the GENERAL RADIO Experimenter

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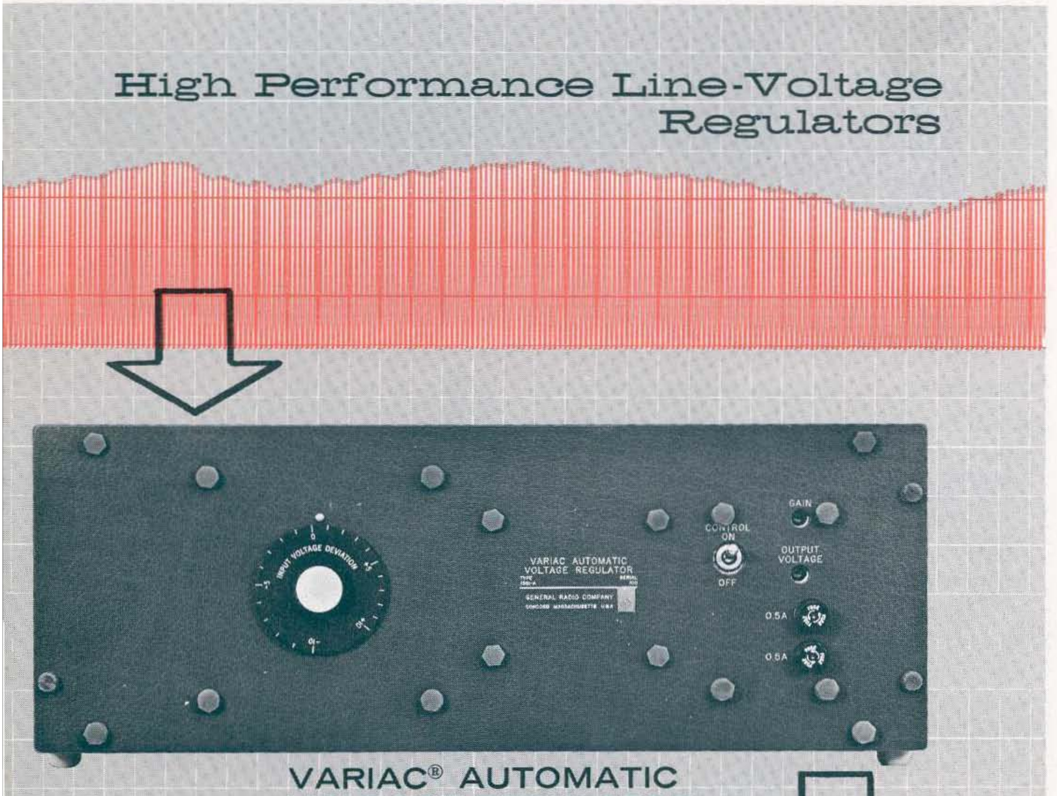

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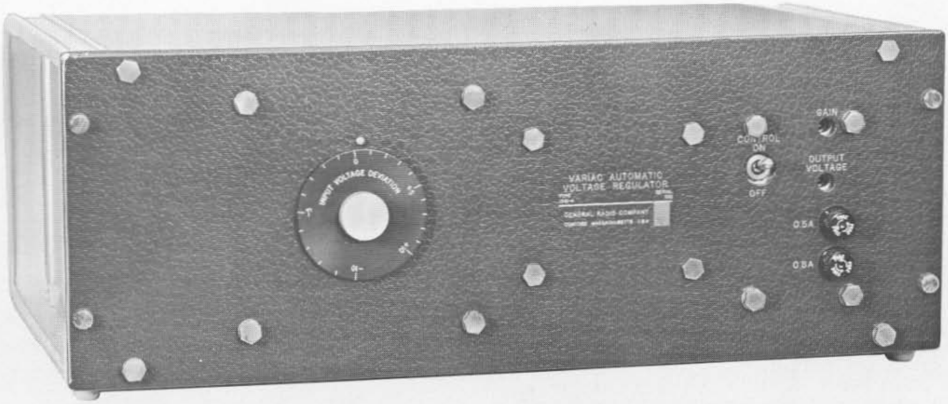

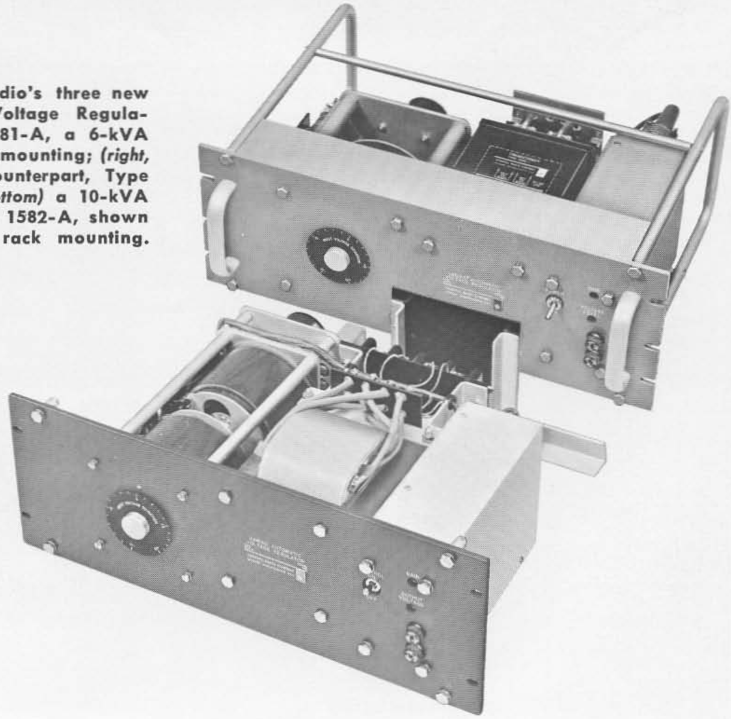

Figure 1. General Radio's three new Variac® Automatic Voltage Regulators: *(above)* Type 1581-A, a 6-kVA unit, shown for bench mounting; *(right, top)* its militarized counterpart, Type 1571-A; and *(right, bottom)* a 10-kVA commercial unit, Type 1582-A, shown without cabinet, for rack mounting.

