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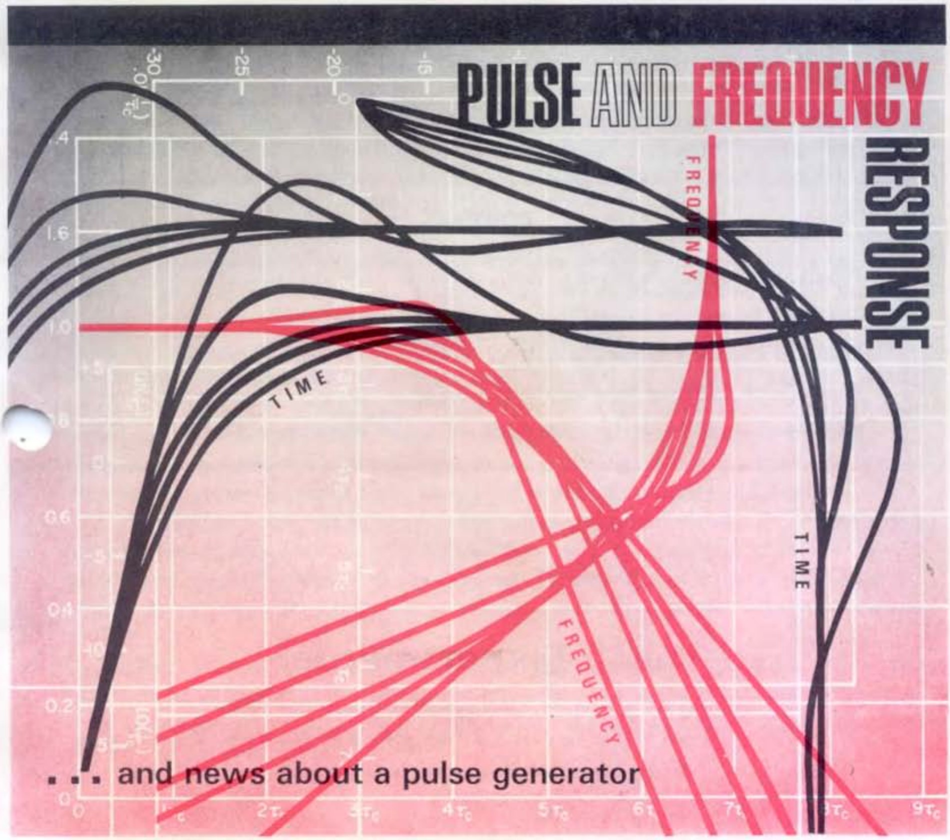

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- Two new Variac® autotransformers

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the GR Experimenter

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PULSE AND FREQUENCY RESPONSE

Some Useful Relations

by J. K. Skilling

If a fast pulse is applied to a wideband circuit, and the output has a rise time of 50 nanoseconds and no overshoot, the upper cutoff frequency is approximately 7 MHz. If the same circuit exhibits 25 percent droop in its response to a 1-second pulse, its low-frequency cutoff is 0.05 Hz.

It is often useful to be able to infer the frequency response of a device or circuit from measurements of pulse response. For one thing, instrumentation for pulse work is generally simpler than that for sine-wave measurements. To measure directly the two cutoff frequencies in the above example, one would probably need two sine-wave signal generators to cover the range from 0.01 Hz to 10 MHz. Measurements at several frequencies would have to be taken. A sweeper would be more convenient — although it would give no phase information — but would entail an even larger commitment to instrumentation. On the other hand an economical, general-purpose pulse generator and oscilloscope are the only instruments needed for the rapid pulse testing of wideband circuits and devices.

And there are other advantages to the use of pulse techniques. Pulse excitation allows independent control over signal level and average power in the device under test. Also, the use of an oscilloscope to view the output immediately uncovers gross distortion, clipping, oscillation, etc that might otherwise escape notice.

On the debit side, the present state of instrumentation does not allow the observation of waveforms with the accuracy that characterizes sine-wave testing. As a result, pulse tests give less accurate information about frequency response than do direct measurements.

MODELS

Correlation of pulse and sine-wave responses is generally done, as most active-circuit calculations are done, with models. One calculates both pulse-response waveforms and frequency-response curves for a given model circuit with given component values. The assumption is that if the observed waveform from a real device is the same as a waveform that one has calculated for the model, then the device's frequency response will be the same as that calculated for the model. This is born out, at least to a good approximation, by most common wideband, linear devices and circuits that have reasonably smooth rolloffs at both low and high frequencies.

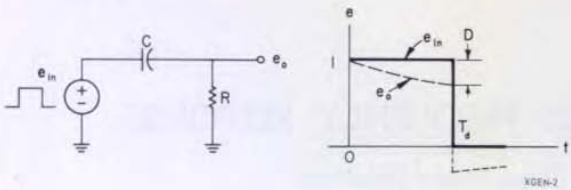


Figure 1. Low-frequency model and input and output waveforms.

LOW-FREQUENCY RESPONSE

The low-frequency response of the circuit governs the amount of "droop" in the output pulse. It is usually adequate to choose a simple RC high-pass circuit as a model for low-frequency behavior, as shown in Figure 1. If the amount of droop is less than 30 percent, the low-frequency cutoff frequency f_1 is given in terms of the droop D (percent) and the pulse-duration time T_d by the approximate formula

$$f_1 \doteq \frac{0.159}{T_d} \cdot \frac{D}{100}$$

In practice, one might adjust the pulse duration until a droop of 25 percent is observed. When $D = 25\%$, the low-frequency cutoff can be calculated with a formula that is a little more accurate than the one given above:

$$f_1 \doteq \frac{0.0456}{T_d} \qquad (\text{droop} = 25\%)$$

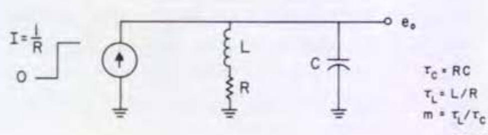
HIGH-FREQUENCY RESPONSE

The high-frequency behavior of the circuit governs the way in which the output waveform responds to an input step. A model suitable for high frequencies is somewhat more complicated than the low-frequency model we have just discussed. The one usually chosen is the series-parallel circuit shown in Figure 2.[1] The source is a current step whose magnitude I is such that $IR = 1$. The ratio m measures the amount of peaking or, inversely, the amount of damping. (It is equal to the square of the circuit's Q at the resonant frequency of L and C.)

Figure 3 shows calculated output waveforms for various values of m. Note that the time is measured in units of τ_C. Also shown in Figure 3 are the definitions of two parameters having to do with the circuit's step response: the overshoot r and rise time t_r.

[1] Valley and Wallman, *Vacuum Tube Amplifiers*, M.I.T. Radiation Laboratory Series, No. 18, Boston Technical Publishers, Inc., Lexington, Mass., 1964.

Figure 2. High-frequency model. The source is a current step of magnitude *I*, and the circuit elements are in a series-parallel configuration.



