from 1700 to 4200 MHz. Lumped LC circuits usually end up with larger inductance than is desired in the face of the inevitably significant minimum capacitances established by the active elements required for power generation or amplification. For many years GR has successfully employed the butterfly circuit,[1] an integral LC tuner, to provide tuning ranges up to 5:1 in the frequency range up to about 1000 MHz, but this circuit does not lend itself readily to major extension of frequency range by bandswitching.

These considerations led to a reexamination of tuning-capacitor design during the development of the 1026 Standard-Signal Generator (see page 3). The result was a new design, a translatory-motion tuning capacitor consisting of a pair of stators and a sliding plunger in place of a rotor (Figure 1). Given this configuration and its sliding action, it could not escape being nicknamed "guillotine." As far as we know, previous efforts to build this type of capacitor have never been carried to successful commercial realization, presumably because of the mechanical problems inherent in building a suitable carriage to support the plunger and in driving it.

[1]E. Karplus, "The Butterfly Circuit," *General Radio Experimenter*, October 1944.

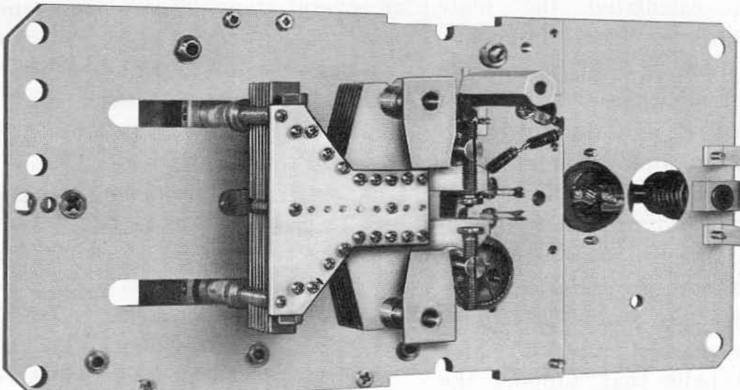


Figure 1. The guillotine capacitor, as used in the 1026 Standard-Signal Generator.

Nevertheless, our preliminary work showed that we could build a one-band oscillator to tune to at least 700 MHz with the guillotine, compared with a top frequency of about 500 MHz with a rotary capacitor. This was the margin we sought in order to design for a 500-MHz top frequency in our new signal generator. The margin was needed to permit the added inductance required for bandswitching coil contacts and to allow for the trimmers necessary to provide tracking of oscillator and amplifier circuits without the need for operator trimmer adjustment. Furthermore, the plates could be shaped to provide a linear relation of frequency to dial indication without intermediate cams for law conversion, and the plates could be supported at their extremities to minimize microphonics. These advantages of the guillotine design persuaded us to tackle the associated mechanical problems.

The guillotine is a balanced structure suitable for both single-ended-oscillator and push-pull-amplifier service with but minor differences. Andrew P. Lagon, of our engineering staff, who suggested that we use the guillotine configuration, calculated the plate shapes to produce the linear tuning law, which most users find attractive because channel allocations within any one service are of uniform width and spacing. Four ceramic rods of high-strength alumina support the pair of stators on a sturdy aluminum base plate, which has been stabilized prior to final machining of critical surfaces. The base plate also carries a pair of hardened, vee-grooved rails with three stainless-steel balls that support the plunger carriage. One of the rails is spring-loaded to eliminate vertical and horizontal play in the carriage. The carriage is coupled to a rack, also supported by the base plate. The plunger and stators are soldered assemblies of precision-tolerance flat plates and grooved spacer rods.

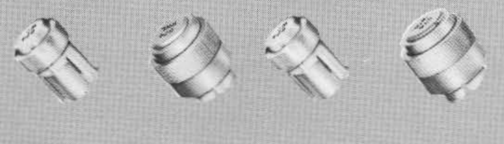
A reasonably large air gap (0.022 inch) and careful soldering to minimize plate distortion make possible very uniform capacitance characteristics from one unit to the next. Nevertheless, the very tight tracking requirements require adjustment of each capacitor to a predetermined schedule using a segmented top plate on the plunger. Trimming for tube-capacitance variations is accomplished by an auxiliary balanced capacitor whose stator plates are supported directly by the main guillotine stator support rods. In order to minimize connection inductance, the tube plate connectors are integral with the guillotine, and the spring contacts that mate with the coil turret contacts are supported at the narrow end of the stators.

The stability and reproducibility required of the capacitors for successful application in the 1026 Signal Generator were verified by extensive tests at several stages of evolution and on substantial numbers of capacitors built for the first lot of signal generators. This low-inductance design results in such good tracking of the three tuned stages in the new signal generator that true single-dial frequency control is now available to 500 MHz.

— G. P. McCouch

ACKNOWLEDGMENTS

The successful design of the guillotine capacitor was the result of cooperation between Andrew Lagon, Richard Mortenson, David Foss, and the author.



NEW GR874 TERMINATIONS

Figure 1. *(left to right)* **Types 874-WO and -WOL Open-Circuit Terminations and Types 874-WN and -WNL Short-Circuit Terminations.**

In coaxial measurements, it is often necessary to set up a short or open circuit at a specific point in a coaxial line. How accurately that point, called the reference plane, is known is one of the most important characteristics of a short- or open-circuit termination. The GR874 short- and open-circuit terminations (Figure 1) have been redesigned to establish the reference plane more accurately and with less variation with frequency. Locking versions have also been added.

The objective is to place the reference plane exactly at the front face of the support bead of the GR874 connector on the unknown under test (see Figure 2). How well this objective has been met is clearly shown in Figure 3, which compares the reference-plane deviations of the old and new open-circuit terminations.

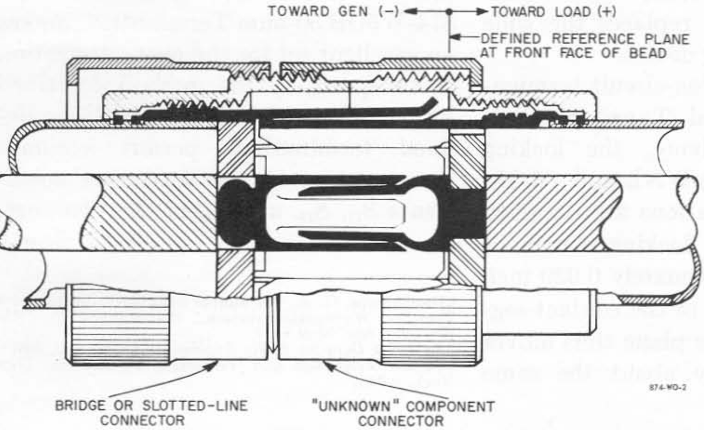


Figure 2. Cross-section of mated GR874 connectors defining reference plane.

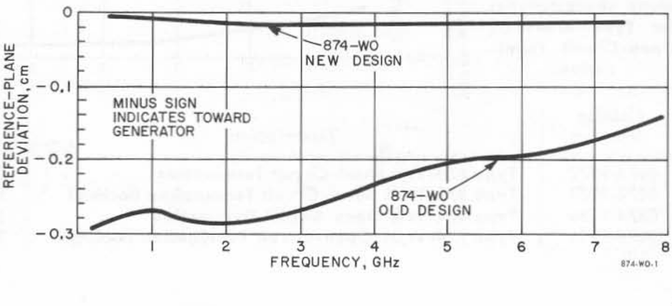


Figure 3. Typical reference-plane-deviation vs frequency for old and new Type 874-WO Open-Circuit Terminations.

the GR Experimenter

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In the open-circuit termination, the key to the improved performance is a built-in cylinder made of high-density polyethylene. This cylinder both provides electrical lengthening and compensates for errors in characteristic impedance.

In the redesigned short-circuit termination, a solid disk replaces the wide strip as the shorting device.

The short- and open-circuit terminations are designated TYPES 874-WN and -WO, respectively; the locking versions are the 874-WNL and -WOL. The locking terminations are designed so that the mating locking connector is disengaged approximately 0.020 inch to prevent jamming of the contact segments. The reference plane thus moves toward the load by about the same amount. Reference-plane characteristics for the 874-WOL appear in Figure 4.

Applications

Beyond the usual applications for these terminations, the combination of 874-WN and -WO, together with the 874-W5OB 50-ohm Termination, makes an excellent set for the characterization of two-ports by the method described by DesChamps[1],[2]. The accurately defined terminations permit accurate measurement of the scattering coefficients S_{11}, S_{12}, and S_{22} of the two-port.

— J. ZORZY

[1] DesChamps, G. A., "A Simple Graphical Analysis of a Two-Port Waveguide Junction," *Proceedings of the IRE*, No. 42, p 859, May 1954.
[2] *Reference Data for Radio Engineers*, 4th Edition, International Telephone and Telegraph Corporation, New York, p 649.

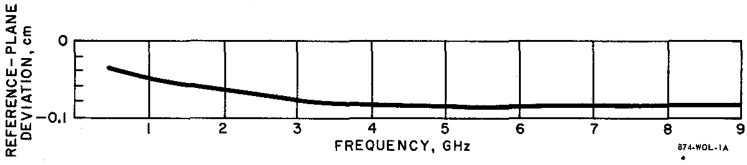


Figure 4. Reference-plane characteristics for Type 874-WOL Open-Circuit Termination.

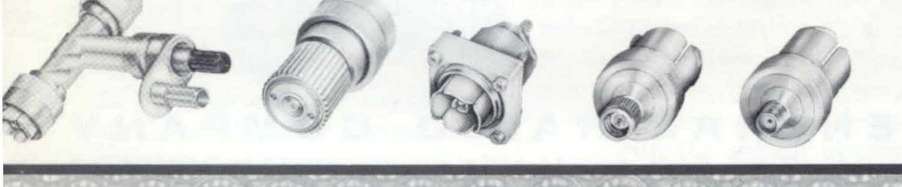
Catalog Number	*Description*	*Net Weight*	*Price in USA*
0874-9970	**Type 874-WN Short-Circuit Termination**	1 oz (30 g)	$4.50
0874-9971	**Type 874-WNL Short-Circuit Termination (locking)**	1½ oz (45 g)	5.75
0874-9980	**Type 874-WO Open-Circuit Termination**	1 oz (30 g)	4.00
0874-9981	**Type 874-WOL Open-Circuit Termination (locking)**	1½ oz (45 g)	5.25

THE GENERAL RADIO

Experimenter

NEW GR900

AND GR874



COMPONENTS

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RESISTIVE VOLTAGE DIVIDERS

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REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

COAXIAL MICROWAVE NEWS

NEW GR900 COMPONENTS

The GR900 and GR874 lines of coaxial devices continue to expand. The GR900 14-mm precision connector is already backed up by the most extensive line of precision components available, and we now add a precision ell, 13-inch sections of inner-conductor rod for use in fabricating precision air lines, and adaptors to 7-mm precision connectors.

To the long-popular GR874 line we add a bias insertion unit, rod and tube for fabricating air lines, a keyed panel connector, and several new adaptors.

PRECISION ELL

Figure 1.

In precision coaxial measuring systems, the simple matter of going around corners is not so simple, and very careful design is necessary to achieve a right-angle turn that will not introduce reflections. With the introduction of a GR900 precision ell, it is now possible to make a 90-degree turn with a residual VSWR of less than 1.01 at 1.5 GHz and less than 1.02 at 4 GHz.

The change in direction takes place in a transmission line whose axis describes a 90° circular arc. The uniformly varying change in direction along the arc results in an essentially uniform characteristic impedance.

In the curved region of the ell, the conductors have square cross sections. Where these sections join the standard 14-mm line of round cross sections, coplanar compensation is employed. The ell is, of course, equipped with GR900 precision connectors.

The electrical length of the ell is nominally 10 cm, but, because of the finite curvature, the electrical length increases with increasing frequency.

Uses

The ell is especially useful in systems involving complex interconnections where it is necessary to minimize reflections and to maintain phase linearity as, for instance, in precision phase- and attenuation-measuring systems.

For measurements of dielectric properties with the 900-LB Slotted Line,[1] it is not always convenient to connect the sample holder directly to the slotted line. If, for instance, the dielectric to be measured is a liquid, the sample holder usually must be vertical. Vertical orientation is also often necessary for sample holders placed in environmental chambers. In such applications the ell connects the sample holder to the slotted line with very little loss in accuracy.

[1] J. F. Gilmore, "Measurements of Dielectric Materials with the Precision Slotted Line," *General Radio Experimenter*, May 1966.

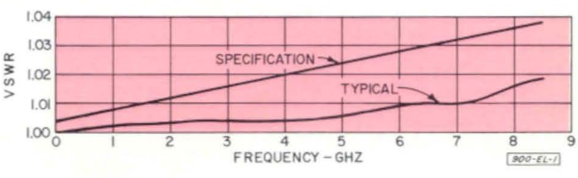


Figure 2. VSWR characteristics of precision ell.

SPECIFICATIONS

Frequency Range: Dc to 8.5 GHz.

Characteristic Impedance: 50 Ω ±0.4% at frequencies where skin effect is negligible.

VSWR: Less than $1.004 + 0.004 f_{GHz}$. See curve.

Electrical Length: $[10.00 + 0.0014 (f_{GHz})^2 \pm 0.02]$ cm.

Insertion Loss: Less than $0.017\sqrt{f_{GHz}}$ dB.

Maximum Voltage: 1500 V peak.

Maximum Power: 10 kW up to 1 MHz; $10\ kW/\sqrt{f_{MHz}}$ above 1 MHz.

Mating Dimensions: 2.066 in. (5.246 cm) from center line of one connector to reference plane of second connector.

Over-all Dimensions: 2 11/16 by 2 11/16 by ⅞ in. (68 by 68 by 22 mm).

Net Weight: 10 oz (280 g).

Catalog Number	*Description*	*Price in USA*
0900-9527	**Type 900-EL Precision 90° Ell**	**$180.00**

ADAPTORS TO 7-MM CONNECTORS

Two new adaptors permit interconnection of GR900 and 7-mm coaxial connectors.

The Type 900-QAP7 mates GR900 with the Amphenol APC-7 connector, and the Type 900-QPF7 mates GR900 with the Rohde & Schwarz Precifix A and Dezifix A Connectors.

SPECIFICATIONS

Frequency Range: Dc to 8.5 GHz.

Characteristic Impedance: 50.0 Ω nominal.

VSWR: Less than $1.003 + 0.002 f_{GHz}$.

Maximum Voltage: 1000 V peak.

Maximum Power: 6 kW up to 1 MHz; $6\ kW/\sqrt{f_{MHz}}$ above 1 MHz.

Dimensions: Length 2⅛ in. (54 mm); max dia 1¹⁄₁₆ in. (27 mm).

Net Weight: 3½ oz (100 g).

Catalog Number	*Description*	*Price in USA*
0900-9791	**Type 900-QAP7 Adaptor** (GR900 to APC-7)	**$110.00**
0900-9793	**Type 900-QPF7 Adaptor** (GR900 to Dezifix/ Precifix A)	**110.00**

PRECISION INNER-CONDUCTOR RODS

Precision rod and tube have been available from GR for some time for the custom fabrication of precision air lines, sliding loads, sample holders, short and open circuits, etc. The 27-inch precision rod formerly supplied is now replaced by 13-inch lengths, available in pairs by catalog number 0900-9507. The shorter lengths allow us to maintain more rigid control of critical dimensions.

A limited supply of the old 27-inch rods (No. 0900-9508) is still on hand, and these longer rods will be shipped on order until the supply is exhausted.

Precision outer-conductor tube (No. 0900-9509) will continue to be sold in 27-inch lengths.

SPECIFICATIONS

Materials: Centerless-ground, stress-relieved, silver-layered brass rod (two supplied).

Outer Diameter: 0.24425 inch.

Accuracy of Diameter: ± 65 microinches.

Uniformity of Diameter: ± 25 microinches.

Surface Finish: 20 microinches, max.

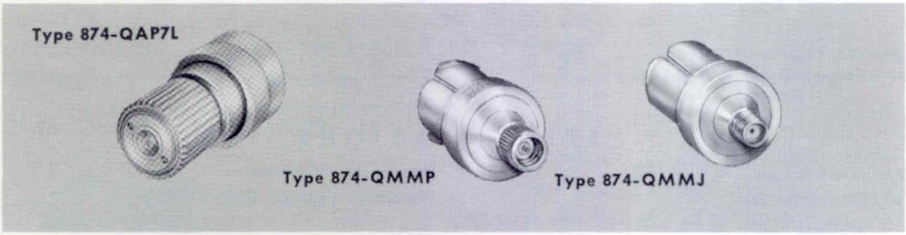
Straightness: 0.003 TIR, max.

Length: 13 inches (330 mm). **Weight:** 7 oz (200 g).

Catalog Number	*Description*	*Price in USA*
0900-9507	**Precision Coaxial Rod**	**$22.00 per pair**

NEW GR874 COMPONENTS

NEW ADAPTORS



Adaptors are available from GR874 to OSM-type miniature connectors and to Amphenol APC-7 precision 7-mm connectors.

Four GR874-to-OSM adaptors cover the mating requirements to both male and female OSM's in locking and nonlocking versions. These adaptors mate with the following connector types: ASM, BRM, ESCAM, MICRO, MOB-50, NPM, OSM, SRM, STM, and GRM. The VSWR characteristics (Figure 3) apply only when the mating connector has the mating dimensions shown in Figure 4.

The new TYPE 874-QAP7L Adaptor mates a locking or nonlocking GR874 connector with an Amphenol APC-7 7-mm connector. The typical VSWR of a single adaptor is shown in Figure 5.

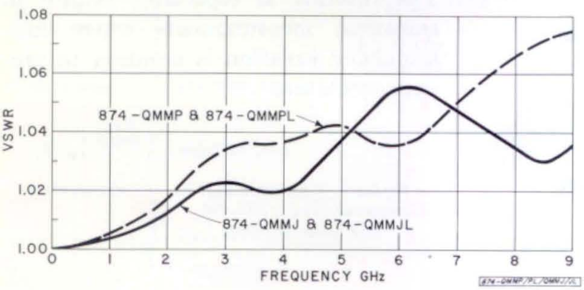


Figure 3. VSWR characteristics of Types 874-QMMJ and -QMMP Adaptors. (Values are typical for single adaptors.)

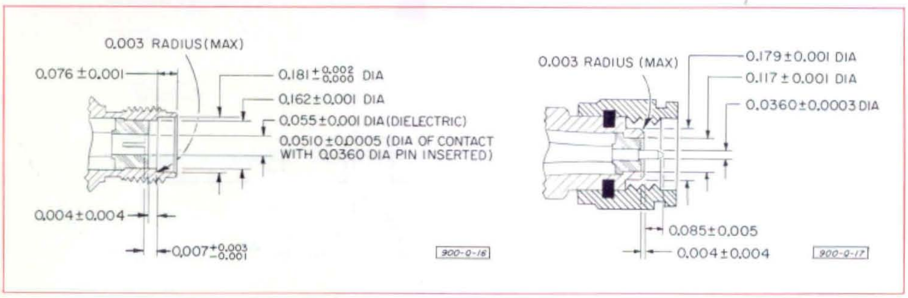


Figure 4. Critical mating dimensions for Types 874-QMMJ and -QMMP Adaptors.

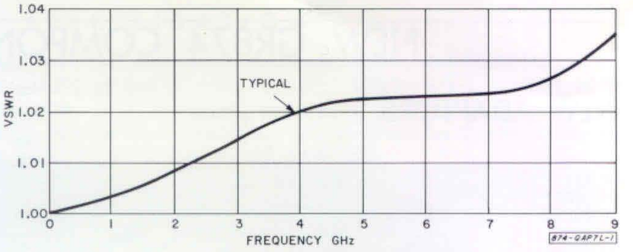


Figure 5. Typical VSWR characteristics of Type 874-QAP7L Adaptor.

SPECIFICATIONS

Frequency Range: Dc to 9 GHz.

VSWR: See curves.

Impedance: 50 ohms.

Maximum Voltage: 1000 V peak.

Catalog Number	*Description*	*Includes*	*Mates With*	*Price in USA*
0874-9722	**Type 874-QMMJ Adaptor**	OSM jack	OSM plug	**$16.00**
0874-9723	**Type 874-QMMJL Adaptor, Locking**	OSM jack	OSM plug	**17.25**
0874-9822	**Type 874-QMMP Adaptor**	OSM plug	OSM jack	**22.00**
0874-9823	**Type 874-QMMPL Adaptor, Locking**	OSM plug	OSM jack	**23.25**
0874-9791	**Type 874-QAP7L Adaptor**	Amphenol APC-7		**60.00**

BIAS INSERTION UNIT

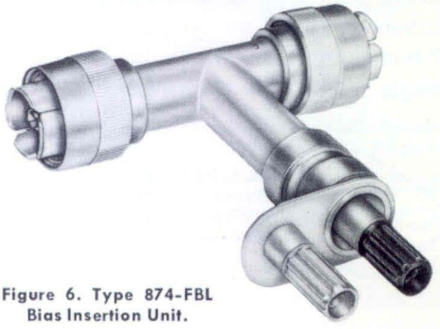


Figure 6. Type 874-FBL Bias Insertion Unit.

The TYPE 874-FBL Bias Insertion Unit (Figure 6) is used to bias coaxial devices and is a valuable component in slotted-line immittance measurements on semiconductors. It consists of a coaxial tee with dc blocking in one arm and filtering in the other (see Figure 7). The filtering is especially helpful in transistor measurements where low-frequency isolation is required to prevent oscillation.

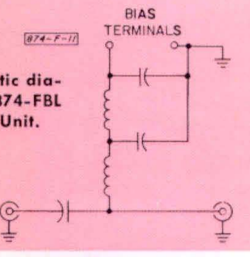


Figure 7. Schematic diagram of Type 874-FBL Bias Insertion Unit.

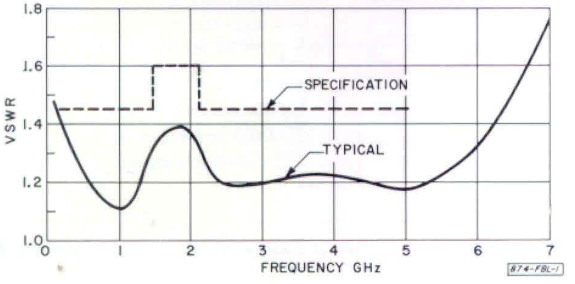


Figure 8. VSWR characteristics of Type 874-FBL Bias Insertion Unit.

SPECIFICATIONS

Current Rating: 2.5 A.
Voltage Rating: 400 V.
VSWR: See curve.
Insertion Loss: Typically less than 1.7 dB from 300 MHz to 3 GHz (except 2 dB at approx 1.8 GHz), less than 0.8 dB from 3 to 5 GHz.
Dimensions: 4⅜ by 3⅞ in (115 by 99 mm).
Net Weight: 6½ oz (185 g).

Catalog Number	*Description*	*Price in USA*
0874-9759	**Type 874-FBL Bias Insertion Unit**	**$75.00**

NEW KEYED PANEL CONNECTOR

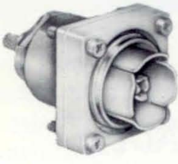


Figure 9. Type 874-PBRL58A Keyed Panel Connector.

A new recessed, locking GR874 panel connector has been designed for those who require positive, fixed orientation of the connector with respect to the panel. In the new 874-PBRLA family of connectors, a key on the screw-mounted panel flange engages a keyway on the side of the connector, so that when the assembly is completed the connector cannot rotate, even if the clamping nut loosens.

Five models of this new connector are available to cover a wide range of cable sizes. General specifications are the same as those for other GR874 panel connectors listed in our catalog.

Catalog Number	*Type*	*Fits*	*Price in USA*
0874-9481	**874-PBRLA**	874-A2 cable	**$6.60**
0874-9483	**874-PBRL8A**	(50-ohm) RG-8A/U, -9B/U, -10A/U, -87A/U, -116/U, -156/U, -165/U, -166/U, -213/U, -214/U, -215/U, -225/U, -227/U; (non 50-ohm) RG-11A/U, -12A/U, -13A/U, -63B/U, -79B/U, -89/U, -144/U, -146/U, -149/U, -216/U cables	**6.60**
0874-9485	**874-PBRL58A**	(50-ohm) 874-A3, RG-29/U, -55/U (series), -58/U (series), -141A/U, -142A/U, -159/U, -223/U cables	**6.60**
0874-9487	**874-PBRL62A**	(non 50-ohm) RG-59/U, -62/U (series), -71B/U, -140/U, -210/U cables	**6.60**
0874-9489	**874-PBRL174A**	(50-ohm) RG-174/U, -188/U, -316/U; (non 50-ohm) RG-161/U, -179/U, -187/U, -298/U cables	**7.45**

ROD AND TUBING

Rod and tubing are now available for fabricating custom GR874 components and air lines. The tubing is a 15⅞-inch section of Alballoy-plated brass, with an outer diameter of 0.624 (+0.000, −0.002) inch and an inner diameter of 0.5625 (±0.0010) inch. The rod, also 15⅞ inches long, is of high-conductivity silver-plated brass, with a diameter of 0.24425 (±0.00025) inch. Characteristic impedance of a coaxial line made up of this rod and tubing is 50 ohms ±0.375%. Ends are already machined to accept 874-B or -BBL connectors, and machining instructions are included for shorter sections.

Catalog Number	*Description*	*Price in USA*
0874-9508	**Inner-Conductor Rod**	**$4.00**
0874-9509	**Outer-Conductor Tubing**	**4.00**

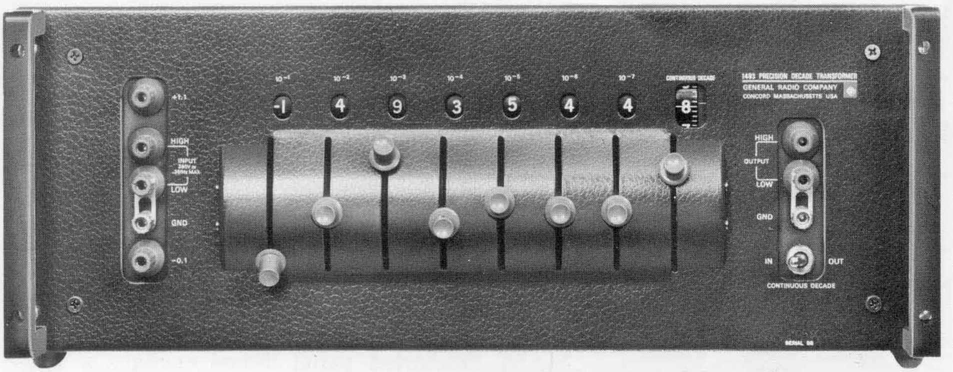


Figure 1. Type 1493 Precision Decade Transformer.

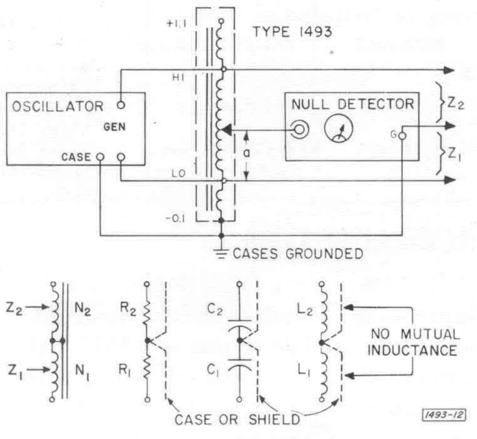
THE TYPE 1493 PRECISION DECADE TRANSFORMER

The new TYPE 1493 Precision Decade Transformer (Figure 1) is much more than just another "ratio box." Its accuracy (± 2 digits in the 10^{-7} decade), range (-0.1111111 to $+1.11111110$), resolution better than 1 part in 10^9 (with auxiliary equipment of appropriate sensitivity) and convenient, lever-switched, in-line readout set it apart from conventional ratio transformers.

Ratio transformers have been around for a long time. Most of our readers are probably familiar with their fundamental use in the measurement of an unknown turns ratio or of the magnitude ratio between two similar impedances (Figure 2). The addition of the resistor and capacitor shown in Figure 3 permits a closer measurement, by narrowing the null through phase balance. More complex circuits permit ratio measurements with a repeatability of a few parts in 10^9.

Acceptable for calibration by the National Bureau of Standards, the 1493 can be used as a primary standard to calibrate other ratio transformers. In educational and experimental laboratories it can be used as two of the adjacent ratio arms in many different

Figure 2. Diagram illustrating use of the 1493 in measurement of the ratio between the magnitudes of two impedances.



$$a = \text{ratio} = \frac{N_1}{N_1 + N_2} = \frac{R_1}{R_1 + R_2} = \frac{C_2}{C_1 + C_2} = \frac{L_1}{L_1 + L_2} = \frac{Z_1}{Z_1 + Z_2}$$

$$\frac{a}{1 - a} = \frac{N_1}{N_2} = \frac{R_1}{R_2} = \frac{C_2}{C_1} = \frac{L_1}{L_2} = \frac{Z_1}{Z_2}$$

(Reversal of C subscripts caused by $X_C = \frac{1}{j\omega C}$.)

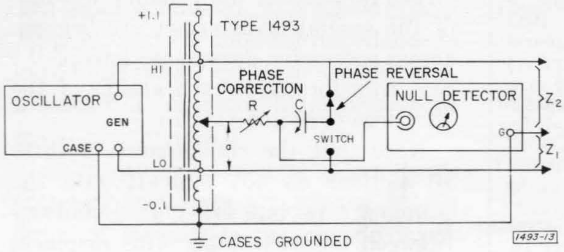


Figure 3. Setup similar to that of Figure 2, but with resistor and capacitor added for phase balance.

transformer bridge circuits for accurate impedance measurements.

Since the 1493 accuracy is basically determined by fixed turns ratios and by the relatively invariant properties of magnetic cores, no appreciable degradation of accuracy with time should occur. Calibration should literally last a lifetime, barring accidents.

The departures from tradition can be sensed from a look at the front panel (Figure 1).

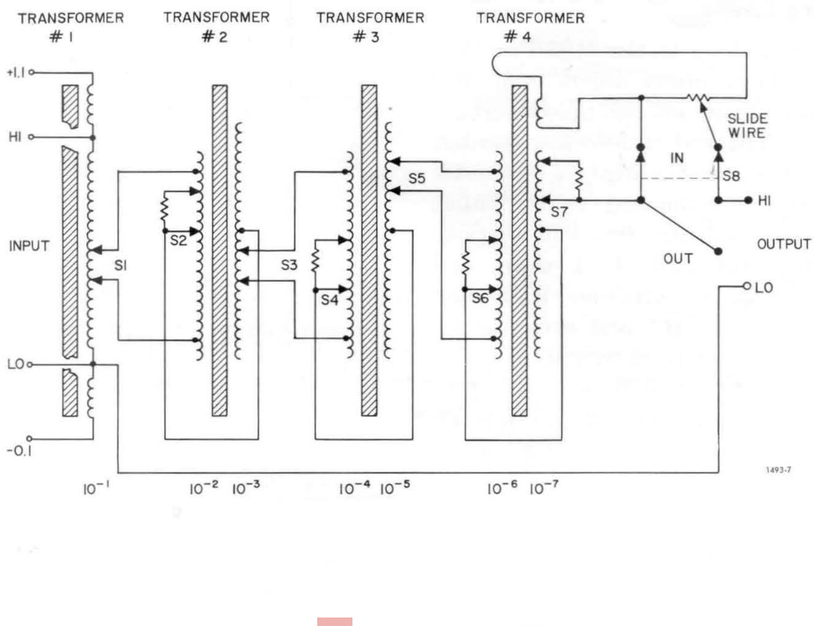
The decade switches are the fingertip-lever type introduced by General Radio on the Type 1615-A Capacitance Bridge. Though they have been modified to meet the requirements of the 1493, they retain the convenient short-throw traverse from -1 to $\times$ (or 10), as well as the easy-to-read, horizontal, in-line, digital display.

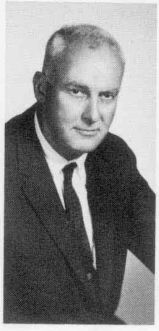
Seven of the levers control step switches; the eighth controls a continuous slide-wire decade that can be switched in to yield essentially infinite resolution.

The Transformers

The 1493 is an assembly of four separate transformers interconnected to produce seven switched decades (see Figure 4). Each transformer winding uses a multifilar cable whose individual conductors are all taken from the same

Figure 4. Schematic diagram showing interconnection of the four transformers.





Gilbert Smiley came to General Radio in 1943 from General Control Company, where as Chief Engineer he was active in industrial automation development. An engineer in GR's Industrial Instruments Group, he has specialized in the design of transformers and other iron-core devices and is largely responsible for the Duratrak brush-track coating used on Variac® autotransformers.

wire spool to ensure equality of resistance. The cable is randomly disposed and lightly twisted and then is wound on an unusually large high-permeability core, with uniform spacing over 360 degrees. Individual conductors are connected end-to-end, aiding, and taps are brought out from certain junctions and ends. After testing, the four transformers are hot-sealed in a magnetically shielded catacomb. The sealant immobilizes the transformers and their connections and keeps out moisture. A second complete performance test follows sealing.