IN THIS ISSUE

A NEW SERIES OF HIGH-PERFORMANCE LINE-VOLTAGE REGULATORS

GR's Type 1570 series of automatic line-voltage regulators has won wide acceptance since its introduction 11 years ago. Now a new series, including 26 different models, makes its debut. Major improvements in the new regulators: all-solid-state design, faster response, wider line-voltage ranges, increased power-handling capacity, and availability of all models in 400-cycle versions.

Regulated voltage is a necessity wherever voltage-sensitive elements are present — for instance, in many laboratory calibrations and measurements, industrial processes, and test programs. For such applications, as well as for more routine work in places where line-voltage regulation is poor, the automatic line-voltage regulator is indispensable. By regulation of the ac line, a number of critical load voltages — ac or dc — can be controlled with a single instrument.

CONSIDERATIONS IN VOLTAGE REGULATION

Three fundamental characteristics of ac line voltage are its frequency, magnitude, and waveform. Most commercial power systems hold line frequency within very close tolerance, and frequency instability is seldom a serious problem. Magnitude is the most critical parameter of line voltage, the one most generally subject to deviation, and the *raison d'être* for the line-voltage regulator. Although the primary function of the regulator is to stabilize the magnitude of the voltage, it is desirable that it do this without distorting the waveform.

An obvious measure of the performance of a line-voltage regulator is the accuracy with which it holds the magnitude of the output voltage to the nominal value. Accuracy can be specified as a long-term figure, if we disregard brief voltage fluctuations that exceed the regulator's response capability and thus pass from input to output undiminished. The ability of a regulator to react to very brief fluctuations is related to its response speed, another important measure of a regulator's performance.

Magnitude of an ac voltage requires further definition as peak, average, or root-mean-square, and the line-voltage regulator must choose one of these values to hold constant. Most regulators stabilize the rms or average magnitude, because these are the values most loads respond to.

What the regulator does to the waveform of the voltage it regulates is significant wherever the regulator and the load respond to different values of voltage magnitude (as, for instance, where an rms regulator is regulating voltage into a peak-responding device).

Another possible source of trouble is the power factor of the load. Specifications for many regulators assume a resistive load (power factor = 1.0), and departures from this condition can cause these regulators to shift output voltage by an amount well in excess of the stated accuracy.

Accuracy, response speed, introduced distortion, vulnerability to load power-factor — these, plus the practical factors of cost, reliability, and size, are the chief factors in the selection of a regulator.

TYPES OF LINE-VOLTAGE REGULATORS

There are three principal types of line-voltage regulators: electronic, magnetic, and electromechanical (servo). All operate by sensing the output voltage and adding voltage to, or subtracting it from, the input to restore the output to its nominal value.

The Electronic Regulator

The electronic (amplifier-type) regulator approaches the ultimate in ac-voltage-regulation devices. The regulated output voltage is continuously compared with a pure-sine-wave reference, and any errors are electronically removed. Such a regulator can actually remove distortion on the incoming line and can reduce, if not completely eliminate, transients on the regulated output voltage. Disadvantages are a sensitivity to line frequency and load power factor, modest power-handling capacities, and high cost per kVA.

The Magnetic Regulator

The several types of magnetic regulators are characterized by moderate correction speed, good to excellent reliability, a tendency to introduce distortion in the voltage waveform, and sensitivity to load current, power factor, and line frequency. Magnetic regulators also are usually quite heavy.

One type of magnetic regulator — the ferroresonant — is completely passive and boasts the highest reliability and the lowest cost. It is inherently short-circuit-proof, which adds to its reliability but which also severely limits its ability to handle large starting currents. Its output voltage level is factory set and cannot be changed.

Another type, the saturable-reactor regulator, uses an electronic feedback loop to achieve higher performance than the ferroresonant type. Harmonic filters are often included to reduce the considerable distortion introduced by the regulator.

A third type of magnetic regulator uses silicon-controlled rectifiers in an electronic switching scheme. Unlike other magnetic types, the SCR regulator operates satisfactorily with load power factors from 1.0 to 0 lagging, although its performance is not specified for leading power factors. The distortion introduced by this type of regulator characteristically exhibits itself as a large change in the peak value of the output voltage, which may persist independently of the regulator's normal rms- or average-correcting action.

Electromechanical Regulators

The electromechanical line-voltage regulator combines the power-handling capabilities of a motor-driven variable autotransformer with the fast response and accuracy of an electronic feedback loop. This type of regulator introduces no distortion, is totally unaffected by changes in load power factor from 0 leading to 0 lagging, and is insensitive to load current. With proper design, it can hold output voltage constant in the face of wide swings of line frequency, a factor important in the regulation of voltage from unstable sources, such as emergency generators. On a kVA-per-pound basis, it is the lightest of all automatic voltage regulators. It is well suited for applications involving heavy starting currents and can safely withstand transient overloads of up to 10 times the rated output current. Electromechanical regulators are practical over a wide range of power ratings, from one to hundreds of kVA.

THE NEW VARIAC® AUTOMATIC VOLTAGE REGULATORS

General Radio has developed an entirely new line of electromechanical regulators, consisting of 26 electrically different models, with militarized versions and several mounting options running the total to 114. Of the 26 electrically different models, 13 are for 60-cycle operation and the other 13 are designed to regulate 400-cycle voltage (with no 60-cycle power required). Each group of 13 regulators includes four 115-volt units, six 230-volt units, and three 460-volt units, covering an over-all range of power-handling capacity from 2 to 20 kVA.

These regulators can also be used in combination on three-phase systems.