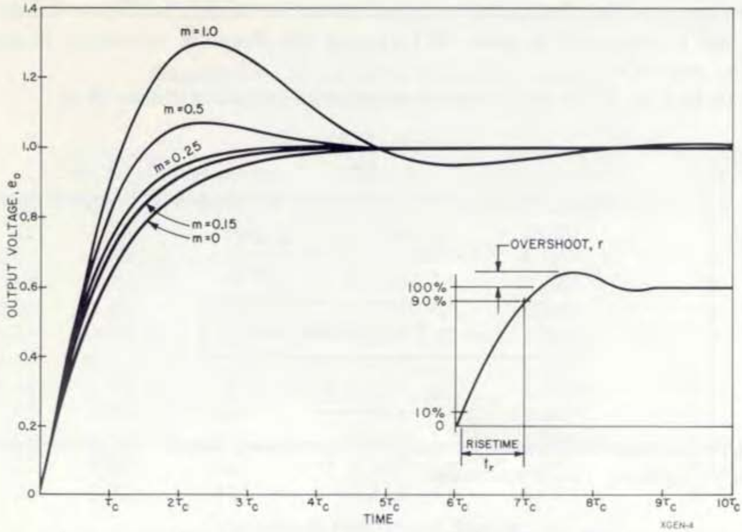


Figure 3. Step response of the series-parallel high-frequency model of Figure 2. The time axis is set off in units of τ_c and the waveforms are functions only of *m*. The illustration in the lower right corner shows how overshoot and risetime are defined.

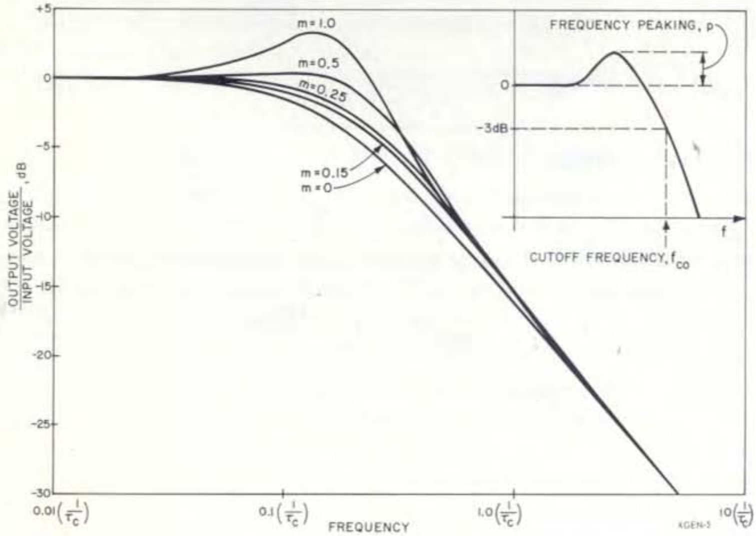


Figure 4. High-frequency response of the series-parallel model of Figure 2. These curves are the companions of those in Figure 3. The frequency axis is marked off in units of $1/\tau_c$, and the shapes of the curves, like the shapes of those in Figure 3, depend only on *m*. The amount of peaking *p* (dB) and the cutoff frequency f_{co} are defined by the sketch in the upper right corner.

The corresponding frequency-response curves are shown in Figure 4. Here the frequency is measured in units of $1/\tau_C$ and the shape of the curve is again a function only of m.

Let us look at a few simple relations for representative values of m.

$\underline{m = 0}$

There is no inductance, and the model reduces to the simple RC low-pass network.

$$t_r = 2.2\,\tau_C, \qquad f_{co} = \frac{0.159}{\tau_C}$$

$$\boxed{t_r f_{co} = 0.35 \text{ (golden rule)}}$$

$\underline{0 < m < 0.25}$

Now there is some inductance present. As m increases, the rise time decreases and the high-frequency cutoff increases.

$\underline{m = 0.25 \text{ (critical damping)}}$

This is the condition for the shortest rise time without overshoot (although one can do better with a more complicated "peaking circuit"[2]).

$$t_r = 1.53\tau_C, \qquad f_{co} = \frac{0.225}{\tau_C}$$

$$\boxed{t_r f_{co} = 0.344}$$

The golden rule survives.

$\underline{0.25 < m \leq 0.50}$

The rise time continues to decrease and the cutoff frequency continues to increase, but overshoot starts to build up and so does frequency peaking. At $m = 0.50$,

$$t_r = 1.12\tau_C, \qquad f_{co} = \frac{0.286}{\tau_C}$$

$$\boxed{t_r f_{co} = 0.32}$$

The golden rule is still barely tarnished.

$\underline{m > 0.50}$

Rise time continues to decrease, but overshoot becomes very large and reproduction of the pulse waveform is consequently poor. The cutoff frequency con-

[2] See Valley and Wallman, *op. cit.*, p 75.

Table 1
Characteristic quantities for the series-parallel high-frequency model

m	normalized rise time t_r/τ_c	overshoot r	normalized cutoff frequency $f_{co} \cdot \tau_c$	rise-time-cutoff-frequency product $t_r \cdot f_{co}$	frequency peaking p
0	2.20	0%	0.159	0.350	0 dB
0.07	2.03	0	0.172	0.349	0
0.15	1.82	0	0.192	0.349	0
0.20	1.68	0	0.208	0.349	0
0.25	1.53	0	0.225	0.344	0
0.35	1.31	1.0	0.258	0.335	0
0.50	1.12	6.7	0.286	0.320	0.25
0.70	1.01	16.	0.294	0.297	1.5
1.0	0.940	30.	0.289	0.276	3.3
2.0	0.865	68.	0.268	0.232	7.8
5.0	0.825	148.	0.245	0.202	15.

tinues to increase until $m = 0.7$, where the bandwidth of the circuit is maximum. The frequency peaking at maximum bandwidth is 1.5 dB. For larger values of m, the model becomes a relatively high-Q resonant circuit with frequency peaking that is excessive for a wide-band device.

The relations discussed above are tabulated as a function of m in Table 1.

AN ALTERNATIVE HIGH-FREQUENCY MODEL

We can gain some confidence in the model-building method if we compare the above results with those for a different high-frequency model, the series circuit shown in Figure 5. The behavior of the series model is shown in Figures 6 and 7 and Table 2.

The response is similar to that of the series-parallel model. The improvement in rise time and cutoff frequency with increasing m is not as marked — the improvement is 10 percent less at $m = 0.25$, for example. The cutoff frequency maximizes at a lower value of m than with the series-parallel model. But the most

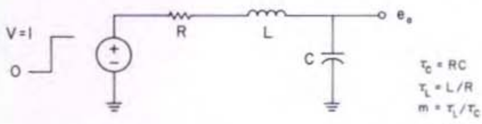


Figure 5. Another high-frequency model. The source in this circuit is a voltage step of unit magnitude and the circuit elements are in series.