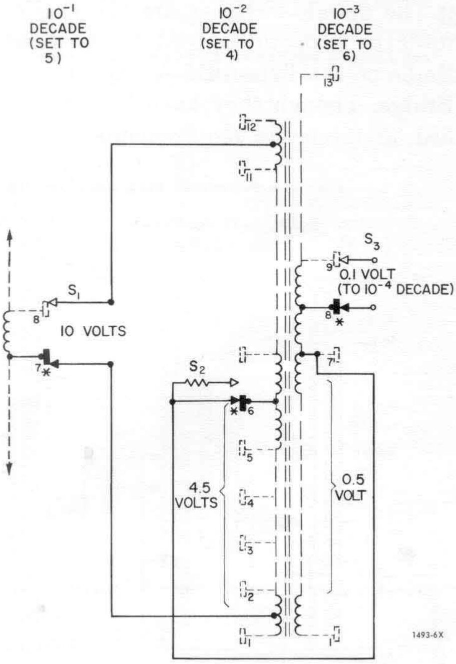
Switching Scheme

The switching of the 1493 Precision Decade Transformer differs from that of other units in several important respects. The first transformer has but one, 12-section winding. Its 10 center sections are connected to the input terminals, and the two end sections provide −0.1 and +1.1 over- and under-voltage connections. With only one winding on the first core, we are able to use the largest wire size capable of meeting the requirements for core excitation. Large wire means low resistance and reduced regulation error from loading.

Ten sections of the primary winding of the second transformer are bridged across successive decade sections of the first transformer by the action of the 10^{-1} decade switch (Figure 5).

Note that the connections to these 10 sections do not coincide with the primary taps but are placed midway between adjacent taps. This displaces the normal voltage per tap by one-half the section voltage. To compensate for this displacement, the secondary winding connection is made to the midpoint rather than to one end. The double offset results in a correct ratio indication from the switch settings. This switching scheme, unique to the 1493, offers two major advantages: (1) Maximum internal impedance (resistance

Figure 5. Diagram showing bridging connections of adjacent transformers.



*Position of ← switch determines setting of significant digit.

and leakage inductance) is substantially reduced, and (2) internal impedance is kept more nearly constant versus ratio, a matter of considerable importance in many measurements.

As one proceeds from input to output, accuracy restraints are relaxed by an order of magnitude per decade. Consequently, transformers 3 and 4 have fewer turns on smaller cores. The reduction in turns lowers the contribution of these two transformers to internal impedance.

One conductor of the multifilar secondary winding on transformer 4 is isolated from the winding proper to form a tertiary winding, which is traversed along the secondary by the 10^{-7} switch. This winding, supplemented by one additional turn around the core, aiding, drives the slide wire. The aiding turn, plus an adjustable resistance in series with the slide wire, allows the slide-wire voltage to be made equal to one step on the 10^{-7} switch. The tertiary winding and aiding turn connect to the slide-wire zero through a knife-edged contact to minimize zero ambiguity. A two-position switch permits operation of the 1493 with or without the slidewire, as desired.

— G. Smiley

SPECIFICATIONS

RANGE: −0.1111111 to +1.11111110 with 7 step decades and continuous slide-wire decade in 10^{-8} position. Each step decade adjustable −1 to X (10). Continuous decade adjustable 0 to X.

ACCURACY

Linearity: Indicated ratio, measured at 100 V, 1000 Hz, with a resolution of $\pm 1 \times 10^{-9}$, agrees with a standard calibrated by the National Bureau of Standards to within their limits of uncertainty, stated as ±2 digits in the 10^{-7} decade. At frequencies from 50 Hz to 2 kHz, ratio accuracy is approx ±1 digit in the 10^{-6} decade. Incremental accuracy of last 4 step decades will be better than ±2 parts in 10^5. Continuous decade accurate to ±1%.

Phase Error (at 1 kHz): < ±6 microradians for ratio settings from 0.1 to 1.0; < ±40 μrad for 0.01 to 0.1; < ±125 μrad for 0.001 to 0.01.

INPUT

Max Voltage: 350 V; below 1 kHz, $0.35 f_{Hz}$ V.

Impedance: > 150 kΩ at 1 kHz; > 20 kΩ from 100 Hz to 10 kHz.

Direct Current: No dc should be applied to input.

OUTPUT

Impedance (dependent on ratio setting): Max: 3.5 Ω, 62 μH; min: 0.5 Ω, 6 μH. With slide-wire decade switched out, max resistance is reduced to 2.7 Ω.

Max Output Current: 1 A.

GENERAL

Terminals: Gold-plated GR 938 Binding Posts.

Accessories Available: Recommended generator and null detector for precise comparison or bridge applications: the 1311-A Audio Oscillator and 1232-A Tuned Amplifier and Null Detector or the combination 1240-A Bridge Oscillator-Detector.

Cabinet: Rack-bench. End frames for bench mount or rack-mounting hardware included.

Dimensions (width x height x depth): Rack, 19 x 7 x 8⅜ in. (485 x 180 x 215 mm); bench, 19 x 7⅜ x 10¾ in. (485 x 190 x 275 mm).

Net Weight: Rack, 28 lb (12.7 kg); bench, 30 lb (13.6 kg).

Shipping Weight: Rack, 41 lb (18.7 kg); bench, 43 lb (19.6 kg).

Catalog Number	*Description*	*Price in USA*
1493-9801	**Type 1493 Precision Decade Transformer,** Bench Model	**$1100.00**
1493-9811	**Type 1493 Precision Decade Transformer,** Rack Model	**1100.00**

As we went to press we learned, with deep regret, of the sudden death of Mr. Gilbert Smiley, the author of the above article.

NEW VOLTAGE DIVIDERS OFFER RESOLUTION TO 10 PPM

Our popular resistive voltage dividers have passed through another stage of evolution. The "new breed" is the 1455 series, of which there are five versions. Beyond an obvious improvement in cosmetics (Figure 1), the new dividers boast tighter specifications and two five-dial units that extend resolution down to 10 parts per million.

Because these dividers are being used increasingly as adjustable elements in measurement and control systems, we have slimmed the package down to 3½ inches and offer both bench and relay-rack models. Also, connections can be made at the rear as well as at the front, and the readout is in-line.

The five versions are the TYPES 1455-A, -AH, and -AL, and the TYPES 1455-B and -BH. The three -A dividers are four-dial units with a ratio range of 0.0001 to 1.0; the two -B's have five dials and a ratio range of 0.00001 to 1.0. An H in the suffix indicates a high impedance rating and consequently greater voltage-handling ability (up to 700 volts); the 1455-AL is a low-impedance divider useful at radio frequencies.

Uses

The decade voltage divider is an established means of obtaining accurately known voltage ratios. Among its many uses are the calibration of voltmeters, linearity measurements on continuously adjustable autotransformers and potentiometers, measurement of gain and attenuation, precise measurement of frequency-response characteristics of audio-frequency networks, and the determination of transformer turns ratios.

The new 1-kΩ 1455-AL will be found useful in testing voltmeters at low radio frequencies. It has a 3-dB response to 7.5 MHz and an output error of less than 1% at 1 MHz. Its low output impedance will make it attractive in many other areas where its low input impedance and voltage rating are not restrictive.

Some Comments on Accuracy Specifications

When these Kelvin-Varley dividers first appeared, we specified accuracy as a percent of reading. Later, the

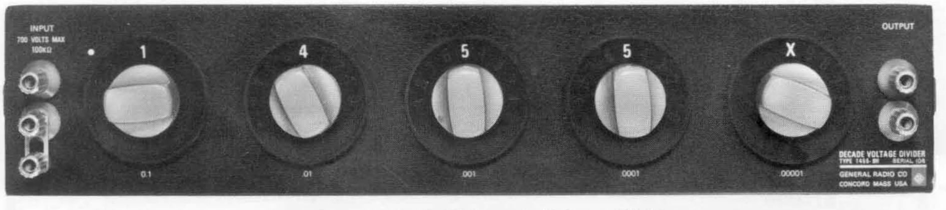


Figure 1. Type 1455-BH Decade Voltage Divider.

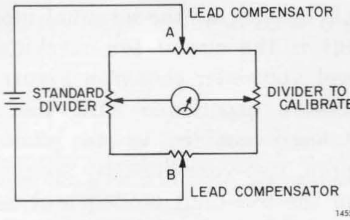


Figure 2. The comparison of two dividers with lead compensators.

use of the term "linearity" became popular for both resistive and ratio-transformer-type voltage dividers. Linearity here is essentially a percent-of-full-scale accuracy specification and equals the percent-of-reading accuracy times the indicated value. Because the indicated value is always less than 1, the linearity specification is a lower number than the percent-of-reading specification and therefore appears at first glance to represent greater accuracy.

We recognized the trend and included an over-all linearity specification, but we kept the percent-of-reading specification because it was the more stringent at low settings. Unfortunately, this created some confusion, for, although we meant both specifications to apply at all settings, some people felt that the percent-of-reading specification was overriding and that there were settings where the linearity specification did not hold. In order to clear up this point, we have gone over to the linearity specification completely and have extended it to all dial settings. This results in somewhat amazing numbers for the linearity of settings for which the first few dials are set to zero.

(Actually, the old accuracy-of-reading specification was more stringent at many settings but didn't sound as good. For those who prefer an accuracy-of-reading specification and understand that it in no way negates the linearity specification, our old statement of $\pm 0.04\%$ *of reading* still applies.)

This linearity specification is somewhat complicated by the fact that the accuracy at very low settings depends somewhat on the method of connection, because of the residual resistance in the internal wiring and switches. Absolute linearity is determined with respect to the voltage output at zero setting; thus by definition there is no error at the zero setting. This type of linearity applies when the zero value of the measured device can be set equal to that of the divider. Consider, as an example, the comparison of two dividers using lead compensators (Figure 2), which are adjusted to bring the zero and unity points of both to coincidence.

When the divider is used as a simple three-terminal device, as shown in Figure 3, the voltage drop in the divider switches and wiring causes a small residual error (see Specifications). Although this error is quite low because of the silver-overlay multiple-contact switches used, at very low settings it can become a limitation on performance. When the input and output circuits do not have a common ground

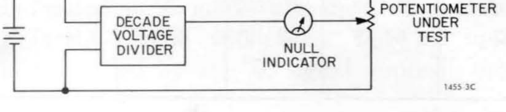


Figure 3. Divider connected to measure potentiometer linearity.

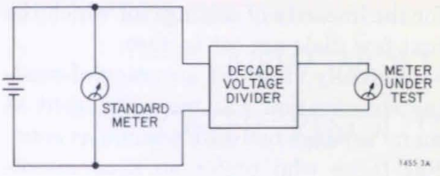


Figure 4. Divider connected to test low-level voltmeter.

and the output is taken between the output terminals, this voltage drop can be largely compensated for and a tighter specification is possible. An example of such a four-terminal measurement is the circuit for checking a low-level voltmeter shown in Figure 4.

Extensive experience with the resistors used enables us to place a ±20-ppm, two-year linearity specification on the five-digit models and ±30 ppm on the higher-impedance four-digit ones. The -AL model is specified at ±50 ppm.

— H. P. Hall

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

SPECIFICATIONS

Type:	**1455-AH**	**—A**	**—AL**	**—BH**	**—B**
Dials:	4	4	4	5	5
Input Resistance:	100 kΩ	10 kΩ	1 kΩ	100 kΩ	10 kΩ
Input Voltage Rating: May be 20 ppm linearity change at full rating (see below)	700 V	230 V	70 V	700 V	230 V
Frequency Response (f_o at 3 dB down): (unloaded, at max output resistance setting)	85 kHz	850 kHz	7.5 MHz	69 kHz	690 kHz
Resolution (in ppm of input):	100	100	100	10	10
Linearity					
Absolute Linearity (in ppm of input): Output taken with respect to output zero setting at low audio frequencies with input voltage < ½ rating.					
Ratio					
0.00001 to 0.00010	—	—	—	± 0.02	± 0.03
0.00010 to 0.00100	± 0.2	± 0.3	± 0.7	± 0.2	± 0.3
0.00100 to 0.01000	± 2	± 2	± 3	± 2	± 2
0.01000 to 0.10000	±15	±15	±20	±10	±10
0.10000 to 1.00000	±30	±30	±50	±20	±20
Terminal Linearity (in ppm of input) (add to absolute linearity):					
Four-terminal (output with respect to low output terminal):	± 0.004	± 0.04	± 0.4	± 0.004	± 0.04
Three-terminal (low terminals common or output with respect to low input terminal):	± 0.02	± 0.2	± 2	± 0.02	± 0.2
Max Output Resistance: (input shorted)	27.9 kΩ	2.79 kΩ	333 Ω	28.8 kΩ	2.88 kΩ
Effective Output Capacitance: (typical, unloaded)	67 pF	67 pF	67 pF	80 pF	80 pF

Frequency Characteristic:
Acts like simple RC circuit below f_o so that

$$\frac{E_o}{E_{in}} \approx \frac{\text{reading}}{\sqrt{1 + \left(\frac{f}{f_o}\right)^2}}.$$

Tabulated value of f_o is at setting that gives max output resistance so that f_o at all other settings is higher. At $0.044 f_o$, response is down $< 0.1\%$.

Accuracy of Input Resistance: $+0.015\%$, except for 1455-AL, which is $+0.025\%$.

Temperature Coefficient: $<$ 20 ppm for each resistor. Since voltage ratios are determined by resistors of similar construction, net ambient temperature effects are very small.

Dimensions (width × height × depth): Rack models, 19 × 3½ × 4⅝ in. (485 × 89 × 120 mm); 4-dial bench models, 14¾ × 3½ × 6 in. (375 × 89 × 155 mm); 5-dial bench models, 17⁵⁄₁₆ × 3½ × 6 in. (455 × 89 × 155 mm).

Net Weight: Bench models, 4-dial, 6¾ lb (3.1 kg); 5-dial, 7¾ lb (3.6 kg).

Shipping Weight (est): Bench models, 4-dial, 7½ lb (3.5 kg); 5-dial, 8½ lb (3.9 kg).

Add approx 1 lb (0.5 kg) to net and shipping weights for rack models.

Catalog Number	*Description*	*Price in USA*
	Type 1455 Decade Voltage Dividers	
	Bench Models	
1455-9700	1455-A, 4-dial, 10-kΩ	$215.00
1455-9702	1455-AH, 4-dial, 100-kΩ	215.00
1455-9704	1455-AL, 4-dial, 1-kΩ	215.00
1455-9706	1455-B, 5-dial, 10-kΩ	255.00
1455-9708	1455-BH, 5-dial, 100-kΩ	255.00
	Rack Models	
1455-9701	1455-A, 4-dial 10-kΩ	222.00
1455-9703	1455-AH, 4-dial, 100-kΩ	222.00
1455-9705	1455-AL, 4-dial, 1-kΩ	222.00
1455-9707	1455-B, 5-dial, 10-kΩ	260.00
1455-9709	1455-BH, 5-dial, 100-kΩ	260.00

PULSES FROM THE TONE-BURST GENERATOR

We are indebted to Mr. R. S. Caddy, Senior Lecturer at the University of New South Wales, for pointing out the useful properties of the General Radio Tone-Burst Generator, TYPE 1396-A, as a pulse generator. If the instrument is connected as for its usual application of producing tone bursts (interrupted or gated sine waves) and if the sinusoidal signal is removed from the gate input and replaced with a dc signal, the output becomes a pulse. The output is limited to about 7 volts positive or negative behind 600 ohms. The prf is controlled by the ac signal applied to the TIMING INPUT terminals. If this timing-signal frequency is swept, the timing circuits of the 1396 will maintain a constant duty ratio by keeping the output-pulse duration a fixed number of periods of the timing signal.

If one alternately produces pulses and tone-bursts by switching the gate input from a dc to a sinusoidal signal source, the output signals from the tone-burst generator will have approximately the same shape frequency spectrum, but it will be centered at dc in the former case and at the sinusoidal frequency in the latter.

the GR Experimenter

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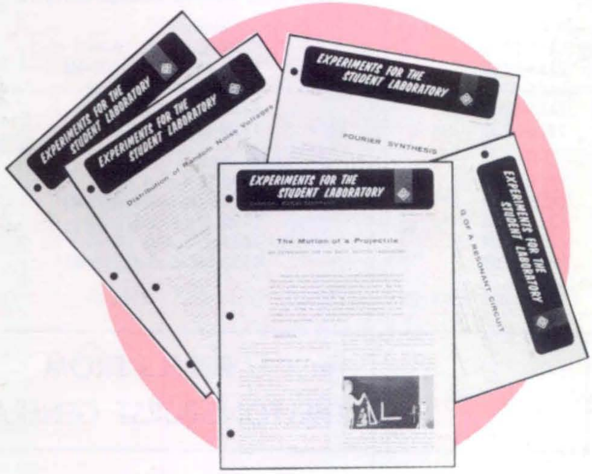
NEW PUBLICATIONS FROM GENERAL RADIO



A new series of applications notes, *Experiments for the Student Laboratory*, has been initiated by GR, and five of these educational aids are already in print. They are "The Motion of a Projectile," "Q of a Resonant Circuit," "Fourier Synthesis," "Distribution of Random Noise Voltages," and "Electronic Voltmeters." Any or all are free on request.

For those concerned with sound and vibration measurement, current and choice reading from GR now includes a new quarterly (*Noise Measurement*), a new series of application notes (*Noise Notes*), and a major revision of the renowned *Handbook of Noise Measurement*. Price of the *Handbook* remains $1.00; the other publications are available for the asking.

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IN THIS ISSUE

THE NEW GENERAL RADIO MICROPHONE

Figure 1. Types 1560-P5 (left) and 1560-P6 Microphones.

If you think that piezoelectric microphones are too temperature-sensitive for precision measurements, the following article has a few surprises for you. New lead zirconate-titanate ceramic microphones have performance characteristics formerly associated only with expensive condenser microphones.

In the continuing search for improved accuracy in sound measurements much effort has gone into improving the characteristics of microphones, because the microphone used is often regarded as the limiting factor in the accuracy of the measurement. Because of the long and continued development effort applied to the condenser microphone, particularly at the Bell Telephone Laboratories,[1] and the extensive work on its calibration in many laboratories all over the world, the condenser microphone very early gained wide acceptance as the best microphone for accurate sound measurements. This acceptance was achieved in spite of the inherent disadvantages of the condenser microphone as compared with the piezoelectric microphone, which is also widely used for sound measurements. These disadvantages are the need for a very high-input-impedance preamplifier supplying an accurately known, high dc polarizing voltage, smaller dynamic range, and greater sensitivity to the effects of humidity. There are, however, several reasons for the wide acceptance of the condenser microphone. Those units that have been carefully designed and built, such as the Western Electric 640AA Condenser Microphone, have a

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[1] E. C. Wente, "A Condenser Transmitter as a Uniformly Sensitive Instrument for the Absolute Measurement of Sound Intensity," *Physical Review*, Vol 10, 1917, pp 39–63.
E. C. Wente, "Sensitivity of the Electrostatic Transmitter for Measuring Sound Intensities, "*Physical Review*, Vol 19, 1922, pp 478–503.
L. O. Sivian, "Absolute Calibration of Condenser Microphones," *Bell System Technical Journal*, Vol 10, No 1, January 1931, pp 96–115.
M. S. Hawley, "The Condenser Microphone as an Acoustic Standard," *Bell Laboratories Record*, Vol 33, No 1, January 1955, pp 6–10.

This month's cover — GR tests every one of its microphones for frequency response, sensitivity, capacitance, dissipation factor, leakage resistance, and linearity to 150 dB.

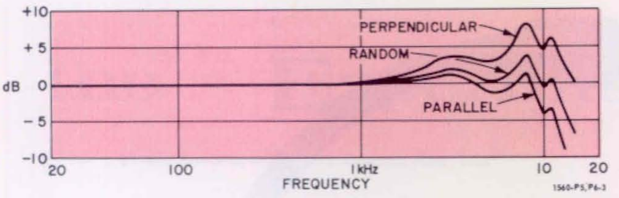

Figure 2. Typical response curves for random, parallel, and perpendicular incidence.

smooth frequency response, and the temperature coefficient is small. If they are carefully handled, their calibration is stable with only moderate aging effects. Furthermore, condenser microphones of this type can be calibrated by fundamental techniques in an acoustic cavity, and they have a high acoustic impedance.

Recent developments in piezoelectric microphones have demonstrated, however, that all these desirable features can now be obtained by the proper design of the piezoelectric microphone and at a lower over-all cost. Because of the additional inherent advantages of the piezoelectric microphone, we believe it will become the preferred microphone to use for acoustic measurements.

Most of the early piezoelectric microphones for sound measurements used a Rochelle-salt crystal as the sensitive element. Although stable and physically rugged, those microphones did not have a high acoustic impedance, and they could not be satisfactorily calibrated in an acoustic cavity. They also had more serious faults, because Rochelle salt is easily damaged by exposure to only moderately high temperatures (56°C), and the microphone capacitance was highly dependent on temperature.

A great advance in piezoelectric microphones was made possible by the development of the lead zirconate-titanate [2] piezoelectric ceramic element. The use of this material eliminated the disadvantages of Rochelle salt, since this new material can be obtained with excellent stability to 100°C, and the capacitance of the ceramic element is relatively independent of temperature.

The TYPES 1560-P3 and 1560-P4 Microphones,[3] supplied on our sound-measuring equipment until recently, use this material, and, as a result, they are stable and rugged, and their characteristics are essentially independent of normal temperature variations. They can also be calibrated by reciprocity techniques in a closed coupler.[4]

When these microphones became available, the most important reasons for using the condenser microphone were eliminated, and the TYPE 1560-P3 Microphone has been widely used for sound measurements. Many measurements, however, have been standardized to use the condenser microphone, which had a diameter of 0.936 inch in contrast with the 1⅛ inches of the TYPE 1560-P3.

Therefore, the next step in the development of the piezoelectric measurement microphone was the adoption of this standard diameter. At the same time it was found possible to obtain a significant improvement in the response characteristic.

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[2] B. Jaffe, R. S. Roth, and S. Maryallo, "Properties of Piezoelectric Ceramics in the Solid-Solution Series Lead Titanate-Lead Zirconate-Lead Oxide: Tin Oxide and Lead Titanate-Lead Hafnate," *Journal of Research of the National Bureau of Standards*, Vol 55, No 5, November 1955, pp 239–254.

[3] E. E. Gross, "TYPE 1551-C Sound-Level Meter," *General Radio Experimenter*, August 1961.

[4] B. A. Bonk, "Absolute Calibration of PZT Microphones," *General Radio Experimenter*, April–May 1963.

The result of this development is the new TYPE 1560-P5 Microphone, shown in Figure 1, manufactured by the General Radio Company, and now used on GR acoustical instruments. It has the same diameter and essentially the same acoustic impedance as the condenser microphone, so that it is suitable for applications where heretofore only condenser microphones have been used. Its frequency response is better than that of any previously available sensitive piezoelectric microphone and is similar to that of the WE 640AA Condenser Microphone up to 15 kHz.

Details of Construction

The microphone is enclosed in an outer brass shell with a brushed chromium plating, which maintains the same high quality of finish after years of use. The aluminum diaphragm directly behind the protective front grid drives a ceramic piezoelectric element, and the electrical output appears at pin terminals at the back of the cartridge. The air leak is also at the back.

The cartridge is available, connected directly to a 3-pin male audio connector, as the TYPE 1560-P5 Microphone or, when connected through a gooseneck to such an audio connector, as the TYPE 1560-P6 Microphone Assembly, also shown in Figure 1. It is now also supplied on the TYPES 1565-A and 1551-C Sound-Level Meters and as part of the TYPE 1560-P40K Preamplifier and Microphone Set.[5]

Frequency Response

The microphone is designed to have a nearly flat response to sounds of random incidence. Figure 2 shows a typical response curve of the microphone for random, parallel, and perpendicular incidence. Most of these microphones follow the random-incidence curve within ±1 dB from 20 to 7000 Hz.

Directivity

The new microphone, since it has the same size as a type L laboratory standard microphone and is similar in construction, maintains the same good omnidirectional characteristics. Up to 1000 Hz the variation in output with angle of sound incidence is small. Above 1000 Hz diffraction causes the microphone to respond more to sounds arriving normal to the diaphragm (0° or

[5] C. A. Woodward, "A New, Low-Noise Preamplifier," *General Radio Experimenter*, June 1965.

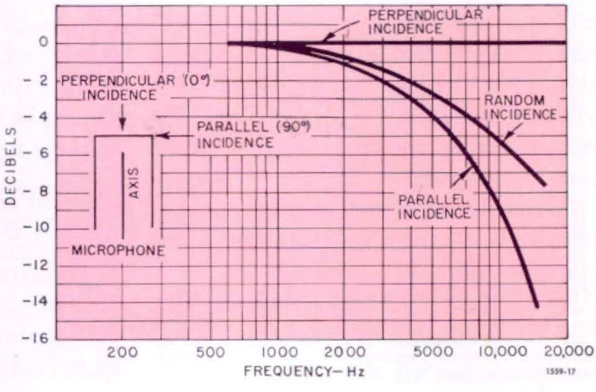


Figure 3. Sensitivity variation with frequency for random, parallel, and perpendicular incidence.

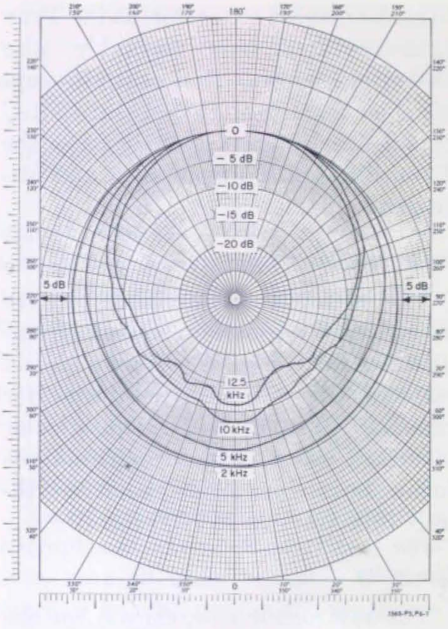


Figure 4. Polar responses at various frequencies.

perpendicular incidence) than to sounds from other directions. Figure 3 shows the extent of the variation in sensitivity as a function of the direction of sound incidence. Figure 4 shows polar responses of the microphone at different frequencies.

Acoustic Impedance and Use in Couplers

The acoustic impedance presented by the microphone in a coupler is equivalent to that of an air volume of 0.45 cm^3 when referred to the front edge of the protecting grid. Of this equivalent volume, about 0.2 cm^3 is due to the compliance of the diaphragm and the ceramic element.

The mechanical resonance of the diaphragm and ceramic element structure is damped acoustically behind the diaphragm. This arrangement corrects one of the deficiencies of the earlier piezoelectric microphones, since the damping element in the new microphone is protected from dirt and mists, whereas, in the earlier ones, the damping element became clogged after long exposure to oil mists and the response of the microphone was affected. Furthermore, the response of the new microphone is not significantly affected by the gas in front of the diaphragm. It can, consequently, be used in coupler calibrations with gases other than air.[6]

As a result of the use of the standard diameter and the elimination of any significant frequency-determining element from in front of the diaphragm, this new microphone can be used in standard reciprocity calibration couplers and earphone calibration couplers designed for the type L laboratory standard microphone.[7]

Although the microphone can withstand sound-pressure levels of 160 dB re 20 $\mu N/m^2$ without damage, pressures much higher than this may ruin it. Such excessive pressures can occur if the microphone is inserted in a coupler without an adequate pressure release. This pressure release is usually a small-diameter hole in the coupler, which can be plugged after the microphone is inserted. But some devices have been built without this release, and these should be modified before use with any of these microphones.

Electrical Impedance—Temperature and Humidity Effects—Cables

The nominal impedance of the Type 1560-P5 Microphone corresponds to a capacitance of 380 pF, in contrast to the 50 to 70 pF that is characteristic of con-

[6] S1.10 — 1966, *USA Standard Method for the Calibration of Microphones*, USA Standards Institute, New York, N. Y.
[7] Z24.8 — 1949, *American Standard Specification for Laboratory Standard Pressure Microphones*, USA Standards Institute, New York, N. Y.

denser microphones. Owing to the lower impedance of the new microphone, its output voltage is less affected by the connection of a cable than is the condenser type. The variation of capacitance with temperature is shown in Figure 5. With a 25-foot cable, this variation produces a change in actual signal voltage of only about 0.025 dB/°C. With a very long cable, the temperature coefficient increases to about 0.04 dB/°C. When the microphone is attached directly to the TYPE 1560-P40 Preamplifier, this effect is eliminated. The remaining temperature coefficient is that of open-circuit output voltage, which is less than ±0.01 dB/°C.

The sensitivity of a condenser microphone is directly proportional to the applied dc polarizing potential, and the stability of the system can therefore be no better than that of the polarizing supply voltage. Accurate monitoring of this voltage is essential for assurance that the sensitivity of the microphone system does not change. Such precautions are not necessary with the piezoelectric microphone, because no polarizing voltage is required.

The high polarizing voltage needed with the condenser microphone also makes it particularly sensitive to the effects of humidity. If moisture provides a conductive path for the polarizing voltage at the microphone, the resulting leakage current introduces excessive noise into the signal path, and the system can easily become inoperative. Because of the heat produced by the vacuum-tube amplifiers commonly used, the humidity at the microphone terminals is kept below the ambient humidity, and the above-mentioned effects of humidity have not been as widely observed as they will be when solid-state preamplifiers come into general use.

The effects of humidity on the piezoelectric microphone are much less serious, because of the absence of a polarizing voltage and because its impedance is only about one-tenth that of a condenser microphone. Its impedance is still relatively high at low frequencies, however, and prolonged exposure to extremely high humidity should be avoided. If it is exposed to 100% humidity for a matter of hours, some loss in sensitivity below 100 Hz may occur, and some increase in low-frequency background noise will accompany this loss.

When the piezoelectric microphone is used at the end of a long cable with no preamplifier, the upper sound pressure at which it can be used is not affected. The distortion from a condenser microphone, however, is affected by the impedance that terminates it. If a cable is connected between the condenser microphone and a high-impedance preamplifier, the distortion is greater than when the condenser microphone works

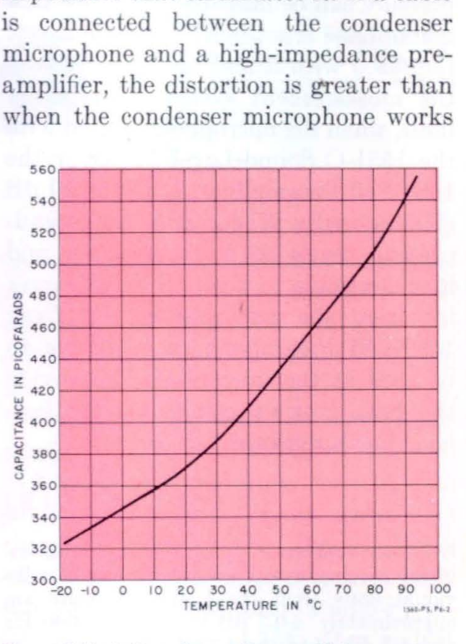


Figure 5. Variation of capacitance with temperature.

Basil A. Bonk received his BSEE and MSEE degrees from MIT in 1960, then joined General Radio as a development engineer in the Audio Group. There he has specialized in the development of microphone calibration systems and in the design of measurement microphones.

directly into the preamplifier. Because of this effect and the very serious loss in signal level that results when a condenser microphone is used with a cable, the microphone is almost always used directly on a preamplifier. The TYPE 1560-P5 Microphone, however, can be used over a wide range of sound-pressure levels without a preamplifier.

Dynamic Range

The lower limit at which sound-pressure can be measured with this new microphone is set by the electronic system used with it and the bandwidth of the measurement system. This lower limit, when the microphone is used with the 1551-C Sound-Level Meter or the 1560-P40 Preamplifier, is about 20 dB (C-weighted). When very low sound-pressure levels are to be measured and it is advisable to position the observer far from the microphone, the TYPE 1560-P40 Preamplifier is available to be used at the microphone to provide the best possible conditions for low-level measurements.

The upper limit of linear operation of the microphone is set by distortion. The distortion in the electrical output is less than 1% when the microphone is exposed to a sound-pressure level of 150 dB re 20 $\mu N/m^2$. The total dynamic range is then about 130 dB for C-weighting, which is significantly larger than the dynamic range of condenser microphones.

Conclusion

The characteristics of this new microphone are so good that the limitations on the accuracy of a practical sound measurement will almost always come from factors other than the behavior of the microphone. Some of these other factors that limit accuracy are the following: the effects of the room; of interfering objects, particularly the observer; of stray pickup; of ambient noise; of the placement and mounting of the noise source; of the microphone positions used; and of the particular space and time averaging techniques used.

The TYPE 1560-P5 Microphone can be used in all the ways that the highly respected condenser microphone has been used in the laboratory, and, in addition, it is well suited for use in portable field-type sound-measuring systems.