Principles of Operation

The diagram of Figure 2 shows how these line-voltage regulators work. A voltage deviation at the output activates a servo feedback loop, consisting of a control unit, a two-phase motor, a VARIAC® autotransformer, and a step-down buck-boost transformer. The deviation is thus translated into a correction voltage that is added to or subtracted from the input to restore the regulated voltage to its correct value.

Control Unit

The solid-state control unit converts any small deviation in output voltage into a proportional electrical signal to drive the motor. The deviation is first sensed by an rms detector, whose dc output, after filtering, is compared with a constant 9-volt reference derived from a Zener diode. The resultant difference voltage is chopped into an ac error signal whose magnitude is propor-

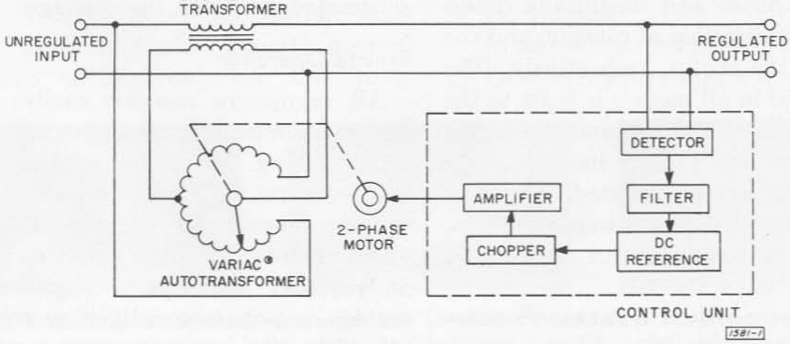

Figure 2. An elementary diagram of General Radio's electromechanical regulator.

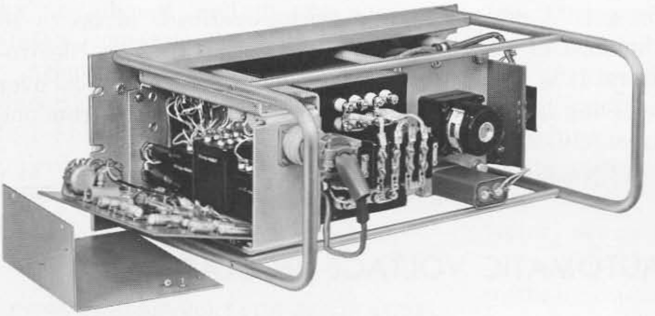

Figure 3. Rear view of a regulator with dust cover removed and etched board swung out for access to components and wiring.

tional to the output voltage deviation and whose phase is determined by the direction of this deviation. A solid-state amplifier converts the small ac error signal into the large ac voltage required by the motor.

An important feature of the General Radio regulator — and one that distinguishes it from most other electromechanical units — is the completely proportional control system. That is, the control unit continuously feeds a proportional correcting voltage to the motor. This technique helps give these regulators greater accuracy and higher response speed than are possible with a simple on-off motor control.

Motor and Transformer Circuits

The ac error signal from the control unit drives a two-phase servo motor, with the phase and magnitude determining the direction of rotation and the speed of the motor, respectively. The motor used in all models is built to the rigid specifications demanded for the militarized units (for instance, the stator is epoxy encapsulated). The low-inertia rotor contributes importantly to the excellent response and accuracy of the regulator.

The motor drives a VARIAC® autotransformer through a heavy-duty, instrument-type gear train. The autotransformer is positioned either side of a center tap (depending on the sense of the correction) by an amount proportional to the output-voltage deviation. All autotransformers are ball-bearing models with take-off brushes that are designed for the unusual demands imposed by regulator service.

The autotransformer output voltage is stepped down by a buck-boost transformer. Thus the full adjustment range of the autotransformer can be used to produce the relatively narrow range of correction voltage, and the current rating of the autotransformer is effectively multiplied. The phase of the voltage applied to the buck-boost transformer depends on the position of the autotransformer brush with respect to the center tap and determines whether this correcting voltage is added to or subtracted from the line voltage.

Remote Operation

All regulators can be easily connected for remote sensing, detection, or programming. Remote sensing at the load corrects for the voltage drop on the line between the regulator and the load. Remote detection permits use of an external detector to regulate dc voltage or a characteristic of ac voltage other than rms. For remote programming, the regulated output voltage level can be adjusted by an external variable resistance.

Physical Characteristics

All regulators are single, self-contained units seven inches high and 19 inches wide. The commercial models are available for bench, rack, or wall mounting. (A fourth option allows the customer to buy a rack model without cabinet, at a saving in cost.)

Militarized models are designed to meet or exceed the general requirements of specifications MIL-E-4158B and MIL-E-16400C. They are constructed on seven-inch U-shaped extruded aluminum channel to meet military shock and vibration requirements and are supplied for relay-rack installation.

The mechanical design, based on a modular approach, provides easy access to all components. Removal of six screws permits the dust cover to be removed and the etched board to be swung out (see Figure 3), exposing every wiring connection and component.

The two screwdriver adjustments (gain and output voltage), the manual-automatic control switch, and the control-unit fuses are all accessible at the front panel. Also on the front panel is a dial that indicates the percent difference between input and output voltages. This dial also permits manual voltage adjustment when the front-panel switch is thrown to MANUAL. There are no internal operating controls or adjustments.

C. E. Miller received his Bachelor of Engineering degree from Yale in 1960. He then joined the General Radio Engineering Staff as a Development Engineer in the Industrial Group, where he has been concerned with the design of automatic voltage regulators and stroboscopic equipment.

Performance

The introduction of General Radio's new automatic line-voltage regulators establishes the electromechanical regulator as comparable in performance to magnetic types, but lower in cost.

Accuracy and response speed are the best available in electromechanical regulators. Accuracy is 0.25 or 0.5 percent, depending on model, and is independent of load current, power factor, and frequency and voltage changes in the line. Response speed is 20 to 160 volts per second, depending on model, and is comparable to the speed of magnetic regulators of equivalent ratings. Figure 4 shows the actual performance of regulators responding to step changes in line voltage.