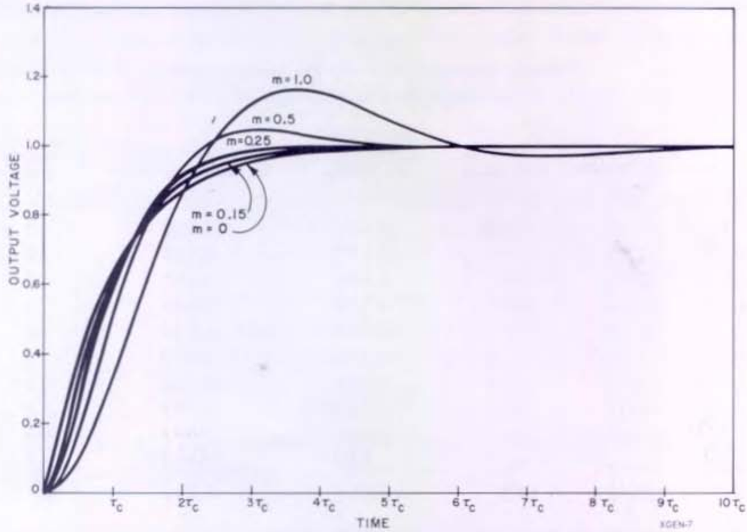


Figure 6. Step response of the series high-frequency model. The improvement in rise time with increasing *m* is a little less than in the series-parallel model, but otherwise these curves are very similar to those of Figure 3.

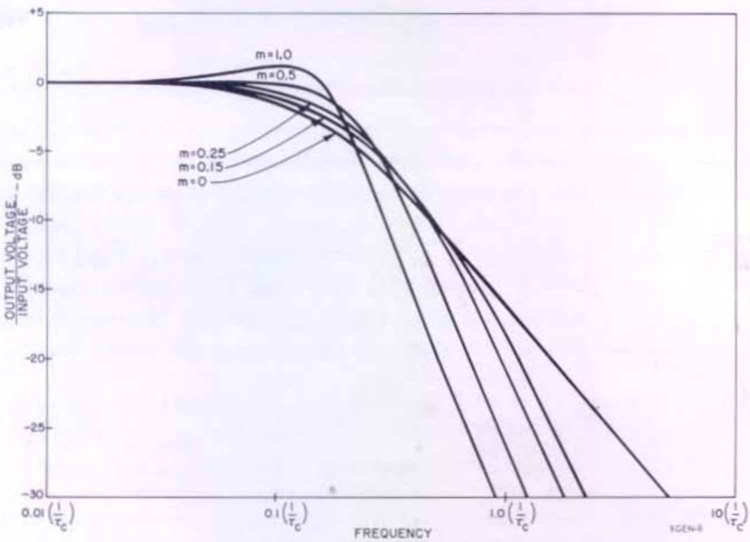


Figure 7. High-frequency response of the series model.

Table 2
Characteristic quantities for the series high-frequency model

m	normalized rise time t_r/τ_c	overshoot r	normalized cutoff frequency $f_{co} \cdot \tau_c$	rise-time-cutoff-frequency product $t_r \cdot f_{co}$	frequency peaking p
0	2.20	0%	0.159	0.350	0 dB
0.07	2.04	0	0.171	0.349	0
0.15	1.87	0	0.186	0.348	0
0.20	1.77	0	0.196	0.347	0
0.25	1.68	0	0.205	0.344	0
0.35	1.57	0.6	0.218	0.342	0
0.50	1.52	4.3	0.225	0.342	0
0.70	1.55	9.6	0.219	0.339	0.74
1.0	1.64	16.3	0.202	0.331	2.49
2.0	1.97	30.5	0.159	0.313	7.18
5.0	2.75	48.6	0.107	0.293	14.42

important thing to notice is that this circuit, too, obeys the golden rule: the rise-time-bandwidth product is very close to 0.35 for small m.

CASCADED STAGES

Cascaded stages with no overshoot have a combined rise time that is approximately the root-sum-square of the individual rise times:

$$t_{r\ combined} = \sqrt{t_{r1}^2 + t_{r2}^2 + t_{r3}^2 + \ldots}$$

This useful relation also allows one to calculate the effect of the measuring system on the rise time. For example, if a pulse generator and oscilloscope by themselves show a rise time of 20 nanoseconds, and if when they are used to measure the rise time of a device the result is 80 nanoseconds, the above relation gives a rise time of 77.5 nanoseconds for the device itself.

Bear in mind that the formula given above applies only in cases of no overshoot. Valley and Wallman[3] give the following rules for cascaded stages with overshoot: "For stages having very small overshoot (1 or 2%) the overshoot grows extremely slowly or not at all as the number of stages increases. . . . For stages having overshoots of about 5 to 10% the overshoot increases approximately as the square root of the number of stages, and the rise time increases substantially less rapidly than as the square root."

[3] *Op. cit.*, p 78.

CONCLUSIONS

The low-frequency cutoff can be determined from the measured amount of droop in the response to a long pulse. The prediction of high-frequency performance is a little more complicated because two cases have to be distinguished. In the overdamped zero-overshoot case ($0 \leq m \leq 0.25$) it is difficult to determine m from an observed pulse waveform. But it is surprising to note that the cutoff-frequency-rise-time product changes by less than 2 percent over the range $m = 0$ to $m = 0.25$. If the modeling procedure is reasonable, the prediction of cutoff frequency from rise time or vice versa anywhere in the overdamped region is accurate enough for engineering purposes.

In the underdamped case ($m > 0.25$), the cutoff-frequency-rise-time product depends upon m. But in this case the amount of overshoot, which one can measure with fair accuracy, gives a good estimate of m.

Acknowledgement

Computation of the data contained in the figures and tables was programmed by Bruce J. Jay.

BIBLIOGRAPHY

John G. Truxal, *Automatic Feedback Control System Synthesis*, McGraw-Hill Book Co. Inc., New York, 1955, p 37 ff.

Jacob Millman, Herbert Taub, *Pulse, Digital, and Switching Waveforms*, McGraw-Hill Book Co., New York, 1965, Chapter Two.

John J. D'Azzo, Constantine H. Houpis, *Feedback Control System Analysis and Synthesis*, McGraw-Hill Book Co., New York, 1966, p 80 ff.

Frederick Emmons Terman *et al*, *Electronic and Radio Engineering*, McGraw-Hill Book Co., New York, 1955, p 288 ff.

Truman S. Gray, *Applied Electronics*, John Wiley & Sons, Inc., New York, 1954, p 563 ff.

Lawrence Baker Arguimbau, *Vacuum-Tube Circuits*, John Wiley & Sons, Inc., New York, 1948, Chapter IV.

George E. Valley, Jr. and Henry Wallman, *Vacuum Tube Amplifiers*, Boston Technical Publishers, Inc., Lexington, Massachusetts, 1963, Chapter Two.

James K. Skilling received his BSEE degree from the University of California at Berkeley in 1953 and his MSEE degree from Johns Hopkins University in 1963.

Before joining General Radio in 1959, he held an engineering position at Douglas Aircraft and served as an electronics instructor at the United States Naval Academy.

At GR he has been primarily involved with the development of pulse equipment and digital system techniques. He is a member of IEEE, A.A.A.S. and SIGMA XI.

