the GENERAL RADIO Experimenter

GR Experimenter 1969 Contents & Index

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THE
GENERAL RADIO

Experimenter

VOLUME 43
NUMBERS 1, 2
JANUARY/FEBRUARY 1969

A DIGITAL FREQUENCY DIVIDER AND DELAY GENERATOR
NEW COUNTER FEATURES 35-MHz BANDWIDTH
1-GHz-BANDWIDTH AMPLIFIER DESIGN

The Cover: Prize-Winning Sculpture Uses GR Strobes

Experiments in Art and Technology (E.A.T.), a group involved in the interactions of artists and engineers, recently awarded a $1000 second prize to Mr. Frank Turner of Western Union for his engineering contribution to artist Wen-Ying Tsai's work "Cybernetic Sculpture."

The piece, which is based on the idea of harmonic motion, is made up of groups of nine-foot-high stainless-steel rods illuminated by stroboscopic light. Each rod is vibrated at its base, and the vibrations excite standing waves in the rods. When the surroundings are quiet, the strobes are synchronized with the mechanical-vibration frequency and the standing waves appear frozen. But sounds made by viewers frequency-modulate the flashing rate, causing the frozen standing waves to spring into shimmering, graceful undulations.

Mr. Tsai's work was recently on view at the New York Museum of Modern Art's exhibition, "The Machine as seen at the End of the Mechanical Age."

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The 1399 Digital Divider/Period and Delay Generator.

THE DIGITAL DIVIDER

A SYNTHESIZER FOR PERIOD AND DELAY

The 1399 Digital Divider/Period and Delay Generator bears the same relation to a conventional delay generator that a frequency synthesizer bears to a signal generator. With a frequency synthesizer we can obtain, for example, output frequencies up to 70 MHz in 10-Hz steps with the stability of the reference frequency. Analogously, the 1399 provides periods from 0.3 to 9,999,999.9 μs in increments of 0.1 μs when it is driven by its internal 10-MHz reference oscillator, and the accuracy and stability of the synthesized period are those of the reference.

Basically, the 1399 is a divider; it can divide any frequency from about 10 Hz up to more than 10 MHz by any selected integer from 3 to 99,999,999 – from a fractional frequency to a standard integral frequency, or vice versa. Since the instrument's circuitry is digital, practically the sole source of jitter is the derivation of the clock pulse from zero crossings of the reference signal. The output period consequently has remarkably little jitter; the stability of a 1-second output signal obtained by dividing a 1-volt, 10-MHz reference by 10^7 will typically be better than 10^{-12} second! The jitter in the output period would most likely be determined in this case by the second-to-second stability of the reference signal.

Simply throwing a front-panel switch converts the 1399 from a divider to a digital delay generator. While digital delay generators are now common laboratory tools, the unconventional circuitry of the 1399 makes it unique in this capacity: it is a digital delay generator *without recovery time.* It can, for example, produce 0.999,999,9-second delays at a 1.000,000,0-second rate! Ordinarily, the delay mode uses the precision internal time base, but, as in the divider mode, an arbitrary external signal can serve as the reference.

The 1399, like most recent GR instruments, has full external programming capabilities. All functions can be controlled remotely, most of them by low-current contact closures. The divider ratio or delay time, normally set on front-panel thumbwheel switches, can be externally controlled by switches, relays, or saturated NPN-transistor switches. The control-data format is either 1-2-4-8 or 1-2-4-2 BCD. To program a given decade externally, the corresponding thumbwheel is simply set to zero.

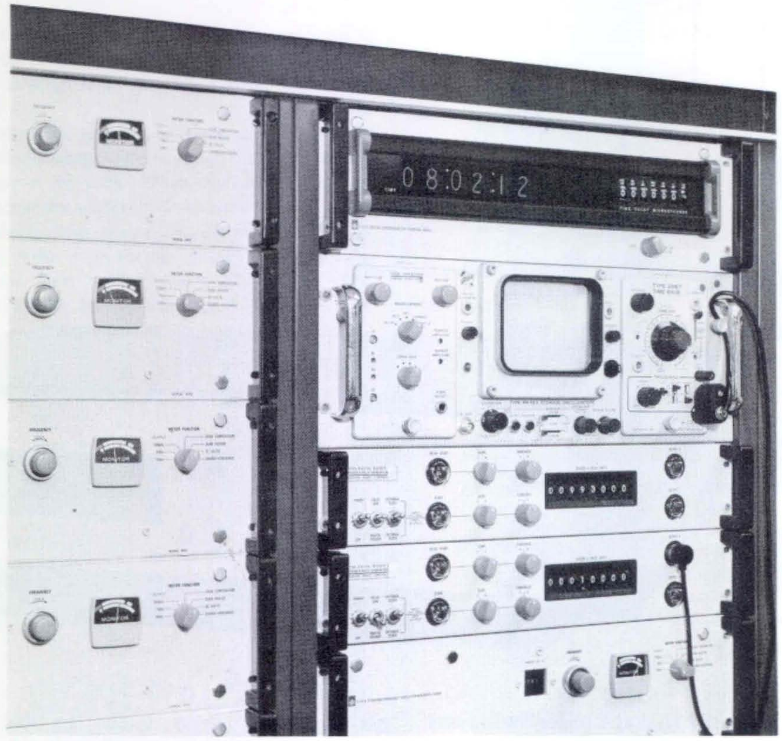


Figure 1. The 1399 and Loran C. An 1123 Syncronometer® digital time comparator, an 1124 Receiver, and two 1399's used to calibrate 1115-C Standard-Frequency Oscillators. The 1399's enable the system to derive time differences from a Loran system with arbitrary base rate.

APPLICATIONS

Delayed Sweep

It has been our experience that a device such as the 1399 is an essential oscilloscope accessory in examining time relationships in digital computers. Delay-sweep oscilloscopes ordinarily use analog delay systems, which exhibit excessive jitter when generating long delays. A coherent digital delay system does not suffer from this defect. The 1399, timed from the computer's clock, can be used to trigger a fast oscilloscope sweep. When the delay is initiated by one event, a subsequent event can be displayed on the oscilloscope screen with minimum jitter.

Counter-Readout Testing

The 1399 has proved to be the economical solution to a problem in our own calibration laboratory. In the final checkout of our counters, we switch on every digit sequentially in every decade of the readout in order to check the data-output wiring and the gas readout tubes. To do this, the 1399 is programmed to generate successive intervals of 1,111,111.1 μs, 2,222,222.2 μs, etc. While a frequency synthesizer could do the same job, the 1399 ties down a smaller investment in instrumentation.

Counter Accessory

As an accessory to a counter, the 1399 can be used to scale down the standard frequency by a ratio chosen at will in order to provide an arbitrary interval for the count. Or the 1399 can be used to count events and to gate the counter, which then measures elapsed time.

Digital Frequency Synthesizer

The development of the dividing scheme that has been used in the 1399 was undertaken originally in connection with a project to design a digital frequency synthesizer. Special-purpose synthesizers based on digital division are now quite common. Their strong point is the small number of tuned circuits required; thus, they are compact and require almost no setup adjustment.

A frequency-synthesizer scheme based on digital division is shown in Figure 2. The oscillator can be phase-locked to harmonics of the standard frequency f_s. If f_s were one hertz, for example, the 1399 would lock the oscillator at every hertz from about 10 Hz to over 10 MHz. Coarse tuning of the oscillator can be done manually, or it can be done automatically by adding a discriminator at the scale-of-N output.

The disadvantage of the digital-divider synthesizer of Figure 2 is the long settling time when the minimum frequency increment is small. Since the bandwidth of the filter in the phase detector's output has to be small compared with f_s, the phase-lock loop must have a time constant that is long compared with $\frac{1}{f_s}$ – perhaps 30 seconds if f_s were 1 Hz.

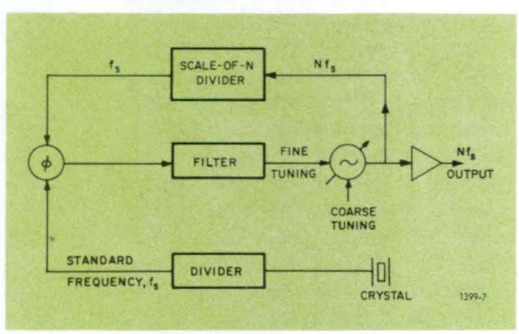


Figure 2. A digital-divider frequency synthesizer.

SPECIFICATIONS

Frequency Divider Ratio: 3:1 to 99,999,999:1. Delay range, 3 to 99,999,999 clock time-intervals, i.e., 0.3 μs to 9.999,999,9 s with 10-MHz clock.

Delay Accuracy: Delay interval varies from 0 to 1 clock interval when clock and start signals are not coherent.

INPUT CHARACTERISTICS Clock and delay-start inputs are identical except for max frequency (rate).

Rate: Delay-start input, 100 Hz to 2.5 MHz. Ext clock input, max > 10 MHz, typically 12 MHz; min, 100 Hz for 1-V pk-pk sensitivity, lower frequency with reduced sensitivity.

Sensitivity: 100 mV rms; will accept waveform of arbitrary shape.

Input Impedance: Approx 100 kΩ‖30 pF.

Trigger Threshold: ±1 V dc offset.

Trigger Polarity: Positive or negative, switch-selected.

INTERNAL CLOCK OSCILLATOR

Frequency Control: 10-MHz third-overtone quartz crystal in proportional-control oven.

Temperature: <1 ppm, 0°C to 50°C.

Warmup: Within 1 ppm from room temperature in 10 min.

Short-Term Stability: 1×10^{-7} for 1-s sampling interval.

Long-Term Stability: 1×10^{-6} per year; with oscillator running continuously, $< 3 \times 10^{-9}$ per day after one month of operation.

Internal Clock Output: 1 V rms into 50 Ω.

Output Pulse: 5 V behind 50 Ω positive and negative available simultaneously. Duration approx 15 ns

PROGRAMMABILITY All functions and control settings, except trigger threshold, controlled by single contact closures to chassis ground. Max current, 2 mA through closed contact; max voltage drop, 150 mV across closed contact.

Divider/Delay Control: 1-2-4-8 BCD; DTL logic levels or contact closures.

Trigger Threshold: 0 to +10 V into approx 100 kΩ produces −1 to +1-V threshold detection.

Power Required: 100 to 125 or 200 to 250 V switch selected, 50 to 400 Hz, 20 W.

Accessories Supplied: Power cord, spare fuses, and mounting hardware with the rack model.

Mounting: Bench model (in metal cabinet) or rack model.

Dimensions (width × height × depth): Bench, 19½ × 4⅞ × 17 in. (495 × 125 × 435 mm); rack, 19 × 3½ × 16 in. (485 × 89 × 410 mm).

Net Weight: Bench, 28 lb (13 kg); rack, 21 lb (10 kg).

Shipping Weight (est): Bench, 43 lb (20 kg); rack, 36 lb (16.5 kg).

Catalog Number	Description
	1399 Digital Divider/Period and Delay Generator
1399-9801	Bench Model
1399-9811	Rack Model

HOW IT WORKS: GET READY; GET SET; GO!

A functional diagram of the 1399 is shown in Figure 3. A 10-MHz third-overtone crystal in a proportional oven produces the 0.1-μs internal clock signal. The clock can also be an external signal of a few hertz to over 10 MHz. A gate determines whether the divider will run continuously, for the divide mode, or start on command of an external signal, for the delay mode. Both the delay-start and external-clock inputs have slope and threshold controls for establishing the trigger pulses from well-defined portions of the input signals. These input-trigger circuits are similar to those in most counters. The 1399's output circuit produces brief (15-ns), high-energy (±5 V behind 50 ohms) pulses that mark off the controlled interval.

The criteria for an arbitrary-scale divider were set down a number of years ago in a project at General Radio to develop a digital frequency synthesizer.[1,2,3] What was needed, simply, was a system that would produce one output pulse for every Nth input pulse. It was hoped that the more significant decades could be slower, as they are in a counter – an important design objective, because fast flip-flops were, and still are, more expensive, more power hungry, and less reliable than slower ones. Ideally, the maximum counting rate of the scale-of-N should not be much less than the resolution of the first flip-flop in the fastest-counting decade. These were the goals. As the design of an actual circuit proceeded, there became apparent two rather fundamental obstacles to the use of conventional counting schemes in a frequency divider.

[1] R. W. Stuart, "A High Speed Digital Frequency Divider of Arbitrary Scale," *1954 IRE Convention Record,* Part 10.

[2] R. W. Frank, "A Computer Type Decade Frequency Synthesizer," *1954 IRE Convention Record,* Part 10.

[3] Work done under Contract DA 36-039 SC-15542, Signal Corps Engineering Laboratories, Ft. Monmouth, New Jersey.

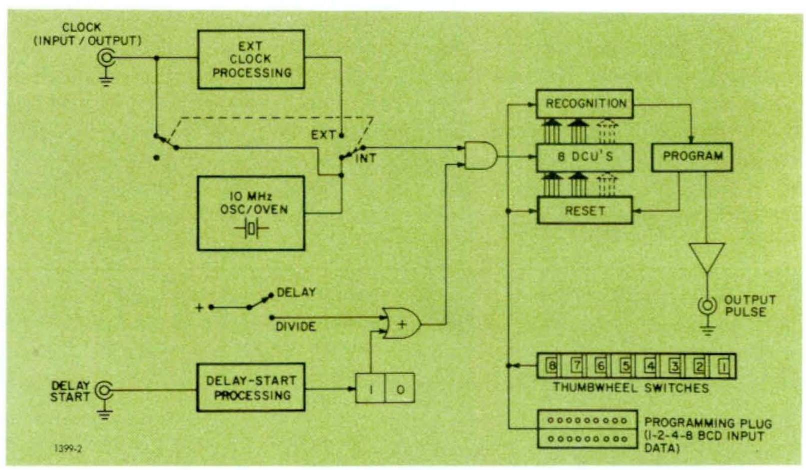


Figure 3. Functional diagram of the Type 1399 Digital Divider/Period and Delay Generator.

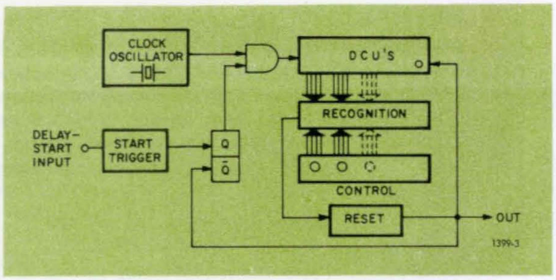


Figure 4. Simple delay generator based on digit recognition.

Figure 4 shows a simple delay generator based on digit recognition. At the first clock pulse after the gate is opened, the decade counting units (DCU's) begin to accumulate. They fill to the recognition state established by the controls, whereupon an output pulse is generated and the DCU's are reset to zero. If the duty ratio of the delay is not very high, the reset interval can last long enough to reset the slowest decade.

The problem with this simple delay scheme is the time required for propagation through the registers. If, by the time the last decade has reached its recognition state, the first one has moved on, recognition will not occur. This limit is reached rather quickly; five consecutive flip-flops, each with a delay of 20% of its resolution (a typical value), will cause a recognition failure at the maximum clock frequency even if all flip-flops are equally fast (which would be a violation of the first of our design criteria).

Figure 5 shows a scale-of-N divider. It is similar to the delay circuit just discussed, but it must divide continuously. This requirement compounds our problems: not only do we have to achieve reliable recognition, we must also complete the reset in less than one clock period. The situation has one hopeful circumstance, however. *If there is time to reset the highest-speed decade, then there is plenty of time to reset the others before they start to accumulate counts (10 clock periods for the second decade, 100 for the third, etc).*

There is another well-known way to achieve scale-of-N division when clock rates are not too fast. Figure 6 shows a circuit that recognizes only one state, namely, the output carry of the last decade. The division integer is determined by the *reset* state of the DCU's: the decades are reset to the complement of the desired divisor. Here there is no worry about time delays interfering with recognition; but we have traded that problem for another: there is not enough time for resetting. Suppose our divisor ends in the digits 001. The corresponding three DCU's would have to be preset to the complement 999. The first clock pulse after reset would carry straight through these three digits, changing them to zeros, and spill over into the fourth. The trouble is that this would have to happen before the slower second, third, and fourth DCU's had had time to reset. There is not even one full clock period available for resetting, because the reset pulse cannot occur until the output-initiating clock pulse (which changes *all* the digits from nines to zeros) has propagated through the entire counting register.

By now it will be clear that each of the two basic scale-of-N dividing schemes is plagued by a different problem. The reset-one-state-and-recognize-ten-states system suffers from recognition failure due to cumulative time delay, while the reset-ten-states-and-recognize-one-state system suffers from inadequate resetting time. One might wonder if a combination of the two methods could yield a workable system – and in fact it does.

Each DCU must have ten distinct states. We have talked about systems in which the DCU's have ten recognition states and one reset state, and systems in which they have one recognition state and ten reset states. Table 1 shows an intermediate scheme of 2 reset and 5 recognition states. In this system, a decade has to accumulate at least five counts following a carry before it is in a recognition state. Therefore every decade after the first remains in a recognition state for at least 5 preceding-decade pulses, waiting for the preceding decade to reach recognition. Furthermore, since decades are reset to no more than 5, every decade after the first will have five preceding-decade pulses in which to be reset. The reader might like to work out for himself the details of the reset-5-recognize-2 system shown in Table 2. It turns out to be only a little less efficient than the reset-2-recognize-5 system just discussed.

These examples of counting systems with mixtures of reset and recognition states demonstrate that an appropriate choice of coding can solve the recognition and reset problems that are inherent in the basic divider arrangements. We have seen that, with either of the two codings just described, first, the slower DCU's are waiting at their recognition states for the faster ones to catch up, and, second, all the DCU's (except the first) have plenty of time to reset before they have to begin counting.

The remaining limitation on the counting speed is the delay associated with carries between flip-flops in the first decade. Actual delays between

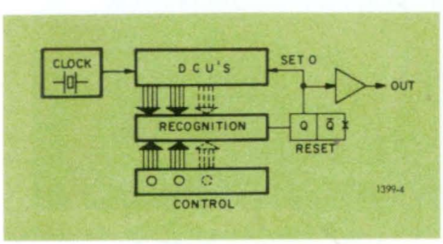


Figure 5. Simple scale-of-N divider. Like the delay generator of Figure 4, this circuit depends on digit recognition.

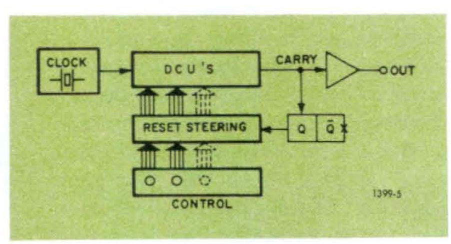


Figure 6. An alternative scale-of-N divider. This circuit recognizes only one state: all DCU's at zero. The count starts with the DCU's preset to the complement of the divisor.

Table 1
Reset-2-recognize-5 system

number	reset	recognize
0	5	5
1	5	6
2	5	7
3	5	8
4	5	9
5	0	5
6	0	6
7	0	7
8	0	8
9	0	9

Table 2
Reset-5-recognize-2 system

number	reset	recognize
0	4	4
1	3	4
2	2	4
3	1	4
4	0	4
5	4	9
6	3	9
7	2	9
8	1	9
9	0	9

Table 3
Reset-4-recognize-3 system. This is essentially the coding used in the 1399.

number	reset	recognize
0	5	5
1	4	5
2	5	7
3	4	7
4	1	5
5	0	5
6	1	7
7	0	7
8	1	9
9	0	9

the various digits will depend on the logical design of the DCU, but the source of delay is always the same: it is the carry generated by the 1-to-0 transition of an individual flip-flop. Now, since *odd* digits are counted by a 0-to-1 transition of only the *first* flip-flop, there is no delay problem with the odd digits. With our reset-2-recognize-5 coding, for example, serious delays occur at the numbers 6 and 8. If we could find a reset-recognition scheme in which only odd numbers have to be recognized, we would have attained our goal of a system whose counting speed is limited only by the resolution of the first flip-flop.

Table 3 shows a reset-4-recognize-3 coding scheme that is essentially similar to the one adopted for the 1399. Like the 2-5 and 5-2 systems, this system resets to no number greater than 5 and recognizes no number less than 5, so that it, too, overcomes the reset-time and propagation-delay difficulties. In addition, this system requires the recognition of only odd numbers.

Let us see how a system with 4-3 coding deals with the awkward divisor 00 000 003. The seven more significant DCU's will be reset to their recognition states of 5, and the first DCU will be reset to 4 (0010). The clock will now advance the first DCU to 7 (1110), and the 6-to-7 (0110-to-1110) transition will establish recognition and reset the first DCU to 4. Since recognition and reset involve transitions of the first flip-flop from 0 to 1 and back to 0 all within one clock period, the system will run half as fast as the first flip-flop – if there are no reset delays. In practice, reset delays would cause a further slight reduction in the speed of the system.

A modification[4] of the 4-3 system permits the 1399 to count at very nearly the speed of the first flip-flop. The trick is to delay the start of the count by two clock periods to allow enough time for the first flip-flop to reset. Refer to the 1399's programming diagram, shown in Figure 7. While counting is in progress, *FF1* is in its Q state, the clock gate is open, and the output gate is closed. Recognition operates *FF1* to Q, thus closing the clock gate, opening the output gate, and triggering the reset-pulse generator. The first clock pulse following recognition passes through the output gate and switches *FF2* to Q. The next clock pulse switches *FF2* back to $\bar{Q}$, thereby triggering the output pulse, resetting *FF1* to Q, closing the output gate, opening the clock gate, and beginning the next count. Counting has been interrupted for two clock periods, allowing plenty of time to reset the first decade. The two missed counts are compensated for by making the first decade recognize 3's, 5's, and 7's instead of 5's, 7's, and 9's. The small price we have to pay for the additional speed is that the system is unable to divide by two, a trivial operation that can be done by one flip-flop. (As a matter of fact, two clock periods are more than enough time for resetting, but skipping just one pulse would require that even numbers be recognized in the first decade.)

[4]U.S. Patent No. 3,050,685.

–R. W. Frank

A brief biography of Mr. Frank appeared in the November-December, 1967 issue of the *Experimenter.*

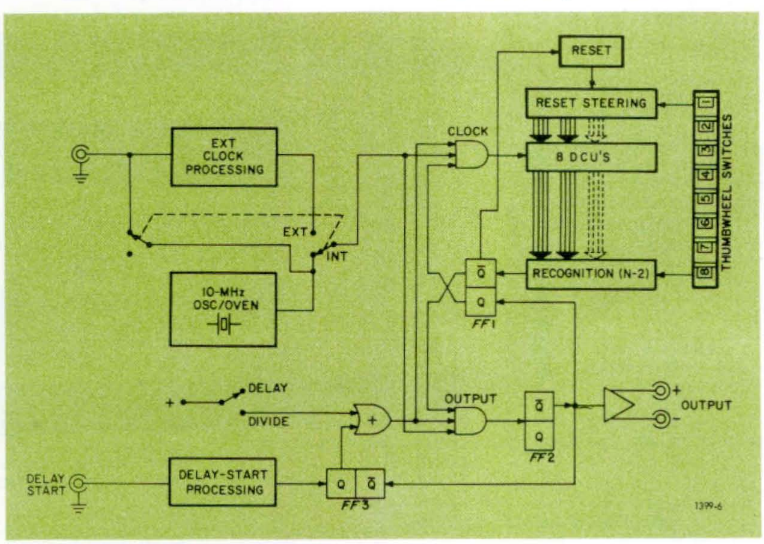
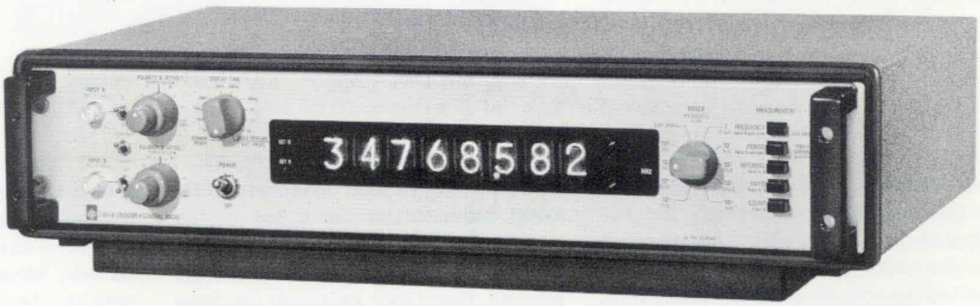


Figure 7. Internal programming of the 1399.

The new 1191-B Counter.



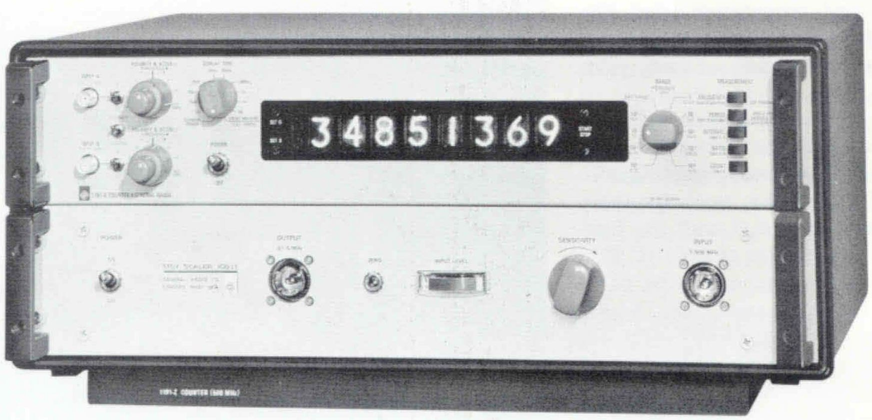
1191 COUNTER NOW FEATURES 35–MHz BANDWIDTH

• "Strobed" single-period measurement • 1-μs data holdoff • Improved time-interval mode • Crystal-oven standby power

The 1191 is an integrated-circuit counter/timer for measuring frequency, period, average period, frequency ratio, and time interval. (See the November-December 1967, *Experimenter* for a complete description.)

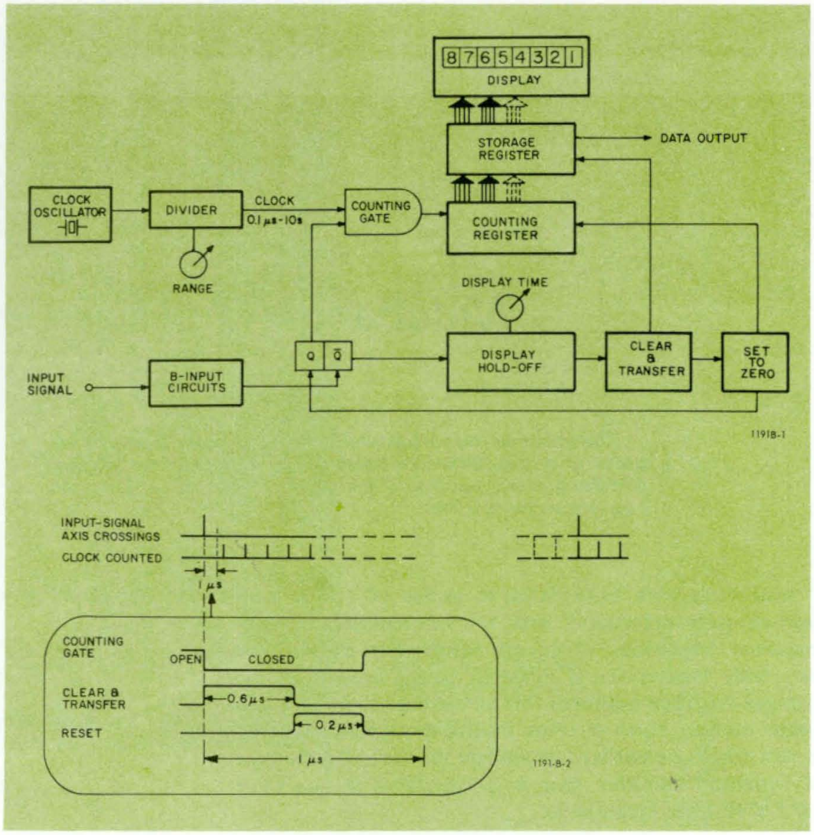
The new 1191-B offers several significant improvements over its predecessor, and at the same base price. Foremost among these are the increased frequency range – the upper limit is now 35 MHz – and a unique "strobed" single-period mode, in which the 1191-B makes period measurements with virtually a 100-percent duty cycle.

Like its predecessor, the 1191-B is available in combination with either of two GR scalers as the 1191-Z. The combination offers all the features of the counter alone, plus operation up to 500 MHz. The bench version of the 1191-Z has the counter and scaler in a single cabinet.



The 1191-Z Counter. The 500-MHz model is shown here in its bench version.

Program of "strobed" period measurement.



"Strobed" Period Measurement

Many measurements require that the counter extract information from the input signal at the maximum possible rate. Ordinarily, a counter measurement of the period of a signal is inefficient from an information-gathering point of view. The period mode is conventionally programmed so that a pair of consecutive signal axis crossings open and close the counting gate. The closing of the gate initiates a hold-off period (often as long as 0.1 second) during which the data is transferred to storage. At the end of the hold-off period the counter is ready to make a new measurement. The duty ratio of such a program can be no more than 50 percent, and it can be this high only if the hold-off period can be made brief enough to allow counting during alternate cycles.

The unique "strobed" period mode of the 1191-B permits the new counter to gather data during almost 100 percent of the time. The internal program of this measurement is shown in the accompanying block and timing diagrams. *Every* input-signal axis crossing initiates an operation in which the accumulated count is transferred to storage, the counting register is reset to zero, and a new count is begun.

If the last-counted clock pulse in a given measured period has to propagate through all eight decades of the counting register, approximately 0.6 μs will elapse after the counting gate closes before the counting register has settled and the data are in storage. Another 0.2 μs is needed to reset the counting register to zero. Thus, with the DISPLAY TIME set to its minimum position (1 μs), the total dead time – the time at the beginning of a measured period before the counting gate opens – is about 0.8 μs. Except at the fastest clock rate (0.1 μs), only a single clock pulse may not get counted because of the dead time, and the probability of missing a pulse decreases by an order of magnitude as the clock rate decreases by an order of magnitude. One can easily determine the precise amount of dead time by measuring an accurately known period with the internal clock set at 0.1 μs. The discrepancy between the count and the known period will be between 0.7 and 0.9 μs.

Catalog Number	Description
	1191-B Counter
1191-9710	Bench Model
1191-9711	Rack Model
1191-9712	Bench Model with Data-Ouput Option
1191-9713	Rack Model with Data-Output Option
1191-9714	Bench Model with High-Precision Time-Base Option
1191-9715	Rack Model with High-Precision Time-Base Option
1191-9716	Bench Model with both Options
1191-9717	Rack Model with both Options
	1191-Z Counter (100 MHz)
1191-9900	Bench Model with both Options
1191-9901	Rack Model with both Options
	1191-Z Counter (500 MHz)
1191-9902	Bench Model with both Options
1191-9903	Rack Model with both Options
1158-9600	**P6006 Probe**, Tektronix Catalog No. 010-0127-00 (not sold separately)

Wideband Amplifier Design

by M. Khazam

This article derives the formulas (in terms of y-parameters) for constructing transistor constant-gain circles on a load-admittance Smith chart. An input-admittance grid constructed on the same chart provides a direct readout of the input admittance. A design example using transmission lines as matching elements is given.

Amplifiers with bandwidths in excess of 1 GHz have become quite practicable with the introduction of high-frequency transistors. Feedback equalization of gain over such wide bandwidths is difficult, partly because of the complex feedback networks that would have to be used and partly because there is often insufficient open-loop gain. In this article we outline a procedure for designing the type of wideband amplifier that is incorporated in the Type 1237 VHF/UHF Preamplifier.

Two-port Power Gain

The y-parameters of a two-port network are defined by the well-known relations

$$\left.\begin{aligned} I_1 &= y_{11} V_1 + y_{12} V_2 \\ I_2 &= y_{21} V_1 + y_{22} V_2 \end{aligned}\right\} \qquad (1)$$

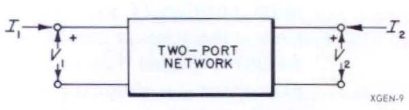


Figure 1

where the signs of voltages and currents are those of Figure 1. If Y_L is the load admittance and Y_S the source admittance, then the input and output admittances are given by

$$Y_{\text{in}} = y_{11} - \frac{y_{12}\, y_{21}}{y_{22} + Y_L} \qquad (2)$$

$$Y_{\text{out}} = y_{22} - \frac{y_{12}\, y_{21}}{y_{11} + Y_S} \qquad (3)$$

The power flowing from the source into the network is

$$P_{\text{in}} = |V_1|^2 \quad \mathcal{R}e\, Y_{\text{in}}$$

and the power flowing out of the network into the load is

$$P_{\text{out}} = |V_1|^2 \left|\frac{y_{21}}{y_{22} + Y_L}\right|^2 \quad \mathcal{R}e\, Y_L$$

The power-gain ratio is thus

$$\mathcal{G} = \frac{P_{\text{out}}}{P_{\text{in}}} = \frac{\mathcal{R}e\, Y_L}{\mathcal{R}e\, Y_{\text{in}}} \left|\frac{y_{21}}{y_{22} + Y_L}\right|^2 \qquad (4)$$

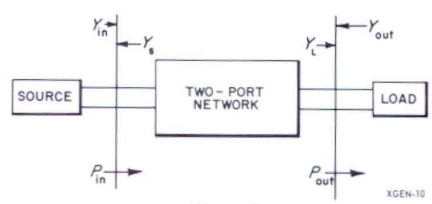


Figure 2

The more significant transducer gain $\mathcal{G}_t$, defined as (power to load) ÷ (source's available power), is related to the power gain $\mathcal{G}$ by

$$\mathcal{G}_t\,(\text{dB}) = \mathcal{G}(\text{dB}) - \text{input mismatch loss (dB)} \qquad (5)$$

Gain Circles On the Load-Admittance Smith Chart

We define a normalized load admittance y by

$$y = g + jb \equiv \frac{\mathcal{R}e\, Y_L}{\mathcal{R}e\, y_{22}} + j\, \frac{\mathcal{I}m\, Y_L + \mathcal{I}m\, y_{22}}{\mathcal{R}e\, y_{22}} \qquad (6)$$

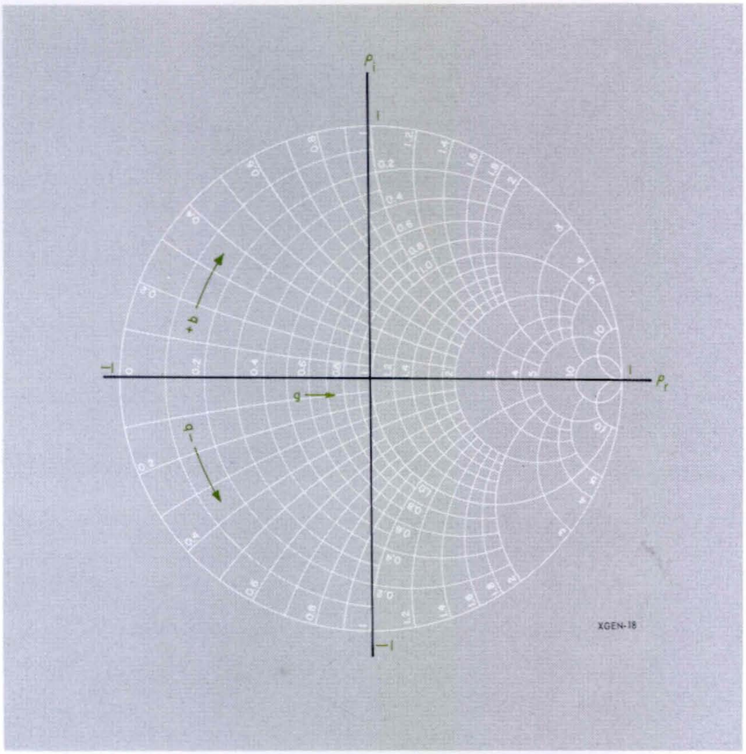


Figure 3. The relation between y and ρ.

Note that the imaginary part of y includes $\mathscr{I}m\, y_{22}$ along with $\mathscr{I}m\, Y_L$. Since g and b may vary over wide ranges of values, we would find ourselves constructing rather unwieldy charts if we continued to work directly with y. Accordingly we make a transformation to a new variable ρ:

$$\rho = \frac{y-1}{y+1} \tag{7}$$

(Note that ρ is the negative of the reflection coefficient, as conventionally defined.) If we write ρ_r and ρ_i for the real and imaginary parts of ρ, the expression (4) for the power-gain ratio can be recast in the following form:

$$\left(\rho_r + \frac{g_t \mathscr{G}}{|y_{21}|^2}\right)^2 + \left(\rho_i - \frac{b_t \mathscr{G}}{|y_{21}|^2}\right)^2 = r^2 \tag{8}$$

where

$$r = \sqrt{1 - 2\mathscr{G}\,\frac{2g_{11}g_{22} - g_t}{|y_{21}|^2} + \mathscr{G}^2\,\frac{|y_{12}|^2}{|y_{21}|^2}} \tag{9}$$

We have used the notation $y_{11} = g_{11} + jb_{11}$, $y_{22} = g_{22} + jb_{22}$, and $y_{12}y_{21} = g_t + jb_t$.

Equation 8 represents a family of constant-gain circles on the ρ-plane. When $\mathscr{G} = 0$, the radius r is unity and the circle's center is at the origin of the ρ-plane (the circle coincides with the zero-conductance circle on the Smith chart). Provided the two-port is unconditionally stable, the radius decreases with increasing gain until a value of $\mathscr{G}$ is reached at which the radius is zero. The gain at this point, $\mathscr{G}_{\max}$, is the maximum that can be achieved without external feedback, and the corresponding point on the ρ-plane represents a conjugate match between the two-port and the load. The centers of the circles,

$$\rho_r(\text{center}) = -\frac{g_t \mathscr{G}}{|y_{21}|^2}, \quad \rho_i(\text{center}) = \frac{b_t \mathscr{G}}{|y_{21}|^2}, \tag{10}$$

fall on the straight line through the origin of the ρ-plane that makes an angle

$$\phi = \tan^{-1}\left(-\frac{b_t}{g_t}\right) \tag{11}$$

with the positive ρ_r-axis. The distance from the centers to the origin of the ρ-plane is

$$d = \frac{|y_{12}|}{|y_{21}|}\,\mathscr{G} \tag{12}$$

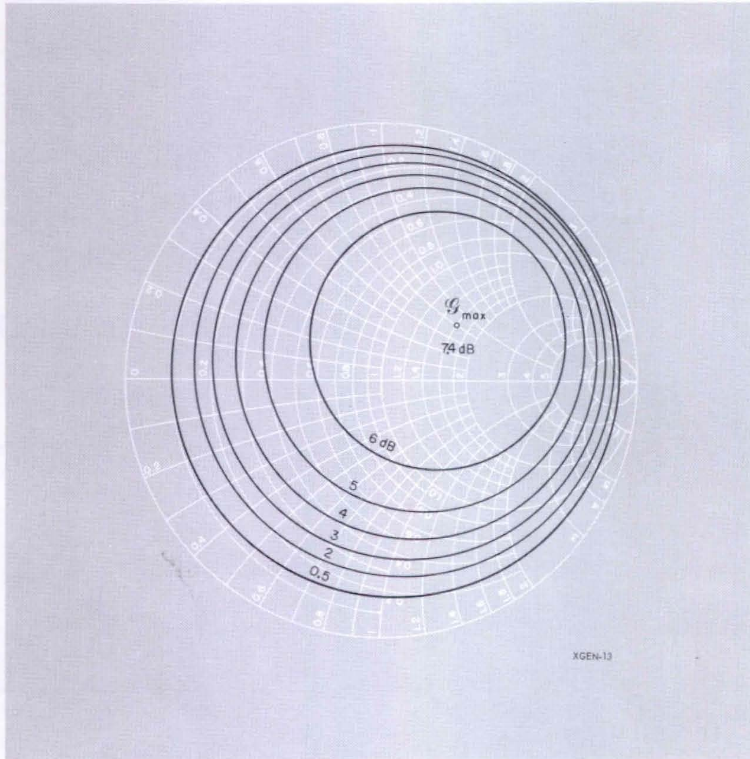


Figure 4. A set of constant-gain circles for a 2N3478 transistor at 900 MHz plotted on the load-admittance Smith chart.

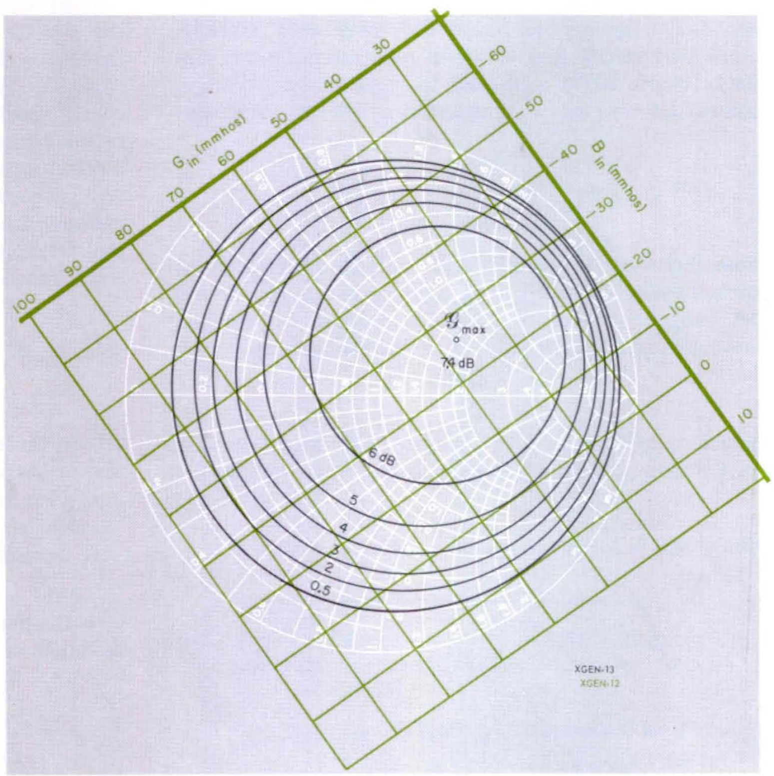


Figure 5. An input-admittance grid has been added to the chart of Figure 4.

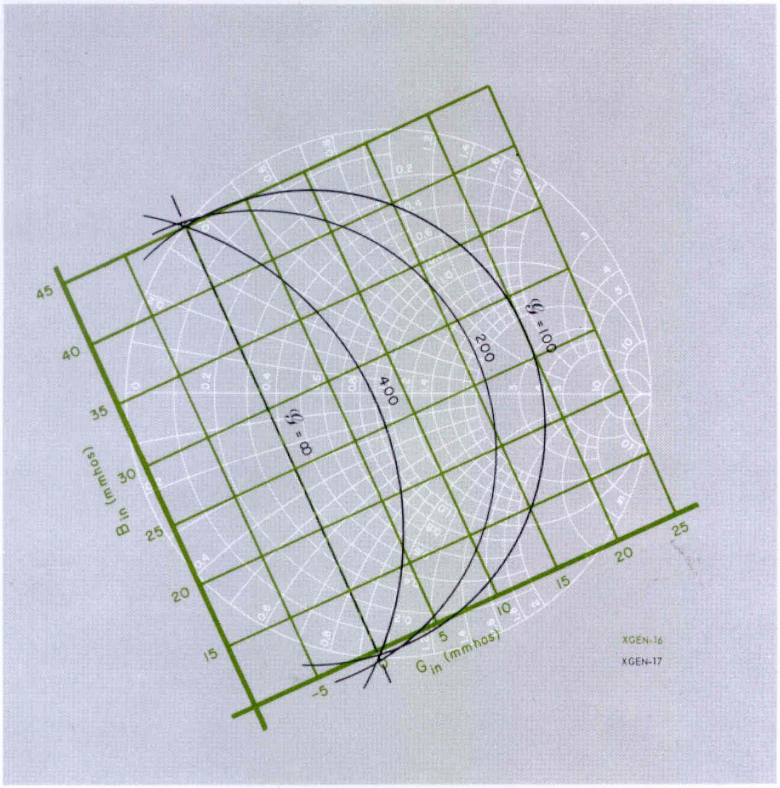


Figure 6. A potentially unstable situation. If the load admittance falls in the lower left of the chart beyond $\mathcal{G} = \infty$, the input conductance is negative and the two-port can oscillate.

Figure 4 shows a set of constant-gain circles for a 2N3478 transistor at 900 MHz, plotted on a Smith chart. The chart's normalized grid corresponds to our admittance variable y.

Input-Admittance Grid

To find the input admittance of the two-port, we return to equation 2:

$$Y_{in} \equiv G_{in} + jB_{in} = y_{11} - \frac{y_{12}\,y_{21}}{y_{22} + Y_L} \tag{13}$$

If we replace Y_L on the right-hand side of (13) by the normalized load admittance y, defined by (6), and then make the transformation (7) to the ρ-plane, we get

$$\left.\begin{aligned} G_{in} &= \left(\frac{g_t}{2g_{22}}\right)\rho_r - \left(\frac{b_t}{2g_{22}}\right)\rho_i + \left(g_{11} - \frac{g_t}{2g_{22}}\right) \\ B_{in} &= \left(\frac{b_t}{2g_{22}}\right)\rho_r + \left(\frac{g_t}{2g_{22}}\right)\rho_i + \left(b_{11} - \frac{b_t}{2g_{22}}\right) \end{aligned}\right\} \tag{14}$$

These expressions represent a rectangular grid of G_{in} = constant and B_{in} = constant lines on the ρ-plane. We can construct this grid on our load-admittance chart if we will note the following relations between the G_{in}– and B_{in}–lines and the ρ_r- and ρ_i-axes. The spacing $\Delta\rho_r$ between the intercepts of consecutive G_{in} = constant lines with the ρ_r-axis is

$$\Delta\rho_r = \frac{2g_{22}}{g_t}\,\Delta G_{in} \tag{15}$$

where ΔG_{in} is the chosen increment in G_{in}. The spacing $\Delta\rho_i$ along the ρ_i-axis between B_{in}-line intercepts is given by the same relation (the G_{in}– and B_{in}–axes have the same scale factor):

$$\Delta\rho_i = \frac{2g_{22}}{g_t}\,\Delta B_{in} \tag{16}$$

The angle that the B_{in} = constant lines make with the ρ_r-axis is

$$\tan^{-1}\left(\frac{\Delta\rho_i}{\Delta\rho_r}\right)_{B_{in}\,=\,\text{const}} = \tan^{-1}\left(-\frac{b_t}{g_t}\right) \tag{17}$$

which is just the angle ϕ, given by (11), of the line of centers of the constant-gain circles. The Y_{in}-grid has still to be positioned on the ρ-plane, and this requires the calculation of one point.

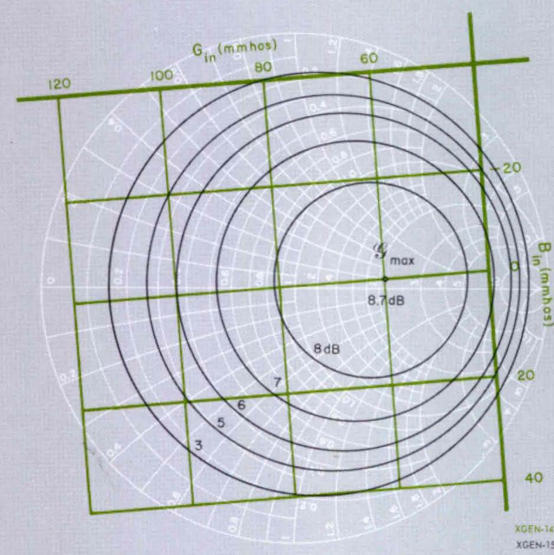


Figure 8. Design chart for the 2N3478 at 600 MHz.

Figure 5 shows an input-admittance grid superimposed on the gain-vs -load-admittance chart of Figure 4.

Stability

The constant-gain circles of Figures 4 and 5 are characteristic of a two-port that is unconditionally stable. An unconditionally stable two-port is one whose input admittance has a positive conductive component whenever the output admittance does, and vice versa.

A potentially unstable situation is illustrated in Figure 6, which shows a load-admittance-Smith-chart plot of the gain circles and input-admittance grid for a 2N3478 transistor plus feedback capacitor at 400 MHz. Notice that in this case the gain circles get larger with increasing gain and that there are negative values of G_{in} within the g = 0-circle, where the load conductance is positive. The $\mathcal{G} = \infty$ "circle" coincides with the G_{in} = 0 line, and load admittances that fall to the right of this line will cause oscillation unless the source conductance is large enough that the total conductance at the input is not negative.

One can inspect the output admittance as a function of source admittance by constructing a rectangular output-admittance grid on the source- admittance Smith chart. The procedure is the same as that outlined for constructing the G_{in}– B_{in}–grid except that the subscript 1's and 2's on the y-parameters are interchanged.

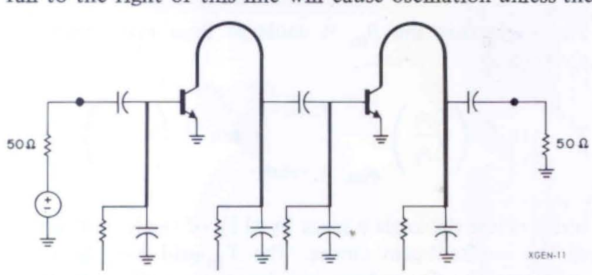


Figure 7. Two-stage wideband amplifier. Matching is accomplished by L-networks formed by series and shunt sections of line.

A Design Example

Let us design a two-stage amplifier using 2N3478's. The analysis of a wideband amplifier by the method described in this article generally involves sampling the performance at a number of frequencies over the passband. Transistor parameters have to be measured and charts constructed at each frequency. For the sake of a simple example, though, let us set ourselves a fairly limited objective: more or less constant gain from 900 MHz down to 600 MHz or possibly lower. We will see what we can do with the circuit of Figure 7, in which the matching networks are simple series-and shunt-line L networks, and we will analyze the performance at only the two frequencies 900 and 600 MHz. We shall assume 50-ohm source and load impedances.

The design chart for 900 MHz is the one shown in Figure 5, and the chart for 600 MHz is shown in Figure 8. We find from the charts that the load admittance for maximum gain is 0.83 - j5.6 mmho at 900 MHz and 1.1 - j3.9 mmho at 600 MHz. Assuming that the matching networks will be etched on a circuit board, we select 125 ohms as a practical value for the characteristic impedance of the line sections.

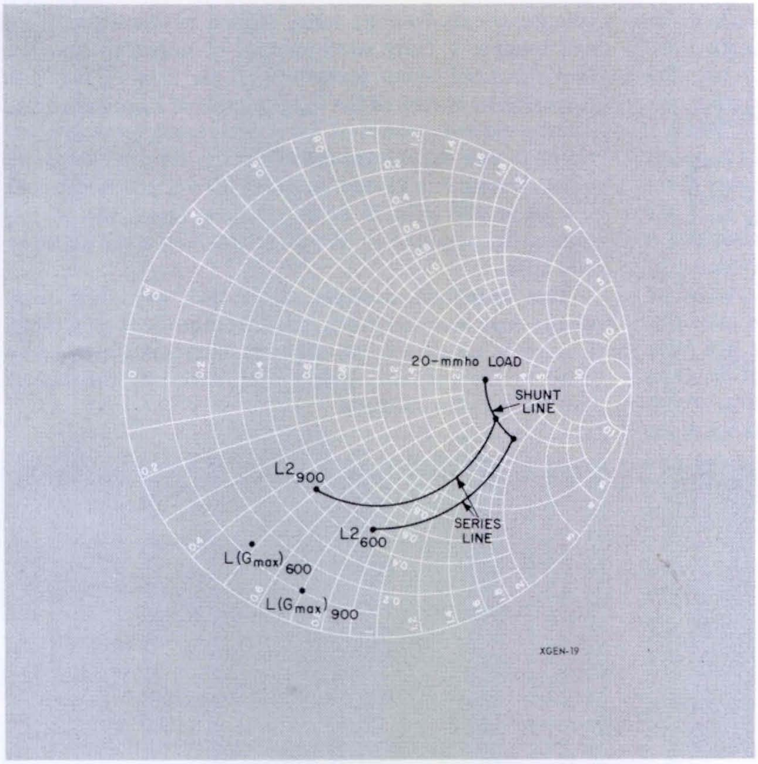


Figure 9. Transformation of the load admittance by the output network. The admittance grid is normalized to 8 mmhos (125 Ω).

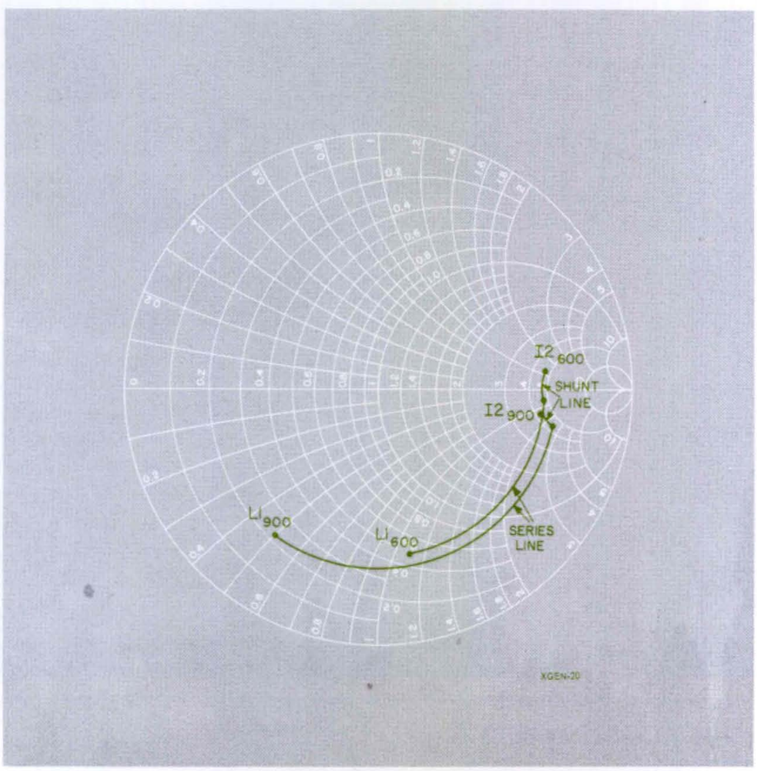


Figure 10. Admittance transformation by the interstage network.

Through a pencil-and-paper cut-and-try procedure, we arrive at an output network with a shunt-line length of 4.2 cm (electrical) and a series-line length of 4.7 cm. The admittance transformation performed by the output network is plotted on the Smith chart of Figure 9, which is normalized to 8 mmhos (125 ohms). The points $L2_{600}$ and $L2_{900}$ represent the loads that the network presents to the transistor at 600 and 900 MHz, and $L(\mathscr{G}_{max})_{600}$ and $L(\mathscr{G}_{max})_{900}$ are the load admittances for maximum gain at the two frequencies. When the admittances represented by $L2_{600}$ and $L2_{900}$ are renormalized to y_{22} and plotted on the charts of Figures 8 and 5, respectively, the gains that are found are 6 dB at 600 MHz and 5.1 dB at 900 MHz. Although we could probably do a better job of matching with higher-impedance line sections or with a more complicated network, we shall be content, for the purposes of our example, with these results.

We select an interstage network with shunt-line lengths that are again 4.2 cm and a series-line length of 5.25 cm. The transformation is plotted in Figure 10. The points $I2_{600}$ and $I2_{900}$ are the input admittances of the second stage, found from the G_{in}-B_{in}-grids of Figures 8 and 5 and normalized to 8 mmhos. The points $L1_{600}$ and $L1_{900}$ are the load admittances presented to the first transistor. When we renormalize $L1_{600}$ and $L1_{900}$ to y_{22} and plot them on Figures 8 and 5 respectively, we find gains of 4.9 dB at 600 MHz and 5.7 dB at 900 MHz for the first stage.