No waveform distortion is introduced by these regulators, and transient over-

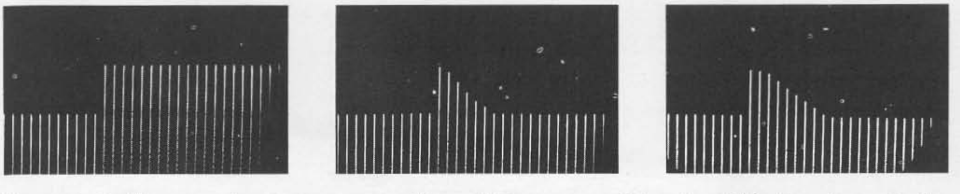

Figure 4. Oscillograms showing response of two of the new regulators to a 2% step change in line voltage. *Left,* the unregulated input; *center,* the response of a 6-kVA model; *right,* the response of a 10-kVA model.

loads up to 10 times the nominal ratings can be handled safely.

The 114 variations of GR's new family of regulators span wide ranges of voltage, correction range, frequency, and power rating, with several mounting options. The use of modular construction and interchangeable parts, moreover, suggests a further proliferation to meet special customer requirements.

— C. E. MILLER

Acknowledgment

The author wishes to acknowledge the work of M. J. Fitzmorris, who was associated with the development of these regulators in the early stages of the project.

COMMERCIAL MODELS

Regulators are available in the following ranges for either 400-cycle or 60-cycle service. By means of a change in connections, 60-cycle models can cover the range of 50 to 60 c/s, with only a slight reduction in correction range. Both Type 1581-A and Type 1582-A can be supplied for bench, wall, or rack mounting, or without cabinet.

TYPE 1581-A

*Output Volts**	*Correction Range %*	*kVA*	*Accuracy (% of output voltage)*
115	90 to 110	5.8	0.25
115	82 to 124	2.9	0.5
230	95 to 105	9.2	0.25
230	90 to 110	4.6	0.25
230	82 to 124	2.3	0.5

*Adjustable, ± 10%.

PRICES: from $495 to $575.

TYPE 1582-A is available in the same ranges of operation but has approximately twice the kVA rating. Additional models, connected for 460-volt service, have kVA ratings approximately ¾ those for 230-volt service.

PRICES: from $555 to $635.

MILITARIZED MODELS

TYPE 1571-A is a militarized version of the Type 1581-A. It is designed to meet the requirements of MIL-E-4158B and MIL-E-16400C.

PRICES: $650 for 60-cycle service
$695 for 400-cycle service

For Complete Listings with Specifications, Consult General Radio Catalog S, pages 223 to 226, or Call or Write Your Nearest GR Sales Office.

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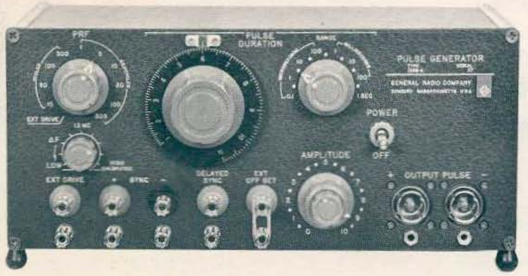
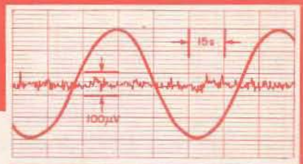

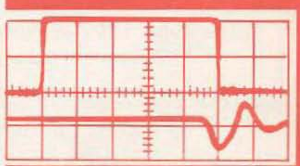
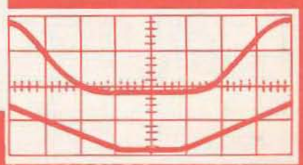

In this issue • • •

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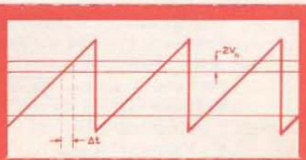
- Type 1398-A Pulse Generator
- Type 1397-A Pulse Amplifier

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REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

INPUT NOISE

ITS INFLUENCE ON COUNTER AND PULSE-GENERATOR PERFORMANCE AND ITS MEASUREMENT

The significance of input noise in counters, pulse generators, and other broadband switching circuits and a method for its measurement are the subjects of the first of three short articles featured this month. The second and third articles describe a new pulse generator and amplifier, both designed for general-purpose laboratory applications.

In an externally driven pulse generator, noise introduced in the input circuit determines the amount of jitter in the output pulse; in a frequency counter, the input-circuit noise affects the accuracy of the instrument, particularly in low-frequency period measurements. Clearly, a knowledge of the input-noise characteristics of such instruments is essential to an understanding of their accuracy specifications.

Techniques for measuring input noise were devised at General Radio during the development of the Type 1151 Digital Time and Frequency Meter[1], the only counter, as far as we know, that carries a direct input-noise specification. The same measurement techniques were applied to our pulse generators; Figure 1 is the result of such a measurement of our popular Type 1217-C Unit Pulse Generator. Inasmuch as we will specify input noise, where appropriate, for future instruments, we will explain in this short article how we make the measurements.

[1] R. W. Frank, "Zero to 300 kc with Five-Digit Accuracy," *General Radio Experimenter*, June 1963.

The Significance of Input Noise

The operation of a typical input switching circuit is shown in Figure 2. This is the characteristic of a simple Schmitt trigger circuit, with the triggering and resetting points defining the hysteresis voltage region. Figure 3 magnifies the triggering point to show the area of uncertainty and to indicate how this area is influenced by noise on the triggering threshold and on the signal. It is apparent that the error in the time of triggering depends on the slope of the signal voltage and the magnitude of the noise. If a signal of

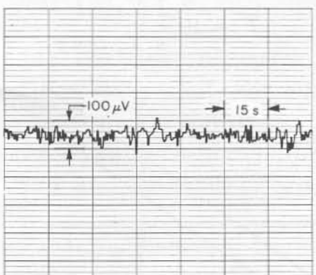


Figure 1. Noise of a Type 1217-C Unit Pulse Generator driven by a 2.05-V ramp with a 204.8555-millisecond period. The last three digits are recorded. One major division corresponds to 100 µV, peak-to-peak, noise.