THE 1340 PULSER

Versatility and price make this the "best buy" of the general-purpose pulse generators.

When a new instrument improves on the performance of products already on the market, it represents progress; when it offers conveniences that the others don't, it is attractive; when at the same time its price meets or beats the competition, it makes news. So the GR 1340 Pulse Generator is news. It is a general-use pulse generator that offers a total combination of performance and features that is not surpassed even by high-priced generators.

0.2 Hz to 20 MHz

Repetition periods from 50 nanoseconds to 5 seconds will interest users in fields from high-speed circuit testing to seismological research. A single pulse for one-shot tests can be triggered manually from a front-panel pushbutton.

The 1340 also offers eight decades of pulse-duration times, from 25 nanoseconds to 2.5 seconds. The maximum duty ratio that can be attained varies from 70 to over 95 percent, depending on the combination of repetition-rate range and duration-time range that is in use. In case of an inadvertent setting that exceeds the allowable duty ratio, a panel lamp flashes to alert the operator that the output waveform may not be trustworthy. In any event there is no risk of a circuit overload if the duty-ratio limit is exceeded.

The repetition and duration controls each combine range switching and continuous adjustment in single easy-to-read dials. The repetition dial reads both prf and period, along with the correct units.

A convenience feature of the 1340 is the "squarewave" position of the pulse-duration control. In this setting the output consists of 50%-duty-ratio pulses, regardless of the prf. Square-wave testing with the 1340 is thus fast and simple, since there is no need to reset the duration when the prf is changed.

10-Volt Output, ± 1-Volt Offset

Both positive and negative pulses are available simultaneously at two low-reflection GR874® coaxial connectors. Each output is a high-impedance (about 1 kΩ) current generator with a maximum source current of 200 mA and a saturation voltage of better than 10 volts. Fifty-ohm resistors are switched across the outputs when reflectionless sources are required. With the resistors switched in, the 1340 can deliver 5-volt pulses to a 50-ohm load.

Another 1340 convenience feature is the output offset. Independently adjustable biases can be applied to the two outputs to offset the pulses up to ± 1 volt (or ± 20 mA). The offset feature eliminates the need for auxiliary equipment, for example, to ensure positive triggering in the presence of noise or small biases. It is a feature that will be of particular interest to those concerned with integrated-circuit testing.

The 1340's synchronizing trigger is a square wave, rather than the narrow pulse that is available at most pulse-generator sync outputs. The square wave has three principal advantages. 1) Even when the 1340 is being used to generate long pulses, the sync pulse is not too short to provide an effective trigger. 2) The square wave provides a stable trigger as the prf is changed, since its dc level does not vary. 3) By changing the scope's trigger level, one can synchronize the scope one-half period ahead of the pulse.

PPM, PDM, PAM

The period, duration, negative amplitude, and positive amplitude can each independently be linearly modulated by externally applied modulating voltages. Voltages between −0.5 and −5 volts sweep the period and duration over the decade ranges selected on the range switches. Voltages between 0 and +5 volts and 0 and −5 volts, respectively, vary the amplitudes of the positive- and negative-going pulses over the entire 10-volt range.

The 1340 has other modes of electronic control. A gating input permits an external switch closure or positive-going pulse to inhibit the 1340's period generator. Gating the oscillator means that a pulse that is already underway is not chopped by the gating signal. Whether one talks about modulation, sweeping, or remote programming, the 1340's electronic-control features mean greater usefulness.

A description of some of the 1340's novel circuitry follows the specifications.

SPECIFICATIONS

PULSE PERIOD (PRF)

Internally Generated: 50 ns to 5 s (20 MHz to 0.2 Hz) in 8 decade ranges. Single-pulse push button on panel.

Externally Controlled: 1 Hz to 20 MHz; triggers on any waveform of >3 V pk-pk. Input resistance approx 100 kΩ. Output pulse is started by negative-going transition. Period control acts as input trigger-level control in external mode.

OUTPUT-PULSE CHARACTERISTICS

Duration: 25 ns to 2.5 s in 8 decade ranges, or square wave.

Rise and Fall Times: 5 ns ±2 ns at 5 V, 50-Ω load, and 50-Ω source resistance.

Amplitude: Positive and negative ground-based pulses available simultaneously with independent amplitude and offset control. Source current continuously adjustable to at least 0.2 A (i.e., across 50-Ω load, 10 V from high source resistance or 5 V from 50-Ω source).

Offset: Continuously adjustable source current from −20 to +20 mA.

Source Resistance: 50 Ω, or high (approx 1 kΩ) for 50-Ω loads.

Distortion: Preshoot, overshoot, ringing, etc, <0.5 V (5% of max output).

Duty Ratio: Duty ratios of over 70% can be obtained on all ranges except decreasing to approx 50% at 50-ns period in 50-to-500-ns range.

SYNCHRONIZING PULSE

Waveform: Square wave. Negative transition precedes start of output pulse by approx 35 ns; positive transition can be used for half-period pretriggering.

Amplitude: 2.5-V pk-pk positive square wave behind 500-Ω source impedance.

MODULATION AND GATING

Modulation: Period and duration are linearly controllable by an external voltage between −0.5 and −5.0 V. Amplitude of the positive-pulse output is linearly controllable by an external voltage of 0 to +5 V, the negative-pulse output by 0 to −5 V. Period and duration are modulatable over the decade range set by range switches; amplitude can be modulated over its full range. Amplitude modulation can be used for noncoherent gating of output pulse.

Gating: An impedance of ≤600 Ω to ground inhibits output; +4 to +8 V allows normal output; 1340's can be used to gate 1340's.

GENERAL

Power Required: 100 to 125 or 200 to 250 V, 50 to 400 Hz, 30 W.

Accessories Supplied: Spare fuses, power cord.

Accessories Available: GR874® coaxial components, attenuators, terminations, tees, etc.

Mounting: Convertible-Bench Cabinet.

Dimensions (width × height × depth): Bench 8½ × 5⅝ × 13 in. (220 × 145 × 330 mm); rack, 19 × 5¼ × 11¼ in. (485 × 135 × 290 mm).

Weight: Net, 9¼ lb (4.2 kg); shipping, 13 lb (6.0 kg).

Catalog Number	*Description*	*Price in USA*
	1340 Pulse Generator	
1340-9700	Bench Mount	$395.00
1340-9701	Rack Mount	417.00

Yech!

NOVEL VOLTAGE-VARIABLE SCHMITT CIRCUIT

The 1340's period generator makes use of the voltage-variable Schmitt circuit, shown in Figure 1.

Q1 and *Q2* are regeneratively connected to form a Schmitt-type multivibrator. The common emitter current supplied by R_E flows entirely in either *Q1* or *Q2* when the circuit is in one of its two stable states. Consider the case where *Q2* is conducting (on) and *Q1* and *Q3* are off. Under these conditions the voltage divider *R1*, *R2*, and *R3* sets a voltage, E_U, at the base of *Q2*. In order to keep *Q1* and *Q3* off, their base voltages must be more negative than E_U.