Figure 7 does not show a series-line section at the input of the amplifier. This is because we can do without an admittance-transforming series section at the input if we will tolerate a fairly small amount of mismatch loss. We find the admittances presented by the base of the first transistor by means of the G_{in}-B_{in}-grids of Figures 8 and 5. Adding the susceptance of the 4.2-cm input shunt line, we calculate that the mismatch-loss penalty for connecting the input directly to a 50-ohm source is 0.5 dB at 600 MHz and 0.7 dB at 900 MHz. If we accept these losses, the over-all gain of the amplifier is 10.4 dB at 600 MHz and 10.1 dB at 900 MHz.

This simplified example shows that, even with very simple matching networks, amplification over quite wide frequency ranges can be achieved with gains reasonably close to the optimum gain of the device at the highest frequency in the passband.

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

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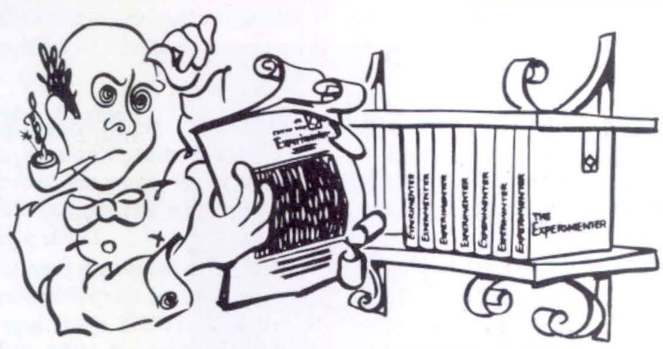

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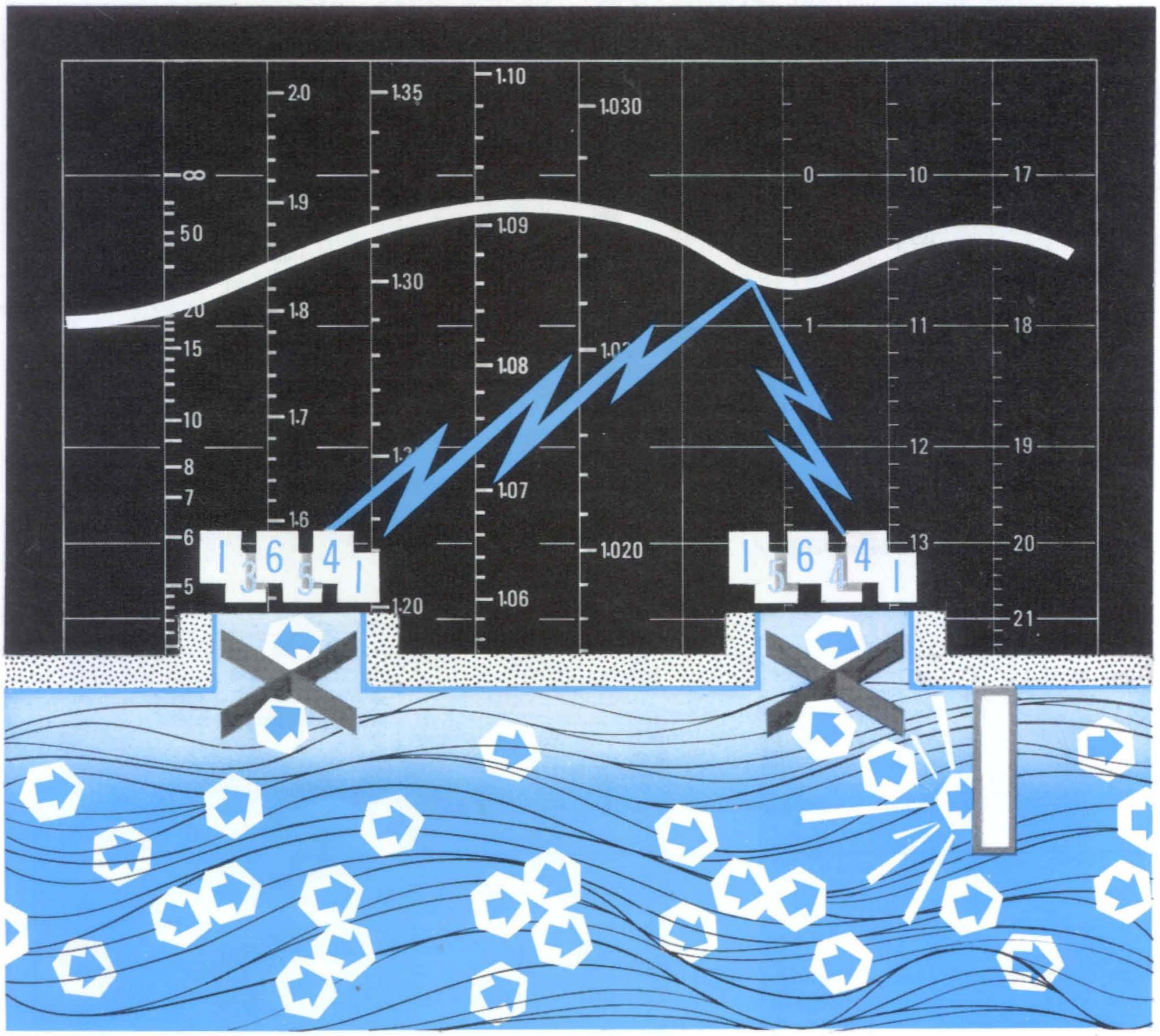

THE GENERAL RADIO
Experimenter

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

THE COVER Broadband testing, using sweep-frequency reflectometry techniques, has been accepted as a time-saving, cost-reducing tool in research and industry. Our cover illustrates, in an abstract way, the ease of applying reflectometry techniques, the simplicity of the directional coupler which transmits the basic information, and the clarity of the observed parameter in the form of an oscillogram upon the graticule.

THE NEW LOOK

We hope you were pleasantly surprised at the look of the new *Experimenter* when it arrived at your desk in February. The change in size and format is evidence that General Radio recognizes its obligation to readers of the *Experimenter* to present information related to new measuring instrumentation, or improvements in previously issued models, in a form both provocative and attractive. An obligation exists also to supply technical information that is worthy of retention in your files.

Looking back through the years and reading Vol. 1 No. 1 issued in June, 1926, I was pleased to note that the *Experimenter* was stipulated to be a new General Radio *service* to experimenters in home laboratories. Service still is our theme but the direction has changed from the home laboratory of the radio hams to the laboratories of science and industry, to the production areas serving commerce and trade, and to the several echelons of measurement accuracy represented by standards and calibration laboratories.

The *Experimenter* has a significant role to play in advancing the state of the art of measurement engineering. At GR we are aware that many of our readers are part of the task forces that generate so much of what is new, useful, and available in the measurement and control fields. We hope to draw upon the experiences of our readers for some of the source material from which editorials will be generated in the future. The Editor's office at GR is prepared to receive any comments you feel will be of benefit to our readers. Spring is here – let's get together as a team and "Play Ball!"

C. E. White
Editor

The New Sweep-Frequency Reflectometer

An integrated system for direct-reading reflectometry measurements, which is remotely programmable, capable of simultaneous dual-channel presentations, and easy to operate. Expandable SWR and loss scales permit in-depth exploration of resonance bandwidths, perturbations, and residual reflections. Use of GR900® components significantly reduces inherent system errors and establishes an extremely low residual SWR.

by T. E. MacKenzie, J. F. Gilmore, M. Khazam

INTRODUCTION

The GR 1641 Sweep-Frequency Reflectometer operates in the *frequency domain*. Using a sine-wave signal to the unit being tested, it measures total reflection (SWR) and transfer (insertion loss) as a function of frequency.

This method should not be confused with the pulse-echo method (time-domain reflectometry), originally used for detection and location of faults in cables. The pulse-echo method employs a visual display which presents individual reflection locations as *magnitude* versus *distance*. The frequency-domain-reflectometer display presents *net reflections* versus *frequency*.

The introduction of low SWR connectors and the advance in performance of directional couplers have encouraged the move from fixed frequency measurements in coaxial circuits to sweep-frequency techniques. Sophisticated systems now operate over broader bandwidths and to tighter specifications. Some complications remained, however, including the need to establish calibrated reference levels for each measurement and a lack of direct SWR readout.

The GR 1641 was designed to eliminate such faults. It provides broad frequency coverage, 20 MHz to 7.0 GHz, in two bands by means of two integrated rf units. Band changing is simple; no time is lost in dismantling or reassembling components. Initial self-calibration is not repeated for an ensuing series of SWR or loss measurements within the instrument's range. High order of accuracy is assured by directivity greater than 43 dB at 1.0 GHz and 37 dB at 7.0 GHz plus SWR (for equivalent source and for the detector) less than 1.03 at 1.0 GHz and 1.06 at 7.0 GHz. All this is available at moderate cost.

For test purposes, only two panel controls require adjustment by the operator – the functional DISPLAY switch and the RANGE switch. The former controls measurements of SWR, INSERTION LOSS, or both; the latter establishes SWR readings at full scale equal to ∞, 2.0, 1.35, 1.10, and 1.03 or full-scale losses of 0, 10, 17, 27, and 37 dB. The controls are programmable for remote operation.

Test data are read directly from the panel meter in the operational modes corresponding to fixed frequency, stepped frequency or slow sweep. In the continuous sweep mode, direct and simultaneous indications of SWR and loss are presented on an auxiliary oscilloscope.

The 1641 is shown in use in Figure 1 with commercially available auxiliary equipment types which are in the normal

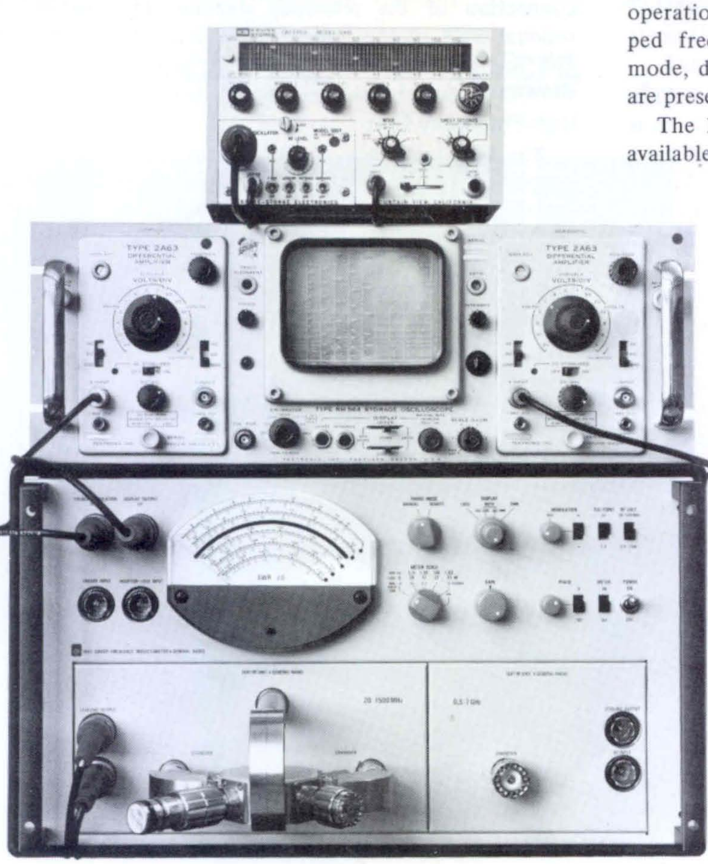


Figure 1. A test arrangement using GR 1641 Reflectometer.

Figure 2. Reflectometer panel.

complement of a testing laboratory. The assembly comprises the 1641, an external rf source, and an external oscilloscope. Figure 2 shows the panel of this reflectometer, with the low-frequency rf unit (20 to 1500 MHz) on the bottom left, the high frequency rf unit (0.5 to 7.0 GHz) on the bottom right; both are mounted in the main frame with the indicator unit. The transfer detector is separate for flexibility. The device to be measured is connected between the UNKNOWN port of the 1641 and the transfer detector, the detected output of which is connected to the 1641 through a cable.

SYSTEM DESCRIPTION

The interconnection of the various components of the reflectometer system for SWR and loss measurements is shown in Figure 3. The sweep source is modulated by a 10-kHz square-wave signal from the 1641. The detected return-loss and insertion-loss signals are fed to the indicator, basically a 10-kHz tuned amplifier, the output of which is displayed on the oscilloscope after rectification. The blanking pulse of the sweep source is fed to the indicator to provide a zero-retrace output level, and also to trigger the channel switch for an alternate display of return loss (or SWR) and insertion loss with each sweep.

Low-Frequency Unit *

The low-frequency unit consists of a balun and a summing junction, which together form a bridge circuit. Leveling is accomplished by means of a detector at the balun junction. A second detector provides the SWR information. Simple conversion of the bridge for comparison measurements is made possible by the external connection of the reference standard termination. The residual SWR of the bridge is less than 1.015 to 1.0 GHz and less than 1.02 to 1.5 GHz. A schematic diagram is shown in Figure 4a.

High-Frequency Unit

The high-frequency rf unit consists of two directional couplers, two detectors, and appropriate interconnecting transmission lines. A schematic diagram of this unit is shown in Figure 4b. The first directional coupler (the leveling coupler) couples part of the input signal to the leveling detector, which levels an rf source when the 1641 is

*Patent Pending.

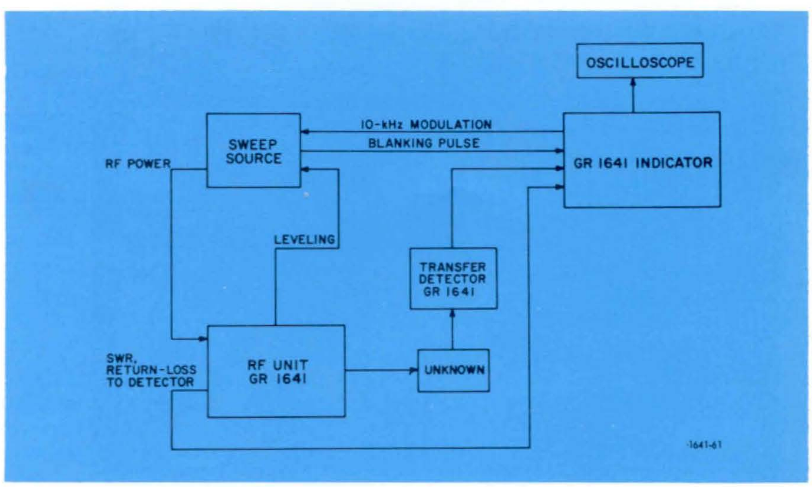


Figure 3. Interconnections of test equipment.

operating in a sweep-frequency mode. The remainder of the input signal goes to the measuring coupler where part is coupled to the UNKNOWN connector and the rest is absorbed by the built-in termination.

The directivity of the measuring coupler is greater than 40 dB from 0.5 to 4.0 GHz, and greater than 37 dB from 4.0 to 7.0 GHz. The coupling characteristics of the measuring and leveling couplers are identical so that the magnitude of the signal reaching the UNKNOWN connector follows the frequency response of the leveling detector.

The high directivity, wide-band directional couplers are the heart of the rf unit. These directional couplers have an asymmetric coupling region of three sections. The theoretical design of such units to achieve a desired coupling characteristic is well known and results have been tabulated.[1]

While achieving a desired coupling response is a relatively simple matter, the high directivity of these coupler units requires the utmost care in design and fabrication. The wide bandwidth covered by these units makes it inevitable that any characteristic impedance errors or discontinuities will add at some frequency, so all components must be designed and made properly. Thus the built-in termination, support beads, connector and line-size transition must all have extremely low SWR's, and the coupling-region dimensions must be nearly perfect to achieve this level of performance. These characteristics have been attained in the 1641, resulting in a measuring coupler with substantially higher directivity than is available in any other commercial multi-octave coupler.

[1] Levy, R., "Tables for Asymmetric Multi-Element Coupled-Transmission-Line Directional Couplers," *IEEE Transactions on Microwave Theory and Techniques,* May 1964, pp 275-279.

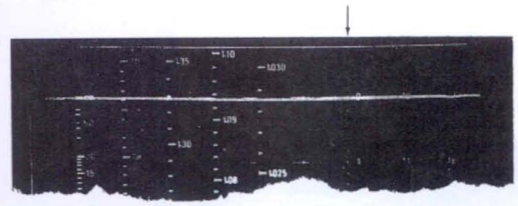


Figure 5. Typical calibration curve for insertion-loss measurements in 2.0- to 4.0-GHz range; full scale = 0 dB.

Transfer Detector

Both the external transfer detector, used for insertion-loss measurements, and the built-in SWR detector are selected to have the same frequency response as the leveling detector. Both have a 10-dB attenuator built in to improve impedance match. The maximum power level at the detectors for an on-scale indicator reading is approximately −10 dBm. It is impossible, while maintaining an on-scale reading, to drive the detectors outside the square-law region. The leveled power required from the source is −10 dBm plus the dB-coupling factor of the directional coupler.

Data Presentation

A typical calibration curve for insertion-loss measurements is shown in Figure 5. The flatness of this curve depends mainly upon the similarity of the coupling characteristics of the leveling and measuring couplers and upon the tracking of the leveling and the transfer detectors.

Typical calibration curves for SWR measurements are shown in Figure 6. The over-all flatness of these responses

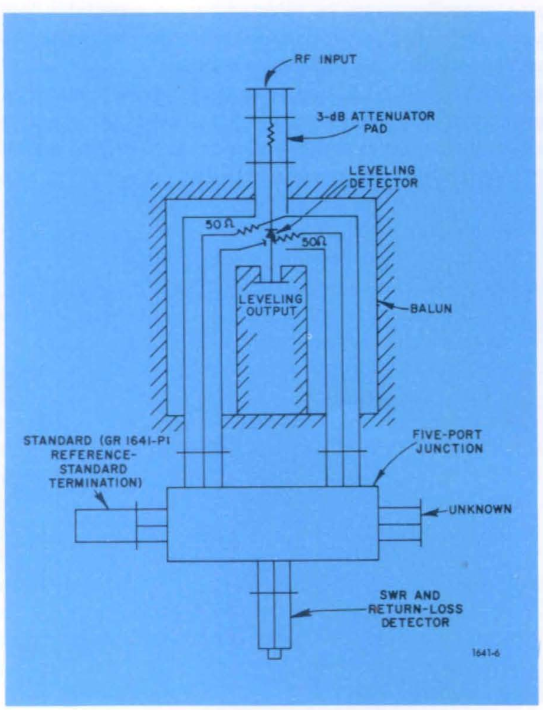


Figure 4a. Schematic diagram of low-frequency rf unit.

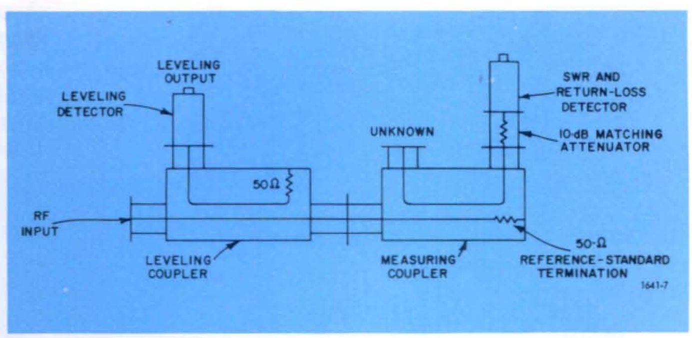


Figure 4b. Schematic diagram of high-frequency unit.

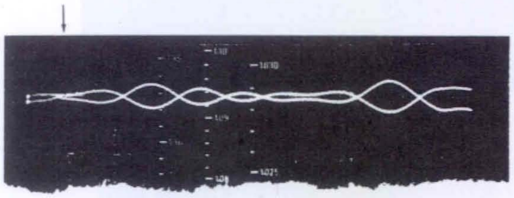


Figure 6. Typical calibration curves for SWR measurements in 2.0- to 4.0-GHz range: full scale = ∞.

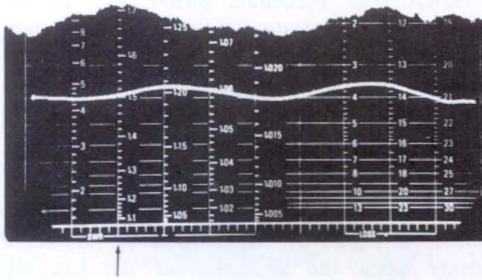


Figure 7. SWR measurement of 1.5:1 mismatch in the 2.0- to 4.0-GHz range; full scale = 2.0.

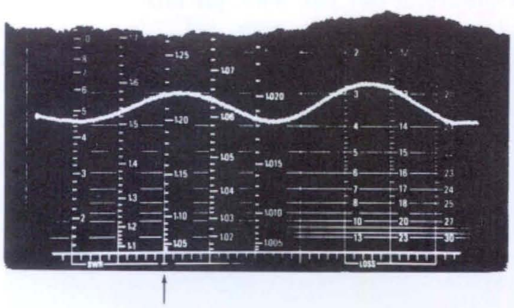


Figure 8. SWR measurement of 1.2:1 mismatch in the 2.0- to 4.0-GHz range; full scale = 1.35.

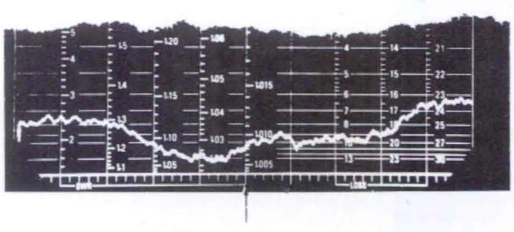


Figure 9. Residual SWR of typical rf unit from 2.0 to 4.0 GHz; full scale = 1.03.

depends upon the similarity of the coupling characteristics of the two couplers and upon the tracking of the SWR and the leveling detectors. The ripple in these traces and the difference between the open- and short-circuit curves depend mainly on the equivalent source match which, at the UNKNOWN connector, is less than $1.01 + 0.007 f_{GHz}$.

SWR measurements of typical standard mismatches are shown in Figures 7 and 8. These curves illustrate typical system performance at SWR levels of 1.5 and 1.2 respectively. The extremely low residual SWR of a typical high-frequency rf unit is shown in Figure 9.

Design

Figure 10 shows a block diagram of the GR 1641 Indicator. Two identical preamplifiers are connected to the main amplifier chain through the channel switch. The channel selection is accomplished through the DISPLAY switch, which has three positions: SWR, LOSS, and BOTH. In the BOTH position, the channel-switching circuit controls the channel switch; blanking pulses from the sweep generator cause alternate presentation of the preamplifiers to the main amplifier chain. The result is an alternate display of the SWR and LOSS functions with subsequent sweeps.

The detector is a synchronous detector of the sample-and-hold type. Gate pulses, derived from the 10-kHz oscillator, connect a charging capacitor to the output of the final amplifier for a duration of approximately three microseconds. A phase-shift network in the amplifier chain is adjusted so that the gate pulses coincide with the positive peak of the 10-kHz signal at the final amplifier. The value of the charging capacitor determines the response time of the detector and the over-all bandwidth of the indicator, since this is a synchronous detector. The value of the charging capacitor must be selected for an acceptably low noise-output level; the sweep speed must then be limited to a value compatible with the response time.

The METER SCALE switch controls the attenuator in the main amplifier chain and also the value of the charging capacitor. This relieves the operator from having to search for the optimum compromise between output noise and sweep speed; it is only necessary to set the sweep time to a value greater than or equal to the minimum recommended sweep time indicated by the meter-scale control.

Usually in systems that use diode detectors, for example, the 1641, the operator must take special precautions to prevent the detectors from functioning outside the square-law regions, thereby introducing gross errors. This is not so with the 1641 indicator. The indicator's sensitivity range is selected so that, for on-scale meter readings, the source output power must be set to a level compatible with that required for detector square-law operation. If excess power is available, the measurement range can be extended by this amount by recalibrating the instrument for SWR, with a standard mismatch other than an open or short, or for insertion loss, with a precision attenuator. For recalibration with a standard mismatch of 2:1 SWR ratio, a switch is provided that modifies the attenuator switching sequence of the indicator to maintain the meter SWR calibration for extended range operation.

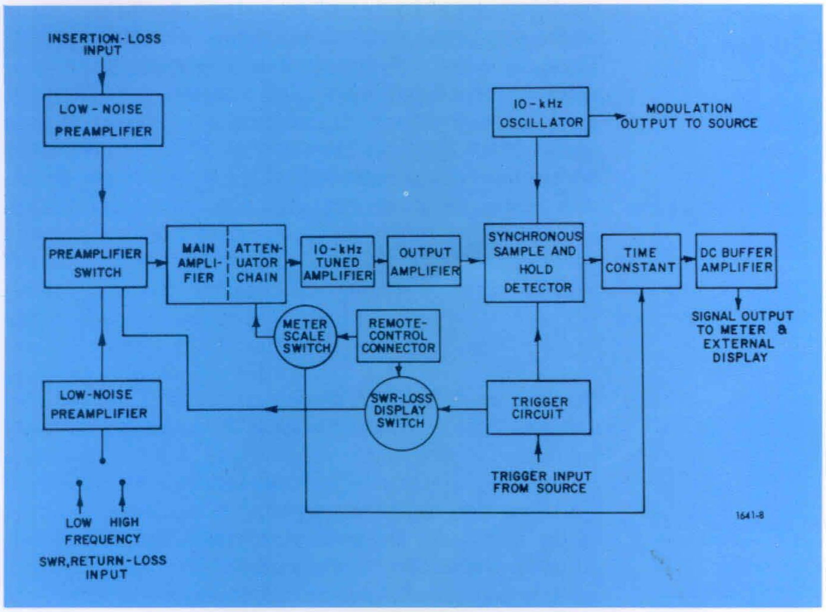


Figure 10. Block diagram of 1641 Indicator.

The diode detector, as generally used with low-frequency tuned amplifiers, has a useful operating range of approximately 30 dB. This range is limited at low signal levels by the noise of the system and at high signal levels by the deviation from square-law operation. One method to extend the square-law range is to shunt the detector output with a properly chosen resistor. Figure 11 shows the measured deviation from square-law for a diode detector as a function of input power for three values of the load resistance. The optimum value of the load resistor is approximately $\frac{1}{2}R_d$, where R_d is the dynamic resistance of the diode. When a low resistance dc return path is present, as through a choke, the optimum load resistance is approximately equal to R_d. The advantage of an increase in dynamic range, brought about by this method, is offset somewhat by the disadvantage that there is a reduction in sensitivity caused by the addition of the shunt resistor.

In the 1641 indicator design, a dc return through a choke is presented at the input. The optimum load resistance is provided by the adjustable input impedance of the preamplifier. The disadvantage mentioned above is thus bypassed. The wide dynamic range of the 1641 indicator system is achieved by the extension of the detector square-law range at high signal levels and by the reduction of the indicator bandwidth to a very small value in the most sensitive measurement range. This narrow bandwidth is made possible through the use of the synchronous-detector technique previously described.

The SWR-loss display and meter-scale control functions can be remotely controlled by contact closures to ground

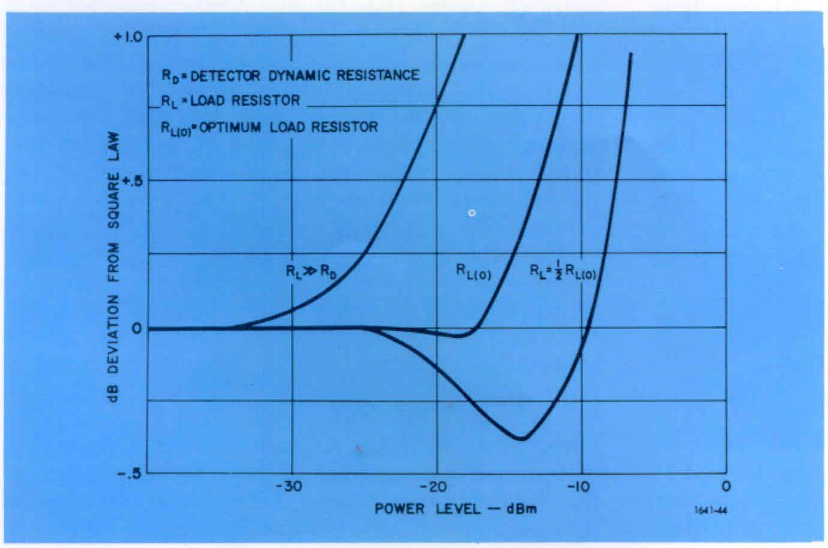


Figure 11. Diode detector — square-law range extension.

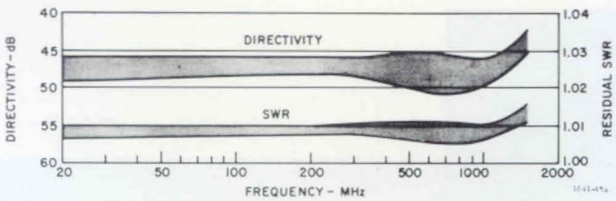


Figure 12a. Typical spread of directivity and SWR data for low-frequency rf unit.

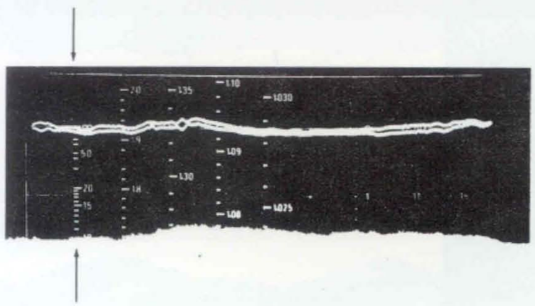


Figure 12b. Dual presentation of typical response data, using open- and short- circuit terminations in 0.5- to 1.0-GHz range; full scale = ∞.

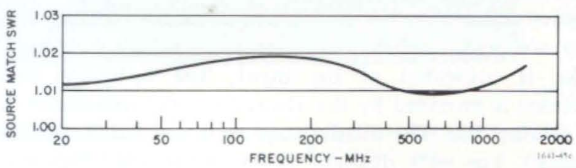


Figure 12c. Residual source match to unknown based on measurements of a unity reflection coefficient with varying reflection phase — low-frequency rf unit.

through electronic or mechanical switches. This allows the 1641 to be used in automatic measuring systems for sorting or qualifying devices.

The effect of ground currents in a measuring system that requires interconnections of several instruments had to be considered in the design of the indicator. Relatively large 10-kHz currents can flow through parts of the ground circuit, causing small voltage differences between various ground points. If the ends of the input cable shield are connected to two such points through low-impedance circuits, the resulting voltage difference between the ends of the input cable shield will appear in series with the measuring signal at the amplifier input. This can cause large errors in the measurement of low SWR or high insertion-loss values. This error is reduced to a negligible level in the 1641 indicator by connecting only one end of the input cable shield directly to the frame and by keeping the impedance between the other end and the frame high with respect to the shield resistance.

ACCURACY CONSIDERATIONS

The high degree of accuracy that has been obtained by use of the 1641 is a result of using components based on the precision of the GR900® connector. In fact, the total error of the 1641 system is no greater than the component error introduced by use of general-purpose connectors in other commercial systems. A consideration of the system errors, however, will provide illustrations of the limits of performance, as well as providing a greater insight into the 1641 system operation.

The main error contributions for reflection and transmission measurements with the 1641 are described approximately by:

$$|\Gamma_i| = k\,|\Gamma_x(1 + \Gamma_s\Gamma_x) + \Gamma_0 + \Gamma_\ell\,\tau_x\,\tau_x'| \qquad (1)$$

$$|\tau_i| = k\,|\tau_x\,(1 + \Gamma_s\Gamma_x + \Gamma_\ell\,\Gamma_x') + \tau_0| \qquad (2)$$

in which Γ_x is the true reflection coefficient.

τ_x is the true transmission coefficient of the measured device.

Γ_i and τ_i are the reflection and transmission coefficients indicated by the 1641 system.

Γ_0 and τ_0 are the residual reflection and transmission coefficients of the system respectively.

Γ_s is the equivalent source match and

Γ_ℓ is the transfer-detector match of the system, both expressed as reflection coefficients.

Γ_x' and τ_x' are the output reflection coefficient and the reverse transmission coefficient of the device being measured.

k is the frequency response characteristic of the system normalized to unity.

The relations above are based upon a two-port device, but they are easily expanded to apply for multiports. For a one-port device, the third term in equation (1) goes to zero and equation (2) is inapplicable.

The directivity or equivalent residual SWR of the system (Γ_0 when expressed as a reflection coefficient) is important to all reflection measurements; the residual detector match to the unknown (Γ_ℓ when expressed as a reflection coefficient) is important to low-reflection measurements on low-loss devices and to transmission measurements on high-SWR devices. (For the most accurate low-SWR measurements on multiports, auxiliary reflectionless terminations may be employed.) The equivalent residual source match to the unknown (Γ_s when expressed as a reflection coefficient) is important to both reflection and transmission measurements on high-SWR devices. Figures 12a, b, and c show typical data for the low-frequency rf unit (20 to 1500 MHz). Values of Γ_0, Γ_s, and Γ_ℓ are given in the published specifications of the 1641.

The residual-transmission coefficient (τ_0) is essentially the noise level of the system. It is dependent on the available source power and, for a source power of 30 mW, τ_0 is typically less than 0.005 (46 dB). The system flatness or leveling, characterized by k in equations (1) and (2) is dependent upon both the 1641 and the rf source employed. The principal causes of deviations in k from unity are distortion of the system modulation in the source leveling circuitry and tracking errors in the frequency responses of the 1641 detectors. Figure 13 shows a typical trace of system flatness for the low-frequency range.

The above discussion of accuracy assumes negligible harmonic content in the source signal. Presence of harmonics will affect the 1641 system in at least two ways; leveling across a band of frequencies is directly dependent upon the harmonic content and the leveling and measuring detectors respond to the net source signal (fundamental plus harmonics). The relative rf phases of the fundamental and of the harmonics at the two detectors may differ, however, and this can cause variations in a normally flat trace across the observed frequency band. The variations usually are not smooth but appear as an amplitude perturbation of the order of a few tenths of a decibel. The use of low-pass filters in the source-signal circuit, when necessary, will eliminate this problem.

A second effect of the presence of harmonics is apparent in the measurement of devices such as band-stop networks that have high loss at the fundamental frequency and low loss at the harmonic frequencies. The effect is also observed in the measurement of band-pass networks that have a low SWR at the fundamental frequency and a high SWR at the harmonic frequencies. The ratio of harmonic level to signal level, in these cases, is increased by the characteristics of the device being measured and results in a requirement for more stringent filtering of the source signal.

TYPICAL APPLICATIONS

The GR 1641 Reflectometer has wide application in production-test facilities; development, research, and calibration laboratories; incoming inspection and quality-control activities. It measures one-port, two-port and multiport devices, both passive and active, with bidirectional or unidirectional properties. With GR900 precision coaxial adaptors, accurate measurements can be made on devices equipped with a wide variety of connectors. Information can be obtained on a fixed- or swept-frequency basis from a meter, oscilloscope or recorder presentation, with go-no-go limits easily established. Since the 1641 is programmable, it can be controlled from a remote station, either manually or via a computer.

All the measurements made with the reflectometer fall into either the reflection-coefficient (SWR) or transmission-coefficient (insertion loss) category, but these two categories encompass an extensive list of characteristics. The reflection-coefficient or SWR category includes return loss and percent impedance deviation. The transmission-coefficient or insertion-loss category includes isolation, attenuation, coupling, directivity, gain, and frequency response. Most of the important characteristics of filters, antennas, isolators, circulators, switches, power dividers, couplers, amplifiers, attenuators, cables, and terminations are included in these two categories.

The 1641 provides a convenient means to measure the SWR of antennas, to search for resonances, and to characterize components or semiconductor devices. The following paragraphs describe some specific measurement examples and areas of application.

In 1962 T. E. MacKenzie joined GR's Microwave Group. Previous employment was with the Alford Manufacturing Company from 1954 to 1962. He holds degrees of BSEE (1958) and MS in Physics (1963) from Northeastern University. His work at GR has been primarily in the development of microwave instruments, components, and standards.

J. F. Gilmore joined General Radio as an engineer in the Microwave Group in 1963, after receiving his BSEE in 1961 and MSEE in 1963 from Northeastern Univeristy. He is presently engaged in microwave circuit and component design. He is a member of IEEE.

After receiving his EE degree in 1957 from the Delft University of Technology in Holland, M. Khazam was a project engineer with the Laboratory for Electronic Developments for the Dutch Armed Forces. He joined GR's Engineering Department in 1962 and presently is specializing in the development of uhf-vhf instruments and components.

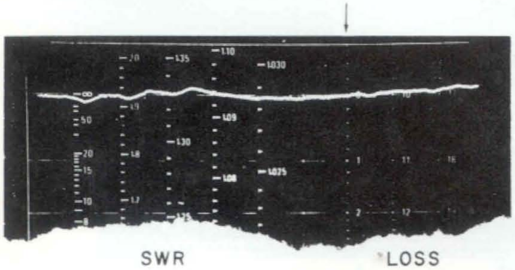


Figure 13. 1641 system-response flatness — low-frequency 0.5- to 1.0-GHz range; full scale = 0 dB.

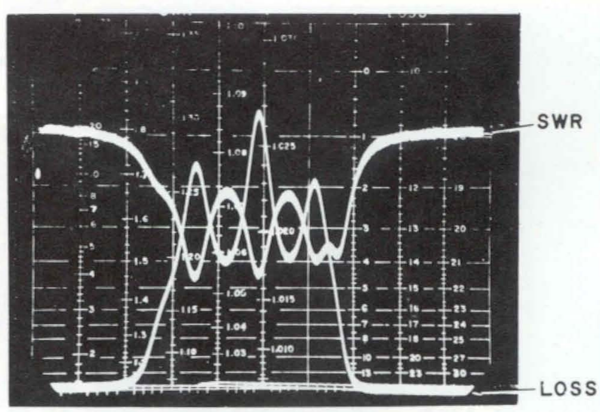


Figure 14. SWR and loss characteristics of a 5-stage bandpass filter in the 463- to 473-MHz range; full scale: ∞ (SWR) and 10 dB (loss).

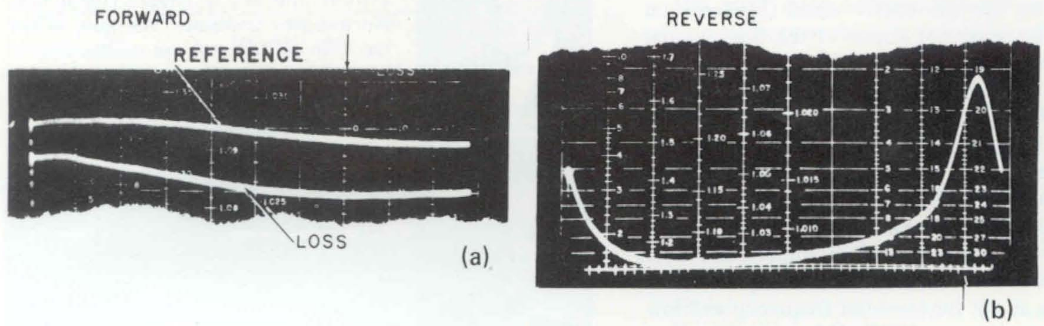


Figure 15. Insertion loss for isolator in the 2.0- to 4.0-GHz range; full scale: 0 dB in 15a and 17 dB in 15b.

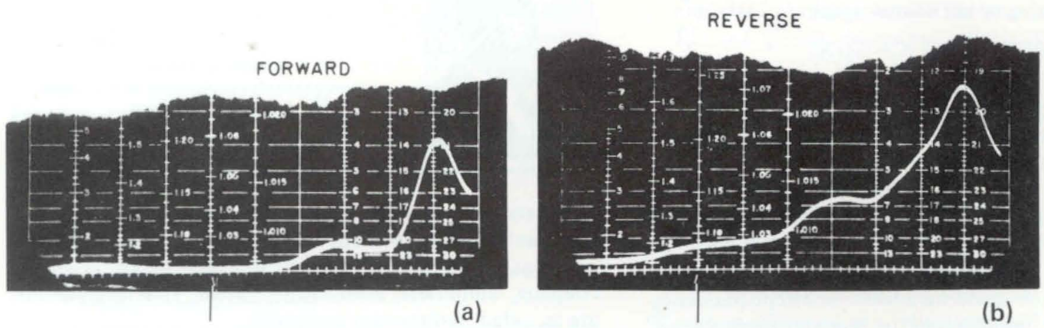


Figure 16. SWR for isolator in the 2.0- to 4.0-GHz range; full scale = 1.35.

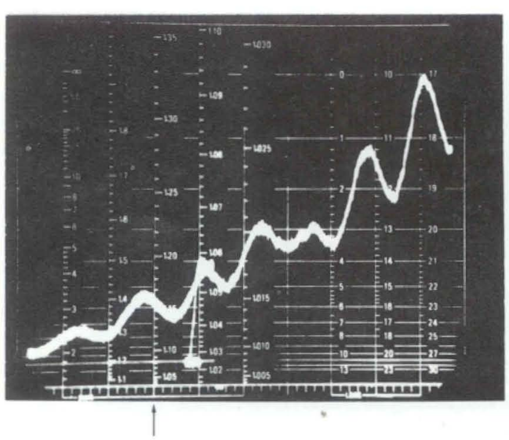


Figure 17. Video detector SWR in the 2.0- to 7.0-GHz range; full scale = 1.35.

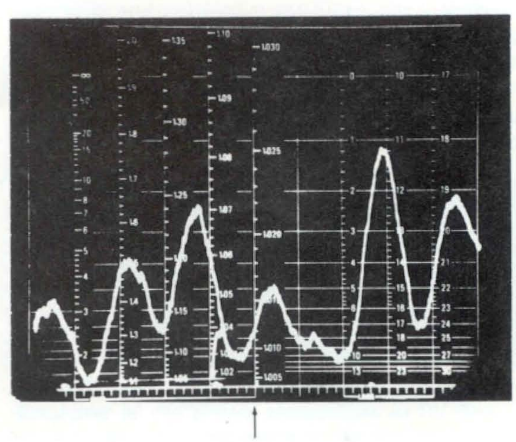


Figure 18. Termination SWR in the 2.0- to 7.0-GHz range; full scale = 1.03.

Microwave Components Measurements

Figure 14 shows the measured passband characteristics of a five-stage bandpass filter over the 463- to 473-MHz band. Since both the transmission-response and SWR characteristics can be displayed simultaneously, the effects of peaking in the individual sections can be quickly and easily determined.

Figures 15 and 16 show the measured characteristics of an isolator over the 2- to 4-GHz band. The forward characteristics were measured with the isolator input connected to the rf unit UNKNOWN port. The reverse characteristics were measured with the isolator output connected to the rf unit UNKNOWN port. These measurements are typical of those required on circulators, switches, power dividers and attenuators.

Figure 17 shows the SWR of a video detector over the 2- to 7-GHz range and Figure 18 shows the SWR of a termination over the same range. The full-scale SWR range in the first case is 1.35, in the second 1.03. The ripples in the traces indicate that the system residual SWR is less than 1.02.

Figure 19 shows the reject-band loss of a nominally 2-GHz strip-line low-pass filter. The spurious responses in the vicinity of 5 GHz are the result of leakage coupling. The full-scale loss is 17 dB.

The data shown in Figures 17, 18, and 19 were obtained using the system of Figure 20, which comprises the 1641, an Alfred Model 9510 Multi-octave Sweep Oscillator and a Tektronix Type 564 Storage Oscilloscope. The rf connections between the reflectometer and the oscillator plug-ins were made through a coaxial switch. To coordinate the triggering of the oscillator plug-ins and the rf switch with the horizontal-sweep voltage to the oscilloscope, the oscillator-blanking output was used (through a separate switching network). It was also used to switch the leveling connections. The modulation connections were made in parallel. The system was operated continuously with a sweep time of 10 seconds.

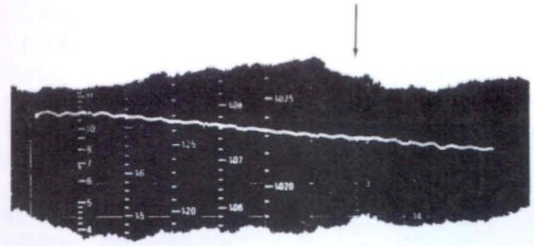


Figure 21. Insertion loss for 12.5 feet of RG-58/U cable; full scale = 0 dB.

Cable Measurements

Figure 21 shows the attenuation characteristic of an RG-58/U cable assembly about 12.5 electrical feet long. Figure 22 shows the SWR characteristic. These measurements were taken by inserting the cable between the UNKNOWN port and the transfer detector of the 1641. The frequency range of measurement is 1 to 2 GHz. The attenuation characteristic decreases uniformly from 1.4 dB to 2.1 dB. The ripple in the SWR characteristic indicates that the main sources of SWR error are about 12.5 feet apart (a half wavelength at the ripple-rate frequency); in fact, the SWR is principally caused by the discontinuities at the cable-connector junctions.

In many instances, it is not possible to attach a detector to the far end of a cable – for example, if the cable or transmission line is long or passes through a bulkhead or is

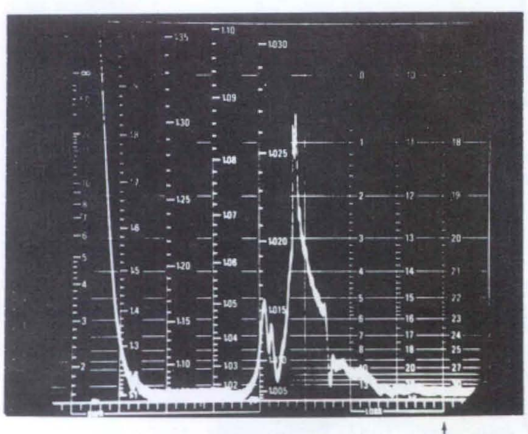


Figure 19. Reject-band loss of 2-GHz strip-line low-pass filter; full scale = 17 dB.

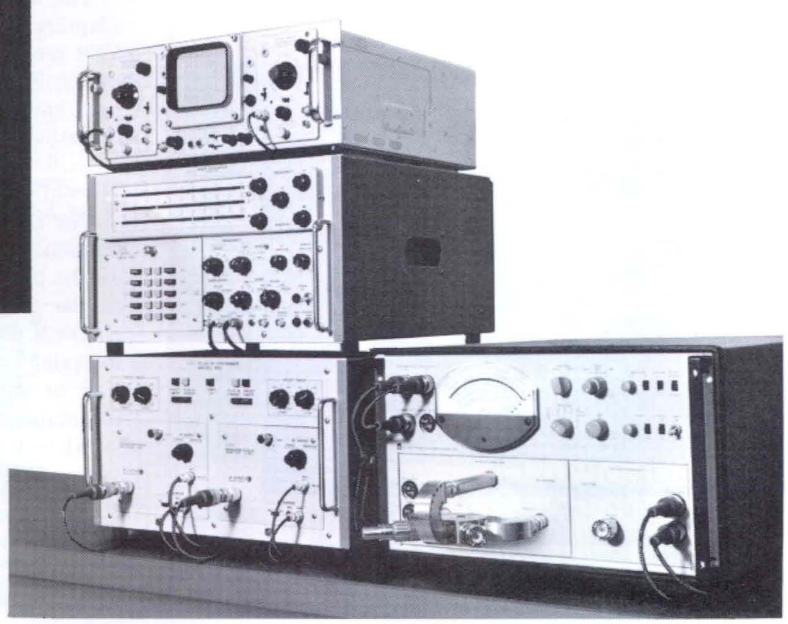


Figure 20. Test assembly for Figures 17, 18, and 19.

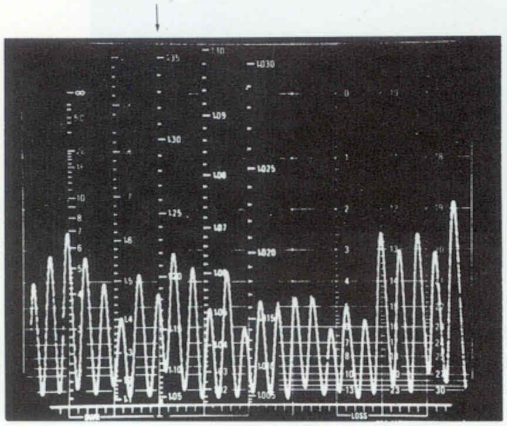


Figure 22. SWR for 12.5 feet of RG-58/U cable; full scale = 1.35.

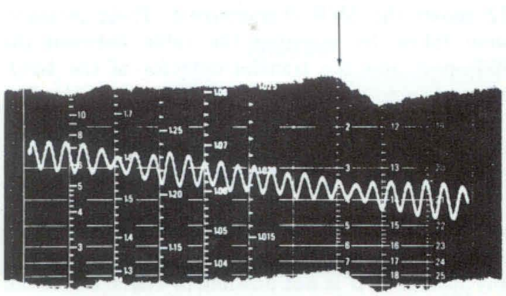


Figure 23. Return loss for 12.5 feet of RG-58/U cable open circuited; full scale = 0 dB.

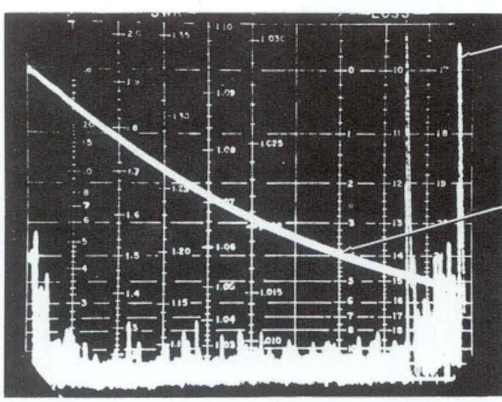


Figure 24. Insertion loss and SWR for 50 feet of RG-214/U cable in the 2.0- to 4.0-GHz range; full scale = 1.10 (SWR) and 10 dB (loss).

attached to an antenna mast. Measurements of the loss and SWR characteristics still are of interest. Such measurements can be accomplished by leaving the far end of the cable open-circuited (or short-circuited) and by looking at the return-loss or reflection characteristic only. This is illustrated in Figure 23 for the same cable as that of Figures 21 and 22.

The attenuation is just one-half of the average indicated return loss because the measured signal travels down and back the length of the cable. The ripple in the return-loss characteristic is a measure of the cable input SWR, which is given approximately by the voltage ratio corresponding to the ripple excursion in dB. This measurement is limited to the degree that the value given by twice the cable attenuation must be at least 10 dB less than the terminated-cable return loss. If this is not the case, then the attenuation information is lost. For example, if the cable attenuation is 15 dB, twice the cable attenuation is 30 dB; the attenuation information therefore is submerged in the cable return loss. In fact, for long, lossy cables (20-dB attenuation or greater), the SWR can be read directly with the cable not terminated. In such cases, the cable attenuation can only be determined by utilizing the transfer detector at the far end of the cable.

Figure 24 shows the attenuation and SWR characteristics of a 50-foot length of RG 214/U cable over the frequency range of 2 to 4 GHz, measured directly with the transfer detector. The cable is equipped with GR900 cable connectors, and the SWR is very low except at the points of resonance. The resonances are caused by periodicity in the cable itself. The characteristics in the vicinity of one of the resonances are shown in detail in Figure 25. Long cables, such as this, exhibit SWR's that vary quite rapidly with frequency, and resonances such as the one illustrated are not uncommon.

Low-SWR Measurements

The SWR's of connectors, whether they are on cables, adaptors or air lines, are usually quite low, and appreciable SWR resolution is required to achieve meaningful measurements. Figure 26 shows a series of SWR curves of adaptor pairs and cables, all on the 1.03 full-scale SWR range. The frequency band is 250 to 500 MHz.

Tuned Reflectometry

For the most accurate low-SWR measurements at fixed frequencies, the system residual directivity or residual SWR can be tuned out with an impedance matching tuner (such as the 900-TUA or -TUB Tuner) attached to the UNKNOWN port of the 1641. The tuner is adjusted so a null is indicated on the 1641 while measuring a standard termination or while using quarter-wavelength or other air-line techniques.[2]

When a source is employed that delivers approximately 100 mW leveled power (into 50 ohms), additional SWR

[2] MacKenzie, T. E. "Some Techniques and Their Limitations as Related to the Measurement of Small Reflections in Precision Coaxial Transmission Lines," *IEEE Transactions on Instrumentation and Measurement*, Vol. IM-15, No. 4, December 1966. (Available from GR as Reprint No. A133)

resolution can be obtained by calibrating initially at a 2.0-SWR level and utilizing a 1.01-SWR full-scale range. Under these conditions, the residual noise level is reduced to an equivalent SWR approximating 1.002. (Note – the 1.01 scale is obtained by using the 1.10 scale and manually inserting a zero after the test-data decimal point.)

If speed is required, in addition to the high accuracy described herein, pre-set tuners can be employed. In this manner, production measurements at discrete frequencies can be made with a minimum of set-up time, and with direct readout.

Active Devices

Use of the 1641 to measure the characteristics of active units, such as solid-state semiconductor devices, is illustrated in Figure 27, with measurements on a GR 1237 VHF/UHF Preamplifier. The gain and input SWR measurements of Figure 27a were obtained by connecting the amplifier input directly to the 1641 UNKNOWN port, and

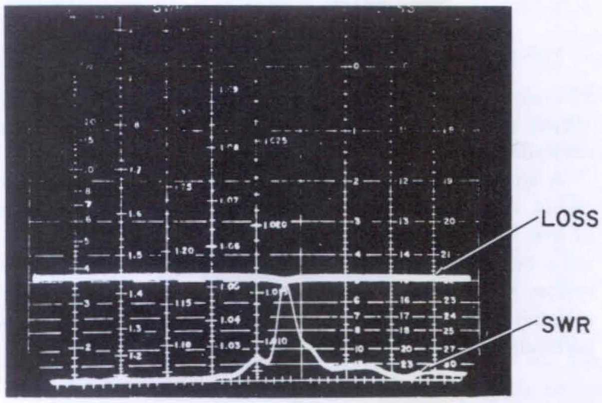


Figure 25. Detailed expansion of 3.75-GHz resonance in Figure 24; full scale = 1.35 (SWR) and 10 dB (loss).

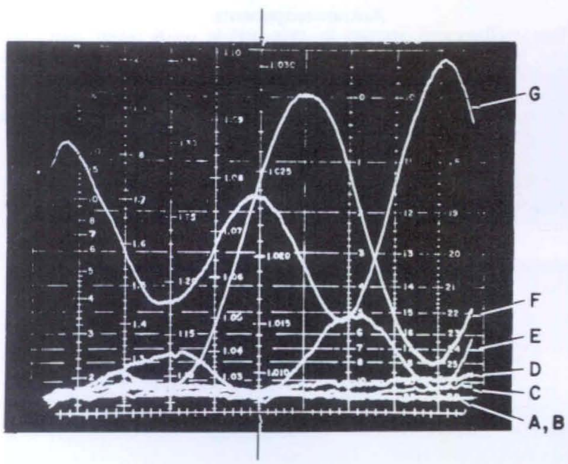


Figure 26. SWR curves for assorted adaptor pairs and cables in the 250- to 500-MHz range; full scale = 1.03. Curve A shows the response with 900-W50 Termination connected at UNKNOWN port of rf unit, curve B the response with a pair of 900-Q874 Adaptors inserted between termination and UNKNOWN port. Curve C indicates response after substitution of 900-QN Adaptors for 900-Q874 Adaptors, curve D after substitution of 900-QMM Adaptors for 900-QN units. Curve E shows the response of 3 feet of RG-214/U cable equipped with GR874® locking connectors, specially chosen to illustrate optimum response obtainable; curves F and G illustrate normal responses obtainable for 874-R20LA and 874-R22LA Patch Cords (3-feet) respectively, each equipped with GR874 locking connectors.