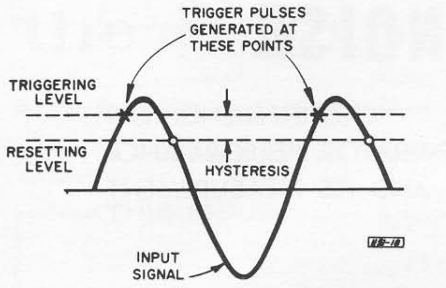


Figure 2. Operation of a typical input circuit.

precisely known period and very low noise is applied to a noisy threshold, the error in a period measurement will be as shown in Figure 4.

If the signal applied in Figure 4 were a sine wave, then the slope of the signal voltage would be the time derivative of the signal as it passes the triggering area. This leads to the familiar figures for period-measurement error due to signal-to-noise ratio. At zero crossings:

$$\Delta T = \pm \frac{1}{\pi} \frac{V_n}{V_s}$$

If triggering occurs not at the zero crossing but at an angle θ in the input-voltage cycle, then:

$$\Delta T = \pm \frac{1}{\pi} \frac{V_n}{V_s} \frac{1}{\cos \theta}$$

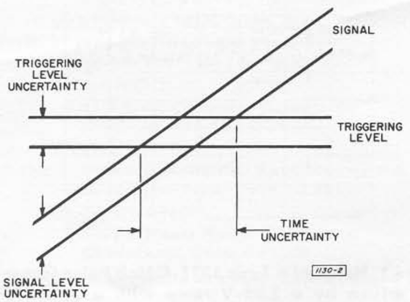


Figure 3. The effects of uncertainty in signal on triggering level.

In both the above equations, V_n/V_s is the noise-to-signal ratio. If this figure and the crossing angle are known, then the error in a period measurement can be determined. This is the basis for the accuracy specifications of most counters, *with no assumptions made about where the noise V_n comes from.* To make a valid measurement, the user must concern himself with two quantities: First, he must know the signal-to-noise ratio of the signal he is measuring; second, he must know the equivalent input noise of the measuring instrument.

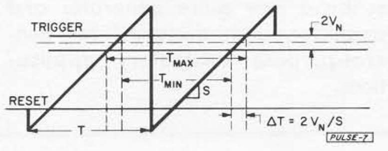


Figure 4. A periodic signal of slope S volts/second and period T traverses the hysteresis region with a superimposed noise signal of peak value V_n. By inspection it can be seen that the maximum time-interval fluctuation, ΔT, will be $\pm$ 2 V_n/V_s.

As an example of the practical implications of these factors, consider a single-period measurement of a 10-volt signal with an 80-dB signal-to-noise ratio. If the counter introduces more than 1 millivolt of noise (and most counters do), that noise, rather than the noise on the signal, will control the accuracy of measurement.

Measuring Input Noise

The system used at General Radio for quantitative measurement of the performance of switching circuits is shown in Figure 5. The task that this system must perform can be deduced from Figure 4. It must produce a

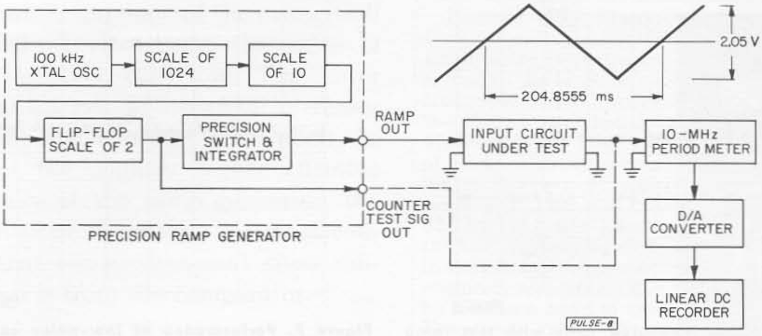


Figure 5. The precision ramp and the system for generating it. The quartz-crystal oscillator output is divided by 20,480 to produce a precise time interval. The crystal is set low in frequency by about 25 Hz to produce final period of 204.8555 milliseconds. Dc-coupled output is set to exactly 2.05 V, peak-to-peak, so the slope is nearly 20 V/s in both directions.

completely clean signal of exactly known period, feed this signal into the input circuit to be evaluated, and measure the fluctuations in period of the triggers produced. If very high signal-to-noise ratios are to be attained, frequency stability must be excellent; therefore the system uses a crystal oscillator. The reference signal is triangular; if it were sinusoidal, either a separate measurement would be necessary to determine that the input circuit was truly switching at zero crossing or corrections to the signal-to-noise figure for signal slope at different triggering levels would have to be made.

In the test setup shown, the ramp is highly linear, so that no corrections for phase are necessary. The ramp slope is symmetrical, so that the noise for both negative-going and positive-going threshold detectors can be evaluated.

The ramp has a period of 204.8555 milliseconds (the last digits chosen to place a plot at midscale on a strip chart) and an amplitude of 2.05 volts, or positive and negative slopes of 20 volts/second.

This ramp is fed to the input circuit under test, and the period of the output triggers from the triggering circuit is measured on any 10-MHz period-measuring counter. In the complete absence of noise, the counter would indicate 204.8555 milliseconds. The left-hand digit corresponds to volts, the second to tenths of a volt, etc. Thus the final digits (555) correspond to hundreds, tens, and units of microvolts of noise. The $\pm$ 1 count error of the counter therefore corresponds to an ultimate system resolution of $\pm$ 1 μV. The counter reading is fed through a D/A converter to a strip-chart recorder.

The big question, of course, is — how good is the ramp? For what we actually measure are time-interval fluctuations due to noise in the combination of the ramp and the input circuit under test.