Transfer of conduction to *Q1* occurs when the input voltage E_{IN} on the base of *Q1* is raised to a value close to E_U, at which point *Q1* comes on enough to start the turn-off of *Q2*. Regenerative transfer of conduction to *Q1* results,

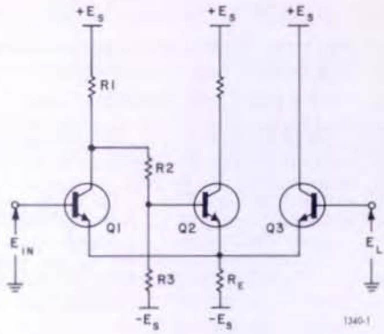


Figure 1. Simplified schematic diagram of the voltage-variable Schmitt circuit, part of the 1340's period generator.

and, with *Q1* on, the base of *Q2* is driven more negative than the base of *Q3*.

In order to transfer the conduction back to *Q2*, E_{IN} is now lowered to a value near E_L. At this point *Q3* begins to draw current, and this reduces the current in *Q1*, causing *Q2* to begin to turn on. The regenerative interconnection of *Q1* and *Q2* again causes the emitter current to shift, this time back to *Q2*.

Conduction is therefore shifted back and forth between *Q1* and *Q2* by cycling E_{IN} first up to an upper trigger level E_U and then down to a lower trigger evel near E_L. The significant advantage of this circuit over other multivibrators is that the lower trigger level is controlled in a very simple manner by an externally supplied voltage E_L. The lower trigger level is, in fact, almost exactly equal to E_L.

Two more transistors, *Q4* and *Q5*, turn the voltage-variable Schmitt circuit into the complete voltage-variable period generator, Figure 2. *Q5* is a constant-current source, supplying a current I to the timing capacitor C. *Q4* is a switched constant-current source, supplying no current when off and $2I$ when on.

When *Q1* is on, both *Q2* and *Q4* are off and the current into C is $-I$; the

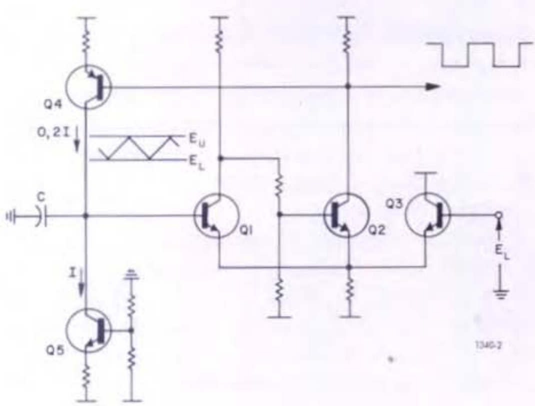


Figure 2. Simplified schematic of the complete period generator. Voltage control of the period is achieved by varying E_L, which governs the capacitor-voltage swing.

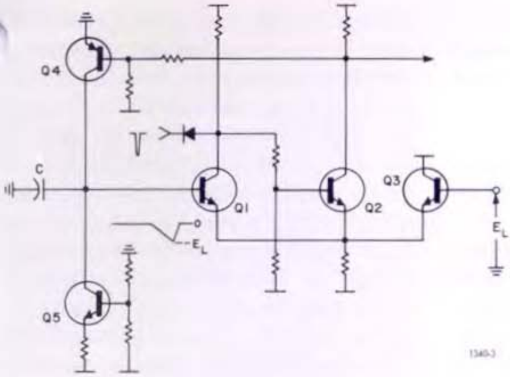


Figure 3. The pulse-duration generator, simplified. The control voltage E_L determines how long the circuit stays in its unstable state.

voltage across C drops linearly with time. When this voltage reaches E_L, the voltage-variable Schmitt shifts to the *Q1*-off, *Q2*-on state, and *Q4* comes on. The net capacitor current is now $+I$, and the capacitor voltage increases linearly. When the capacitor voltage reaches E_U, the Schmitt shifts states back to *Q1*-on, *Q2*-off. In this fashion the capacitor alternately charges and discharges so that the capacitor voltage is a triangular wave, and a square wave is developed at the collector of *Q2*.

Voltage control of the period is effected by varying E_L, which controls the peak-to-peak amplitude of the capacitor voltage. Since the charging currents are constant, the period of the square wave is directly proportional to the peak-to-peak swing of the capacitor voltage.

The duration generator uses the voltage-variable Schmitt circuit in the monostable configuration shown in Figure 3. In this circuit *Q5* is again a constant-current generator. In the stable state, *Q2* is on; *Q2* holds *Q4* on, and the capacitor voltage is clamped to ground. In monostable operation, the base of *Q2* is slightly above ground when it is on; thus the capacitor voltage, which cannot go positive, does not turn *Q2* off. A negative trigger, applied to the collector of *Q1*, turns off *Q2*, and *Q4* turns off with *Q2*. The constant-current source now charges C, and the capacitor voltage starts a negative climb. When it reaches E_L, the

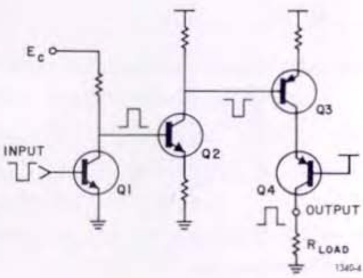


Figure 4. Simplified diagram of the positive-pulse output circuit. The two stages following *Q1* are nonsaturating, so that E_c controls the amplitude of the output pulse.

circuit reverts to the stable $Q1$-off, $Q2$-on state. The duration of the output pulse is determined by the amplitude of the voltage change on the timing capacitor and therefore is linearly proportional to the control voltage E_L.

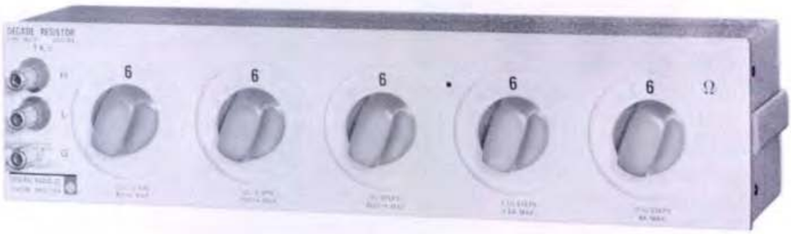
The output circuit for the positive pulse is shown in Figure 4.

The first transistor $Q1$ is normally saturated and is turned off by the input pulse. The amplitude of the pulse at the collector of $Q1$ is changed by adjusting the supply voltage E_c of this stage. $Q2$ is a nonsaturating amplifier that provides current gain to drive the output stage. The cascode (grounded-emitter driving grounded-base) output stage is also nonsaturating and has a very fast rise time. In order to handle the high output current and yet maintain high switching speed, $Q3$ and $Q4$ are each a pair of parallel transistors. All the amplifier stages following $Q1$ are nonsaturating so that the output amplitude is proportional to the control voltage E_c. This allows voltage control of the pulse amplitude.

The negative-pulse amplifier circuit is the same as the positive amplifier except that negative supply voltages and complementary transistor types are used.