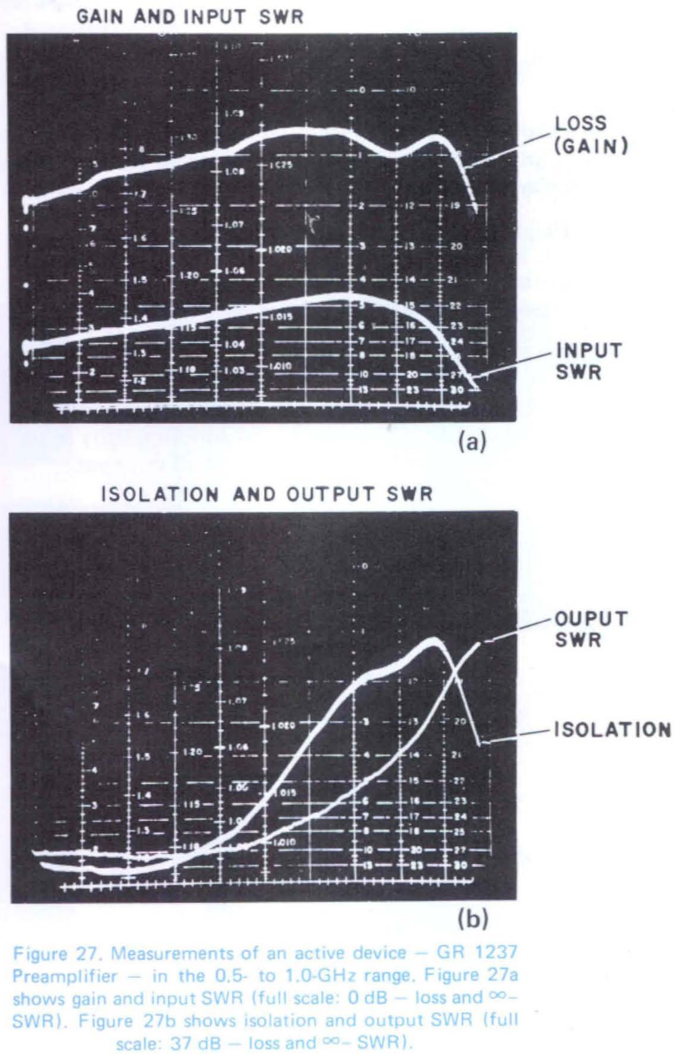


Figure 27. Measurements of an active device – GR 1237 Preamplifier – in the 0.5- to 1.0-GHz range. Figure 27a shows gain and input SWR (full scale: 0 dB – loss and ∞– SWR). Figure 27b shows isolation and output SWR (full scale: 37 dB – loss and ∞– SWR).

by inserting 13-dB attenuation between the amplifier output and the 1641 transfer detector. The attenuation combines with the amplifier gain to introduce a net loss that can be measured directly. The amplifier gain is determined from the measured loss by subtracting the amount of attenuation used. Thus, in this example, subtracting the 13-dB attenuation used from the measured loss of 0.5 to 2 dB across the band gives a –11- to –12.5-dB *loss* or a 11- to 12.5-dB *gain*. The attenuation was added behind the amplifier so the input SWR could be directly measured with no reconnections. The amplifier isolation and output SWR, as shown in Figure 27b, were obtained by reversing the amplifier connections end for end and connecting the amplifier input directly to the transfer detector.

The magnitudes of the s-parameters, $|s_{ij}|$, are obtained directly from the measurements described above and are expressed in dB as return loss ($|s_{ii}|$, $|s_{jj}|$) and insertion loss ($|s_{ij}|$, $|s_{ji}|$). Similar measurements can be made on transistors, diodes, etc, by the use of appropriate mounts and bias-insertion units. If dc bias is used with the device to be measured and no internal dc blocks are provided, a dc block is required between the device and the 1641 UNKNOWN port. The GR 1641 Transfer Detector is internally dc blocked.

ACCESSORIES REQUIRED

Although the 1641 depends on an external source of rf, the source requirements are not stringent. Specifically, the requirements are:

1. 10-mW minimum leveled rf output into 50 ohms

2. capability of being driven by external 10-kHz square-wave modulation, on-off, in the range of –15 to +15 volts and of presenting a minimum load impedance of 1 kΩ to the external modulation source (the 1641)

3. capability of being leveled from an external sampling detector that delivers a control signal with sensitivity in the 50- to 550-mV range for a 30-mW source power input

4. a blanking pulse of ± 3 to 50 volts into 30 kΩ.

Increasing rf output to 200 mW makes greater resolution possible. The modulation and leveling circuitry should not introduce appreciable differential distortion of the 10-kHz waveform. Use of the blanking pulse specified in (4) is required only for triggering alternate (simultaneous) measurements of SWR and insertion loss.

Specific requirements for the oscilloscope are:

1. both axes dc-coupled

2. vertical sensitivity of 0.1 V/cm or greater

3. horizontal sensitivity consistent with the sweep-signal output available from the source (0.1 V/cm is usually more than adequate).

A storage oscilloscope is not necessary for all measurements. It is particularly useful, however, for measurements of low SWR and high insertion loss where the sweep speed must be reduced below flicker speed to accommodate the system response time. The Tektronix storage oscilloscope is recommended, and the 1641 calibrated graticule fits this particular unit.

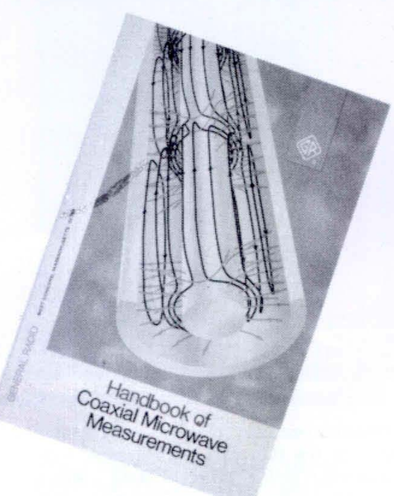
Design and development responsibilities for the GR 1641 Reflectometer were as follows:

T. E. MacKenzie – Project direction, general systems design, and low-frequency rf unit.
J. F. Gilmore – High-frequency rf unit.
M. Khazam – General-system, detectors and indicator design.

Acknowledgements

The oscillograms shown in this article were taken using the Alfred Electronics 650-series of sweep oscillators for frequencies above 0.5 GHz and the Kruse-Storke Electronics 5000-series of sweep oscillators for frequencies below 0.5 GHz. The oscilloscope used was the Tektronix Inc. Model 564 Storage Oscilloscope with two Type 2A63 Differential-Amplifier Plug-Ins.

Complete specifications and prices for the GR 1641 were distributed with the January/February 1969 issue of the ***Experimenter.***



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THE VIABLE VHF/UHF PREAMPLIFIER

A low-noise, low-level, wideband amplifier operating in the frequency range 150 kHz to 1 GHz is as essential to the laboratory worker as bread to the growing child. In the home it could be used to boost the level of television signals, if a correct impedance match between antenna and receiver is established. The new GR 1237 transistor amplifier probably will not be used in the home but its place in the laboratory is assured by its simplicity and adaptability.

The GR 1237 amplifier was designed for general laboratory use in 50-ohm systems as an amplifier, preamplifier, or isolator. It finds application as part of a system to form a sensitive, broadband detector system such as that of Figure 1. Here, a demodulator, such as the GR 874-VQ, is operating as a video or envelope detector, employing a narrow-band, sensitive 1-kHz indicating amplifier such as the GR 1232 or 1234. The circuit suffers from lack of sensitivity and selectivity (or susceptibility to harmonics) as compared with a heterodyne system, but introduction of the 1237 unit improves the sensitivity, as evident in Figure 2, while use of a low-pass filter (GR 874-F) at the detector input reduces susceptibility to harmonics. The resultant system replaces the heterodyne as a null detector and eliminates need for a local oscillator.

Another use involves the familiar GR 1607 Transfer-Function and Immittance Bridge. The 1237, directly connected to the bridge terminals, effectively blocks the local-oscillator signal from the bridge and provides a means for connecting a mixer into the circuitry via the output terminals of the 1237. Bridge measurement errors, caused by local-oscillator feedthrough, are eliminated, sensitivity is improved and bridge performance enhanced, particularly below 100 MHz. Another advantage is assurance of small signal operation of semiconductor devices under test, because of the large reverse loss of the 1237.

Some operating characteristics of the 1237 amplifier are shown in Figure 3. Physical appearance is as shown in Figure 4, and specifications are detailed below for reader convenience.

SPECIFICATIONS

Frequency Range: 150 kHz to 1 GHz.

Gain: >10 dB (see typical curve, Figure 3).

Reverse Attenuation: >33 dB; below 700 MHz, >43 dB.

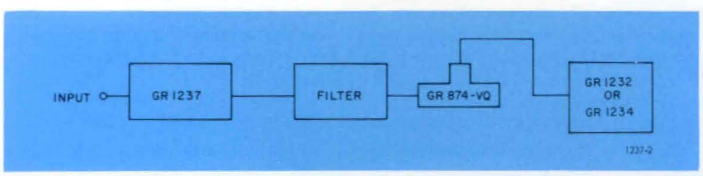
Noise Figure: See typical curve, Figure 3.

Terminals: Input and output, GR874® locking coaxial connectors.

Power Required: 100 to 125 or 200 to 250 V, 50 to 400 Hz, 1.5 W; or 9 V dc, 18 mA.

Catalog Number	Description	Price in USA
1237-9700	1237 VHF/UHF Preamplifier	$195.00

Figure 1. Detector system using the Type 1237.



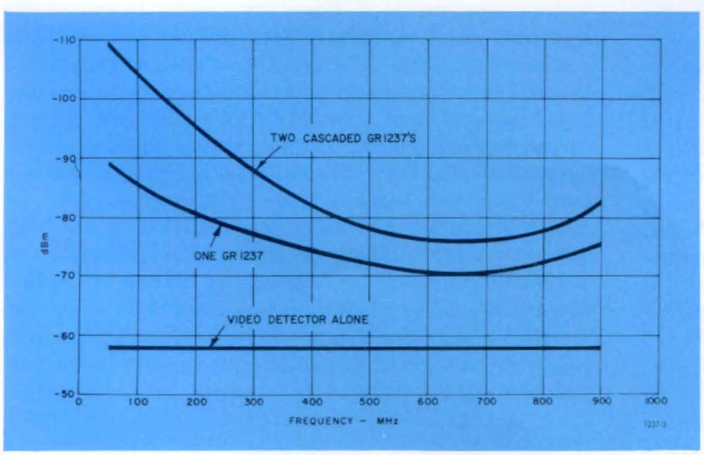
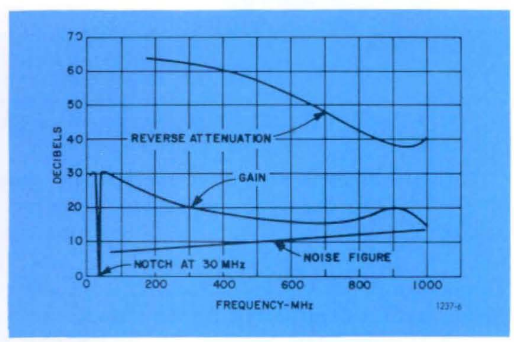


Figure 2. Relative improvement in sensitivity, using the Type 1237.

Figure 3. Typical noise figure, gain and reverse attenuation characteristics.



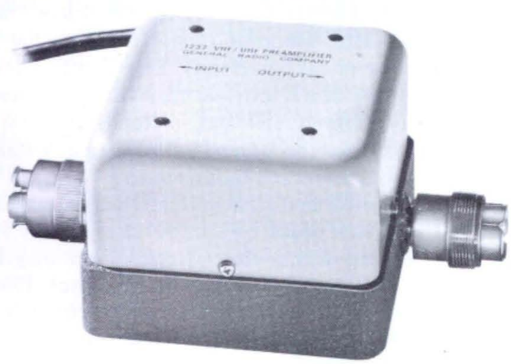
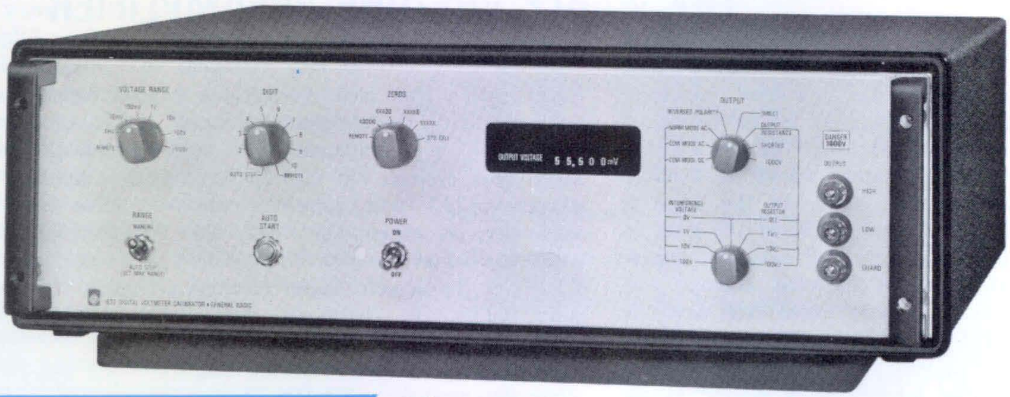
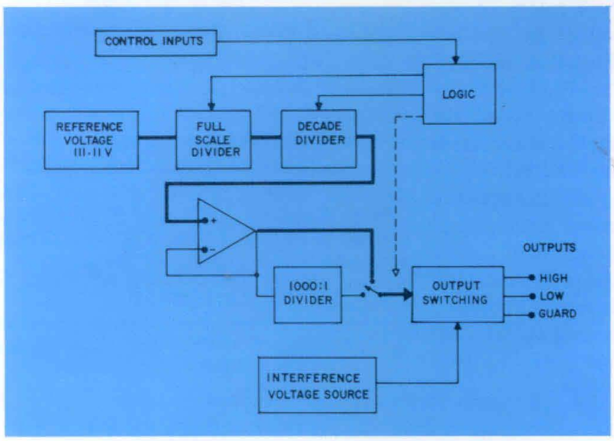


Figure 4. Type 1237 UHF/VHF Preamplifier.



Type 1822 Digital Voltmeter Calibrator.



Block diagram of the calibrator.

The Semiautomatic DC DVM Calibrator

by Ralph P. Anderson

Improvements in performance of digital voltmeters have encouraged increased use, with a correspondingly increased calibration work load. Standards being generated will specify calibration procedures to assure accurate performance of the meters. With these standards in mind, General Radio has designed the 1822 Digital Voltmeter Calibrator, which uses semiautomatic techniques to simplify calibration tests and to save calibration man-hours.

The Need

Digital voltmeters have taken the accurate measurement of dc voltages from the standards laboratory to the engineer's workbench and also to the production line, where relatively untrained personnel can make precision measurements quickly and accurately. Like any measuring instrument, digital voltmeters must be calibrated periodically to assure proper operation within published specifications. Increasing use of the DVM has strained the facilities of many calibration laboratories to the breaking point, because of the complexity, large number, and lack of uniform testing of the instruments involved. To standardize specification format and test methods, the USASI has created, and is in the process of accepting, a standard on automatic voltmeters and ratiometers.

Causes of error in digital-voltmeter measurements can be divided into two broad categories. The primary errors are those generated within the voltmeter, and they are independent of the character of the voltage source. These errors are caused by voltage offsets, reference-voltage drifts, hysteresis effects, nonlinearities, and range scaling errors. Traditional calibration methods, which have been developed to detect and to correct for these sources of error, require the generation of a large number of accurately known voltages.

Secondary errors are those that depend on the specific manner in which the voltmeter is connected or used and, to a large extent, upon the source characteristics in relation to the input characteristics of the voltmeter. The finite input impedance and input current of a DVM cause loading errors during the measurement of voltages with significant source impedances. Similarly, the finite isolation impedance can cause errors when common-mode voltages are present.

The traditional calibration of the DVM leaves these secondary sources of error undetected. It is in this area that the USASI proposed standard, C39.6, will have its greatest impact. It clearly defines these sources of error and preferred testing methods so both the maker and user can agree not only on what a specification means but also on how to verify it.

The Concept

The ideal digital voltmeter would have an infinite input impedance, zero input current and, of course, similar characteristics between each of its terminals and ground. In approaching these characteristics, digital voltmeter manufacturers have come very close to the ideal. The degree to which they have succeeded, however, may be easily defeated when the digital voltmeter is placed in a hostile environment. Insidious settlement of dust and grime onto even the best of insulators will eventually change them from high to low resistance and possibly into a current generator. Similarly, a small disturbance in the balance of the input circuits of the digital voltmeter can cause an offset current that is large with respect to that which the user might expect from reading the data sheet. Since the measurement accuracy of a DVM may well include the effects of these secondary sources of error, it is imperative that tests for them be included in every calibration cycle and that the users be kept informed of the results.

R. P. Anderson presently is Acting Group Leader of the Engineering Department's Industrial Group. He received his MS degree in Applied Physics from Harvard in 1962 after earning his BS in Engineering at Brown University in 1958. During the interim period 1958 to 1962, he served with the US Navy and worked for Ford Aeroneutronics, joining GR in 1963. He is a member of IEEE and Sigma Xi.

The Instrument

The General Radio Type 1822 Digital Voltmeter Calibrator has been designed to test for most of the sources of error as defined by the proposed USASI Standard. Basically the Type 1822 is a remotely programmable voltage supply, with an internal program. Its primary purpose is to make convenient and practical the tests indicated in the standard. It will automatically present a series of low source impedance, precision dc voltages (from 100 μV to 111.11 V) to the output terminals to check for the primary sources of error. Included also are means for checking the input characteristics, the common-mode rejection, the normal-mode rejection, and manual selection of voltages up to 1111.1 volts.

All precision voltages generated by the 1822 are ultimately referred to a single specially selected and aged reference diode, which is housed in a proportionally controlled oven with fast warmup characteristics. The reference voltage, 111.11 volts, is scaled up from the diode voltage by a resistive divider. The resistors in each of the dividers are precision wire-wound resistors, wound from the same spool, heat treated to relieve manufacturing stresses, and placed, uncoated, in a common oil bath to achieve, as nearly as possible, identical changes in temperatures. These precautions result in a ±1.5 ppm temperature coefficient for the division ratio and provide excellent long-term stability.

A photochopper-stabilized amplifier provides the voltage amplification needed for the kilovolt output range and, on the lower ranges, serves as a high-input-, low-output-impedance buffer isolating the dividers from the effects of varying loads.

The VOLTAGE-RANGE switch sets maximum values in decade ranges from 1 millivolt to 1000 volts; in each range the value set by a second control, the digit switch, multiplies the range by 0.11111, 0.22222, 0.33333,..... 1.11110. The ZEROS control permits the last 1, 2, or 4 digits to be replaced by zero to match the resolution of the DVM readout. An automatic stepping mode cycles the digit setting upward to maximum and then back down to one-tenth of full scale for a quick check of all the indicators and of linearity. After one cycle it can be set to step down one range and cycle through the same sequence. Direction of count, range incrementing, and stepping can all be remotely controlled. Because the voltages generated in these sequences have easily recognized numbers and are cycled automatically, the operator's attention can be focused on the DVM under test.

The output switching can reverse the output polarity, short the terminals, and modify the basic dc voltage by the addition of common-mode or normal-mode interference voltages or by a change in the output impedance. This switching also provides a safety interlock that prevents dangerous output potentials in the 1000-volt range from being selected or programmed accidentally.

The 1822 can be completely calibrated against a primary voltage standard by conventional methods, and its inherent stability will permit a six-month recalibration cycle. For a simple check for changes in the internal reference, dividers, or output amplifier, a "simulated standard-cell output" is available that can be compared by a simple null measurement against a saturated standard cell.

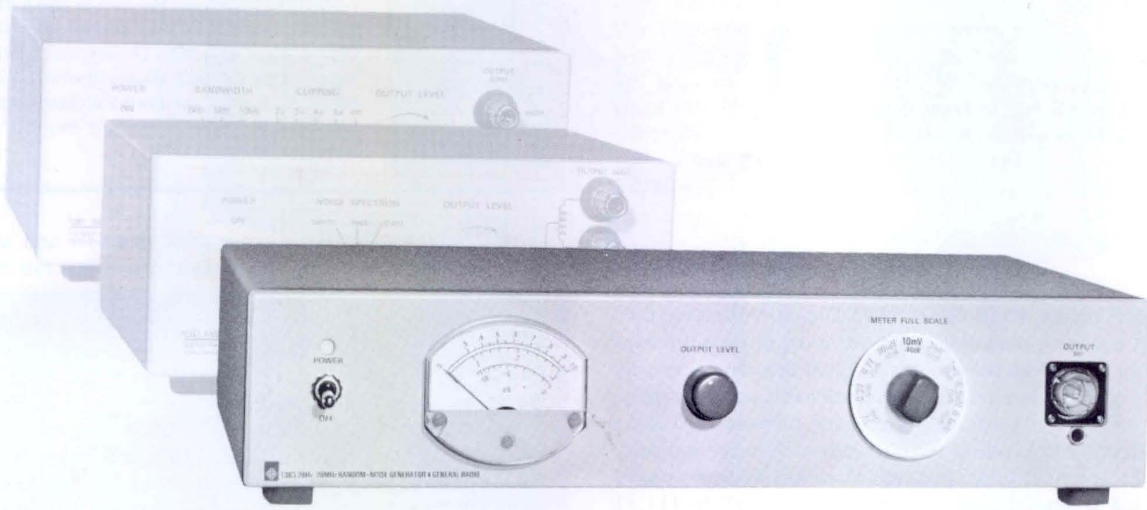
For voltages that are not generated in the sequences of similar digits and zeros, the user may easily substitute an external voltage divider in place of the internal, automatically switched dividers of the 1822. Either a simple resistive divider or a Kelvin-Varley Divider may be used to generate any voltage from microvolts to 1111.1 volts, with the divider itself varying only between 1 millivolt to 111.11 volts.

Its Applications

While the 1822 is ideally suited, because of its mobility, to calibrating DVM's on location in the laboratory, production line, or inspection department, it is not limited to this use. Any precision instrument that requires, transmits, or transduces analog voltages can be calibrated and its susceptibility to interference determined readily by the 1822. Analog-to-digital converters, telemetry systems, and analog voltmeters are typical examples.

As a component in the complete testing system, the 1822 can be programmed by the system to establish precise voltage levels for periodic calibrations of other sections of the system.

The 1822 is a convenient, stable transfer standard that moves easily from the standards laboratory to wherever the DVM's are used. Because of its flexibility, time savings, and ease of use, it can be used economically more frequently than conventional calibration instruments, thus assuring more reliable results from all voltmeters. With the 1822, testing for secondary error sources is more practical than ever before.



NOISE, NOISE, AND MORE NOISE!

- 20 Hz to 20 MHz, ±1 dB
- 30-μV to 1-V output, open circuit

The GR 1383 Random–Noise Generator is a white-noise signal source of contant spectrum level, operating in the frequency range 20 Hz to 20 MHz. Output level is at one volt into an open circuit, derived from a 50-ohm source impedance. Level is adjustable over a range of 80 dB. This instrument complements random-noise generators GR 1381 (audio frequency) and GR 1382 (sub-audio frequency), described in the *GR Experimenter* issued January, 1968.

What's It Good For?

Experimenters will use it as a source of controlled background noise while studying signal-detection and -reception systems. Measuring noise temperature or noise figure of the input stages of an amplifier, within the specified frequency range, is greatly simplified.[1] A simple reminder – the 1383 becomes a 50-ohm source of noise, whose equivalent temperature is the ambient temperature when the power is not turned on.

Use it for checking intermodulation distortions[2] – apply a noise spectrum to the amplifier input, remove a band of noise frequencies from the input by means of a band-stop filter; check the output of the amplifier, using a band-pass filter of the same frequency range as the band-stop filter to determine how much intermodulation product has appeared in the vacant band. This is a useful check for general quality of performance.

[1] Mumford W. W., and Scheibe, E. H., *Noise Performance Factors in Communication Systems*, Horizon House-Microwave, Inc., 1968.

[2] Icenbice, P. J. Jr., and Fellhaur, H. E., "Linearity Testing Techniques for Sideband Equipment," *Proceedings of the IRE*, Vol. 44, pp 1775-1782, December 1956.

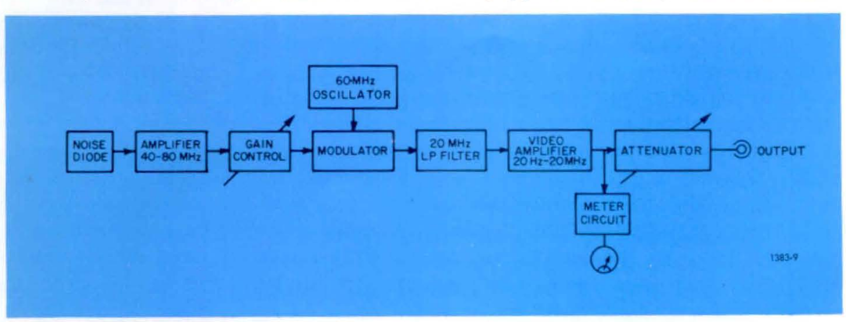


Figure 1. Block diagram of the GR 1383.

J. J. Faran, Jr. is a graduate of Washington and Jefferson College (AB-1943) and Harvard University (MA-1947 and PhD-1951). As a Research Fellow at Harvard, Dr. Faran worked on correlation techniques as applied to acoustic receiving systems. In 1952 he joined GR's Audio Group and has since developed a variety of instruments, including recorders, voltmeters, and analyzers.

Check interference in multichannel communication systems. The broad bandwidth of the 1383 encourages its use at intermediate and radio frequencies. Interference produced in channels adjacent to the test channel may be useful for investigating interference-mitigation methods.

How about checking effective-noise bandwidth of filters? Apply a noise of known bandwidth to the filter input, and compare noise amplitude at input and output. The reduction in level is proportional to the square root of the ratio of the effective noise bandwidths. Note – the bandwidth of the noise output of the 1383 is 20 MHz ±5%.

Do You Need Special Noises?

The noise from the 1383 can be used to modulate a signal generator to produce sidebands of noise at almost any frequency. The modulation bandwidth will, in most cases, be limited by the modulation amplifier in the signal generator rather than by the 20-MHz bandwidth of the 1383. If broader bands of noise are required, and the carrier frequency is undesirable, mix the noise with a sine wave from a signal generator in any one of several balanced mixer units which are commercially available.

Narrow bands of noise are easily produced by building tuned circuits into GR 874-X Insertion Units for insertion into 50-ohm lines; more complicated filters designed for operation in 50-ohm lines may be inserted in series with the output.

What Makes It Work?

The noise source is a temperature-limited thermionic diode, commonly accepted as a random-noise standard. A feedback system, shown in the block diagram of Figure 1, controls the filament current while maintaining a constant plate current. The generated noise voltage level is stabilized in this manner, since it is dependent upon the square root of the dc plate current. Noise current from the diode is amplified in the frequency range 40 to 80 MHz and heterodyned against a 60-MHz oscillator signal, producing noise in the dc to 20-MHz band. The "video" amplifier produces an output noise level of one volt rms, open circuit. The sharp cutoff filter following the mixer provides very accurate definition of the effective bandwidth of the generated noise and helps reduce oscillator leakage into the output.

For Old Friends

We have redesigned the high-frequency section of the old 1390-B Random-Noise Generator, extending the upper limit of the bandwidth from 5 MHz to 20 MHz; tightened tolerances on spectral flatness; made the amplitude distribution symmetrical; built in a constant and more practical output impedance; and restyled the cabinet. We believe you'll find the 1383 quite useful.

– J. J. Faran, Jr.

SPECIFICATIONS

Spectrum: Flat (constant energy per hertz of bandwidth) ±1 dB from 20 Hz to 10 MHz, ±1.5 dB from 10 MHz to 20 MHz.

Waveform: Table shows amplitude-density-distribution specifications of generator compared with the Gaussian probability-density function, as measured in a "window" of 0.2σ, centered on the indicated values:

Voltage	Gaussian Prob. Dens. Function	Amplitude-Density Dist of 1383	
0	0.0796	0.0796	±0.005
$\pm\sigma$	0.0484	0.0484	±0.005
$\pm 2\sigma$	0.0108	0.0108	±0.003
$\pm 3\sigma$	0.000898	0.000898	±0.0003

(σ is the standard deviation or rms value of the noise voltage.)

Output Voltage: Full output 1.0 V rms min, open circuit.
Output Meter: Indicates open-circuit output voltage ahead of 50 Ω.
Output Impedance: 50 Ω. Can be shorted without causing distortion.
Amplitude Control: Continuous control and 8-step, 10 dB-per-step attenuator.
Output Terminals: GR874® coaxial connector that can be mounted on either front or rear panel.
Accessories Supplied: Spare fuses, lamp, power cord.
Power Required: 100 to 125 or 200 to 250 V, 50 to 400 Hz, 40 W.

Catalog Number	Description	Price in USA
	1383 Random-Noise Generator	
1383-9700	Bench Model	**$775.00**
1383-9701	Rack Model	**795.00**

Figure 1. GR 1442
Coaxial Resistance Standard.

STABLE SERIES OF COAXIAL RESISTANCE STANDARDS

The GR 1442 Coaxial Resistance Standards (Figure 1) are the most recent addition to the line of General Radio impedance standards with coaxial connectors. As in the capacitors described previously,[1,2] the GR900® connectors make it possible to have two-terminal impedance standards for the radio-frequency range, which can be calibrated to a high degree of accuracy.

The GR 1442 resistor consists of two GR900 connectors with a short length of outer conductor forming the housing for the resistor. The resistance element, consisting of a cluster of metal-film resistors, is connected between the two inner contacts. The length of the assembly has been made as short as possible in order to keep both the inductance and capacitance low.

The two GR900 connectors permit use of the 1442 resistor in two different ways. One use is as an individual resistance standard, obtained by shorting one end with a GR 900-WN Short Circuit Termination and using the other end for the terminals. The other use is as a resistor in series with other impedance standards having GR900 connectors. This provides a wide range of combinations for calibrations and other measurement purposes.

The characteristics that make the GR900 connector so useful at microwave frequencies also make it a desirable means of connection for two-terminal devices at lower frequencies. The low resistance and the accurately defined reference plane, plus the precisely repeatable inductance and capacitance when connected to different instruments, make it possible to construct and to calibrate impedance standards for a wide range of frequencies and uses. Precision coaxial connectors are essential for the highest accuracy when radio-frequency resistance calibrations are made by the National Bureau of Standards.[3]

When shorted at one end, the 1442 can be used to calibrate high-frequency resistance bridges, such as the GR 1606-B, and similar instruments. It can be used also to calibrate a wide range of instruments for dissipation factor, conductance, or Q when connected in series with a coaxial capacitance standard. This can be particularly useful in materials laboratories and in similar activities that do not have immediate access to the services of a high-frequency electrical standards laboratory. A few 1442 resistors and GR 1405 or 1406 capacitors will provide a number of dissipation-factor combinations in the range of samples

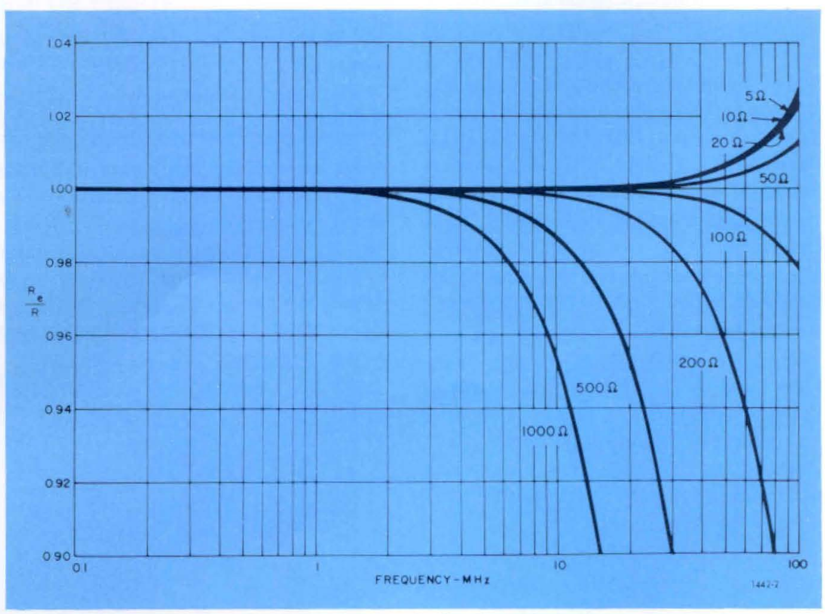


Figure 2. Typical ratios of effective series resistance
to dc resistance, with a
GR 900-WN Short Circuit at one end.

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

[2] Orr, R. W., and Zorzy, J., "More Coaxial Capacitance Standards," *General Radio Experimenter*, May 1968.

[3] *Calibration and Test Services of the NBS, Special Publication 250* – 1968 Edition, p 7.12 and 7.30.

normally measured. These can be used for routine calibrations or as a check when the results of a measurement on a sample appear questionable.

The change in effective series resistance with frequency for 1442 resistors is shown in Figure 2. The change in reactance with frequency is shown in Figure 3. Substantially, all the change in effective resistance for the range shown by the curves is due to the inductance and the capacitance of the resistor. This is explained by reviewing the equation for effective series resistance R_e of the resistor whose equivalent circuit is shown in Figure 4. For analysis, one end of the resistor is shorted.

$$R_e = \frac{R}{1 - \omega^2 \, (2LC - R^2C^2 - \omega^2 L^2 C^2)}$$

The denominator has been rearranged somewhat from the usual form to make it easier to see the effects of the individual parameters. The value of L in this equation is 9 nH; the capacitance 3.4 pF, the sum of 2.5 pF across one end of the resistor assembly and 0.9 pF across the resistor proper. These inductance and capacitance values are substantially the same for all resistance values. At the lower values of resistance, R^2C^2 is less than $2LC$, making the denominator less than one, and the effective resistance rises as the frequency increases. In the higher resistance values, R^2C^2 is greater than $2LC$, making the denominator greater than one, and the effective resistance decreases with frequency. The term $\omega^2L^2C^2$ is relatively small for these resistors at the frequencies shown in Figure 3. A more complete analysis of resistor performance at radio frequencies is given in an article by D. B. Sinclair.[4]

– R. W. Orr

[4]Sinclair, D. B., "Type 663 Resistor – A Standard for Use at High Frequencies," *General Radio Experimenter*, January 1939.

R. W. Orr received his BS in EE from Texas A&M in 1928. He held technical positions with General Electric Company, RCA, Erie Resistor Corporation, AMP Inc., and Aerovox Corporation prior to joining General Radio in 1964. Presently he is a member of the Engineering Department's Impedance Group. Mr. Orr is a member of IEEE and ASTM. He is Chairman of ASTM Committee D-9 on Electrical Insulating Materials, member of Committee D-27 on Electrical Insulating Liquids and Gases, and Associate of the NAS/NRC Conference on Electrical Insulation and Dielectric Phenomena.

SPECIFICATIONS

Initial DC Accuracy: ±(0.1% + 0.3 mΩ).

Stability: ±0.05% per year.

Dissipation: 1 W max.

Capacitance (Inner to outer conductor): 5 pF, typical.

Inductance: 9 nH, typical.

Temperature Coefficient of Resistance: ±50 ppm/°C, except ±100 ppm/°C for 1442-F.

Catalog Number	Description	Resistance	Price in USA
1442-9705	1442-F	5Ω	$65.00
1442-9706	1442-G	10Ω	65.00
1442-9707	1442-H	20Ω	65.00
1442-9708	1442-J	50Ω	65.00
1442-9709	1442-K	100Ω	65.00
1442-9710	1442-L	200Ω	65.00
1442-9711	1442-M	500Ω	65.00
1442-9712	1442-N	1000Ω	65.00

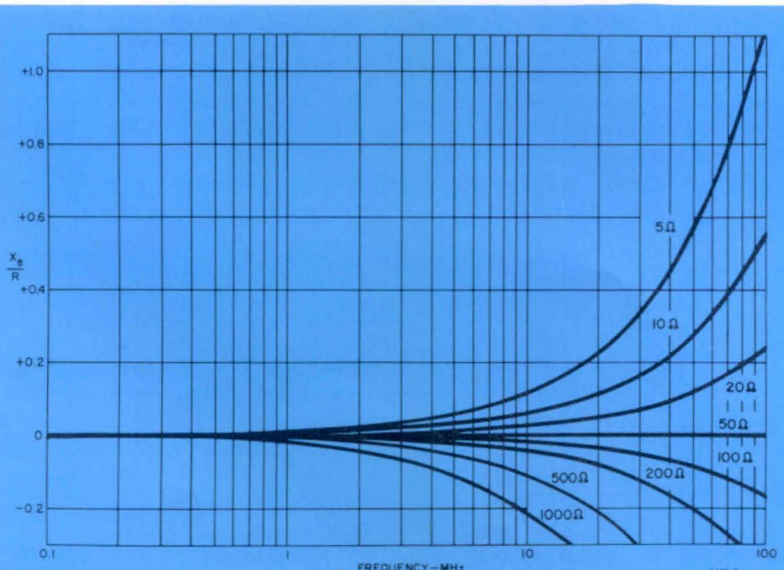


Figure 3. Typical ratios of effective series reactance to dc resistance, with a GR 900-WN Short Circuit at one end.

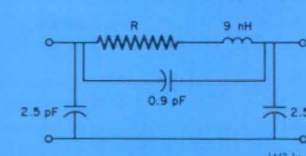


Figure 4. Equivalent circuit for GR 1442 resistance standards up to 100 MHz.

NEW

Type 1863 Megohmmeter – Inspection Model

Type 1864 Megohmmeter – Laboratory Model

A familiar standby for GR customers – the popular 1862-C Megohmmeter – has been replaced by *two* new units, Types 1863 and 1864, which require fewer adjustments and are easier to use.

The 1863 is recommended for production and inspection testing of insulation at five test voltages in the range 50 to 500 volts. Resistance ranges from 50 kΩ to 20 TΩ will cover most insulation specifications.

The 1864 will answer the need for a laboratory-type instrument; but also, its use is recommended for production testing at voltages not available in the Type 1863. This new instrument meets such requirements as 1000-volt checking, leakage testing of semi-conductors and capacitors to as low as 10 volts, and testing for reverse leakage of rectifiers. The highest range of the instrument – 200 TΩ – is most useful for research and development work involving insulating materials.

Leakage testing of paper, plastic, ceramic, and mica capacitors traditionally has been accomplished by measurement of insulation or leakage *resistance* of the capacitor. Testing of electrolytic capacitors of the aluminum- or tantalum-plate type usually is specified as a measurement of the leakage *current*, a method also employed in the inspection of semiconductor devices. Either megohmmeter, plus Ohm's Law, will accomplish the specified test.

Stability of calibration is maintained by use of a four-transistor unity-gain amplifier with FET input circuitry. In addition, no warmup drift is encountered, and high zero stability is maintained during operation. Human engineering has not been overlooked, as evidenced by the warning light that is activated by application of the test voltage. Also continued are use of the MEASURE-CHARGE-DISCHARGE switch and provision for performing grounded and ungrounded measurements. System engineering application of the instrument is encouraged by providing rear terminal access to an output voltage that is inversely proportional to R_x. This proportional voltage may be used to trigger a limit indication or an actuating mechanism in an automatic system. Its value varies from 0 to 4 volts when test voltages of 100 to 1000 are used; the value is proportionally less at lower test voltages.

SPECIFICATIONS

Voltage and Resistance Ranges:

Voltage	R_{min} Full Scale	R_{max} 10% of Scale	R_{max} 2½% of Scale	Useful Ranges
Type 1863				
50, 100 V	50 kΩ	500 GΩ	2 TΩ	7
200, 250, 500 V	500 kΩ	5 TΩ	20 TΩ	7
Type 1864				
10 to 50 V	50 kΩ	500 GΩ	2 TΩ*	7*
50 to 100 V	200 kΩ	5 TΩ	20 TΩ	8
100 to 500 V	500 kΩ	5 TΩ	20 TΩ*	7*
500 to 1000 V	5 MΩ	50 TΩ	200 TΩ	8

*Recommended limit.

Resistance Accuracy: ±2 (meter reading + 1)% on lowest 5 ranges (min reading is 0.5). For higher ranges add:

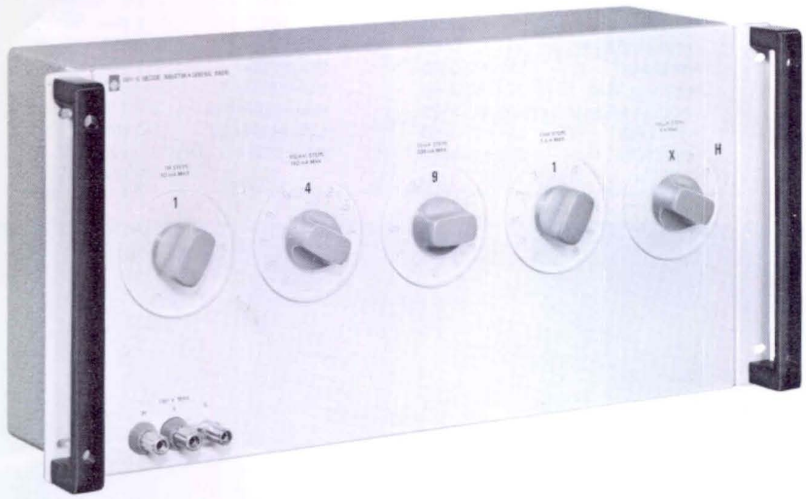
	sixth	seventh	eighth
1863	2%	4%	–
1864	2%	3%	5%

Voltage Accuracy (across unknown): ±2%.

Short-Circuit Current: 5 mA approx.

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

Catalog Number	Description	Price in USA
	1863 Megohmmeter	
1863-9700	Portable Model	**$385.00**
1863-9701	Rack Model	**385.00**
	1864 Megohmmeter	
1864-9700	Portable Model	**485.00**
1864-9701	Rack Model	**485.00**



NEW

The Type 1491 Decade Inductor incorporates several GR940 Decade–Inductor Units within, and insulated from, a single metal housing. The assembly offers moderately precise standards of inductance, with values of Q much higher than air-core coils at audio and low radio-frequency operation. Complete specifications are available from GR Catalog T.

Catalog Number Bench	Catalog Number Rack	Description	Inductance Total	Inductance Steps	940's Included	Price in USA Bench	Price in USA Rack
		Decade Inductor					
1491-9701	1491-9711	**1491-A**	0.111 H	0.0001 H	DD, E, F	**$615.00**	**$630.00**
1491-9706	1491-9716	**1491-F**	1.111 H	0.0001 H	DD, E, F, G	**795.00**	**815.00**
1491-9703	1491-9713	**1491-C**	1.11 H	0.001 H	E, F, G	**595.00**	**610.00**
1491-9707	1491-9717	**1491-G**	11.111 H	0.0001 H	DD, E, F, G, H	**995.00**	**1015.00**
1491-9704	1491-9714	**1491-D**	11.11 H	0.001 H	E, F, G, H	**795.00**	**815.00**
1491-9702	1491-9712	**1491-B**	11.1 H	0.01 H	F, G, H	**615.00**	**630.00**

NEW

General Radio announces the first broadly applied quantity discount policy in the electronics industry; it applies for all customers and to all products.

Since it costs us less per unit when several, even two, items are ordered at a time, GR will share this savings with its customers. Anticipate your needs, consolidate purchases, and save your money. The full discount schedule shown below became effective March 2, 1969.

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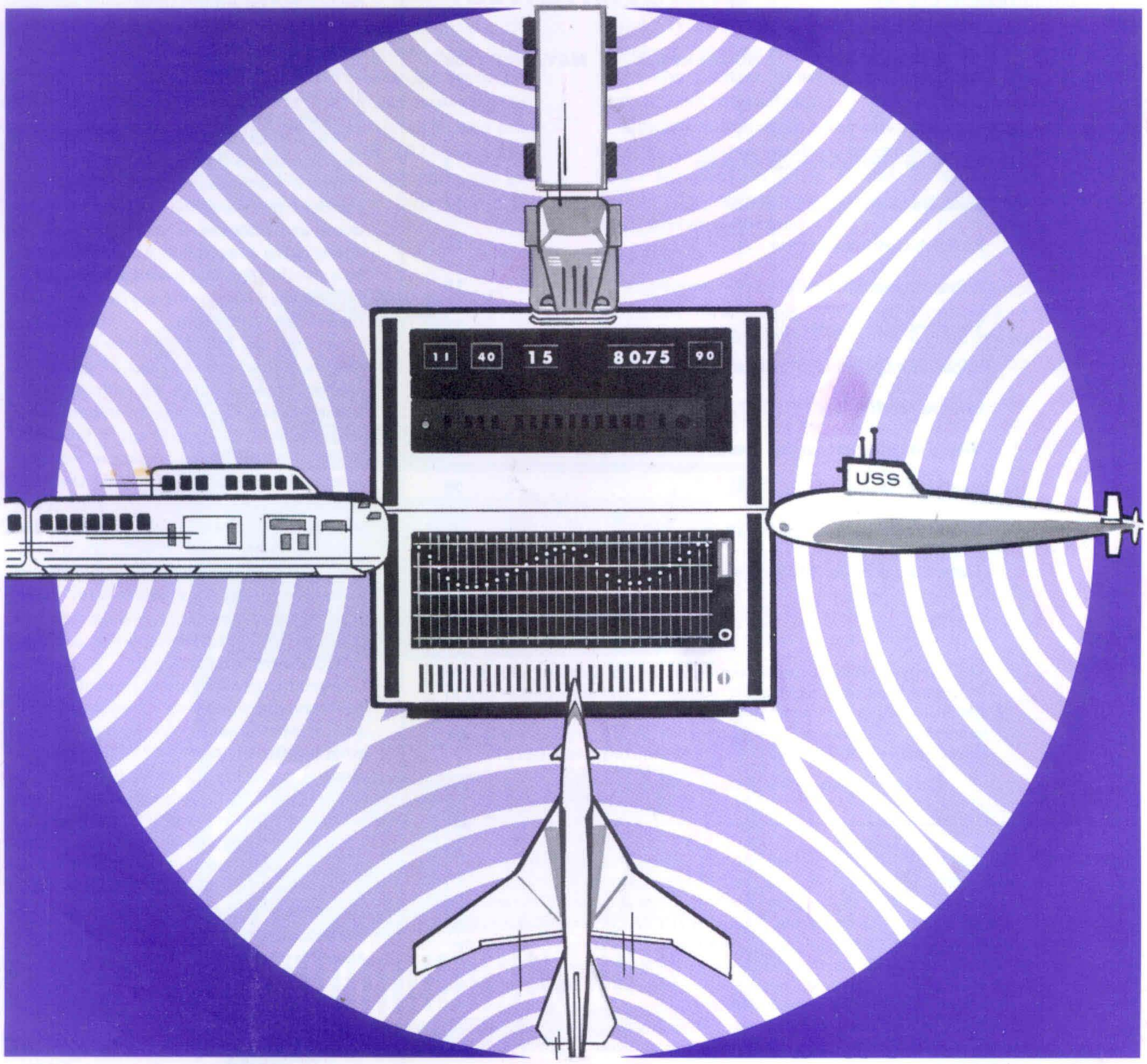


THE
GENERAL RADIO
Experimenter

VOLUME 43 NUMBERS 5, 6 MAY/JUNE 1969

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

THE COVER Experts in the field of sound and vibration are face to face with a problem peculiar to twentieth-century civilization — NOISE. We hear, from all directions, of the effects of noise pollution upon the well-being of the general public. We recognize also that noise can be an indicator of trouble, an element of danger, or a mask to cover desired information. In order to mitigate noise, we must evaluate its effects by performing measurements and studying the efficiency of noise-reduction efforts. Most desirable in noise studies is an ability to make instant evaluations through real-time measurements. Our cover depicts some areas of concern to the public, to engineers, and to the government. GR hopes that its latest contribution, the GR 1921 Real-Time Analyzer, will help establish new standards for allowable noise levels, improve mechanical designs, and contribute to the health and safety of all people.

A recent article in the National Conference of Standards Laboratories ***NCSL Newsletter*** asked the question "Do specifications really specify?" Editors of several trade journals also have been beating the drums to start a parade for clear and undeceptive specifications. A most important point raised was reflected in the comment ". . . bad specs drive out good ones."

The subject of specifications is one that is very sensitive to the economic feelings of instrument manufacturers. Specifications are the links that form the bond between producer and customer. That bond is only as strong as the weakest link, which could easily be a specification that is not attainable or can be met only under controlled (but not specified) circumstances.

A link can be found weak for other reasons, one of which is use of unclear or strange language. Lack of definitions for special terminology employed in some fields of instrumentation technology is, too often, the root of the evil of ambiguous wording. Another weakness may arise in the choice of tolerances; these ***can*** be based upon actual production test results with a clear understanding, or expression, of the standard deviations of the basic test data. Unfortunately, no standardized approach exists with which manufacturers may comply and the customer is left, too often, to determine how accurately a tolerance has been derived.

More thought is required, both of customer and producer, which will lead to a recognition of what is good, what is necessary, and what is sufficient in presenting specifications. General Radio has pondered the problem of specifications for many years and, in its own way, has tried to present to the customer what it believes is a true specification for instrument performance. We wonder sometimes whether we might say less, or more, in presenting a parameter for instrument performance? Perhaps our readers have some thoughts on the subject?

C E White

C. E. White
Editor

New-Generation Acoustical Analyzer

A sound and vibration analyzer, conceived as a third-generation answer to analysis and interpretation of an increasingly complex environment on land, under sea, and in the air. This prodigy gives 1/3-octave spectrum analysis from 3.15 Hz to 80 kHz and employs a unique digital-detection technique to achieve performance unattainable with analog instruments.

by W. R. Kundert, J. A. Lapointe, and G. R. Partridge

THE BROAD VIEW

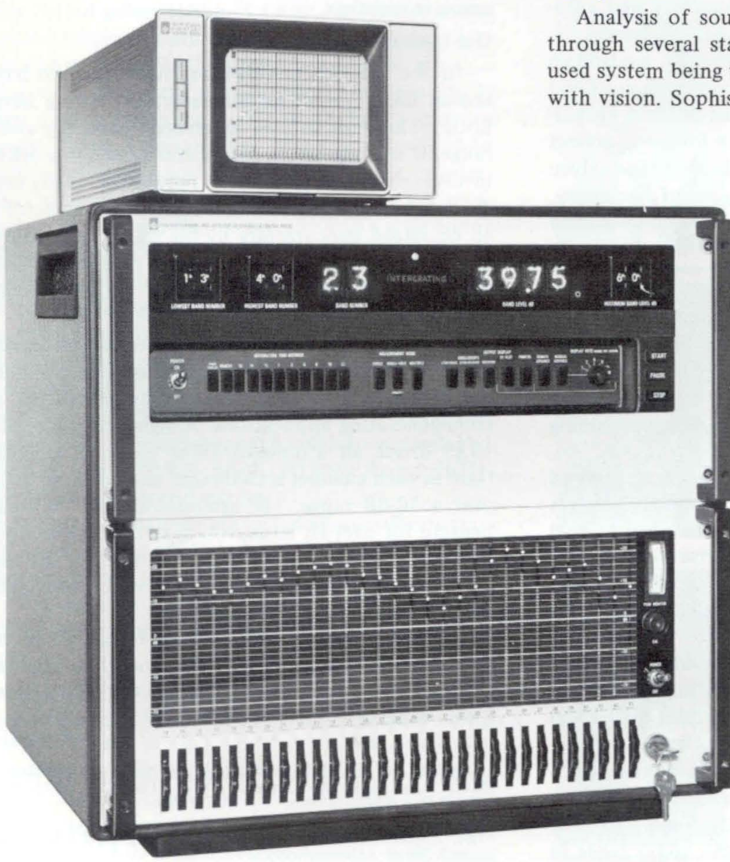
Long Time or Real Time?

This hurried world of ours substantiates the maxim that "Time is of the essence." The pressures placed upon design and development activities leave little doubt that solving engineering problems in real time is a capability most engineering activities must have, in order to remain virile and competitive. Consider the fact that the GR 1921 Real-Time Analyzer will measure an unknown noise spectrum and give you the corresponding Stevens loudness level, all in 0.5 second! Or that acoustic recordings of swiftly-moving vehicles, even planes, are accomplished so quickly that instructions for corrective measures can be transmitted to the vehicle, received and effected, and the results recorded while vehicle and station are still in a line of sight! This analyzer is a major contribution to active and meaningful research, development, and production.

Quickly summarizing, we can say that, when an analysis must be completed on-line, when large quantities of data are to be analyzed, or when a series of contiguous spectra is desired (e.g., automobile and airplane pass-by studies), real-time analysis is essential.

Background

Analysis of sound and vibration phenomena has passed through several stages, the most primitive and still widely used system being that of the human auditory senses coupled with vision. Sophisticated electronic systems in the second



Type 1921 Real-Time Analyzer, with accessory Type 1921-P1 Storage Display Unit.

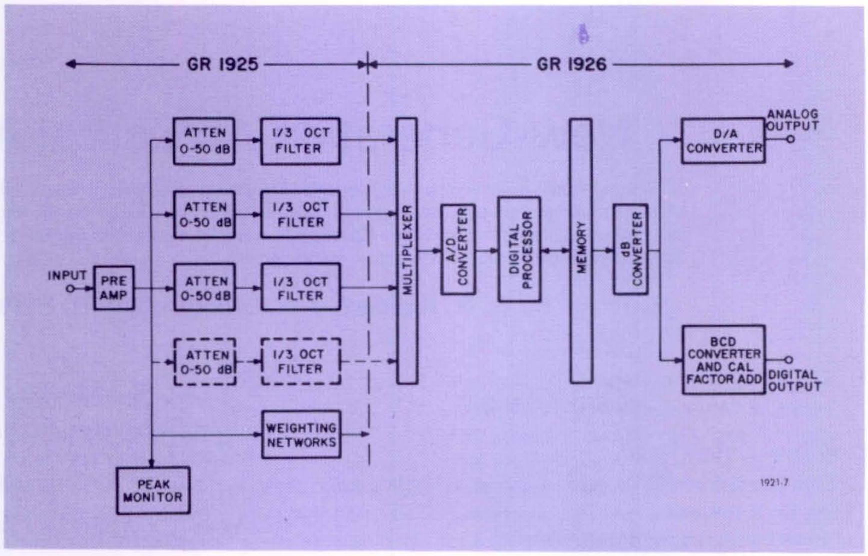


Figure 1. GR 1921 Block diagram.

stage helped to generate records of tests that were far more useful and durable. These systems, still employed, include sound-level meters, vibration meters, and swept (or serial) analyzers of many types. The advance from primary to second stage, unfortunately, did not provide a fundamental and necessary asset – the ability to analyze rapidly and automatically.

Looking back through the past three decades, we find an increased emphasis upon the vital need to cope with noise and vibration problems arising from technical advances in propulsion and other fields. Engineers have been forced to collect huge amounts of data, often laboriously, in order to analyze the complex problems involving various types of machinery, vehicles, and the disturbances that contribute so heavily to noise pollution of our living environment.

Some Design Views

An analysis of a time-varying signal, conducted with a conventional 1/3-octave serial analyzer, can employ a tape loop, which allows the test signal to be played back repeatedly as the analyzer is stepped to, or swept by, each center frequency. This method of analysis is quite time consuming and is not suitable for "on-line" operation.