A linear amplifier can be inserted between the ramp generator and the input circuit under test to increase the ramp slope. If the slope is increased by a factor of 10 (corresponding to 20 dB of inserted gain), the ramp-amplifier noise will quickly predominate over

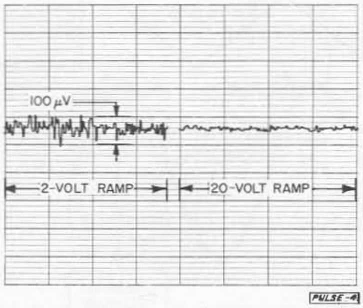


Figure 6. Noise measured (*left*) with test ramp applied directly to a Type 1217-C and (*right*) with low-noise amplifier inserted between ramp generator and a Type 1217-C.

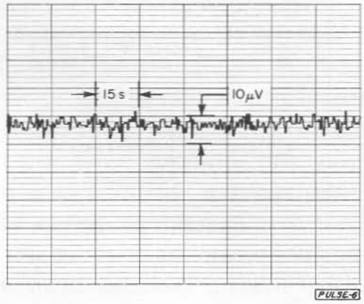


Figure 7. Performance of low-noise comparator with precision ramp. System noise appears to be about 6 μV, peak-to-peak.

that of the input circuit alone. The ramp noice can then be measured, as long as the amplifier noise is negligible.

Figure 6 compares the input noise measurement of Figure 1 with a similar measurement made with a 20-dB low-noise amplifier inserted between ramp generator and input circuit. The sharp reduction in noise as the slope is increased by an order of magnitude is convincing proof that most of the noise is in the input circuit, not in the ramp-amplifier combination.

Figure 7 shows how the best balanced amplitude comparator[2] we have yet built (Figure 8) responds to the test ramp. One would conclude from Figure 7 that the ramp-comparator combination has a peak-to-peak noise no worse than 10 μV (3 μV, rms). The true rms value for any of these measurements can be either computed from the successive counter readings or measured with a low-frequency rms meter.

Again the question must occur: Is the primary contributor to this over-all noise the ramp or the comparator? With such a low-noise comparator, it is not possible to insert an amplifier with any assurance that its input noise is lower than that of the comparator.

[2] R. W. Frank, "How to Kill Time — Accurately!" *General Radio Experimenter*, December 1958, p 8.

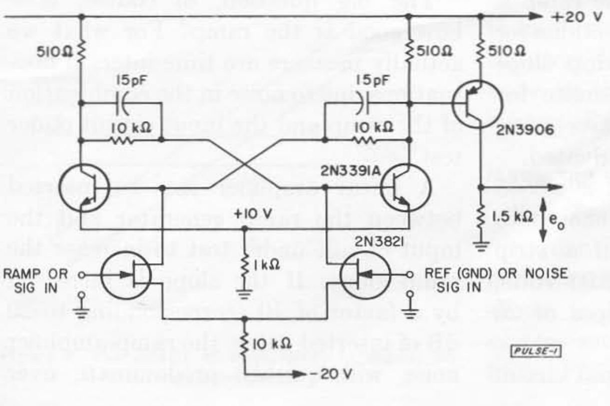


Figure 8. Schematic diagram of low-noise comparator. Balanced input circuit uses TI 2N3821 N-channel field-effect transistors. Hysteresis is 0.3 V at either gate. Comparator maintains constant hysteresis to over 500 kHz.

In Figure 9, the slope of the ramp itself is varied by changes in the value of the integrating capacitor. The lower value of integrating capacitor increases the ramp slope to 30 volts/second. In spite of the slightly higher effective impedance of the ramp generator, the over-all noise decreases. We must conclude that the predominant noise contribution is from the comparator.

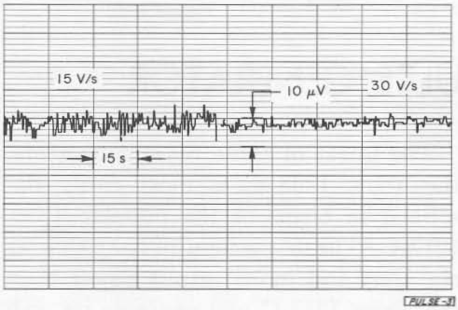


Figure 9. Performance of FET comparator as ramp slope is changed from 15 to 30 V/s.

In Figure 10, the trace begins showing the system noise of less than 10 μV. A 1-kHz signal of 30 μV, rms, amplitude is then injected into the comparator; finally, 30 μV of noise at 20-kHz and 500-kHz bandwidths is applied for several minutes. Note that 30 μV of noise in the 20-kHz band does occasionally attain 100 μV, peak-to-peak, amplitude but doesn't look like 30 μV rms. With the 500-kHz bandwidth, the rms value of the noise looks even smaller. This proves that the comparator functions as a time-domain filter.

From such measurements we conclude that the input noise of a Type 1217-C Pulse Generator is about 20 μV, that of the Type 1398-A about 50 μV. Since most counters have noise figures probably in the 1-to-10-millivolt region, this means that the accuracy of a single-period measurement can be improved by more than an order of magnitude if a Type 1217-C or 1398-A Pulse Generator is used to trigger a counter.

— R. W. Frank

Richard W. Frank, Group Leader in the GR Engineering Department served in the U. S. Navy from 1942 to 1945. After the war he attended the Massachusetts Institute of Technology, receiving his SB in electrical engineering in 1950 and his SM in 1951. He joined the GR engineering staff in 1951 as a development engineer, and since 1957 he has been head of the Frequency and Time Group.

Figure 10. Performance of the ramp-comparator system with various noise signals applied. Trace starts with 30 seconds with no noise signal. Then 30 μV, rms, at 1kHz is applied for about 45 seconds, showing that the ramp slope is about correct and that one division equals 100 μV, peak-to-peak. Then a Type 1390-B Random-Noise Generator provides 30 μV, rms, of random noise in 20-kHz and 500-kHz bands. With the 500-kHz bandwidth, one would have to wait for a very long time for a full deviation error. Note the similarity of the noise generator trace to that showing system performance, an order of magnitude lower in level, in Figure 7.