— B. A. BONK

Editor's Note

The design of this microphone was started by B. B. Bauer and A. L. DiMattia of CBS Laboratories and was completed by the author at the General Radio Company.

SPECIFICATIONS

Frequency Response: Typical response is shown in the accompanying plot. Deviations of individual units from the typical response are approximately ±0.3 dB from 20 to 1000 Hz and ±1 dB up to about 7000 Hz.

Sensitivity: −60 dB re 1 V/μbar nominal.

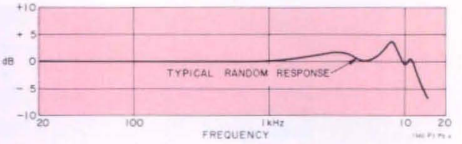
Temperature Coefficient of Sensitivity: Approximately −0.01 dB/°C.

Internal Impedance: Capacitive; TYPE 1560-P5, 390 pF at 25°C, nominal; TYPE 1560-P6, 425 pF at 25°C, nominal. Temperature coefficient of capacitance: 2.2 pF/°C over range of 0 to 50°C.

Environmental Effects: Microphone is not damaged by temperatures from −40 to +60°C and relative humidities of 0 to 100%.

Terminals: Microphones fit 3-terminal microphone cable connector. For hum reduction both microphone terminals may be floated with respect to ground.

Cartridge Dimensions: Diameter 0.936 ± 0.002 in. (23.7 mm), length 1⅛ in. (29 mm).



Net Weight: Type 1560-P5, 2 oz (60 g); Type 1560-P6, 8 oz (0.3 kg).

Shipping Weight: Type 1560-P5, 1 lb (0.5 kg); Type 1560-P6, 3 lb. (1.4 kg).

Catalog Number	*Description*	*Price in USA*
1560-9605	**1560-P5 Microphone**	**$60.00**
1560-9606	**1560-P6 Microphone Assembly**	**85.00**

NEW FIVE-FREQUENCY SOUND-LEVEL CALIBRATOR

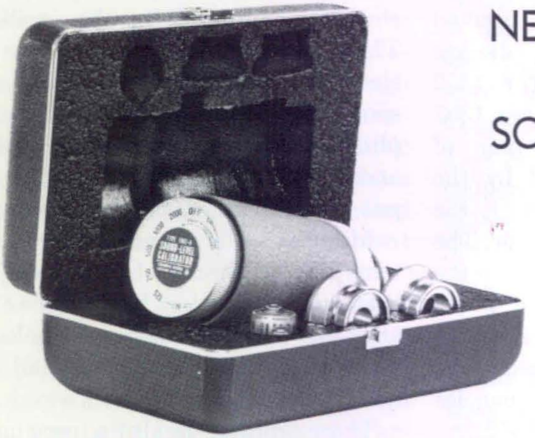


Figure 1. Type 1562-A Sound-Level Calibrator in storage case, with snap-in adaptors.

One of our lighter-spirited publications suggests that, for a day-to-day check on sound-level-meter calibration, one may hold the instrument at arm's length and say, in an even voice, "I feel rather foolish talking to a sound-level meter," repeating this announcement daily and noting any variation in indicated level. Those who prefer a more reliable and less attention-getting approach will be interested in GR's new Type 1562-A Sound-Level Calibrator, a small, transistorized oscillator-speaker-coupler unit designed for the calibration of most commonly used sound-measuring microphones and systems.

Much is to be gained from accurate calibration of an acoustical measurement system. The better the calibration accuracy, the closer one can approach allowed performance specifications, the more consistent his comparison measurements will be, and the more confidence he can have in his measurements.

Acoustical measuring instruments can be calibrated in many ways. The simplest procedure is the amplifier self-check, provision for which is built into many General Radio sound-measuring instruments. At the other end of

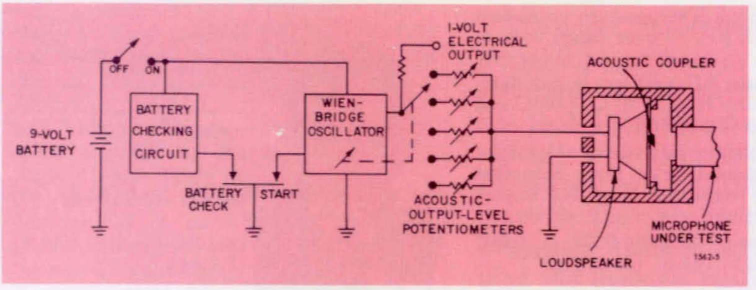


Figure 2. Diagram showing principal elements of the calibrator.

the scale is the precise calibration, from 20 to 8000 Hz, of a microphone or system by means of the Type 1559-B Reciprocity Calibrator.[1]

Probably the most commonly performed calibration has been a simple over-all system check at a specified frequency, for which General Radio has listed the combination of Type 1552 Sound-Level Calibrator and Type 1307 Transistor Oscillator.[2] This pair of instruments is now succeeded by the small cylinder shown in Figure 1, the Type 1562 Sound-Level Calibrator. The new calibrator offers, in addition to the obvious convenience of the new packaging, specifications far superior to those of its predecessor. (For example, the 1562 has five calibration frequencies vs the single frequency of the 1552.)

The new sound-level calibrator contains a solid-state oscillator, an electroacoustic transducer, and an acoustic coupler, all enclosed in a cylindrical housing only 2¼ inches in diameter and 5 inches long. In normal operation, it is placed over the microphone of the system to be calibrated, and the frequency selector switch is set to one or more of the five available test frequencies (125, 250, 500, 1000, and 2000 Hz). The calibrator is factory-adjusted to develop a sound-pressure level of 114 dB re 20 $\mu N/m^2$ (0.0002 μbar).

An important additional feature is the availability, at a phone jack, of the electrical output from the oscillator. Thus, for example, one can use the electrical output to measure the response of a system without its microphone, then connect the microphone and apply the calibrator's acoustic output to verify the microphone response.

Many types of sound-measuring systems can be checked by the new calibrator. With five test frequencies available, one can check frequency-selective instruments such as octave-band analyzers.

The calibrator is also a very useful accessory for the Type 1525 Data Recorder.[3] Not only can the acoustic output be used for system calibration, but the electrical output can be used to adjust recorder bias voltage and to produce electrical test signals on tape.

DETAILED DESCRIPTION

The principal elements of the calibrator are shown in Figure 2. The oscillator drives a loudspeaker, which generates high-level acoustic signals in a coupler that fits over the microphone to be calibrated. The electrical output of the oscillator is available at a phone

[1] "A Reciprocity Calibration for the WE640AA and Other Microphones," *General Radio Experimenter*, December 1964.
[2] E. E. Gross, "An Improved Sound-Level Calibrator," *General Radio Experimenter*, June 1955.
[3] Arnold Peterson, "Magnetic Tape Recorder for Acoustical, Vibration, and Other Audio-Frequency Measurements," *General Radio Experimenter*, October 1966.

jack on the side of the calibrator housing. This jack is built into a tubular nut, which secures the outer shell of the instrument and which also keeps the calibrator from rolling off tables.

At the top of the calibrator is a rotary switch and dial combination with seven positions: the five operating frequencies, a battery-check position, and a power-off position.

The Oscillator and Amplifier

The oscillator is a Wien-bridge circuit first described by Fulks.[4] The key to its stable operation is a thermistor in the negative feedback path, which automatically adjusts its resistance to the value needed to maintain oscillation. Its time constant is short enough to correct rapidly for amplitude variations, yet long enough to cause little distortion at low frequencies. It operates at a high temperature, in an evacuated bulb, to minimize the effects of ambient temperature.

The amplifier uses four transistors in a single, direct-coupled feedback loop. Enough negative feedback is used to achieve a transfer characteristic that is substantially independent of transistor characteristics.

The Output System

The loudspeaker is a controlled-reluctance magnetic transducer with a very low temperature coefficient and long-term stability proven by years of successful operation in the TYPE 1552 Calibrator. Similarity with the older calibrator ends, however, with the output coupler, which is designed to accommodate the 1⅛-inch-diameter piezoelectric ceramic microphone now in use on thousands of sound-level meters. Two snap-in adaptors are provided, one for the new 15/16-inch-diameter sound-level meter microphone (see page 3, this issue) and type L laboratory standard microphones such as the WE 640AA, and the other for the ⅝-inch microphone used with the TYPE 1551-P1 Condenser Microphone System. Figure 3 is a cross-section drawing of the coupler in place on a 1⅛-inch microphone, and Figure 4 shows the coupler plus snap-in adaptor in place on a 15/16-inch microphone.

PRINCIPLES OF CALIBRATION

The 1562 Calibrator develops a constant sound-pressure level of 114 dB

[4] Fulks, R. G. "Novel Feedback Loop Stabilizes Audio Oscillator," *Electronics,* Vol 36 No 5, Feb. 1963. Available from General Radio as Reprint A-107

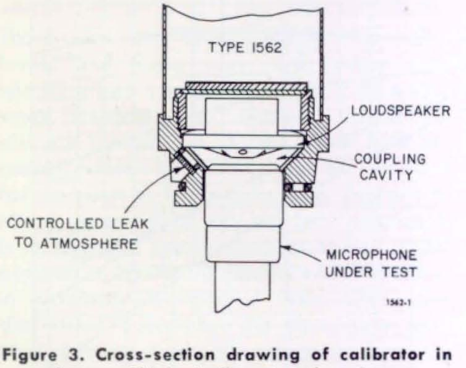


Figure 3. Cross-section drawing of calibrator in place on 1⅛-inch-diameter microphone.

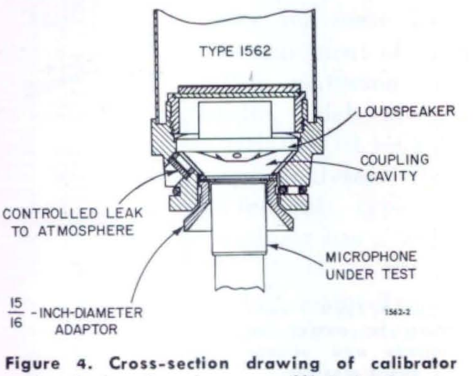


Figure 4. Cross-section drawing of calibrator and adaptor in place on 15/16-inch-diameter microphone.

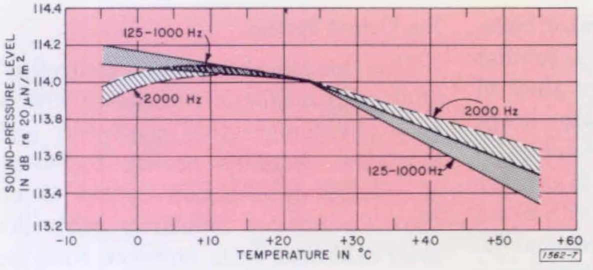


Figure 5. Typical and specified variation in output vs ambient temperature.

re 20 $\mu N/m^2$ at each of five frequencies (125, 250, 500, 1000, and 2000 Hz) when its acoustic coupler is placed over a high-acoustic-impedance sound-measuring microphone. This level is established at General Radio in terms of a carefully maintained laboratory standard microphone (WE 640AA) with a pressure calibration determined by reciprocity and traceable to the National Bureau of Standards.

The calibrator's constant output vs frequency is in contrast with the characteristic of a sound-level meter, which is designed with weighted frequency response in accordance with international standards. Furthermore, the microphones used on most sound-level meters are adjusted for a flat response to sounds of *random incidence* in a free field. Therefore, to determine exactly what a sound-level meter should indicate when the calibrator is coupled to its microphone, one must correct for the random-incidence characteristic of the microphone and for the weighted response of the sound-level-meter amplifier. These corrections are small for the new $^{15}/_{16}$-inch microphones, but they should be taken into account where extreme accuracy is needed. Detailed correction tables are included in the instruction manual for the calibrator.

ENVIRONMENTAL EFFECTS

The calibrator output level is established at a temperature of 23°C and an atmospheric pressure of 760 mm of mercury. As long as the battery voltage is at least 6 volts (a 9-V battery is used),

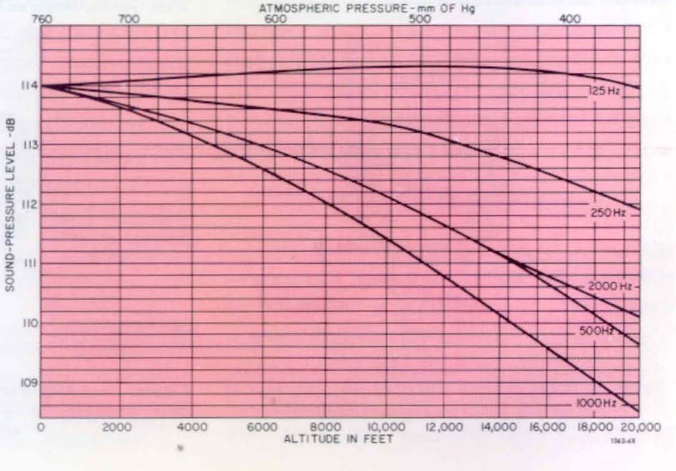


Figure 6. Typical variation in output vs altitude and atmospheric pressure.

normal variations in temperature and barometric pressure will have negligible effect on the sound-pressure level developed. Figure 5 shows the variation in output that may be expected as the ambient temperature departs from 23°C.

Large changes in barometric pressure due to altitude changes as one moves about the country do produce appreciable variations in output, but even these are generally smaller than the changes that occur in a closed coupler with high-acoustic-impedance transducers. For exampie, the sound-pressure reduction in a closed coupler at 15,000 feet altitude is about 6 dB. The corresponding loss for the 1562 is only about 3 dB (Figure 6).

— E. E. Gross

A brief biography of Mr. Gross appeared in the October 1966 *Experimenter.*

SPECIFICATIONS

ACOUSTIC OUTPUT

Frequencies: 125, 250, 500, 1000, and 2000 Hz, ±3%.

Sound-Pressure Level: 114 dB re 20 μN/m².

Accuracy (at 23°C and 760 mm Hg):

	at 500 Hz	*other frequencies*
WE 640AA or equivalent	±0.3 dB	±0.5 dB
Other microphones	±0.5 dB	±0.7 dB

Temperature Coefficient: At 125, 250, 500, and 1000 Hz: −0.01 to −0.025 dB/°C from 23°C to 50°C, 0 to −0.01 dB/°C from 0 to 23°C. At 2000 Hz: −0.01 to −0.015 dB/°C from 23°C to 50°C, 0 to −0.01 dB/°C from 0 to 23°C.

Pressure Correction: Chart supplied.

ELECTRICAL OUTPUT

Voltage: 1.0 V ±20% behind 6000 Ω.

Frequency Characteristic: Output is flat ±2%.

Distortion: <0.5%.

Connector: Jack to accept standard phone plug.

GENERAL

Operating Environment: 0 to 50°C, 0 to 100% relative humidity.

Accessories Supplied: Carrying case, adaptors for 15/16-in.- and 5/8-in.-diameter microphones (fits 1 1/8-in. microphones without adaptor). Battery included.

Battery: One 9-V Burgess PM6 or equivalent. 120 hours use.

Dimensions: Length 5 in. (130 mm); diameter 2 1/4 in. (55 mm).

Weight: Net, 1 lb (0.5 kg); shipping 4 lb (1.9 kg).

Catalog Number	*Description*	*Price in USA*
1562-9701	**1562-A Sound-Level Calibrator**	**$195.00**

A DIAL DRIVE FOR STEPPED OR SWEPT ANALYSIS

Those who must specify acceptable noise levels and those who must follow such specifications are understandably happiest when dealing with discrete values at discrete frequencies. At the same time, of course, enough information must be included for the noise to be meaningfully described. Many people think that the best approach is a stepped third-octave analysis, which presents a significant amount of information in an easily interpreted form. GR's new dial drive automates the procedure and also permits no-hands analysis by the traditional swept technique.

Modern test codes for noise frequently involve a measurement of the noise spectrum, either continuous or at specified frequencies, which can be conveniently made with the GR 1564-A Sound and Vibration Analyzer.[1] Designed specifically for this type of measurement, the analyzer has a wide frequency range (2.5 Hz to 25 kHz), can be operated either manually or automatically, and has both one-tenth-

[1] W. R. Kundert, "New Performance, New Convenience With the New Sound and Vibration Analyzer," *General Radio Experimenter*, Sept–Oct 1963.

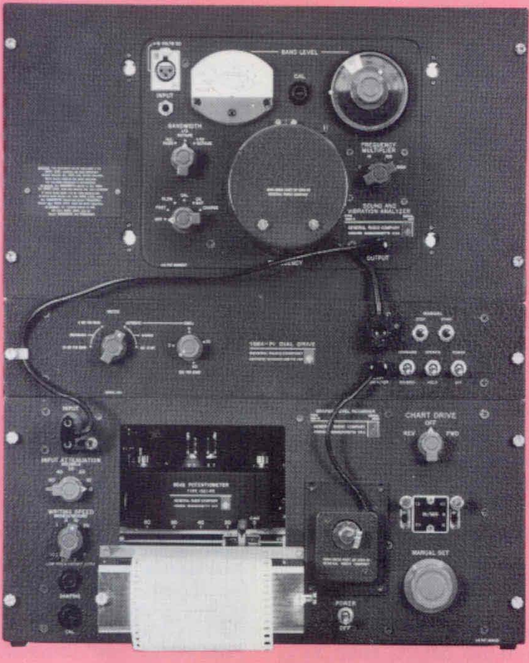


Figure 1. Type 1564-P1 Dial Drive mounted on 1564-A Sound and Vibration Analyzer and synchronized with 1521-B Graphic Level Recorder.

octave and one-third-octave bands. For automatic operation, it is used with the Type 1521-B Graphic Level Recorder, and the combination of analyzer and recorder is available, completely assembled, as the Type 1911-A Recording Sound and Vibration Analyzer.

Recently a growing interest in stepped $\frac{1}{3}$-octave analysis has placed new demands on many acoustical laboratories. To adapt the 1564 for automatic stepped as well as continuous analysis, we now offer the Type 1564-P1 Dial Drive.

The dial drive is essentially a device for automatically moving the frequency dial of the Type 1564 Sound and Vibration Analyzer from one third-octave center frequency to the next, with adjustable dwell time at each step. As a matter of convenience, the dial drive also provides for continuous rotation of the analyzer frequency control. Thus the combination of 1564-A Analyzer and 1564-P1 Dial Drive presents three possibilities: stepped $\frac{1}{3}$-octave, continuous $\frac{1}{3}$-octave, and continuous $\frac{1}{10}$-octave analysis. The complete analysis setup will often also include a GR 1521-B Graphic Level Recorder, and here the dial drive also contributes to convenience by replacing the usual chain linkage between recorder and analyzer with electrical synchronization.

CONTINUOUS VS STEPPED ANALYSIS

A continuous plot of the spectrum is very helpful in the evaluation of a noise if the noise limits are specified as continuous functions of frequency, which can be entered on the spectrum chart to check compliance. Some test codes (e.g., MIL-STD 740 and ASHRAE 36A-63) have tried to simplify the analysis procedure by specifying acceptance levels in discrete third-octave bands. The selected center frequencies of these bands are those of the preferred frequency series (S1.6-1960, American Standard Preferred Frequencies for Acoustical Measurements, and ISO-R266-1962). These frequencies include 1000 Hz and the frequencies spaced above and below 1000 Hz in third-octave steps.

With stepped analysis, the presentation for each third octave is a single reading that can be quickly compared with the requirements of a code or with a similar reading taken at another time. This simplification is achieved at the expense of some information; where one is *not* bound by specification he can choose either the single-valued approach of stepped analysis or the detail that only a continuous analysis can provide.

Figures 2, 3, and 4 describe, better than can words, the difference between

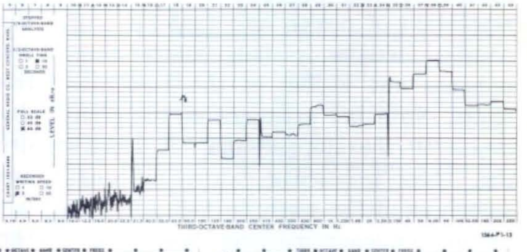


Figure 2. A stepped ⅓-octave analysis of vibration of defective motor. Dwell time was set at 10 seconds, full-scale level was 80 dB, and recorder writing speed was 3 in./s. Chart paper is GR 1521-9460, specially designed for stepped ⅓-octave analysis.

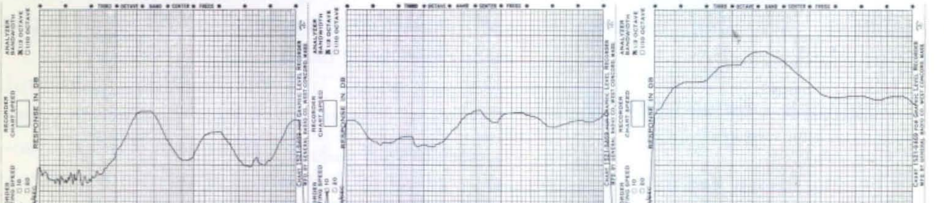


Figure 3. Continuous ⅓-octave analysis of the same vibration analyzed in Figure 2. Detail is greater, but ⅓-octave levels are harder to read.

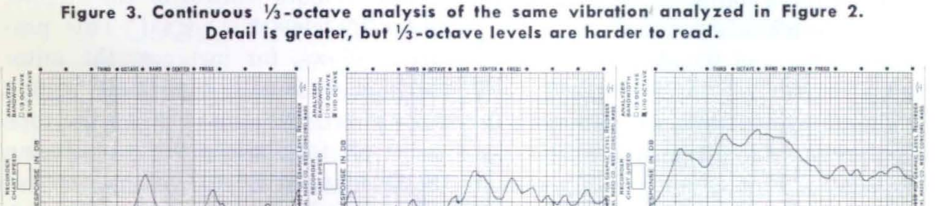


Figure 4. Continuous $\frac{1}{10}$-octave analysis of the same vibration analyzed in Figures 2 and 3. Note presence of components not visible on broader-band analyses.

stepped and continuous analysis. These are chart recordings of the vibration of a motor with a damaged sleeve bearing. Figure 2 shows a stepped ⅓-octave analysis; Figures 3 and 4 are continuous ⅓- and 1/10-octave analyses, respectively. All were made with the combination of 1564 Sound and Vibration Analyzer, 1564-P1 Dial Drive, and 1521-B Graphic Level Recorder. In all cases, the recorder was set for a writing speed of 3 inches per second and was equipped with an 80-dB potentiometer.

It is immediately apparent that the stepped-analysis recording (Figure 2) is the easiest to interpret, that the continuous recordings give far better resolution of peaks, and that the 1/10-octave analysis reveals many components not visible in the ⅓-octave chart. Note the existence of the transients on the stepped-analysis recording. These transients, caused by range switching of the analyzer, can serve as useful frequency markers, because the ⅓-octave switching frequencies are accurately known.

DETAILED DESCRIPTION OF THE DIAL DRIVE

The dial drive consists of a stepper motor, which mounts on and drives the analyzer; a contactor, which mounts on and provides synchronization with the recorder; and the electronic control unit.

The stepper motor is driven by pulses from a ring-of-four counting decade in the control unit. Four pulses applied in the proper sequence to the four drive coils of the motor advance the motor 7½ degrees per pulse. The 10-to-1 gear ratio between motor and analyzer frequency control divides these steps into ¾-degree increments.

The ring-of-four counting decade provides the necessary logic to sequence pulses for either forward or reverse operation and also to ensure that the motor never stops with one of its coils energized. The rate at which the motor, and thus the analyzer frequency dial, rotates is a function of the pulse train that is applied to the ring-of-four counting decade. For stepped operation, these pulses come from a free-running multivibrator, at a rate that moves the analyzer dial 30 degrees, or from one ⅓-octave-band center to the next, in about 0.35 second. For continuous operation, they are derived from the power-line frequency through dividers. (Since the 1521-B Recorder is driven by a synchronous motor, the analyzer is automatically synchronized with the recorder in continuous operation.) Two speeds are available in the continuous mode: 6 or 20 seconds per ⅓-octave band.

The control unit includes start and stop buttons, forward and reverse switches, a continuous-automatic selector switch, and a dwell-time control. A connector at the rear of the unit permits remote control of the stepper motor by electrical signal. This provision allows, for instance, the automatic analysis of a tape loop, with a piece of reflective material on the tape and a photoelectric pickoff[2] producing the signals to advance the stepper motor.

Synchronization between the analyzer and the recorder chart paper is accomplished by means of a contactor assembly, which mounts on the recorder. This is a cam-actuated switch,

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[2] G. Partridge, "A Simple Way to Synchronize Magnetic Tape With Oscilloscope Trace," *General Radio Experimenter*, October 1966.

which causes the dial drive to move the analyzer to the next ⅓-octave band at each ¼-inch travel of the chart paper. Chart paper 1521-9460 is specially calibrated for this application.

SUMMARY

The Type 1564-P1 Dial Drive greatly simplifies stepped ⅓-octave analysis and permits changeover to continuous ⅓- or 1/10-octave continuous analysis at the turn of a switch. Those who are called upon to make such analyses should find appreciable savings in time and convenience through the use of this accessory.

— B. A. Bonk

SPECIFICATIONS

STEPPING CHARACTERISTICS

Stepping Motion: 0.75°/step; 40 steps (30°) per ⅓ octave; controlled to step in sequence of 4 pulses = 3°.

Stepping Time: In stepped mode, approx 0.35s/30°; in continuous mode, 6s/30° or 20s/30°, both synchronized to 60-Hz line.

Dwell Time (⅓-octave band): Dwell time + stepping time is 1, 3, 10, or 30 s, when controlled by 1521-B Graphic Level Recorder with 60-rpm motor. These times can be increased 2 × or 4 × with cam adjustment. Dwell time can also be set by front-panel control from approx 1 to 60 s.

GENERAL

Temperature Range: Operating, 0 to 50°C; storage, −40 to +70°C.

Humidity Range: 0 to 95% RH.

Synchronization: To 1521 Graphic Level Recorder in both stepped and continuous modes.

Recording System: Output from 1564 analyzer can be connected to any recorder with input impedance of 10 kΩ or more and sensitivity of at least 10 mV (1521-B Recorder recommended).

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

Accessories Supplied: Adaptor-cable assembly, power cord, spare fuses, end frame set (bench model) or rack-support set (rack model).

Accessories Available: Chart paper for 1521 Recorders: 1521-9460 for stepped analysis, 1521-9469 for continuous analysis.

Dimensions (w × h × d): Relay-rack section, 19 × 3½ × 12½ in. (485 × 89 × 320 mm); stepper motor, 4¼ (dia) × 5⅛ in. (110 × 135 mm); contactor assembly, 3 × 4 1/16 × 2⅛ in. (77 × 105 × 54 mm).

Weight: Net 16½ lb (7.5 kg); shipping, 36 lb (16.5 kg).

Catalog Number	*Description*	*Price in USA*
1564-9771	**1564-P1 Dial Drive,** Bench Model	$720.00
1564-9772	**1564-P1 Dial Drive,** Rack Model	720.00
1521-9460	**Chart Paper (stepped mode)**	2.75
1521-9469	**Chart Paper (continuous mode)**	2.75

OSCILLATOR FOR AUDIOMETER CALIBRATION

With the introduction of a standard earphone coupler[1], General Radio greatly simplified the use of its sound-level meters and analyzers in the calibration of audiometric equipment. To round out the calibration system, we are now offering a low-distortion audio oscillator with switch selection of 12 frequencies commonly used in audiometry. Among these frequencies is the octave series based on 125 Hz, which is incorporated in specifications of the USA Standards Institute.[2]

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

[2] *American Standard Specification for General Diagnostic Purposes, Z24.5 — 1951*, USA Standards Institute, 70 E. 45th St., New York, N. Y.

Type 1311-A

The 1311-AU is an all-solid-state, Wien-bridge oscillator, which uses extensive negative feedback to attain very low distortion (typically under 0.1%) and high degrees of amplitude and frequency stability. Except for its output frequencies, it is identical to the TYPE 1311-A Audio Oscillator, described in the August-September 1962 isssue of the *Experimenter*.

SPECIFICATIONS

FREQUENCY

Range: 12 fixed frequencies, 125, 250, 400, 500, 750, 1000, 2000, 3000, 4000, 6000, and 8000 Hz. ΔF control provides ±2% adjustment.

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

Frequency Stability: 0.1% typical, long-term, after warmup.

Synchronization: Telephone jack provided for external synchronizing signal. Locking range is about ±3% for 1-V rms reference signal. ΔF control can be used for phase adjustment.

OUTPUT

Power: 1 W into matched load (taps provide at least 0.5 W into any resistive load between 80 mΩ and 8 kΩ).

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

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

Impedance: Between one and two times matched load, depending on control setting. Output circuit is isolated from ground.

Amplitude Stability: Better than 1% long term, 0.01% short term, typical after warmup.

Synchronization: High-impedance, constant-amplitude, 1-V rms output for use with oscilloscope, counter, or other oscillator.

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

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

GENERAL

Terminals: TYPE 938 Binding Posts. Separate ground terminal holds shorting link, which can be used to ground adjacent OUTPUT binding post.

Power Required: 105 to 125 or 210 to 250 V, 50 to 400 Hz. 22 W.

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

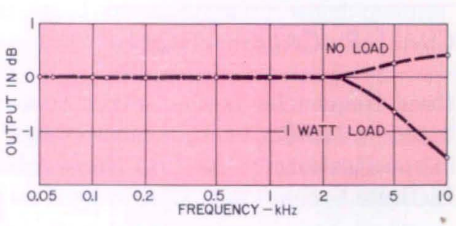
Accessories Available: Rack-mounting set (panel 5¼ in. high).

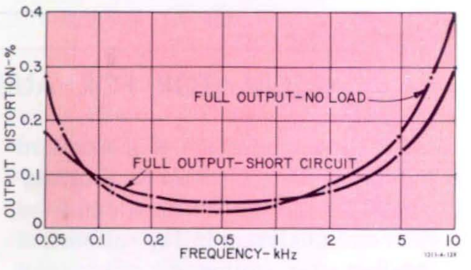
Mechanical Data: Convertible-bench cabinet.

Dimensions (width × height × depth): 8 × 6 × 7¾ in. (205 × 155 × 200 mm).

Weight: Net, 6 lb (2.8 kg); shipping, 9 lb (4.1 kg).

Output characteristics.





Catalog Number	*Description*	*Price in USA*
1311-9703	**1311-AU Audiometric Oscillator**, 115 volts	**$230.00**
1311-9704	**1311-AU Audiometric Oscillator**, 230 volts	**230.00**
0480-9638	**480-P308 Relay-Rack Adaptor Set**	**7.00**

GR Product Notes

DECADE RESISTORS

Two six-dial decade resistors have been added to the 1434 series. The 1434-B (Catalog No. 1434-9702) has a total resistance of 1,111,110 ohms and a minimum per-step resistance of 1.0 ohm. The 1434-X (Catalog No. 1434-9724) has a total resistance of 111,111 ohms, a minimum per-step resistance of 0.1 ohm. Prices in USA are $135 and $116, respectively.

SOUND—VIBRATION

A 100-foot extension cable (1560-P72B) is now available for use between the 1560-P40 Preamplifier and a microphone or analyzer. Catalog No. is 1560-9977. Price in USA, $29.00.

STROBOSCOPES

The inexpensive 1539-A Stroboslave, when coupled with the 1531-P2 Flash Delay and the 1536-A Photoelectric Pickoff, is enough stroboscope for most people who want only to look at or to photograph objects moving at high speed and who do not have to measure speed. For easy ordering, we now offer the combination of all three items as the 1539-Z Motion Analysis and Photography Set. Catalog No. is 1539-9900 for 115-V systems, 1539-9901 for 230 V. Price in USA is $435.00.

COAXIAL

An adaptor is now available between our two coaxial connector series, GR900 and GR874. The 874-Q900L Adaptor contains a 900-AB connector and a locking GR874. VSWR is typically under 1.04 to 9 GHz. Catalog Number is 0874-9709. Price in USA is $15.

Erratum

In Figure 3, page 9, of the April *Experimenter*, the case of the 1493 decade transformer should not have been shown connected to the −0.1 tap. In this setup the −0.1 tap is not used.

the GR Experimenter

GENERAL RADIO COMPANY
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WHEELS OF PROGRESS

General Radio's fleet of traveling exhibits has a new flagship, a spacious, air-conditioned Cortez outfitted with operating displays of GR instruments. The new van will roam the western and southwestern regions of the U.S.

Other GR "road shows" travel by specially equipped station wagons and are set up on invitation in or near industrial plants and laboratories.

Our line of acoustical instruments has its own vehicle, called GRAIL (General Radio Acoustical Instrument Laboratory). In Europe, two Mercedes vans cruise the autobahns for General Radio (Overseas).