Several methods of making real-time 1/3-octave analyses are possible. One technique employs conventional 1/3-octave analog filters followed by a set of analog detectors, or signal averagers, and a scanner. A dynamic range of no more than 30 to 40 dB is possible in an analog square-law detector (range changing in a real-time analyzer is not possible), and this is inadequate for many applications.

Digital filtering and the discrete Fourier Transform have been known for some time, and the science of digital signal processing is now rapidly advancing. The Fourier Transform is a natural basis for designing a spectrum analyzer; unfortunately, though it is easy to apply and to understand for a simple periodic signal, its use in analyzing complex or random signals is not fully understood and processors are only now available with sufficient speed to cover the audio range in real-time. Though a purely digital, real-time, 1/3-octave analyzer with wide-dynamic range is within the state of the art, it would be a very expensive device. So called recursive digital filtering must be ruled out because present-day processors are still too slow for full audio range 1/3-octave processing in real-time.

Our Design Choices

In the GR 1921 Real-Time Analyzer, we have chosen analog filtering and digital detection, taking advantage of both techniques to give us greatest accuracy and dynamic range. It is designed for sound and vibration work that often involves signals which may be random or totally unspecified. As a real-time analyzer it has many filter-detector channels, all energized concurrently by the signal to be analyzed. Our packaging permits independent use of the two basic units – the GR 1925 Multifilter[1] and the GR 1926 Multichannel RMS Detector. (See Figure 1.)

The GR 1925 Multifilter supplies summed and scanned outputs, in addition to its parallel output, providing for its use in a variety of spectrum-shaping, equalizing, and spectrum-generating applications. A signal connected to the GR 1925 drives all attenuator-filter channels simultaneously. Gain in each channel is calibrated and adjusted in 1-dB steps over a 50-dB range. The attenuators may be used to compensate for over-all measuring-system response errors. They also serve to de-emphasize, or pre-whiten, the input spectrum, thereby increasing the effective dynamic range of the GR 1921 system to as much as 95 dB.

One-third-octave filters, ranging from 3.15 Hz to 80 kHz and complying with current IEC and USA Standards,* are available in regular versions of the GR 1921. Octave-band filters ranging from 4 Hz to 16 kHz are also standard and

[1] Kundert, W. R., "A Calibrated Spectrum Synthesizer," *General Radio Experimenter*, October 1968. The GR 1925 may be purchased as a separate unit.

*IEC Publication 225-1966 and USA Standard USAS S1.11, 1966 Class 3 (High Attenuation).

available; narrow-band filters ranging from one-third to one-tenth octave are available on special order. The multi-filter chassis will accommodate up to 30 filters. Peak monitoring against overload is a standard feature, as well as mechanical key-locking of attenuator settings.

The GR 1926 detector can be used independently to measure as many as 45 signals simultaneously, which may be derived from multiple transducers or other sources. It can be considered as a 45-channel voltmeter with near-ideal characteristics. The detector operates by sampling each input channel and feeding these samples to the digital circuits that are time shared on all channels. It is available in 30- or 45-channel versions.

The detector simultaneously computes the rms level for each filter channel in the GR 1925. Up to 1024 samples are taken from each channel; the samples are converted to a digital binary number and squared. The squared values are accumulated in a memory register where spaces are provided for all channels.

Single integration or measurement periods are adjustable in nine octave steps ranging from 1/8 to 32 seconds. For integration periods of 1 second and longer, 1024 samples are taken from each channel during the integration period. For shorter integration periods the number of samples is proportionately reduced, with a minimum of 128 samples being taken in a 1/8-second integration period. At the end of an integration period, the sum-of-squares value is converted to decibels for output presentation. A single answer can be fed to a high-speed receiving device in about 15 μs, and all channel levels can be presented at the output in less than 1 ms.

Output data are presented simultaneously in digital (BCD format) and analog forms, and on a front-panel visual numeric display. Control signals and format are designed to permit output interfaces with a digital computer, printer, oscilloscope, automatic dc step-chart recorder, or X-Y plotter. A panel control allows the operator to add a scale factor to the digital output. The detector requires an input signal level of 1 volt rms for full scale maximum output. The output indication corresponding to a 1-volt input signal is adjustable from 60 to 159 dB in 1-dB steps.

Band numbers are presented with each output level and are set to correspond to the USA Standard Band Numbers for 1/3-octave filters (USAS S1.6 - 1967) for the particular version of the GR 1921. The number of channels to be displayed is controlled by the LOWEST BAND NUMBER and HIGHEST BAND NUMBER control settings.

Two extra internal channels are included, to calibrate zero level and full scale on the 1926. They can be measured during each integration period to monitor calibration continuously, if desired. The calibration channels are also intended to be used to set up analog output equipment, such as a reader or oscilloscope.

DIGITAL DETECTION ADVANTAGES

Wide Dynamic Range

There is really no fundamental limitation on dynamic range once a signal is converted to digital form, as there always is in the analog world. In the GR 1921 analyzer, only the multiplexer and A/D converter stretch analog techniques,

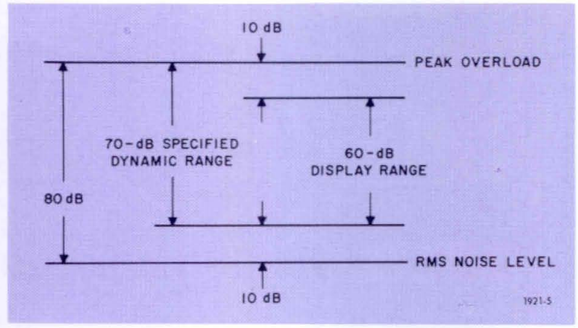


Figure 2. Dynamic range capability.

and these circuits have a dynamic range, measured from overload to noise level, of more than 80 dB. In the digital circuits it is only necessary to carry enough significant digits to attain the desired dynamic range.

The dynamic range situation for the GR 1921 analyzer is illustrated in Figure 2. On a per channel basis, the rms noise level is more than 80 dB below the peak-overload level. The specified dynamic range, the range of instantaneous levels for which signal-to-noise ratio is greater than 10 dB, is 70 dB. The display range extends from a level at least 10 dB above the noise floor to a level 10 dB below the peak-overload level.

W. R. Kundert joined GR's Audio Group in 1959, between Northeastern University degrees (BSEE - 1958 and MSEE - 1961). He was made Group Leader of the Acoustics/Signal Analysis Group in 1968. He is a member of the IEEE, Audio Engineering Society, Acoustical Society of America, and Eta Kappa Nu.

J. A. Lapointe holds degrees from Northeastern University (BSEE-1957 and MSEE-1959). At Raytheon Company he performed radar system circuit design, was responsible for systems design and project management at Sanders Associates, and acted as consultant at Signatron. In 1967 he joined GR, specializing in digital signal processing techniques. He is a member of IEEE and Eta Kappa Nu.

G. R. Partridge received his PhD(EE) degree from Yale in 1950 and taught at Purdue University during 1950-55 (Associate Professor EE). He joined Raytheon Company in pulse-communication work in 1955, transferring to GR in 1962 for design work in pulse generators and amplifiers. In 1958 he published ***Principles of Electronic Instruments*** (Prentice-Hall) and is author of numerous technical papers. His memberships include IEEE, Acoustical Society of America, Tau Beta Pi, Eta Kappa Nu, Sigma Xi, and he is a registered Professional Engineer in Massachusetts.

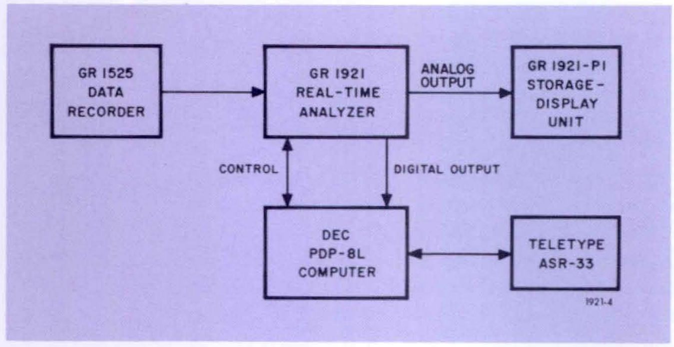


Figure 4. Block diagram of typical analyzer/computer system.

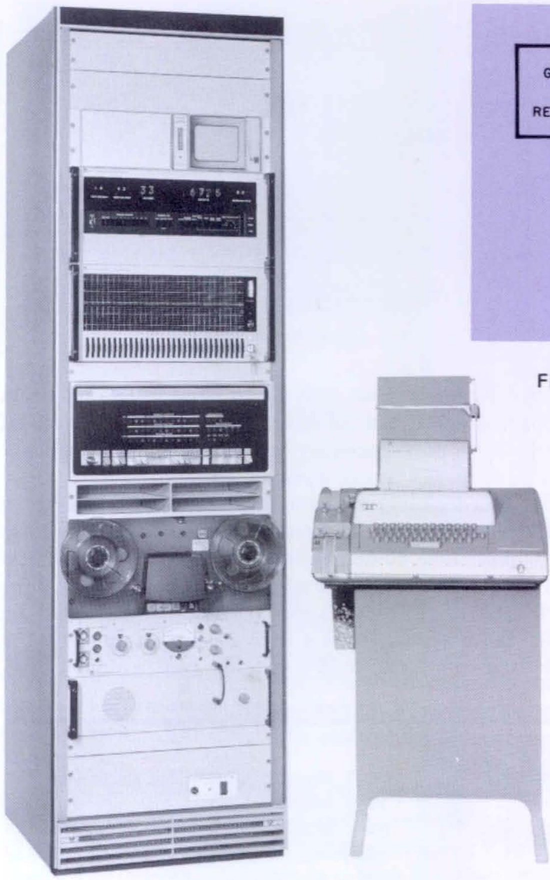


Figure 3. Example of analyzer/computer system.

This ensures a crest-factor capacity, even at full scale, of 10 dB. Though instantaneous levels (samples) over a range of more than 80 dB are taken into account in computing band levels, answers in the lowest 10 dB are considered as noise and are therefore suppressed. Similarly, answers in the top 10 dB are displayed as overload, having encroached too closely upon the crest factor.

Computed band levels that exceed the upper limit of the display range are identified in the output by adding 800 dB to the erroneous digital signal with a resultant amplitude jitter in the corresponding analog level.

Accuracy

The digital approach used in the GR 1921 circumvents the accuracy limitations of analog methods. The computations can be performed to any desired resolution and, of course, all signals are processed with the same computer, so channel-to-channel uniformity is not a consideration.

True Integration Characteristics

The samples taken from each filter channel are squared and averaged to determine the band levels. When the "start" button is pushed, the detector samples the filter channels until the pre-selected integration time has elapsed. Samples are taken only during this period, and all samples are weighted equally in the result. The integration is thus linear, and the time period during which the measurement was made is accurately known. It is easy to vary the average rate at which samples are taken to produce any desired integration period. The GR 1921 has a choice of nine integration periods, ranging from 1/8 to 32 seconds.

By comparison, analog systems have several disadvantages. Long-time analog integrators in the quantity needed for a detector are expensive to the point of being impractical. To change integration time requires that a component in each detector channel be switched. This, too, is expensive, and electronically cumbersome. In an analog design, one must settle for a narrow choice of integration or averaging periods. Also, the only practical averaging method is simple RC smoothing. When presented with a signal, the output from this type of averaging circuit tends toward a finite saturation level. Events that occur early in the time signal are "forgotten" or leak off, and it is impossible to know exactly to what degree each event in the time signal has affected the answer. Several time constants are needed to bring this detector to saturation, the level for which it is calibrated, and this means that a longer time signal is needed for a given degree of statistical accuracy when the signal is random.

SYSTEMS AND SOFTWARE

The GR 1921 can produce data in prodigious quantities in a short time, posing the problem of data reduction. More important, it usually happens that the concern of the engineer is with a single number or possibly a suggested course of action, and not with the raw spectral information that is the product of the analysis process.

A small digital computer used in conjunction with the GR 1921 can perform any or all four basic functions:

- Analyzer Control - the computer can be programmed to select automatically all functions that are normally controlled manually by front-panel controls.
- Rapid Data Storage - the computer can ingest the results of a measurement in about one millisecond, thus freeing the analyzer to make another measurement. The stored results of the series of measurements are then typed out from the computer.

```
CLEAR


ID: WRK8
LOW BAND #: 14
HIGH BAND #: 43
INTEG TIME (SEC): 4

# MEASUREMENTS: 6
TIME BETWEEN MEAS (SEC): 2
OUTPUT: OCT, OCT MAX, TOCT MAX, SONES, PHONS, SONES MAX, SIL

START


WRK8

OCT BANDS
MEAS  15     18     21     24     27     30     33     36     39     42
   1  66.9  100.1  107.5  108.0  106.8  107.8  108.4  103.9   70.3   65.8
   2  64.4   99.4  103.9  107.4  105.9  104.4  102.5   96.5   62.8   65.5
   3  62.0   87.1   65.0   66.3   58.8   55.1   54.9   54.9   57.4   65.4
   4  61.6   96.5   94.5   89.1   62.4   55.1   54.9   54.9   57.8   65.5
   5  64.9  102.4  107.9  106.8   96.6   65.6   54.9   54.9   57.8   65.4
   6  96.3  108.4  106.4  105.3  105.1  103.4   93.0   62.6   58.0   65.9
 MAX  96.3  108.4  107.9  108.0  106.8  107.8  108.4  103.9   70.3   65.9

TOCT
BANDS
 MAX
  14  72.0   89.0   95.2   98.0  102.7  106.2  104.5  103.2  103.0  103.5
  24 103.2  102.5  100.5  102.7  102.2  102.0  103.2  103.2  102.2  105.7
  34 101.5   91.0  101.0  100.2   69.2   62.2   56.5   57.5   60.7   63.2

SONES MEAS
   1     2      3      4      5      6    MAX
 OVRLD OVRLD  24.9   44.6  116.7  191.5  OVRLD

PHONS MEAS
   1     2      3      4      5      6
 124.2 119.6  86.4   94.8  108.8  115.9

SIL   MEAS
   1     2      3      4      5      6
 107.6 104.2  56.2   57.4   72.3  100.4
```
1921-10

Figure 5. Data print-out.

- Comparison - the computer can compare the results of a measurement with a stored reference spectrum and make a decision depending upon their relation. For example, a noisy device can be rejected when its spectrum exceeds a preset limit. The computer also can be programmed to tell what limits have been exceeded, by what amount, and probable cause.
- Data reduction - the computer can be used to reduce the spectral information from the analyzer to obtain various acoustical ratings. For example, programs can be supplied to find loudness level, speech-interference level (SIL), and perceived-noise level (PNL).

Figures 3 and 4 show an analyzer/computer system and its corresponding block diagram. Data are fed to the analyzer from the GR 1525-A Data Recorder. The analog output from the analyzer drives the GR 1921-P1 Storage Display Unit, which displays each spectrum as it is transferred to the computer. The computer accepts digital data from the analyzer, and another link transmits control signals between the analyzer and computer. All communication between the system and the operator is via teletype.

This system and the operator work together in a "conversational" mode. The computer presents a series of questions and the operator replies; the questions concern desired frequency range, integration time, time between measurements, and quantity of measurements. After the operator has supplied this set-up information, the computer asks which of a series of standard data-reduction operations it should perform. This information is supplied by the operator and then, at the touch of a key, the system performs the required measurements and types out answers in a standard format. The system is capable of making a rapid series of 1/3-octave measurements with selectable time delay between measurements. It computes octave band levels from 1/3-octave levels, and it determines the maxima from corresponding band levels in the series of spectra using either octave or 1/3-octave data. It can also define maximum-loudness and maximum-speech-interference levels.

An example of the typed-information output from the system is shown in Figure 5. The underlined data are instructions supplied by the operator; all other data are computer-controlled output.

Basic software packages are available to compute octave-band levels, SIL, PNL, Stevens loudness level, and, if desired, the maximum of these in a sequence of different measurement intervals. Software packages to control the analyzer have been developed for use in custom systems. Other programs can be made available for custom analyzer/computer systems.

APPLICATIONS

The one-third-octave analyzer is the heart of most acoustic noise-analyzer systems. This relatively broad, constant-fractional-bandwidth method of analysis is, for a variety of reasons, well suited to airborne noise analysis and control work.

On-line operation is made practical by the high speed and accuracy of the GR 1921. As mentioned in the introduction, measurements of aircraft fly-by and automobile pass-by are so rapidly assimilated that corrective measures can be accomplished during the test period. This approach, which is not possible using a tape-loop serial system, can greatly reduce the cost of testing.

Time-varying signals are handled by the analyzer so fast as to permit liberal application of spectrum analysis. It is possible to produce three-dimensional plots, with suitable display devices, to present spectrum versus time. Or, a computer system might be programmed to present long-time average spectra in the section of a time-varying signal, where the changes are not significant, while retaining the fine detail in other areas.

Volume-data-reduction work is economically accomplished through intelligent use of an analyzer/computer system. This is especially true when compared to the conventional method of employing a swept analyzer and recorder and reducing data by means of tables and a desk calculator. A comparison table, showing relative times to measure a spectrum and then to compute the corresponding Stevens loudness level, is shown below.

Conventional Means		1921/PDP-8L System	
Time for analysis	20 s	Time for analysis and computation of loudness level	0.5 s
Read chart levels and look up loudness indices	300 s		
Calculate loudness level	180 s		
Total	500 s	Total	0.5 s

The saving in time is clearly evident; one year's work by conventional methods is now accomplished in a few hours! Even when a general-purpose computer is used to find loudness level from spectral information, considerable time is required to prepare data for insertion in the computer. In addition, waiting for availability of a general-purpose computer greatly reduces the attractiveness of this approach.

Production-line product testing has seldom employed spectrum analysis because of the long time required to obtain accurate measurements. The real-time analyzer eliminates this objection and potentially makes it a powerful production-line tool for fault diagnosis. The use of a small computer in a production-line system can increase effectiveness even further. The computer could be programmed to compare measured spectra with reference spectra, suggesting possible faults and repairs to be made to the rejected product.

Transmission-loss measurements, in which dynamic range and signal-to-noise-ratio may be critical, are simplified through use of the calibrated adjustable channel attenuators on the GR 1925 Multifilter. A block diagram of a system for transmission-loss testing is shown in Figure 6. The GR 1382 Random-Noise Generator and the multifilter provide a source-side spectrum, shaped to provide a uniform signal-to-noise ratio versus frequency on the receiving side. The multifilter on the receiving side is set to have a transmission pattern that is approximately the inverse of the transmission pattern on the sending-side multifilter, taking into account frequency-response errors of the transducers. This results in a direct-output plot of transmission on the recorder, while minimizing the source-power requirements and maximizing tolerance to interfering signals on the receiving side.

Radiated power measurements can be made by dividing the same integration period among a specific number of microphones. A switch closure to ground, at the appropriate control line on the GR 1926 Detector, divides the selected integration period into 2, 4, 8, or 16 segments, corresponding to the number of microphones. Alternatively, the position of a moving microphone can be synchronized with the measurement interval in order to obtain a space integration.

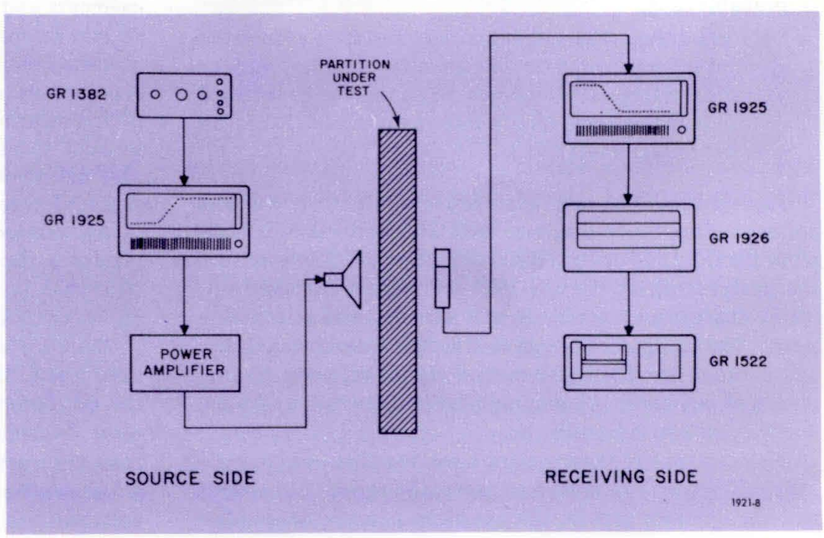
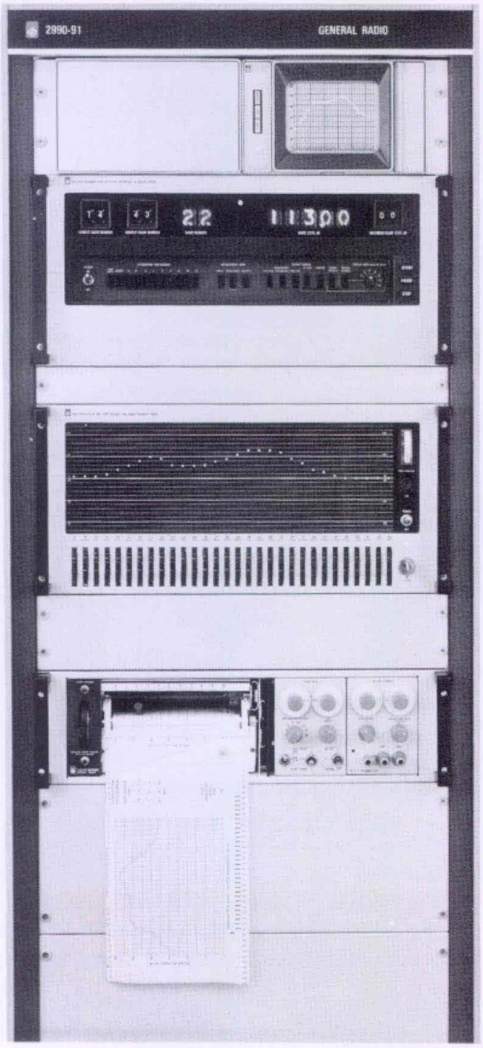


Figure 6. Block diagram of transmission-loss measurements.



AFFINITY FOR ACCESSORIES

The GR 1921 analyzer has great flexibility in operation with accessory equipment.

It can be expanded to 45 channels by use of the 45-channel version of the GR 1926 detector and two GR 1925 multifilter units.

It supplies power to and interfaces with the GR 1560-P40 Preamplifier, suitable for high-impedance transducer inputs.

It operates with the GR 1525 Data Recorder and similar tape recorder units that have good dynamic and frequency ranges and stable characteristics. Of particular interest is use of the maximum-level dB control on the analyzer to restore the calibration factor lost in the recording process. This control is used to make the analyzer read out in decibels, with the reference level the level of the original signal.

It is compatible with the new GR 1522 DC Recorder* (bottom unit, Figure 7a), an analog recorder much faster than conventional X-Y plotters. The new recorder operates synchronously with the GR 1921. A short dwell period as each band level is selected allows the recorder pen to settle, producing a neat bar graph with standard scale factor. The analyzer/recorder combination and a sample chart record are shown in Figure 7.

It can supply output data at rates up to 360 band levels per second. The MDS Series 800 Hi-Speed Digital Printer is available on special order from General Radio. The printer should be used when most accurate numerical data are required. Note that the GR 1921 is capable of driving a printer and analog recorder simultaneously.

*Refer to page 16.

Figure 7a. Visual/record assembly including, from top to bottom, GR 1921-P1 Storage Display, GR 1926 Multichannel RMS Detector, GR 1925 Multifilter, and GR 1522 DC Recorder.

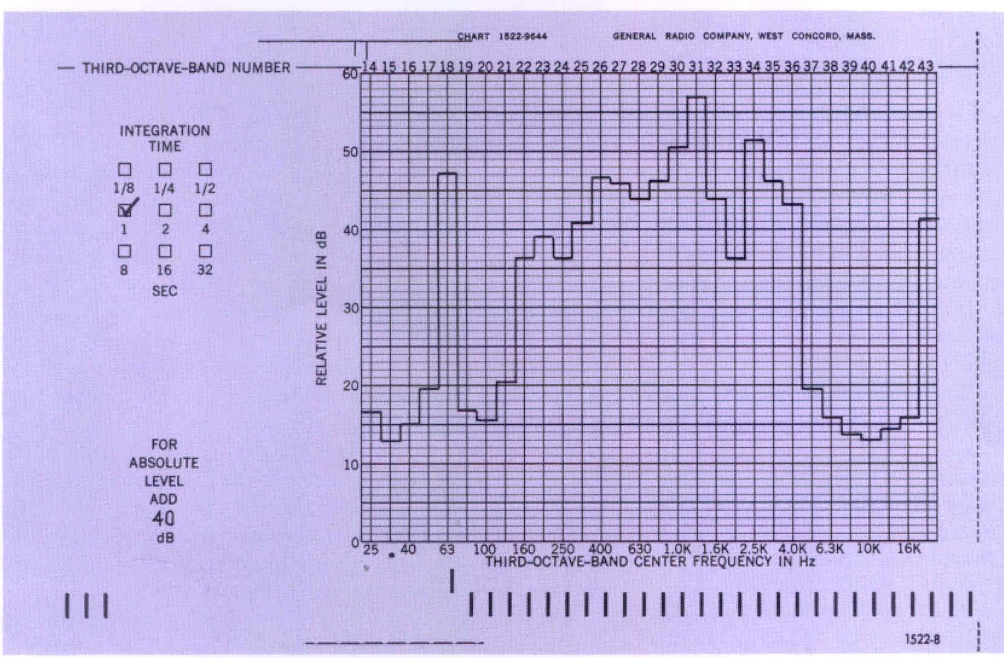


Figure 7b. Typical GR 1522 graphic recording.

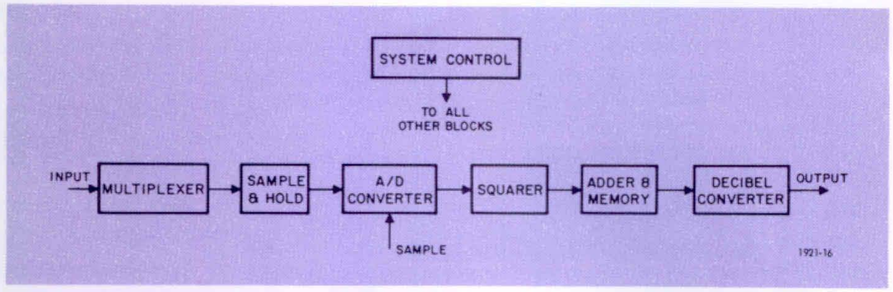


Figure 8. Block diagram of digital detector.

The Houston Instrument Series 6400 (with 024 option) Omnigraphic Recorders are available on special order from General Radio when an X-Y recording format is desired. These recorders use fan-fold charts that load automatically, thereby taking advantage of the speed of the 1921 units. Conventional X-Y recorders are not recommended because of slow loading time.

The GR 1791 Card Punch Coupler is available to couple the GR 1921 units to an IBM 526 Printing Summary Punch, permitting transfer of band numbers and levels to standard IBM cards.

The GR 1921-P1 Storage Display Unit (top unit, Figure 7a) will provide a rapid visual display of spectra from the analyzer. It is a slightly modified version of the Tektronix Type 601 Storage Display Unit. Three CRT display modes can be selected by the GR 1921 – NON-STORE, STORE-ERASE, and STORE-NO ERASE. When any of the three scope modes is selected at the GR 1921, data are fed out at a rate of one band per millisecond. When X-Y PLOT or DATA PRINTER output modes are selected, the scope remains in operation but is fed data at the rate set by the DISPLAY RATE control. In this manner, the unit continues to monitor output when recording devices are used.

A CLOSER LOOK AT THE DETECTOR

A unique design deserves some elaboration in order to stress its qualities and advantages. This section expands upon our previous presentation of design details and refers to Figure 8.

The *Multiplexer* is an electronic switch with up to 45 inputs and one output. The input channels are switched on in sequence by command from the system control, using addresses coded in an 8 X 6 matrix. Three extra input channels are available; one is left unused, the other two are reserved for calibration purposes. Calibration of zero level is achieved by addressing a channel with grounded input; the second calibrate channel is supplied with a precision dc voltage to produce a full-scale signal level of 60 dB for this channel.

The *Sample-Hold Circuit* takes two samples of its input simultaneously. One microsecond after the samples are taken, a decision is made to select the one sample that is within the linear operating range. This decision, designated "coarse-range," is also used by the A/D converter and the squarer. Polarity inversion is switched in, if needed, so that a positive voltage is always delivered to subsequent circuits.

The A/D Converter uses floating-point binary arithmetic to achieve a constant-percentage accuracy. After the coarse-range decision (as noted above), fine ranging is performed to locate the signal in a 2:1 amplitude group. Next a 3-bit group of digits completes the process, for a total effective conversion range of 15 bits.

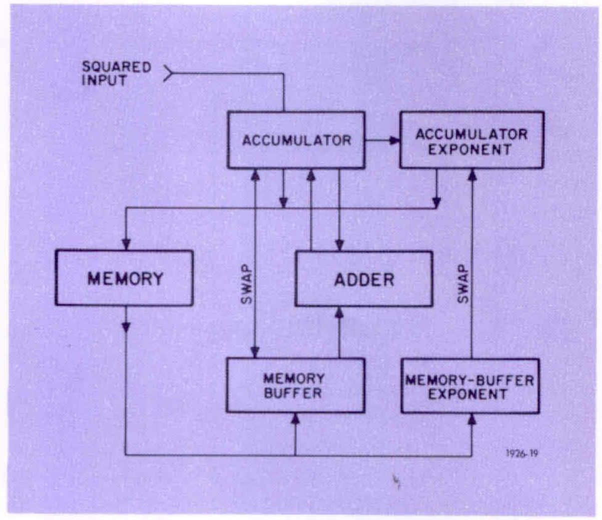


Figure 9. Block diagram of computational mode.

The *Squarer* accepts binary inputs from the sample-hold circuit and the A/D converter to generate the squared value of each input sample. Its output is in a floating-point format that is consistent with the arithmetic of the computer.

The computational mode is dominated by the *ADDER AND MEMORY* block of Figure 8, shown in more detail in Figure 9. This represents a very simple, but true, digital computer. The squared numbers are delivered to the accumulator in two parts: a mantissa and an exponent. Simultaneously, the number in memory is taken into the memory buffer as a mantissa and an exponent; the two exponents are then compared. Since only the accumulator has provision for shifting digits, the number with the smaller exponent must be in the accumulator. If a swap is necessary to get the number with the smaller exponent into the accumulator, a logic circuit orders the swap made. Next the exponents are made equal. The accumulator bits are shifted to the right by one place and the accumulator exponent is increased in value by one digit.

Once the exponents in the accumulator and memory buffer are equal, the digits in corresponding stages of the mantissas have the same weight and may then be added.

The output of the adder is returned to the accumulator for one last check. If the result of the addition involves a "carry",

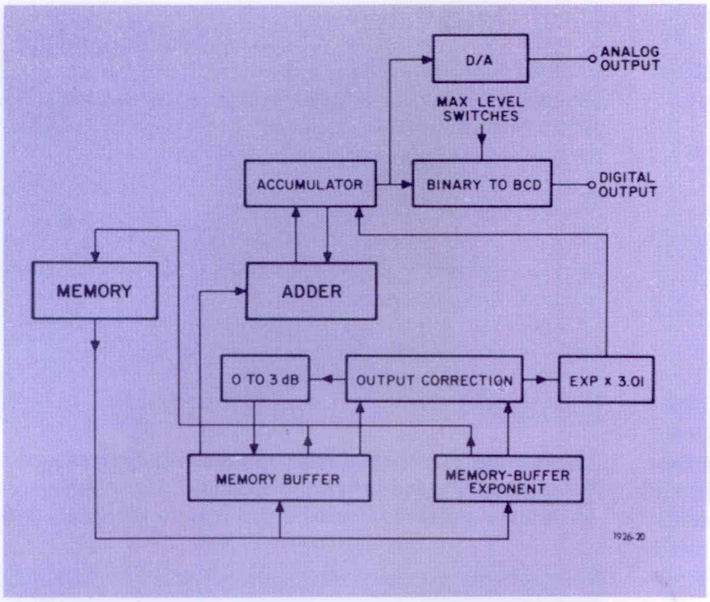


Figure 10. Block diagram of output mode.

so that the accumulator contains 16 rather than 15 bits, then the mantissa must be cut back to size. This is done simply by shifting the mantissa one place to the right and increasing the exponent again by one digit. Therefore, if there is a carry to 16 places, one more shift and add will be needed. If there is no carry, the accumulator is finished with its work on this particular squared-input number.

The contents of the accumulator and accumulator exponent are put back into the memory unit. Another squared number is presented to the accumulator, and the whole process repeats.

The output mode begins when all data gathering is completed. Figure 10 shows the sequence of data flow. In the first step, the stored sum-of-the-squares is drawn out of the memory into the memory buffer. This number is fed to a block marked *OUTPUT CORRECTION*. If the time of integration was less than 1 second, the sum of squares is a smaller number than if the time had been 1 second or longer (as the time is reduced below 1 second, fewer samples are taken). The *OUTPUT CORRECTION* block multiplies the sum of the squares by 2, 4, or 8 if the integration time is 1/2, 1/4, or 1/8 second, respectively. The numbers derived from the mantissa in the sum of squares give a contribution of between 0 and 3 dB to the final answer. This contribution is determined in the *0 TO 3 dB* block of Figure 10. The numbers derived from the exponent in the sum of the squares must be multiplied by 3.01.

The circuits in the *MEMORY-BUFFER* and the *MEMORY-BUFFER EXPONENT* blocks are clocked JK flip-flops. They can accept new data into storage as the previously held data are read out. The next step in the output sequence takes advantage of this fact; the sum of the squares, in binary format, is returned to the memory for future use at the same moment that the result of the *0 TO 3 dB* determination is fed into the memory buffer. Simultaneously, the product of *EXP* X *3.01* is deposited in the accumulator. The accumulator and memory buffer are then connected to the adder, which gives the sum of the *0 TO 3 dB* and the *EXP* X *3.01* contributions. The adder output is the binary representation, in decibels, of the sum of the squares that was just taken from, and returned to, memory.* The adder output is then transferred to the accumulator, which now contains the decibels in binary form. This result is converted to a binary-coded decimal (BCD) in 1-2-4-8 format. If the front-panel MAXIMUM BAND LEVEL dB control calls for the range of answers to be other than 0 to 60 dB (e.g., 20 to 80 dB), an appropriate constant scaling factor in decibels is added at this time.

The final output consists of

(1) BCD representation of the decibels calculated from the sum of the squares plus the calibration factor, if any, added by the MAXIMUM BAND LEVEL dB control, and

(2) An analog voltage proportional to the number of decibels, ***not including*** the calibration factor.

If the exponent in the sum of the squares is too large, the instrument concludes that an overload condition existed in the input signal. It then causes the answer to be reported as 800-some or 900-some dB's. Answers up to 62.75 dB (not counting any additional decibels added by the MAXIMUM BAND LEVEL dB control) are reported correctly.

*Note that the square root is never taken explicitly; the conversion is directly from the sum of the squares to decibels.

Acknowledgment

The authors gratefully acknowledge the support of a number of General Radio engineers who have contributed greatly to this development program. Jim Esselstyn was responsible for the mechanical design and packaging of the entire GR 1921 system. Matt Fichtenbaum participated in early work on the detector and might properly be described as the computer section architect. Dave Nixon was involved in the early cost and feasibility study on the detector. Ralph Anderson designed all the power supplies for the system. Jim Faran designed the output D/A converter for the detector and did the original work on the input A/D converter. Carl Woodward has been involved in interfacing accessories with the system. Arnold Peterson acted as consultant during the entire program.

Complete specifications for the GR 1921 are included with this issue as a tear sheet, removable for insertion in GR Catalog T.

Some Notes on Digital Detection

A digital detector computes rms voltage, starting with the sample variance formula

$$\sigma^2 = \frac{1}{n} \sum_{i=1}^{n} V_i^2 \qquad (1)$$

where n is the number of samples included in a measurement and V_i are the input voltages that occur at discrete instants of time. To obtain an output in decibels, it is necessary to find

$$\mathrm{dB} = 10 \log_{10} \left(\frac{\sigma}{\sigma_0}\right)^2 \qquad (2)$$

where σ_0 is the desired reference level.

The input signal to be measured is a continuous function of time. It is sampled only at discrete instants of time when it is accepted for digital processing. Two questions naturally arise. How many samples are needed for a measurement, and when should they be taken? To answer these questions, we must examine both random and periodic input signals.

Random Inputs

First, consider the case of a stationary random input signal with a zero mean voltage and a Gaussian amplitude distribution.

Curves of measurement repeatability versus independent signal samples are shown in Figure 11. These curves are used in pairs and represent the upper and lower limits of expected measurement variations. They are commonly known as confidence limit curves, where the confidence in percent specifies the probability that a measurement lies within the area bounded by the curves. The most efficient sampling scheme for purely random inputs is uniform sampling, with a sample spacing equal to the correlation time of the input. Samples must be statistically independent to be effective for measurement purposes and, hence, must not be taken at intervals closer than the correlation time.

For example, a white noise limited to the frequency range of 0 to BW hertz has a correlation time of $\frac{1}{2BW}$ second This corresponds to the Nyquist rate, which is the required sampling rate for signal reconstruction. A lower sampling rate is perfectly acceptable for variance measurement.

Now consider the measurement of Gaussian noise with the GR 1921 analyzer. The GR 1925 filter bands have a constant fractional bandwidth and hence the correlation time will vary with the center frequency. For a 1-second integration time, 1024 independent samples will not be available from all bands. The number of independent samples available from a particular band will vary depending on the integration period selected. These factors were included, along with the results of Figure 11, to generate the composite confidence limits for 1/3-octave-filter center frequency and are shown in Figure 12.

Periodic Inputs

For stationary random inputs, we simply require enough measurement samples to obtain statistical stability (repeatability). With periodic inputs, however, we shall see that the sample rate is also a critical parameter. Consider the case where the input $V(t)$ is a sinusoid of unit variance, unknown frequency f, and phase ϕ.

$$V(t) = \sqrt{2} \cos (2\pi f t + \phi) \qquad (3)$$

With n samples of the input and a uniform sample spacing of δ seconds, the variance estimate is

$$\sigma^2 = \frac{2}{n} \sum_{i=1}^{n} \cos^2 (2\pi f i\delta + \phi) \qquad (4)$$

Errors in the variance estimate are primarily due to the number of samples n and the sample spacing δ. Phase angle ϕ can be regarded as a random variable that is uniformly likely to be

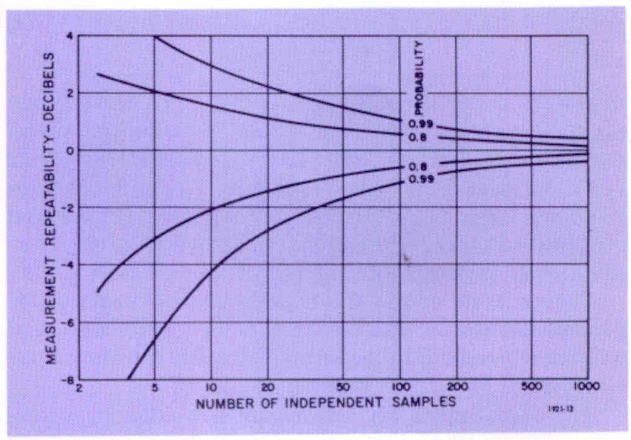


Figure 11. Confidence limits of measurements (in general).

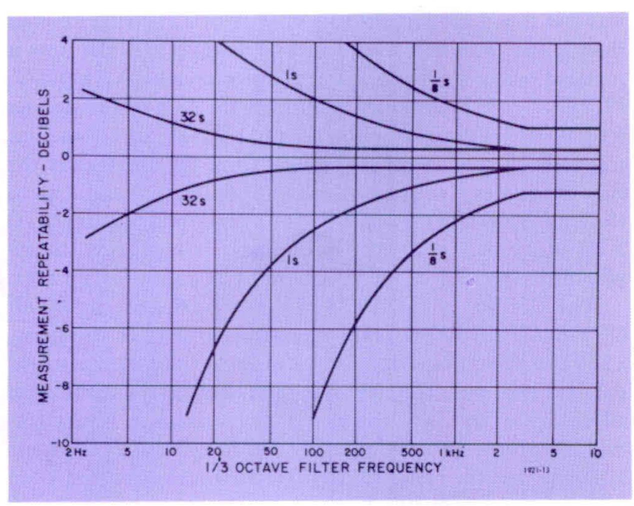


Figure 12. Composite 99% confidence limits of measurements (1/3 octave).

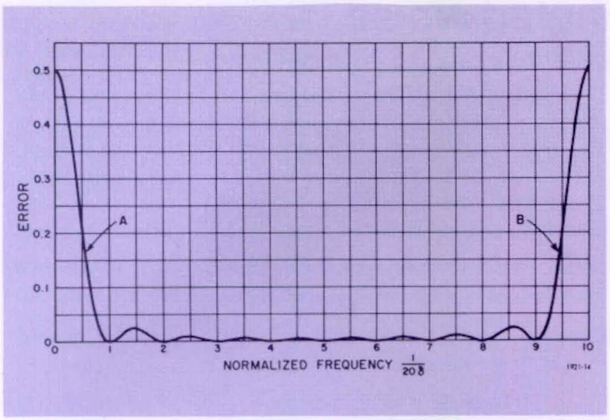


Figure 13. Measurement errors expected for fixed sampling rate.

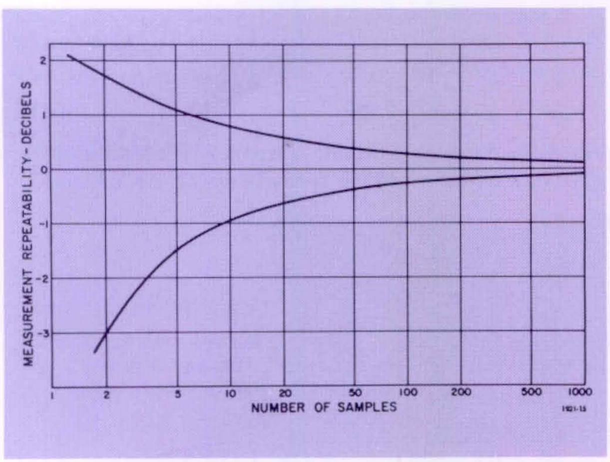


Figure 14. Range of repeatability of measurements.

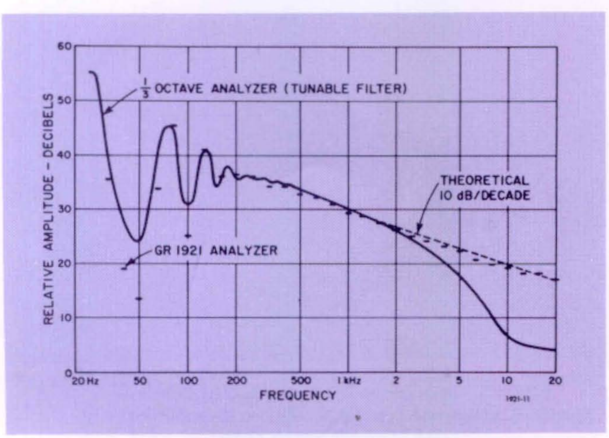


Figure 15. Measurements comparison at 25 Hz -- Tunable 1/3-octave filter versus GR 1921.

between 0 and 360 degrees. Averaging the effects of ϕ in equation (4) and subtracting the true variance, we have the mean-squared error result:

$$\epsilon^2 = \frac{1}{2}\left(\frac{\sin 2\pi n f}{n \sin 2\pi f}\right)^2 \qquad (5)$$

Equation (5) is plotted in Figure 13 for the case of $n = 10$, to show the effects of input frequency f relative to sample spacing δ. The largest errors occur in the frequency ranges A and B. The remaining error lobes are relatively small in magnitude.

The large error lobes repeat at input frequencies that are multiples of one-half the sample rate. Because these frequencies lie within the input-frequency range of the GR 1926 detector, uniform sampling is unsatisfactory.

Modulating the sampling rate will cause the frequencies to shift. A shift of the sampling frequency during a measurement interval will shift the frequencies at which peak errors occur, thus distributing the error. This principle is incorporated in the 1926 detector design. The sample rate is swept during each measurement period, over a 2:1 frequency range. With 1024 samples for measurement, the worst-case error is reduced to about 0.2 dB.

The measurement repeatability for various values of n, assuming a modulated sampling rate, can be approximated by integrating curves like that of Figure 13. The results are plotted in Figure 14, which shows the dependence of repeatability upon n, the number of samples per measurement. These curves represent the 1σ limits of the range of measurement repeatability.

So far we have considered only the special case of a sinusoidal input signal. Other periodic signals should also be considered. A gated sinusoid (tone burst), for example, is measured with less than 1024 samples. The sample number reduction is in proportion to the duty factor. The measurement repeatability is determined by the number of effective samples and is specified by the curve of Figure 14.

A Difficult Spectrum

Consider the output of a 1/3-octave filter set with a square-wave input. At low frequencies there is only one harmonic in the the output of each 1/3-octave filter. At high frequencies, however, the filters' outputs are periodic sequences consisting of the step-function responses of the filters. The results of analyzing a 25-Hz square wave with a tunable 1/3-octave filter and analog detector are compared with the results of measuring this same signal with a GR 1921 Real-Time Analyzer in Figure 15. A dashed line with a slope of 10 dB/decade shows the theoretical spectrum shape.

At low frequencies there is slight disagreement because of a difference in filter shapes. At a frequency of 10 kHz, a measurement error of about 13 dB is made by the analog detector because of the high crest factor of the filter output signal, while the expected measurement repeatability from the GR 1921 (and indeed the measured variation) is less than 1 dB.

–J. A. Lapointe

Rayleigh-Distributed Noise

A Method for Reproducing a Singular Type of Interfering Background Noise

Noise that appears at the output of envelope detectors such as those in some radio or radar systems typically has the Rayleigh amplitude distribution rather than the Gaussian. The Gaussian density distribution function (Figure 1) is the familiar symmetrical bell-shaped curve; the Rayleigh distribution function (Figure 2) is quite nonsymmetrical, being zero for negative values of noise voltage. In experiments on signal detection or intelligibility, it may be important to reproduce accurately the type of interfering background noise. Because the Rayleigh distribution function is very small where the noise voltage is near zero, it cannot be approximated accurately by full-wave rectification of Gaussian noise.

One of the most common examples of a voltage having Rayleigh distribution is that seen at the output of a detector whose input is narrow-band random noise. Generating Rayleigh noise in this manner involves amplification, filtering, detecting, and smoothing. This necessarily results in considerable reduction of the bandwidth of the Rayleigh noise compared to that of the original Gaussian noise.

There is a procedure for starting with Gaussian noise and generating Rayleigh noise which does not require filtering and smoothing and the attendant bandwidth reduction. Start with two statistically-independent Gaussian random noises, square each and add, and take the square root of the sum. These analog operations are indicated in Figure 3. Proof that this procedure yields the Rayleigh distribution is given below. The argument is similar to that used in solving the famous problem of the random walk in two dimensions (References 1, 2). The joint probability distribution function of two Gaussian random variables x and y, having rms values of 1 (that is, $\sigma = 1$), is

$$p(x, y)\, dx\, dy = \frac{1}{2\pi} e^{-\frac{x^2 + y^2}{2}}\, dx\, dy.$$

Consider the distribution of the variable r where $r^2 = x^2 + y^2$. In terms of r,

$$p(r)\, r\, dr\, d\theta = p(x, y)\, dx\, dy = \frac{1}{2\pi} e^{-\frac{r^2}{2}}\, r\, dr\, d\theta$$

(where the element of integration $dx\, dy$ has been replaced by its counterpart in polar coordinates, $r\, dr\, d\theta$). Integrating over all angles,

$$p(r)\, dr = \frac{1}{2\pi} e^{-\frac{r^2}{2}}\, r\, dr \int_0^{2\pi} d\theta$$

$$= re^{-\frac{r^2}{2}}\, dr, \qquad r > 0. \tag{1}$$

This is the Rayleigh distribution function. The procedure for generating r from x and y is precisely that shown in Figure 3.

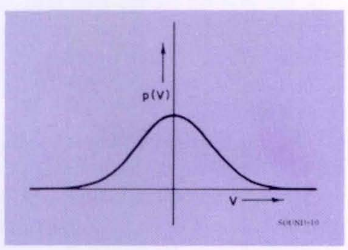


Figure 1. The Gaussian probability density distribution function.

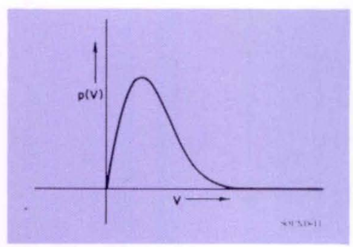


Figure 2. The Rayleigh probability density distribution function.

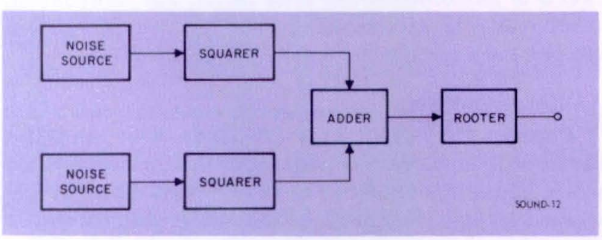


Figure 3. System for producing noise having a true Rayleigh distribution.

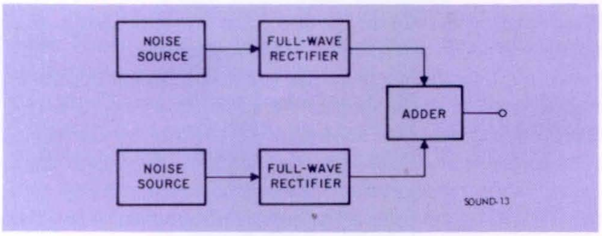


Figure 4. Simpler system for producing noise having approximately the Rayleigh distribution.

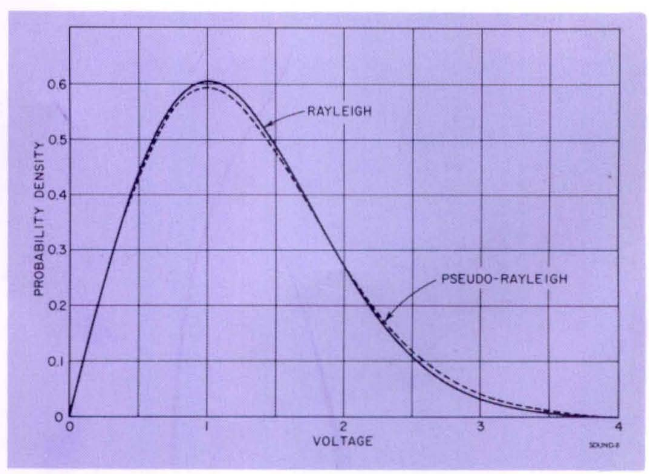


Figure 5. Comparison of amplitude distribution of pseudo-Rayleigh noise (suitably adjusted in amplitude) with that of true Rayleigh noise.

There is a much simpler procedure for generating a close approximation to the Rayleigh distribution. Shown in Figure 4, it avoids the use of the squaring circuits (squarer) and the square-root circuit (rooter).

To determine the actual amplitude density distribution function of the output of this system (pseudo-Rayleigh noise), we start with the Gaussian distribution function for each noise source,

$$p(x)\,dx = \frac{1}{\sqrt{2\pi}}\, e^{-\frac{x^2}{2}}\, dx,$$

where again, for simplicity, σ has been taken equal to 1. After full-wave rectification (mathematically, taking the absolute value) the distribution becomes, for each noise source,

$$p_1(x)\,dx = \sqrt{\frac{2}{\pi}}\, e^{-\frac{x^2}{2}}\, dx, \qquad 0 \leqslant x.$$

The probability that the sum $(x+y)$ will have a value between x_0 and $(x_0 + dx_0)$ is

$$p_A(x_0)\,dx_0 = \int_0^{x_0} p_1(x)\,p_1(x_0 - x)\,dx \bullet dx_0,$$

where the integration is taken over all possible values of x which could lead to a sum x_0. (Remember that $x > 0$ and $(x_0 - x) > 0$.) Substituting for $p_1 x$, and dropping dx_0,

$$p_A(x_0) = \frac{2}{\pi}\int_0^{x_0} e^{-\frac{x^2}{2}}\, e^{-\frac{(x_0 - x)^2}{2}}\, dx$$

$$= \frac{2}{\pi}\, e^{-\frac{x_0^2}{4}} \int_0^{x_0} e^{-\left(x - \frac{x_0}{2}\right)^2}\, dx.$$

Now, letting $y = x - \frac{x_0}{2}$,

$$p_A(x_0) = \frac{2}{\pi}\, e^{-\frac{x_0^2}{4}} \int_0^{\frac{x_0}{2}} e^{-y^2}\, dy$$

$$= \left(\frac{2}{\sqrt{\pi}}\, e^{-\frac{x_0^2}{4}}\right)\left(\frac{2}{\sqrt{\pi}}\int_0^{\frac{x_0}{2}} e^{-y^2}\, dy\right) \quad (2)$$

$$= H'\!\left(\frac{x_0}{2}\right) H\!\left(\frac{x_0}{2}\right),$$

where H' and H are precisely the functions tabulated in Reference 3. When the values of x_0 are multiplied by 0.8 and the values of $p_A(x_0)$ are multiplied by 1.25 we find the agreement with the true Rayleigh distribution shown in Figure 5. These multiplications simply amount to changing the amplitude of the pseudo-Rayleigh noise; the area under the probability curve remains equal to unity. In the figure $p_A(x_0)$ is compared with the function $p(r)$ from Equation 1. The pseudo-Rayleigh noise has an amplitude distribution which is extremely close in form to the true Rayleigh distribution.

Examination of Equation 2 shows that the behavior of $p_A(x_0)$ when x_0 is very small is the same as for the Rayleigh distribution; however, the approximation behaves as $\exp\left(-\frac{x_0^2}{4}\right)$ instead of $x_0 \exp\left(-\frac{x_0^2}{2}\right)$ for large values of x_0. The latter behavior hardly shows on the linear plot for Figure 5. This difference does result in a detectable change in the number of peaks above a given level.

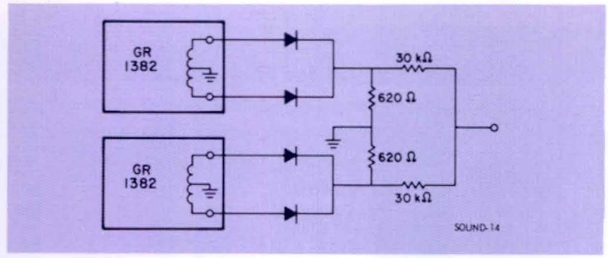


Figure 6. System for producing pseudo-Rayleigh noise.

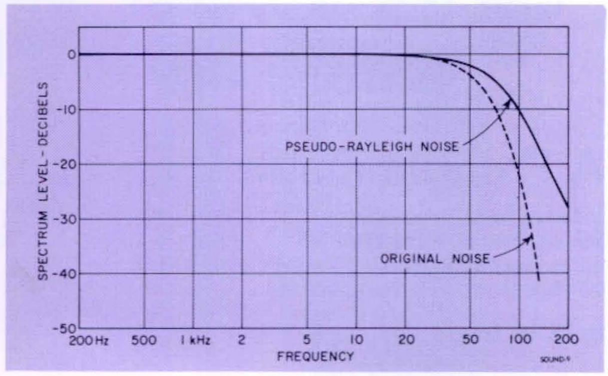


Figure 7. Comparison of pseudo-Rayleigh noise spectrum with the output spectrum of the GR 1382 generator.