— J. K. Skilling

IMPROVED DECADE RESISTORS



The GR line of precision decade resistors has a new look, better performance "under the hood," and a new type number.

General Radio decade resistors have been constantly improved since they were first introduced in 1917. Often these changes occurred as new techniques permitted, unheralded by outward appearance changes or new type-number designations. A review of published specifications, however, would reveal the steady improvements that have taken place. Most recently, both appearance and performance have gotten better, in recognition of which the precision 1432 line has been renamed the 1433. Improvements represented by this new number include:

In-line digital readout; reading errors are far less likely.

Models with from four to seven dials give required resolution.

All models rack mountable.

Better accuracy, ±0.02% two-year accuracy for high values.

"Over-all" as well as "incremental" accuracy specifications now given.

Better ac performance at high resistances, owing to reduced stray capacitance.

Rear-panel terminals for system installation.

"Over-all" accuracy versus "incremental" accuracy

Historically, GR has specified the accuracy of each separate resistor in a decade box rather than that of the total resistance. This practice has reflected the fact that each resistor is individually adjusted. We called this specification "accuracy of resistance increment" because the difference in value between any two positions of the same switch has the same accuracy as a resistor on that switch. Knowledge of the incremental accuracy is needed in many uses, and we are not abandoning it; but more often the absolute accuracy of any given setting of the box — the "over-all" accuracy — is desired.

If all the resistors in a box had the same accuracy specification, any combination of them would have this same specification. Lower-value resistors, however, cannot be adjusted as easily and generally are not as stable as higher-value resistors. So our over-all specification has two terms: ± (0.02% + 2 mΩ).

A single over-all specification is intended to apply to any and all possible settings of the decade box, but there are obviously just too many to permit testing every one. Instead, we adjust the lower-value units to much tighter tolerances than the published ± 2 mΩ, and test the assembled boxes at carefully selected values with limits substantially closer than the catalog specifications. To substantiate the validity of this method, 300 boxes were "made" on a computer under the assumption of a very poor tolerance distribution. These were first computer "tested" against our laboratory specifications and then computer "measured" at *every* setting. No box that passed the simulated laboratory test was in error by more than 9/10 of the catalog specification at *any* setting.

Such testing, whether real or simulated, checks only the *initial* tolerance of the decade box. Confidence in the two-year accuracy of GR 1433's is assured by stability records of the resistors and by the use of exceptionally good switches.

Switches

Decade-box accuracy depends on the resistors and on the switches. In the over-all specification, switch resistance plays a particularly important part. Commercial rotary switches with solid-silver contacts, such as we use in many instruments, can vary 1 milliohm or more in contact resistance. Obviously, a five-decade box can display a variation of 5 milliohms or more, which would have to be accounted for in the specification. The GR-designed and manufactured switches in our Type 1433 decade boxes have a low and very repeatable contact resistance. Critical contacts (on lower decades and zero settings) use multiple-leaf wipers

The use of multi-leaf wipers on silver-overlaid studs assures a contact resistance that varies by less than 100 microhms.

against silver-overlaid studs (see photo). As a result, contact resistance changes by no more than 100 microhms and hence has but slight effect upon the specification. Our less expensive line, the 1434, uses commercial switches with double contacts to give a ± 0.05% ± 5 mΩ over-all specification.

Resistors

For added dimensional stability, the 1-ohm through 10-kilohm units are now wound on ceramic rather than fiber-glass or mica cards as before. While the 100-kilohm units are unchanged, the 1-megohm units are now wound on bobbins.

For years, we've held that flat, single-layer card resistors are far superior to multi-layer bobbin types, as the latter have much higher inductance and shunt capacitance that degrade their performance in ac operation. Indeed, we still feel this is true, but only for single resistors. When 1-megohm cards are wired together in a decade box, the capacitances between cards and from the cards to the box become the main cause of reduced ac performance. These capacitances are lower when bobbins are used, and the ac performance is better in spite of the higher individual shunt capacitance. We have now swallowed our pride and learned to wind precision bobbin resistors, for, besides being better for decade boxes, they're a lot less costly to make and actually permit lower prices on some decades.

Separate Decades

The improved accuracy of the 1433's is available in the individual Type 510 Decades, which are sold separately. These units are ideal for assembly into production-test instruments, bridges, and other experimental or permanent equipment where only one or two decades are needed. The 510's are completely shielded, both mechanically and electrically, by an aluminum housing, and a knob and dial plate are supplied with each unit.

Decade resistors are difficult to specify adequately, but it is our intention to describe the 1433's performance in the specifications below as completely and truthfully as possible. A discussion of some of the terms we use and why we use them is given in the October, 1965 *Experimenter*, and further information on ac performance is given in the instruction sheet.

— H. P. Hall

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

Acknowledgement

Mr. Walter J. Bastanier is responsible for improvements in the resistors that have made possible the new, tighter specifications.

SPECIFICATIONS

Long-Term Accuracy: Our two-year warranty applies to the tolerances given below unless the resistor is damaged by excessive current. These tolerances apply for low-current measurement at dc or low-frequency ac (see below).

Over-all Accuracy: The resistance difference between that at any setting and at the zero setting is equal to the indicated value $\pm(0.02\% + 2\ m\Omega)$.

Incremental Accuracy: See table. This is the accuracy of the change in resistance between any two settings on the same dial.

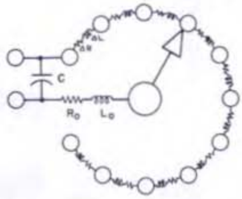
Max Current: The max current for each decade is given in the table below and also appears on the panel of each decade box and on the dial plate of each decade resistance unit.

Frequency Characteristic: The accompanying plot shows the max percentage change in effective series resistance, as a function of frequency for the individual decade units. For low-resistance decades the error is due almost entirely to skin effect and is independent of switch setting, while for the high-resistance units the error is due almost entirely to the shunt capacitance and its losses and is approx proportional to the square of the resistance setting.

The high-resistance decades (510-E, -F, -G, and -H) are very commonly used as parallel resistance elements in resonant circuits, in which the shunt capacitance of the decades becomes part of the tuning capacitance. The parallel resistance changes by only a fraction (between a tenth and a hundredth) of the series-resistance change, depending on frequency and the insulating material in the switch.

Characteristics of the 1433's are similar to those of the individual 510's modified by the increased series inductance, L_o, and shunt capacitance, C, due to the wiring and the presence of more than one decade in the assembly. At total resistance settings of approx 1000 ohms or less, the frequency characteristics of any of these decade resistors are substantially the same as those shown for the 510's. At higher settings, shunt capacitance becomes the controlling factor, and the effective value of this capacitance depends upon the settings of the individual decades.