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THE GENERAL RADIO

Experimenter

ALSO IN THIS ISSUE

- New 30-MHz I-F Amplifier
- fastrak Markers for the Graphic Level Recorder

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THIS ISSUE

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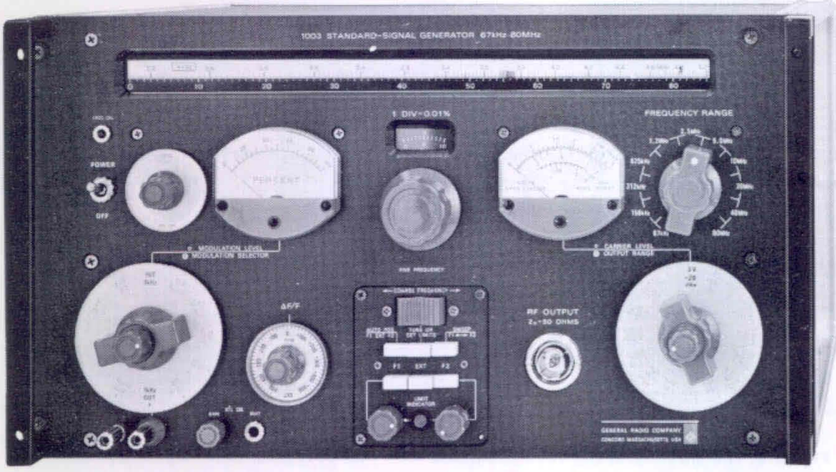


Figure 1. Type 1003 Standard-Signal Generator.

THE 1003 STANDARD-SIGNAL GENERATOR

It is infrequent that one sees major innovation in an art as mature as signal-generator design. Thus the subject of this month's feature is particularly noteworthy, for the 1003 is based on a truly innovative idea for achieving dramatic improvements in frequency stability, resolution, and accuracy. Freshness of approach marked the entire development, and the result is an interesting new chapter in the history of one of the most important of all electronic instruments.

A new generation of GR standard-signal generators began with the introduction, last March, of the 1026,[1] which upgraded many performance characteristics by an order of magnitude or more. Now the 1026 is joined by the lower-frequency (67 kHz–80 MHz) 1003, an all-solid-state signal generator that will probably be the ultimate in this class of instrument for some time to come.

The 1003 is distinctly different from the conventional signal generator. It is different in the way it generates frequencies (by a single-range oscillator, with dividers to produce the lower frequencies) and in the degree to which it maintains frequency, typically within a part per million per 10 minutes. Like the 1026, the 1003 was designed to be the highest-performance signal generator available in its frequency range, and test results indicate that it does in fact enjoy a wide margin over other signal generators now on the market.

[1] G. P. McCouch, "A New 500-MHz Standard-Signal Generator," *General Radio Experimenter*, March 1967.

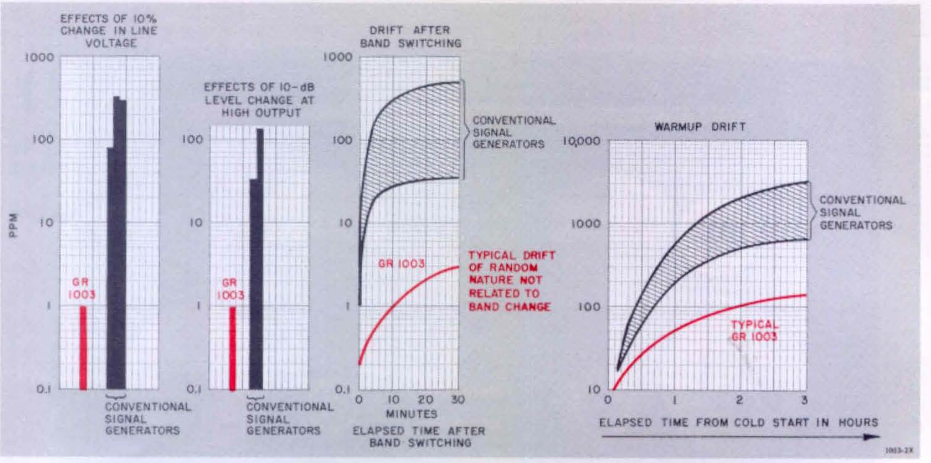


Figure 2. The stability of the 1003 compared with that of typical other signal generators.

For example, Figure 2 illustrates the stability of the 1003 compared with that of typical signal generators of conventional design. One of the chief reasons for the 1003's great advantage in stability is its new approach to frequency generation: Its oscillator is optimally designed for the highest range, and frequency dividers are switched in to produce the lower ranges, imparting the stability of the top range to all other ranges without deterioration. Another result of this approach is a calibration accuracy of ¼%, which is well beyond the reach of other signal generators.

The 1003 uses a motorized dial drive for tuning, sweeping, and programming. For fast, coarse tuning, pushing a rocker switch in the center of the front panel sends the indicator gliding along the slide-rule main frequency dial at about 7% frequency change per second. After using this motor drive to reach the right neighborhood, the user fine-tunes by means of a large rotary control, with each dial division corresponding to 0.01% of the main-dial setting. If this isn't precise enough, the ΔF/F front-panel control provides electronic, backlash-free settability to a few parts per million over a 1000-ppm range.

Both of the fine-tuning controls are fully calibrated in relative terms, so that the user can detune from a given point by a precisely known amount anywhere on the dial.

It is evident from the foregoing that the frequency stability, calibration accuracy, and resolution of the 1003 permit many more meaningful measurements in very narrow-band systems and devices (e.g., ssb receivers, crystal filters), where older signal generators are either marginal or useless because of resolution and drift problems. In such instances the user has had to use synchronizing schemes or synthesizers to provide a stable enough signal, and in the process he has encountered new problems, such as spurious signals, reduction in shielding efficiency, loss of calibration accuracy, to say nothing of the added tuning inconvenience.

The availability of a motor-driven frequency control presents obvious opportunities for both local and remote automatic tuning, and these are exploited by a programmable automatic-frequency-control device. With this unit, one can sweep between adjustable frequency limits and can automatically tune to preset frequencies. The 1003 can be purchased with or without the auto-control unit installed.

The 1003 has a full complement of auxiliary outputs, including a unique F/N monitor that is a byproduct of the frequency-divider method of rf generation. The F/N output frequency is an exact integral fraction 1/N of the actual output, always falling between 67 and 156 kHz. The value of N appears on the dial of the selected frequency range. The constant-level, unmodulated F/N output can be used in many ways, one of which almost suggests itself: measuring or monitoring output frequency indirectly by means of an inexpensive low-frequency counter, even with full modulation.

The main rf output frequency is available at the rear-panel F-monitor connector, which is fully isolated when not in use.

OPERATING CHARACTERISTICS

The 1003 covers its 67 kHz-to-80 MHz range in 10 bands, each somewhat over an octave wide. Over the entire range the instrument can deliver 180 milliwatts of leveled cw power into a 50-ohm load. This is equivalent to 6 volts behind 50 ohms. When the carrier is 95%-modulated, the maximum available carrier level is 3 volts. Envelope distortion and incidental fm are minimized.

The entire warmup frequency drift is typically about 0.01%, and frequency changes due to band switching and to variations in line voltage, load, and level are generally less than 1 part per million (see Figure 2).

The precision 10-dB-per-step attenuator maintains both accuracy and impedance match over the entire 140-dB stepping range. Attenuator error is less than 0.1 dB per step. with a maximum accumulation of 0.5 dB. The attenuator and the continuously adjustable carrier-level control provide an over-all range of 155 dB.

The all-solid-state 1003 draws only 20 watts from the power line. As a result, temperatures are low and components are not under stress. All active devices are operated very conservatively, and the power supplies are short-circuit-proof.

HOW IT WORKS

(See Elementary Diagram, Figure 3)

Oscillator and Power Amplifier

A single-range (34 to 80 MHz) master oscillator is the source of all output frequencies. The key to the instrument's excellent frequency stability is thus the success with which this oscillator was made insensitive to temperature variations and to the influence of the following stages.

A varactor diode permits incremental tuning (ΔF/F) over a limited range. A compensation scheme is used to obtain constant fractional resolution, permitting calibration of the ΔF/F control in ppm. The electronic tuning circuit is also the means by which the signal generator can be frequency-modulated or phase-locked to an external signal, when the ultimate in accuracy and stability is desired.

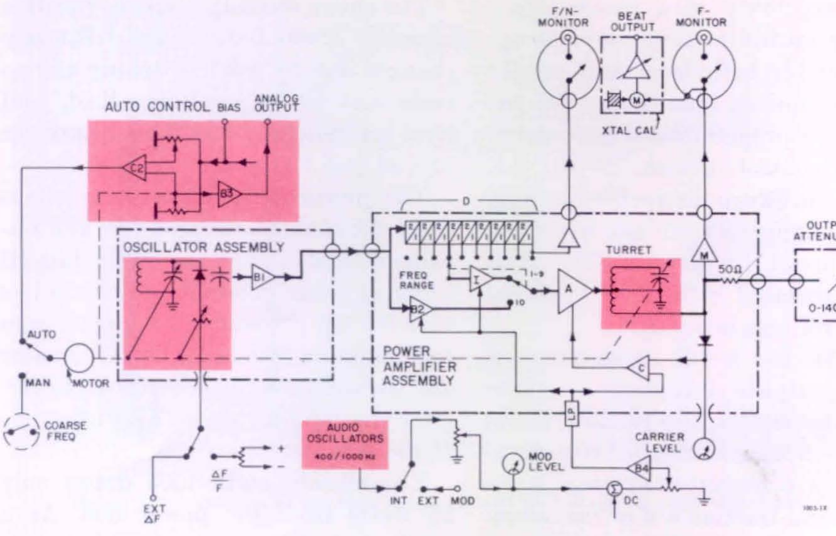


Figure 3. Elementary block diagram.

The oscillator output, after passing through untuned buffer *B1*, enters the power-amplifier unit. On the highest-frequency range (34 to 80 MHz), the rf signal passes through an additional untuned buffer *B2* to the main amplifier *A*. For all lower-frequency ranges, the signal is applied to a series of frequency dividers and thence through untuned buffer (*I*) to the power amplifier. The nine 2:1 dividers give a maximum divisor of 512. Accordingly, the lowest frequency range, produced by the entire cascaded divider chain, is the highest range divided by 512, or 67 kHz to 156 kHz. This low-range output is available as the F/N monitor output, mentioned earlier.

A high degree of isolation between the oscillator and the power amplifier under all conditions practically eliminates all frequency-pulling effects from changes in operating and loading conditions at the output stage. Furthermore, range-switching effects are virtually nil, as Figure 2 shows very clearly, since the same oscillator is used on all bands. Thus no time is wasted in waiting for the frequency to restabilize after band switching, as is typical with other signal generators.

When a particular range is selected, the appropriate number of dividers is activated, and a turret connects the appropriate tank circuit to the power transistor. The tank-circuit variable capacitor is ganged with the oscillator variable capacitor by a non-slip steel cord.

The power amplifier is a 2N3375, whose base voltage controls modulation and output level.

Output System and Leveling

The power-amplifier control voltage is supplied by comparator circuit *C*, which is part of a feedback control system. The other elements of the feedback loop are the tuned amplifier *A* and the detector circuit, whose dc output is compared against a composite reference signal. Any difference between these two signals generates an amplified correction voltage, which makes the rectified output follow the reference voltage. The regulating action is further enhanced by a secondary control path, which varies the drive level and thereby increases the dynamic range of modulation.

Because the stability of the reference voltage is essential to the maintenance of a constant carrier level, all circuits associated with the generation of this reference voltage are supplied with highly stabilized bias voltages. The results of such careful design are evident in Figure 4, which shows the carrier level varying well under 0.01 dB as the line voltage is swung ±10 percent.

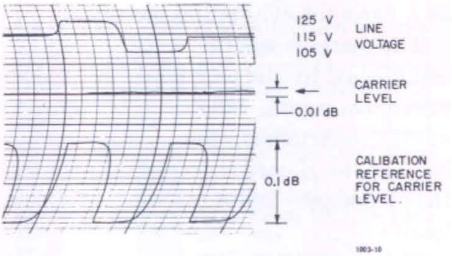


Figure 4. Effects of ± 10% line-voltage swing on carrier level.

The detected rf is measured and displayed by the carrier-level meter, which is calibrated in open-circuit volts (i.e., the voltage behind the 50-ohm source impedance) and in dBm of available power. Since the rf level at the sampling point is kept constant by the control circuit, this point can be considered to be a zero impedance source; a 50-ohm series resistor provides the true 50-ohm source impedance.

The carrier-level control varies the reference voltage of the feedback loop and thus provides continuous adjustment of the leveled output, over a range of 15 dB. The precision step attenuator covers a range of 0 to 140 dB in 10-dB steps.

Modulation

The basic modulating function is performed in the power-amplifier stage by the base voltage on the 2N3375 transistor. This function is linearized through the feedback action, which makes the detected envelope essentially identical to the composite reference signal. In Figure 5 ,which is an X-Y display of a 90% modulated rf signal vs the modulating signal, one can judge the linearity by observing the straightness of the sloped sides of the trapezoid. Another, novel type of presentation (Figure 6) shows the sum of the modulated and modulating signals. Ideally this should produce a horizontal baseline. Departures from the ideal serve

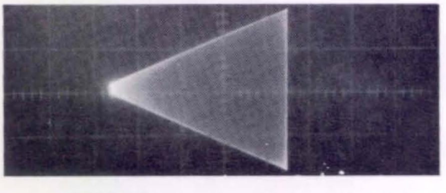


Figure 5. X-Y display of a 90% modulated rf signal (6.5 MHz) vs the modulating signal (400 Hz).

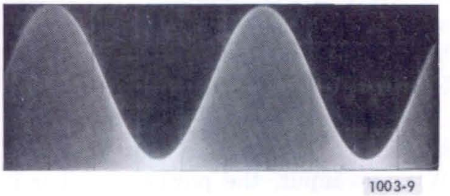


Figure 6. Oscillogram showing addition of modulated (6.5-MHz) and modulating (400-Hz) signals at 90% modulation. Horizontal baseline indicates lack of distortion.

as a basis for evaluating distortion.

The various modes of operation are established by the nature of the applied reference signal, whose instantaneous value determines the instantaneous level of the rf carrier (within, of course, the response limits of the feedback loop).

There are two internal modulating frequencies, 400 Hz and 1 kHz. At either frequency, the modulating signal is highly stable and has very low distortion. The amplitude of these modulating signals can be adjusted by the MOD LEVEL control for up to 95% modulation. The modulation level is monitored in terms of the audio modulating voltage but is calibrated directly in percent. A compensation circuit ensures that a given modulation setting is kept constant over the range of the carrier-level control.

External modulation can be applied with either ac or dc coupling. In the EXT AC mode, any audio-frequency signal can be accepted, controlled, and monitored in the same way as for internal modulation. With sinusoidal waveforms, the modulation passband is flat within 1 dB from 20 Hz to 10 kHz. The ultimate upper limit is the 20-kHz nominal cutoff frequency of the low-pass filter used to feed external signals into the power-amplifier enclosure. On the lower-frequency ranges, however, the rf-amplifier bandwidth also affects the highest usable modulation frequency and percentage modulation.

In the EXT DC mode, the input jack is coupled directly to the amplifier. With no input, the power amplifier is turned off, and a positive-going voltage is required to turn it on. In the off condition, the carrier is down by 50 to 60 dB. Internal limiters protect against excessive modulation input voltages. This mode of operation is particularly useful for remote-control applications and for low-frequency square-wave modulation.

Crystal Calibrator (See Figure 7)

A 1-MHz crystal oscillator is the basic reference source for the optional crystal calibrator. Two more frequencies, 200 kHz and 50 kHz, are derived by division and are thus coherent with the 1-MHz signal. Even the lowest marker frequency can be used up to the highest carrier frequencies.

Since the rf sample for the crystal calibrator is taken from the F-monitor channel (see Figure 3), a high degree of isolation is realized, providing a reverse attenuation well over 100 dB between crystal calibrator and main output. As a result, the crystal calibrator can be used without fear of contaminating the main output with spurious sidebands.

When the F-monitor output is switched on, it is possible to feed an external reference signal through the F-monitor jack and to use portions of the crystal calibrator circuitry as a

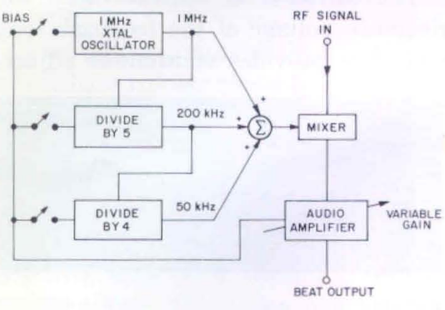


Figure 7. Elementary diagram of the crystal calibrator.

heterodyne frequency meter. In this case, only the mixer-amplifier part of the crystal calibrator is activated.

Auto-Control Unit

The auto-control unit permits a number of automatic tuning operations by either local or remote control. For automatic tuning, the standard frequency-control motor becomes part of a servo positioning system (see Figure 8). An analog dc voltage, proportional to tuning-shaft position, is compared against a reference voltage in a differential amplifier. The amplified error voltage actuates one of two relays, depending on the polarity of the error signal. The appropriate relay energizes the motor to bring the error to zero, and the relay then drops out and turns the motor off. Simultaneously, a dc pulse from a charged capacitor is applied across the motor windings to bring the motor to an abrupt stop. Resolution and accuracy are adequate to permit resettability to within 0.1%.

Rudi Altenbach received his Dipl. Ing. degree in EE from Karlsruhe Technical University in 1948. After three years as development engineer with Siemens and Halske in Germany, he came to Canada, and later to the U.S. From 1951 to 1963 he was engaged in various capacities in the design and development of radar, radio relay equipment and related devices at Canadian Marconi Company, Hermes-Itek Company, and Raytheon Company. In 1963 he joined the GR's Development Engineering staff and has since been working primarily on signal-generator development. He is a member of the IEEE.

The zero-error position is indicated by a neon lamp on the auto-control panel. This lamp is used in the setting of the reference potentiometers to a desired tuning position or limit and also serves as a frequency or position marker. Two internal multiturn high-resolution potentiometers (F1 and F2) permit continuous adjustment of the auto-tune positions or sweeping limits. Many more additional tuning points

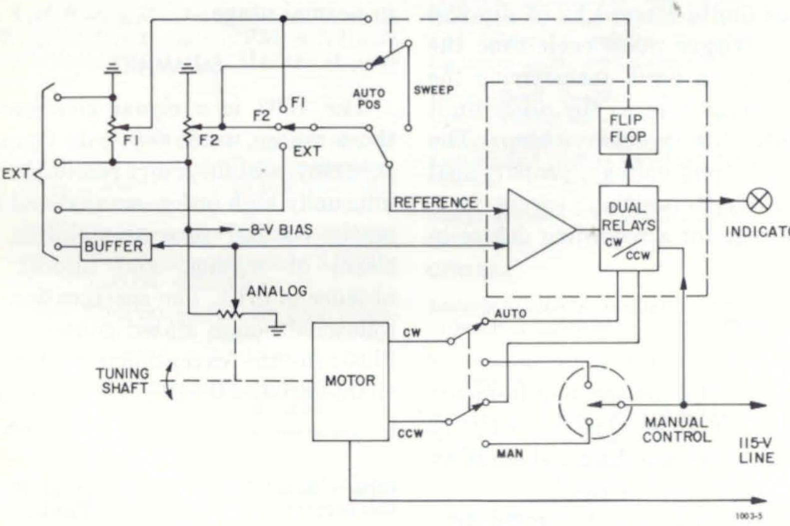


Figure 8. Elementary diagram of the auto-control unit.

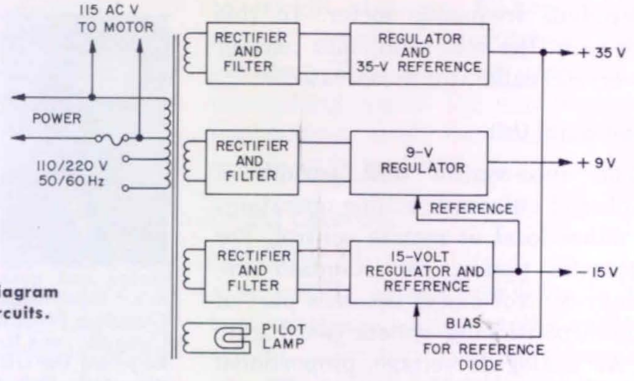


Figure 9. Elementary diagram of the power-supply circuits.

can be added by means of sequentially switched reference signals through an extension socket. An external reference may be either a voltage between 0 and −8 volts or a potentiometer connected to the extension socket. The latter method is preferable for minimum drift. Up to 5 mA can be drawn from the 8-volt bias source, equivalent to over thirty 50-kilohm potentiometers in parallel.

In sweep operation the motor is driven repetitively between the two adjustable limits, F1 and F2. A flip-flop receives a trigger pulse each time the motor reaches a limit, transferring the reference connection to the other limit to actuate the reverse sweep. The analog dc output voltage, proportional to tuning shaft position, serves as a sweep voltage for a recording device in this mode.

Power Supplies (See Figure 9)

Since the total power requirements are very small, it is relatively easy to obtain excellent regulation and stability together with very low ripple.

Especially critical is the regulation of the −15-volt supply that feeds the oscillator section; variations in this supply are kept to a few millivolts under all adverse conditions by use of a temperature-compensated reference diode in a high-gain series regulator circuit. The other two bias voltages (+9 and +35 V) are also stabilized by series regulators. All active elements are silicon, and protection against accidental damage or burnout is achieved through current limiting. Total dissipation, even under continuous short-circuit conditions, is within safe limits in normal usage.

SUMMARY

The 1003 is a signal generator for those whose work demands frequency accuracy, stability, and resolution of an unusually high order, manual and automatic tuning, programmability, precision of setting, and almost total absence of drift. The specifications that follow, although stated conservatively, illustrate the exceptional performance characteristics that have been achieved.

— R. Altenbach

Editor's Note: The basic concept of the 1003 was suggested by A. Noyes, Jr. The instrument was developed by the author, with J. K. Skilling providing the divider circuitry.

SPECIFICATIONS

FREQUENCY

Range: 67 kHz to 80 MHz in 10 ranges: 67 to 156, 135 to 312, 270 to 625, 540 to 1250 kHz, 1.08 to 2.5, 2.16 to 5, 4.32 to 10, 8.64 to 20, 17.28 to 40, and 34.56 to 80 MHz.

Calibration Accuracy: ±0.25%, typically ±0.1%; scale logarithmic, 140 in. total length. Logging scale with vernier, 8500 div, 0.01%/div.

Crystal Calibrator (optional): Markers at 50-kHz, 200-kHz, and 1-MHz intervals, accurate to 20 ppm.

Mechanical Tuning: Fast motor drive, manually or externally controlled; manual fine tuning, 1% per revolution, calibrated, resettable to 0.01%.

Auto-Control Tuning (optional): 0.1% positioning accuracy. Motor drive sweeps between preset limits or tunes on command to preset frequencies (two internally, additional from external dc voltages or dividers). Sweep rate approx 7%/s.

Electronic Tuning: Internal, ±500 ppm, calibrated, settable to better than 2 ppm; external, approx 60 ppm /volt up to ± 1000 ppm typical, limited fm capability. Max input ±15 V into 15 kΩ (+ volts increase frequency).

Stability: After warm-up < 5 ppm per 10 min, typically 1 ppm. Frequency will vary less than 1 ppm as a result of ±10% line-voltage changes, range switching (instant restabilization), rf-level adjustments, or load variations. Warmup drift typically 150 ppm in 3 h at 20°C.

Temperature Coefficient: < 20 ppm/°C, typical.

Carrier Distortion: < 5%, typical.

Noise: A-M, hum and noise sidebands down at least 80 dB relative to carrier. FM, < 3 Hz pk at high-frequency end, < 1 Hz pk at low-frequency end.

RF OUTPUT

Range: CW, 0.1 μV to 6 V behind 50 Ω, 180 mW into 50 Ω (−133 to + 22.6 dBm); modulated, 0.1 μV to 3 V behind 50 Ω, 45 mW into 50 Ω (−133 to + 16.6 dBm).

Source Impedance: 50 Ω. SWR is < 1.02 with attenuator set for 0 dBm or less, < 1.05 for + 10 dBm, < 1.20 for + 20 dBm.

Level Control: Total range, 155 dB. Step attenuator, 140 dB in 10-dB steps; continuously adjustable level control, > 10 dB additional.

Accuracy of Leveled Output Power: ±1 dB at any frequency and termination. Attenuator, ±0.1 dB per 10-dB step, max accumulated error ±0.5 dB.

Level Stability: Warmup drift < 0.3 dB, temperature effects < 0.01 dB/°C, line-voltage variations < 0.02 dB.

Meter: Reads open-circuit volts and dBm.

MODULATION

Level: 0 to 95%, continuously adjustable. Stable within ±1 dB independent of carrier or modulation frequency (within modulation bandwidth) and output level.

ModulationBand width: At 100-kHz carrier, max modulation frequency is 500 Hz for 95% a-m and 2 kHz for 30% a-m. Above 1-MHz carrier, max is 5 kHz for 95% and 10 kHz for 30%.

Meter: Reads 0 to 100%. Accuracy ±5% of full scale, 0 to 95% to 10 kHz within stated modulation bandwidth.

Incidental Angle Modulation: < 0.1 radian pk at 30% a-m.

Internal

Frequency: 400 and 1000 Hz, ±0.5%. Output of 2 V behind 100 kΩ available at panel connector.

Envelope Distortion: < 1% at 50% a-m, < 2% at 70% a-m.

External

AC-Coupled: 20 Hz to 20 kHz, 2 V into 2.5 kΩ for 95% modulation.

Direct-Coupled: DC to 20 kHz. Carrier off with 0-V input; 3-V rf output with +5 V into 10 kΩ. Max input 10 V peak.

AUXILIARY MONITORING OUTPUTS

Main-Output Frequency: At least 0.5 V pk-pk into 50 Ω (CW) at output carrier frequency.

Subharmonic Frequency: At least 0.3 V pk-pk (approx square wave) behind 150 Ω. Frequency (between 67 and 156 kHz) is coherent with and integrally related to carrier frequency by factor N shown on main dial.

Tuning-Shaft Position (with auto-control option): Analog dc voltage proportional to shaft position and logging number. Approx −7.5 V max behind 7500 Ω, or 90 mV for 1% frequency change.

Range Indicator: Contact closure through rear connector.

GENERAL

Leakage: Effects negligible on measurements of receiver sensitivity down to 0.1 μV.

Environment: 10 to 50°C ambient for specified performance.

Accessories Supplied: 874-R22LA Patch Cord, power cord, 12-terminal connector for external controls, spare fuses, hardware for both bench and rack mounting.

Power Required: 105 to 125, 195 to 235, or 210 to 250 V, 50 to 60 Hz, 20 W (33 W with motor operating).

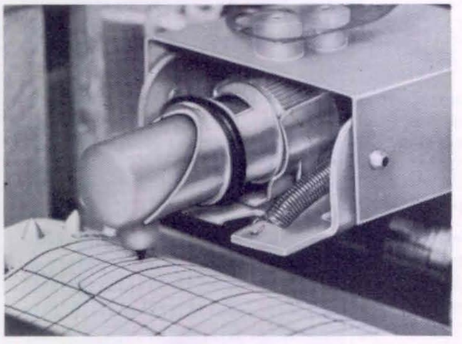
Mounting: Rack-bench cabinet.

Dimensions (width x height x depth): Bench model, 19 x 11 x 15¼ in. (485 x 280 x 390 mm); rack model 19 x 10½ x 12¾ in. (485 x 270 x 325 mm).

Weight (approx): **Net,** 64 lb (30 kg); **shipping,** 87 lb (40 kg).

Catalog Number	*Description*	*Price in USA*
1003-9701	**1003 Standard-Signal Generator**	**$2795.00**
1003-9704	**1003 Standard-Signal Generator Complete with Auto-Control Unit and Crystal Calibrator**	**2995.00**

NEW CARTRIDGE PENS FOR GRAPHIC RECORDER



***fastrak* recorder marker in place on GR 1521 Graphic Level Recorder.**

We have developed a new, cartridge-type, disposable recorder pen for the popular Type 1521 Graphic Level Recorder. Trade-named the *fastrak* recorder marker, the new pen eliminates the problems of ink loading and tip cleaning.

The tip is specially designed for graphic recording, with only 2 grams of force required for proper operation. Each cartridge has about twice the life of one old-style pen refill and can outlast three rolls of chart paper. When the cartridge is empty, replacement is clean, quick, and easy. The entire *fastrak* cartridge, including the writing tip, is disposable. The ink is fast drying, and the marker performs well at all recording speeds. Each marker has a protective cap that prevents drying when the pen is not in use.

fastrak recorder markers are boxed in sets of 12 pens of one color (red, green, or blue) or in an assortment containing four each of the three colors.

The graphic level recorder will henceforth be supplied equipped with *fastrak* markers. A conversion kit is available to adapt existing Type 1521 recorders to the *fastrak* marker.

Catalog Number	*Description*	*Price in USA*
1521-9439	***fastrak* Recorder Marker Conversion Kit** (includes marker-holder set, installation instructions, and combination marker set)	**$25.00**
1521-9449	***fastrak* Combination Marker Set** (includes 4 red, 4 green, 4 blue markers)	**15.00**
1521-9446	***fastrak* Marker Set,** 12 red markers	**15.00**
1521-9447	***fastrak* Marker Set,** 12 green markers	**15.00**
1521-9448	***fastrak* Marker Set,** 12 blue markers	**15.00**

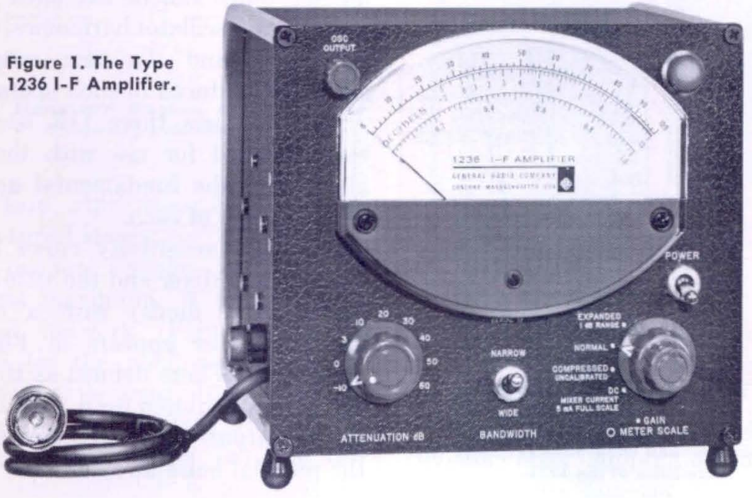


Figure 1. The Type 1236 I-F Amplifier.

A NEW 30-MHz AMPLIFIER WITH TWO BANDWITHS

The 30-MHz amplifier is a popular instrument that goes under a variety of names. It is an important element in a precision heterodyne receiver, and it is sometimes called, somewhat loosely, a receiver. Since it often serves, in combination with a local oscillator and mixer, as a detector for bridge measurements, it is also known as a null detector. No matter what its name, it is practically indispensable for a great many measurements.

The new GR 1236 is a low-noise, high-gain 30-MHz tuned amplifier with two switch-selected bandwidths, giving the user a choice of a "narrow" band of 0.5 MHz or a "wide" band of 4 MHz. One would typically use the narrow band for operation at lower frequencies, switching to the wide bandwidth at higher local-oscillator frequencies where frequency stability is often a problem. The narrow- and wide-band response characteristics are shown in Figure 2.

A six-inch taut-band meter with calibrated linear and decibel scales gives excellent resolution. The top 10 percent of the scale can be expanded to give a full-scale range of 1 dB with a resolution of 0.02 dB per small division. When the meter scale switch is set to COMPRESSED, the agc loop compresses the meter scale to about 50 dB. This feature is almost

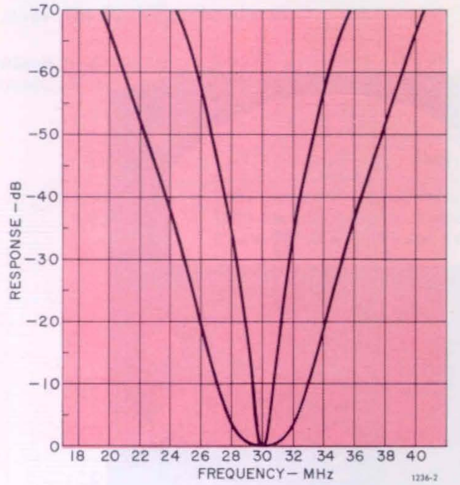


Figure 2. Narrow- and wide-band response characteristics of the 1236.

indispensable when the instrument is used as a null detector in a bridge measuring system.

The attenuator covers a range of 70 dB in 10-dB steps, with an accuracy of $\pm$ (0.1 dB + 0.1 dB/10 dB). The accumulated error will generally not exceed 0.3 dB. Because of the excellent repeatability of the attenuator, it is entirely practical to calibrate it against an external standard, thus reducing the attenuator error to that of the standard.

The 1236 combines easily with the new, highly sensitive Type 874-MRAL Mixer (see page 19) and one of the GR line of oscillators to form a wide-range measuring receiver. The 1236 includes a separate adjustable regulated power supply for the local oscillator.