A schematic diagram of how pseudo-Rayleigh noise can be produced from two GR 1382 Random-Noise Generators [1] is shown in Figure 6. High-conductance germanium diodes, such as 1N455 or 1N695, should be used for low forward voltage drop. A dc amplifier will be needed if it is necessary to reduce the output impedance level or to provide more power.

The effect on the spectrum of this method is shown in Figure 7. This is a comparison of the measured input and output spectra of the system shown in Figure 6. In Figure 7, the spectra are normalized to the same level at low frequencies. This method of approximating Rayleigh noise simply spreads the spectrum a little towards higher frequencies. Similar spreading probably will occur with the true Rayleigh system shown in Figure 3.

This easily assembled system for generating pseudo-Rayleigh noise includes only noise generators, diodes, and resistors. It produces noise having a close approximation to the Rayleigh amplitude distribution.

—J. J. Faran

A brief biography of Dr. Faran appeared in the March/April, 1969 issue of the ***Experimenter***.

[1] Faran, J. J., "Random-Noise Generators," *General Radio Experimenter*, January 1968.

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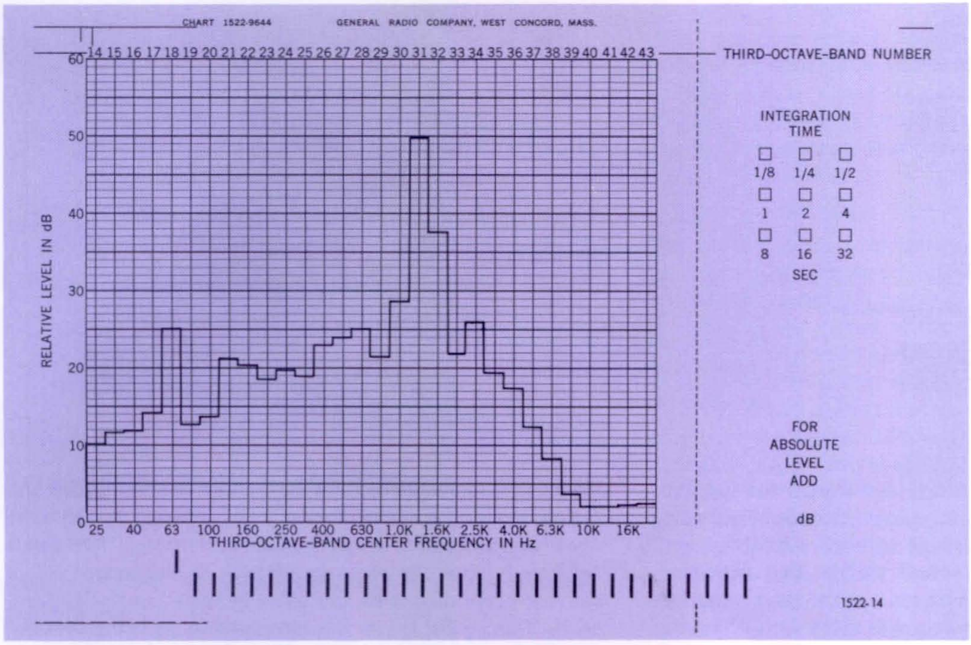

Figure 1. Typical recording programmed by control marks.

excursions beyond preset tolerances, and of time relationships between charts made on several recorders simultaneously.

- Reinforced mounting sheets (8½″ by 11″) are available for GR 1522 charts. Two adhesive strips are provided on the sheets to facilitate the mounting of recorded charts, which can then be filed in data notebooks.
- The Y-axis recording action of the GR 1522 is much faster than that of the X-Y recorder-65 inches/second as compared with 20 inches/second for the X-Y recorder.
- The X-axis of the GR 1522 is expandable manually or automatically, with local or external programming; it can range from 10 to 100 inches to permit detailed physical and temporal waveform investigation. The X-Y recorder is limited to 10 to 20 inches.
- Control of the X-axis motion is by digital rather than analog signals. For this reason it is possible to convert angle or shaft position to chart position directly, using an optical or commutator encoder.
- Isolated grounding provisions avoid ground loops between the GR 1522 and other instruments.
- External sync signals are available to operate several recorders synchronously.
- The GR 1522 is capable of use with small computers, provided proper interfacing is supplied.

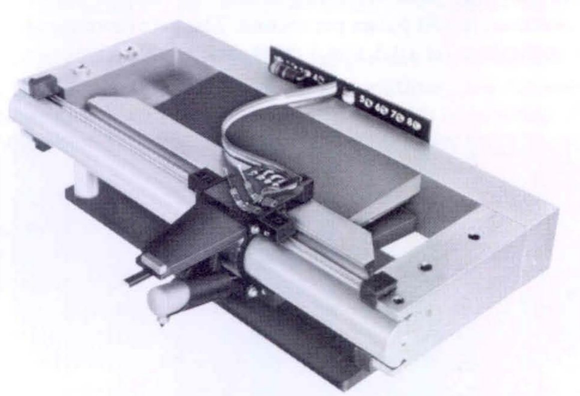

Figure 2. Servo-motor pen drive.

Some Design Facts

Backbone of the 1522 is a field-proven dependable servo motor,* Figure 2, consisting of an aluminum-wire coil wound on a light-weight plastic form. This coil assembly is supported by four miniature ball bearings gliding in a V-groove formed in a longitudinal iron bar. The iron bar is part of the magnetic path produced by a large permanent magnet that maintains a uniform magnetic field in the gap through which the coil assembly moves. When a current is applied through the coil, a force is generated perpendicular to the current path and magnetic field. This force moves the coil assembly to right or left, depending on the polarity of the current, imparting a true linear motion. The ball bearings cause very little backlash and have very little friction, even though the force is off center. (Force is applied in the magnetic gap, not in line with the bearings.) The pen and potentiometer wiper arm are mounted directly on the motor coil; this technique eliminates backlash between pen and wiper arm of the feedback potentiometer.

The operation of the servo loop (Figure 3) starts with the introduction of a dc signal through input circuitry designed for a wide range of voltages and currents. The scaled input is compared, within a high-impedance amplifier, to the feedback-potentiometer position voltage. The output of the am-

*US Patent 2,581,133 owned by General Radio Company.

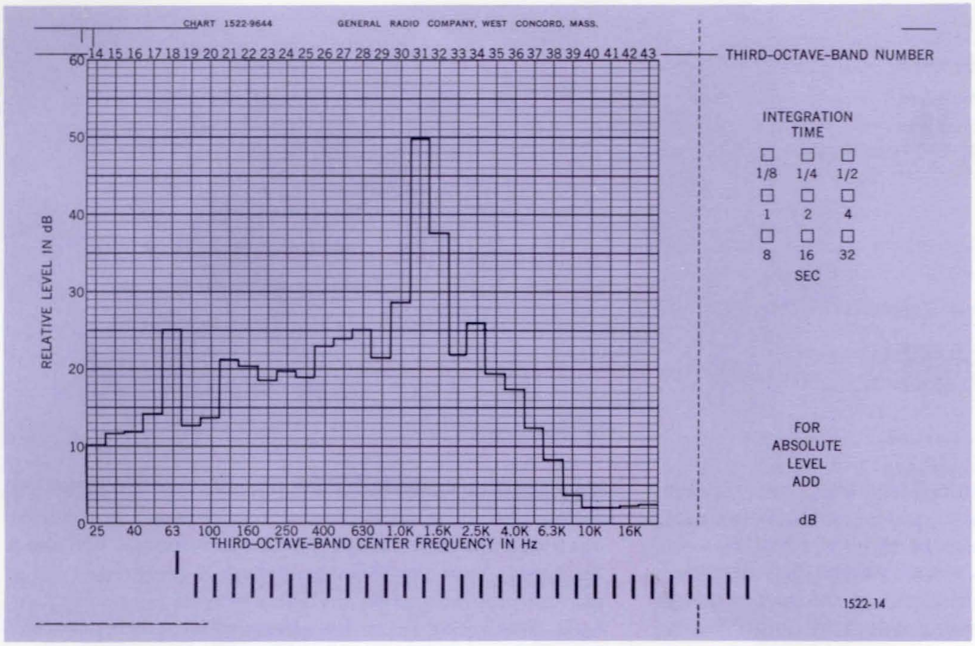

Figure 1. Typical recording programmed by control marks.

excursions beyond preset tolerances, and of time relationships between charts made on several recorders simultaneously.

- Reinforced mounting sheets (8½″ by 11″) are available for GR 1522 charts. Two adhesive strips are provided on the sheets to facilitate the mounting of recorded charts, which can then be filed in data notebooks.
- The Y-axis recording action of the GR 1522 is much faster than that of the X-Y recorder-65 inches/second as compared with 20 inches/second for the X-Y recorder.
- The X-axis of the GR 1522 is expandable manually or automatically, with local or external programming; it can range from 10 to 100 inches to permit detailed physical and temporal waveform investigation. The X-Y recorder is limited to 10 to 20 inches.
- Control of the X-axis motion is by digital rather than analog signals. For this reason it is possible to convert angle or shaft position to chart position directly, using an optical or commutator encoder.
- Isolated grounding provisions avoid ground loops between the GR 1522 and other instruments.
- External sync signals are available to operate several recorders synchronously.
- The GR 1522 is capable of use with small computers, provided proper interfacing is supplied.

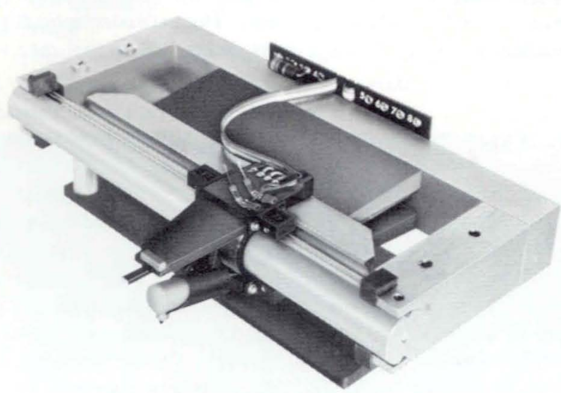

Figure 2. Servo-motor pen drive.

Some Design Facts

Backbone of the 1522 is a field-proven dependable servo motor,* Figure 2, consisting of an aluminum-wire coil wound on a light-weight plastic form. This coil assembly is supported by four miniature ball bearings gliding in a V-groove formed in a longitudinal iron bar. The iron bar is part of the magnetic path produced by a large permanent magnet that maintains a uniform magnetic field in the gap through which the coil assembly moves. When a current is applied through the coil, a force is generated perpendicular to the current path and magnetic field. This force moves the coil assembly to right or left, depending on the polarity of the current, imparting a true linear motion. The ball bearings cause very little backlash and have very little friction, even though the force is off center. (Force is applied in the magnetic gap, not in line with the bearings.) The pen and potentiometer wiper arm are mounted directly on the motor coil; this technique eliminates backlash between pen and wiper arm of the feedback potentiometer.

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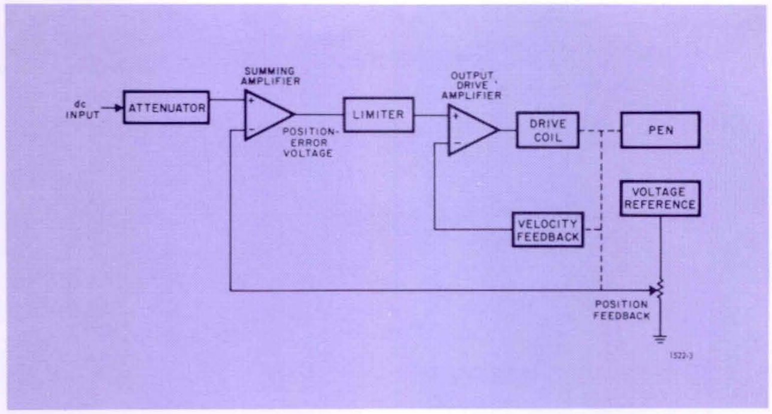


Figure 3. Block diagram of servo system.

plifier is the position-error voltage and is zero for the null position. If the input dc signal increases, a position-error voltage will be generated, which is fed through a limiter, to be compared with the velocity-feedback voltage. This difference is amplified to become the drive voltage for the pen motor assembly. The pen and wiper arm will move until the position-feedback voltage equals the input voltage, and the position-error voltage again is zero.

The velocity-feedback-voltage signal permits high feedback-loop gain at the null point (no velocity feedback), resulting in high static accuracy; dead-band error is less than 0.15%. The loop gain is effectively lowered during dynamic operations, keeping the position-feedback loop stable at high speeds.

Velocity feedback is derived from an unusual bridge technique (see Figure 4). Note that the output-drive amplifier serves both the linear servo motor and another fixed coil wound on a similar iron core. These two coils, together with two resistors, form a bridge network that can be balanced for ac as well as dc signals. When drive current is fed into this bridge network and the pen is stationary, the bridge is balanced; the output from the bridge and differential amplifier is zero. If the servo motor moves, the servo-motor coil develops a voltage proportional to the velocity; the fixed coil does not. Thus, there is a voltage output from the differential amplifier proportional to velocity.

Increasing the velocity feedback can simulate the action of a low-pass filter. The open-loop frequency response of this type of servo system is shown in Figure 5. As the velocity feedback is increased beyond the amount required for stabilization, the loop gain is decreased. The frequency at which the open-loop and closed-loop responses pass through 0 dB also is decreased. Thus the servo bandwidth is compressed by an amount proportional to any extra increase in velocity feedback. This action filters the effects of noise and undesired fluctuations in recording.

The position-feedback potentiometer is situated directly above the motor-drive coil. It is made of conductive plastic which assures low maintenance and extremely long life, greater than 20 million cycles, with negligible increase in noise or wear.

Full control of the chart drive, without mechanical clutches, is assured by use of three motors. Two induction motors, operating in a stalled condition, provide a constant holdback for chart supply and take-up and are instantly responsive to forward-or-reverse-direction instructions. The third motor, the stepper type referred to earlier, drives the sprockets that engage perforations in the chart, imparting a dependable motion to the chart at all times. Without the use of gears in the drive train, the recorder has eighteen chart speeds, from 0.5 s/in. to 20 h/in.

Operation of the stepper motor is controlled by a multivibrator, phase-locked to the ac power frequency, either 50 or 60 Hz. Its regulation and stability are similar to those of a synchronous motor. The maximum rate at which the motor will respond, without losing a step for instant start-stop operation, is 300 pulses per second. This rate corresponds to 2 inches/second (0.5 s/in.) chart speed. The actual rate of

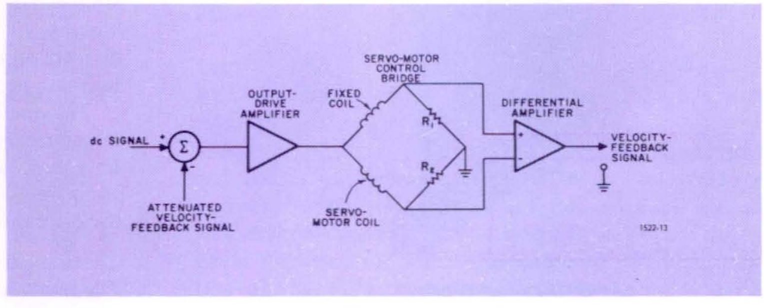


Figure 4. Schematic of velocity-feedback derivation.

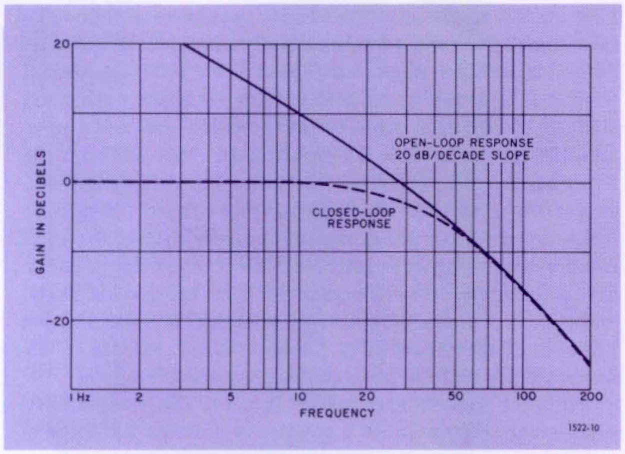


Figure 5. Servo-loop response.

rotation is locally or remotely controlled through integrated-circuit divider chains.

The output from the divider goes through a gating circuit to the stepper-motor driver circuit. This gate is controlled by a flip-flop for instant start-stop operation. The input to the flip-flop is capacitively coupled to both the front-panel switch and to the external programming input. Either input can be overridden by the other; the last signal determines the logic state. Logic memory is important for chart control; only a momentary pulse is necessary to set the state. Thus, when the photocell is used to change chart direction, the memory keeps the direction fixed after the mark is no longer under the photocell.

A forward-reverse flip-flop controls the direction of the stepper motor. Motor direction can be controlled locally or remotely

Output-control information, for use in system or automatic applications, is available from the recorder. As they pass over a photocell, the black timing marks on the chart (Figure 1) trigger output controls. Two photocells control as many as three output drives. When two in-line black marks pass the photocells at the same time, the paper motion is stopped. This method is used to signal the end of the chart roll; the ending of the roll is also verified by a printed note on the paper.

A switch is provided that lifts the pen and moves the chart at the fastest paper speed for scan operations; pen lift also occurs during scanning of the chart. In the AUTO position of the pen lift, the pen is dropped to the paper only in the RECORD mode and while the chart is moving.

Maximum and minimum limit switches are adjustable to any point along the chart. They provide electronic control for external operations such as sorting, inspecting, or sounding an alarm.

Situated next to the feedback potentiometer is a linear take-off potentiometer that can provide an external voltage proportional to the position of the pen. This voltage, and commercially available multiple integrated-circuit comparators, can establish limit-set positions controllable by an external potentiometer or voltage. The limit stops can be programmed externally, if a control voltage is used.

Another programming feature is pen blanking; when the servo system is turned off, pen motion freezes. The pen can be stopped anywhere on the chart during a scan operation, or during any operation that normally would force the pen off-scale during switching.

System Compatability

There are many possibilities for assimilating the GR 1522 into measurement and control systems. You are already aware of the functions performed by the GR 1921 Real-Time Analyzer (page 3) and of the value attached to permanent records of noise and vibration signals. The GR 1522, designed as a systems recorder, is a perfect complement to the GR 1921. A detailed description of the close working relationship between the recorder and analyzer may trigger readers' thoughts of other useful applications.

The 1926 detector stores the level of all the signal-input channels. It then feeds this information to the recorder at a rate that is controlled by the recorder's ability to accept and digest the data. Chart movement is continuous until the photocell control stops the recorder and changes channels in the detector. A fixed time delay of 0.1 second in the detector allows the pen to reach the signal level of the new channel before the detector sends a control signal to the recorder to start the chart. The pen then records the level of this channel at the fastest paper speed. The resultant plot has square corners, closely resembling the actual scope display. The start-stop-mark sequence repeats until the last channel selected by the analyzer is recorded. The pen automatically lifts, and the next chart moves rapidly into position. At this time, chart motion stops and the recorder does not function until commanded by the analyzer. It takes approximately 8 seconds to complete one chart and to advance to the next chart.

Table 1 shows the time required to make such a recording on three types of recorders: GR 1522, automatic X-Y recorder, and typical X-Y recorder.

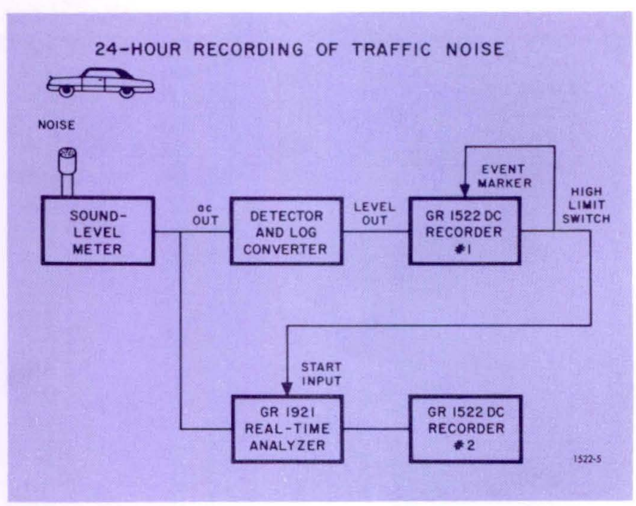


Figure 6. Typical long-time recording/analysis assembly.

Table 1

Recorder Type	Recording Time (seconds)	New Chart Advance Time (seconds)	Total Time (seconds)
GR 1522	6	1-2	7-8
Automatic X-Y	15	1	16
Typical X-Y	15	45	60

It is readily apparent that the 1522 takes one-half the time for each recorded spectrum, as compared with the automatic X-Y recorder, and is far superior to the typical X-Y unit.

Typical Applications

A practical application of the 1921/1522 combination, using two recorders (Figure 6), is the real-time spectrum analysis of band levels above a preset acoustical level. This might be a 24-hour study of traffic noise, aircraft fly-over noise, or similar patterns. During this interval we record, on Recorder No. 1, the average level of noise picked up and processed through a detector and log-converter circuit. Obviously, we don't want a spectrum every eight seconds because we would have 10,000 charts of recorded data in a 24-hour period. To save chart space and to compress accumulated data, the recorder is set at a slow chart speed. The 1921 analyzer is triggered by a GR 1522 limit switch to make a spectrum analysis of signals above a preset threshold level and also to identify the occurrence by printing an event mark on the Recorder No. 1 chart. After the GR 1921 is triggered, Recorder No. 2 receives instructions from the 1921 system to make the spectrum recording that corresponds to the indicated event. Upon completion of the study, a complete record is available, from Recorder No. 1, of the average sound level, and accumulated spectrum analyses of noise above a preset limit are presented by Recorder No. 2. All this can be accomplished without personnel in attendance.

Another system application involves the temperature-coefficient (Figure 7a) of a practically unlimited number of components. A scanner and the GR 1654 Impedance Comparator can measure 100 components consecutively, feed the information to the GR 1522, and plot the percent deviation from a standard at room temperature. A second plot can be

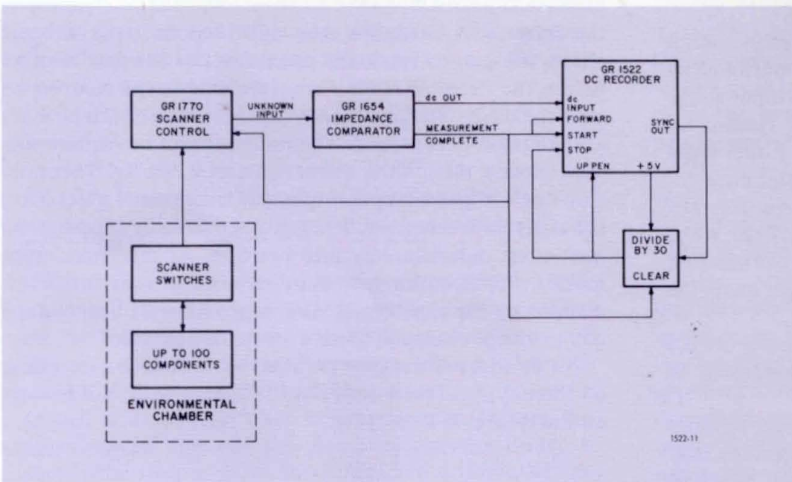


Figure 7a. Block diagram for multi-component temperature-coefficient measurements.

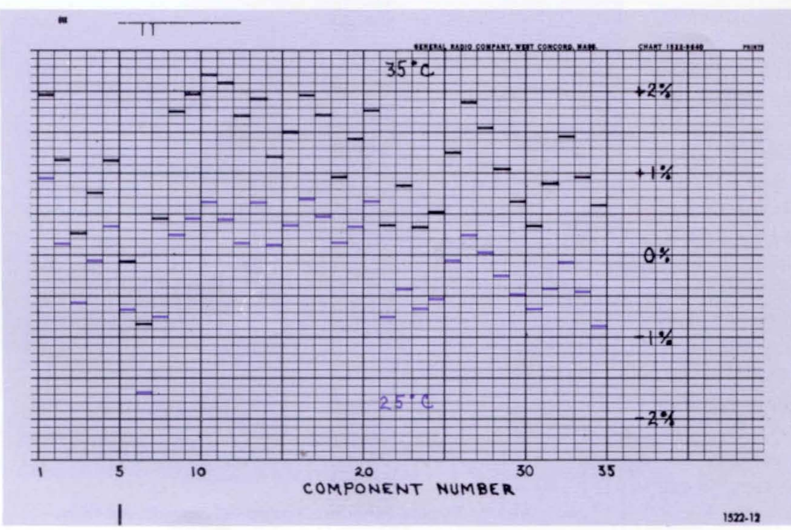


Figure 7b. Component test results. Chart motion is controlled by external divider circuit.

made using a second-color pen for a different temperature (Figure 7b). A measuring sequence is started by a switch on the scanner control. Upon switching to a component, the comparator waits for balance and then gives a measurement-complete command. This time delay allows the pen to move to the next component-recording position. The measurement-complete signal from the GR 1654 starts the recorder and paper motion. This particular chart paper has no control marks; the chart motion is controlled by external means. An external flip-flop divider divides, by 30, a sync signal used to drive the stepper motor. After 30 pulses, the chart has been moved a recording interval of 0.2 inch. At this instant, motion stops and the pen lifts. The same measurement-complete command is fed back into the scanner control, and the entire record cycle is repeated after the short delay period to achieve balance.

Remote programming for process control facilitates unattended recorder operation. A signal level, beyond preset limits, from a control transducer, triggers accelerated chart motion for greater detail, operates the scanner control to permit observation of other significant transducers, and then stops the process.

In another typical application, we can monitor a zener-diode voltage output. Here, both long-term drift and short-term noise are important. Photocell-control marks every 11 inches are used to trigger a flip-flop circuit that permits a long-term performance recording for 110 minutes (10 min/in.). This recording is followed by an accelerated recording of noise for 55 seconds (5 s/in.). Each recording is alternately and automatically presented in a continuous sequence of 11-inch charts.

To encourage specific application of the recorder to the GR 1921 analyzer, four different charts to cover various channel combinations and bandwidths used in the GR 1921 will soon be available from stock. Recorder channel widths of 0.208, 0.250, and 0.500 inch are available, dependent upon spacing of the photocell control marks on the chart edge. An unlimited variety of charts using various placement of the marks can be supplied on special order to customers.

M. W. Basch received his degrees at MIT (BSEE and MSEE-1958) and joined GR in 1958 as a development engineer in the Industrial Instruments Group and later transferred to the Acoustics/Signal Analysis Group. He has specialized in design of recorders and associated instrumentation. His memberships include IEEE, Tau Beta Pi, and Eta Kappa Nu.

ACKNOWLEDGMENT

The author gratefully acknowledges the help of M. C. Holtje in several critical areas and the support of R. P. Anderson for the power-supply design, of G. E. Neagle for the mechanical design, and of P. A. d'Entremont for the industrial design.

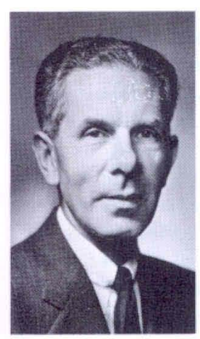
Complete specifications for the GR 1522 are included with this issue as a tear sheet, removable for insertion in GR Catalog T.

The Honorable Society

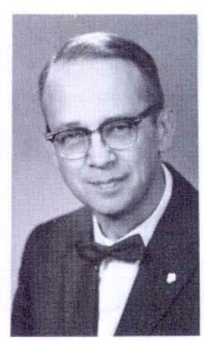
Long-time readers of the *Experimenter* will note with interest honors accorded two of their friends.

On May 1, 1969 Paul K. McElroy was awarded the annual Contribution Award of the IEEE Parts, Materials, and Packaging Group, at ceremonies conducted during the Electronics Components Conference of the Electronic Industries Association. "PK", now retired from GR, but hardly less active professionally than while he was with us, came to the attention of our readers for the first time in 1926. It was then he published a two-part article entitled "Design and Testing of Plate Supply Devices." He prepared a steady stream of written articles for almost forty years, terminating his output with "A 100-μF Decade Capacitor" in July, 1965. Our debonair retiree resides at 58 Gayles Road, Belmont, Mass.

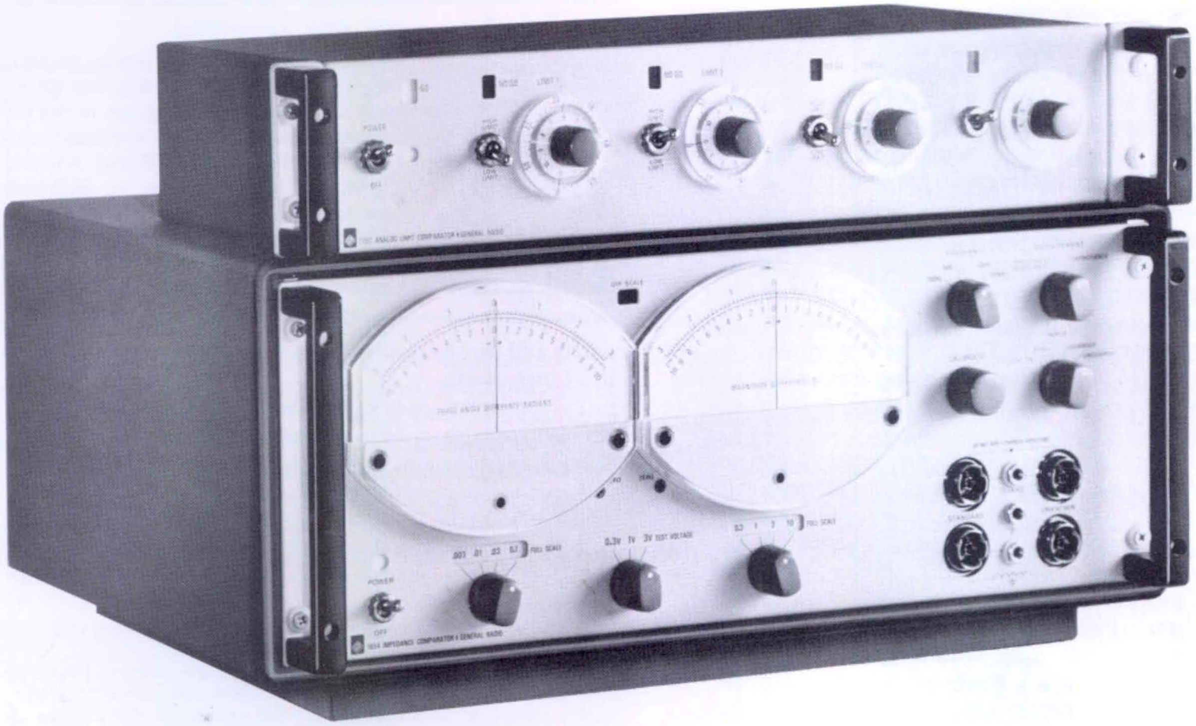
The other recipient of honors was Dr. Arnold P. G. Peterson, GR consultant in the areas of sound and vibration. As part of ceremonies conducted to mark the twentieth anniversary of the foundling of the Audio Engineering Society, Dr. Peterson received the John H. Potts Memorial Award on October 23, 1968. The bronze medallion named for the deceased publisher and editor of "Audio Engineering" was given in recognition of outstanding achievement in the field of audio engineering. The citation noted Dr. Peterson's "many prominent contributions to the design of audio and acoustic instrumentation." Some readers may remember the first *Experimenter* article by APGP entitled "The Type 757-A UHF Oscillator" in the August 1941 issue; more may recall his latest article, "A Magnetic Tape Recorder For Acoustical Vibration And Other Audio-Frequency Measurements" in October 1966. It appears to be time for another article!



P. K. McElroy



A. P. G. Peterson



Type 1654 Impedance Comparator with its companion accessory, Type 1782 Analog Limit Comparator.

IMPEDANCE COMPARISON SPRINTS AHEAD

Mobility in measuring impedances, coupled with accuracy, wide range, and ability to control!

The Goal

Over the years, GR has tried to relieve the tedium and monotony of passive component inspection and testing by introducing functional-designed test equipment. For instance, we feel that technicians are able to remain alert for longer periods of time when test data are presented in simple GO/NO GO patterns. Eliminating the time normally required to read panel-meter indications also results in a saving in labor costs. In our estimation, the design of the new GR impedance comparator has successfully joined the concept of complex, programmed testing with the concept of simple, individual component measurements, at no sacrifice in accuracy or reliability.

The first comparison bridge engineered by General Radio for production testing was the manually-operated GR 1604 Comparison Bridge.[1] Several years later, it was followed by the GR 1605 Impedance Comparator,[2] incorporating greater precision and versatility, simultaneously indicating phase-angle and magnitude differences between two external test impedances over an extended frequency range, and requiring no bridge balancing. This bridge was semi-automatic in operation.

In 1964 the GR 1680-A Automatic Capacitance Bridge[3] brought with it a component test rate better than two per second. The inspection-rate bottleneck was broken, and the unit's acceptance by industry encouraged introduction of another digital unit - the GR 1681 Automatic Impedance Comparator System.[4] The customer received digital readout of phase-angle and magnitude differences, higher accuracy and resolution, and greater compatability with automatic component- and data-handling equipment.

Many analog-measurement operations and customers still existed. The need to extend and expand the design of analog comparators was the incentive to develop the latest type of GR comparator, which incorporates automatic features.

The Achievement

The GR 1654 Impedance Comparator supersedes the GR 1605 comparator and, with its companion accessory the GR 1782 Analog Limit Comparator, provides a semi- or fully-automatic system at about one-third the cost of a similar digital system. Measurement features include

- Operating range from 100 Hz to 100 kHz in four steps.

[1] Holtje, M. C., "A New Comparison Bridge for the Rapid Testing of Components," *General Radio Experimenter*, December 1952.

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

[3] Fulks, R. G., "The Automatic Capacitance Bridge," *General Radio Experimenter*, April 1965.

[4] Leong, R. K., "The Automatic Impedance Comparator," *General Radio Experimenter*, June-July 1968.

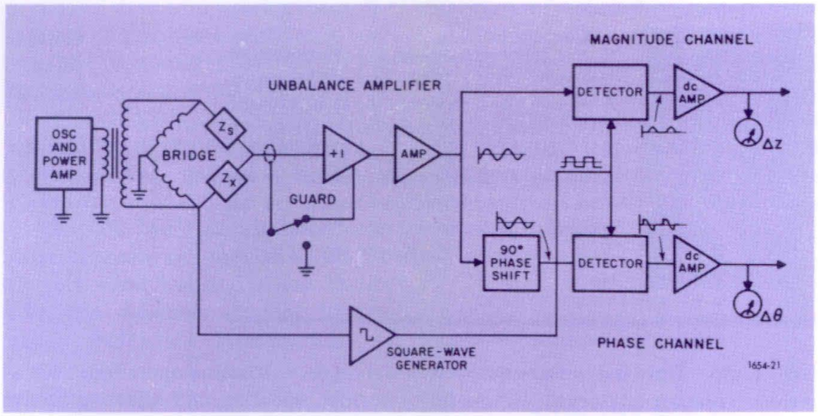


Figure 1. Block diagram of GR 1654 Impedance Comparator.

- Comparison precision to 30 parts per million.
- Impedance range from 2 Ω to 20 MΩ at 100 Hz.
- Capacitance range as low as 0.1 pF direct reading, with a modified substitution method.*
- Six deviation ranges from 0.1% to 30% full scale.
- Test voltages from 0.3 to 3 volts, for easy voltage-coefficient tests.
- Operating rate up to 4 units per second, in the automatic test mode.
- Positive indication of direction of overload.
- Indications on large panel meters.
- Analog output voltages.

*Refer to Circuit Notes at end of article.

The block diagram of the 1654 comparator, Figure 1, illustrates use of a tightly-coupled 1:1 ratio toroidal transformer as two arms of the bridge; the standard and test impedances complete the circuit. The unbalanced-output signal of the bridge is fed through a guarded circuit, which effectively reduces cable capacitance by three orders of magnitude. This permits measurement, with negligible error, of test items as remote as thirty feet. From the amplifier system the signal is passed to the magnitude and phase channels.

Input to the magnitude-channel phase detector is direct. Input to the phase channel first undergoes a 90° phase shift, before connection to the phase detector, in order to bring the test frequency signal in phase with the error voltage component due to any phase difference. The phase detector is fundamentally a switch operated in synchronism with the test frequency. Exact switching is controlled by a square wave derived from the zero phase signal of the bridge.

The rectified voltage is that component of the error voltage which is in phase with the controlling square-wave voltage. The detected output is fed to a stable dc operational amplifier which provides the required analog output voltage that is interpreted as magnitude difference or phase-angle differ-

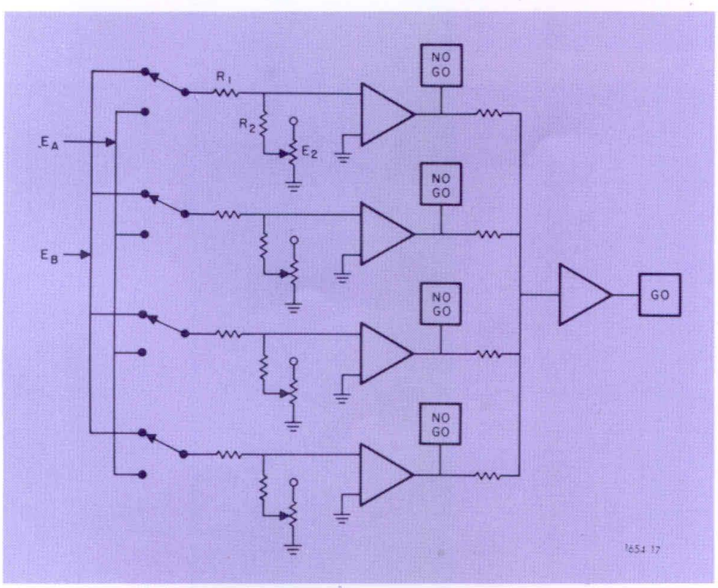


Figure 2. Block diagram of GR 1782 Analog Limit Comparator.

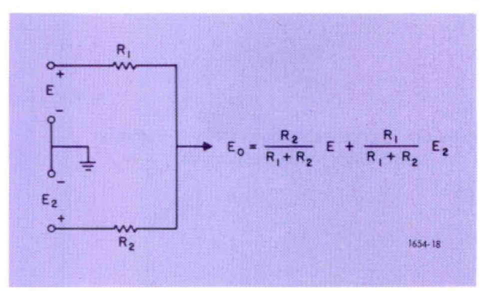


Figure 3. Schematic of limit comparison function.

ence between the test item and the standard.

The 1782 accessory unit (Figure 2) incorporates front-panel lamps to indicate GO or NO GO conditions. Each NO GO lamp indicates a single test limit, established as a preset voltage E_2. The input voltage E, derived from the magnitude or phase channels of the 1654, is compared with E_2, as shown in the schematic diagram, Figure 3. Unbalance voltage E_o, if it exists, is amplified by an operational amplifier with sufficient positive feedback to cause the amplifier to switch off or on. Its output triggers the NO GO lamp drivers; if the comparison is out of established tolerance, the NO GO lamps will light. Within-tolerance conditions for all limits will trigger the GO lamp. Optional relay-equipped models operate external automatic sorting devices.

R. K. Leong holds degrees from Northeastern University (BSEE-1960 and MSEE-1962). After two years' service in the U. S. Army as 1st Lt., coordinating instrumentation for upper atmosphere research, he joined GR in 1964 as a development engineer in the Low-Frequency Impedance Group. He has specialized in bridge and comparator design. He is a member of Tau Beta Pi and Eta Kappa Nu.

Applications

On its own, the GR 1654 is a work horse on the inspection line or in the laboratory. Routine operations such as sorting, selecting, and adjusting passive components (R, L, and C), and any complex assembly of these components, are accomplished as quickly as you can make connections to the bridge. The GR 1680-P1 Test Fixture

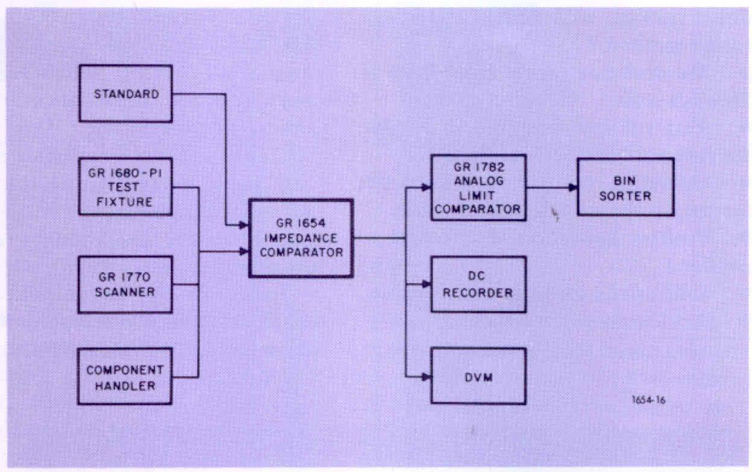


Figure 4a. Typical instruments and devices for automatic component-measurement system.

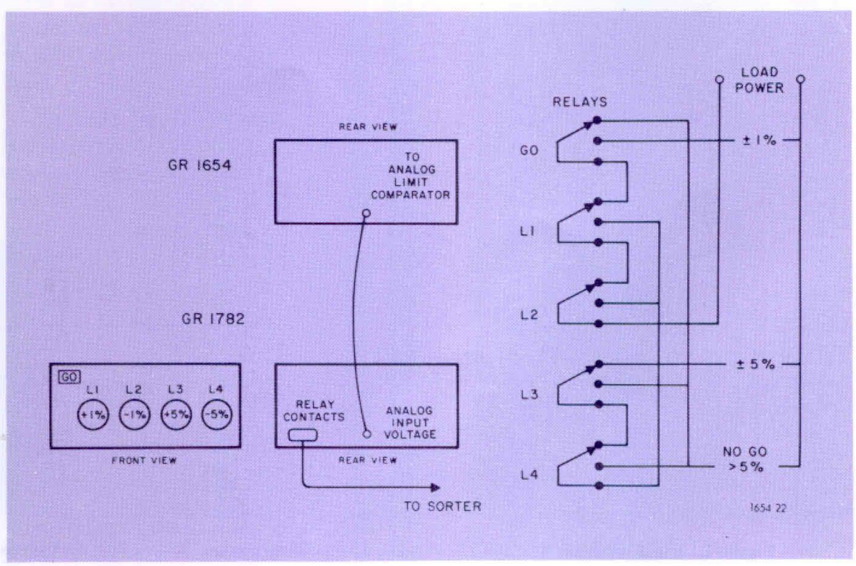


Figure 4b. Typical sorting system - schematic and relay interconnections.

is an ideal device for holding manually-inserted components. For faster sorting, the comparator and supporting instruments and accessories are an efficient automatic component-measuring system, as shown in Figure 4a. A typical sorting system is shown in Figure 4b.

The GR 1654 also is suitable for measurements of capacitors of extremely small values, low-loss capacitors, and dielectric samples; temperature-coefficient measurements of components tested within environmental chambers are a routine matter. Production testing of back-biased diodes and transistor-collector junctions is also routine; testing and adjusting of ganged capacitors and potentiometers for desired tracking within tolerances is another useful application.

–R. K. Leong

Complete specifications for the GR 1654 and 1782 instruments are available in the 1969 supplement to GR Catalog T.

Catalog Number	Description	Price in USA
	1654 Impedance Comparator	
1654-9700	Bench Model	$1300.00
1654-9701	Rack Model	1250.00
	1782 Analog Limit Comparator	
	Bench Model	
1782-9700	without relays	550.00
1782-9702	with relays	625.00
	Rack Model	
1782-9701	without relays	570.00
1782-9703	with relays	645.00

All prices subject to quantity discount.

CIRCUIT NOTES

The GR 1654 bridge circuit (Figure 5) measures impedance difference as a percentage of the average of the standard (Z_s) and unknown (Z_x) impedances. The real part of small phase-angle differences can be derived from the equation

$$\text{Re}\ \frac{E_o}{E} = \frac{|Z_x| - |Z_s|}{|Z_x| + |Z_s|} \quad (\text{When } \theta_x - \theta_s \leqslant 0.1 \text{ radian})$$

Introduction of nonlinear networks into the magnitude channel results in a linear indication of the impedance difference as a percent of the standard

$$\text{Re}\ \frac{E_o}{E} = \frac{|Z_x| - |Z_s|}{|Z_s|} \times 100\%$$

and reduces the number of scales for total measurement range. Measurements of R, L, and C furnish magnitude differences as percentages:

$$\frac{R_x - R_s}{R_s} \times 100\%, \quad \frac{L_x - L_s}{L_s} \times 100\%, \quad \frac{C_x - C_s}{C_s} \times 100\%$$

The imaginary part of small phase-angle differences can be obtained from the equation

$$\text{Im}\ \frac{E_o}{E} = \frac{\theta_x - \theta_s}{2}$$

For measurements of R, L, and C, the phase-angle difference indication is a measure of

$$\Delta D \text{ of } C \text{ and } L, \text{ or } \Delta Q \text{ of } R.$$

Measurement of low-value capacitances usually is limited by input-terminal capacitance C_{IN}. The capacitance effect is mitigated by the connection of the standard C_A, as shown in Figure 6. If the value of C_A is chosen as approximately $10 \times C_X$, or greater, the magnitude difference readout is proportional to $\frac{2C_X}{C_A + C_{IN}}$. For example, if C_{IN} is typically 1 pF and $C_A + C_{IN} = 200$ pF, for a meter reading of 0.1% the value of $C_X = 0.1$ pF. Note that the bridge is *direct* reading.

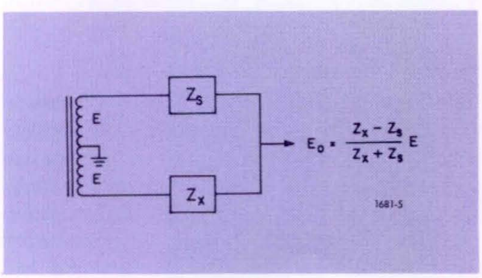


Figure 5. GR 1654 Bridge circuit.

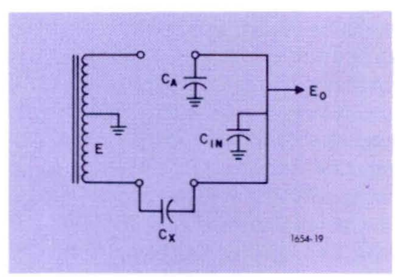


Figure 6. Bridge schematic for small-capacitance measurement.

Information Retrieval

Indexes for Volumes 41 and 42 (1967, 1968) of the *Experimenter* are available upon request to the editor.

VHF and UHF Attenuation Measurement to 140 dB[1]

This paper describes a system assembled for rapid measurement comparisons of attenuators with values ranging to 140 dB, with basic resolution of 0.01 dB and with maximum errors of 0.08 dB. A simplified system for measurements to 110 dB at certain selected frequencies is described also, as well as refinements permitting spot frequency measurements to 170 dB.

by S. Brown Pulliam

Introduction

The present-day design of commercially available standard-signal generators is highly sophisticated.* Quality assurance testing by the manufacturers must rely upon state-of-the-art techniques. An example is the testing of output-signal attenuators, which control the signal level over a range of 0 to 140 dB in 10-dB steps. Maximum allowable error is 0.1 dB per step; maximum cumulative error is 0.5 dB. The frequency range for these requirements is from dc to 500 MHz. This article deals with the test measurement system that was designed to assure instrument performance to specifications.

The System Choice

R. W. Beatty of NBS has classified the various attenuation-measurement systems;[2] our system falls in the general category "Direct Substitution." Several techniques are available under this system, including series and parallel substitution. The major advantage of the series substitution technique is that, as the test attenuator is adjusted in level, the reference attenuator is similarly adjusted but in the opposite direction. The total circuit attenuation therefore remains constant and, the signal level into the detector being relatively constant, nonlinearity is not a problem. In the parallel-substitution technique, the test signal is split into two paths, one through the test attenuator and the other through the reference attenuator, and is recombined in a detector. This has the advantage that any instability in the output level of the source, or in the gain of the detector, has no effect upon measurement accuracy. However, in order to make the system work, either an auxiliary rf switch or a precision phase shifter is required.

Most commercially available attenuation-measurement systems use an i-f substitution method in which the reference attenuator is calibrated at one frequency only. These systems are limited to a maximum single-step range of approximately 100 dB due to the dynamic range of the mixer.

There is no definite basis for a choice between the series- and parallel-substitution techniques; however, with sufficient level stability of signal source and detector, the series-substitution technique is simpler to implement.

Any practical system for measuring large values of attenuation involves detection of a weak signal. Most engineers are aware of the fact that increased sensitivity involves narrowing of the receiver bandpass but may overlook the fact that some receiver systems contain elements which have a "threshold effect." One such element – the second detector of the usual single-conversion superheterodyne receiver – will lose an applied signal if it is buried in random noise,[3] and no amount of post-detection filtering can retrieve it. Fortunately, in most attenuation-measurement applications the available test signal is large enough, using conventional superheterodyne techniques, to obtain detection resolution equal to that obtainable by very sophisticated synchronous detection schemes.

System Design Considerations

Little of the literature concerning attenuation measurement systems is devoted to the problem of ease or rapidity of readout versus system sensitivity. Measurement range can be increased by narrowing the bandwidth, and, similarly, resolution can be increased by lengthening the measurement time. Decisions must be reached on what resolution or precision is required in a particular measurement and what maximum attenuation must be inserted in the measurement channel. The remaining significant variables are time, generator power, and detector noise figure. Time

*Examples are General Radio Types 1003-A and 1026-A.

[1] Abstracted from paper of same title scheduled for publication in the March, 1969 issue of the *IEEE Transactions on Instrumentation and Measurement.*

[2] Beatty, R. W., "Microwave Attenuation Measurements and Standards," *NBS Monograph 97,* April 1967.

[3] Downing, J. J., *Modulation Systems and Noise*, Chapter 4, Prentiss-Hall Inc., 1964.

is the most difficult variable to resolve and involves factors which are quite subjective. An experimental conclusion reached was that peak-to-peak meter jumps should be less than 0.02 dB in a measurement interval less than 5 seconds. This in turn required an equivalent signal-to-noise ratio in the receiver of at least 50 dB in a 1-Hz bandwidth. Assuming a total attenuation of 140 dB to be measured, the ratio of input signal to output noise is 190 dB. Signal generator power was calculated to be +21 dBm, considering a theoretical available thermal noise level of −174 dBm for a 1-Hz bandwidth (300 K operating temperature) and assuming a receiver noise figure of 5 dB. Available powers from the signal generators under test were +27 dBm (Type 1026-A) and +22.5 dBm (Type 1003-A), presenting no problem.

The signal versus noise requirements outlined above, combined with the need to avoid "threshold-effects" losses, resulted in selection of a pre-detection bandpass of 1 kHz. This immediately imposed stringent frequency-stability requirements upon the measurement system.

Working in the uhf spectrum with an LC tuned signal generator, which could easily drift more than 1 kHz, required an auxiliary phase-lock loop. The dual phase-lock system shown in Figure 1 was developed for the uhf tests. A synchronizer phase locks the source frequency, and a second loop locks the second i-f signal to a 50-kHz crystal-controlled reference frequency. This permits optional use of a wave analyzer or selective voltmeter if extension of the attenuation measuring range to 170 dB is required. A lock-loop bandwidth of 500 Hz is more than adequate to correct for any frequency drift in the crystal-controlled converter, when the 50-kHz phase-detector output is used to control the synchronizer reference frequency.

The basic readout instrument of the attenuation-measurement system is a 1-dB expanded-scale meter at the i-f amplifier output. It was slightly modified to provide a one-second time constant. An auxiliary readout recorder has become an integral part of the measurement system because it reduces operator fatigue considerably.

S. B. Pulliam joined the engineering staff of General Radio in 1959 as a member of the Signal-Generator Group. He had received the BS degree in Physics from Union College in 1951, served in the US Navy for three years as Electronics and Communications Officer, and also had been associated with the Naval Ordnance Laboratory as Electronic Scientist. Mr. Pulliam is a member of IEEE.

The vhf-measuring system developed is identical to the original uhf system with the exception that the converter is replaced by a 79-MHz converter. No synchronizer is required in this system. The frequency stability of either signal generator is more than adequate to stay within the 500-Hz locking range of the secondary loop, and the dc control is connected directly to the signal-generator varactor terminals.

System Performance Evaluation

Tests of the drift in level of the combined signal generator and 30-MHz amplifier indicate a stability of ±0.01 dB, for a time interval of 20 seconds. Further tests have indicated drift does not exceed ±0.015 dB for intervals up to 10 minutes.

Signal levels at the receiver input as low as −143 dBm produce successful phase locking. For useful measurements at this level, however, a second i-f bandpass on the order of 10 Hz is required. This may be provided by the wave analyzer as shown in Figure 1.

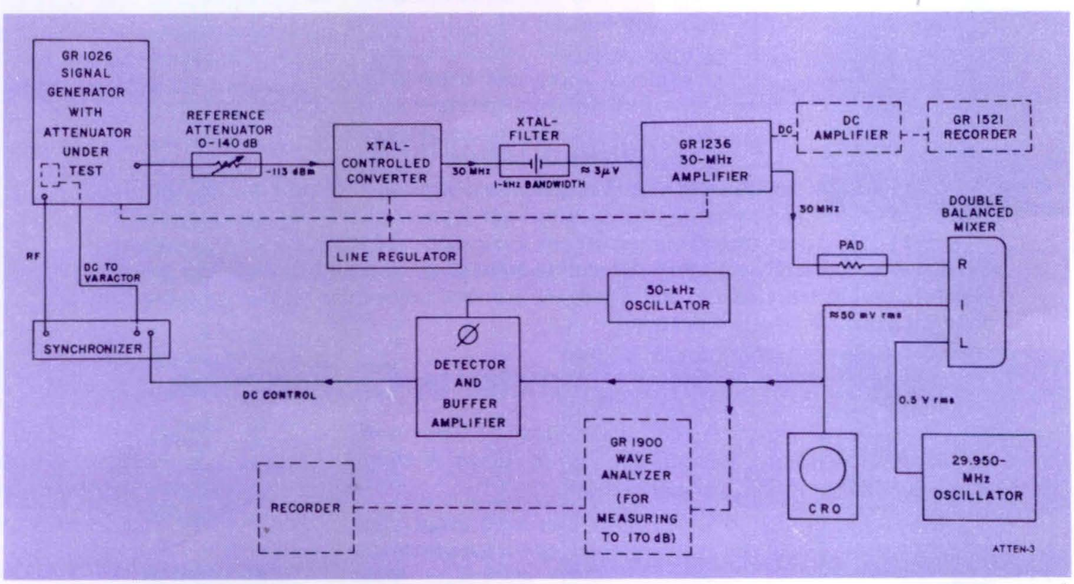


Figure 1. Block diagram of attenuation-measuring system.

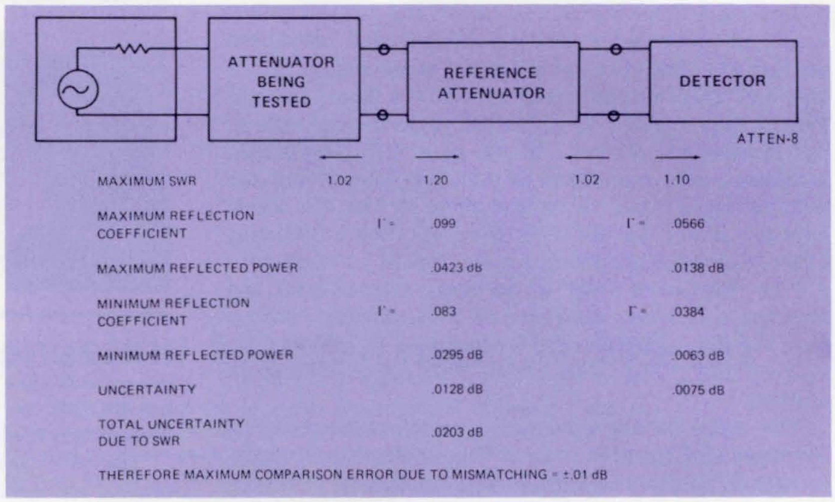


Figure 2. Significant reflection errors for interior steps of test and reference attenuators.