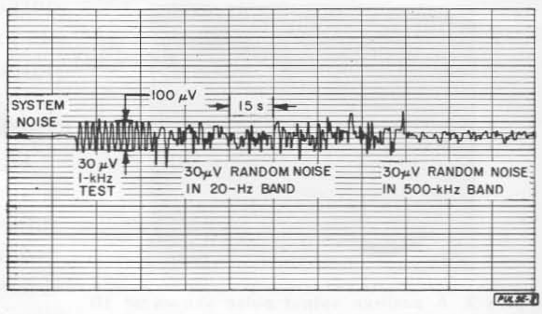


Figure 1.

THE TYPE 1398-A PULSE GENERATOR

An instrument, no matter how ingeniously designed and efficiently made, is never so good that it can't be improved. That is the story of GR's general-purpose pulse generator, which our engineers designed over 10 years ago and have not been able to leave alone since. Each successive model represented unequalled value for this class of instrument, and customer response has always been excellent. But the campaign continued to squeeze just a few more ounces of performance out of the design, while holding the cost in line.

[1] R. W. Frank, "Improved Performance from the Unit Pulse Generator," *General Radio Experimenter*, December 1964.

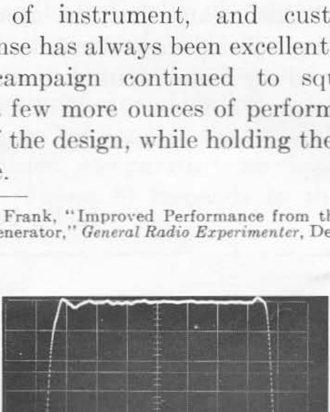


Figure 2. A positive output pulse shown at 10 ns/cm writing speed. The fall time is about 3 ns.

The latest of these pulse generators, the Type 1398-A, is most easily described in terms of the popular Type 1217-C.[1] The new generator has a shorter rise time and more output power than the Type 1217-C and contains its own regulated power supplies. For those who don't need the faster pulse and the extra power and who have their own power supply, the Type 1217-C remains available.

To review the general specifications of the Types 1217-C and the 1398-A: In both, prf range is 2.5 Hz to 1.2 MHz and duration is 100 nanoseconds to 1 second. In the Type 1398-A, transition times have been dropped from 10 to 5 nanoseconds (see Figure 2), and output current has been boosted from 40 to 60 mA, so that the open-circuit output is 60 volts behind 1 kilohm.

Circuit

Figure 3 is a block diagram. In the timing and output circuits, vacuum tubes are used for their high input impedance and ruggedness; the pulse-control circuits are transistorized for maximum switching speed. In the Type

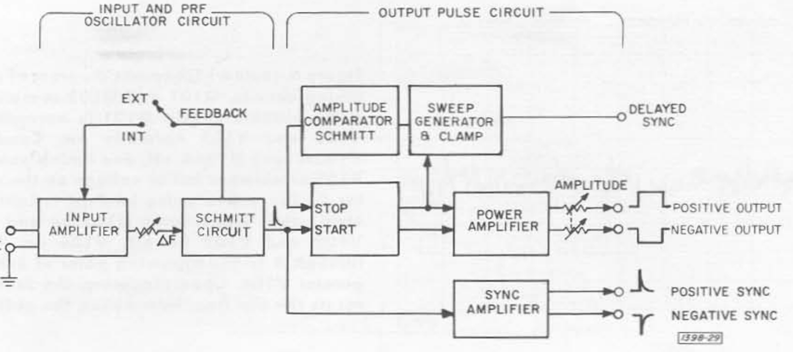


Figure 3. Block diagram of the pulse generator.

1398-A, new silicon npn transistors and double-frame-grid power output tubes further extend the capabilities of the hybrid arrangement.

The input circuits, shown in Figure 4, require only three active devices, which are switched into operation as either a prf oscillator or an aperiodic trigger circuit, depending on whether internal or external drive is desired. Equivalent input noise (as measured by the techniques described in the preceding article) is about 50 μV, rms (see Figure 5), which is much lower than that of many trigger-type input systems. This low input noise leads to very low prf jitter with external drive.

The push-pull output circuit (Figures 6 and 7) offers many advantages: It presents a constant load to the power supply; it delivers both positive-going and negative-going pulses simultaneously; the pentodes used are linear current sources, which produce the same source impedance for either polarity; the output is short-circuitable, and there are no duty-ratio restrictions; and finally, since the output terminals are direct-coupled to the output stage plates, there is never any rampoff.

Another characteristic of this circuit is that it retains a dc component negative with respect to chassis ground. A feature of the new generator is provision

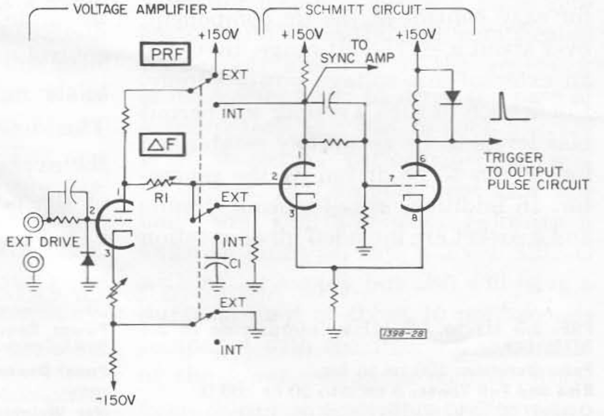


Figure 4. Schematic diagram of input circuits. Switched as shown, for external operation, the Schmitt circuit is driven by the input dc amplifier and is a fast 2-MHz-to-dc trigger circuit. When the PRF control is set for the internal mode, the same components are switched into a stable oscillator that can be injection-synchronized from the external-drive terminals. The input tube becomes a current source to translate the left-hand plate swing to center on the Schmitt circuit. The circuit oscillates within its hysteresis region. C1 establishes frequency range, R1 gives continuous frequency adjustment.