With a given oscillator, the frequency range can be extended by use of the local-oscillator harmonics, though sensitivity and dynamic range are somewhat reduced in such operation.

Table 1 lists three GR oscillators recommended for use with the 1236, along with the fundamental and harmonic ranges of each.

A typical sensitivity curve for the 874-MRAL Mixer and the 1236 (in the narrow-band mode) with a suitable local oscillator appears in Figure 3. Sensitivity is here defined as the input signal level required for a 3-dB increase in the output of the i-f amplifier over the residual noise level.

Circuit

A low-noise preamplifier uses two Nuvistors in cascode in the input stage and a third Nuvistor in the output stage. The heater supply of the Nuvistors is regulated to achieve high gain stability vs line-voltage changes.

The preamplifier output is fed to a ladder-type step attenuator, which covers 70 dB in 10-dB steps. The output meter is used for interpolation between steps.

The postamplifier consists of one untuned and three tuned stages. The gain of the untuned stage is controlled by a front-panel control, with coarse and fine adjustments.

When the METER SCALE switch is in the COMPRESSED position, it activates

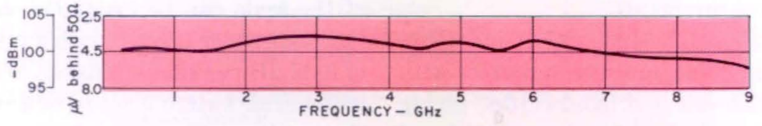


Figure 3. Typical sensitivity curve for receiver system comprising 1236, local oscillator, and Type 874-MRAL Mixer.

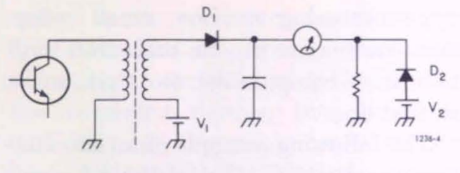


Figure 4. Elementary diagram of the detector linearizing network.

M. Khazam received his degree in Electronic Engineering from the Delft University of Technology, Holland, in 1957, and from 1957 to 1960 was a project engineer with the Laboratory for Electronic Developments for the Armed Forces, in Holland. He joined General Radio in 1962 as a development engineer in the Microwave Group and has since specialized in the development of vhf-uhf instruments and components.

the agc loop, which controls the gain of the two tuned stages.

The output voltage is about 2 volts rms maximum. A temperature-stabilized network (Figure 4) compensates for the nonlinear characteristics of the detector diode. In this network, *V1* is adjusted for a linear response in the upper part of the meter scale and *V2* is adjusted to optimize the lower part. Figure 5 shows the response with and without compensation.

The measured deviation from a linear response of a compensated detector circuit is plotted in Figure 6. A full-scale meter deflection corresponds to 2 volts rms rf voltage. Point *A* is the reference point, in this case 100% meter deflection, *B* and *C* are points of zero error; their positions are determined by *V1* and *V2*. The three points of zero error may be positioned for minimum error over either the whole range or part of the range.

The power supply consists of a Nuvistor plate supply, a supply for the transistors and for the Nuvistor heaters, and a local-oscillator plate and heater supply. All voltages except the local-oscillator heater supply are regulated.

Applications

The 1236 will be widely used with a local oscillator and mixer as a sensitive null detector for bridges, such as GR's 1602 UHF Admittance Meter and 1607 Transfer-Function and Immittance Bridge. Its excellent performance char-

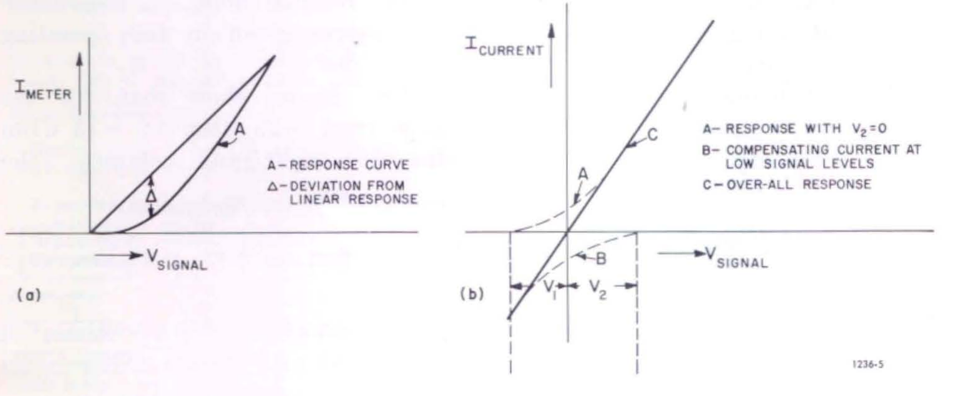


Figure 5. Uncompensated (*left*) and compensated (*right*) response of detector circuit.

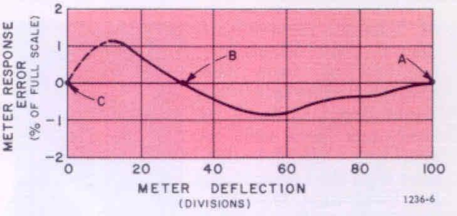


Figure 6. Measured deviation of compensated detector circuit from linear response.

acteristics suggest that it will also be a popular relative-signal-level meter in attenuation measurements, SWR measurements at low signal levels with slotted lines, reflection-coefficient measurements with hybrids or directional couplers, etc. SWR meters consisting of a tuned detector and a high-gain low-frequency amplifier often require a signal level that is too high for measurements on nonlinear devices. The heterodyne detector, with its much higher sensitivity, is the preferred SWR meter in such instances, and it is also recommended in general for precision measurements of both high and low SWR.

Measurements of small reflection coefficients with a directional coupler or a hybrid reflectometer are restricted by the directivity of the coupler or the balance of the hybrid and by the dynamic range and sensitivity of the detector. By the use of precision tuners and such terminations as those available in the GR900 line, the directivity or balance can be made almost perfect at any one frequency. Then, with a heterodyne-measuring receiver, small reflection coefficients can be measured with accuracy comparable to that of a slotted line.

The following example of an attenuation measurement using the i-f series substitution method indicates the accuracy and dynamic range attainable with this system.

The measurement setup is shown in Figure 7. The receiver consists of a 1236 I-F Amplifier, a 1208 Oscillator (40–530 MHz), and an 874-MRAL Mixer. The measuring frequency is 500 MHz.

The 1236 output reading (attenuator setting plus meter indication) is noted with and without the unknown attenuator in the circuit. The difference of the two readings is the measured attenuation. These measurements are repeated at different signal levels to determine the useful dynamic range of the system.

The results appear in Table 2. The top two rows give the 500-MHz signal level at the detector. The third row gives the attenuator values as measured, while in the fourth row the numbers are corrected for the 1236 attenuator errors. The numbers in the fifth row are corrected for the error caused by the residual noise, in accordance with curves given in the operating instructions.

These figures show that, for the range from -73 dBm to -13 dBm (the five right-hand columns), the

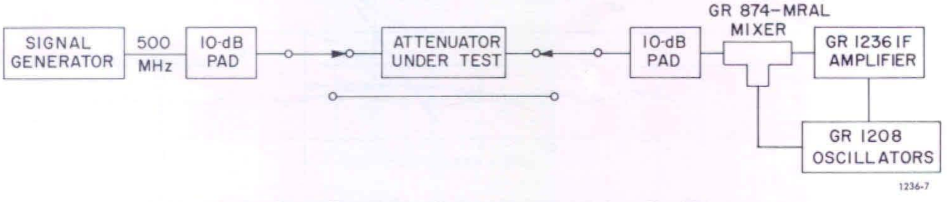


Figure 7. Setup for attenuation measurement described in text.

spread in the uncorrected attenuation figures is 0.19 dB, with a maximum deviation from mean of 0.11 dB. For the corrected figures, the spread is 0.04 dB and the maximum deviation from mean 0.02 dB. With the corrections for residual noise applied, the spread over the range from −83 to −13 dBm is again 0.04 dB, with a maximum deviation from the mean of 0.02 dB. Here the accuracy is perchance considerably better than that given in the specifications.

— M. Khazam

TABLE 1

Local Oscillator Type	*Frequency Range, MHz*			
	Fundamental	*2nd harmonic*	*3rd harmonic*	*4th harmonic*
1208-C	40-530	100-1030	165-1530	230-2030
1209-C	220-950	470-1870	720-2790	970-3710
1218-B	870-2030	1770-4030	2670-6030	3570-8030

TABLE 2

Min signal level at detector in dBm	−95	−83	−73	−63	−53	−43	−33
Max signal level at detector in dBm	−73	−63	−53	−43	−33	−23	−13
Measured attenuation in dB	17.2	19.05	19.73	19.69	19.61	19.75	19.80
Measured attenuation corrected for attenuator error in dB	17.4	19.25	19.78	19.76	19.76	19.80	19.79
Measured attenuation corrected for attenuator error and residual noise in dB	20.1	19.8	(noise not a factor)				

SPECIFICATIONS

Center Frequency: 30 MHz.

Bandwidth: Wide band, approx 4 MHz; narrow band, approx 0.5 MHz, selectable by panel switch.

Noise Figure: Typically 2 dB.

Sensitivity: From a 400-Ω source, for a 3-dB increase in meter deflection, < 9 μV (wide band) or < 3.5 μV (narrow band).

Meter Characteristics

Normal Scale: −2 to 10 dB. Linearity ±0.2 dB over 0 to 10-dB range.

Expanded Scale: 1-dB full scale. Linearity ±0.03 dB.

Compressed Scale: 40-dB min range.

Attenuator

Range: 0 to 70 dB in 10-dB steps.

Accuracy: ± (0.1 dB + 0.1 dB/10 dB) at 30 MHz.

Continuous Gain Control: 10-dB min range.

Video Output (Modulation): 1.5 V max; 1-MHz bandwidth.

I-F Output: 0.5 V max into 50 Ω.

Power-Supply Output: 150 to 300 V dc, adjustable, at 30 mA, regulated; 6.3 V ac at 1 A.

Power Required: 105 to 125, 195 to 235, or 210 to 250 V, 50 to 60 Hz, 22 W (without oscillator).

Accessories Supplied: Power cord, spare fuse.

Accessories Available: As local oscillator, GR 1208, 1209-C, 1209-CL, 1215, 1218, and 1361; 874-MRAL Mixer; GR874 low-pass filters, attenuators, adaptors, etc.

Mounting: Convertible-bench cabinet.

Dimensions (width x height x depth): 8 by 7⅜ by 8 in. (205 x 190 x 205 mm).

Weight: Net, 12½ lb (6 kg); **shipping,** 14¾ lb (7 kg).

Catalog Number	*Description*	*Price in USA*
1236-9701	**1236 I-F Amplifier**	**$675.00**

GR Product Notes

CARD-PUNCH COUPLER

The new 1791 Card-Punch Coupler converts the binary-coded digital output of the GR 1680-A Automatic Capacitance Bridge into the 10-line decimal-coded contact closures required by an IBM 526 Printing Summary Punch. Up to 22 digits of parallel data can be accepted from one or more sources. Since the coupler is a systems component, in some instances requiring custom treatment of connections, price will be quoted on an individual basis.

GR900 ADAPTOR AND AIR-LINE SETS

Now available are complete sets of GR900 precision adaptors and reference air lines, mounted in mahogany cases with foamed-plastic inserts. The 0900-9451 GR900 Precision-Adaptor Set includes all the adaptors needed to mate GR900 connectors with male and female BNC, C, N, SC, OSM*/BRM, TNC, Amphenol APC-7, Precifix 7 mm, and GR874 connectors.

The 0900-9452 GR900 Reference Air-Line Set contains one each of the six lengths (5, 6, 7.5, 10, 15, and 30 cm) of Type 900-LZ Reference Air Lines, plus a 900-WN4 Short-Circuit Termination and a 900-WO4 Open-Circuit Termination, both of which are commonly used with the air lines.

The storage case alone is also available, for those who would like to give their GR900 components the maximum protection against damage and dirt.

*OSM is a registered trademark of Omni-Spectra, Inc.

AMPLITUDE-REGULATING POWER SUPPLY

The Type 1263 Amplitude-Regulating Power Supply is now the 1263-C, the new suffix denoting a regulated dc heater supply for improved oscillator performance and a relocated output-rectifier connector for more convenient installation in relay racks.

GR Product Notes

(continued)

GR874 MIXER

The new 874-MRAL Mixer is an improved version of the 874-MR, with significantly better sensitivity and with GR874 locking connectors. Frequency range is 10 MHz to 9 GHz, with a maximum i-f of 60 MHz. A natural partner for the 1236 I-F Amplifier described earlier in this issue.

GR874 ATTENUATORS

The 874-G14 14-dB fixed attenuator described in the October 1965 *Experimenter* is now available in a locking version, the 874-G14L. GR874 single-section, T-type resistance pads are now offered in 3-, 6-, 10-, 14-, and 20-dB sizes, locking and non-locking.

Catalog Number	*Description*	*Price in USA*
0900-9451	**GR900 Precision-Adaptor Set**	**$1210.00**
0900-9452	**GR900 Reference Air-Line Set**	**682.00**
0900-9450	**GR900 Storage Case**	**35.00**
1263-9703	**1263-C Amplitude-Regulating Power Supply**	**485.00**
0874-9561	**874-G14L 14-dB Fixed Attenuator**	**32.50**
0874-9947	**874-MRAL Mixer**	**65.00**

WESCON

All the new instruments described in this issue can be seen at Wescon, at the Cow Palace, San Francisco, August 22 through 25, 1967. In addition, the GR booth (No. 3015–3018) will feature operating displays of GR frequency synthesizers, automatic component-measuring systems, recording wave analyzers, and the Type 1026 Standard-Signal Generator described in the March, 1967 *Experimenter*.

Development engineers from our Concord and Bolton plants will join engineers from our San Francisco and Los Angeles offices in staffing the GR Wescon booth. We look forward to meeting you there.

GENERAL RADIO COMPANY
WEST CONCORD, MASSACHUSETTS 01781

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NBS TO CALIBRATE BOLOMETERS FITTED WITH 14-mm PRECISION COAXIAL CONNECTORS

The National Bureau of Standards Radio Standards Laboratory, Boulder, Colorado, announces a calibration service for the measurement of effective efficiency * of coaxial bolometer units fitted with 14-mm precision connectors (e.g., GR900), over a continuous frequency range from 4 to 8.5 GHz. Use of 14-mm precision connectors permits greater accuracy of measurement at radio frequencies than was possible with the older type N connectors, according to the NBS announcement. At present the calibration service is available for measurement at a nominal power of 10 milliwatts and for bolometer units fitted with thermistor-type elements having a nominal operating resistance of 200 ohms.

* The effective efficiency of a bolometer unit is the ratio of substituted dc power in the unit to the rf power dissipated within the bolometer unit.

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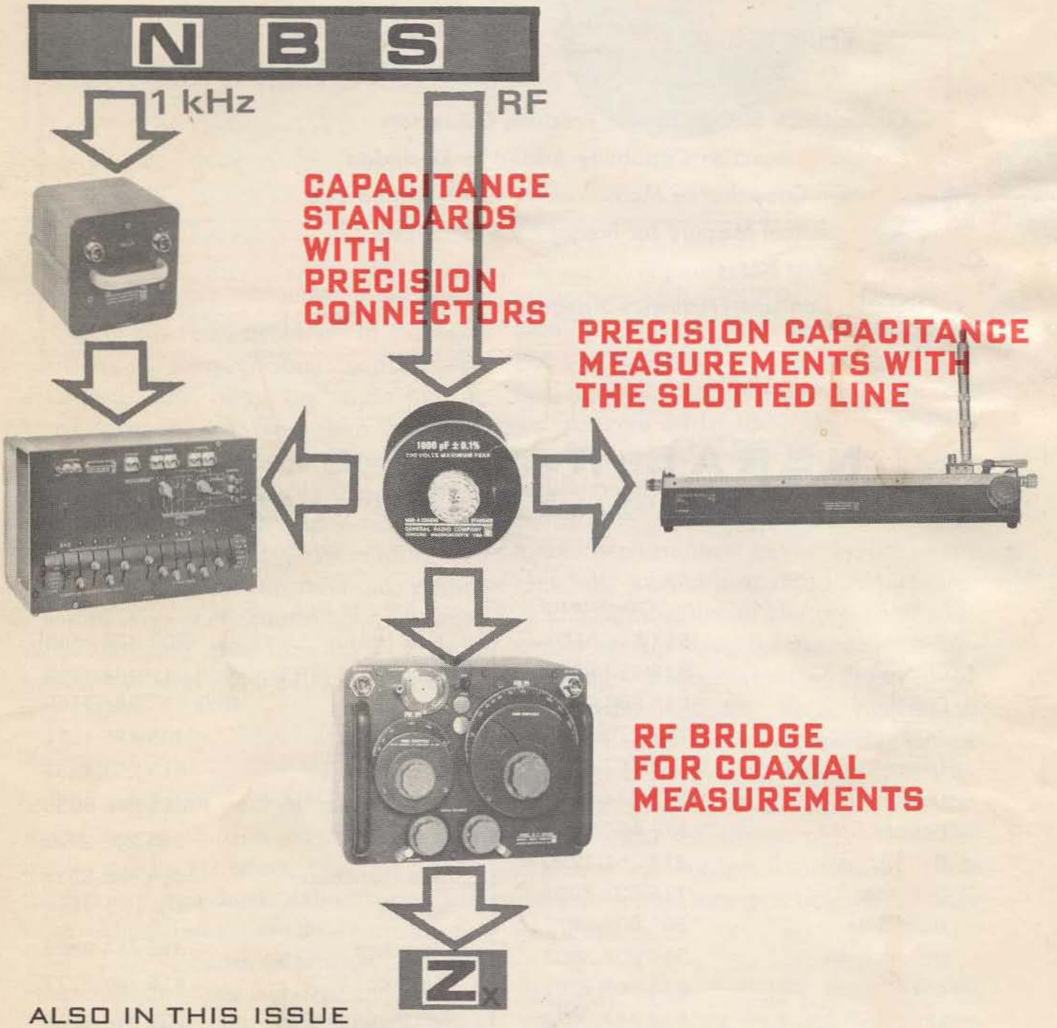
Experimenter



ALSO IN THIS ISSUE

PROGRAM PRESET UNIT FOR FREQUENCY SYNTHESIZERS

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REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

CAPACITANCE STANDARDS WITH PRECISION CONNECTORS

One of the chief problems in the calibration of two-terminal capacitance standards has been the variation in stray capacitance from one binding-post connection to another. Significant improvement in this area had to await the development of a coaxial connector with high repeatability, low inductance, and a precisely known reference plane. Fortunately, such connectors are now available, and they are put to good use on the new TYPE 1406 Capacitance Standards.

The precision coaxial connector has made its reputation on its low SWR in the gigahertz region, and few if any microwave standards laboratories are now without benefit of these connectors. The well justified attention to uhf characteristics has tended to obscure the fact that the precision coaxial connector is also a powerful tool at radio frequencies and all the way down to dc.[1]

The introduction of the TYPE 1406 Coaxial Capacitance Standards, equipped with GR900® precision coaxial connectors, focuses attention on the usefulness of such connectors at low frequencies. The 1406, available in five values from 50 to 1000 pF, would be a good standard capacitor even with ordinary connectors. With the highly repeatable, low-inductance GR900, it offers performance never before commercially available at radio frequencies.

The significance of the precision connector on the standard capacitor is far-reaching. The National Bureau of Standards will no longer calibrate capacitors with binding posts or other unshielded terminals[2]; the Bureau further states[3] that maximum calibration accuracy can be offered only when the capacitor is equipped with precision coaxial connectors meeting the IEEE Standard.[4]

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[1] R. N. Jones and L. E. Huntley, "Precision Coaxial Connectors in Lumped Parameter Immittance Measurement," *Proceedings of the IEEE*, Vol 55 No. 6, June 1967.

[2] Federal Register, Vol 32, No. 15, 24 January, 1967.

[3] R. N. Jones and R. L. Jesch, *High Frequency Immittance Calibration Services of the National Bureau of Standards*, U. S. Department of Commerce, National Bureau of Standards, October 1, 1965.

[4] D. E. Fossum "Progress Report of the IEEE Instrumentation and Measurement Group Technical Subcommittee on Precision Coaxial Connectors," *IEEE Transactions on Instrumentation and Measurement*, Vol. IM-13 No. 4, December 1964.

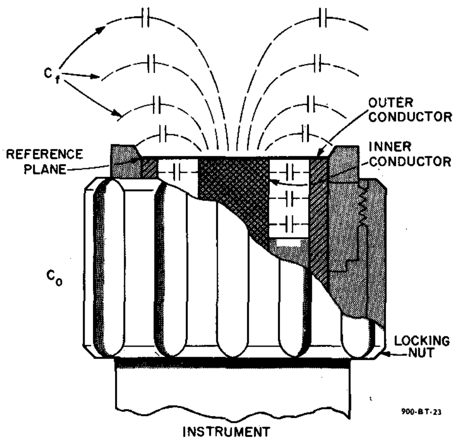


Figure 1. Open GR900® precision coaxial connector showing fringing capacitance (C_f).

The emphasis of NBS and of others on coaxial connection is easy to understand. Two-terminal capacitance standards are usually calibrated in terms of the capacitance added to the bridge terminals. When these terminals are binding posts, this added capacitance is not clearly defined, since the addition of the standard changes the stray capacitance between the binding posts and also adds capacitance from the capacitor case to the high binding posts.[5] Different bridge and binding-post configurations mean different strays and consequent variations in the net capacitance added to different instruments by the same standard.

The use of a GR900 precision connector for the terminals changes all this. The connector has a precisely known reference plane, and capacitance can be defined and measured as the internal capacitance up to this reference plane.

An open GR900 connector on an instrument has a fringing capacitance (Figure 1) consisting of the total stray capacitance beyond the reference plane.

[5] John F. Hersh, "A Close Look at Connection Errors in Capacitance Measurements," *General Radio Experimenter*, July 1959.

When a capacitor with a GR900 connector is added to the open connector, the fringing capacitance is eliminated, and the net increase in capacitance equals the value of the capacitor as measured at its reference plane minus the value of its fringing capacitance. The fringing capacitance of a 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. 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 low over-all inductance of the new standard means that the capacitance change with frequency is negligible up into the megahertz range (see Figure 2). The difference between the 1-kHz and the 1-MHz values of capacitance is, for the lower capacitance values, smaller than the uncertainty of the 1-kHz measurement on a precision bridge. Assuming that the useful upper-frequency limit is the point where the effective capacitance differs from its 1-kHz value by 10 percent, the 1000-pF capacitor is useful to 16 MHz, the 50-pF unit to 83 MHz.

Because of its wide frequency range and its acceptability by the National Bureau of Standards for calibration above 30 kHz, the 1406 capacitor is expected to be used chiefly as a standard for the calibration of two-terminal bridges and other impedance-measuring instruments. Of course, the most convenient arrangement and the most accurate measurements result when the bridge is equipped with a GR900 connector, as is the rf bridge described elsewhere in this issue. The next best thing is a precision adaptor, and two

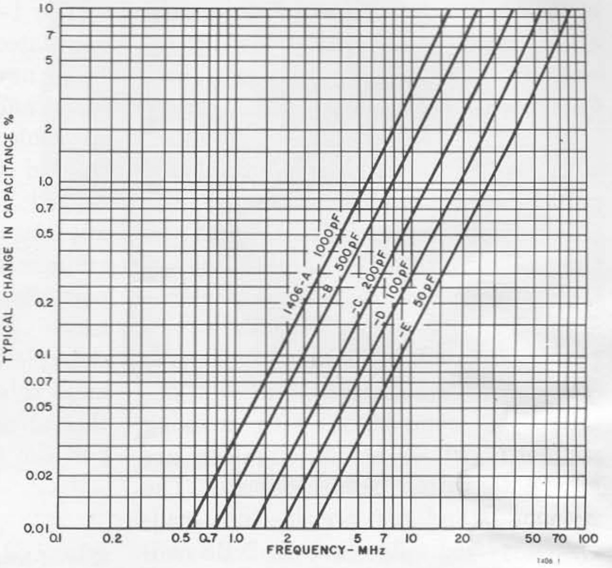


Figure 2. Typical percent increase in capacitance of Type 1406 Coaxial Capacitance Standards with frequency.

Figure 3. Type 1615-P2 Coaxial Adaptor.

of these have been designed specifically for use with bridges. One, the Type 1615-P2 (Figure 3), 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 (Figure 4) mates with binding posts on ¾-inch spacing, such as are used on the GR 716 Capacitance Bridges, with other posts

Figure 4. Type 900-Q9 Adaptor.

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.

Description

The air capacitor in the Type 1406 Standard is a rigid assembly of parallel plates mounted in a shielded cylindrical enclosure, with a GR900 precision connector providing the terminals. Metal parts are aluminum to minimize stresses caused by differences in thermal expansion. The capacitor plate assembly is insulated from the mounting plate by cross-linked thermoset polystyrene insulators, treated to reduce the effects of humidity. The heavy cover plate is threaded to screw into the cylindrical housing; the resulting rigid assembly eliminates mechanical instabilities that could cause small variations in capacitance.

Calibration

A calibration certificate supplied with each capacitor gives the measured value of capacitance at 1 kHz, the temperature and relative humidity at the time of measurement, and the calculated effective capacitance of the unit at 1 MHz. The 1-kHz capacitance applies at the well defined reference plane of the GR900 connector, and it is obtained by comparison with working standards whose absolute values are known to within ±0.01 percent. The working standards are in turn calibrated periodically against NBS-calibrated reference standards.

The 1-MHz value of capacitance is calculated from the measured low-frequency capacitance and the known inductance of the capacitor. Of considerable importance is the fact that, due to the low inductance and the rugged construction of the standard, any change in its 1-MHz capacitance value will be accompanied by a proportional change in its 1-kHz capacitance value. Thus a standard can, for example, be NBS-calibrated at 1 MHz, and this high-frequency value can thereafter be monitored by means of periodic 1-kHz measurements.

— R. W. Orr

A brief biography of Mr. Orr appeared in the August 1966 *Experimenter.*

SPECIFICATIONS

Calibration: A certificate of calibration is supplied with each unit, giving the measured capacitance at 1 kHz and at a specified temperature and relative humidity. The measured capacitance is the capacitance at the reference plane of the GR900 connector. This value is obtained by comparison, to a precision better than ±0.01%, 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 0.05%/year.

Accuracy: Capacitance is adjusted to within 0.1% of nominal value.

Residual Parameters: See table below. Dissipation factor is given for 40% RH and varies as the 3/2 power of frequency above about 100 kHz.

Insulation Resistance: Greater than $10^{12}\ \Omega$ at 23°C and less than 50% RH.

Temperature Coefficient of Capacitance: Typically 10 to 20 ppm/°C between 20 and 70°C.

Accessories Available: Type 1615-P2 for convenience in calibrating with 1615-A Capacitance Bridge; Type 900-Q9 Adaptor for connection to GR 716 Capacitance Bridge, Boonton Q Meter, and other binding-post-equipped instruments.

Terminal: GR900 precision coaxial connector.

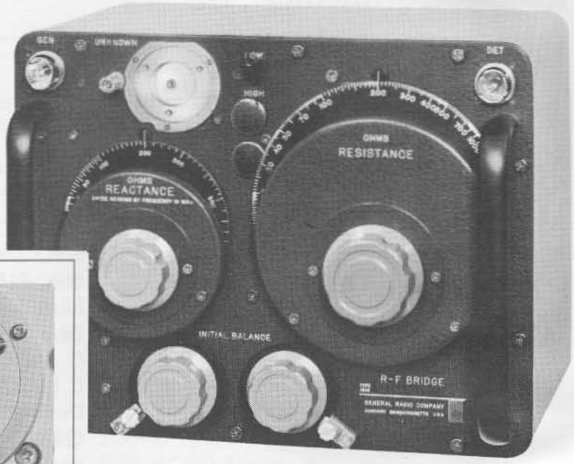
Mounting: Aluminum panel and cylindrical case.

Dimensions: 3 by 5¼ in. (77 by 135 mm).

Weight: Net, 1¾ lb (0.8 kg); **shipping** (est) 5 lb (2.3 kg).

Catalog Number	*Type*	*Nominal Capacitance*	*Peak Volts*	*Typical Dissipation Factor* 1 kHz (40% RH)	*Typical Dissipation Factor* 1 MHz	*Typical Inductance*	*Price in USA*
1406-9701	**1406-A**	1000 pF	700	3×10^{-6}	50×10^{-6}	8.6 nH	**$120.00**
1406-9702	**1406-B**	500 pF	900	5×10^{-6}	30×10^{-6}	8.4 nH	**115.00**
1406-9703	**1406-C**	200 pF	1200	20×10^{-6}	25×10^{-6}	8.1 nH	**110.00**
1406-9704	**1406-D**	100 pF	1500	30×10^{-6}	20×10^{-6}	7.6 nH	**105.00**
1406-9705	**1406-E**	50 pF	1500	50×10^{-6}	15×10^{-6}	6.7 nH	**100.00**
1615-9602	**1615-P2 Coaxial Adaptor,** GR900 to binding post						**60.00**
0900-9874	**900-Q9 Coaxial Adaptor,** GR900 to binding post						**50.00**

COAXIAL-CONNECTION CAPABILITY ADDED TO RF BRIDGE



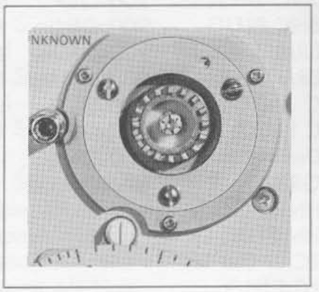


Figure 1. Type 1606-B RF Bridge. Inset shows GR900® adaptor in place.

The new Type 1606-B RF Bridge (Figure 1) differs from its predecessor by the simple but important fact that its "unknown" terminals can be quickly and easily converted to accept a GR900® precision coaxial connector. This means that the bridge can be precision-calibrated against GR900 components or against the Type 1406 Coaxial Capacitance Standards described elsewhere in this issue. It also means that components and networks equipped with GR900 connectors can be accurately measured on the bridge. Moreover, almost any type of coaxial connector can be connected to the bridge terminals through one of the many GR900 adaptors available.

To review the principal characteristics of the 1606 RF Bridge: Resistance and reactance are indicated directly in ohms, over a frequency range from 400 kHz to 60 MHz. Measurements are made by a series-substitution method, in which the bridge is first balanced with its unknown terminals short-circuited, then balanced with the unknown impedance connected. Since its introduction, the bridge has been very widely used in rf measurements on antennas, transmission lines, networks, and components.

Conversion for Coaxial Connection

The Type 1606-P2 Precision Coaxial Adaptor Kit (Figure 2) converts the Type 1606-B RF Bridge either to the standard GR900 configuration or to a 14-mm flange connection. In addition to the necessary GR900 and flange connector hardware, the kit includes both GR900 and GR874 50-ohm compensating lines, for matching out the bridge terminal capacitance.

With the GR900 adaptor in place, the bridge can be quickly converted for use with any of the popular coaxial connectors, via GR900 precision adaptors. With the GR874 50-ohm compensating line in place, the extensive line of GR874 adaptors is added to the compatibility list. The many connection possibilities are given in Table 1.

The most precise measurements can be performed with the direct GR900 connection. In this case the unknown is

Figure 2. Type 1606-P2 Precision Coaxial Adaptor Kit.

assembled with a GR900 connector for the direct measurement on the bridge.

Precise Bridge Calibration

Perhaps the most important advantage of the precision coaxial connector is that it serves as the essential link between the bridge and the most accurate available calibration standards, including new coaxial capacitance standards (see page 3) acceptable for calibration by the National Bureau of Standards.

With the coaxial connection, a precision calibration is a quick and easy matter. The bridge is balanced with the GR900 adaptor (from the 1606-P2 kit) in place and with the standard connected to it. Then, if the bridge is to be used without the GR900 adaptor, the known capacitance of the adaptor is subtracted out. In addition, the 1606-P2 kit contains a spacer sleeve that can be added to duplicate the GR900 adaptor capacitance in open-terminal measurements (e.g., measurements where non-coaxial components are connected directly to the bridge).

TABLE I

Adaptors from 1606-B Bridge (with 1606-P2 Kit) to Various Coaxial Systems

Connector on Unknown	*Through GR900**	*Through GR874***
APC-7 (7mm)	900-QAP7	874-QAP7L
BNC JACK	900-QBP	874-QBPA
BNC PLUG	900-QBJ	874-QBJL
BRM JACK	900-QMMP	874-QMMPL
BRM PLUG	900-QMMJ	874-QMMJL
C JACK	900-QCP	874-QCP
C PLUG	900-QCJ	874-QCJL
Dezifix	900-QPF7	
GR874	874-Q900L	
HN JACK		874-QHPA
HN PLUG		874-QHJA
LC JACK		874-QLPA
LC PLUG		874-QLJA
LT JACK		874-QLTP
LT PLUG		874-QLTJ
Microdot JACK		874-QMDP
Microdot PLUG		874-QMDJL
OSM† JACK	900-QMMP	874-QMMPL
OSM† PLUG	900-QMMJ	874-QMMJL
Precifix A	900-QPF7	
SC JACK	900-QSCP	874-QSCP
SC PLUG	900-QSCJ	874-QSCJL
TNC JACK	900-QTNP	874-QTNP
TNC PLUG	900-QTNJ	874-QTNJL
UHF JACK		874-QUP
UHF PLUG		874-QUJL

* For 50-ohm compensation, add 1606–3060 unit, supplied with 1606-P2 kit.
** For 50-ohm compensation, add 1606-3070 unit, supplied with 1606-P2 kit. Otherwise, add 874-Q900L adaptor.
† Registered trademark of Omni Spectra, Inc.