Figure 2 presents the interface reflections for the interior 100 dB of the attenuators and results of calculations for the interior steps. Attenuation comparison error for all interior steps is calculated to be ±0.01 dB. For the extreme steps an additional reflection pair between the generator and the attenuator being tested causes a maximum error that is calculated to be 0.06 dB.

Much attention was directed to rf leakage around the attenuator path because of the large amounts of attenuation involved. Exceptional care was taken to shield the source from the detector. A leakage signal which is coherent with the detected signal can cause an error of 0.01 dB if its level is 60 dB below that of the detected signal. A measurement at the 140-dB level would require shielding effectiveness of 200 dB. In addition to the requirement for well shielded signal sources and detectors, interconnections must also be well-shielded by solid-sheath coaxial cable. Tests made with the system described in this article showed that in no case was leakage found to be a disturbing factor for measurements to 140 dB.

The summary of the errors, shown in Table 1, indicates a worst-case comparison error of 0.08 dB for attenuations to 140 dB, and 0.57 dB to 160 dB.

Table 1
Summary of comparison errors

Source	Interior Attenuator Step	Extreme Values to 120 dB	to 140 dB	to 160 dB
Recorder resolution and system drift	±.01 dB	±.01 dB	±.01 dB	±.01 dB
SWR	±.01	±.06	±.06	±.06
Noise fluctuation	0	0	±.01	±0.2
Meter error (1000)	–	–	–	±0.3
Worst-case error	±.02 dB	±.07 dB	±.08 dB	±0.57 dB

Editor's Note – Copies of Mr. Pulliam's paper will be available from General Radio Company, 300 Baker Avenue, West Concord, Massachusetts 01781.

Seminar Scheduling

A new seminar has been added to General Radio's Customer Seminar Program. Its title is Real-Time Sound and Vibration Measurements and it covers theory and application of the GR 1921 Real-Time Analyzer, input/output equipment, and analyzer/computer systems. Two 3-day seminars were scheduled during June at the Concord facility. This fall, the seminars will be repeated both at Concord and in several major cities. Contact your local GR District Office for more information.

Recent Technical Articles by GR Personnel

"Instruments and Techniques of Sound Measurements," W. M. Ihde, *National Safety News,* December 1968.*

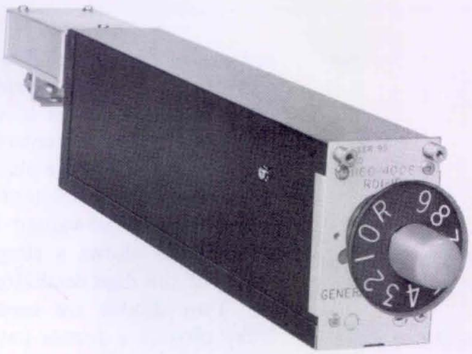
"Simplified Pure-Tone Audiometer Calibration," R. P. LaJeunesse, *National Hearing Aid Journal,* March 1969.**

"Tone-Burst Testing of Underwater Transducers," D. M. Lloyd, *Undersea Technology,* April 1969.*

"Recent Advances in Coaxial Components for Sweep-Frequency Instrumentation," T. E. MacKenzie, *Microwave Journal,* June 1969.

"Ratio Transformer Techniques for Precise Gain and Phase Measurements," T. J. Coughlin, *Electronic Instrument Digest,* June 1969.

*Reprints available from General Radio.
**Expanded Instrument Note (IN114) available from General Radio.



Type 1160-RDI-1B Digit Insertion Unit.

FASTER SWITCHING FOR 1160-SERIES SYNTHESIZERS

Why and How

The General Radio 1160-series of frequency synthesizers is significantly improved, primarily in programmed applications, with the introduction of the GR 1160 RDI-1B Digit Insertion Unit. It offers higher switching speed and practically unlimited switch life by replacing reed relays with semiconductor diodes. It retains and expands the remote-control capabilities of the original 1160 RDI-1 unit.[1]

The use of synthesizers has increased considerably since their introduction to the fields of electronic measurement and instrumentation. The synthesizer transfers the accuracy and stability of a frequency standard to a broad range of frequencies with almost unlimited resolution.[2] Additional capabilities, such as electronic frequency selection from an external program, have opened up many applications not anticipated previously and have caused a demand for increased switching performance. The new digit insertion unit is expected to meet this demand.

The modular arrangement of the General Radio synthesizer family allows us to use identical digit insertion units in a chain to control the successive digits in the number defining the output frequency.[3] One group produces frequency increments from 100 kHz to 0.01 Hz, depending upon its position in the chain. This group of units is identified as RDI-1 (the letter R indicates remote-control capability). The early programmable units contain reed relays. Switching is relatively slow (2 ms) and typical life approximates 10^8 operations. It should be noted that a number of this magnitude could be exceeded in relatively short time in an application involving continuous high-speed switching.

The new RDI-1B unit eliminates both disadvantages. Switching time now is less than 200 μs (an order of magnitude improvement), and life expectancy is now almost unlimited.

Figure 1 shows a transition from 90 kHz to 20 kHz. No phase discontinuity is visible with an oscilloscope sweep rate of 50 μs/cm. It is important to note that this kind of presentation permits no assessment of the accuracy of the newly-established frequency immediately after switching. Synthesizer systems employing any kind of narrow-band devices after the point of frequency switching will have stabilizing times and frequency errors which may not be insignificant in relation to the normally high accuracy of synthesizers. Higher-resolution measurement techniques are therefore required.

Figure 2 shows an example of the method that was employed to obtain repetitive switching sequences from 900 kHz to zero frequency, using a dc-coupled phase detector for observation. From inspection of the oscillograph, it is evident that the proper phase relationship is obtained in most cases in less than 100 μs (horizontal scale is 200 μs/cm). The lower trace represents the programming signal (voltage on "0" programming line).

Precision sweep capabilities, with the continuously-adjustable decade, account for numerous applications of GR synthesizers in network characterization. In addition to these capabilities, it is now possible to simulate a

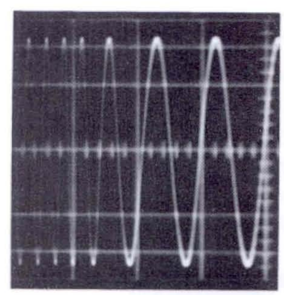


Figure 1. Switching transition, 90 to 20 kHz.

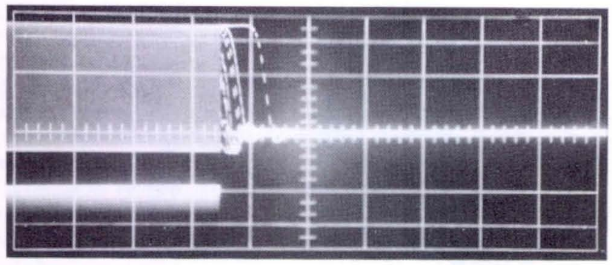


Figure 2. Switching sequences - 900 kHz to zero Hz.

[1] Lohrer, G. H., "Remote Programming for GR Synthesizers," *General Radio Experimenter*, May 1965.

[2] "Applications for Coherent Decade Frequency Synthesizers," General Radio Company, October 1968. Copies available on request.

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

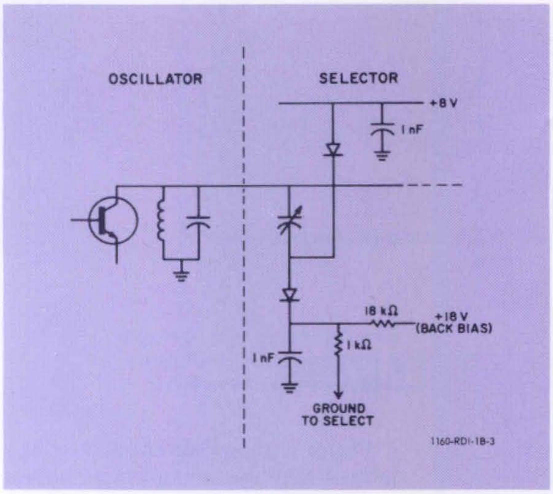


Figure 3. Simplified diagram of diode-switched oscillator - one section of selector only.

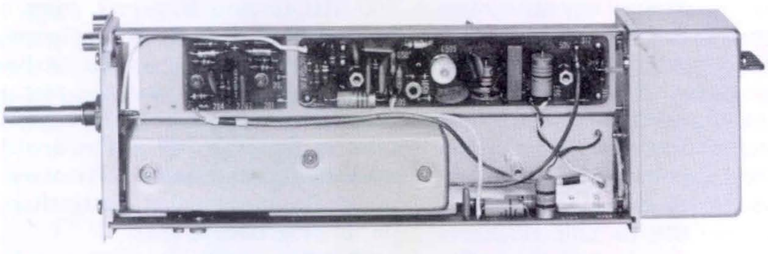


Figure 4. Interior details of RDI-1B unit.

sweep by remotely controlling frequency in suitably small increments, with no restrictions imposed by relay-contact life. The GR RDI-2, -3, and -4 insertion units already are programmable by the electronic control in steps of 1 MHz or 10 MHz.

The new RDI-1B unit replaces both the DI-1 and the RDI-1 units. Synthesizer models without remote frequency programming are still available; in the future they will be supplied with the RDI-1B module but without remote-control filtering. Filter plugs and remote cables are available separately, on special order, for remote-control programming, if desired at a later date. A 10-line code is normally provided but 1-2-4-8 BCD coding is available on special order.

Design Details

An important aim of the new design was circuit simplification. This was accomplished by using a simple transistor input mixer, a new phase-lock circuit design, and a divide-by-ten circuit differing considerably from the previous design. A higher switching speed was desired in the phase-lock loop that selects the proper digit from a picket fence of frequencies. We obtained it, and also an economic advantage, by employing silicon diodes to connect the different frequency-control capacitors and by redesigning the phase-lock loop for highest speed consistent with proper attenuation of unwanted frequencies.

Figure 3 shows a simplified schematic of the digit oscillator and selector. Two diodes are used per digit. They provide a double path for the rf current to ground in the "on" state and obviate the need for an rf choke in the "off" state. The energized end of the capacitor is isolated by means of the diode back bias. The digit oscillator is part of the active filter that selects the proper digit from ten available frequencies.

To illustrate further the changes in design, we divide 50-51 MHz by 10 with an injection-locked oscillator, directly synchronized by the high-frequency signal. A circuit design, which guards against free-running of the oscillator, causes the output amplifier to turn off automatically when an input signal has not reached the divider.

Improvements in reliability and in performance of the synthesizer-digit insertion unit combination have been attained through reduction of some spurious outputs and susceptibility to magnetic disturbances. Most significant, however, is the reduction of switching time from 2 ms to 200 μs.

You will note, from Figure 4, that the outer appearance of the RDI-1B is very similar to that of the RDI-1 and DI-1 units, indicating interchange compatability. The outward appearance, however, is all that is similar. Four circuit boards of identical dimensions greatly facilitate semi-automatic alignment and testing. Functional compartmentizing and shielding have improved serviceability and performance of the unit.

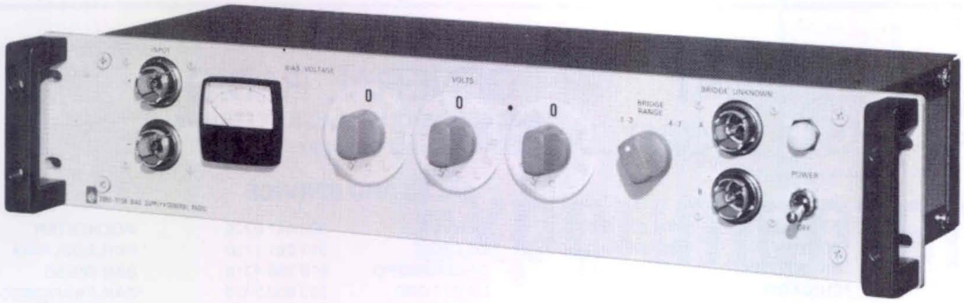
–G. H. Lohrer

G. H. Lohrer is a 1952 graduate of Technische Hochschule (Dipl Ing EE), Karlsruhe, W. Germany. His first employment was with G. Lorenz A. G.; he joined Canadian GE in 1953, Philips Electronics in 1955, National Company in 1959, working in various capacities in the VHF-UHF field. In 1961, he joined General Radio where he has specialized in synthesizer work. A senior member of IEEE, he is also a member of the Ontario Professional Engineers Association.

Complete specifications for the GR 1160 RDI-1B are available in the 1969 Supplement to General Radio Catalog T.

Catalog Number	Description	Price in USA
	1160-RDI-1B Digit Insertion Unit	
1160-9485	less filter plug	**$455.00**
1160-9480	with filter plug	**505.00**

All prices subject to quantity discount.



2995-9158 Bias Supply.

The measurement of capacitors under dc bias has become more than an occasional task for the Type 1680 Automatic Capacitance Bridge, and we have responded to this and other requirements with a companion bias supply, the 2995-9158.

This supply provides a bias voltage adjustable from 0 to 50 volts* in increments as small as 0.1 volt. Voltages can be remotely programmed or manually set, and a panel meter serves to indicate the output level. The 2995-9158 also protects the GR 1680 from large current surges by instantaneously limiting the current whenever a charged or shorted capacitor is inadvertently connected for measurement.

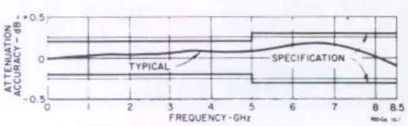
An additional feature of this bias supply is the very low series impedance it presents to the bridge test signal. This eliminates the need for corrections to the GR 1680 readings on all but the highest capacitance range. Engineering development was by R. P. Anderson, Assistant Group Leader, Impedance and Networks Group.

*Several supplies can be used in series to permit a combined bias voltage of 150 V.

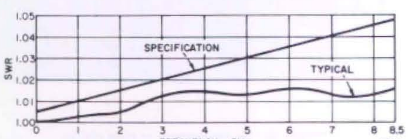
Voltage Range: 0 to 50.0 V.
Accuracy: (0.2% + 10 mV), typical.
Stability: (0.1% + 1 mV), typical for 8 hours.
Impedance: Less than 0.2 Ω from 0 to 1 kHz.
DC Current Limit: 40-mA positive, 10-mA negative.
Transient Current Limit: Less than 100 mA within 2 μs.
Added 3-Terminal Capacitance: Less than 1 pF.
Programming: 100 Ω/V.

Catalog Number	Description	Price in USA
2995-9158	**Bias Supply**	**$800.00**

All prices subject to quantity discount.



Type 900-G6, -G10 Attenuation Accuracy.



Type 900-G6, -G10 SWR.

GR 900-G6 and GR 900-G10 Precision Fixed Attenuators are equipped with GR900® connectors and have values of 6 dB and 10 dB respectively. These attenuators are much lower in SWR and have a more uniform attenuation over a wide frequency range than units previously obtainable for 50-ohm lines.

GR 900-D20 Adjustable Short Circuit is a coaxial sliding short circuit equipped with the GR900 connector, usable as a tuning and matching element or as a reactance standard.

GR 900-LK10 Precision Adjustable Line is a line stretcher with very low SWR and uniform characteristic impedance and insertion loss, equipped with GR900 connectors.

Complete specifications for the four units above are available in the 1969 Supplement to General Radio Catalog T. Engineering development was by John Zorzy, Group Leader, GR Engineering Department.

Catalog Number	Description	Price in USA
	Precision Fixed Attenuators	
0900-9850	900-G6 (6 dB)	**$185.00**
0900-9851	900-G10 (10 dB)	185.00
0900-9430	**900-D20 Adjustable Short Circuit**	160.00
0900-9570	**900-LK10 Precision Adjustable Line**	245.00

All prices are subject to quantity discount.

Type 900-G10

Type 900-D20

Type 900-LK10

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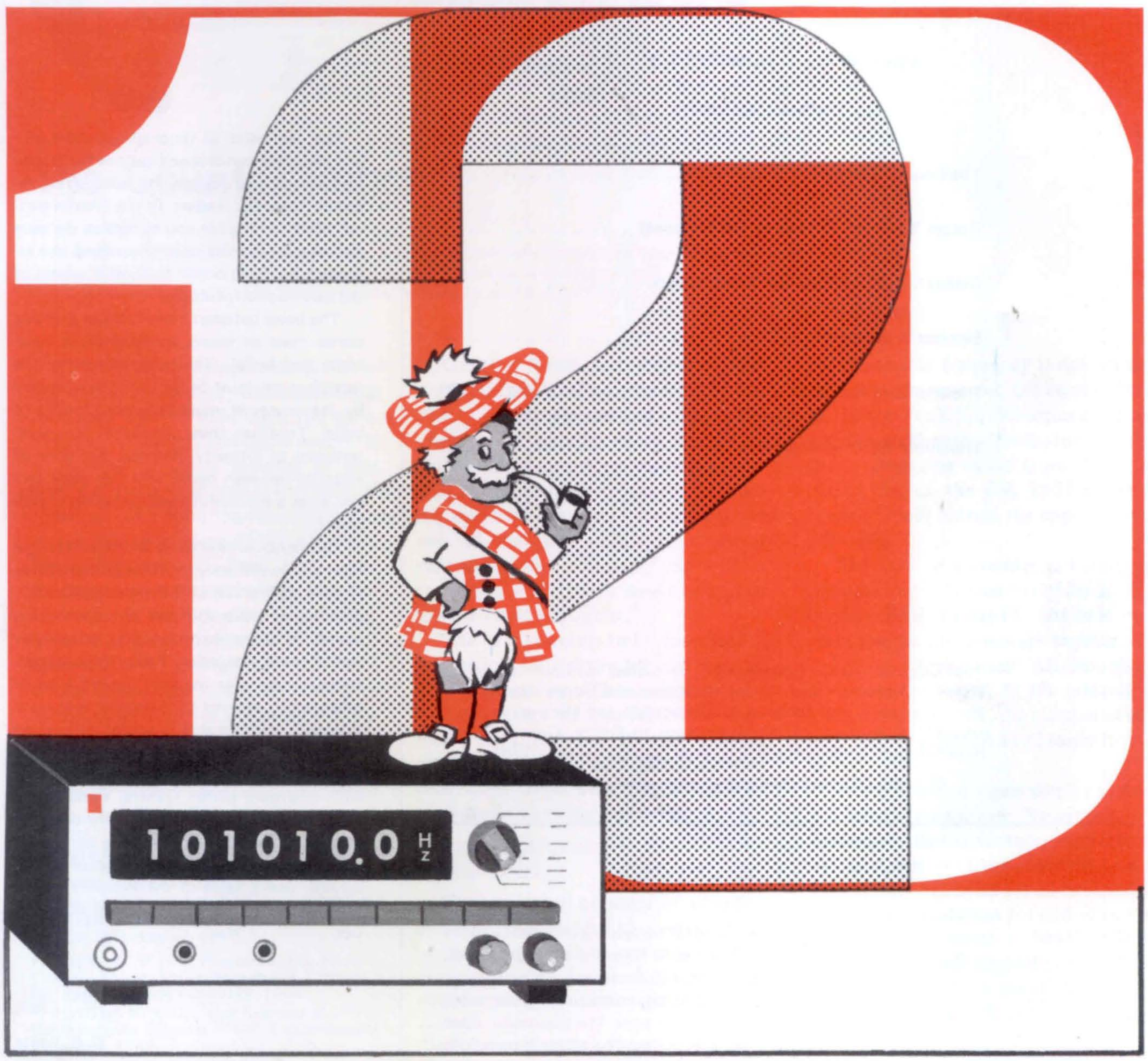


THE
GENERAL RADIO
Experimenter

VOLUME 43 NUMBERS 7, 8 JULY/AUGUST 1969

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

THE COVER — The multiplicity of numbers is the seed for computing processes and the basis for science in general. In prehistoric times, with his 10 fingers as a group control, man counted by using twigs and pebbles. The Greeks and Romans supplied graphic control with their symbolic number-figures. Modern technology confounds the centuries-old use of 10 as a number base by adopting 2 as its number base. The conversion from decimal to binary system is the subject of our cover and, in a not-too-subtle manner, our little friend draws the attention of our readers to our line of new, inexpensive counters.

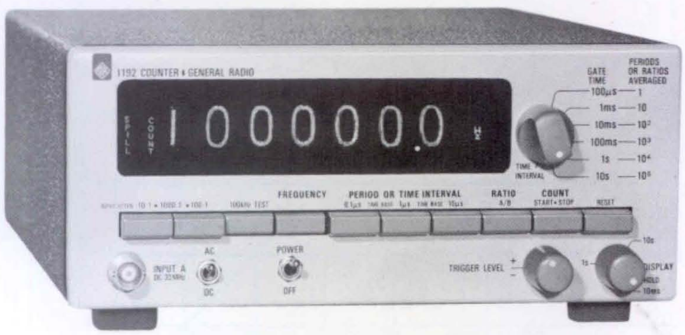
Our last editorial touched upon the subject of performance specifications for instruments and was designed to elicit some response from our readers. In the interim period, while waiting for you to receive the issue and to consider the questions raised, our attention has been drawn to another aspect of the specification problem.

The bond between manufacturer and customer must be based, in equal parts, upon *truth* and *belief*. The presentation by the manufacturer must be truthful; the reception by the customer must be with a feeling of belief. Together, these conditions denote attainment of integrity. Without that bond of integrity between maker and user there cannot exist a successful manufacturing organization.

Our personal sensitivity on this subject has been sharpened lately by a change in position. Recent assumption of the duties and responsibilities of editorship has not diluted our memories of, and experiences as, a customer rather than a supplier. From these experiences we are able to draw upon a fund of information related to customer problems. Formerly we were able to accept nothing we saw or heard in the manufacturing world without question. Our working relationship with engineers within General Radio, however, has been established upon the common ground of mutual respect and a continuing desire to give to our readers information, service, and a faith in the integrity of our engineering and in the quality of our product.

C E White

C. E. White
Editor



GR 1192 Counter

The Counter-Punch

Diminutive in size and cost, the GR 1192 Counters successfully match their bigger brothers in versatility, sensitivity, and immunity from noise. These instruments can interface with computers; they count up to 500 MHz (using the GR 1157-B Scaler) and offer precisions of 5, 6, or 7 digits.

The digital electronic counter is today a widely accepted, basic measuring tool used side by side with the oscilloscope, voltmeter, and signal source. It is also the youngest member of that family of tools inasmuch as we find first mention[1] of a counting-rate meter in a report from a meeting of the American Physical Society in April 1936. The development of the meter is an interesting study in evolutionary engineering, from its original application as an averaging device to register the output of a Geiger-Muller tube counter[2] to today's omnipresent tool for counting components and for measuring frequency and time. Performance, operability, and reliability have consistently improved while cost has steadily decreased.

General Radio has made notable contributions to this progress. The GR 1130 Counter was the first to apply the principle of parallel entry storage to prevent the intermittent read-in and display flicker of earlier counters.* Within the past two years we have introduced one of the first all-integrated-circuit general-purpose counters,[3] as well as creating a completely new concept in low-frequency counters.[4]

The new counters to be described in this article reflect the present state of device technology. These counters take full advantage of the latest developments in integrated circuits, display devices, packaging, and manufacturing procedures to produce a high degree of performance at a modest price.

Economics of Resolution and Number of Digits

The majority of users of counters apparently consider the measurement of frequency to be one of the counter's most important functions. They want a counter that operates over the maximum possible frequency range consistent with the cost and the current state of the art. The GR 1192-series of counters, designed with a recognition of this fact, covers the range from dc to 32 MHz. Use of the GR 1157-B Scaler (page 13) will extend the upper limit to 500 MHz.

The cost of a counter is highly dependent upon the number of digits in its display. Each displayed digit calls for a counting decade, a storage register, and the display device itself. All these items are relatively costly. In the interest of maximum economy, the number of digits is varied in the GR 1192 series from 5 to 7.

The number of digits displayed does *not* affect resolution. Resolution for frequency measurements, determined by the duration of the counting gate, is a maximum of 0.1 Hz for all counters of the series. Resolution for period and interval measurements is determined by the internal clock frequency (10 MHz), corresponding to a maximum resolution of 0.1 μs for each of the 1192-series instruments.

[1] Gingrich, et al, "A Direct-Reading Counting Gate Meter for Random Pulses," *Review of Scientific Literature,* December, 1936.

[2] Bousquet, A. G., "A Counting-Rate Meter for Radioactive Measurements," *GR Experimenter,* July-August, 1947.

[3] Frank, R. W., "A Programmable 20-MHz Counter-Timer Using Integrated Circuits," *GR Experimenter,* June-July, 1968.

[4] Westlake, N. L. Jr., and Bentzen, S., "The Recipromatic Counter," *GR Experimenter,* June-July, 1968.

*U. S. Patent No. 3,328,564

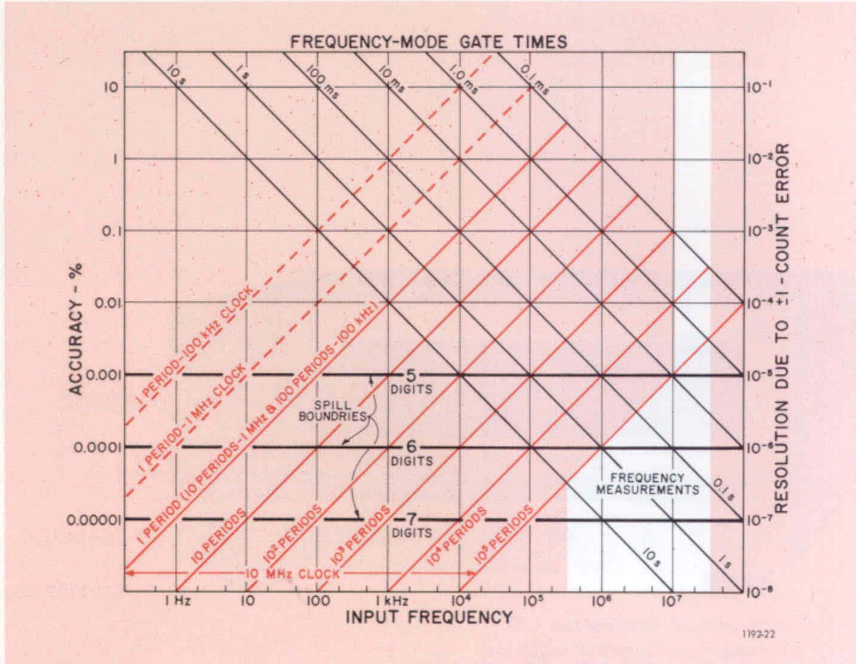


Figure 1. A summary of the GR 1192 resolution and display characteristics in its PERIOD and FREQUENCY modes. Within the white area resolution is highest in the FREQUENCY mode. The 1-ms GATE TIME prevents spill-over at the highest counting frequency; for period measurements, the 100-kHz counter clock permits up to 1-s periods to be measured without spill-over in the 5-digit counter, while the 10^5-PERIODS control permits parts-per-million resolution at an input frequency of 1 MHz.

More About Resolution

The wide range of counting-gate times (100 μs to 10 s) and period-measurement clock frequencies (0.1 μs to 10 μs) permit even the 5-digit counters to display the most significant figures of a measurement without spill-over (Figure 1). All the counters incorporate a lamp indicator to warn that, in the interest of increased accuracy, the more significant figures have been spilled-over from the register.

Since the best time resolution of a counter is established by a maximum counter-clock rate, period measurements become less accurate as the input frequency is increased. For example, in a single period, a 1-MHz signal will produce a reading of 10 counts. In order to increase the accuracy of time measurements, up to 10^5 periods of the input signal may be averaged. This time-averaging process has the side benefit of reducing the effects, in the displayed data, of noise on the input signal by approximately the amount of the averaging (20 dB per decade of averaging).

The resolution of the GR 1192 for measurement of different input signals of good waveshapes, using frequency-, period-, or multiple-period measurements, is shown in Figure 1. The figure is relatively complex; fortunately, all the complexity is in the figure because the GR 1192 has both an automatically-positioned decimal point and a display of the dimensional unit of the measurement. Note that the same resolution is obtainable whether you measure with the 5-, 6-, or 7-digit counter because the largest numbers are displayed by using sufficiently short gating times. Conversely, the smallest numbers are displayed by using relatively long gating times. Since each counter in the 1192 series has identical gating-time controls, resolution is identical from one counter model to another.

Other Characteristics

In addition to making single- or multiple-period frequency measurements, the GR 1192 performs the other basic counter functions of serial accumulation or time-interval measurements, or it can establish non-decimal time-base ratios. As with the measurement of frequencies, the wide range of clock ratios (and time-base scaling) permits full resolution while accommodating even the least-costly 5-digit display.

- In serial accumulation or simple counting, a measure is continuous as long as the operator permits the counting-gate in the instrument to remain open. In the GR 1192, control of the gate is either manual or remote by means of start and stop pulses, or by a remote signal. Often it is desired that the counter present a total count over an extended interval, rather than the repetitive short interval counts. Such requirements exist in production control, or when intervals such as pulse duration are measured. A total count can

S. Bentzen received his BSEE degree from Indiana Institute of Technology in 1962, joined General Radio the same year, and completed work for his MSEE degree from Northeastern University in 1966. Presently he is a development engineer in the GR Frequency Group. He is a member of IEEE.

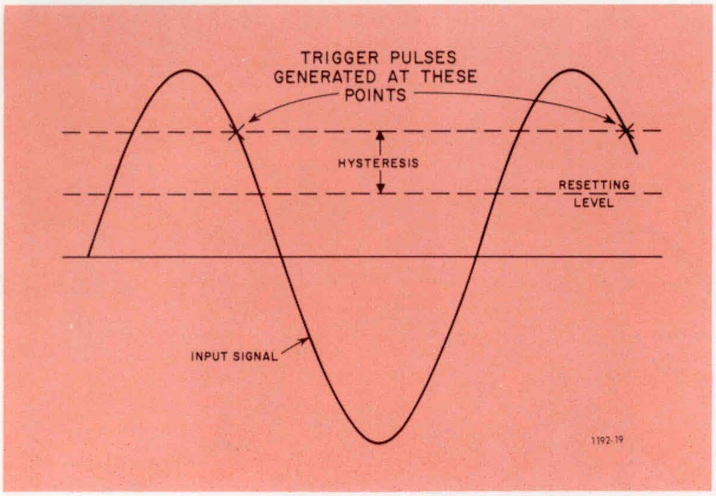


Figure 2. Illustration of hysteresis applied to trigger pulses.

be retained by the counter memory until the counter is directed to present it or to erase the memory.

- Measurement of a time interval can be defined by the duration of a single pulse. In the time-interval mode it is important that the internal clock frequency be high for good resolution, but not so high that a long time interval would cause register spill-over. The highest obtainable resolution in interval measurements is 0.1 μs in the 1192 series. In order to prevent loss of the more significant digits, the counter clock frequency can be reduced to 100 kHz, which permits the measurement of an interval as long as 100 seconds without spill-over.

- Another useful measurement mode is that of RATIO. The frequencies of two signals A and B are related: A/B. The B signal is used effectively to establish a new time base for the counter, as in these examples: A 100-Hz signal connected to the B INPUT can produce a gating time as long as 1000 seconds. Ratios of a non-decimal relationship, e.g., 60 ms, can be obtained when a stable input frequency of 16.6 kHz is fed into the B INPUT. Such a ratio is useful if you desire to display rpm. In a similar manner, it is possible to establish ratios that permit displays of flow in gpm, velocity in mph, and other parameters.

Some Design Notes

There are no panaceas in the design of counter input circuits. The thoughtful designer gives the user what he wants – high input impedance, good sensitivity, and low internal noise. It is necessary, however, that the user understand, and use intelligently, the input controls designed into the GR 1192 series.

The input circuit of the GR 1192 feeds a level detector that produces a pulse when the signal to be measured passes through a predetermined level. The level detector has a 10-mV hysteresis magnitude, referred to the input terminals, which effectively prevents

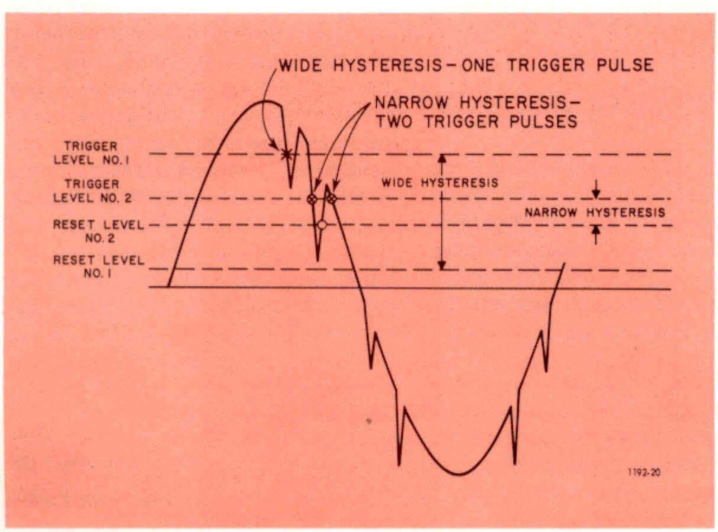


Figure 3. Example of hysteresis widening to reduce false triggering.

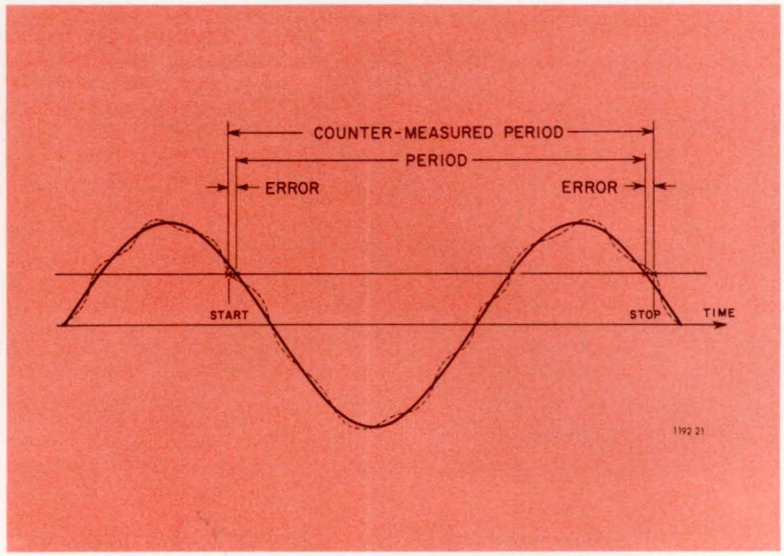


Figure 4. Effect of internal counter noise upon trigger point.

any false triggering actions by an input signal containing noise (Figure 2). Pulse triggering by noise may be prevented by expanding the hysteresis magnitude to such an extent that the hysteresis is greater than the superimposed noise (Figure 3).

Good practice indicates that the hysteresis magnitude should always be kept as wide as practical to overcome the effects of additive noise. In the GR 1192, hysteresis magnitude can be made great enough, by adjustment of the INPUT ATTENUATOR control, to overcome the triggering effects of unwanted noise levels from 10 mV to 10 V. If noise is superimposed on a waveform, the TRIGGER LEVEL control can be adjusted to move the triggering level to a point on the waveform which has less noise.

Optimum performance in the face of superimposed noise is obtained when the counter triggers on the maximum slope position of the input waveform. The TRIGGER LEVEL control, in combination with the INPUT ATTENUATOR control, permits threshold adjustment over a range of ±100 volts. Since the negative transitions are usually faster, we have chosen a negative triggering wave slope in our design.

The input impedance of a counter should be high enough so that very little loading of the input circuit takes place under operating conditions. The input impedance of the GR 1192 is 1 MΩ shunted by 27 pF. These values permit the operator to use an input probe such as the GR P6006 which has an impedance of 10 MΩ shunted by 7 pF.

In a good counter, the internal input noise should be very low and usually will be insignificant compared to the noise that has already been imposed upon the input-signal waveform. Internal noise adds to the signal and will cause the triggering point to vary in time (Figure 4).

Error in a period measurement due to noise impressed upon a signal can be expressed as:

$$\epsilon = \frac{N}{\pi S n} \times 100\%,$$

in which N is the noise level and S the signal level in the same units of measurement, and n is the number of periods averaged. If we assume a signal-to-noise ratio of 40 dB, the error in a single-period measurement is calculated to be: $\epsilon = 0.318\%$. This value is well within the resolution of the counter.

When the input signal is very, very clean, the limit of measurement accuracy is established by the noise in the input circuitry of the counter. The effective input noise of the 1192 counter typically is of the order of 50 to 100 μV. Thus the external triggering error in microseconds is less than $\frac{0.0002}{\text{signal slope in } \mu s}$. Interested readers can refer to the December 1962 issue of the *GR Experimenter* for a general article discussing error sources encountered in counter measurements, and to the February 1966 issue for an article specifically discussing noise-produced errors.[5,6]

The accuracy of any counter is almost completely dependent upon the accuracy and stability of its time-base oscillator. Economical compromises involved consideration of low-cost ovens or crystals that operate over the instrument ambient-temperature range. The GR 1192 employs a very stable 5-MHz crystal, operating in the ambient-temperature range of the counter, for internal-frequency control. Its temperature coefficient is less than 3×10^{-7} in the range 0 to 55°C. Total frequency shift due to temperature is less than 4×10^{-6} while long-term drift is less than 2×10^{-6} per month. If higher stability is desired in the counter, it can be locked to an external frequency standard, with the attendant gain in accuracy and stability.

[5] McAleer, H. T., "Digits Can Lie," *GR Experimenter*, December 1962.

[6] Frank, R. W., "Input Noise," *GR Experimenter*, February, 1966.

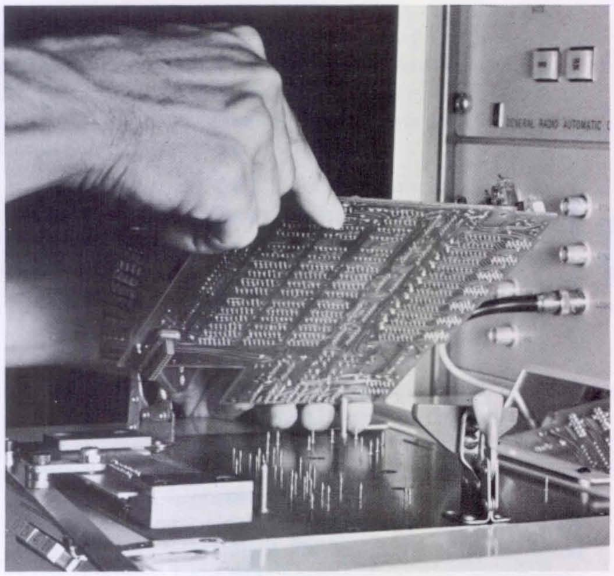


Figure 5. Automatic testing system for the GR 1192 etched-board subassemblies.

The physical size of counters has been decreasing steadily and this GR unit is no exception. In height, the progress has been from 16 inches to 3-1/2 inches, with a corresponding shrinkage to one-half rack width.

A Word About Production Control

The GR 1192 incorporates numerous computer-type integrated circuits mounted upon three etched-circuit boards. In order to test each circuit board, GR constructed an automatic computer-controlled test assembly that receives each type board in an individual test jig (Figure 5). A total of 90 points is tested on one board. The associated computer is programmed to perform more than 300 independent tests after it has first determined that the circuit has been conditioned for testing. The test assembly identifies failures and reports the failures in a printed record, using an associated teletypewriter. Upon completion of the automatic test program, the boards, okayed by the computer, are removed by the operator and installed in the instrument for a final over-all functional check.

–S. Bentzen

ACKNOWLEDGMENT

The author gratefully acknowledges the assistance of B. Sargent in the electrical design of this instrument and of J. McCullough for production engineering.

Complete specifications for the GR 1192 are available on the catalog page, included as a tear sheet inside the back cover of this issue, removable for insertion in GR Catalog T.

Recent Technical Articles by GR Personnel

"On Estimating Noisiness of Aircraft Sounds," R. W. Young and A. Peterson, *Journal of the Acoustical Society of America*, April, 1969.*

"Spectrum Analysis of Stationary Noise Signals," W. R. Kundert and A. P. G. Peterson, *Sound and Vibration*, June 1969.*

"High-Gain Phase Detector Circuit Uses No Transformers," C. C. Evans, *Electronic Design*, May 24, 1969.

"Microwave Tuners," T. E. MacKenzie, *Electronic Instrument Digest*, April 1969.

*Reprints available from General Radio.

A major instrument we employ to promote service to our readers is, of course, the *Experimenter*. Do not overlook, however, the existence of the supplementary handbooks, application notes, and reprints of technical articles, such as those above, available to our readers. A listing of these publications will be mailed to you upon request to the Editor. Other questions related to GR instruments or engineering are also welcomed.

Calibration of High-Voltage Transformers

This article, by an Australian reader, is presented as a calibration note of general interest to readers in the high-voltage and power fields.

Standard voltage transformers for comparison with the unknown transformer at voltages above 110 kV are expensive and rarely available. An absolute method must then be used.

The ratio and phase angle of the transformer under test are compared, at rated voltage, with those of a stable and loss-free capacitance divider of negligible voltage coefficient of capacitance. The ratio of the divider is then measured at low voltage. It is generally held that the ratio should be determined as a ratio, that is, derived by a transfer process and not by the measurement of individual quantities. This applies when you are striving for the limit of attainable accuracy while using a special divider with gas-dielectric three-terminal capacitors in both arms. For industrial purposes an uncertainty of measurement of 0.03% in ratio and 1 minute in phase angle is sufficient.

It is convenient to measure the two capacitances with the GR 1620-A direct-reading, 0.01% capacitance-measuring assembly. The ratio is obtained from

$$k = \frac{|V_P|}{|V_S|} = \frac{C_2}{C_1}\left(1 + \frac{\phi^2}{2}\right)$$

wherein ϕ is the phase angle of the transformer in radians and C_1, C_2, V_P, and V_S are as given in Figure 1. The expression for ratio differs from C_2/C_1 by less than 2 parts in 10^5 for phase angles smaller than 20 minutes.

The transformer under test is connected in series aiding, and the ratio balance is obtained by adjusting C_2. The capacitor C_1 is a three-terminal compressed-gas type of suitable voltage rating. Commercially available models have capacitance values of 50 or 100 pF. Reputable makes with clean, conditioned, and well-centered cylindrical electrodes have a loss angle of less than 10^{-5} radian and insignificant voltage- and frequency-coefficients of capacitance. It is therefore sufficient and convenient to measure C_1 at 1000 Hz.

If C_1 has values of 50 or 100 pF, the capacitor C_2 may have values to 0.5 microfarad, for ratios between 1000 and 5000. The dielectric of C_2 is polystyrene for the decade steps and air for the variable part. In tests carried out so far on three-phase transformers rated up to 330 kV/110 V, 3 phase, the capacitor C_2 was home built; a GR 1412-BC is an excelent commercial substitute. The losses of polystrene-dielectric capacitors are very low; for calibration purposes it is sufficient to use a constant correction of +0.3 minute. Tests have shown that the voltage coefficient up to 100 volts is less than 2 parts in 10^5 and thus negligible for the purpose. The capacitance of C_2 was measured at 2½ times the power frequency, viz 125 Hz, to avoid errors due to stray fields at the

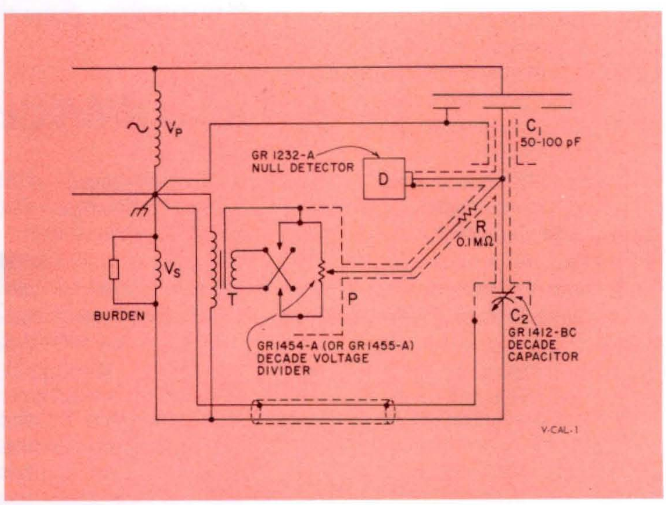


Figure 1. Schematic diagram of high-voltage transformer calibration circuit.

low voltage level employed in the GR 1620-A assembly. For polystyrene capacitors, the frequency error of capacitance is negligible between 50 and 125 Hz. Errors due to temperature changes were minimized by measuring C_2 immediately after the calibration of the unknown transformer and without shifting its physical position.

The battery-operated GR 1232-A tuned amplifier is a suitable null indicator (D in Figure 1) and presents no problems with double earthing, as would be the case with a mains-operated device.

The phase angle is adjusted by means of P and is obtained from

$$\phi = \left[\frac{1}{\omega C_2(R+r)} \propto \frac{3438}{10}\right] + 0.3 \text{ min},$$

wherein R is the measured resistance of the nominal 100,000-ohm 1-W high-stability, carbon current-injection resistor, r is the output resistance of the GR 1454-A 10,000-ohm voltage divider* and α is the dial reading of the 1454-A divider.

The permalloy-core stepdown transformer T has a ratio of 10/1; its primary impedance exceeds 100,000 ohms and thus presents a negligible burden to the transformer under test; its resistances referred to the secondary are less than 1 ohm.

If the transformer to be calibrated is designed for a small burden, the loading due to C_2 may cause an error which is not negligible. To correct for this, the ratio and phase angle are first measured in the normal way (k_1, PA_1). The measurements are repeated (k_2, PA_2) with an auxiliary capacitor of approximately the same value as C_2 connected across the burden. This will make the ratio smaller and the phase angle more negative. The corrected results are then:

$$k = k_1 \left(1 + \frac{k_1 - k_2}{k_1}\right); \quad PA = PA_1 + (PA_1 - PA_2).$$

Direct-reading measurement was not sought because calibrations are done infrequently, and generally only on the first transformer of a production run. This unit serves as a standard for the remainder.

While all work was carried out at 50 Hz there is no question that the arrangement is suitable for the calibration of voltage transformers at 60 Hz.

* Type 1455-A replaces Type 1454-A.

L. Medina is an EE graduate of the Technical University of Vienna, 1932, with an ME awarded him in 1959. He recently retired from the High Voltage Laboratory, which he helped establish, of the Commonwealth Scientific and Industrial Research Organization, Sydney, Australia. Presently he is teaching part time at the University of New South Wales and at Sydney University, while acting as consultant in high-voltage engineering.

Seminar Scheduling

A series of lecture-seminars on the subject of manufacturing/plant noise will be presented by Laymon N. Miller of Bolt, Beranek, and Newman. The series will be conducted in six cities on the dates tentatively scheduled below.

Seattle, Washington	October 3, 1969
Denver, Colorado	October 22, 1969
Houston, Texas	November 14, 1969
Atlanta, Georgia	November 22, 1969
Charlotte, North Carolina	December 3, 1969
Washington, D. C.	December 10, 1969

The lectures will deal with practical acoustics and noise control. Readers interested in attending the lectures can contact the GR District Office closest to the locations listed above for information related to exact location and fee.

Reports from the Field

MATING THE GR 1680-A WITH THE IBM 1800

Another successful marriage of the GR 1680-A Automatic Capacitance Bridge Assembly to the digital computer is illustrated by an example supplied us by Tom King of the IBM Data Acquisition and Control System (DACS) Center in San Jose, California. The center has successfully interfaced the IBM 1800 computer, using the Multiprogramming Executive (MPX) Operating System, with several asynchronous test stations for demonstration purposes. The computer has been able to control, at one time, such diverse manufacturing devices as the GR 1680-A, a relay test station, a carburetor test station, a distributor test station, and a dynamometer. As demonstrated at the 23rd Annual Conference of the Instrument Society of America in New York, the tests were run asynchronously, with output reports recorded on the IBM 1443 printer, IBM 1053 and 1816 keyboard printers, and on the Tektronix 611 storage oscilloscope. The same bridge program now is being run at IBM DACS.

The 1680-A assembly operates in conjunction with an Atescar mechanical handler (Figure 1). Test pallets, which contain inductors, capacitors, or resistors in lots of 25 or 50 units, are subject to test specifications recorded in an IBM 1810 disk file. When the test process is started, the 1800 program calls for a record of the pallet and lot numbers, retrieves the related test specification data from the memory disk, and instructs the handler to position the components sequentially for test by the GR 1680-A. Upon completion of a measurement, the data are fed back to the IBM 1800 program for comparison and temporary storage prior to calculations.

Upon completion of tests of 25 or 50 positions, the IBM 1800 executes calculation of the output programs, meanwhile preparing for another test run by the GR 1680-A. Output reports and scope displays are generated from the variable-core storage area, making it possible to overlap tests by the GR 1680-A with reports of previous runs.

Typical data that can be economically stored in disk files are test specifications, life-test data, and other historical information. Required reports can be generated later from this stored information.

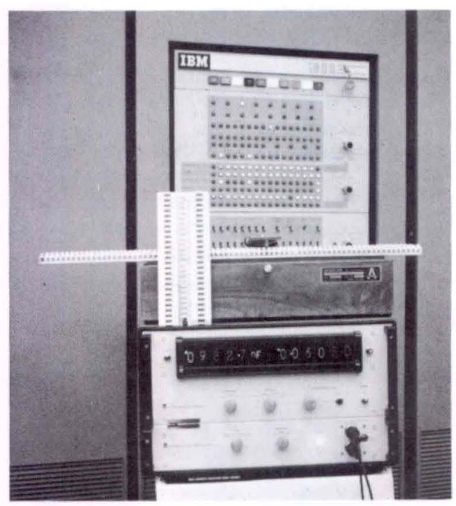


Figure 1. Control of GR 1680-A bridge assembly by IBM 1800 computer.

Photo courtesy of IBM.

PROGRAMMING TESTS OF COMMUNICATION CABLES

We have had a report that the GR 1680-A assembly is involved in tests of communication cables.[1] The requirements here are measurements of mutual capacitance between the two wires of a pair, of unbalanced capacitances between conductors of one pair and those of another pair, and of unbalance of a pair to ground. Cables may range in size to 100 pairs; two-pair measurement combinations reach 4950. Previously, to carry out a manual check of pair-to-pair unbalance was impractical; measurements were made only on pairs known to be physically close.

[1] For complete details refer to "An Automatic Computer-Controlled System for the Measurement of Cable Capacitance," R. G. Fulks and J. Lamont, *IEEE Transactions on Instrumentation and Measurement*, December 1968.

The automated system has the ability to alter its measurement sequence. The computer quickly determines whether the first direct capacitance of a combination is less than a programmed value. This result indicates the pairs are physically separated and more precise measuring is eliminated. At a rate of four measurements per second, the over-all automatic measurement time is reduced to 15% of the manual time by incorporating overlapping operations into the sequence of operations. The bridge can balance itself, the computer can calculate capacitance, and the typewriter can type the results – almost simultaneously!

Measurements are normalized to nanofarads per mile and typed out in a table showing the distribution of measurements. Average and standard deviations are calculated also. Test tolerances are determined automatically from two values typed in by the operator – number of pairs to be tested and cable length.

The assembly performing this task is shown in Figure 2. It includes the GR 1680-A assembly, PDP-8/S digital computer, scanner, teletypewriter, and an interface unit to perform signal-translation and data-storage functions. The assembly is self-calibrating; a test employing a standard capacitor is performed before and after each test sequence, to verify correct operation of the bridge and scanner assembly. Cable-measured values outside programmed tolerances trigger an alram to alert the operator to check for the cause of trouble.

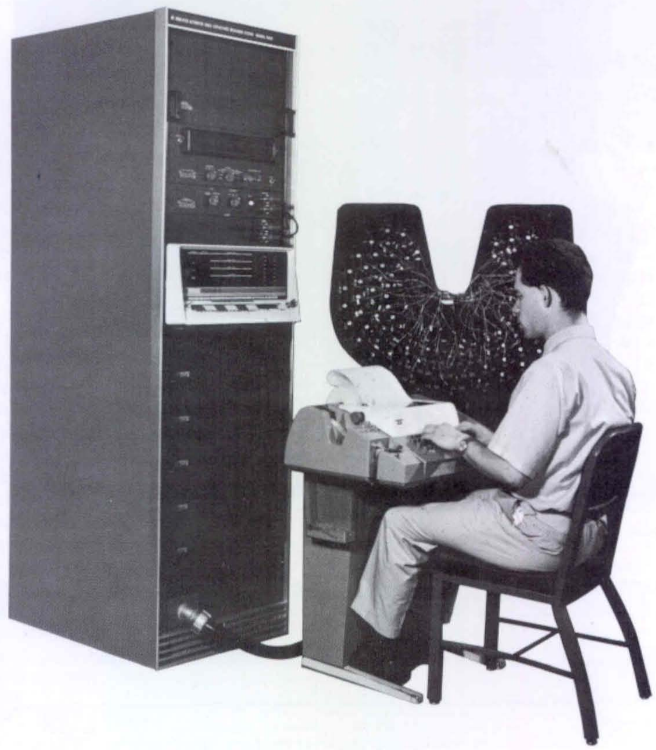


Figure 2. Automatic test assembly for communication cable inspection.

DESIGNING FIRE-RESISTANT NAVY CABLE

From time to time we at GR are made aware of the extreme environmental conditions to which "ordinary" electrical products are subjected and of the part played by our instrumentation during extraordinary inspections. Recently we learned of the rigorous development tests of Navy cabling used on board US Navy ships for transmission of audio and data information. These cables must perform, without failure, during exposure to shipboard fire. Failure could result in severe compartment damage, loss of personnel, and even destruction of the ship because of the loss of an instruction or command.

The Naval Applied Science Laboratory, Brooklyn, N.Y. has responsibility for design of fire-resistant, interior-communications, ship-board cabling. Part of the task of proving a successful design is exposure of cable samples to flame while electrical measurements of cable capacitance and conductance are made. An assembly that performs such measurements (Figure 3) includes the GR 1680-A Automatic Capacitance Bridge Assembly plus the GR 1521-B Graphic Level Recorder and GR 1136-A Digital-to-Analog Converter. Engineers know that rapid fluctuation of the capacitance of twisted-pair communication cables, during exposure to simulated-fire conditions, is difficult to monitor manually. Use of the automatic measuring assembly simplifies the measurement task considerably. Mr. M. DeLucia, Project Engineer at the Naval laboratory, notes that a test response, similar to that shown in Figure 4, is typical of improved-cable response to exposure to extreme environment, represented by the gas flame in Figure 3.