Typical Values of R_o, L_o, and C for the Decade Resistors:

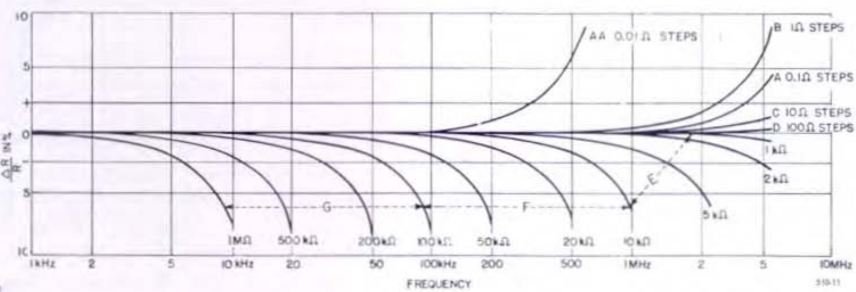


Equivalent circuit of a resistance decade, showing location and nature of residual impedances.

Zero Resistance (R_o): 0.001 Ω per dial at dc; 0.04 Ω per dial at 1 MHz; proportional to square root of frequency at all frequencies above 100 kHz.

Zero Inductance (L_o): 0.1 μH per dial + 0.2 μH.

Effective Shunt Capacitance (C): This value is determined largely by the highest decade in use. With the low terminal connected to the shield, a value of 15 to 10 pF per decade may be assumed, counting decades down from the highest. Thus, if the third decade from the top is the highest resistance decade in circuit (i.e., not set at zero), the shunting terminal capacitance is 45 to 30 pF. If the highest decade in the assembly is in use, the effective capacitance is 15 to 10 pF, regardless of the settings of the lower-resistance decades.



Maximum percentage change in series resistance as a function of frequency for Type 510 Decade-Resistance Units.

SPECIFICATIONS (continued)

Temperature Coefficient of Resistance: Less than ±10 ppm per degree C for values above 100 Ω and < ±20 ppm per degree C for 100 Ω and below, at room temperatures. For the 1433's the box wiring will increase the over-all temperature coefficient of the 0.1- and 0.01-Ω decades.

Switches: Quadruple-leaf brushes bear on lubricated contact studs of ⅜-in. diameter in such a manner as to avoid cutting but yet give a good wiping action. A ball-on-cam detent is provided. There are eleven contact points (0 to 10 inclusive). The switch resistance is less than 0.0005 Ω. The effective capacitance is of the order of 5 pF, with a dissipation factor of 0.06 at 1 kHz for the standard cellulose-filled molded phenolic switch form and 0.01 on the mica-filled phenolic form used in the 510-G and 510-H units.

Max Voltage to Case: 2000 V pk.

Type 1433 Decade Resistors

Catalog Number								*Price in USA*	
Bench	*Rack*	*Type*	*Total Ohms*	*Ohms per Step*	*No. of Dials*	*Type 510 Decades Used*		*Bench*	*Rack*
1433-9700	1433-9701	**1433-U**	111.1	0.01	4	AA, A, B, C		**$120.00**	**$128.00**
1433-9702	1433-9703	**1433-K**	1111	0.1	4	A, B, C, D		**122.00**	**130.00**
1433-9704	1433-9705	**1433-J**	11,110	1	4	B, C, D, E		**125.00**	**133.00**
1433-9706	1433-9707	**1433-L**	111,100	10	4	C, D, E, F		**120.00**	**128.00**
1433-9708	1433-9709	**1433-Q**	1,111,000	100	4	D, E, F, G		**154.00**	**162.00**
1433-9710	1433-9711	**1433-T**	1111.1	0.01	5	AA, A, B, C, D		**146.50**	**154.50**
1433-9712	1433-9713	**1433-N**	11,111	0.1	5	A, B, C, D, E		**144.00**	**152.00**
1433-9714	1433-9715	**1433-M**	111,110	1	5	B, C, D, E, F		**147.50**	**155.50**
1433-9716	1433-9717	**1433-P**	1,111,100	10	5	C, D, E, F, G		**182.50**	**190.50**
1433-9718	1433-9719	**1433-Y**	11,111,000	100	5	D, E, F, G, H		**247.50**	**255.[illegible]**
1433-9720	1433-9721	**1433-W**	11,111.1	0.01	6	AA, A, B, C, D, E		**168.50**	**176.50**
1433-9722	1433-9723	**1433-X**	111,111	0.1	6	A, B, C, D, E, F		**166.50**	**174.50**
1433-9724	1433-9725	**1433-B**	1,111,110	1	6	B, C, D, E, F, G		**210.00**	**218.00**
1433-9726	1433-9728	**1433-Z**	11,111,100	10	6	C, D, E, F, G, H		**276.00**	**284.00**
1433-9729	1433-9730	**1433-F**	111,111.1	0.01	7	AA, A, B, C, D, E, F		**191.00**	**199.00**
1433-9731	1433-9732	**1433-G**	1,111,111	0.1	7	A, B, C, D, E, F, G		**229.00**	**237.00**
1433-9733	1433-9734	**1433-H**	11,111,110	1	7	B, C, D, E, F, G, H		**303.50**	**311.50**

Type 510 Decade-Resistance Units

Catalog Number	*Type*	*Total Resistance Ohms*	*Resistance Per Step (ΔR)Ohms*	*Accuracy of Resistance Increments*	*Max Current 40° C Rise*	*Power Per Step Watts*	ΔL *μH*	C^{**} *pF*	L_o *μH*	*Price in USA*
0510-9806	**510-AA**	0.1	0.01	±2%	4 A	0.16	0.01	7.7-4.5	0.023	**$23.00**
0510-9701	**510-A**	1	0.1	±0.4%	1.6 A	0.25	0.014	7.7-4.5	0.023	**17.50**
0510-9702	**510-B**	10	1	±0.1%	800 mA	0.6	0.056	7.7-4.5	0.023	**26.00**
0510-9703	**510-C**	100	10	±0.04%	250 mA	0.6	0.11	7.7-4.5	0.023	**27.00**
0510-9704	**510-D**	1000	100	±0.02%	80 mA	0.6	0.29	7.7-4.5	0.023	**25.00**
0510-9705	**510-E**	10,000	1000	±0.02%	23 mA	0.5	13	7.7-4.5	0.023	**20.50**
0510-9706	**510-F**	100,000	10,000	±0.02%	7 mA	0.5	70	7.7-4.5	0.023	**21.00**
0510-9707	**510-G**	1,000,000	100,000	±0.02%	2.3 mA	0.5	—	7.7-4.5	0.023	**61.00**
0510-9708	**510-H**	10,000,000	1,000,000	±0.02%	0.7* mA	0.5	—	7.5-4.5	0.023	**92.00**
0510-9604	**510-P4**	Switch only	(Black Phenolic Frame)							**11.00**
0510-9511	**510-P4L**	Switch only	(Low-Loss Phenolic Frame)							**12.00**

* Or a max of 4000 V, pk.
** The larger capacitance occurs at the highest setting of the decade. The values given are for units without the shield can[illegible] place. With the shield cans in place, the shunt capacitance is from 0 to 20 pF greater than indicated here, depending on whether the shield is tied to the switch or to the zero end of the decade.

The 1218-BV lockable microwave oscillator.