An alternative practice is to leave the GR900 adaptor in place and to add to it a 900-Q9 adaptor, which converts from GR900 to binding posts. The parasitic inductance and capacitance values are furnished with this adaptor.

Some of the many standards that can be used to calibrate the bridge are listed in Table 2.

— J. Zorzy

A brief biography of Mr. Zorzy appeared in the August 1966 *Experimenter*.

TABLE 2

GR900-Equipped Standards

Capacitance	
1406-A Coaxial Capacitance Standard	1000 pF
1406-B Coaxial Capacitance Standard	500 pF
1406-C Coaxial Capacitance Standard	200 pF
1406-D Coaxial Capacitance Standard	100 pF
1406-E Coaxial Capacitance Standard	50 pF
900-L10 Precision Air Line	6.6 pF
900-L15 Precision Air Line	10 pF
900-L30 Precision Air Line	20 pF
900-LZ5 Reference Air Line	3.3333 pF
900-LZ6 Reference Air Line	4.0000 pF
900-LZ7H Reference Air Line	5.0000 pF
900-LZ10 Reference Air Line	6.6667 pF
900-LZ15 Reference Air Line	10.000 pF
900-LZ30 Reference Air Line	20.000 pF
900-WO4 Open-Circuit Termination	2.67 pF
Resistance	
900-W50 Standard Termination	50 Ω
900-W100 Standard Termination	100 Ω
900-W200 Standard Termination	200 Ω
900-WR110 Standard Mismatch	45.45 Ω
900-WR120 Standard Mismatch	41.67 Ω
900-WR150 Standard Mismatch	33.33 Ω

SPECIFICATIONS

RANGES OF MEASUREMENT

Frequency: 400 kHz to 60 MHz. Satisfactory but somewhat less accurate operation can be obtained at frequencies as low as 100 kHz and somewhat above 60 MHz.

Reactance: ±5000 Ω at 1 MHz. This range varies inversely as the frequency; at other frequencies the dial reading must be divided by the frequency in MHz.

Resistance: 0 to 1000 Ω.

ACCURACY

Reactance: At frequencies up to 50 MHz, $\pm 2\% \pm (1 + 0.0008\,Rf)\,\Omega$ where R is the measured resistance in ohms and f is the frequency in MHz.

Resistance: At frequencies up to 50 MHz,

$$\pm \left[1\% + 0.0024 f^2 \left(1 + \frac{R}{1000} \right) \% \pm \frac{10^{-4} X}{f} \Omega + 0.1\Omega \right]$$

(where X is the measured reactance in ohms). Subject to correction for residual parameters.

GENERAL

Generator: External only (not supplied) to cover desired frequency range. Recommended, Type 1211-C and Type 1215-C Unit Oscillators, Type 1330-A Bridge Oscillator, Type 1310-A Oscillator, Type 1003 Standard-Signal Generator.

Detector: External only (not supplied). A heterodyne detector, Type DNT-6, is recommended for use above 3 MHz. A well shielded radio receiver is also satisfactory.

Accessories Supplied: 2 leads of different lengths to connect unknown impedance to bridge terminals; ½-in. spacer and ¾-in. screw to mount component to be measured directly on bridge terminals; 874-R22LA Patch Cord.

Accessories Available: Luggage-type carrying case, 1606-P2 Precision Coaxial Adaptor Kit.

Mounting: Welded aluminum cabinet.

Dimensions (width × height × depth): 12½ × 9½ × 10¼ in. (320 × 245 × 260 mm).

Weight: Net, 23 lb (10.5 kg), with case, 29 lb (13.5 kg); **shipping,** 30 lb (14 kg), with case, 31 lb (14.5 kg).

Specifications for 1606-P2

Capacitance Added: By adaptor to GR900, 0.38 pF at reference plane (less fringing capacitance); by flange adaptor, 0.18 pF.

Weight: Net, 10 oz (270 g); **shipping,** 12 oz (340 g).

Catalog Number	*Description*	*Price in USA*
1606-9702	**1606-B R-F Bridge**	**$1050.00**
1606-9601	**1606-P1 Luggage-Type Carrying Case**	**25.00**
1606-9602	**1606-P2 Precision Coaxial Adaptor Kit**	**95.00**

PRECISION CAPACITANCE MEASUREMENTS WITH A SLOTTED LINE

The precision slotted line, normally considered a microwave instrument, can be used to measure the capacitance of any high-Q capacitor from about one-third its self-resonant frequency to nearly the self-resonant frequency. Such measurements are made at General Radio in connection with the calibration of new coaxial capacitance standards (see page 3). From the basic capacitance measurements, the self-resonant frequency is determined by extrapolation, and the parasitic inductance of the capacitor is easily calculated. Since the departure from the "dc" capacitance reaches 12.5 percent at one-third the resonant frequency, the slotted-line measurement covers a very important frequency range.

Principle of Measurement

This unusual application of the slotted line results from the fact that, for a small capacitive reactance, the minimum position of the slotted-line probe is relatively close to the unknown terminals, in terms of wavelength. In lumped-parameter terms, the inductance of the coaxial slotted section is generally great enough to resonate the capacitor being measured. The slotted-line probe, in the active sense, behaves like an adjustable short-circuiting wiper whose position along the line corresponds to the minimum of the voltage standing-wave pattern. The slotted line thus acts as a precision variable inductor.

At higher frequencies, usually above 50 MHz, this lumped representation is no longer accurate, and distributed transmission-line formulas must be used for capacitance determination.

Procedure

The use of a GR900® precision coaxial connector is highly recommended for maximum accuracy and is essential if the measurement is to be made on a Type 900-LB Precision Slotted Line. The Type 1406 Coaxial Capacitance Standards are equipped with GR900 connectors and can thus be directly connected to the precision slotted line for high-frequency calibration. For capacitors not so equipped, the Type 900-Q9 Adaptor (see page 5) offers a convenient binding-post connection.

The measured values are slotted-line minimum position and exact frequency. The capacitance is calculated

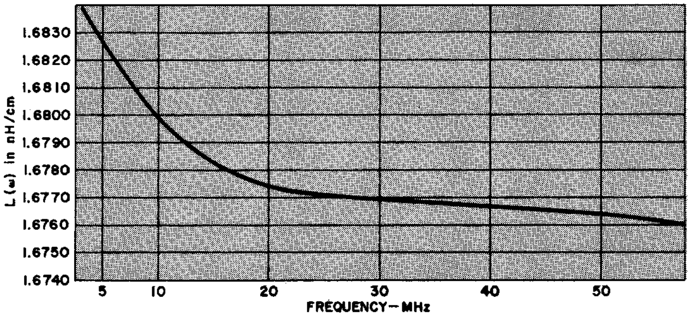


Figure 1. Typical inductance per unit length for Type 900-LB Precision Slotted Line.

from: $$C_x = \frac{1}{\omega^2 L(\omega) \ell}$$

where ω = angular frequency ($2\pi f$)
$L(\omega)$ = slotted-line inductance per unit length
ℓ = physical length from unknown reference plane to slotted-line minimum.

The inductance per unit length of the slotted line is not easily calculable because the conductors are usually made of one material and plated with another. Skin-effect calculations are not reliable because the plating thickness is not known accurately enough. This inductance has been measured for a number of TYPE 900-LB Precision Slotted Lines; the results are shown in Figure 1, which may be used to determine $L(\omega)$. These measurements were made by means of a solid-silver TYPE 900-LZ Reference Air Line, whose characteristics are determined on an absolute basis from its dimensions and known values of conductivity. The dimensions are known to within about ±0.01%; conductivity is known to within about 10%. A high degree of accuracy for conductivity is not required in this calibration; the skin-effect correction is, for example, less than 1% above 1 MHz, and the resulting error in the skin-effect correction would be only about 0.05% in this case.

The physical length from the unknown reference plane to the slotted-line minimum position can, with care, be measured to an accuracy of about 0.05 percent. A final measurement accuracy of about 0.3 percent is possible, taking into account the variation of $L(\omega)$ from one slotted line to another.

The residual inductance of the capacitor can be determined from the measured data, if a low-frequency capacitance measurement is performed. The TYPE 1615-A Capacitance Bridge can be used to measure capacitance (C_o) at 1 kHz to ±0.01 percent. Then the inductance (L_o) required to resonate C_o at the high measurement frequency is determined from

$$L_o = \frac{1}{\omega^2 C_o}.$$

The inductance that actually resonated the capacitor at the high frequency ($L\ell$) is then subtracted from L_o to give the residual inductance of the capacitor.

— J. ZORZY
M. J. MCKEE

"We do not wish to belabor the point, but do want to make a convincing case for using precision connectors wherever accurate and precise immittance measurements are required. . . . The spectacular increase in accuracy of lumped immittance measurements in the past few years has resulted almost entirely from improved standards, instruments, and techniques made possible by precision connectors. This accuracy cannot be realized by the ultimate user of the measurement unless he is willing to accept the small added cost of using precision connectors."

—L. E. Huntley and R. N. Jones, "Lumped Parameter Impedance Measurements," *Proceedings of the IEEE*, Vol 55 No. 6, June 1967.

Figure 1. Type 1163 synthesizer under control of 40-channel program unit. Additional frequencies are available in second 40-channel unit.

A MECHANICAL MEMORY FOR FREQUENCY SYNTHESIZERS

The modern decade-frequency synthesizer can generate precision frequencies defined by many significant figures. This invaluable capability is, unfortunately, accompanied by a minor drawback — the figures must be chosen and set, digit by digit, each time a new frequency is needed.

Whether the frequency is controlled by a series of in-line dials, for good readout, or by a less readable matrix of pushbuttons, manipulation of a series of six or seven or more controls for each desired frequency is a chore, and a mistake in setting is an ever-present danger.

If the synthesizer can be programmed remotely, much of this burden can be transferred to a computer, simple or complex according to the needs of the intended measurements. Less elaborately, when a number of precision frequencies are needed repetitively, storing each of them in a bank of mechanical switches, for instant recall on demand, seems a simple solution to a complex problem. For programmable General Radio synthesizers, this operation is conveniently handled by the new Preset Frequency Program Unit.

The Program Unit has two major parts: an "active housing," cabled to the synthesizer to be controlled, and a "program tray," in which desired frequencies can be set in a convenient array of linear digit switches. A program tray plugged into the housing automatically takes control of the synthesizer. The operator can quickly recover manual control of any or all digits at any time merely by moving the related synthesizer digit dials out of the R (remote) position.

Trays are available with 20 and 40 channels. Of course, any number of trays can be programmed and ready for use when needed. Any tray plugs into the housing in seconds and goes to work instantly.

Figure 1 shows a TYPE 1163 Synthesizer controlled by a program unit. Figure 2 shows the interior of a 40-channel tray, as it appears when it is withdrawn from the housing. In this tray, a matrix of 280 10-position slide-switches is arranged in seven columns and 40 rows. Each row controls a frequency channel. To program a channel frequency, one sets the seven switches in the row to the desired digits by sliding the switch contactors to numbered (0–9) positions on a fiducial strip. A seven-digit frequency can thus be "read into memory" in seconds. Since the trays plug in and out with no permanent wiring connected, a tray can be withdrawn, carried to any convenient location for setting up a program, and restored to use with a minimum of bother.

The two rectangular handles (occupying most of the front of each tray) serve a double purpose: they provide enough leverage to engage or to disengage the multiterminal connectors between tray and housing, and they also carry (under a protective transparent cover) removable cards on which the frequency program set in the tray can be typed or written for instant reference from the front.

In use, any preset channel frequency is selected instantly by the grounding of the appropriate channel line, through a single contact or a transistor switch. Since arrangements for making such

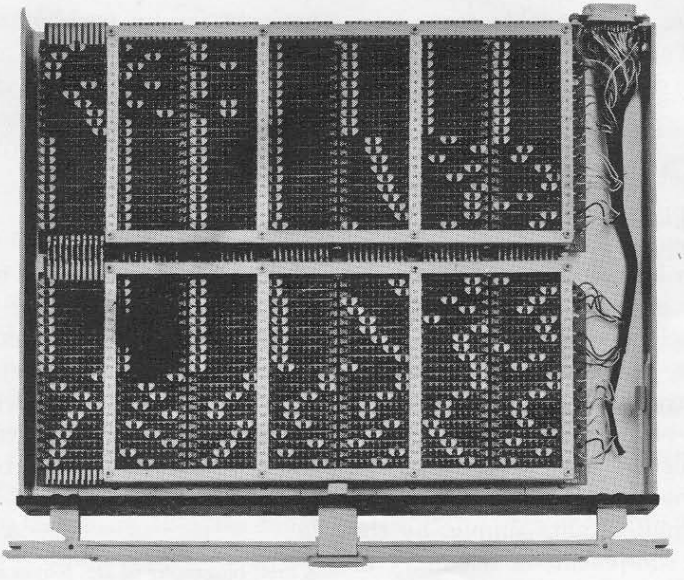


Figure 2. Interior of 40-channel program unit.

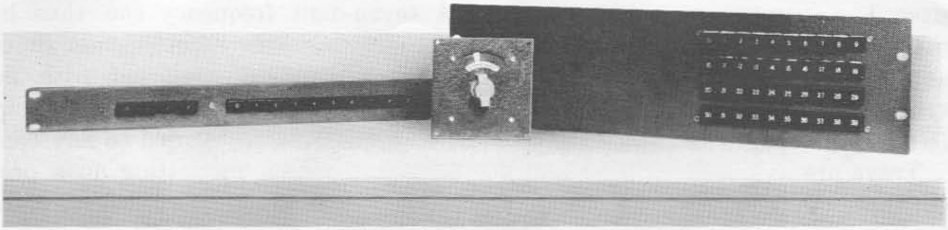


Figure 3. Three experimental program-unit switching assemblies.

contact closures will undoubtedly vary widely depending on user requirements, no standard device for performing this function is presently offered. In the system of Figure 1, a 40-pushbutton switching device controls the selection of any of the 40 channels.

Figure 3 illustrates other pushbuttons and switching devices that have been constructed experimentally. These are not offered as standard products, but control units similar to these can be designed and supplied on special order.

Each digit insertion unit is connected to the Program Unit by a 12-wire cable carrying control and supply lines. These cables are available separately, so one need order only as many of them as there are digits to be programmed.

Applications

The Preset Frequency Program Unit will find application whenever two or more different precision frequencies are expected to be used repetitively, as in production alignment and testing of frequency-selective apparatus such as comb-filters, radio receivers, or transmitters. Physicists engaged in nuclear or atomic resonance research should also find this device very useful.

Ephemeris-predictable Doppler shifts could be handled quite simply, by the use of this equipment, in telemetry or communications systems.

Time-sharing of synthesizers, with suitable programming of output switching, will probably offer important economic advantages in many systems.

Two synthesizers can be simultaneously programmed to different groups of frequencies by two trays controlled by a single set of contact closures. Thus, for instance, simultaneous dual outputs with constant frequency separation can readily be achieved. Such setups are useful in heterodyne measuring systems.

External selection apparatus can, of course, be arranged to step through a series of frequencies in a definite sequence at a controlled rate or, alternatively (as by pushbuttons), to provide random access to any frequency stored. The trays can thus become part of highly automated systems, or they can be used more simply to aid manual testing or research.

Conclusion

The frequency program unit is simple in concept and easy to use; applications are expected to be extremely varied, conceived and executed by engineers and other scientists working in the many fields in which frequency synthesizers are so rapidly becoming indispensable.

— A. Noyes, Jr.

A brief biography of Dr. Noyes appeared in the January 1967 *Experimenter*.

SPECIFICATIONS

Capacity: Stores 20 or 40 preset 7-digit frequencies (depending upon model).

Frequency-Selection Input: Mechanical or solid-state switch closure to ground required for each channel. Each closure must be capable of carrying 70 mA with < 0.5-V drop.

Frequency-Selection Output: Circuit closure on 10-line connection to each controlled digit.

Switching Time: <2 ms, depends only on speed of programmable modules.

Accessory Supplied: Connector assembly (two 24-pin connectors) for connection of external selector switches to 1160-P1.

Accessories Required: Cables from 1160-P1 to synthesizer, one per controlled digit: 2-foot length is standard; select to suit RDI modules to be controlled.

Accessories Available: One empty instrument cabinet, convenient in interchanging trays in systems using more than one tray.

Mounting: Relay-rack cabinet.

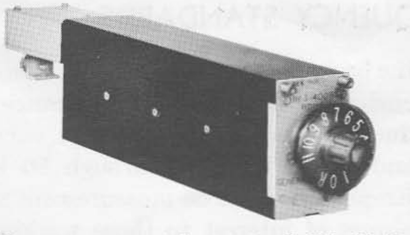
Dimensions (width × height × depth): 19 × 1¾ × 15 in. (485 × 45 × 385 mm).

Weight: 20 channel, net 9 lb (4.1 kg), shipping 20 lb (9.5 kg); 40 channel, net 11 lb (5 kg), shipping 22 lb (10 kg).

Catalog Number	*Description*	*Price in USA*
	1160-P1 Preset Frequency Program Unit	
1160-9620	20 channels	**$1000.05**
1160-9640	40 channels	**1825.00**
1160-9701	**Cable** (2-ft) to any programmable module (1160-RDI-1) up to 100 kHz/step	**75.00**
1160-9702	**Cable** (2-ft) to 1 MHz/step programmable module (1164-RDI-2) in 1164 models	**75.00**
1160-9704	**Cable** (2-ft) to 1 MHz/step programmable module (1163-RDI-4) in 1163 models	**75.00**
1160-9500	**Cabinet**, empty, to store 1160-P1 tray	**50.00**

GR Product Notes

MORE PROGRAMMABLE PLUG-INS FOR SYNTHESIZERS



Two new remotely controllable digit-insertion units extend the advantages of programmability to all digit stations of all GR 1160-series frequency synthesizers, with the single exception of the top digit of the Type 1164.

The 1163-RDI-4 is a plug-in replacement for the DI-4 (× 1 MHz) unit in a Type 1163 synthesizer. This means that the entire 12-MHz range of the 1163 is now programmable.

The Type 1164-RDI-2 is a programmable plug-in for the × 1 MHz station of an 1164 synthesizer.

These two new programmable modules, like the lower-frequency RDI-1 units, are directly interchangeable with their nonprogrammable counterparts, and converting any station for programmability is just a few seconds' work.

Catalog Number	*Description*	*Price in USA*
	Programmable Digit-Insertion Units	
1163-9479	1163-RDI-4, 1 MHz/ step, in 1163 models	**$575.00**
1164-9479	1164-RDI-2, 1 MHz/ step, in 1164 models	**555.00**
1160-9650	**Hook-Up Cable** for all RDI's, 50 ft, 12 conductor, shielded	**15.00**

the GR Experimenter

GENERAL RADIO COMPANY
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SEMINAR ON LOW-FREQUENCY STANDARDS

A seminar on Low-Frequency Electrical Standards will be conducted by the National Bureau of Standards at the Bureau's new site at Gaithersburg, Md., from December 11 to 13, 1967.

The seminar will present information on the accurate measurement of electrical quantities and on the calibration of electrical standards. It will cover the measurement methods used by the Bureau to establish and to maintain the basic electrical units and to calibrate customers' standards of resistance, inductance, capacitance, voltage, current, and power from dc through 30 kHz. Emphasis will be on measurement techniques of interest to those working in standards and calibration laboratories.

Those wishing further information should write to R. F. Dziuba, B-162, Metrology Building, National Bureau of Standards, Washington, D. C. 20234.

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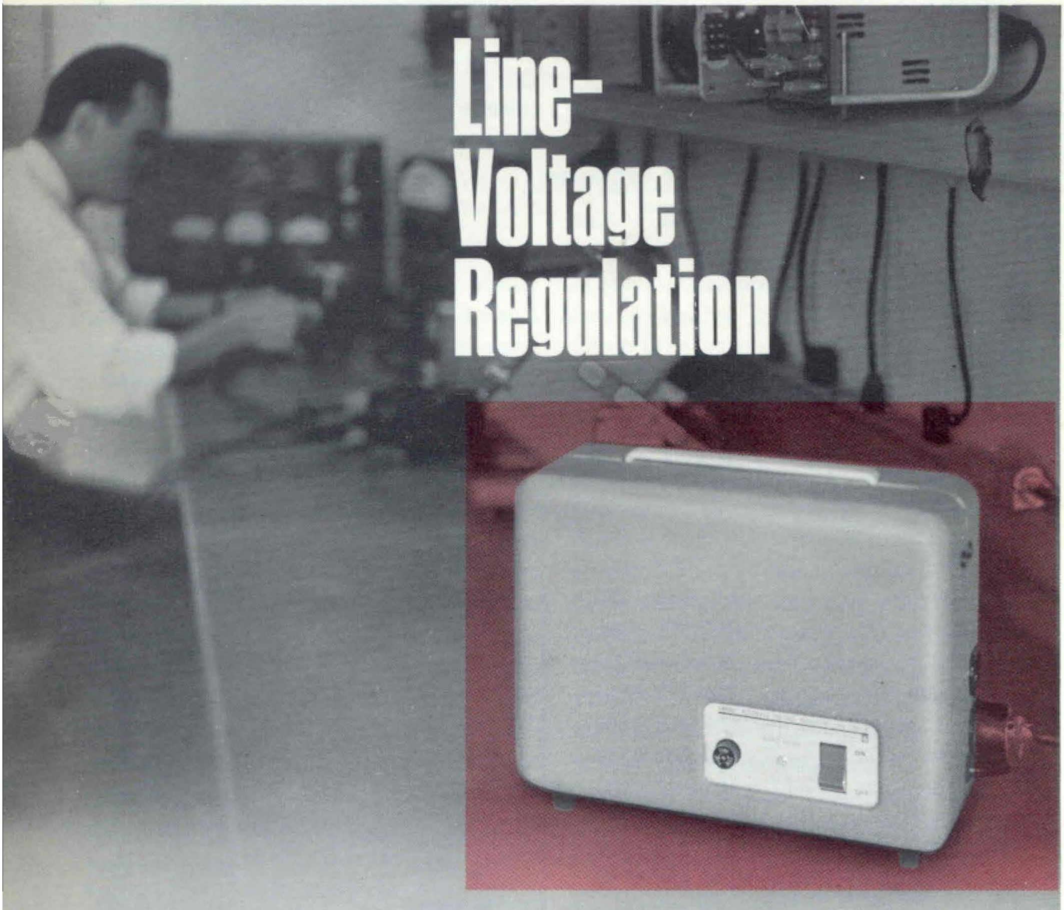

Line-Voltage Regulation

- CONSIDERATIONS IN CHOOSING A VOLTAGE REGULATOR
- AN INEXPENSIVE 1-kVA REGULATOR
- A NEW THREE-PHASE REGULATOR

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THIS ISSUE

CONSIDERATIONS IN THE CHOICE OF A LINE-VOLTAGE REGULATOR

A line-voltage regulator seems a simple enough device, designed to perform a simple task. The temptation is to compare different models merely by looking at one or two of the more obvious specifications. The "simple task," however, is complicated by a number of factors, especially including the characteristics of the load connected to the regulator output. The following article discusses some of the more subtle aspects of this widely used and widely needed family of instruments.

It is widely appreciated that line voltage, as delivered by the public utility, varies in amplitude by at least a few percent (the most widely followed standard is 120 volts $\pm 5\%$) and that much wider swings are likely when the line is subject to wide load variations. It is also fairly well understood that line voltage that is too high or too low — even by as little as a few volts — can decrease the efficiency or shorten the life of many electric devices and can introduce significant errors in measurements made with electrical instruments. For these reasons, many laboratories and manufacturers use automatic line-voltage regulators to hold the line voltage to within a few tenths of a volt or better of a nominal value.

At first glance, the criteria for a voltage regulator seem to be simple, chiefly including the speed with which the regulator can act to restore voltage to a nominal value and the accuracy with which it re-establishes this nominal value. Actually, there is a great deal more involved, and the engineer looking for a line-voltage regulator must be willing to read all the specifications and to assess their impact on one another and on his problem.

Distortion

Take, for instance, the matter of distortion. To appreciate the significance of distortion in a voltage regulator, one must keep in mind that a line-voltage regulator must be designed to detect and regulate a specific characteristic of ac voltage. (The root-mean-square characteristic is usually chosen as the best compromise.) That value may or may not correspond to the needs of the devices operated from the regulator. A lightly loaded capacitive-input dc power supply responds to peak voltage; thermal devices respond to rms; heavily loaded capacitive-input as well as inductive-input power supplies and mechanical systems generally respond to the average value of voltage. A relay rack full of instruments or even a single instrument may well include devices or power supplies with all three response characteristics. The less distortion introduced by the regulator, the better the regulator's ability to track all three characteristics of the voltage while regulating only one.

Thus, the stated accuracy of a regulator must be interpreted in the light of the distortion specification. A $\pm 0.1\%$ accuracy statement usually means that the regulator will hold the rms voltage within $\pm 0.1\%$ of nominal. But if this specification is accompanied by a 3% distortion figure, peak-responding devices operating from the regulator may be faced with as much as a 3% change in input voltage, and an average-responding device may encounter as much as a 1% variation, or 30 or 10 times the maximum error implied by the specification.

The electromechanical regulator, utilizing the continuously adjustable autotransformer, introduces no distortion; in contrast, all magnetic regulators available at present introduce distortion in the process of regulating. Furthermore, the distortion components generated are often at high frequencies, i.e. some portion of the output waveform has a fast rise time, which may raise havoc with some digital equipment.

Overshoot

The response characteristics of voltage regulators frequently display some overshoot, the amount being a function of the degree of damping in the regulator. Although one's instincts might suggest that the optimum amount of overshoot is no overshoot at all, some overshoot usually exists as a byproduct of a fast response characteristic. Moreover, the effect of overshoot can be positive — as, for instance, when the overshoot compensates in part for the effects of a voltage transient on the devices powered from the regulator. There is a delicate balance here; too much overshoot can initiate oscillation and mean an intolerable delay in returning voltage to its nominal value.

Overshoot introduces a problem in the specifying of response time. When can a voltage transient be said to be corrected — when the voltage first comes within specified accuracy limits (even though overshoot subsequently takes it outside those limits) or when the voltage is within the accuracy limits to stay? Common practice ignores the overshoot, but the canny buyer will do otherwise.

kVA Ratings and Overloads

Load rating is another innocent-looking specification that can be misleading. Not that a 1-kVA regulator won't handle 1 kVA; it almost certainly will, but starting surges can momentarily overload a regulator that is otherwise adequately rated. Electromechanical regulators using autotransformers can readily withstand 1000-percent short-term overloads. But magnetic units typically use fast-blow fuses or disabling circuits to protect components from damage, and such fast-acting protective devices can prevent measurements of start-up times of servo or induction motors, high-powered, keyed, or pulsed equipment, etc. In tests at GR, a 1-kVA, 120-volt magnetic regulator, which one would normally expect to control a kilowatt comfortably, blew fuses repeatedly when a 1000-watt incandescent bulb was connected to its output.

High starting surges must be expected in the turn-on of incandescent lamps, induction motors, and transformer-equipped devices.

Response Speed

The time required for a regulator to restore proper output voltage after a line or load disturbance is another important characteristic. Since the correction proceeds essentially exponentially in a magnetic regulator and linearly in an electromechanical regulator, each type expresses response speed or correction time differently. For electromechanical regulators, response speed or correction time has traditionally been stated in volts per second or seconds per volt, respectively, with the given values based on the full slew speed of the servo system. Of course, the system requires some small amount of time to attain this speed, and a more accurate expression for correction time would therefore be in terms of a fixed time plus a time proportional to the magnitude of the correction. We have adopted this method of expression in the specifications for the line-voltage regulators introduced later in this issue.

For a magnetic regulator, the time constant is specified — i.e., the time taken to correct 63 percent of the error. Usually from two to four time constants are required to correct an input disturbance to within the specified accuracy.

How can these two different expressions be compared? Obviously, one must know the magnitude of the input disturbance before any direct comparison can be made. In general, for the usually encountered small but sudden voltage steps superposed on larger but slower voltage excursions, the response speeds of the electromechanical and magnetic regulators are approximately the same. For the less usual case of 10- to 20-percent instantaneous line-voltage steps, neither regulator is very fast, but the magnetic type is definitely the faster of the two.

Load Effects

Common conception has the line-voltage regulator accepting varying voltage at its input and presenting a constant voltage at its output. Changes in load or power factor, as well as the surge effects discussed earlier, can easily create voltage changes as great as those on the unregulated power line. All the magnetic regulators tested at GR were significantly affected by moderate load changes, resulting in a recovery time comparable to that required to correct a large input transient even though the input voltage remained constant. Furthermore, the waveform was modified, changing the peak and average values of output voltage even though the rms level was maintained.

The electromechanical regulator, on the other hand, is little affected by load changes, because of the low output impedance and close coupling between input and output of the autotransformer.

Line-Frequency Effects

The power company generally holds its line frequency very close to the nominal value, and what slight deviations there are do not normally affect regulator performance. In the field, however, a portable generator may well be *off* 50 or 60 Hz more often than it is on, and therein lies a major source of distortion and poor voltage regulation. Line-frequency shifts pose a particular problem for the magnetic regulator, and manufacturers of these caution that the stated performance specifications apply only *at* 50 or 60 Hz. A deviation of 5 percent in frequency increases the distortion to typically 5 percent, a 10-percent deviation to the 8- to 10-percent level. The electromechanical regulator, on the other hand, is essentially unaffected by line-frequency variations, even of the degree encountered with portable generators.

Efficiency

Since a voltage regulator is often a continuous-duty device, efficiency is an important factor, both from the point of view of thermal problems associated with relay racks full of equipment and from that of the increased physical size and weight necessary to dissipate large amounts of power. Small size and weight are, of course, also important for their own sake, especially where the regulator is expected to see portable or field service. For a given power level, the electromechanical regulator is now available at less than one third the weight and at significantly lower cost than comparable magnetic regulators.

Summary

The following table summarizes the important characteristics of the two important classes of line-voltage regulator. A third type, the fully electronic regulator (oscillator-power amplifier) potentially has the highest performance of all but at present is not commercially competitive with the other types.

The table indicates the relative merits of the electromechanical and magnetic regulators.

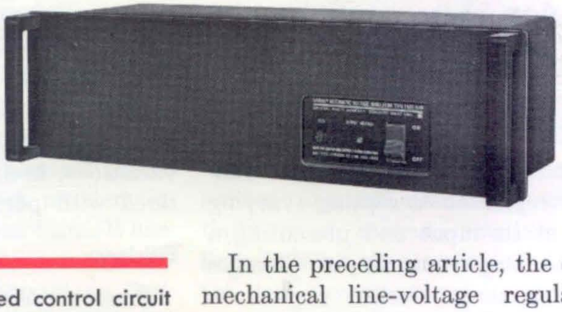
TABLE 1.

Characteristics of Voltage Regulators

	Magnetic	*Electro-mechanical*
ACCURACY	excellent	excellent
RESPONSE SPEED	excellent	excellent
kVA	wide range available	wide range available
DISTORTION	moderate	none added
EFFICIENCY	moderate to high	highest
SENSITIVITY TO LINE FREQUENCY	high	none
SENSITIVITY TO POWER FACTOR	low to high	none
WEIGHT	high	low
COST/kVA	moderate	low

C. Chitouras

A NEW 1-kVA LINE-VOLTAGE REGULATOR



Type 1591-A Automatic Line-Voltage Regulator, available in portable (top) and relay-rack (bottom) versions.

A new, greatly simplified control circuit has allowed us to extend the advantages of the electromechanical line-voltage regulator to the low-power range. The 1-kVA regulator described below (and shown on this month's cover) is a small, lightweight, inexpensive unit scaled to serve the average bench or relay rack.

In the preceding article, the electromechanical line-voltage regulator is seen to have significant advantages over the magnetic regulator, particularly with respect to distortion (hence, ability to maintain the peak and average, along with the rms, levels of voltage) efficiency, insensitivity to line-frequency and power-factor variations, size and weight. For over a decade GR has offered 6-kVA and larger regulators with the above advantages. However, all attempts to offer lower-powered units at commensurate prices have been frustrated by the fact that the manufacturing cost of the control circuitry was essentially independent of the power-handling rating of the regulator; there was thus no incentive for the customer to buy or therefore for GR to offer a low-power unit. Nevertheless, the need for small regulators has increased with time, spurred by the relatively low power requirements of solid-state designs and the increased use of groups of instruments in relay racks. On the performance side, we have seen increased demands for higher levels of accuracy in measurement and control and for the highest reliability consistent with reasonable cost. It is surprising that, in spite of the elementary calculation involved, few realize that a line-voltage change from — 10% to + 10% of nominal represents nearly a 50% increase in power dissipation — a first-order effect on reliability.