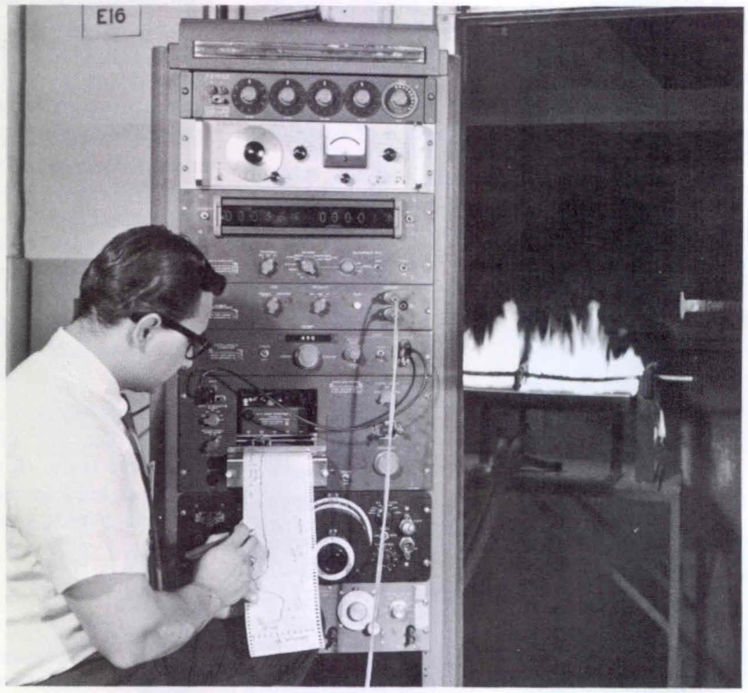
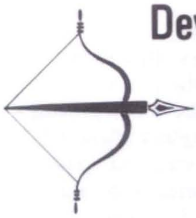


Figure 3. Test assembly for fire-resistant cables.

Photo courtesy of US Navy.

Figure 4. Graph of cable-characteristic response to extreme environmental change.

Art courtesy of US Navy.



Deviations from Accuracy

Our apologies to "PK" McElroy for moving his home location, as included in the article in the *Experimenter* for May/June, page 21. In reality, he still lives at 58 Douglas Road, Belmont, Mass., not at Gayles Road as printed.

Also, we were quite surprised to find that our paragraph which mentioned the honor to Dr. A. P. G. Peterson, same issue and page, could be construed as alluding to APGP as a "foundling" of the Audio Engineering Society. In spite of the Editor's story, Dr. Peterson was *not* abandoned as a child. It was the twentieth anniversary of the *founding* of the Audio Engineering Society that was the occasion for the award to Dr. Peterson. Finally, investigation has proven that Dr. Peterson did not, as reported, publish his first *Experimenter* article in August, 1941. In the October 1937 issue, before joining GR, he printed an article "An Ultra-High Frequency Oscillator" which reported the results of a project jointly sponsored by MIT and GR.

Combination of GR 1157 Scaler and GR 1192 Counter assembled for rack mounting.

NEW

The GR 1157-B Scaler, compatible with the GR 1191- and 1192-series Counters, replaces the popular GR 1157-A Scaler. This model includes improvements such as:

- Greater counter resolution at frequencies that only require prescaling by 10, by the addition of a divide-by-10 function to the divide-by-100 function.
- Smaller size, that permits side-by-side combination with the GR 1192 Counter.

The new scaler-counter combination forms a single, standard-rack-width package (GR 1192-Z) that provides an economical way to obtain high-frequency counting up to 500 MHz (Figure 1).

The signal applied to the 1157-B INPUT terminal is fed to an attenuator that reduces its amplitude by a factor of 1 (no reduction), 2, 5, or 10, so the instrument can handle signals from 0.1 to 7.0 volts rms amplitude. In order to indicate the proper attenuator (SENSITIVITY) control setting, the output of the attenuator is rectified, amplified, and applied to a "green-sector" meter. Generous overlaps in range are provided so that attenuator settings may easily be made to bring the meter reading into the green sector. In order to give a clear indication of the range of proper operation, the upper two-thirds of the meter scale is electronically compressed by the meter amplifier. Thus an actual 7:1 signal range represented by the green sector is compressed into a 3:1 deflection ratio.

The attenuated input signal is amplified and applied to a pulse generator circuit that provides a pulse of 0.5-ns duration, as required by the tunnel-diode binary divider.[1] The binary output, now at one-half the input frequency, is fed to a scale-of-five divider using three integrated-circuit flip-flops in a ring-type configuration. The output of this quinary-divider circuit, now at one-tenth the input frequency, is applied to another integrated-circuit flip-flop for frequency division by two. This signal is finally applied to a second scale-of-five divider, similar to the first, to obtain a continuous end output at one-hundredth the input frequency at the 100:1 SYNC OUTPUT. In addition, a main output is provided that can be switched from the divided-by-100 output to a divided-by-10 output taken from the high-frequency (first) quinary-circuit output.

When the GR 1157-B is used as a prescaler for a counter, the divide-by-10 feature of the prescaler is an advantage since it provides ten times greater frequency resolution in conventional frequency measurements than that provided by the divide-by-100 feature. When the counter's maximum frequency-handling capability is exceeded by the input frequency-divided-by-10 signal, the scaler must be operated in the divide-by-100 mode. Engineering of this product was by J. K. Skilling and B. J. Sargent.

[1] Similar circuitry is described in the GR *Experimenter*, page 13, October 1968 and in *NASA Technical Note D-1337*, "A Tunnel-Diode Counter for Satellite Applications," by E. G. Bush, June 1962.

Complete specifications for the GR 1157-B are available on the catalog page, included as a tear sheet inside the back cover of this issue, removable for insertion in GR Catalog T.

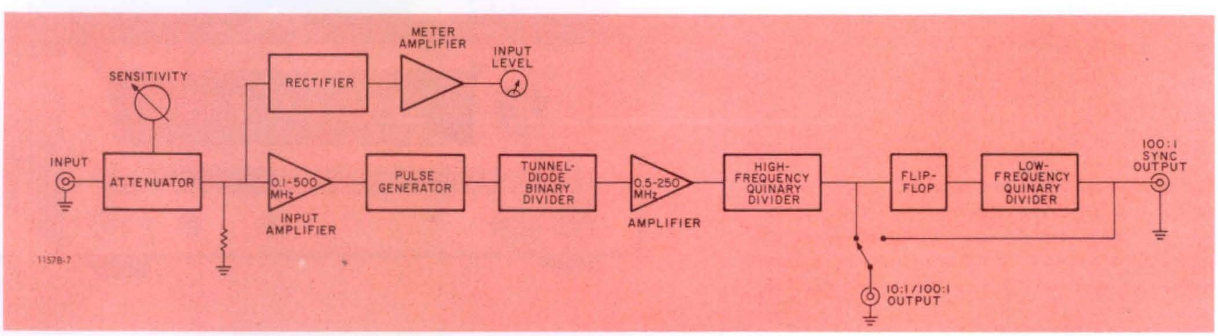


Figure 1. Block diagram of the GR 1157-B.

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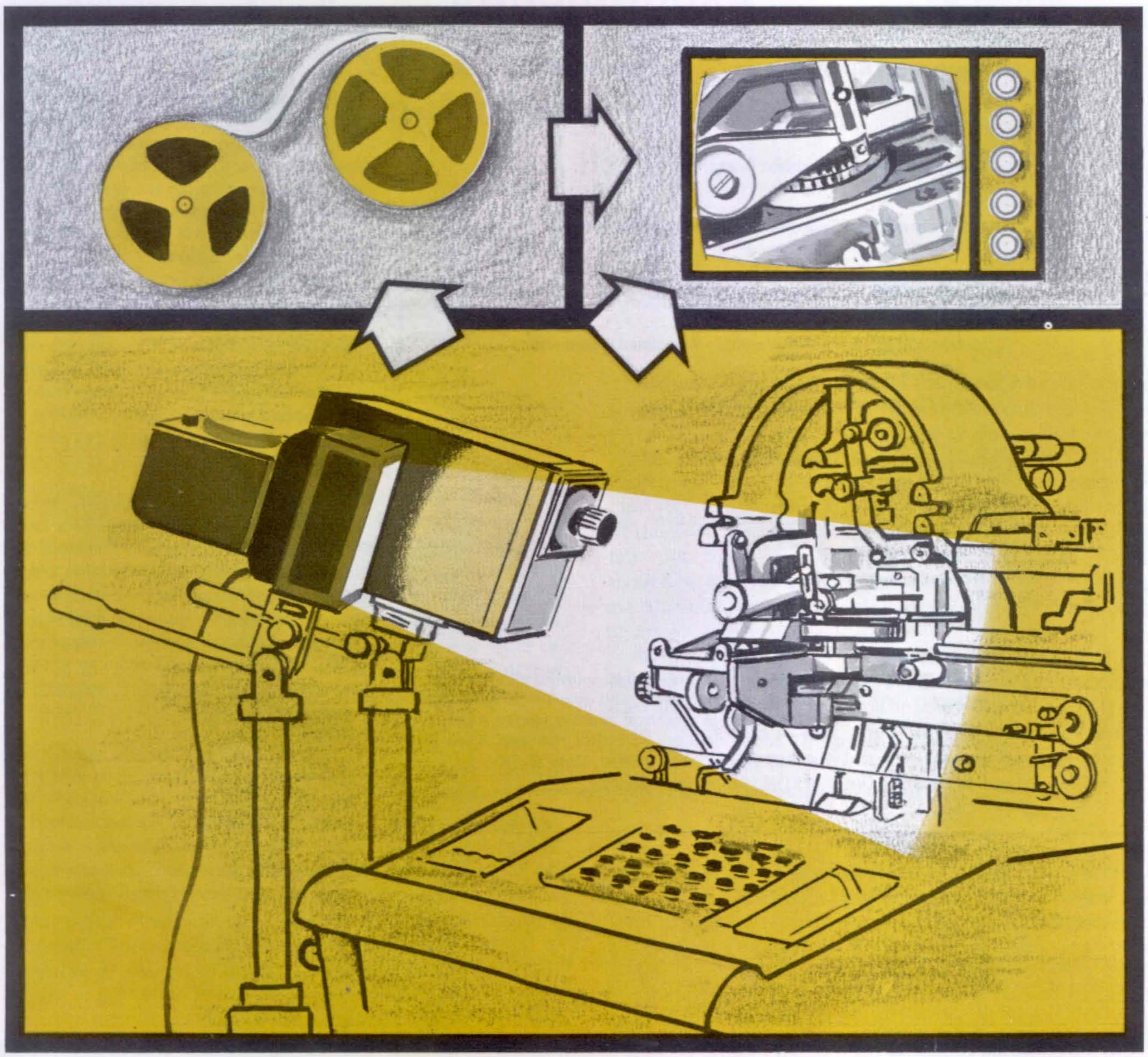

THE
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VOLUME 43 NUMBERS 9, 10 SEPTEMBER/OCTOBER 1969

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

The Cover – Stroboscopic lighting and photography are wonderfully effective tools in the field of mechanical motion analysis. Too often, however, this application is obscured by illustrations of the use of strobe techniques in studies of the human figure in motion. The article in this issue, describing the new GR Strobolume, should convince our readers that this simple but efficacious tool is equally suited to studies of *mechanical* and human motion.

It was my pleasure to spend the last two weeks of August in Ottawa as Scientific Editor for Commission I, International Scientific Radio Union (URSI), during the XVIth General Assembly of URSI. The assembly meets to review the developments that have taken place internationally during the past three years in the many fields of radio science and to explore the areas of research and development urgently required during the ensuing three-year period.

There is, however, a more fundamental purpose behind the meetings. The assembly presents an opportunity for representatives of the national academies and scientific bodies not only to exchange ideas in the technical fields and to provide for arrangements to coordinate experiments but also to meet as persons. The latter perhaps is the one most important factor in the success of these meetings for it removes, for a short time at least, the political and geographical boundaries that separate nations.

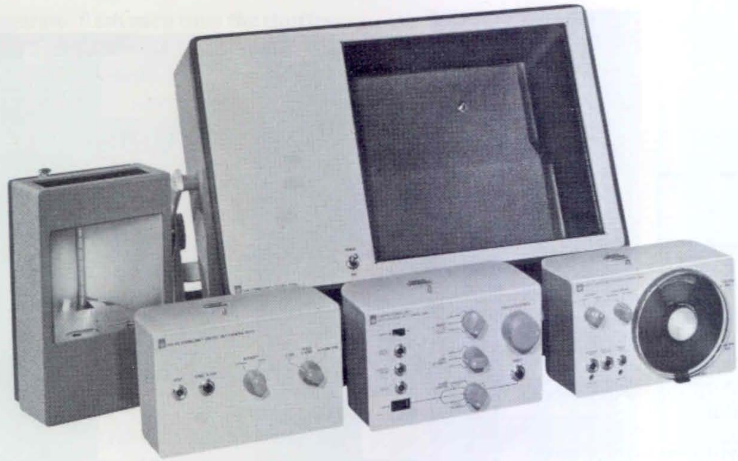
From the international pool of technical information that flowed from the numerous technical reporters, the rising influence of the computer as a vital part of the measurement system was quite apparent. This influence became even more apparent during sessions that involved other commissions, who look to Commission I for help in linking instruments, people, and computers into a coordinated measurement unit.

Unquestionably, the complexity of our ever-expanding technology requires a more rapid adaptation to change. Therefore the interests and concern of Commission I, and of all other activities concerned with measurements and standards, must turn inevitably to technological innovation -- to the coupling of man, instrument, and the computer to reach the goal of purposeful measurement research, from which come new and improved products and services.

The three-year period to the next General Assembly of URSI in 1972 will be an exciting time during which the data monster -- the computer -- will be tamed and managed to perform more meaningful and accurate measurements rather than to deluge man with reams of measurement data of questionable value.

C E White

C. E. White
Editor



GR 1540 Strobolume with lamp head and three available control units.

Detailed Viewing in Ambient Brightness

The number of applications for the stroboscope has increased dramatically in the last few years, severely taxing the performance and features of existing equipment. One of the most important needs has been greater light availability from smaller packages, for operator convenience. The GR 1540 Strobolume electronic stroboscope was developed to meet this need. The GR 1540, now the brightest commercially available, general-purpose stroboscope, combines the features of modular construction with approximately 20 times the light output per flash when compared with previous stroboscopes such as the GR 1531 Strobotac® electronic stroboscope.

A NEW MOTION ANALYSIS TECHNIQUE

The combination of a stroboscopic light and relatively inexpensive closed circuit television equipment is an exciting and powerful new analysis tool that combines aspects of both visual and photographic viewing. With currently available tv equipment, there are no strobe interconnection or synchronization problems. These tv systems employ vidicon camera tubes. The vidicon "sees" and stores the pulsed image in a manner analogous to the human eye. The image is converted to a video signal (electronic) by a scanning beam that can be either played back immediately on a monitor or stored on magnetic tape by a video tape recorder. Suitable video systems (camera, VTR, monitor) today are available for as little as $1,500. The vidicon is quite sensitive and performs well under varied light conditions.

With tv the operator can be at a location remote from the subject, camera, and strobe, which is of value in hazardous experiments such as observing rain erosion on the leading edges of rapidly rotating helicoptor blades. The low cost of video tape and its long playing time, typically 30 to 60 minutes, make it practical to record entire tests such as monitoring development progress in a vibration-reduction program. These recordings may be rerun and reexamined as often as desired long after the test equipment has been torn down or the tested device has been modified. Quite obviously, such records may contain much more useful information than written notes. An additional feature of video taping is the availability of one or two audio channels for data logging.

Because the vidicon can store the pulsed images between flashes, this system makes it practical to extend stroboscopic visual analysis to machines moving so slowly that the eye would otherwise be subject to severe "flicker effect."

Perhaps the most exciting aspect of video storage is the "instant" and "single frame-by-frame" playback capability of the video tape that, unlike movie film, does not have to be sent out to be processed. This gives the operator such flexibility that the system can be used for troubleshooting machine malfunctions in real time or reducing the setup time of complicated machines.

With a slightly more sophisticated system, such as two cameras and two strobe lights, a split-screen technique can be used to view widely separated machine functions or operations with respect to each other, in exact time synchronism! Film becomes useless when the records have no further historical value; the tape may be erased and re-used.

VISUAL MOTION ANALYSIS

The new GR 1540 Strobolume electronic stroboscope is ideally suited as a supplement to visual analysis of almost any form of motion, repetitive or non-repetitive. It can be used as a continuously adjustable flashing source over a range from 30 to 25,000 flashes per minute or for no-contact speed measurements to 250,000 rpm. With a manual oscillator adjustment, the flashing range can be set to be at or near synchronism with a cyclically moving object, to give a visual image of stopped or slow motion.

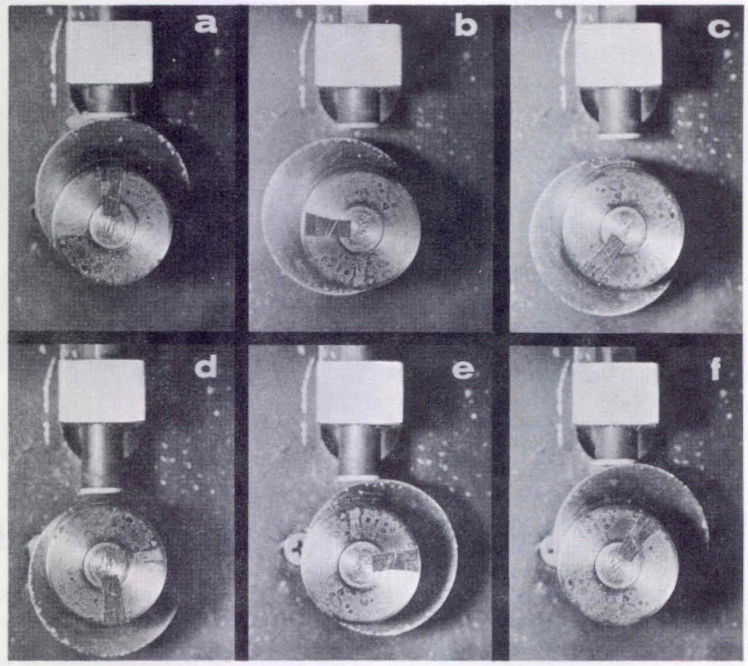


Figure 1. Motion analysis with a still camera. Picture sequence is a permanent record of visual observations made to detect misbehavior of follower in contact with cam rotating at 3000 rpm. The technique combines synchronized flash plus continuous advancement of the time-delay control, to produce a high speed "still movie."

A second method of motion analysis employs a transducer, such as a photoelectric pickoff or a switch contacting mechanism, to sense the position of the object and to "trip" the Strobolume flash. A particularly useful feature is that, by the flip of a switch, the photocell can be made to respond to either light marks on a dark background or dark marks on a light background, or to contact opening or closure. The strobe flash rate follows any deviations in the speed of the moving object to produce a stationary visual image, thus eliminating the need for manual tracking of the oscillator dial. In addition, any point in the motion of the object can be examined if each flash occurs some time later than the corresponding synchronizing signal.

In the GR 1540, this time delay can be manually controlled by a continuously adjustable time-delay circuit. The same knob control used for varying the flashing rate also provides this adjustment over a range of from 100 microseconds to 1 second. The longer a flash is delayed, with respect to the synchronizer's signal, the farther along in its cycle the object will move before being illuminated. The visual image can thus be adjusted through the entire cycle of motion, if desired, and the operator is freed from manually tracking the oscillator to follow any speed variations.

A practical example (Figure 1) shows a cam and its misbehaving follower. Assume that the cam is rotating counter clockwise at 1800 rpm (30 rps or 0.033 second per revolution). With no time delay, the image will appear, as in Figure 1a, at the instant a photoelectric pickoff has sensed the light reflection from a piece of reflective tape mounted on the cam hub. Figure 1b shows what the stopped image would look like when 8.3 milliseconds (1/4 revolution) of time delay have been introduced. As more and more time delay is added, Figures 1c through 1f, the cam can be seen in all phases of its motion. Thus, valuable phase information can be obtained about a moving object's behavior.

PHOTO-INSTRUMENTATION

In many studies, recording requirements can be met only using film as the storage or reference medium. The preceding technique can be used to take high-speed single-flash photographs with conventional cameras. With almost any conventional still camera connected to the X-sync-contact input on the Strobolume, plus the adjustable delay, photos can be taken of a moving subject at a specific point in its cycle. One can, in effect, create a high-speed "still movie" record of fast repetitive events, similar to that shown in Figure 1.

The GR 1540 also has a built-in provision for keying the oscillator manually or with camera contacts to take multi-flash or "flash-burst" photographs of relatively slow events.

Moving-film recording in high-speed photo-instrumentation has chiefly involved the use of high-speed movie cameras such as the Hycam*, with framing rates in the hundreds- or thousands-per-second range. Although the cameras are equipped with internal shutters to pulse the image to the film, a strobe light source such as the GR 1540 is often required to produce images of greater clarity and freedom from distortion. With strobe light as the high-speed shutter, the subject is viewed for microseconds per frame rather than milliseconds. Many high-speed cameras have the

*Red Lake Labs, Inc., Kifer Industrial Park, Santa Clara, California 95051.

necessary sync pulse to trigger the flash each time the shutter is fully open.

A high-speed camera is usually operated at framing rates several times the subject cyclic-motion rate, to obtain many pictures within a single cycle. It is possible, in the case of repetitive motions, to use a stroboscope and a conventional movie camera fitted with an appropriate synchronizing device to take movies at effective framing rates of hundreds of thousands of frames per second. In this case, a sampling technique is employed, similar in concept to the principle employed by a sampling oscilloscope, and each resulting image is of a different cycle. Figure 2a is an illustration of one of a number of techniques (outlined in the *Handbook of High-Speed Photography*) for using the stroboscope (whose "light is the high speed shutter") with a conventional movie camera, while Figure 2b is representative of the results obtainable. The Cinema-Beaulieu* camera in this example was chosen because it uniquely combines the features required of an instrumentation-grade camera, including provision for the addition of the required synchronizer for strobe use.

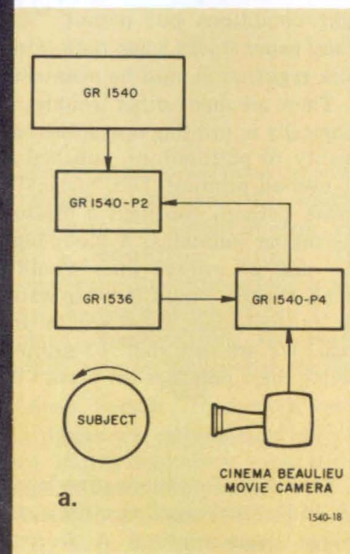
MONITORING QUALITY AT HIGH SPEED

In the graphic arts industry, stroboscopes play a vital role in maintaining printing quality. Today's high-speed presses print at the rate of 1000 feet per minute and faster. At these speeds, it is very important that any printing degradation, such as misregister or misalignment, be caught and corrected as soon as possible in order to minimize paper spoilage. This is particularly true when material such as computer business forms, labels, and recorder chart paper is printed on continuous paper strip. Tearing samples from a roll often is out of the question – even locating the fault is difficult because there is no way of determining how far into the roll the trouble may have gone. But, through the use of stroboscopy, printers are cutting operating costs drastically.

The printing is monitored with a stroboscope usually synchronized to the press, although photoelectric detection directly from the printed material is often practical with the GR 1540. The pressman views the stroboscopically stopped image. At the first sign of a misregister he inserts a paper "flag" into the printed roll where the fault begins, makes his press adjustments and inserts a second flag into the roll at the point where correct printing begins. Later, it is a simple matter to cut out the faults and to splice the rolls.

The modular construction of the GR 1540 makes it well suited to use on presses where the lamp head, operator controls, and power supply must be separated. Paper stretch and shrinkage during continuous printing also can be monitored with a stroboscope. A scale is permanently secured near the edge of the web so that reference can be made to some periodic printed mark on the paper. A reading of the scale is made with a strobe during press make-ready operations. The location of the printed mark with respect to the scale is then a fixed reference for proper operating conditions. This reference is viewed stroboscopically during the print run. If any deviation occurs, the press man knows immediately by how

*Cinema-Beaulieu, Inc., 155 West 68th Street, New York, N. Y., 10023.



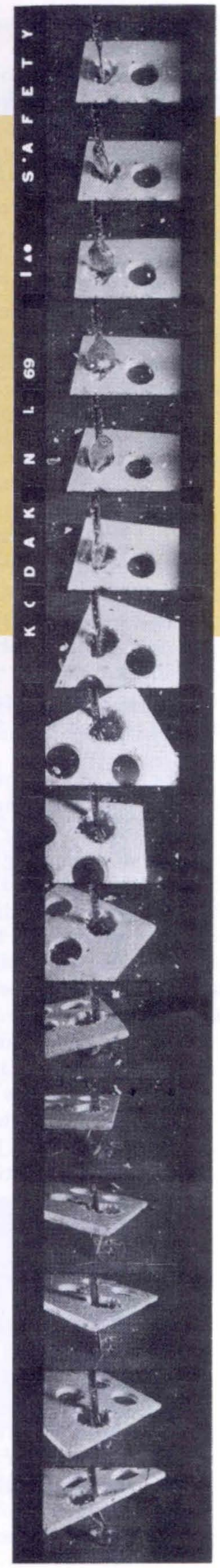


Figure 2. Study of a wood bit drilling an unsecured plastic block. Stroboscope and conventional movie camera uses camera/flash rate of 8 per second to view bit rotating at 2000 rpm. Apparent slow rotation is achieved by gradual increase of flash delay-control setting. Note that the 15-μs effective shutter speed produces the required sharpness of subject to permit detailed examination of individual frames.

b.

much, and whether it is stretch or shrinkage, and can make any required paper-roll tensioning adjustments.

The high light-output levels from the GR 1540 not only make stroboscopic viewing practical under normal ambient light conditions but permit "see-through" inspection of many paper stocks when both sides are printed and front-to-back registration must be maintained.

There are many other troubles that are monitored stroboscopically in printing operations: color register, location and quality of perforations, punched and die-cut holes, quality of over-all printing. Too "mushy" an impression would indicate perhaps too light a pressure or possible trouble in the inking fountains. A blooming or compressing effect at the ends of printed lines would indicate paper misalignment. Hickies would indicate water spotting, and dot break up in half-tone areas would indicate printing cylinder wear. We are sure that the printer for the *Experimenter* is noting these points as the presses roll!

A FEW TECHNICAL DETAILS ABOUT THE GR 1540

The desire for illumination levels well above those of the GR 1531 Strobotac® electronic stroboscope has resulted in unique design solutions. A photographer generally requires the highest possible *light per flash*, which is proportional to the energy in Joules or watt-seconds discharged through the lamp. *Visual intensity*, on the other hand, is approximately equal to the lamp input energy per flash multiplied by the rate at which the lamp flashes.

To satisfy both applications requires a high order of heat dissipation from the strobe lamp. The Xenon-filled lamp supplied with the Strobolume is an efficient converter of electrical energy to light and is protected from overheating by an electronic governor that prevents driving the lamp at rates in excess of the maximum allowable for the corresponding intensity settings. In addition, the quartz envelope of the lamp is forced-air cooled to withstand safely the high locally developed temperatures, thus permitting continuous operation at any flash rate up to 25,000 per minute.

The spectral output of a quartz-envelope Xenon lamp extends from the near ultraviolet, through the visible, and into the near infrared light regions. The GR 1540 is thus an excellent source of pulsed ultraviolet and infrared light when these normally invisible wavelengths are required, such as for detecting fluorescent ink registration marks on printing presses. A plastic faceplate normally covers the reflector and absorbs the ultraviolet light, to prevent possible eye irritation from prolonged exposure.

OTHER FEATURES

There are several new features in the GR Strobolume to extend its flexibility to both experimenters and photographers.

- Mechanical adjustment of the lamp-reflector combination produces either a narrow or broad light beam for such diverse projects as viewing the wide web of an operating printing press by means of a narrow, horizontal beam or photographing a large moving subject by means of the normal rectangular beam pattern.

C. E. Miller was graduated from Yale University in 1960 with a B. Eng. degree and received his MS degree from Massachusetts Institute of Technology in 1966. He joined General Radio in 1960 and is an engineer in the Component and Network Testing Group. He is a member of IEEE, AOA, ATI, SPSE, and holds a patent for a constant offset frequency-generating device to produce slow-motion images.

- Light output is approximately *twenty times* greater than that of the familiar GR 1531 Strobotac under *continuous* operating conditions. It may be increased, at lower flash rates, by using a booster capacitor.
- Three control units presently are available to satisfy a wide variety of applications. Their modular construction permits operating flexibility and provides features required for a particular application at minimum cost; equally important, it permits future expansion as needs dictate.

FURTHER HELP TO EXPERIMENTERS

Most applications for strobe light fall, very broadly, into three categories: speed measurement, motion analysis, and photography. Only a few specific cases have been covered here, but a comprehensive coverage is contained in two GR publications:

Handbook of High-Speed Photography ($1.00 U.S.)
Handbook of Stroboscopy ($2.00 U.S.)

Readers will find both books to be of great value in their respective fields. In addition, a free subscription to a GR periodical, *Strobotactics*, is available upon request. This publication contains descriptions of new and interesting applications, as well as other information of interest to users of stroboscopic equipment.

—C. E. Miller

Complete specifications for the GR 1540 are available on the catalog page, included as a tear sheet inside the back cover of this issue, removable for insertion in GR Catalog T.

A STANDARD-SIGNAL GENERATOR IMPROVES ITS VERSATILITY

Most standard-signal generators do not provide adequate sweep capabilities. On the other hand, sweep-signal generators lack both short- and long-term stability in the cw mode and are therefore of little value in critical point-by-point tests. To meet the increasingly stringent requirements for both sweep and point-by-point testing of high-frequency networks up to 80 MHz, General Radio has made available several new models of the well-established GR 1003 Standard-Signal Generator.[1]

Redesign of the auto-control model of the GR 1003 includes a permanent-magnet dc-motor drive with electronic speed control and a *second* horizontal-sweep output voltage at high level (1 V/1% Δf) to facilitate the display of narrow-band frequency sweeps. The original rf circuit design remains intact, with its high (1 ppm/10 min) carrier-frequency stability and with no need for a restabilization period following range changes.

The carrier-level-meter voltage scales are now calibrated in terms of voltage across a matched 50-ohm termination. This conforms to the current trend, described in this issue of the *Experimenter*, page 12; earlier models were calibrated in "volts behind" the 50-source resistance. If desired, "volts behind" can be obtained by multiplying the meter readings by 2.

The GR 1003 is an excellent instrument for measurement of very selective high-frequency networks by either sweep or point-by-point techniques. The slow-speed sweep, in conjunction with a storage oscilloscope display, is useful for measurement of pass-band characteristics. Two models include an internal crystal calibrator that provides frequency markers at 1 MHz, 200 kHz, and 50 kHz for accurately defining specific points of a display. The high-output capability is free of spurious non-harmonic signals. A highly stable cw signal is available for frequency measurements, at specific values of attenuation, on steep slopes of response curves. Repeatability of the frequency-dial setting is assured by the typical low-frequency drift rate (1 ppm/10 min; residual fm < 3 Hz). An auxiliary rf output jack on the GR 1003 provides a convenient monitor point for a frequency counter, such as the GR 1191 or 1192.

Capabilities and features of the new auto-control model generator include:

- Operation between two front-panel-controlled limits, at continuously-adjustable rates, from $\Delta f/f$ ranges of 0.05%/s to 5%/s. Return sweep is fast and blanks out the rf output.
- Sweep widths with motor drive, varying from 0.2% to over a full octave, provide good transitions to the electronic sweep limits.
- Availability of large sweep voltages, even with narrow sweeps.
- Direct correlation between the sweep voltage and the frequency dial, which is calibrated to 0.25% limit of error, provides a logarithmic display with constant-percentage resolution for wide sweeps. Sweep end points can be determined directly from the corresponding main frequency-dial readings.
- Voltage versus frequency is essentially linear in narrow-band ($\leq 4\%$ sweep width) frequency applications, permitting linear interpolation between the established limits. (Display devices usually have horizontal-axis calibrations in volts per graticule division, thereby permitting direct conversion of the generator's 1V/1%Δf change into hertz.)
- Very narrow sweep widths can be accurately calibrated by use of two different settings of the calibrated $\Delta F/F$ control on successive sweeps (Figure 1). The pattern offset provides direct calibration of the horizontal sweep, measured in parts per million, by taking the difference of the two $\Delta F/F$ settings.
- Tuning by means of the coarse motor drive can be accomplished at any convenient speed.

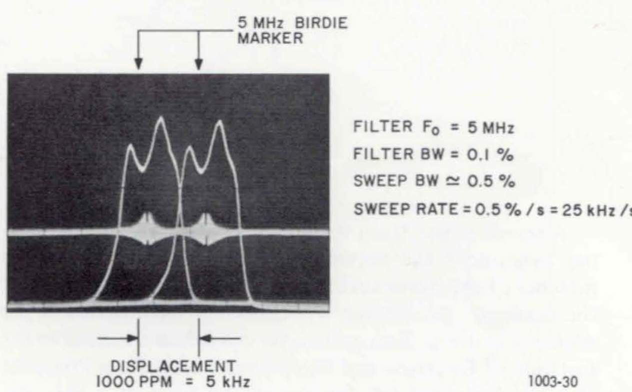


Figure 1. Successive-sweep technique as applied to calibration of very narrow sweep ranges.

[1] Altenbach, R., "The 1003 Standard-Signal Generator," *GR Experimenter*, July-August 1967.

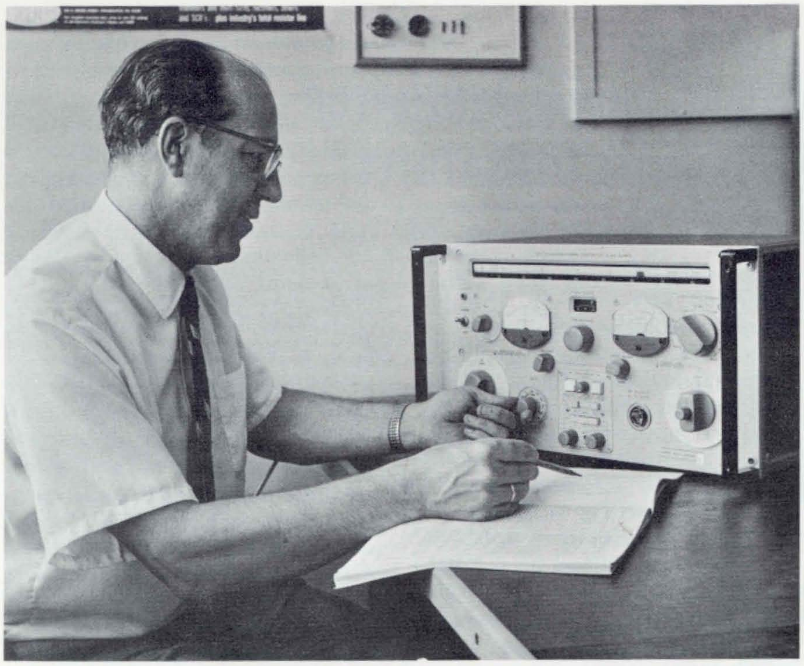


Figure 2. Engineer R. Altenbach adjusts auto-control model of GR 1003.

- Panel indicator lamps show the position of the sweep limit relative to the instantaneous tuning position, in the normal tuning mode. This directional feature greatly simplifies the initial set-up procedures.
- Programmable tuning with a positioning accuracy of 0.1% automatically seeks the programmed frequency at high speed.
- Operation on ac power from 50 to 400 Hz (50 to 60 Hz for basic model).
- Crystal-controlled markers (birdies), generated within two models of the instrument, can be displayed in either sweep mode. The markers can be switched to spacings as close as 50 kHz.

The basic sweep drive functions in a manner similar to that of the original design,[1] with some additions. The NARROWBAND SWEEP VOLTAGE control mechanically engages a separate sweep potentiometer that provides better resolution and higher sweep voltage. In the WIDEBAND mode, the sweep voltage is derived from the analog output through an electrical centering circuit. This allows expansion of the horizontal sweep display for optimum resolution, without restriction of the expansion by the oscilloscope centering control.

A useful relationship between maximum permissible sweep rate (SR_{max}) and steepness of the selectivity characteristic to be measured is expressed:

$$SR_{max} \simeq p\,(\Delta f_{6\ \mathrm{dB}})^2,$$

in which SR_{max} is expressed in Hz/s, $\Delta f_{6\,\mathrm{dB}}$ is the change in Hz for a 6-dB change in response at the steepest part of the response slope, and p is a constant whose value is of the order of unity.

As a practical example, consider a 9-MHz crystal filter with a 6-dB bandwidth of 2 kHz. This filter exhibits a maximum selectivity slope of approximately 100 Hz/6 dB. Maximum sweep rate is calculated to be 10 kHz/s or 0.1% at the 9-MHz center frequency.

Engineering development of this instrument was by R. K. Altenbach, Engineer, Signal Generator Group.

Performance changes for the new GR 1003 are noted on the tear sheet included with this issue.

[1] Ibid.

The Honorable Society

D. B. Sinclair

A recent notice from the National Academy of Engineering announced the appointment of Dr. D. B. Sinclair as member of the Aeronautics and Space Engineering Board of the academy. Dr. Sinclair was elected to membership in the academy in 1965. Recognition for contributions made to the Institute of Electrical and Electronics Engineers as President also was accorded Dr. Sinclair in the form of a specially designed Past President's pin, awarded at the IEEE annual banquet in New York, March 25, 1969.

Notes on FM Distortion in Varactor-Modulated Oscillators

Consider the basic resonant circuit shown in Figure 1, with an inductance L, a fixed capacitance C_o, and a voltage-controlled capacitor C_v. The variable C_v is often made up of 2 varactors connected back to back to reduce rf distortion. However, this has no bearing on the modulation characteristic.

Because of the nonlinear varactor characteristic both modulation sensitivity and distortion terms are dependent on the operating bias V_o and on the ratio p of the fixed C_o to the variable C_v. An analysis of this situation was made to compute the distortion terms and to determine the most favorable operating conditions. The frequency-modulation characteristic is shown with exaggerated curvature in Figure 2. The results are presented here and also in a graph (Figure 3), showing only second-order effects, namely carrier-frequency shifts and second-harmonic distortion of the desired frequency deviation. Third-order effects are generally one or more orders of magnitude lower.

The varactor is characterized by the relationship

$$C_v = \frac{C_1}{V^n} = \frac{C_1}{(\phi + E)^n} \tag{1}$$

where C_1 = capacitance at 1-V effective bias, i.e., 0.4 V applied bias in addition to the 0.6-V contact potential of a silicon device, and

V = effective bias voltage = $\phi + E$
ϕ = contact potential, typically 0.6 V for silicon
E = applied bias voltage
n = coefficient of varactor law.

Based on the resonant condition

$$\frac{1}{\omega^2} = L\,(C_o + C_v) \tag{2}$$

and the series expansion

$$\omega = \omega_o + \Delta\omega_m + \frac{\omega''}{2(\omega')^2}(\Delta\omega_m)^2 + \frac{\omega'''}{6(\omega')^3}(\Delta\omega_m)^3 + \ldots, \tag{3}$$

the following terms were derived:

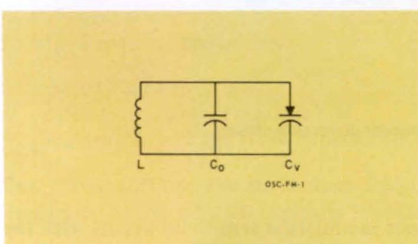


Figure 1. Schematic of basic resonant circuit.

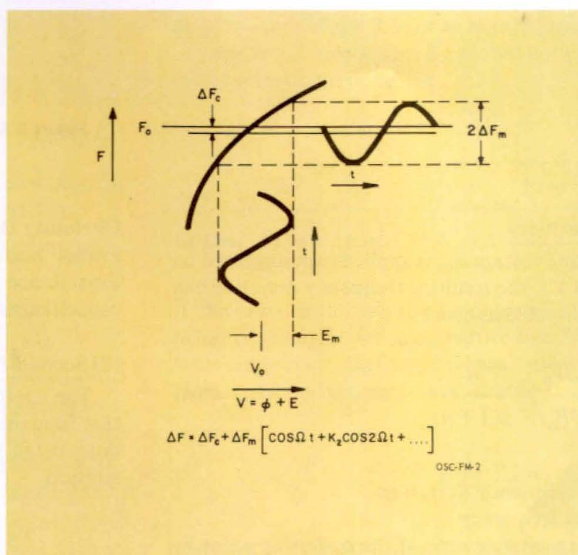
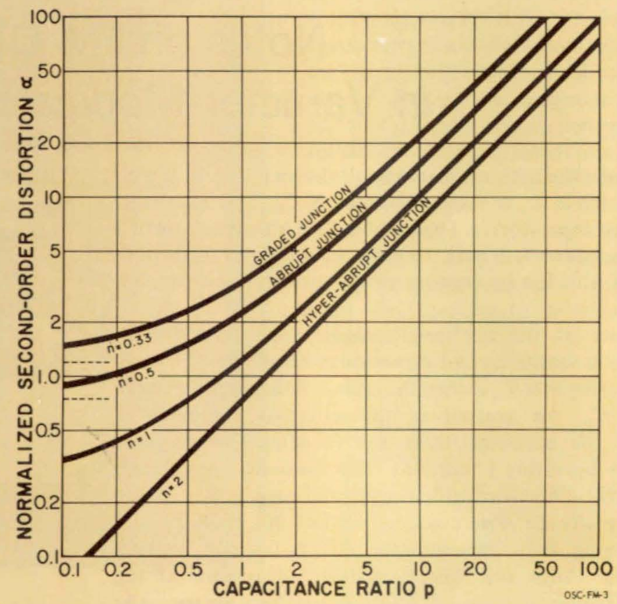


Figure 2. Frequency-modulation characteristic (exaggerated curvature).



$$\alpha = \frac{K_2}{\Delta F_m / F_o} = \frac{\Delta F_c / \Delta F_m}{\Delta F_m / F_o} = \frac{3}{4}\left[\frac{2(n+1)(1+p)}{3n} - 1\right]$$

$$p = \frac{C_o}{C_{V_o}}$$

n = COEFFICIENT OF VARACTOR LAW

K_2 = 2nd HARMONIC DISTORTION

ΔF_c = CARRIER SHIFT DUE TO MODULATION

ΔF_m = PEAK-FREQUENCY DEVIATION

F_o = CARRIER FREQUENCY BEFORE MODULATION

Figure 3. Second-order distortion effects.

(1) Modulation Sensitivity

When a modulating voltage E_m is applied superimposed on an effective dc bias V_o, the resulting frequency deviation can be expressed in a normalized form by

$$S_m = \frac{\Delta F_m / F_o}{E_m / V_o} = \frac{n}{2(1+p)} \tag{4}$$

where ΔF_m = peak frequency deviation
F_o = carrier frequency
p = the capacitance ratio at the operating point V_o,

$$= \frac{C_o}{C_{V_o}} = \frac{C_o}{C_1} V_o^n$$

Obviously the largest modulation sensitivity occurs with the lowest bias and when the varactor represents the only capacitance in the circuit. Because of unavoidable stray capacitances, this condition can never be quite reached.

(2) Second-Harmonic Distortion

The coefficient K_2 is the ratio of the second harmonic to the fundamental of the frequency deviation. When K_2 is normalized to the fractional deviation $\Delta F_m / F_o$, we get the relation

$$K_2 = \alpha \frac{\Delta F_m}{F_o} \tag{5}$$

The function α is shown mathematically and in graphic form in Figure 2 for various values of n and p. Distortion decreases with larger n and with lower fixed capacitance C_o. A hyper-abrupt junction-type varactor with $n = 2$ could make the distortion vanish in the absence of C_o.

With the more generally used varactors, for which $n < 2$, the inherent distortion cannot be reduced below a minimum value.

(3) Carrier Shift with Modulation

The same factor K_2 also describes the carrier shift ΔF_c in a normalized form

$$\frac{\Delta F_c}{\Delta F_m} = K_2 = \alpha \frac{\Delta F_m}{F_o} \qquad (6)$$

The relative shift in terms of the peak deviation is the same as the second-harmonic distortion.

(4) Third-Harmonic Distortion

Generally third-order effects are one or two orders of magnitude lower than second-order terms. For reference the term for the third-harmonic distortion is included here:

$$K_3 = \beta \frac{\Delta F_m}{F_o}$$

$$\text{with } \beta = \frac{1}{6}\left\{\frac{(n+1)\,(n+2)}{n^2}(1+p)^2 - 3\left[\frac{n+1}{n}(1+p) - \frac{1}{2}\right]\right\}$$

Conclusions

If lowest distortion is a main criterion, three conditions should be met simultaneously

– use of high-n varactors
– operation at lowest possible bias
– maintenance of the fixed capacitance C_o at a minimum.

For 1% distortion, the percentage frequency deviation is limited to $(1/\alpha)$%, which is typically 1% or less deviation.

The second condition also corresponds to operating with maximum modulation sensitivity.

Use of the Graph

Let us take an example to illustrate the use of the graph.

Assume an oscillator in the fm broadcast band 88 to 108 MHz. Suppose we want frequency modulation with 75-kHz peak deviation and 1% distortion or less. How much fixed capacitance can we tolerate?

The distortion is largest when the percentage deviation is largest, which occurs at the lowest carrier frequency (88 MHz). There, 1% distortion ($K_2 = 0.01$) makes $\alpha = 0.01 \times 88/0.075 = 11.7$. For this value of α we can read from the graph a value of $p = 7.3$ for an abrupt junction-type varactor, i.e., the fixed capactiance C_o should not exceed 7.3 times the varactor capacitance C_v.

f_c	ΔF	K_2	α	p	Capacitance fixed	Capacitance variable	Capacitance total
88 MHz	75 kHz	0.01	11.7	7.3	C_o	$0.137C_o$	$1.137C_o$

If the same circuit were to be operated at the high end (108 MHz) of the band, we could reduce L to 2/3 of the value at 88 MHz and keep the same capacitances; p and α remain unchanged, and the distortion would be slightly reduced.

$$K_2 = \frac{\Delta F}{f_c}\,\alpha = \frac{0.075}{108} \times 11.7 = 0.00812 \text{ or } 0.81\%$$

The carrier shift with modulation would also be 1% or 0.81% of the frequency deviation, i.e., 0.75 kHz at the low end, 0.61 kHz at the high end.

–R. K. Altenbach

R. K. Altenbach is a graduate of the Technical University of Karlsruhe (Dipl Ingr, Electrical Engineering and Communications, 1948) who held technical positions with Siemens & Halske, Canadian Marconi, Hermes/Itek, and Raytheon. He joined General Radio in 1963 and presently is an engineer in the Signal Generator Group. He is a member of IEEE and also served on EIA Committee TR14.

Signal-Generator Output Calibration

Considerable confusion has been engendered over the years by the existence of two competing methods of output voltage calibration for standard-signal generators, namely the emf or open-circuit or "volts behind" calibration and the matched output or "volts across" calibration.

Fundamental Relationships

Unfortunately no system of output calibration can absolve the engineer or technician of the need to understand his measurement technique and the equivalent circuits to which the metered voltages apply. In principle, it is unimportant which way the generators are calibrated so long as the method is clearly indicated on the panel of the instrument. The basic elements of the situation are shown in Figure 1.

The source voltage E_S is only available at the generator output terminals for a no-load or "open-circuit" condition. It can be shown that, for a *lossless* transmission line, the load voltage E_L is related to the source voltage E_S as follows:

$$E_L = E_S \frac{Z_L}{Z_S\left(\cos\beta\ell + j\frac{Z_L}{Z_o}\sin\beta\ell\right) + Z_L\left(\cos\beta\ell + j\frac{Z_o}{Z_L}\sin\beta\ell\right)} \qquad (1)$$

where Z_S = source impedance
Z_o = characteristic impedance of line
Z_L = load impedance
β = phase constant of line
ℓ = length of line

In general, either Z_S, or Z_L, or both, may be frequency dependent complex quantities, so the calculation is not trivial. The situation is considerably simplified if the transmission line is omitted. Letting $\ell = 0$, Equation (1) reduces to:

$$E_L = E_S \frac{Z_L}{Z_S + Z_L} \qquad (2)$$

In some ways more interesting is the observation that, when Z_S is chosen to be equal to Z_o, Equation (1) again reduces to Equation (2) except for an additional phase shift between E_S and E_L due to the presence of the line. Thus, for $Z_S = Z_o$, the magnitude of the open-circuit voltage at the output end of the transmission line is the same $|E_S|$ that would exist in the absence of any transmission line and is of course *independent of line length*. The result is that the magnitude of E_L depends solely on Z_L and not on some special conspiracy which may or may not provide a conjugate match with a complex Z_S.

For the special case of a purely resistive source $Z_S = R_S$ and $Z_L = Z_o = R_S$ we have a completely matched system whose equivalent circuit is shown in Figure 2a. The generator now delivers to the load the maximum power of which it is capable; this is called the available power, P_{AV}. Defining E_{LM} to be the load voltage under this matched condition, we have:

$$E_{LM} = \frac{E_S}{2} \qquad (3)$$

$$P_{AV} = \frac{E_{LM}^2}{R_S} = \frac{E_S^2}{4R_S} \qquad (4)$$

P_{AV} is usually measured in milliwatts, and generally calibrated in decibels relative to a one-milliwatt reference (dBm). E_{LM} is the value which is calibrated in the "volts across" system, whereas E_S is the value calibrated in the emf or "volts behind" system.

Historic Background

The rivalry between calibration systems has historic roots in the meeting, in the vhf region, of an upward-bound low-frequency technology, which was for many years open-circuit-voltage oriented, and of a downward-moving microwave technology, which was based on power measurements in closed transmission lines of well-controlled impedance. At one time, low-frequency receivers commonly had high-impedance inputs designed for direct connection to the end of a capacitive antenna or to a high impedance open-wire line. Early lower-frequency signal generators had output impedances which were typically less than 10 ohms, so that their open-circuit output was also essentially the terminal voltage for receivers whose input impedance was usually many times higher. These older hf signal generators were generally calibrated in terms of their open-circuit output voltage, and their output-impedance specifications were often rather vague. On the other hand, the basic output calibration of even early microwave signal generators was in

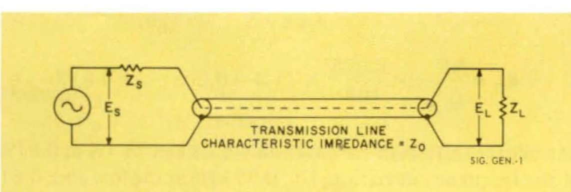


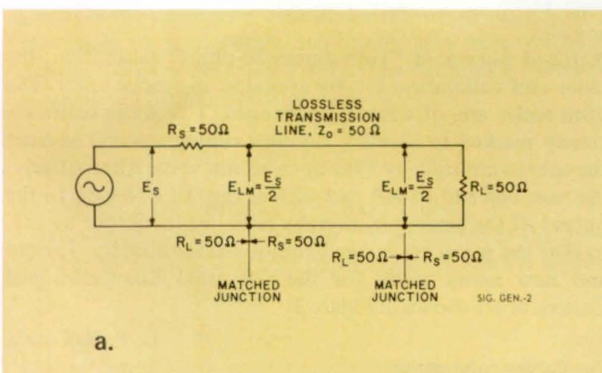
Figure 1. Basic calibration elements.

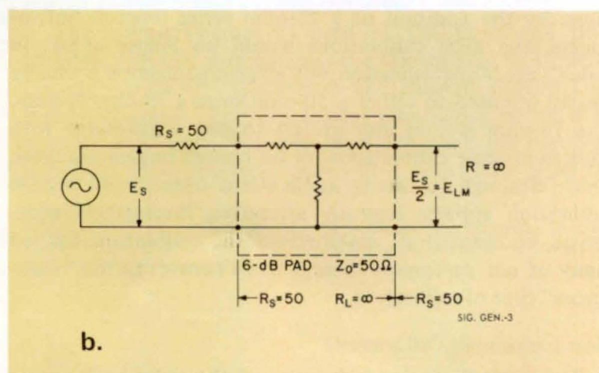
terms of available power, expressed in dBm. The output impedance characteristics were generally specified even when they departed markedly from the ideal. These instruments also carried voltage scales, frequently on a matched-voltage basis (Equation 3) for which the conversion to available power by Equation 4 was obvious. Some microwave generators, such as the GR 1021 series, carried open-circuit voltage scales and dBm scales. In either event, available power was a reasonable basis for calibration of microwave receiver sensitivity because impedance matching to obtain optimum power transfer was facilitated by early standardization on a 50-ohm characteristic impedance for coaxial transmission lines.

Present Situation

With the passage of years, most lower-frequency, open-wire, high-impedance inputs have given way to shielded coaxial circuits designed around 50-ohm or 75-ohm impedance levels, and high-frequency signal-generator source impedances have been standardized at these same levels. At the same time, dBm scales have become widely accepted on hf signal generators as well as on microwave instruments. Unfortunately, there has never been satisfactory standardization of the method of output-voltage calibration; some hf and vhf signal generators are calibrated in terms of emf or open-circuit voltage, and others in terms of matched output voltage. The emf practice has remained prevalent in Europe, while the matched-output practice has been most common on generators specified and procured by the U.S. Department of Defense; these procurements have had a major influence on calibration practices for commercial units as well.

Since both types of signal-generator-voltage calibration have existed in the U.S., the jargon of "hard" and "easy" microvolts has arisen among receiver designers. The term "hard" microvolts applies to the sensitivity measured by a generator with open-circuit calibration. This is because good sensitivity implies a small number of microvolts, clearly more difficult to achieve with this type of calibration. Conversely, sensitivities measured directly without correction, by use of a generator with matched-output voltage calibration, became known as "easy" microvolts. Most receiver test specifications require the use of "hard" microvolts, and the test procedures for use with *matched-output generators* call for insertion of a 6-dB pad at the generator output. The open-circuit voltage at the output of the pad is E_{LM} (Figure 2b), equal to the calibrated output voltage of the matched generator (Figure 2c). Thus the matched-output generator calibration can be read directly in cases where open-circuit values are desired. Direct reading of the dBm scale of a matched signal generator (without the extra 6-dB pad) gives a good measure of sensitivity for modern communication and navigation receivers that are designed for use with well-matched antennas and transmission lines. If the receiver input SWR is high so that it fails to accept the power available from the signal generator, the measured sensitivity suffers just as it would in practice when driven from a well-matched antenna. Since the dBm method of specification avoids the possible confusion of





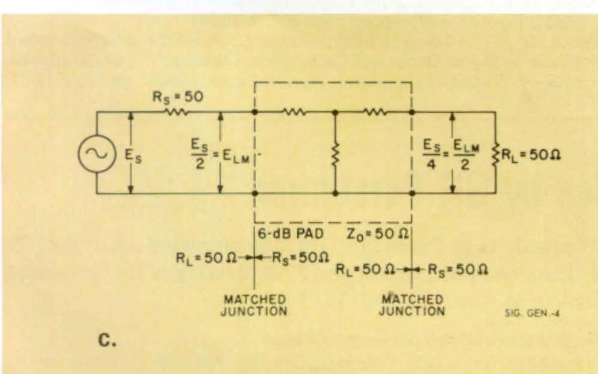


Figure. 2. (a) Basic transmission system relationships; as modified (b) by insertion of 6-dB pad and (c) by addition of matching load.

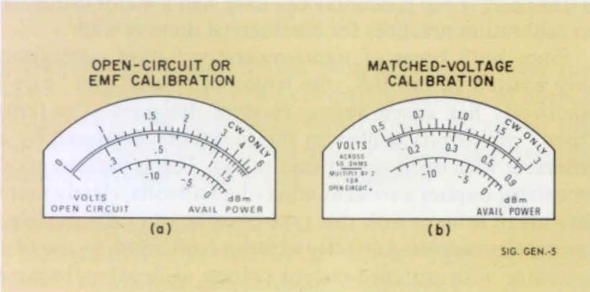


Figure 3. Old and new meter scales for GR 1003.

"hard" versus "easy" microvolts, it should be encouraged. Nevertheless, many sensitivity specifications still call for voltage calibrations, and one of the two types is generally provided on standard-signal generators.