POPULAR UHF OSCILLATOR NOW LOCKABLE

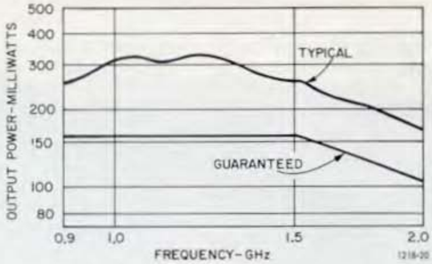
900 to 2000 MHz

The GR 1218 Oscillator has long been valued in many laboratories for its high output power and its spectral purity. Now, the 1218-BV lockable model means that frequency stability too can be achieved by use of a feedback control loop and reference-frequency source.

The oscillator can be phase locked to a stable reference of the same frequency or to a harmonic of the reference signal. Absolute stability of the 1218-BV may be less important than the stability of a difference frequency. In heterodyne systems the 1218-BV, used as the local oscillator, can track small changes in the test frequency. In this way the intermediate frequency is held within the detector passband.

A dc control voltage of ±25 volts will result in a frequency change typically greater than ∓2 MHz. Step-response time is typically under 1 microsecond. As in the 1218-A, output power into 50 ohms is at least 160 mW up to 1.5 GHz, dropping linearly to at least 110 mW at 2 GHz, and, in all other respects as well, performance of the 1218-BV is the same as its predecessor.

The 1218-BV can be used with any of three GR power supplies: 1264-B Modulating Power Supply, 1263-C Amplitude-Regulating Power Supply, or 1267-B Regulated Power Supply. The oscillator is available separately or in combination with any of the above power supplies for bench or relay-rack mounting.



1218-BV power output into 50 ohms.

SPECIFICATIONS

Frequency Range: 900 to 2000 MHz.

Frequency Calibration Accuracy: ±1%.

Warmup Frequency Drift: 0.1% total drift, typical.

Frequency Control: A 4-in. dial calibrated in MHz over 290° (10½-in. scale length); also 800-division logging scale. Slow motion drive of about 8 turns.

ΔF Control (Internal): > ±2 MHz by ⅞ turn of front-panel knob.

Power-Level Pulling (by ΔF control): < ±0.5 dB for ± 2-MHz ΔF.

ΔF Control (Remote): By dc voltage applied at front or rear jacks.

Frequency: > 4 MHz total range for 50-V change.

Voltage: Typical useful range ± 25 V; ΔF control sets center value from +10 to −20 V. Positive-going voltage causes frequency decrease. Applied voltage ±50 V max.

Interface Characteristic: Equivalent to 10 kΩ, 150 pF, and −1.3 mA current source in parallel across terminals, one of which is grounded. Ext source should have < 1000 Ω int impedance; may be ac coupled.

Step-Response Time: < 1 μs, typical.

Output Power (into 50 Ω): >160 mW, 0.9 to 1.5 GHz; drops linearly to >110 mW at 2.0 GHz.

Output Connector: Locking GR874® connector at rear panel. Adaptors available to other connector types.

Level Control: Full output to at least 20-dB attenuation set by uncalibrated front-panel control.

Modulation: AM INPUT jack at front panel for external audio-frequency plate modulation: approx 30 V rms into 6 kΩ required for 30% amplitude modulation. GR 1311 Audio Oscillator recommended as modulator.

Power Supply: Choice of three regulated power supplies. Oscillator is available separately or in combination with any power supply for rack or bench mounting.
The GR 1267-B Regulated Power Supply is suitable for cw operation.
The GR 1263-C Amplitude-Regulating Power Supply automatically holds the output at set level up to 2 V behind 50 Ω, cw or 1-kHz-square wave modulated.
The GR 1264-B Modulating Power Supply provides full power cw or modulated operation from internal 1-kHz square-wave or external pulse up to 100 kHz.

Mounting: Unit cabinet can be mounted with a power supply in single bench assembly or can be rack mounted alone or with power supply by use of appropriate rack adaptor.

Accessories Supplied: 2- and 3-contact phone plugs for modulation and frequency-control inputs.

Accessories Available: GR874® coaxial elements and adaptors.

Dimensions (width × height × depth): Bench, 12 x 7⅝ x 9 in. (305 x 195 x 230 mm); rack (with 1267 power supply), 19 x 7¼ in. (485 x 180 x 185 mm), 1263 or 1264 power supply adds 7 in. to rack height.

Weight (less power supply): Net, 14 lb (6.5 kg); shipping, 25 lb (11.5 kg).

Catalog Number		*Description*	*Price in USA*	
1218-9724		**1218-BV Lockable Oscillator**	$825.00	
Bench	Rack	**Oscillator/Power-Supply Combinations**	Bench	Rack
1218-9901	1218-9902	1218-BV with 1263-C Amplitude-Regulating Power Supply	$1375.00	$1404.00
1218-9903	1218-9904	1218-BV with 1264-B Modulating Power Supply	1240.00	1269.00
1218-9905	1218-9906	1218-BV with 1267-B Regulated Power Supply	1020.00	1046.00

NEW MEMBERS OF THE VARIAC®
FAMILY OF ADJUSTABLE AUTOTRANSFORMERS

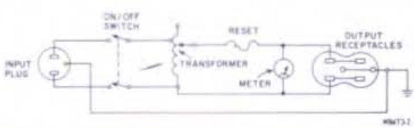
The W8MT3 and its metered companion, the W8MT3VM, provide the most volt-amperes per dollar of any Variac® model, many convenience features, plus the new General Radio "light look."

The W8MT3 and W8MT3VM are cased units that include a power cord and plug, an on-off switch that switches both sides of the input line, a dual receptacle for connection of a multiple load, a manual-reset overload protector, and a carrying handle. The W8MT3VM has a 0-to-150-volt meter that indicates output voltage of the Variac.

A three-wire power cord ties the ground (third) prong on the power plug to the ground-prong receptacle on the load outlet and to the case.

These portable W8-type autotransformers are designed for use at a minimum line frequency of 50 Hz for the stated 120-volt input. They can be operated at up to 400 Hz.

Because of the Duratrak® precious-metal contact surface, these Variac autotransformers can tolerate large momentary overloads without damage and have a comfortable margin-of-safety in normal operation at rated loads.



W8MT3VM circuit. The W8MT3 is the same except that it has no meter.

SPECIFICATIONS

INPUT
Voltage: 120 V.
Frequency: 50 to 60 Hz.
OUTPUT
Voltage: 0 to 140 V.
Rated Current: 10 A; equivalent to 1400 W at max output voltage.
Meter (in W8MT3VM only): 0 to 150 V.
No-Load Loss: 12 W at 60 Hz.
Driving Torque: 10 to 20 ounce-inches.
Replacement Brush: Type VB 2.
Weight: Net, 10 lb (4.6 kg); shipping, 20 lb (9.5 kg).

Catalog Number	*Description*	*Price in USA*
3038-5119	**W8MT3 Variac® autotransformer**	**$38.00**
3038-5015	**W8MT3VM Variac® autotransformer (metered)**	**68.50**
3200-5900	**VB2 Replacement Brush**	**1.00**

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