THE GR TYPE 1591 SERIES OF REGULATORS

As a result of the development of a new, greatly simplified control circuit, it is possible for the first time to produce an inexpensive 1-kVA electromechanical regulator. This new regulator, the portable version of which weighs only 17 pounds and is priced under $300*, is GR's Type 1591.

Four models are available to cover 115- and 230-volt service in both portable and relay-rack packages. The only difference in ratings is that the 230-volt models are rated at 0.8 kVA, resulting from the 20% derating of

* Price applies in U.S.A. only.

Costa Chitouras received his BSEE and MSEE degrees from Massachusetts Institute of Technology in 1955. After service with the U.S. Army Signal Corps Engineering Laboratories, he joined General Radio in 1956. A development engineer in GR's Industrial Group, he has worked on stroboscopes, recorders, frequency meters, and voltage regulators.

autotransformers in going from 115- to 230-volt operation. All models have identical control circuits.

Operating Characteristics

The 1591 will maintain the output voltage at 115 volts (adjustable from 105 to 125 volts) with an accuracy of ±0.2 percent for simultaneous input-line variations from 100 to 130 volts, load variations from no load to full load, power-factor variations from 1.0 to 0 leading or lagging, and line-frequency variations of ±10 percent. Correction takes place within 6 cycles +1.5 cycles per volt.

Output voltage of the 230-volt units is adjustable from 210 to 250 volts and input-line variations from 200 to 260 volts are corrected when the output voltage is set to 230 volts. Correction takes place within 6 cycles +0.7 cycle per volt.

Figure 1 shows recordings of the output voltage of four different 1591's under various combinations of input voltage and load.

Behind the recordings of Figure 1 lies an interesting story. Over a year ago 100 models of the 1591 were constructed. Of the many tests conducted, perhaps the most important was a one-year round-the-clock life test of 40 of these units. There were no failures. From this group of 40, four units were randomly selected and put through an accelerated life test, in which the 115-volt line was modulated at a 3.5-hertz rate, subjecting motor-gear train, Variac® autotransformer, and control circuitry to well over a quarter of a million oscillations per day, while operating at nearly full load rating. As of this writing, the 10-million-cycle mark is at hand. There has been no lubrication or adjustment (the only adjustment possible — internal or external — is the front-panel voltage control). Two of these four instruments, again randomly selected, were used as two of the instruments for the tests in Figure 1. The other two instruments were new. Can you tell the difference?

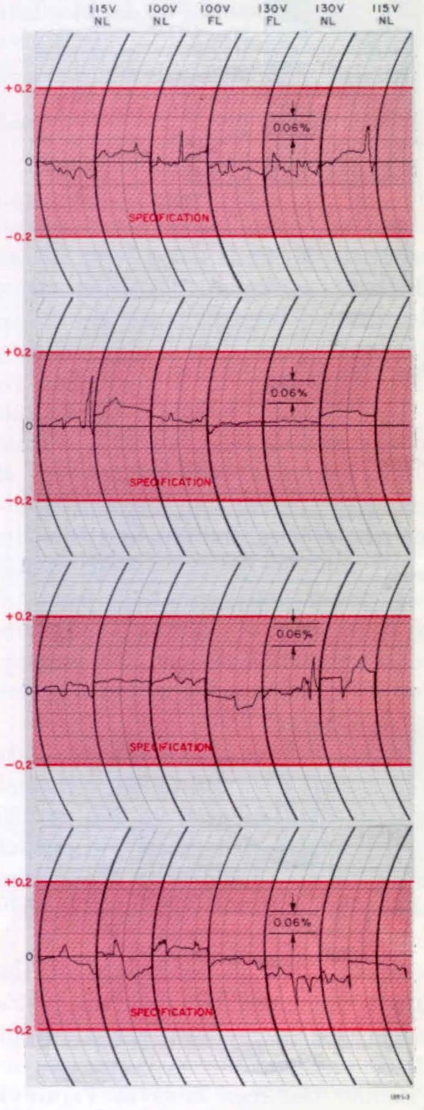


Figure 1. Recordings showing the performance of four different 1591 regulators under changing line and full-load and no-load conditions. Two of the regulators (top and bottom) were new; the other two had been put through a one-year round-the-clock life test.

The two relay-rack versions of the 1591, in addition to meeting the usual GR instruments standards for humidity, storage and operating temperature, etc, passed a test in which they were vibrated from 10 to 55 Hz in one-minute sweeps for 15 minutes in all three planes and at a 30-mil peak-to-peak amplitude. They also passed the Air-Force bench-drop test and the standard 30-g, 11-ms shock test, having been subjected to this three times in six different directions. These tests were conducted while the 1591's were both operating and nonoperating.

The ability of the electromechanical regulator in general and the 1591 specifically to track the average and peak values while actually detecting the rms value is illustrated in Figure 2a. (The resolution of all three voltmeters was 0.1 percent; the band therefore indicates the minimum detectable limits or resolution.) As a comparison, similar plots of two of the best currently available magnetic regulators are shown in Figure 2b and 2c. Figure 2d, 2e, and 2 show the corresponding output waveshapes of these regulators.

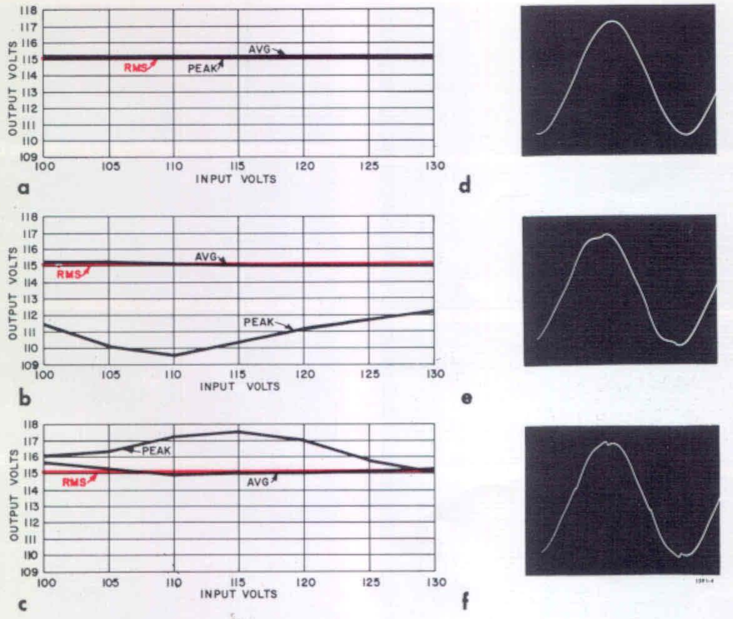


Figure 2. Curves showing the ability (a) of the 1591 and (b, c) of two new magnetic regulators to track peak and average voltage as well as rms. Distortion introduced by the three regulators is seen in the corresponding waveforms at the right (d, e, f).

Figures 3a, 3b, and 3c compare the peak, rms, and average output levels for these three regulators with constant input voltage but varying load current. Varying load current introduces distortion in the magnetic regulators but not in the 1591's. As suggested in the preceding article, there are few instances in which a 1-kVA load, whether it consists of one or many instruments, requires regulation of rms voltage level to within ±0.1 percent and yet can tolerate up to 1- or 3-percent level changes in average and peak voltage.

Speed of Response

The speed with which the 1591 corrects large voltage transients is shown in Figure 4a, an oscilloscope recording of the positive peaks of the output voltage (rms readings at these correction

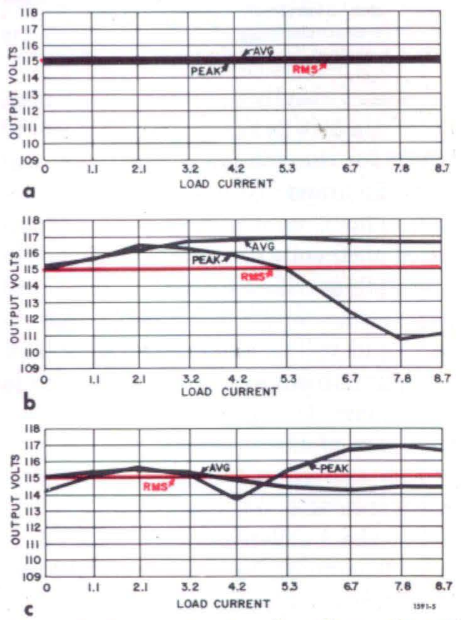


Figure 3. Curves showing the effects of load-current changes on output voltage for (a) the 1591 and (b, c) two magnetic regulators.

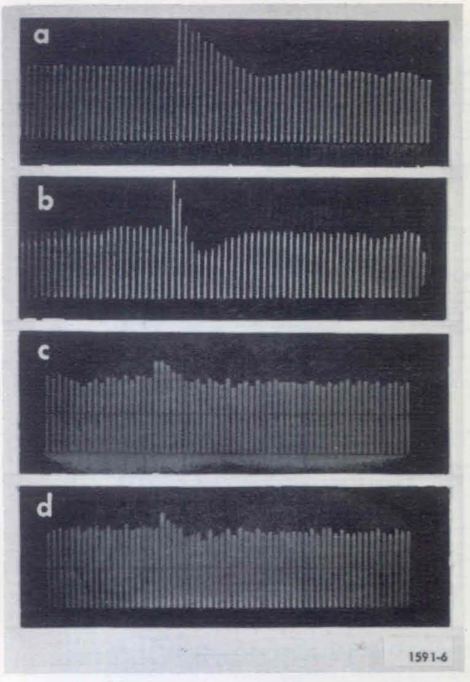


Figure 4. In a and b, we see the response of the 1591 and of a magnetic regulator, respectively, to a 10-V transient. The magnetic regulator corrects faster, but overshoot appears excessive. The response characteristics of the same regulators to a 2-V transient (c and d) show little difference.

rates are virtually impossible to obtain). Note that it takes approximately 10 cycles for the voltage to reach nominal level, followed by a small overcorrection. This is equivalent to a correction speed of 60 volts per second. The correction rate specified for the 1591's takes into account worst-case conditions, such as output voltage set to lower limit, etc.

Figure 4b shows the same test made on a magnetic regulator. Here we see the forte of the magnetic regulator — a high correction rate for large input-line transients.

For the smaller input-line transients more commonly encountered, the results are often different, as shown in Figures 4c and 4d. In this case, the input voltage changed by 2 volts; the response characteristics of the two regulators are now seen to be comparable.

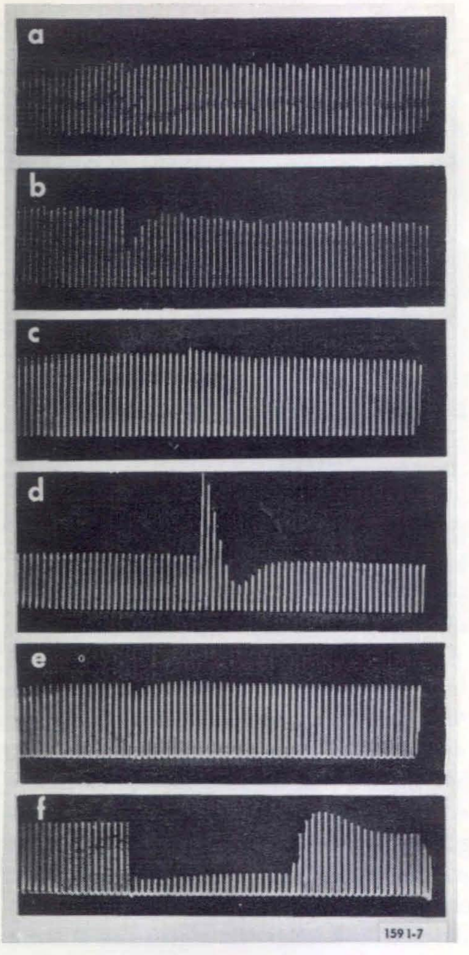


Figure 5. Oscillograms showing the performance of a 1591 (a, c, e) and of a magnetic regulator (b, d, f) in the face of changing load conditions. In a and b, load current was changed from 5 to 6 amperes. In c and d, half the load current was removed. In e and f, a 600-watt incandescent light was connected. Note that the magnetic regulator's protective circuit was triggered by the starting surge of the 600-watt load, even though the regulator was rated at 1 kVA.

Load Effects

The extremely stable characteristics of the 1591 under varying load conditions are illustrated in Figure 5.

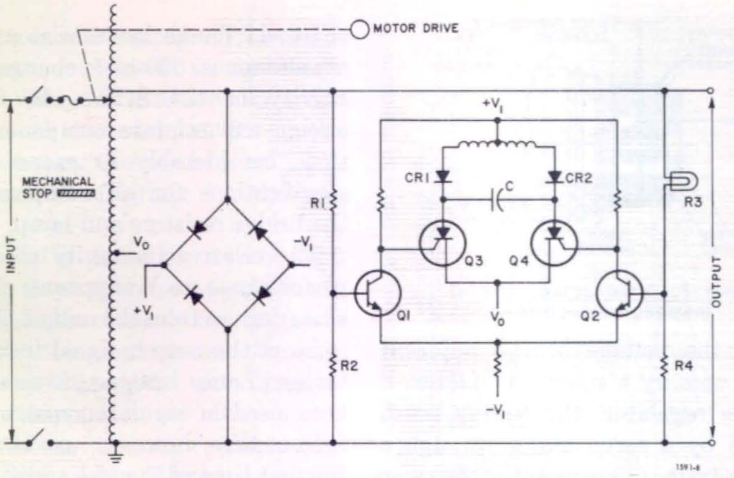


Figure 6. Elementary schematic diagram of the Type 1591 Automatic Line-Voltage Regulator.

Figure 5a shows the output voltage change when the load current on the 1591 is changed by 1 ampere. Figure 5b illustrates a magnetic regulator's performance under the same test. Note that the input line voltage was held constant. Figures 5c and 5d show the output-voltage variation for the 1591 and for the magnetic regulator respectively, when one half of the maximum load current is removed, simulating the turning off of some of the instruments powered by one regulator. Figures 5e and 5f illustrate the performance when a 600-watt incandescent load was applied to both regulators. The protective circuit of the magnetic regulator reduced the output voltage to some low value for approximately one half a second before turning on again. A 1000-watt incandescent load could not consistently be turned on with this regulator, blowing a fuse approximately every third attempt. A 1591 was tested using a 1300-watt lamp load with an on-off cycle of 5 seconds. The test was concluded many cycles later, with not a single interruption of service.

Warmup Drift and Temperature Coefficient

The warmup drift of the 1591, in spite of the fact that it uses a thermal device as a reference, is negligible — typically less than 0.2 percent from a cold start, with 0.1 percent occurring in the first 5 to 10 minutes.

The temperature coefficient of the instrument is specified at less than 0.01%/°C and is typically 50 ppm/°C.

How It Works

Figure 6 is an elementary diagram of the 1591 voltage regulator. To convert the unregulated line voltage to a regulated output, the high side of the input line is connected to the brush in the Variac autotransformer. By precise control of the brush position on the winding, the volts-per-turn and consequently the output voltage are held constant. Figure 1 has indicated that the resolution of the 1591 is often within a few hundredths of a percent, and yet the Variac has an apparent resolution of three quarters of a volt per turn. The brush of a Variac acts as a fine voltage divider, capable of sub-

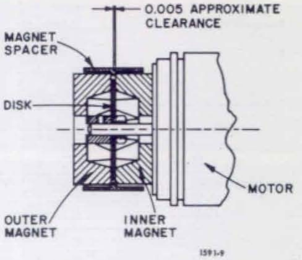


Figure 7. Cross-section view of the eddy-current brake.

dividing the voltage of two adjacent turns of wire by a significant factor.

In this regulator, the Variac brush is driven by a servo motor through a 90:1 gear train. (This motor is the same as that used on the GR 1571 militarized voltage regulator.) To stabilize the servo system, some damping must be introduced. This function is served in the 1591 by an eddy-current brake.

The eddy-current brake (Figure 7) consists of a high-conductance aluminum disk and two magnets. The disk is attached to the motor rotor and revolves between the magnets. At slow motor speeds (minor voltage corrections), no significant hold-back torque is generated by the brake, and full torque is available for brush positioning. When larger corrections (faster motor speed) are required, the brake generates considerable hold-back torque. This torque, which is proportional to velocity, provides the damping necessary to stabilize the system.

The servo motor is driven by a control circuit consisting of a bridge detector (R1 through R4), a differential amplifier (Q1–Q2), and an SCR output stage (Q3–Q4); a total of two transistors and two SCR's form the total active-component complement of this instrument. In spite of the simplicity of this circuit, there is no detectable change in the performance of the 1591 even with a 5-to-1 change in transistor β and a simultaneous 500-to-1 change in sensitivity of the SCR's. In fact, the circuit will tolerate component variations considerably in excess of their specifications for all components but the bridge resistors and lamp.

The relative immunity of regulator performance to component characteristics derives from the unique characteristics of the output signal from a lamp bridge. Lamp bridges have, of course, been used in regulator control circuits before; here, however, use is made for the first time of the two additional zero crossings resulting from the various frequency components generated in a lamp circuit. These components — the 60-Hz bridge voltage, a 90° out-of-phase component, and a third-harmonic component of the line frequency — combine to produce a waveform similar to that shown in Figure 8a. This signal, after amplification, appears as Figure 8b. Note that the positions of the zero crossings are independent of amplifier gain. The amplified signal triggers the two SCR's, whose output waveform is shown in Figure 8c. Note that, while the power applied to the motor remains constant, the direction of rotation and torque are functions only of the phase of the control voltage. The phase is in turn a function of the positions of the zero crossings. At unbalance, the 60-Hz correction signal from the bridge serves to move the zero crossings over a range of about 180° for a 1% change in applied bridge voltage. The result is proportional control of torque, independent of amplifier gain and of characteristics of all components except the lamp and the bridge resistors.

The power-supply voltages are derived from full-wave rectification of the ac voltage at the output of the auto-transformer, referenced to its center tap. The Variac is used "backwards" (input line on the brush, output on the tap), eliminating the need for an additional power transformer for the amplifier.

The use of a pulsating, rather than a filtered, dc supply for the SCR's is required to achieve maximum torque in the motor at the rated temperature rise. (Remember that only the line-frequency component in the motor produces torque; the other components resulting from using filtered dc would be dissipated as heat.)

By isolating the control winding from the SCR commutating capacitor by means of rectifiers CR1 and CR2 (see Figure 6), we ensure that there is always available across the capacitor enough voltage (essentially the peak value of supply voltage) to turn off an SCR, independently of the effects of the inductive servo motor. This technique allows us to enjoy both the efficiency of using pulsating dc for the motor and the reliability of using filtered dc for commutation.

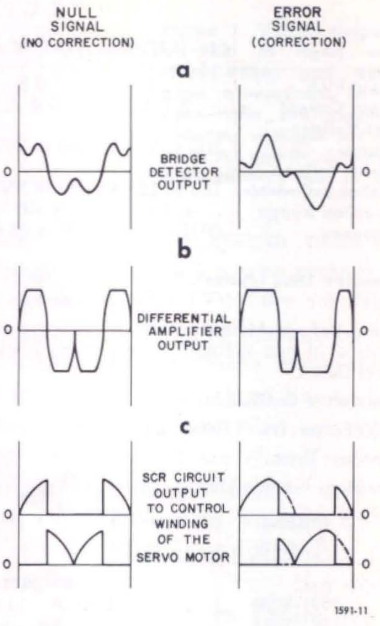


Figure 8. Waveforms present at the bridge detector, differential amplifier, and servo motor for no-error and error conditions. Note that servo action is based on position of zero crossings rather than on amplifier gain.

The Lamp

The lamp used in the 1591 is a GR proprietary design. It exhibits remarkable short- and long-term stability characteristics, brought about by close control of the metallurgical and structural design of the filament and its environment. Its ruggedness is attested by the aforementioned environmental tests that the 1591 has passed. The lamp has a typical temperature coefficient of 20 ppm/°C. It is designed to have a "lifetime" of a few centuries, based on the 10% filament evaporation point — a standard end-of-life rating for incandescent lamps. To test long-term stability, one of these lamps was used as the reference in a dc power supply whose output was compared against a standard cell. The total drift in resistance of the lamp was less than 400 ppm over a three-year period, or an average drift of less than ½ ppm per day.

While it is too costly to measure and guarantee the long-term stability of the 1591's output voltage, our experience indicates that, after a few months of use, the drift of the typical 1591 will be unmeasurable with the usual laboratory ac voltmeters.

C. Chitouras

SPECIFICATIONS

	115-V Models	*230-V Models*
Output kVA	1.0	0.8
Output Current	8.7 A	3.4 A
Input-Voltage Range	100 to 130 V	200 to 260 V
Output-Voltage Range (adjustable)	105 to 125 V	210 to 250 V
Correction Range	± 15 V	± 30 V
Frequency	60 Hz ± 10%	48 to 63 Hz

Correction Time (cycles): 6 c + 1.5 c/V for 115-V models, 6 c + 0.7 c/V for 230-V models.

Output-Voltage Accuracy: ± 0.2% for any combination of line voltage or frequency, load, or power factor.

Temperature Coefficient: <0.01%/°C.

Power Factor: 0 to 1, leading or lagging.

Response: Rms.

Distortion: None added.

Efficiency: 95% at full load.

No-Load Power: Approx 45 W.

ENVIRONMENT

Ambient Temperature (operating): −20 to +52°C, rack model; −20 to +40°C, portable model.

Vibration: Rack model, 30 mils pk-pk at 10 to 55 Hz, three planes, 15 min each plane.

Shock (rack model, operating and nonoperating): AF bench-drop test; 30 g for 11 ms.

GENERAL

Accessories Supplied: Spare fuses.

Dimensions (width x height x depth): Portable, 12¾ × 9½ × 5⅝ in. (325 × 245 × 140 mm); rack, 19 × 5¼ × 6⅜ in. (485 × 135 × 165 mm).

Net Weight: Portable, 17 lb (8 kg); rack, 22 lb (10 kg).

Shipping Weight: Portable, 25 lb (11.5 kg); rack, 31 lb (14.5 kg).

Catalog Number	*Description*	*Price in USA**
	Variac® Automatic Voltage Regulator	
1591-9700	**1591-A,** 115 V, Portable	**$295.00**
1591-9701	**1591-AH,** 230 V, Portable	**320.00**
1591-9712	**1591-AR,** 115 V, Rack	**325.00**
1591-9713	**1591-AHR,** 230 V, Rack	**350.00**

* Quantity discounts available on request.

A NEW THREE-PHASE REGULATOR

With the introduction of the TYPE 1591 Automatic Line-Voltage Regulator (see preceding article), GR can supply regulators rated from 1 kVA (1591) through 20 kVA (1571, 1581, 1582[1]). These instruments are designed primarily to regulate single-phase lines. Obviously, three single-phase regulators can be connected in a wye or closed-delta configuration for use in three-phase systems, and such combinations have been used often, with excellent results. In many installations, however, line voltages and loads are sufficiently well balanced that the use of three separate regulators is more a luxury than a necessity. In such cases, a single control circuit and a single motor could drive

[1] C. E. Miller, "A New Series of High-Performance Line-Voltage Regulators," *General Radio Experimenter*, January 1966.

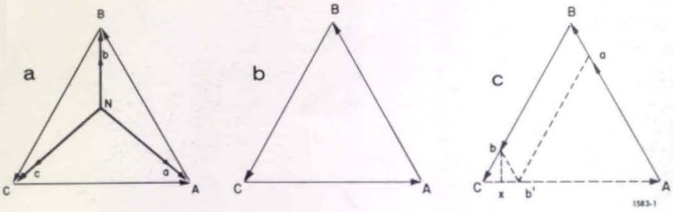


Figure 1. Vector diagrams of (left to right) wye, closed-delta, and open-delta connections. In the open-delta configuration, restoring voltages Aa and Bb to AB and BC automatically restores Ab′ to AC.

an appropriate number of autotransformers. (The autotransformer and a buck-boost transformer are the power-handling elements of a high-powered electromechanical regulator.) The required number of power-handling components is three sets for a wye-connected or closed-delta three-phase system but only two sets from an open-delta system.

The Open Delta

Figure 1 explains vectorially the principles of wye and delta connection for a three-phase line. In the wye connection of Figure 1a, vectors NA, NB, and NC represent the line-to-neutral voltages of a three-phase wye system, and vectors AB, BC, and CA indicate the line-to-line voltages. Assume now that the line-to-neutral voltages drop in amplitude to values indicated by Na, Nb, and Nc. It is clear that restoring Na and Nc to NA and NC is sufficient to correct these two line-to-neutral voltages, along with the line-to-line voltage CA; voltages AB and BC will not be restored, however, until Nb is also corrected to the value NB. Thus all three line-to-neutral voltages of a wye system must be adjusted in order for a wye system to be regulated.

Figure 1b represents the line-to-line voltages of a three-phase closed-delta system, in which all three line-to-line voltages (AB, BC, and CA) are separately corrected. If we choose not to regulate one of the phases and to allow that line-to-line voltage to be determined by the amplitudes of the other two phases, we have the open-delta system shown in Figure 1c. If the nominal voltages are represented by AB, BC, and (by vectorial subtraction) CA, and if these voltages drop to Aa, Bb (which equals ab′), and b′A, it is apparent that if we restore Aa to AB and Bb to BC, b′A will increase to the proper value for CA. (Note that each of the two regulated phases contributes equally to the correction.) Thus only two sets of power-handling components are needed to regulate all phases of a three-phase open-delta system.

The 1583 Regulator

The principles of operation of the 1583 three-phase regulator are similar

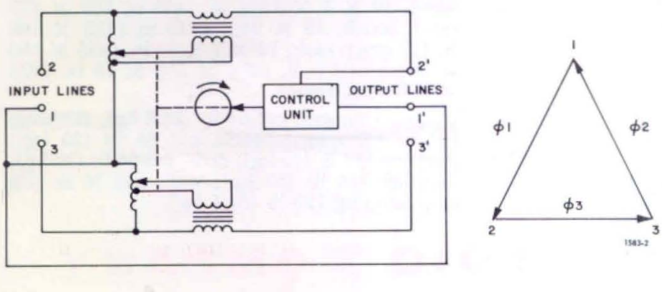


Figure 2. Elementary 1583 Three-Phase Line-Voltage Regulator. ϕ1 and ϕ2 voltages are controlled directly by the two sets of power-handling components, and ϕ3 is automatically corrected as a result.

to those of the single-phase 1581, except for the two sets of power-handling components in the former. Figure 2 is an elementary diagram. In it, two phases ($\phi 1$ and $\phi 2$) are controlled directly by the servo system, and $\phi 3$ is controlled indirectly. A deviation in the $\phi 1$ output voltage activates a servo feedback loop, consisting of a control unit, a two-phase motor, two Variac® autotransformers, and two buck-boost transformers. The deviation in the $\phi 1$ voltage is thus translated into equal corrections applied to $\phi 2$ and $\phi 3$. For a detailed description of the control unit and the motor and transformer circuits, the reader is referred to the January 1966 *Experimenter*.

SPECIFICATIONS

Type	1583-H5	1583-H	1583-H2	1583-LJ	1583-L2J
Input	230 V (line-to-line), 60 Hz			115 V (line-to-line), 400 Hz	
Output	230 V adjustable ± 10%			115 V adjustable ± 10%	
Correction Range* (%)	95 to 105	90 to 110	82 to 124	90 to 110	82 to 124
Line Current (A)	34.0	17.0	8.5	42.5	21.2
Load kVA	13.7	6.8	3.4	8.5	4.2
Correction Time, in cycles (c)	2.5c + 3.0c/V	2.5c + 1.5c/V	2.5c + 0.7c/V	18c + 20c/V	18c + 10c/V
Accuracy (% of output V)	0.25	0.25	0.5	0.25	0.5
Price (depends on mounting)	$620.00 to $655.00			$655.00 to $690.00	

* Ranges listed are for 57- to 63-cycle operation; for 48- to 63-cycle operation, corresponding correction ranges are 95 to 105%, 91 to 109%, and 84 to 119%.

Frequency: 60-Hz models operate from 57 to 63 Hz, and can be modified by a connection change for 48 to 63 Hz; 400-Hz models operate from 350 to 450 Hz.

Response: Rms. **Distortion:** None added.

Efficiency: >98% at full load.

No-Load Power: 45 W.

Ambient Temperature: Operating, −20°C to +52°C; storage, −54°C to +85°C.

Dimensions (width × height × depth): Uncased, 19 × 7 × 14¾ in. (485 × 180 × 375 mm); bench, 19 × 7⅜ × 16 in. (485 × 190 × 410 mm); rack, 19 × 7 × 15 in. (485 × 180 × 385 mm); wall, 19½ × 8⅛ × 16 in. (495 × 210 × 410 mm).

Weight: Uncased, net 54 lb (24.5 kg), shipping 104 lb (47.5 kg); bench, net 64 lb (29 kg), shipping 114 lb (52 kg); rack, net 64 lb (29 kg), shipping 114 lb (52 kg); wall, net 70 lb (32 kg), shipping 120 lb (54.5 kg).

THE GENERAL RADIO

Experimenter

GR'S NEW IC COUNTER

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November-December 1967

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THIS ISSUE

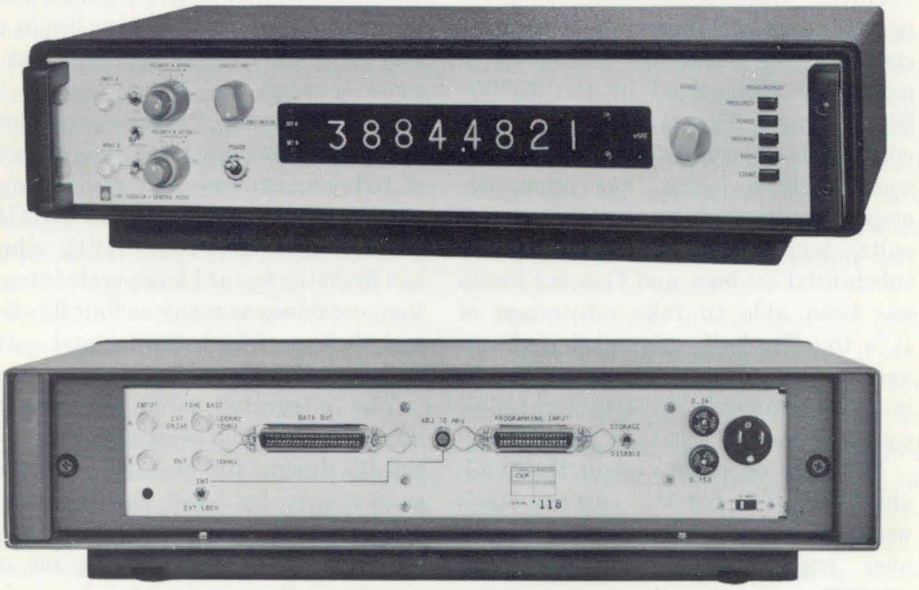


Figure 1. Type 1191 20-MHz Counter, front and rear panel views.

A PROGRAMMABLE 20-MHz COUNTER-TIMER USING INTEGRATED CIRCUITS

The instrument shown on our cover is an eight-digit, programmable, 20-MHz counter capable of measuring frequency, period, frequency ratio, and time interval. It is available with a choice of time bases and optional electrical data output. One of its most outstanding specifications is its price, which serves as evidence that, in at least some areas, technological progress can outrun inflation.

Electronic instruments, as everyone knows, have been getting smaller and smaller — in terms of signal-processing capability per unit volume — since the development of the transistor. Perhaps less widely appreciated is the fact that prices have also been shrinking. Now, with the arrival of integrated circuits, we are witnessing some truly remarkable reductions in instrument costs.

Nowhere is this happy trend more visible than in the area of frequency counters. Only seven years ago, for example, a vacuum-tube 10-MHz counter with period and time-interval capability was priced in the $2500–$3000 range. The transistor brought counter size down to the point where it was hardly a factor, but prices decreased only slightly. Integrated circuits, however, have made possible substantial savings, and General Radio has been able to take advantage of IC's to offer both advanced performance and rock-bottom price, specifically in a programmable 20-MHz, eight-digit counter-timer at less than $1500.*

It is easy enough to credit the availability of IC's for the new counter's wide price and performance margins over vacuum-tube and transistor counters, but we must look further to find out why this counter alone among IC counters offers so much performance at such low cost. The answer lies in the way IC's have been applied. Just as some counters on the market in the early '60's were "transistorizations" of existing vacuum-tube designs, some of today's counters are "integrations" of "transistorizations." Not so GR's new IC Counter (Type 1191), which, in a first step toward large-scale integration, combines as many as four flip-flops and their various buffering and gating stages in a single package.

The integrated circuit shaped the design philosophy behind the 1191. In pre-IC design, for example, one could save money by using slow circuits in noncritical areas. With IC's, extra speed is often available at no cost premium; there may even be savings

* Price applies in USA only.