For many years General Radio has supplied emf or open-circuit voltage calibrations. This is, in fact, the measured quantity at the monitoring or leveling point in most hf and vhf signal generators. Sensitivity measurements made with these calibrations correspond directly to the generally accepted "hard" microvolt type of specification. Calibration by use of open-circuit voltage is attractive for measurements on networks because this voltage is the effective Thévenin source voltage, independent of the length of line used to connect the generator to the network. There is the added advantage that the calibration remains unchanged if the signal-generator output impedance is increased from 50 to 75 ohms by the addition of a 25-ohm series resistor, but, of course, the dBm calibrations would no longer apply. In today's age of specialization, any given organization is usually clearly oriented to either a 50-ohm or to a 75-ohm system. The 75-ohm system user prefers to buy a generator with consistent scale calibrations for his chosen impedance level. Thus, although for many applications open-circuit-voltage calibration appears logically attractive, increasingly widespread acceptance of matched-voltage calibration has led many of our customers to urge us to convert to the "volts-across" type of calibration.

New Instrument Calibrations

The original mechanical design of the output-attenuator dial mechanism in the GR 1003 and 1026 generators made conversion from one calibration system to the other uneconomic. New design features, however, have recently been introduced which allow us to change over quite readily. Current production of these two instruments now carries matched output or "volts-across-50-ohms" calibration; the older emf calibration is only available on special order. The dBm scales are, of course, unchanged. The meter scales are clearly marked to identify the new calibration and instruct the user to multiply by 2 for open-circuit volts. Alternatively, the user can add a 6-dB pad such as the GR 874-G6L to the output of the generator, thereby reducing its output by 2:1, so that the meter reads open-circuit voltage directly. The old and new meter scales for the GR 1003 Standard-Signal Generator are shown in Figure 3.

—G. P. McCouch

G. P. McCouch joined General Radio in 1957, after working with the Harvard Radio Research Laboratory and Aircraft Radio Corporation. After receiving his AB in Physics from Harvard in 1941, he returned as a teaching fellow in 1947-48, to work toward his AM (1948). Presently he is Section Leader of the RF Oscillator Section. He is a senior member of IEEE and has held several offices in IEEE.

For Further Information

Peterson, A. P. G., "Output Systems of Signal Generators," *General Radio Experimenter*, June 1946.

Moore, W. C., "Signal Generator and Receiver Impedance," Boonton Radio Corp. *The Notebook*, No. 3, Fall 1954.

Woods, D., "The Concept of Equivalent Source EMF and Equivalent Power in Signal Generator Calibration, *"Proceedings of the Institution of Electrical Engineers*, Part B, Paper 3356M, January 1961.

Recent Technical Articles by GR Personnel

"Computer-Controlled On-Line Testing and Inspection," P. H. Goebel, *1969 Wescon Technical Papers – Session 8*.*

"True RMS – The Digital Approach," J.A. Lapointe, presented at the 77th Meeting of the Society of America, April 1969.**

"Versatile High-Speed One-Third Octave Band Analyzer," W. R. Kundert, presented at the 77th Meeting of the Acoustical Society of America, April 1969.**

*Reprints available from General Radio.

**Included in article "New-Generation Acoustical Analyzer," *GR Experimenter*, May/June 1969.

Damping Measurements of Resonance Bars

Researchers at Notre Dame University have approached the study of material damping capacity from the beginning, rather than the end. One of the most common procedures for measuring the internal friction of material is the resonance-bar technique, wherein a sample of the material to be studied is cut to the proper dimensions for resonance at some desired frequency in some specific mode of vibration. Damping is determined either with the specimen at resonance or in free decay from resonance. Dr. N. F. Fiore and R. M. Brach of Notre Dame have successfully developed a third variation of the technique in which the damping is determined by the relaxation time (τ_b), characterizing the rate at which the system *builds-up* to steady-state resonance vibration.

A disadvantage of both the steady-state and free-decay measurement approaches is the necessity to bring the material sample to steady-state resonance before the damping can be determined. In this circumstance, rapid physical changes can be taking place within the sample and be lost to the observer, because of the comparatively long time required to reach resonance and to begin serious observations. The approach described here permits the researcher to measure the damping decrement in thousandths of a second while the material sample is vibrated from rest to steady-state resonance.

The time taken by the vibrations of a specimen, under steady-state resonance conditions, to decay to e^{-1} of its maximum or steady-state amplitude, after removal of the driving forces, is designated as τ_d. The damping decrement δ is calculated from the relationship

$$\delta = (f_r \tau_d)^{-1}$$

in which f_r is the resonance frequency in hertz.

Conversely, the time (τ_b) for the amplitude of the driven system to reach $1 - e^{-1}$ of its steady-state value is approximately equal to τ_d. From this relationship the damping decrement can be redefined as

$$\delta = (f_r \tau_b)^{-1},$$

giving a faster and much simpler determination at the first stages of vibration at a time when the sample has been least affected by the driving force.

At steady-state condition, the decrement can be defined as

$$\delta = \frac{\pi(f_2 - f_1)}{f_r},$$

in which f_r is the resonance frequency and f_1 and f_2 are the frequencies at which the magnitude of vibration is 0.707 that at resonance.

The Notre Dame researchers first derived the mathematical relationships between the two methods, as described in the *Journal of the Acoustical Society of America,*[1] and then confirmed their calculations in a series of resonant-bar experiments. They used a Marx composite piezoelectric oscillator in which the input to the system is directly proportional to the input voltage across the driving transducer and the decrement is inversely proportional to the output voltage across a gaging transducer.

The circuit (Figure 1) used by Dr. Fiore and Brach employs two GR instruments, the GR 1162-A7C Coherent Decade Frequency Synthesizer and the GR 1396-B Tone-Burst Generator. The synthesizer generates a resonant frequency, at a controlled voltage, which feeds the tone-burst generator. The generator puts out a controlled pulse that

[1] Fiore, N. F. and Brach, R. M., "Resonance-Bar Damping Measurements by the Resonance Build-up Technique," *Journal of the Acoustical Society of America* (in press).

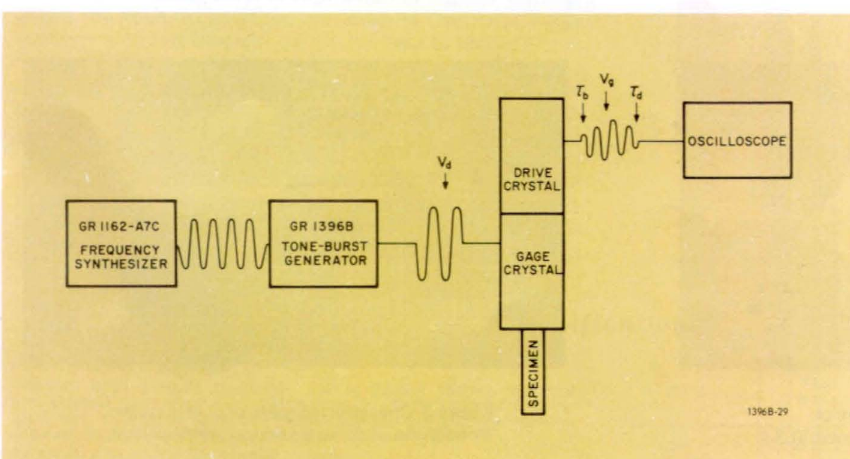


Figure 1. Block/schematic diagram of test equipment for resonance-bars experiments.

energizes the driver crystal; the vibrational magnitude increases from zero to the steady-state value. Resonance buildup is monitored on a storage oscilloscope, which displays the gage crystal output.

The method described above is unique in that it allows one to have his cake and to eat it too. As long as the pulse duration exceeds τ_b, the decrement can be determined directly from the relaxation time. When the pulse ends, the sample-rod driving force is zero; the vibrations and the gage-signal output decay to zero. From the oscilloscope display it is possible to obtain as many as three measurements of the decrement δ. During the time from rest to steady state at resonance, the decrement is defined by the build-up relation time τ_b. At steady state, δ is inversely proportional to the steady state amplitude. During the time from steady state to rest, δ is defined by τ_d. If high-speed measurements are required, the pulse duration can be made less than τ_d; then τ_b can be derived by fitting an expotential curve to the approach envelope as displayed on the oscilloscope.

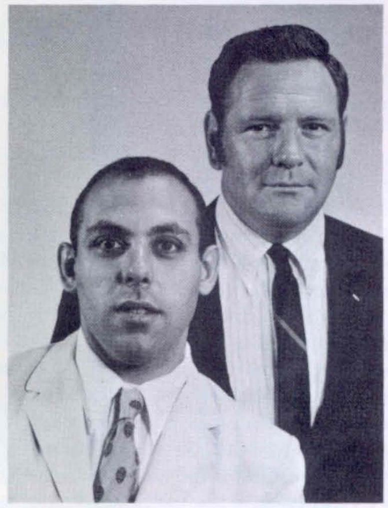
Figure 2 shows the oscilloscope trace for the quartz crystals alone, when driven by a 0.3-second burst in the longitudinal mode, at 50 kHz, and with constant applied

Table 1

Condition	τ_b ms	τ_d ms	δ Buildup	δ Free Decay	δ Steady State
Marx Oscillator	44	49	4.4×10^{-4}	3.9×10^{-4}	3.6×10^{-4}
Marx plus Cu Crystal	2.4	2.5	8.1×10^{-3}	7.9×10^{-3}	8.4×10^{-3}

voltage of 1.8 V rms. Values for characteristic times, taken directly from the figure, give τ_b = 44 ms and τ_d = 49 ms. Values of the decrement as calculated for the three variations of the resonance-bar measurement technique are shown in Table 1.

Figure 3 shows the change in response brought about by mounting an annealed (111) copper single crystal upon the drive crystal and maintaining the applied voltage at 1.8 V rms.



Dr. N. F. Fiore (left) is Associate Professor and Chairman, Department of Metal Engineering and Materials Science, University of Notre Dame. He is a graduate of Carnegie Institute of Technology and holds degrees of BS, MS and PhD in Metallurgical Engineering.

Dr. R. M. Brach (right) is Associate Professor, Department of Aerospace and Mechanical Engineering, University of Notre Dame. He holds degrees of BS and MS from the Illinois Institute of Technology and received his PhD in Engineering Mechanics from the University of Wisconsin.

The large increase in damping causes rapid attainment of steady-state resonance and rapid decay to zero after completion of the tone burst. Calculated values for the decrement, derived by the three variations in technique, are shown in Table 1.

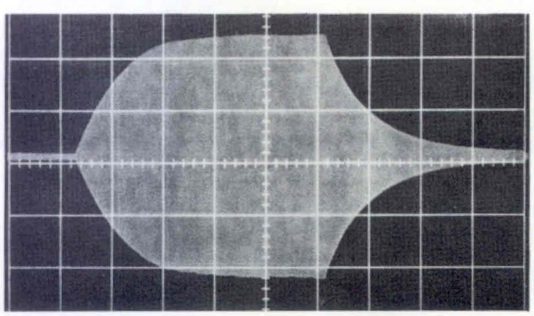


Figure 2. Oscilloscope picture of quartz-crystal response to 50-kHz burst (0.3 s) in longitudinal mode.

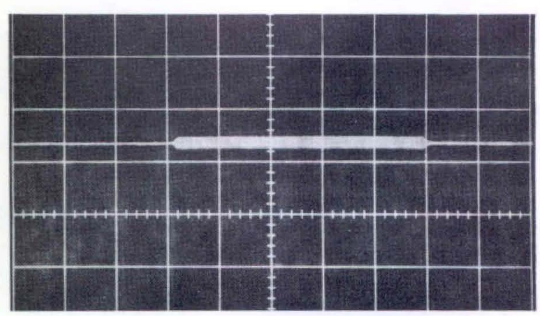


Figure 3. Change in response of quartz crystal by addition of annealed copper single crystal.

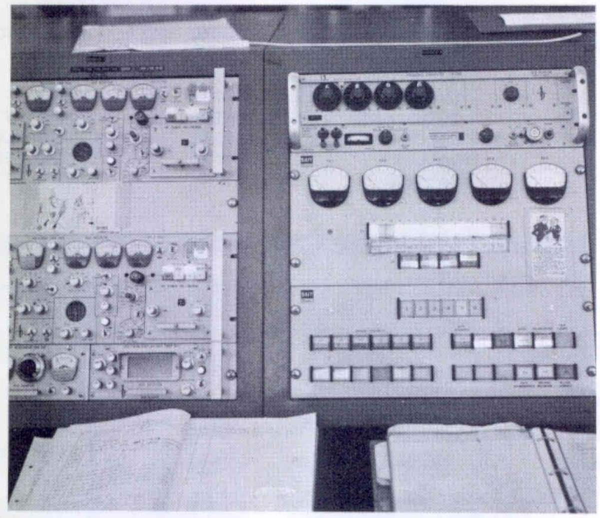
Reports from the Field

FROM THE FAR NORTH

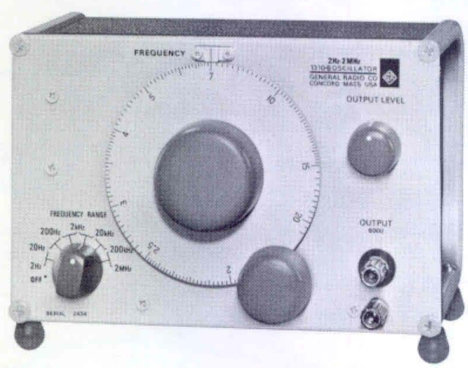
The Kongsfjord Telemetry Station in Spitzbergen, Norway operates as an observation and data-collection point for the European Space Research Organization (ESRO) in addition to handling its routine tasks for the Norwegian government. Engineer Bjorn Myrstad reports that the GR 1163-A4 Coherent Decade Frequency Synthesizer is an integral part of the station's system. It supplies 100-kHz signals for modulation of a test generator and controls three other signals – the telemetry transmitter's second and first local-oscillator frequencies and a test-generator operating frequency.

ESRO is geared for an extensive satellite launch program. Six satellites are scheduled for launch in the next six years, to participate in scientific experiments including cosmic-ray studies. The last satellite is scheduled for launch into a stationary orbit and will contain ten experiments controlled by several European countries.

The photograph contributed by Mr. Myrstad shows the GR synthesizer, upper right, part of the station's system supplied by the Sud Aviation Company.



NEW



Complete conversion of the GR 1310-B Oscillator to solid-state circuitry has not altered the outer appearance of this familiar oscillator. It has, however, enabled GR to lower the price, making the instrument even more attractive to users.

Specifications for the GR 1310-B are the same as for the GR 1310-A in GR catalog T, except for accuracy (± 3% of setting) and price.

Catalog Number	Description	Price in USA
1310-9702	**1310-B Oscillator** (Specify 115-, 220-, or 230-V line operation)	**$275.00**
1560-9695	**1560-P95 Adaptor Cable**	**3.00**
0480-9838	**480-P308 Rack-Adaptor Set**	**10.00**

All prices subject to quantity discount.

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

In the lead article of this *Experimenter* we have taken issue with a view held by many people – that computers are intelligent. The article is presented with the hope that the aura that surrounds computers, in general, will be reduced and that computers will be accorded the respect due them, but only to the extent that it is deserved.

If, however, computers are more wily than we are prepared to admit, we hasten to draw attention to our cover. By assigning to a computer the slight human frailty of repeating itself, and also attributing to it a small sense of human kindness, we have made it serve our purpose. For, unlike our elfin operator, who forgot what season of the year it is, our computer has "remembered" and stirred itself to speak for all of us at General Radio.

To all our readers, wherever you are, we sincerely hope you will join with us in this special season, to count our blessings and to give thanks for all that is good. We wish you, one and all, "SEASONS GREETINGS."

C E White

C. E. White
Editor

Why Computers in Measurement Systems?

"Computers are smart?"* is a question raised quite often even by people who have knowledge of, access to, or control of computers. Of course computers aren't smart -- they are clever *tools* that reflect the intellect of the special group of people who have wedded themselves to these machines. For the past two decades, many in this special group have devoted themselves to the specific task of forging an alliance of computers, humans, and measurement instruments.

Just why should all this effort be expended in this direction? Well, first let's review some of the advantages and disadvantages of the simple combination of measuring instrument and operator. The operator's mind is able to grasp the significance of data derived from the instrument, to recognize patterns in the observed test data, and to react in the appropriate and positive manner. All this can be done, but not for any prolonged interval of time, and not very fast. Fatigue sets in, the operator's mind is less responsive to signals, and the operator's usefullness deteriorates rapidly. On the other hand, a computer is not susceptible to fatique (exclusive of electronic or mechanical deterioration). Properly programmed, the computer can perform all the above functions – faster!

But even with its ability to store *all* the knowledge derived by the operator from training and experience, the computer is merely an electromechanical *slave*, incapable of thinking, analyzing, and reacting to a problem for which we have not supplied instructions. Why then is it so important to utilize the services of the computer? Let's look at a few specific advantages of the combination of computer, operator, and measuring instrument:

- Instrument operation can be faster.
- More data can be derived in a given time interval.
- Comparisons of derived data with built-in reference standards, tolerances, or specifications are quicker by orders of magnitude.
- Provided bilateral interfacing has been built into the system, the operator maintains strict control of test operations.
- Systematic errors due to operator bias are eliminated.
- Test data can be displayed automatically in any of several forms – paper or magnetic tape, punched cards, line printers or teletypewriters, automatic graphic recorders or X-Y plotters, etc.

The advantages inherent in compatible interfacing between the units of the combination are more apparent as we trace the growth in complexity of measurements and realize the need to perform and interpret measurements more quickly.

Measurement Systems are COMPLEX!

Development of manual-measurement techniques has progressed from single-point measurements of simple parameters to swept measurements with automatic readout and

*According to Hughes Aircraft publication *Vector* (First Quarter 1969), "The credentials of any machine that believes 1 + 1 = 10 are suspect."

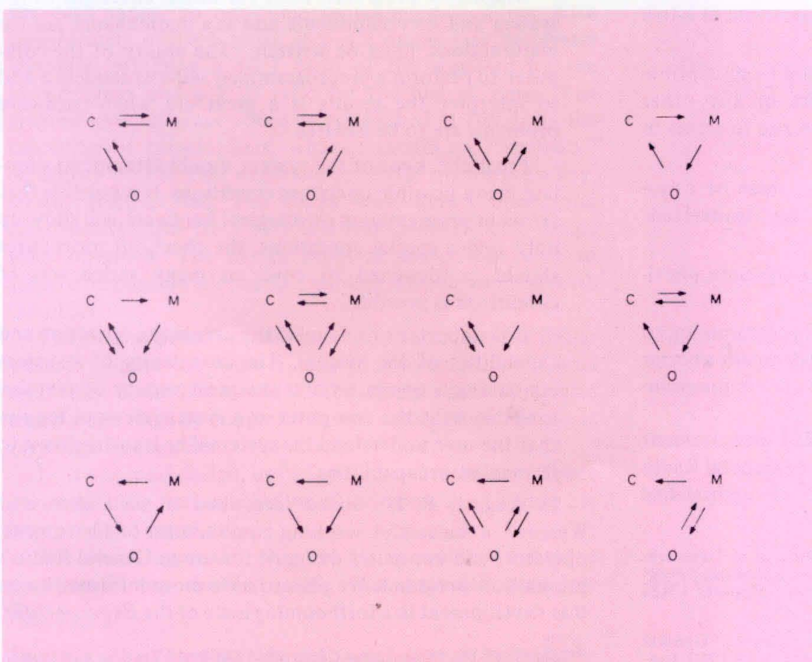


Figure 1. A few of many arrangements for instrument-computer-operator interfaces.

data displays followed by analysis of these data by comparison with recorded standard performance data. The complexity of the latter technique is sufficient to overtax the ability of an operator.

As measurement techniques have evolved there has been a metamorphosis in the systems and techniques of recording and applying measurement data. The simple inspection data sheet that displayed a measurement result vis-a-vis a reference standard, and alerted an operator to make a quality evaluation to accept or reject the inspected item, has now changed into the automatic visual presentation and analysis of data. In such a system, units tested are automatically accepted or rejected, adjusted from out-of-tolerance conditions to within tolerance (if feasible), or the computer may alter the test sequencing for further analysis or action. For specific applications, the interfaces can be unilateral or bilateral[1] and in any number of arrangements, some of which are shown in Figure 1.

At this point it would be wise to define two terms prevalent in the work of coupling computers, instruments, and operators. *Computer-assisted* refers to the general case of the combination of computer and instrument system in which the computer does not exercise control of the test agenda, reacts passively to record data for analysis, and alerts the operator to take certain actions. The term *computer-controlled* signifies both indirect and direct control by the computer over the measurement system and test process. The computer asserts authority over the measurement system by transmitting command pulses to programmed measuring equipment and control units. In each case, assist or control, the computer memory has stored within it all the instructions required to conduct the test program.

Programmable Instruments Don't Just Happen Along

Modern instrument designers must consider problems such as these if they wish to incorporate an instrument into a programmed measurement system:

> Programming circuitry or units must not be susceptible to shock, vibration, temperature, noise or any other environmental condition that could give rise to errors in program signals.
>
> Very short time commands, of the order of nanoseconds, could be delayed effectively by inadvertent introduction of long-line connections.
>
> Interfacing between units must be compatible physically, electrically, and mechanically.

Unless these problems and others are solved, programming an instrument does not necessarily make an instrument a better one. Many measuring instruments, *but not all*, are programmable.

Design of a practical computer-controlled measurement system was outlined by M. L. Fichtenbaum of General Radio recently during a GR measurement seminar.[2] He approached the problem in this manner:

Evaluate the measurement requirements in terms of what can and cannot be done. A system proposal may call for straightforward tasks; it may also demand measurements that are impractical.

Check whether there is available hardware to perform the measurement functions. Is it possible to modify existing instruments, or to make measurements from which the computer can calculate the required information? What would be the cost of developing any special-purpose equipment?

Decide with the user on final system specifications. It is likely that the system designer does not know all about the user's needs; it is also likely that the user does not know all the system designer's limitations and capabilities.

Choose a specific hardware configuration. With the final instrumentation defined, the interface necessary to tie the instruments to the computer can be designed. Typically, an interface will consist of enough standard modules to transfer all necessary data between computer and instruments, plus one or two "special" modules to perform functions specific to the system.

Consider the three main areas in software development. Each instrument in the system has associated with it program segments to control the instrument and to translate data between instrument and computer formats. In most cases, these "instrument module" programs can be developed for use with one instrument and used essentially unchanged in all systems that use the instrument. Each system will have a unique "main-line" program to perform the test functions required. These include the initial setup (of limits, ranges, etc), the sequence through test configurations, and presentation of data in the proper form.

Diagnostic programs, both for initial checkout of the system and its components and as a maintenance aid for malfunctions, must be written. The ability of the computer to perform a preprogrammed self-test sequence and to interpret the results is a great aid when hardware problems are to be isolated.

Final checkout of the system should attempt to simulate many possible operating conditions. It is possible that errors in programming or marginal hardware will show up only under special conditions; the checkout procedures should be designed to cover as many varied sets of conditions as possible.

It is important to educate the user in the operation and capabilities of the system. The complexity of a system containing a computer and the great variety of features available with the computer as a system element require that the user understand the system if he is to employ it to its maximum capabilities.

In August, P. H. Goebel described to an audience at Wescon[3] a successful working combination of instrument, operator, and computer designed for use in General Radio's production programs. We plan to have more information on this development in a forthcoming issue of the *Experimenter*.

[1] Beatty, R. W., "Short Discussion of Error Reduction in Network-Parameter Measurement Through Computerized Automation," *Progress in Radio Science – 1966-1969*, (Commission I Report) URSI XVIth General Assembly, Ottawa, August 1969.

[2] Fichtenbaum, M. L., "Computers in Instrument Systems," General Radio Automatic-Impedance-Measurement Seminar, May 1969 (unpublished).

[3] Goebel, P. H., "Computer-Controlled On-Line Testing and Inspection," *1969 Wescon Technical Papers, Session 8*, August 19-22, 1969.

Search for a Better Transformer

Far back, in the comparative antiquity of June 1926, General Radio was very much interested in the problems of transformer design.[1] During the years that followed we never relaxed our interest, nor our desire to tell our readers about what is considered good design.

Burke noted in 1926 the difficulty of designing ratio transformers for undistorted response over extreme ranges of frequency – 100 to 5000 Hz! Among other things, he demonstrated the importance of matching transformer impedances to the signal source and to the load. These elementary problems are still with us.

A recent problem was to design a ratio transformer for stable operation at 1 MHz and with low leakage impedance. Why? Well – we knew that a transformer could provide an excellent means of decade ranging for the new GR 1682 Automatic Capacitance Bridge.* Since there did not appear to be one available commercially, we continued our research into methods and means of devising such a unit.

During several sessions with GR engineer Dick Sette, the *Experimenter* editor was able to trace the progress made, from the first round-table discussions, through the preliminary design stages, to the final production models of the finished products. Note the use of the plural term – the final result was not one but two devices, each unique for its own function in the capacitance bridge.

The problems, as presented by Sette, were to supply a maximum stability to a wide range of measurements at 1 MHz and to achieve good accuracy at the UNKNOWN position 4 feet away. Thoughts of using computing amplifiers were considered and then dropped because of cost and state-of-the-art development.

It was considered necessary to use different transformers for each range, to overcome switching difficulties. With transformers of 10:1 ratio, it was a simple task to reverse windings, effecting a 10:1 ratio of the standard for the lowest range and a 1:10 ratio for the highest range. The greatest over-all range could be established at 100:1, by the cascading of transformers, without sacrificing "accuracy" capability of 1%. Past experiences of GR engineers Smiley, Hersh, Holtje, Hall, and Fulks were drawn upon, and constructive ideas for a very low leakage-impedance design were obtained. The design problem posed was: how to achieve maximum electrical coupling of one winding with the other.

The first approach considered was a toroidal core, bifilar wound (Figure 1), with a twisted 11-wire cable wound uniformly around the toroid. The result was a 10:1 ratio transformer, tightly-coupled but not satisfactory for use at very high frequencies.

In the second design, use of cup cores in a toroid (Figure 2a) resulted in better coupling. Rewinding, as in Figure 2b, improved the coupling.

The third approach (Figure 3), in which we used a multi-conductor shielded cable that provided very low

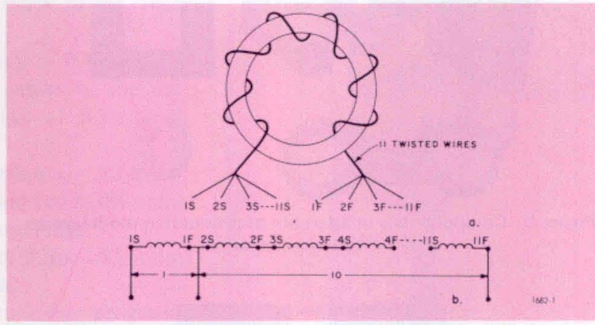


Figure 1. Simple design for 10:1 ratio transformer.

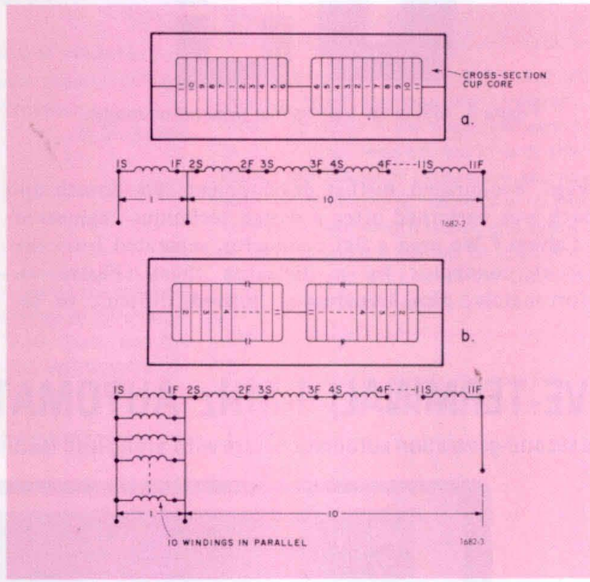


Figure 2. Use of cup cores for an improved design.

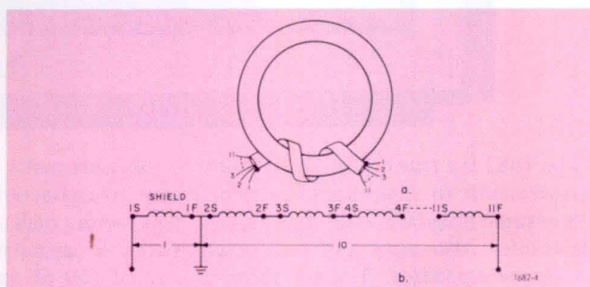


Figure 3. Multi-conductor shielded-cable winding.

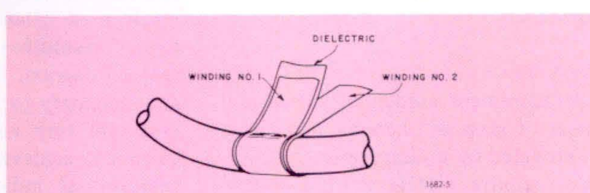


Figure 4. Flat-conductor transformer-winding design.

*See page 6.

[1]Burke, C. T., "Amplifier Ins and Outs," *General Radio Experimenter*, June 1926.

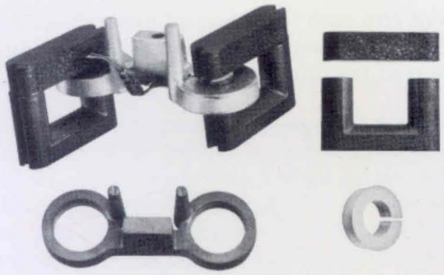


Figure 5. Components and assembly of production-model design of 10:1 ratio transformer.

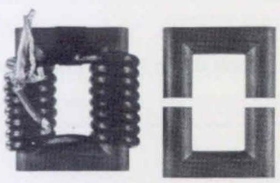


Figure 6. Mid-range transformer production model.

leakage, encouraged further development. The fourth approach was patterned after a design technique engineered by Calvert.[2] We used a flat conductor, separated from the adjacent conductor by a dielectric sheet (Figure 4). Unfortunately, this transformer proved difficult to assemble, and the coil had high capacitance loading. The latter problem was a serious detriment inasmuch as it increased the driving requirements excessively.

Finally, the decision was made to go to a machined-core design. A split, hollow, brass bobbin was made and used as a single turn. The bobbin was silver-plated to minimize skin effects at 1 MHz. To improvise leads for the single turn, the casting that held the bobbins was made with self-circuit leads (horns) that penetrated the mounting board for ease of connection. Ten turns were wound on the bobbin, which was sweated into place in the casting. Figure 5 illustrates the assembly. In order to switch the leads, to provide the desired ratio direction, reed relays were connected to the circuitry.

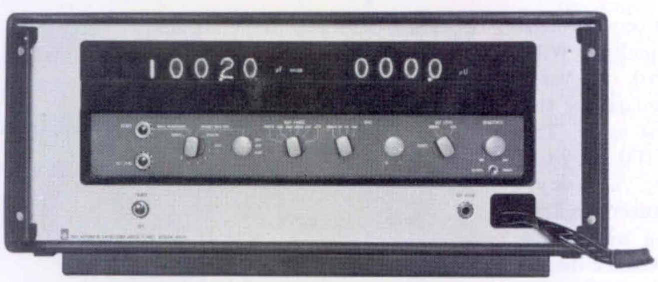
We still had the requirement for the mid-range 1:1 coupling transformer. A simple, balanced, 1:1 ratio transformer was constructed (Figure 6), consisting of a Triax cable wound around a core. The outer and inner shields were used as conductors; the inner conductor was ignored. This technique provided a transformer, simple to manufacture, with one conductor completely surrounded by the second, assuring maximum coupling.

These development chores completed in a satisfactory fashion, manufacturing was able to move out on the final construction of the new GR 1682 Automatic Capacitance Bridge.

[2] US Patent No. 2,659,845 assigned to Wayne-Kerr Laboratories, Ltd.

FIVE-TERMINAL, 1-MHz AUTOMATIC CAPACITANCE BRIDGE

The second-generation automatic bridge with a long-lead feature



The 1682 is a *true* bridge, capable of measurements to high accuracy and with ensured long-term stability. It uses transformer ratio arms and precision admittance standards. The five-terminal-type connection for the unknown minimizes the inaccuracy effects of lead impedances.

Why Measure at 1 MHz?

Measurement guidelines for several classes of glass and ceramic capacitors are provided by military specifications, which require 1 MHz as the measurement frequency.

Measurements at 1 MHz are necessary to characterize low-value capacitors with high shunt conductance. Given a capacitance-conductance circuit of 100 pF and 10 kΩ, a measurement at 1 kHz would yield a large real-current component, which would result in a balance with considerable loss of resolution, as indicated by the vector diagram, Figure 1a. The real and imaginary currents at 1 MHz, however, are very nearly equal, and the balance is achieved with maximum resolution, as indicated by the vector diagram, Figure 1b.

In the integrated-circuit industry, capacitance versus junction-bias-voltage measurements are important sources of analytical data from which several characteristics of semiconductor junctions can be determined. Because present isolation techniques use reverse-biased junctions and thin-oxide films, shunt losses are unusually high for the small-value capacitors being measured. Therefore, measurements at 1 MHz provide maximum resolution and accuracy.

How We Did It

Operation at 1 MHz with long cables to the unknown called for a design to include a ratio-transformer, 5-terminal, ac Kelvin bridge and was, in effect, an extension of a lower-frequency version.[1] The result is the five-terminal configuration shown in Figure 2. To achieve minimum measurement errors due to shunt loading, low-leakage-

[1] Hill, J. J. and Miller, A. P., "An AC Double Bridge with Inductively Coupled Ratio Arms for Precision Platinum-Resistance Thermometry," *Proceedings of the Institute of Electrical Engineers,* February 1963.

impedance ratio transformers were developed.* Loading Z_A appears across the ratio transformer winding N_2, and loading Z_B is across the low input impedance of the bridge preamplifier.

It is extremely important to evaluate the accuracy of a bridge at the terminals of the capacitor being measured, not at the bridge terminals. If a three-terminal measurement were attempted at 1 MHz, the inductance of the go and return paths to the component would cause considerable error, restricting the measurement to the bridge terminals. For example, if a 1000-pF capacitor is measured at the end of two twisted coaxial leads, three feet long, the error in measurement will be approximately 4%.

To provide meaningful measurements of semiconductor-junction parameters, the test-signal level at the unknown is low, 500 mV to 5 mV over the three lowest ranges, and is constant during balance operations in any one range.

The balancing technique is simple, implemented entirely with integrated circuits. The automatic-ranging feature is designed to provide maximum resolution of measurements over the four bridge ranges.

*See page 5.

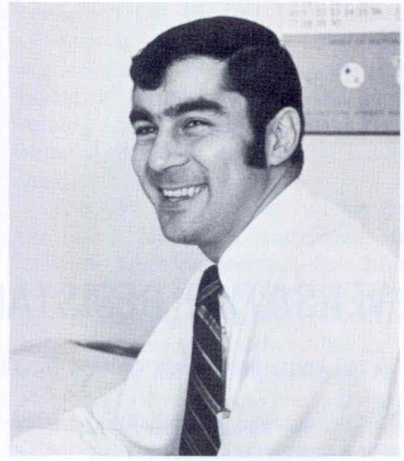
Outstanding Features

- Wide range – 00.001 to 1999.9 pF and 02.00 to 19.99 nF in four ranges.
- High basic accuracy – 0.1% at the end of four-foot cables to the unknown, due to use of 5-terminal connections.
- Rapid automatic balance – 20 measurements per second for ±10% components, on any one range.
- Built-in bias from 0 to 100 Vdc; external to 200 Vdc.
- All functions remotely programmable.
- Various data output and test fixture options.
- Accessory test fixtures available for use of rf capacitance standards* to check bridge performance.

– R. F. Sette

*Typical are GR 1406, 1407, and 1403 Capacitance Standards.

The GR 1682 was developed by the author. D. S. Nixon provided the design of the digital servo; W. A. Montague, D. W. Carey, and A. W. Winterhalter contributed significantly to the mechanical design, etched-circuit layout, and technical support.



R. F. Sette received his BSEE and MSEE degrees from Northeastern University in 1960 and 1962 respectively. He joined GR in 1964 after research work with the AF Cambridge Research Laboratory and service with the US Army at the electronics laboratory at Ft. Monmouth testing and evaluating IC circuits. He is a member of IEEE and Eta Kappa Nu and works with the GR Component and Network Testing Group.

Complete specifications for the GR 1682 are included as a tear sheet at the back of this issue.

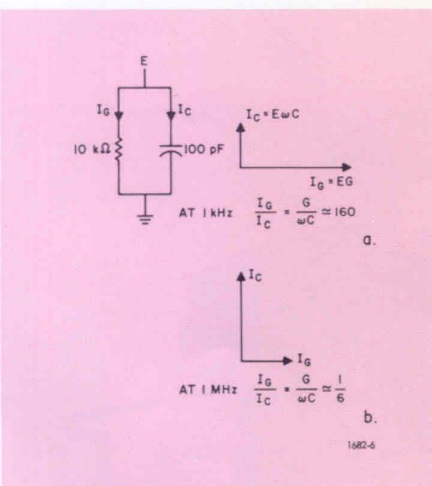


Figure 1. Current vector diagram.

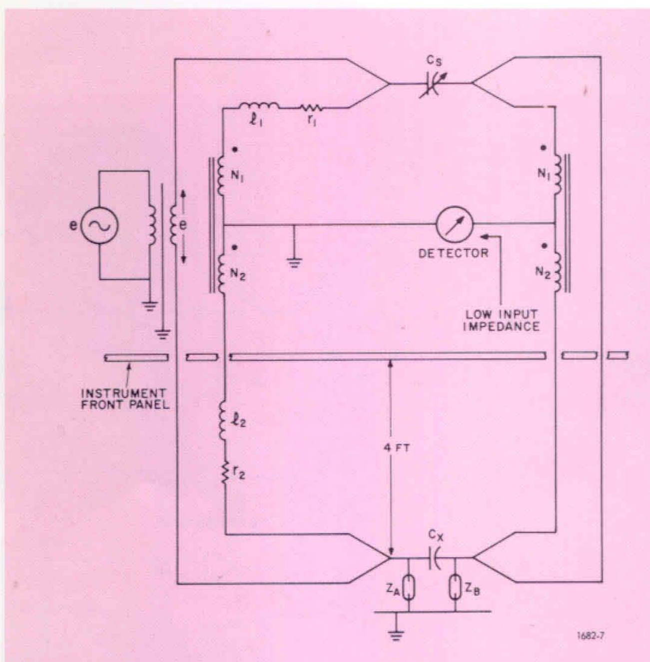


Figure 2. Simplified bridge circuit.

VERSATILE RESISTANCE BRIDGE

A self-contained resistance-measuring system with laboratory accuracy and production speed

The demand for reliability in present complex electronic equipment emphasizes the need for large-volume resistor testing. Measurements of resistors to parts-per-million accuracy formerly were made on a Kelvin or Wheatstone bridge that required manual balancing. Today the old high-accuracy measurement techniques are too slow.

The need for faster measurements was recognized at General Radio more than a decade ago. As a result, the 1652-A Resistance Limit Bridge[1] was developed for use at GR and later offered for general sale because of its versatility. No bridge balance was required. Percentage deviation of the unknown resistor from an adjustable internal standard was indicated but accuracy was limited to 0.2%.

Continued demand for an improved economical resistor-testing device that offered greater versatility, speed, accuracy, and more automatic features than the 1652 was the incentive to develop our latest resistance limit bridge.[2]

[1] Hague, W. M., Jr., "Versatile Resistance Limit Bridge Doubles as Laboratory Standard," *GR Experimenter*, January 1952.

[2] Szpila, R. T., "A Resistance Deviation Bridge Utilizing a Photo Chopper DC Amplifier," MIT Master's Thesis, Electrical Engineering, June 1966.

Some Features

Improved features of the GR 1662 Resistance Limit Bridge, which supersedes the GR 1652-A Bridge, include:

- Resistance range from 1Ω to 111 MΩ.
- Comparison precision to 100 parts per million.
- Five deviation ranges from 0.3% to 30% full scale.
- Internal-resistance-standard limit of error better than 0.02%.
- Operating rate up to 4 measurements per second.
- Four-terminal Kelvin connections.
- Linear analog output voltage to

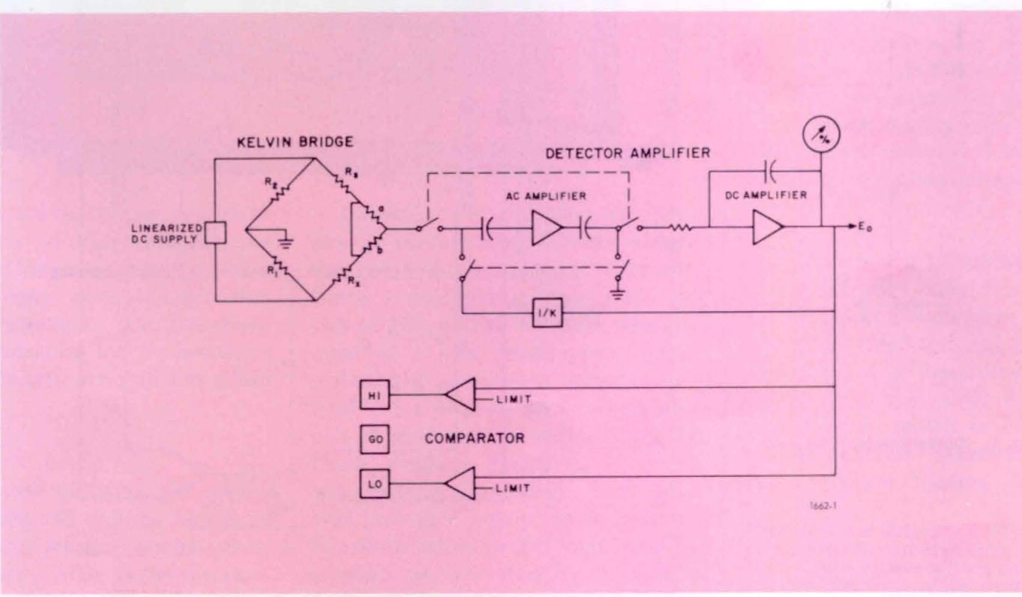


Figure 1. Block diagram of GR 1662 Bridge.

drive DVM's, limit comparators, dc recorders.

- HI-GO-LO limit indication by panel lights.
- Manually-preset or external-programmable limiting.
- Less than 12 mW dissipated in unknown (minimizes self-heating effect on low-power devices).

The block diagram of the GR 1662 bridge, Figure 1, shows the basic elements of the instrument: a dc generator, an active Kelvin bridge, a detector amplifier, and the HI, GO, LO limit circuitry.

The floating dc generator is guarded and shielded to reduce stray leakage paths to ground, thereby preserving bridge accuracy even at the range extremes. The Kelvin bridge consists of exceptionally stable and high-accuracy resistors adjusted to better than 0.01%. The bridge circuit is linearized by means of feedback to the dc source to measure the resistance deviation as a percent of the standard.

Inner Workings

When the ratio R_1/R_2 does not equal R_x/R_s, an unbalance-error signal results. Depending on the unbalance, this error voltage can be quite small. It must be amplified to a useful level so that it can be displayed on a deflection meter as the percent difference between the standard and the unknown resistor. A highly sensitive photo-chopper amplifier is used as the detector amplifier to provide high gain and to minimize drift and noise. The dc input is converted into ac by the photochopper. It is then amplified by a dc amplifier that is also used as a Miller integrator to provide an effective filter for the demodulated signal. The gain of the detector amplifier is stabilized by feedback around the over-all system. The deviation voltage is indicated on a zero-centered meter and by the analog-voltage output.

In addition to the percent-deviation meter readout, a HI-GO-LO indication is also available. The output voltage is fed to a set of analog comparators where it is compared to some preset voltage level to determine if the resistance measured by the bridge is higher, lower, or within the selected tolerance.

Typical Applications

A practical application illustrating the instrument's versatility is resistor sorting at production speed. The meter readout indicates the percentage deviation of the resistor under test from an adjustable internal or external standard. For manual sorting, resistors can be conveniently connected to the bridge by use of the 4-terminal-connected GR 1662-P1 Test Fixture. Completely automatic and faster sorting capability is possible when the bridge is paired with the GR 1782 Analog Limit Comparator[3] and with external handling and sorting equipment.

The GR 1662 is also suitable for use as a laboratory instrument to measure precision resistors to within 200 parts per million. Such precision can be achieved by nulling techniques, as in a conventional Kelvin bridge.

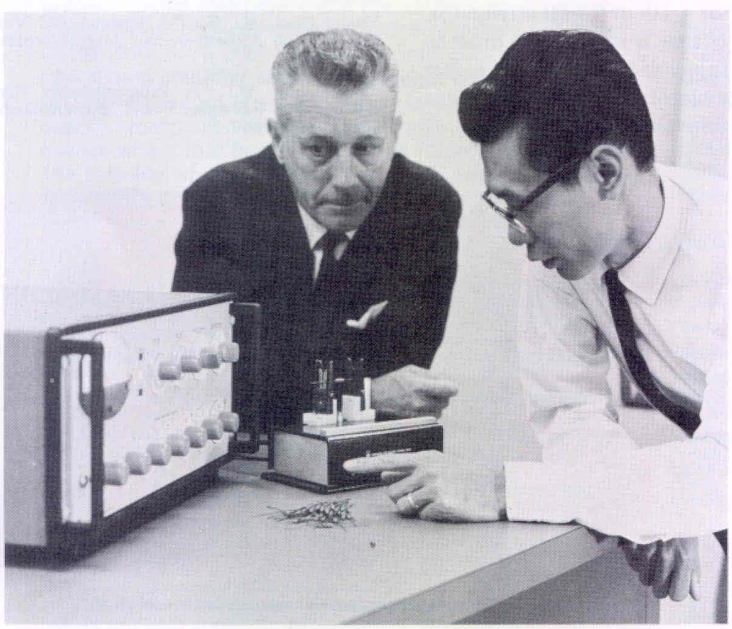
Other interesting applications are matching of resistors, temperature-coefficient measurement of resistors, and trimming of thin-film resistors by controlling production processes with the relay-equipped GR 1782 Analog Limit Comparator that provides four tolerance-limit settings. —R. K. Leong

[3] Leong, R. K., "Impedance Comparison Sprints Ahead," *GR Experimenter*, May/June 1969.

A brief biography of engineer R. K. Leong appeared in the May/June 1969 issue of the *GR Experimenter.*

The GR 1662 was developed by the author, with contributions by R. G. Fulks and R. T. Szpila during the early phases of the development work.

Complete specifications for the GR 1662 are included as a tear sheet at the back of this issue.



Engineer Bob Leong demonstrates GR 1662 bridge to Editor White.

WIDEBAND 20-dB/RANGE AC MILLIVOLTMETER

The ac millivoltmeter is a fundamental tool in electronic measurements. Although many instruments of similar capability are already in existance, the new millivoltmeter from GR makes a unique contribution to ac measurements.

Basically, the GR 1808 is an average-reading voltmeter calibrated to indicate the rms-value of sine waves. But what sets this voltmeter apart from others is its 10-Hz to 10-MHz bandwidth coupled with a 20-dB dynamic range per range. This wide dynamic range makes possible a single voltage scale that, in turn, avoids confusion in reading the meter. The single scale is not only convenient for many amplifier response measurements but is also necessary in some automatic testing or calibration set-ups.

Applications

Because the millivoltmeter is a general-purpose, laboratory- and production-type voltmeter it is difficult to describe a particular application as a "typical" application. Here are some illustrations of interesting applications:

- Most operational amplifiers have the open-loop frequency-response curve shown in Figure 1. Quite often it is desired to know the frequency, f_2, where the second breakpoint occurs in order to maximize the design stability of the amplifier.

  The GR 1808, with its 10-MHz bandwidth, is well suited for this type of measurement.

- The wide dynamic range and wide bandwidth of the GR 1808 encourage its use for attenuator calibration or testing. For a 10- or 20-dB attenuator, no range change is necessary in order to read the input and output. For higher value attenuators, minimum of range changing is involved.

- Frequently, we wish to make ac measurements with higher resolution than the specified accuracy of the available instruments. For example, in tests of the stability of an amplifier with temperature, the absolute value of a measurement is not so important as the change in the measurement as a function of temperature.

  The dc output from the 1808 may be coupled into a GR 1807 DC Microvoltmeter/Nanoammeter[1] to form such a high resolution system. The GR 1807 has an interpolation feature that will enable the user to read the dc output with 0.1% resolution. The system shown in Figure 2 will increase resolution of ac voltages approximately ten times compared with the GR 1808 meter reading. It is also important to note that the dc-voltage output of the GR 1808 can be used to drive a GR 1522 Recorder[2] for a permanent recording of data. The output of the GR 1807 can be connected to the recorder if high resolution recording is desired.

- An important application for the millivoltmeter is voltage measurement from accelerometers, strain gauges, microphones or other similar transducers. In general, such transducers can be reduced to an equivalent-voltage source in series with a capacitance (Figure 3). The voltage source is usually less than 100 millivolts and the capacitor C may be a

(cont. on page 11)

[1] Balekdjian, K. G., "A Unique DC Voltmeter," *GR Experimenter*, August-September, 1968.

[2] Basch, M. W., "A Programmable High-Speed DC Recorder," *GR Experimenter*, May/June 1969.

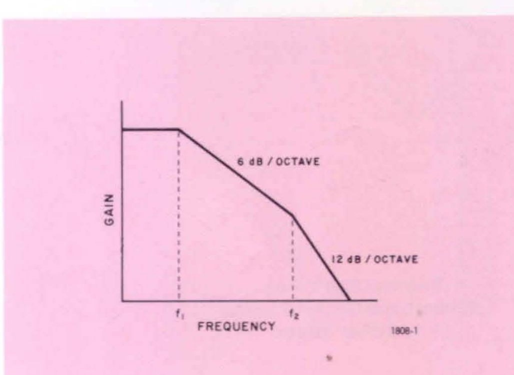


Figure 1. Typical open-loop frequency-response curve.

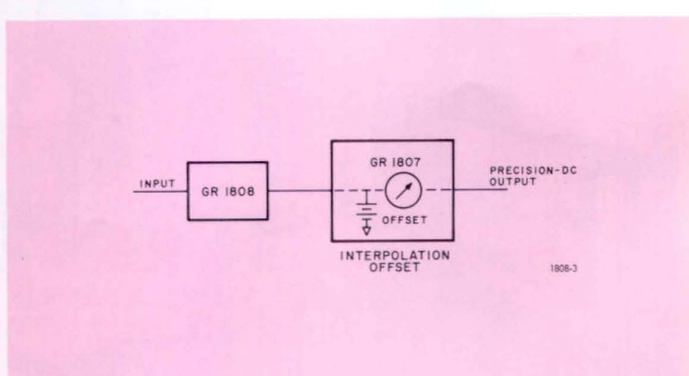


Figure 2. System for increased measurement resolution.

few hundred to a few thousand picofarads.

To measure the output of such a device with reasonable accuracy (1% - 5%), it is essential to have a voltmeter with very low input capacitance. The GR 1808 millivoltmeter with a Tektronix P6008 probe and GR 1808-P1 Probe Adaptor becomes an ideal combination for such measurements. The input capacitance of the probe will be approximately 7.5 pF and the sensitivity of the resulting combination will be 15 mV for full-scale deflection.

Theory of Operation

Figure 4 shows the block diagram of the GR 1808. The input buffer uses a field-effect transistor in order to achieve the high input impedance of the instrument. Both attenuator No. 1 and attenuator No. 2 are resistive type with capacitive frequency compensation. Attenuator No. 1 is used on the 150-V and 15-V ranges, and all switching is done by means of reed relays. This keeps high level ac signals from the sensitive detector circuits.

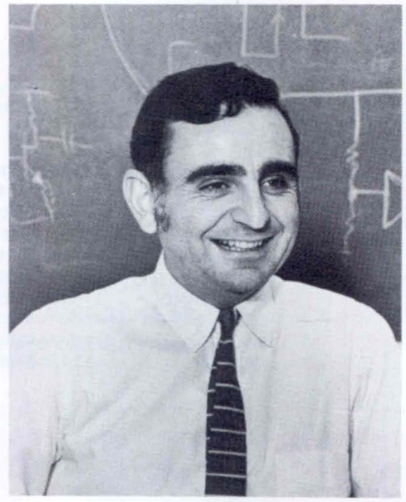
In order to achieve maximum stability, the gain of the 20-dB amplifier is never changed. Instead, attenuator No. 2 provides the proper signal levels for all ranges.

The heart of GR's new voltmeter is the ac-to-dc converter shown in Figure 5. Diodes CR1 and CR2 form a full-wave rectifier circuit. The dc-output voltage (read by the meter) is proportional to the difference between the rectified voltages V_1 and V_2. An important feature of this type of converter is the fact that nonlinear effects due to the diodes are eliminated from accuracy considerations, because the diodes are inside the feedback loop of amplifier A. The unbalance-leakage currents of these diodes are small enough to justify neglecting their effect since only the unbalance leakage current enters the accuracy considerations.

The key to the wideband and wide dynamic range of this converter is amplifier A. It has high open-loop voltage gain even at 10 MHz, to provide sharp rise and fall times at its output. Since most diodes will require 0.3 to 0.5 volt to draw at least 0.1-mA current, sharp rise/fall times are especially necessary at low level and high frequencies if errors are to be avoided.

The GR 1808 AC Millivoltmeter is not just another ac voltmeter; it is a distinct contribution to this basic branch of electronic measurements.

–K. G. Balekdjian



K. G. Balekdjian is a member of the GR Component and Network Testing Group. He received his SB and SM Degrees from Massachusetts Institute of Technology in 1955 and 1957 respectively and joined GR as a development engineer in 1964. George is a member of IEEE, Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

Complete specifications for the GR 1808 are included as a tear sheet at the back of this issue.

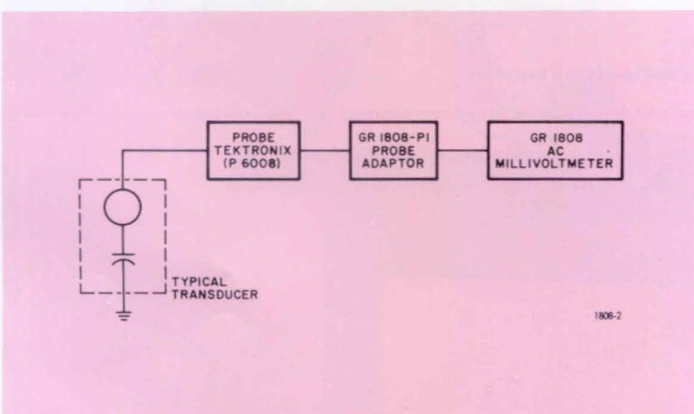


Figure 3. Suggested measurement system for transducer response.

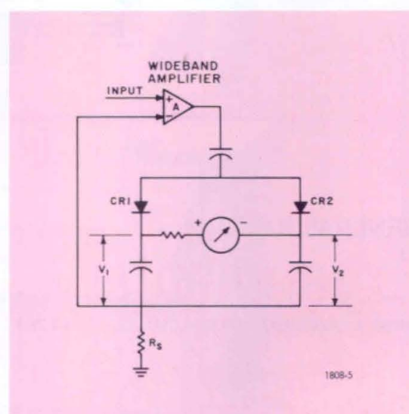


Figure 5. Schematic of ac-to-dc converter.

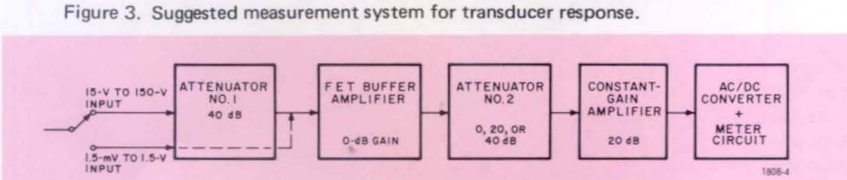


Figure 4. Block diagram of GR 1808 Millivoltmeter.

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