The 1191 counter on this month's cover signals the debut of GR's new "light-look" instruments. The attractive new styling underscores several important design improvements that mean added convenience and longer instrument life.

Panels are now coated with a baked epoxy finish, for long wear and resistance to scratches.

All relay-rack units have handles for easy installation and removal from racks. The handles are at the ends, out of the way of controls and switches.

The new panel color conforms with Federal Standard 595 (gray, 26492), ensuring good match with companion instruments.

Removable dress panels on many rack-mountable instruments facilitate installation of special-color panels for systems requirements.

New glass-fiber-reinforced knobs give solid protection against cracking and chipping. They also snap on and off without setscrew manipulation and give complete protection against hot shafts.

New binding posts are 12-pointed (double hex) for better finger grip. Standard ⅜-inch socket wrenches can be used to make permanent, tight connections.

Panel markings are silk-screened with epoxy for excellent legibility and baked for extra abrasion resistance and permanence.

We hope you like the new "light look."

resulting from the use of many identical fast circuits, and fast circuits can mean more efficient programs for the various measurement functions. In the 1191, for example, the minimum display time is not in the 10's or 100's of milliseconds, as with most other counters, but 100 microseconds. The propagation time through the eight identical decade counting units of the 1191 is only about 0.2 μs, and it takes only about 0.3 μs to transfer data into storage. Thus the 1191 can make a time-interval measurement and in only 0.5 μs be ready for the next measurement. Other counters — even IC designs — typically require at least 50 milliseconds.

Today's counters are increasingly called upon to operate as components in automatic systems, responding to electrical control and presenting electrical outputs for use with computers, printers, recorders, etc. In the 1191, proper application of integrated circuits simplifies remote programming to the point where it does not add significantly to cost. At the relatively small cost of a control input socket and its wiring, one can completely avoid the need to shield and "hot-switch" fast, pulse-type electrical signals. In the 1191, the only hot switching is the choice of ac or dc coupling, the choice of whether the two input channels are connected in common or separately, and the signal attenuation. All other functions and ranges are switched by contact closures, dc voltages, or RC circuits.

The greatest savings arise from the simplicity of the measurement programs, achieved without sacrifice of user convenience or versatility of operation. Switching from one measurement function to another, for example, involves only the restructuring of a system comprising two flip-flops and two gates.

THE COUNTER

The 1911 is a general-purpose 20-MHz counter-timer for measuring frequency, period (single or multiple), frequency ratio, and time interval. All measurement functions, ranges, and most of the secondary controls are programmable. The counter functions are dc-controlled, most by simple contact closures to ground.

It is important to note that the 1191 is capable of true time-interval measurements, with full input circuits for both its start and stop signals. It is not a counter that can be converted into a brute-force timer by injection of start and stop pulses into the main gate.

Readout is eight digits of high-intensity neon indicators, with automatic display of decimal point and of measurement units. Internal storage gives continuous display of rapidly changing data without flicker. The operator has front-panel control of all input trigger-circuit characteristics.

The counter is available with either of two time bases. The less-expensive room-temperature-crystal oscillator has a basic stability specification of less than 0.2 ppm/°C from 0 to 50°C and a drift of less than 2 ppm per month. Those who require greater stability may either phase-lock this oscillator to an external standard or purchase the counter equipped with a high-precision oscillator operating in a proportional-control oven. The stability specification for this oscillator is less than 2 parts in 10^{10}/°C from 0 to 50°C in continuous operation, and drift is less than 1 part per 10^9 per day.

Another important option is the electrical data output, a fully buffered

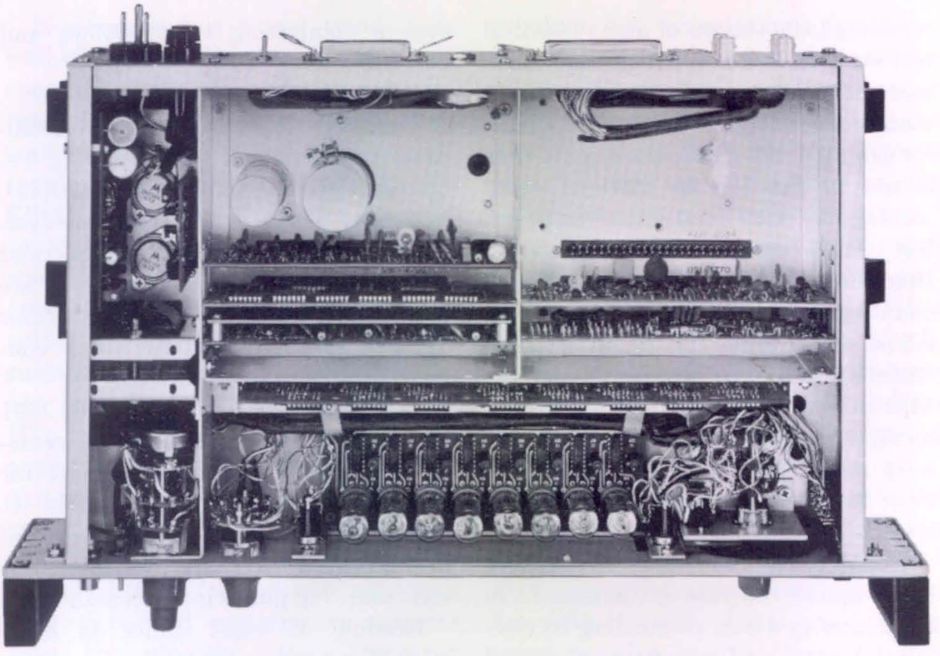


Figure 2. Top interior view.

1-2-4-8 BCD output available at the rear panel of the counter. In addition to the measured data, the output connector carries monitoring data on the main gate and a print-command pulse.

One of the major features of the counter is its "jam transfer" system for holding the displayed or electrical data until the measured data change.* In storage-type counters without jam transfer, all registers momentarily switch off and on between measurements, even where the count in only one of the registers actually changes. Advantages of jam transfer are less rf noise from the counter due to readout tube switching and less noise in the electrical output data.

The simplicity of the counter's design is immediately apparent in an interior view (Figure 2). All circuits except the indicator drivers and the indicators are on plug-in boards, easily removable for repair or replacement. A socket for an extra card is included for specialized systems requirements. Also useful in systems applications are the rear-panel connectors for A and B channel inputs, programming inputs, data-output plug, external-time-base input, and the 10-MHz time-base output.

*US Patent 3,328,564.

HOW IT WORKS

The simplicity of the counter's design is apparent in the following brief descriptions of the circuit programs.

Count Program

The simplest of the programs is the COUNTING mode (Figure 3). Manually operated or electrically programmed switches set a control flip-flop (FF1). When FF1 is in the Q state, the input gate passes pulses generated from

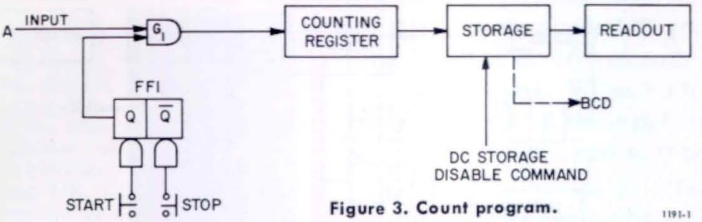


Figure 3. Count program.

the A input signal to the counting register. When a "stop" command resets the control flip-flop to the $\overline{Q}$ state, the total number of pulses accumulated in the counting register is displayed. In the counting mode, the storage units are used as amplifiers.

Frequency Program

For the frequency program, an additional eight-decade counting register, called the timing register, and a crystal-oscillator time base are added to the components used for the counting mode (Figure 4). The function of the additional circuit is to open the counting gate for a precise length of time. If, for example, the counting gate is opened for exactly one second, the number of pulses accumulated in the counting register will be the average number of cycles in one second of the A signal, or the A signal frequency in hertz. The time interval is established by the timing register, which counts up to 10^8 zero crossings of the crystal oscillator (which may, in turn, be phase-locked to an external standard).

The closure of the main gate (G_1) generates commands that move the contents of the counting register into storage, clearing FF2 and closing G_2, and, after a display hold-off interval that can be as short as 100 microseconds, FF2 is reset and the timing sequence is repeated.

Frequency-Ratio Program

The frequency-ratio program is similar to the frequency program, the only

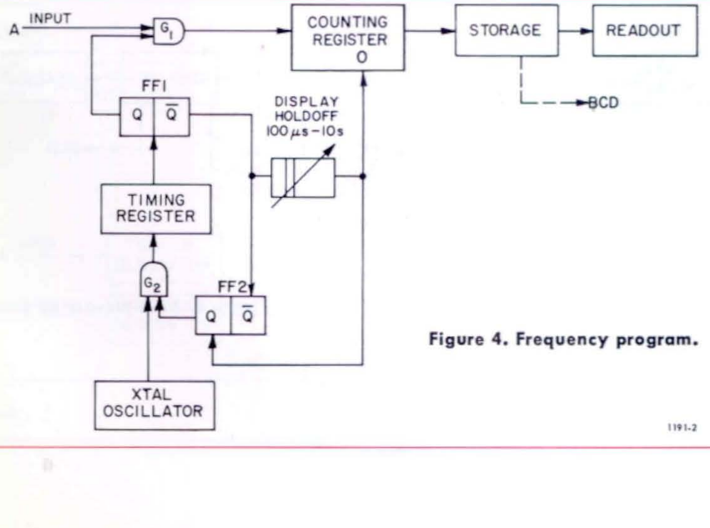


Figure 4. Frequency program.

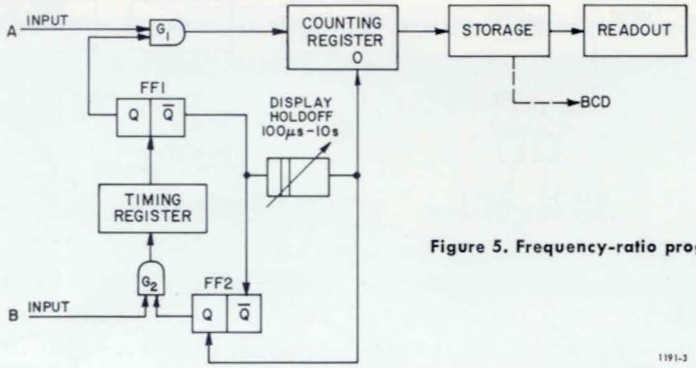


Figure 5. Frequency-ratio program

difference being that the B channel input is substituted for the crystal oscillator (Figure 5). The accumulated count is thus the number of cycles of the A signal during a time interval established by the B signal period.

One very useful application for this program is the use of longer gate intervals than can be obtained with the internal crystal oscillator. If, for example, a 1-MHz standard frequency is available, a 100-second gate will be established with the 10^8 range-switch setting in the ratio program. Similarly, a 1000-second gate can be established with a 100-kHz standard frequency.

Multiple-Period Program

As seen in Figure 6, the multiple-period program uses the same circuits as the frequency ratio program, restructured so that the time-base signal is fed through the input gate to the counting register and the A input is applied to the timing register. The clock interval fed to the timing register is one period of the signal to be measured, and the counted signal is the 100-nanosecond period of the 10-MHz time base. Depending on the position of the range switch, from one to 10^8 periods of the unknown are measured in terms of the time base.

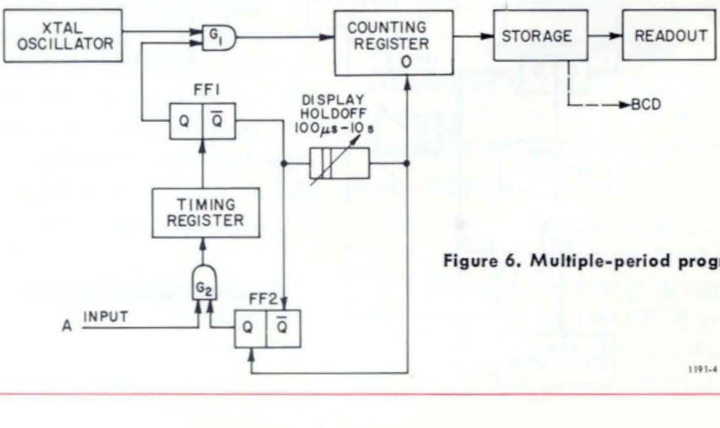


Figure 6. Multiple-period program.

Richard W. Frank, Leader of GR's Frequency and Time Group, received his SB and SM degrees in electrical engineering from the Massachusetts Institute of Technology in 1950 and 1951. He joined GR in 1951 as a development engineer and was appointed Group Leader in 1957.

With the eight-digit counting register and the 10-MHz time base, single periods of signals as low in frequency as 0.1 Hz can be measured in the multiple-period mode. For periods of lower frequency signals, a lower-frequency time base must be counted. Use of the time-interval mode is then indicated.

Time-Interval and Single-Period Program

Figure 7 shows the restructuring of the circuits for the time-interval measurement. The A-channel input circuits are used to generate a pulse that opens G_1 to start the counting of the time-base signal. The B channel is used to produce a stop pulse. Clock frequencies as low as 0.1 Hz can be produced by the timing register, for a maximum measurable time interval of 10^9 seconds (a little over three years). When both A and B inputs are present, the register indicates the time between A and B. If there is no input to channel B, a single-period measurement of the A-channel input is made.

The time-interval mode can be used for pulse-duration measurements. For example, in the measurement of the duration of a positive pulse, the A-input controls are set to start the counter on the positive-going slope and the B-input circuits to produce the stop pulse on the negative-going slope. Maximum resolution for such measurements is 100 nanoseconds.

PROGRAMMABILITY

Almost all the counter's functions are programmable simply by switch closures to ground. The user chooses those functions he wishes to program by setting the corresponding front panel switches to the EXT PROG position. The extent to which the various controls are programmable is indicated by the

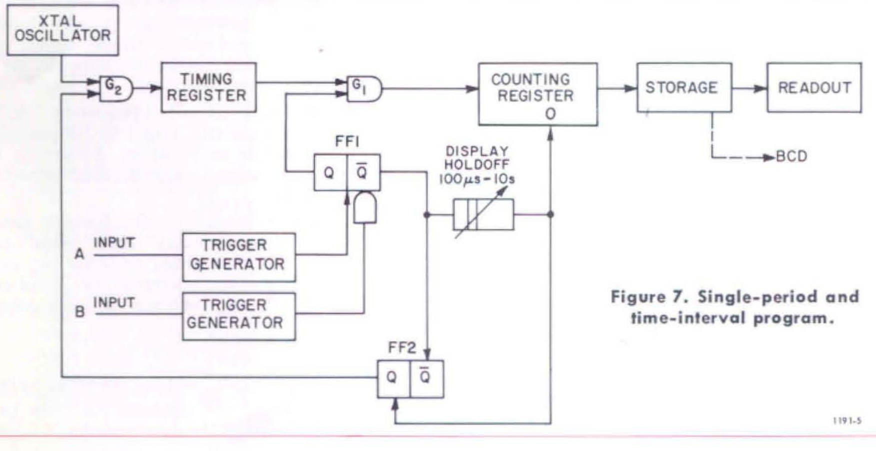


Figure 7. Single-period and time-interval program.

following table. Note that only a few secondary controls are not programmable, and even this limitation can be removed in certain cases on special order.

Function	*Method of Programming*
MEASUREMENT mode	Switch closure to ground
RANGE switch	Switch closure to ground
Trigger-polarity controls	Switch closure to ground
THRESHOLD controls	Dc voltage at desired level
DISPLAY TIME control	RC circuit to ground
START-STOP controls	Switch closure to ground
RESET button	Switch closure to ground
Attenuator controls	not programmable
AC-DC coupling switch	not programmable
SEPARATE-COMMON switch	not programmable

Programming connections are made through a multiterminal connector on the rear panel of the counter. Also available at this connector are several outputs for use with auxiliary equipment. These include a carry pulse from the highest decade, an end-of-display pulse, and a print-command pulse.

ABOVE 20 MHz

The 1191 is easily adapted for use at frequencies to 100 MHz by the addition of the Type 1156 Decade Scaler.[1] The resulting combination is a versatile, programmable, 100-MHz frequency-measuring system priced in the $2000* area.

— R. W. FRANK

ACKNOWLEDGMENT

This design project, extending over the last 10 months, was directed by the author. Such an accelerated development program required the cooperation of too many people, both within and outside the Engineering Department, to make an individual listing of contributors practical; however, particular mention of the contribution of Brian Sargent is in order. Brian contributed to the electrical design and energetically followed the many details necessary for the smooth entry of this instrument into production.

R. W. FRANK

* Price applies in USA only.

[1] D. S. Nixon, Jr., "100-Mc Decade Scaler," *General Radio Experimenter*, September 1965.

SPECIFICATIONS

MEASUREMENT RANGES AND ACCURACY

Frequency: Dc to 20 MHz; 1-μs to 10-s counting gate times. Accuracy, ± 1 count ± time-base accuracy.

Single Period: 1 to 10^9 s measured by counting 0.1-μs to 10-s intervals derived from internal 10-MHz clock. Accuracy (see note) is dependent on the signal-to-noise ratio of the input signal, the counter input noise, and the ± 1-count error.

Multiple Period: 1 to 10^8 periods measured by counting internal 10-MHz clock. Accuracy, see note.

NOTE — Trigger error in time measurements: ±0.3% of one period ÷ number of periods averaged, for a 40-dB input signal-to-noise ratio. This assumes no noise internal to the counter. For input signals of extremely high signal-to-noise ratio, the trigger error in μs will be <0.0005 ÷ the signal slope in V/μs.

Time Interval: 0.1 μs to 10^9 s measured by counting 0.1-μs to 10-s intervals derived from internal 10-MHz clock. Accuracy (see note), ± 1 clock pulse ± trigger error ± time-base stability. Interval is measured between "start" and "stop" pulses driving the input channels independently, or from a single signal with common connection between channels, as for pulse-duration measurements.

Frequency Ratio: 1 to 10^8. Frequency "A," dc to 20 MHz, is measured over 1 to 10^8 periods of frequency "B," dc to 10 MHz. Accuracy, ± 1 count of "A" ± trigger error divided by number of "B" periods.

Count: Register capacity, 10^8. Events are accumulated between "start" and "stop" commands from manual panel buttons or, externally, from contact closures or solid-state switches. In "count," storage is automatically disabled.

INPUT

Frequency: Channel "A," dc to 20 MHz (3 Hz to 20 MHz ac-coupled); channel "B," dc to 10 MHz (3 Hz to 10 MHz ac-coupled).

Sensitivity: 10 mV rms sine wave, 30 mV pk-pk pulse; trigger level variable ±100 mV.

Attenuator: x1, x10, x100 (0, 20, 40 dB); low-capacitance 10:1 probe available.

Voltage Rating: Input voltage should not exceed 150 V on x1 or 300 V on x10 or x100.

Impedance (all attenuator settings): Approx 1MΩ shunted by 35 pF. At rear connectors (supplied mounted, unwired), shunt C increases to approx 70 pF.

Signal Polarity: Front-panel control permits selection of positive- or negative-going signal sense for triggering.

10-MHz TIME-BASE OPTIONS

Room-Temperature Oscillator

Stability: $<2 \times 10^{-7}$/°C from 0° to 50°C. Drift less than $\pm 2 \times 10^{-6}$ per month. With ±10% line-voltage variation, $<2 \times 10^{-8}$.

Manual Adjustment Range: $\pm 1 \times 10^{-5}$ at rear-panel control.

High-Precision Oscillator (in proportional-control oven)

Stability: $<2 \times 10^{-10}$/°C from 0° to 50°C when operated continuously. Drift $\pm 1 \times 10^{-8}$ per week, approx 2×10^{-9} per day after 1 month of continuous operation. With ±10% line-voltage variation, $< 2 \times 10^{-10}$.

Manual Adjustment Range: $\pm 1 \times 10^{-6}$ at rear-panel control.

Time-Base Output: 10-MHz square wave, 2 V pk-pk behind 50 Ω at rear-panel BNC connector.

External Phase-Lock: Both time-base oscillators can be locked to external standard frequency at 0.1, 1, 2.5, 5, or 10 MHz, of at least 1 V rms into 1 kΩ. A front-panel phase-lock indicator lamp is provided.

DATA PRESENTATION

Display: 8-digit display with automatically positioned decimal point and measurement dimensions. High-intensity neon readout tubes.

Storage: Display can be either stored or not. Operator can select from approx 100 μs to 10 s or infinity for display time (in normal mode) and for data holdoff time (in storage mode).

Data Output (in some models): Fully buffered 1-2-4-8 BCD output at standard DTL levels; data zero is 0.5 V max and data 1 approx 5 V behind 6 kΩ.

PROGRAMMING

Input: All instrument functions controllable by closure to ground within capabilities of DTL micrologic (2- to 6-mA sink current required), except:

PERIOD and TIME INTERVAL require approx 50-mA-capacity external closures for added load of dimension-display lamps. Functions controlled by other than contact closure:

Input Threshold: Requires dc voltage of ±100 mV corresponding to desired threshold level.

Display Time: Requires RC circuit to ground. Display/hold-off interval is approx one RC time constant.

Nonprogrammable functions: Input attenuator, input ac/dc coupling, separate/common switch, self-test, internal/external control of time-base oscillator, and frequency adjustment of time-base oscillator.

GENERAL

Environmental: 0° to 50°C operating range.

Power Required: 100 to 125 or 200 to 250 V, 50 to 400 Hz, 32 W.

Accessories Supplied: Rack-mounting hardware set, power cord, spare fuses.

Accessories Available: Input probe, 1156 Decade Scaler for measurement to 100 MHz; 1137 Data Printer, 1136 D/A Converter, and other GR digital-data acquisition equipment.

Dimensions (width x height x depth): Bench model, 19 x 3⅞ x 12¾ in. (485 x 99 x 325 mm); rack model, 19 x 3½ x 11 in. (485 x 89 x 280 mm).

Net Weight (approx): 22 lb (10 kg).

FOR INPUT PROBE — 1158-9600

Input Impedance: 10 MΩ shunted by approx 7 pF when used with 1191 counter.

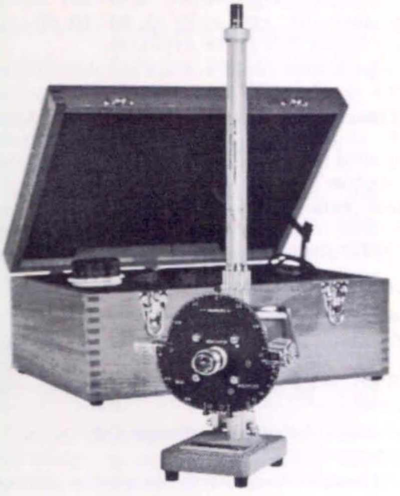
Attenuation: ×10 (20 dB).

Voltage: 600 V dc or ac pk-pk, max up to 5.7 MHz; less at higher frequencies.

Length: 3½ ft.

Catalog Number	*Description*	*Price in USA*
	1191 Counter	
1191-9700	Bench Model	**$1340.00**
1191-9701	Rack Model	**1340.00**
1191-9702	Bench Model with Data Output Option	**1390.00**
1191-9703	Rack Model with Data Output Option	**1390.00**
1191-9704	Bench Model with High-Precision Time-Base Option	**1490.00**
1191-9705	Rack Model with High-Precision Time-Base Option	**1490.00**
1191-9706	Bench Model with both Options	**1540.00**
1191-9707	Rack Model with both Options	**1540.00**
1158-9600	**P6006 Probe,** Tektronix Catalog No. 010-0127-00 (not sold separately)	**22.00**

NEW UHF BRIDGE WITH PRECISION CONNECTORS



In terms of impact on instrument design and performance, one of the most important new products to come from General Radio in recent years has been one of the smallest, the GR900® precision coaxial connector. The GR900 has been the springboard for substantial improvements in many instruments and standards made by GR and other manufacturers. The time-tested admittance meter is the latest beneficiary of GR900 precision, as well as of several other noteworthy improvements.

The development of precision coaxial connectors has led to significant improvement in the performance of many coaxial instruments. The reduction in error introduced at the interface between the unknown and the instrument has made possible not only higher direct-reading accuracy but even greater gains through the use of reference standards equipped with precision connectors. The exceptional repeatability of the GR900® connector has already been put to good use on slotted lines and rf bridges; now users of admittance meters can share in the many advantages of the precision connector.

The TYPE 1602 Admittance Meter has over the years found widespread application for impedance and admittance measurements in the awkward range between lumped-constant and distributed-parameter frequencies. Its frequency range of 20 to 1500 MHz includes frequencies at which slotted lines become long and expensive and conventional bridges suffer from errors due to stray inductances and capacitances. Now a new admittance meter, the 1609, joins the 1602 to bring a new order of accuracy to measurements in this area. (The admittance meter, incidentally, is a null instrument more properly called a bridge, a point we are happy to put right in the naming of the new instrument.)

The 1609 Precision UHF Bridge is a completely new instrument, different in several important respects from the older 1602. In addition to GR900 connectors, the 1609 has more rugged over-all construction, improved, lockable conductance and susceptance standards, and several features that facilitate bridge-arm adjustment. Integral multiplier settings, for example, are detented. The real and imaginary arms can be precisely adjusted by means of a vernier attachment and can be locked into position.

The direct-reading accuracy of the 1609 UHF Bridge is quite good, but truly spectacular results are possible through the use of GR900-equipped calibration standards. For example, through the use of GR900 reference air lines and standard terminations, low SWR's can be measured to within ± 0.002. Recently introduced precision air capacitors, standard resistive terminations, and GR900 air lines, used as C, R, and L standards, similarly improve the accuracy of measurement of capacitance and inductance of lumped- and distributed-parameter circuits to well beyond the direct-reading accuracy.

Precision measurements are not limited to circuits and components equipped with GR900 connectors. Through the use of GR900 precision adaptors, the UHF Bridge connects easily to any of the popular coaxial connectors, including Types N, TNC, BNC, etc. The deterioration in accuracy introduced by the adaptor is usually negligible compared with the inherent uncertainty characteristics of the lower-performance connectors. For measurements on lumped-parameter components with wire leads, the TYPE 900-M Component Mount can be used.

Method of Measurement

Use of the UHF Bridge is simple and straightforward. With a suitable generator and detector and the conductance

New uhf bridge in use measuring SWR of termination, with 7.5-cm reference air line connected as impedance standard.

John F. Gilmore received his BSEE in 1961 and his MSEE in 1963, from Northeastern University. He joined the Microwave Group at General Radio as a development engineer in 1963 and is currently engaged in microwave circuit design.

and susceptance standards (supplied) connected to the bridge, the unknown is connected and the bridge arms are adjusted for a null. At this balance the bridge arms indicate the real and imaginary terms of the normalized admittance of the unknown at a point about 4.9 cm from the GR900 connector face. A short or open circuit is then measured to establish a known reference plane, and a Smith Chart or transmission-line equations are used to determine the impedance or admittance at any reference plane, just as in slotted-line measurements.

Since this is a null instrument, detector response does not affect measurement accuracy, as it can in slotted-line measurements. Multiplier plates are supplied for better resolution of very small or very large values.

SWR Measurements

When only the SWR of a component is of interest, the bridge can be used as a reflectometer. With the component connected to the bridge, the detector output level relative to a calibration level can readily be converted to SWR. Since no balancing is involved, this method is particularly useful for measurement of large numbers of components.

For accurate measurement of low SWR's, a GR900 reference air line can be used as an absolute impedance standard. The bridge then acts as a tunable hybrid. The unknown is connected and the bridge is balanced. The air line is then inserted between the bridge and the unknown, and the detector output is then an accurate indicator (± 0.002) of the SWR of the unknown. — J. F. Gilmore

SPECIFICATIONS

Accuracy: Applies to each term of normalized admittance reading separately.

Frequency	*Larger term <1*	*Larger term 1 to 20*
20–500 MHz	$\pm(0.02\|Y_N\| + 0.01)$	$\pm(0.02\sqrt{M}\|Y_N\| + 0.01M)$
500–1000 MHz	$\pm(0.03\|Y_N\| + 0.01)$	$\pm(0.03\sqrt{M}\|Y_N\| + 0.01M)$
1000–1500 MHz	$\pm(0.05\|Y_N\| + 0.01)$	$\pm(0.05\sqrt{M}\|Y_N\| + 0.02M)$

$|Y_N|$ = magnitude of bridge reading (normalized units)
$= \sqrt{(\text{real term})^2 + (\text{imag term})^2}$
M = setting of multiplier arm, values >1 to 20 required if normalized real or imaginary term is >1.

Impedance accuracy same as above substituting $|Z_N|$ for $|Y_N|$. SWR accuracy $\pm 2\%$ from 20 to 1000 MHz, $\pm 4\%$ 1000 to 1500 MHz, for measurements near unity (matching to 50-Ω system).

Frequency Range: 40 to 1500 MHz, direct reading; down to 20 MHz with correction factor applied to imaginary term.

Measurement Range: 0 to 400 m℧ or 0 to 1000Ω, direct reading; can be extended to 4000 m℧ or 10,000 Ω with multiplier plates (supplied). Instrument measures admittance 4.9 cm inside mating plane of GR900® connector; readings normalized with respect to 20 m℧ (50 Ω). The addition of air line of appropriate length makes instrument direct-reading at any desired reference plane.

Accessories Supplied: 50-Ω conductance standard, adjustable stub and variable air capacitor for susceptance standards, two multiplier plates, 874-R22LA Patch Cord, wooden storage case.

Accessories Required: Generator with 20 mW to 2 W output, detector with better than 10-μV sensitivity. Recommended, GR oscillators, GR Type DNT Detector or 1236 I-F Amplifier, 874-MRAL Mixer, and appropriate oscillator.

Accessories Available: 900-LZ Reference Air Lines as impedance standards, GR900 Standard Terminations and Standard Mismatches for calibration, GR900 adaptors to other connector types.

Dimensions (width x height x depth): 5 x 7¼ x 5½ in. (130 x 185 x 140 mm).

Weight: Net, 16 lb (7.5 kg); **shipping,** 20 lb (9.5 kg).

Catalog Number	*Description*	*Price in USA*
1609-9701	**1609 Precision UHF Bridge**	**$795.00**

NEW RECORDING WAVE ANALYZERS

In the May-June *Experimenter*, we described a new dial drive for automatic stepped or continuous third-octave analysis (as well as continuous tenth-octave analysis) with the Type 1564-A Sound and Vibration Analyzer.[1,2] This dial drive, the analyzer, and the Type 1521-B Graphic Level Recorder are now available in combination, as the Type 1912 Third-Octave Recording Analyzer. Stepped third-octave analysis is finding increasing application, especially through the impetus of military and industry testing standards (e.g., Mil Std 740-B, ASHRAE 36A-63), and this new system makes these measurements simple and convenient.

In addition to the instruments mentioned above, the system includes 10 rolls of chart paper. The recorder is equipped with a 40-dB potentiometer. The system cabinet includes a storage drawer and a system power control that switches the analyzer battery supply as well as the ac line.

Another new system combines the Type 1568-A Wave Analyzer[3] with the graphic level recorder. The 1568 is a 1%-bandwidth analyzer capable of separating very closely spaced frequency components over a frequency range from 20 Hz to 20 kHz. Applications include harmonic-distortion measurements, harmonic analysis (the 1568 will separate about 50 harmonics), and measurements on modulated signals. The analyzer is especially useful in measurements of low-frequency noise or vibration from machinery.

The new Type 1913 Recording Wave Analyzer includes, in addition to the analyzer and the recorder, an 80-dB potentiometer (plus the 40-dB potentiometer supplied with the recorder), 10 rolls of chart paper, and the drive and link units through which the recorder drives the analyzer frequency dial.

[1] B. A. Bonk, "A Dial Drive for Stepped or Swept Analysis," *General Radio Experimenter*, May-June 1967.
[2] W. R. Kundert, "New Performance, New Convenience With the New Sound and Vibration Analyzer," *General Radio Experimenter*, September-October 1963.
[3] W. R. Kundert, "A One-Percent-Bandwidth Wave Analyzer," *General Radio Experimenter*, September 1966.

Catalog Number	*Description*	*Price in USA*
1912-9700	**1912 Third-Octave Recording Analyzer** for 115-V, 60-Hz supply	**$3510.00**
1913-9700	**1913 Recording Wave Analyzer, 1% Bandwidth** for 115-V, 60-Hz supply	**3180.00**
1913-9701	**1913 Recording Wave Analyzer, 1% Bandwidth** for 230-V, 50-Hz supply	**on request**

the Experimenter

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Precision Measurements Association Conference

The first annual conference of the Precision Measurements Association will be held at the Disneyland Hotel in Anaheim, California on January 22, 23, and 24, 1968. Instructional courses, technical sessions, and exhibits will be presented.

General Radio will participate in this program by demonstrating the use of GR900® precision connectors and standards in the calibration of instruments. GR will also demonstrate a technique for calibrating capacitance standards at high frequencies by means of a microwave slotted line equipped with a precision connector.

For further information on the PMA Conference, write to the Precision Measurements Association, 826 N. Victory Blvd., Burbank, California 91502.

General Radio wishes all Experimenter readers a happy and prosperous 1968.

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