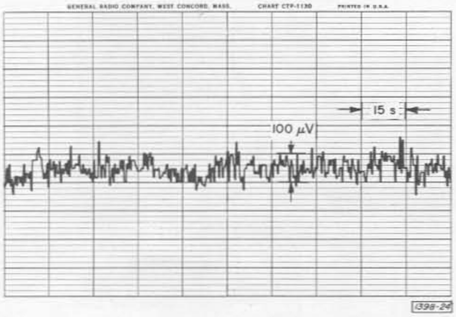
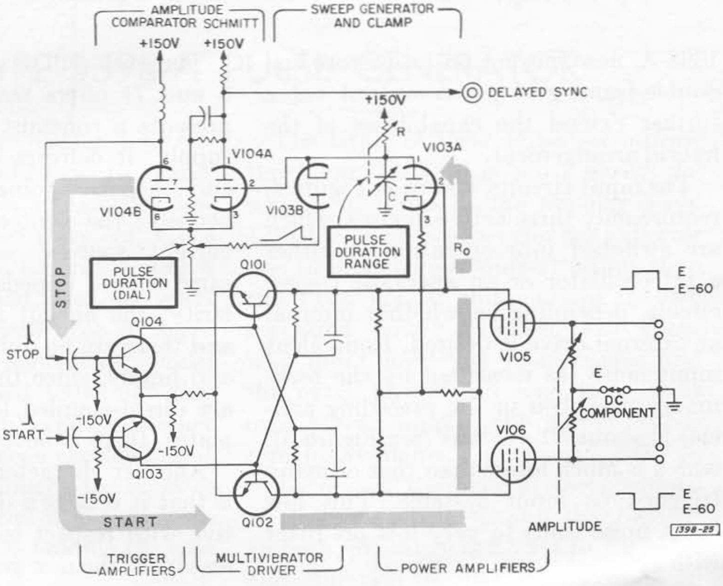


Figure 5. (Above) Plot showing noise generated in input circuit.

Figure 6. (Below) Schematic diagram of output and timing circuits. Q101 and Q102 comprise a transistor bistable circuit. Q101 is normally off and V105 and V103 normally on. Conduction of V103A keep V104A off, and timing potentiometer R125 establishes initial voltage on timing capacitor C. The active pulse interval is initiated by a start pulse, which turns Q101 on and Q102 off. V103 and V105 go off, V106 on. C charges through R to the triggering point of Schmitt comparator V104. Upon triggering, the Schmitt circuit resets the flip-flop, terminating the pulse.



for easy control of the dc component, over about a ± 15-volt range, by use of an external low-voltage power supply. Connection of such a supply will permit bias levels to be accurately established for devices to be driven by the generator. In addition, signal-sensing circuits and a socket are included for connection to a to-be-announced dc-component control unit, which will attach directly to the right-hand side of the generator. This unit will automatically control the average value or either peak value of the positive or negative pulse over a ± 15-volt range.

ABRIDGED SPECIFICATIONS

PRF: 2.5 Hz to 1.2 MHz, internal; dc to 2.4 MHz, external.
Pulse Duration: 100 ns to 1 s.
Rise and Fall Times: 5 ns into 50 or 100 Ω.
Output: 60 mA, positive and negative.
Power Required: 105 to 125, 195 to 235, or 210 to 250 volts, 50 to 60 Hz, 90 W.
Panel Dimensions: 12 × 5¼ inches (305 × 135 mm).
Net Weight: 14½ lb (7 kg).

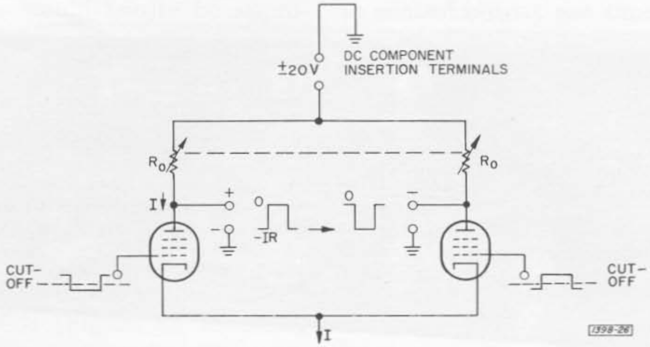


Figure 7. A closer look at the output circuit. The system includes a pair of switched current sources, V1 (normally on) and V2 (normally off). When the active pulse interval switches V1 off, the output voltage at the positive pulse terminal goes from —IR to 0, while the voltage at the negative terminal goes from 0 to —IR. Output impedance is adjusted to control open-circuit output voltage and is unaffected by switching circuits. The dc-component insertion terminals permit use of an external low-voltage power supply to translate the reference from 0 volts to the external supply voltage.

Catalog Number	*Description*	*Price in USA*
1398-9701	**Type 1398-A Pulse Generator**	**$535.00**
0480-9632	**Type 480-P312 Rack-Adaptor Set**	**6.50**

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

For complete specifications, see the current General Radio catalog or write to your nearest GR sales office.

THE TYPE 1397-A PULSE AMPLIFIER

A new pulse amplifier, the Type 1397-A, has been designed as a companion not only for the Type 1398-A Pulse Generator described in this issue but for the Type 1217-C Unit Pulse Generator and the Type 1395-A Modular Pulse Generator[1] as well. The new amplifier increases the relatively low output power available from these instruments to a healthy 50 watts peak.

Why have pulse generator and pulse amplifier in separate packages, anyway? The separation, a long-standing GR practice, makes excellent sense. Why, for instance, saddle a pulse generator with a costly and complicated high-power amplifier if the generator is most often used to drive the base of a transistor? The cost of such overkill is measured in performance as well as dollars. A 1-ampere generator not only will generally cost more than twice as much as, say, a Type 1217-C with power supply but also will have a duration limit of about 10 milliseconds compared with the 1-second maximum of the lower-cost pulse source. Moreover, since few generators producing an

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

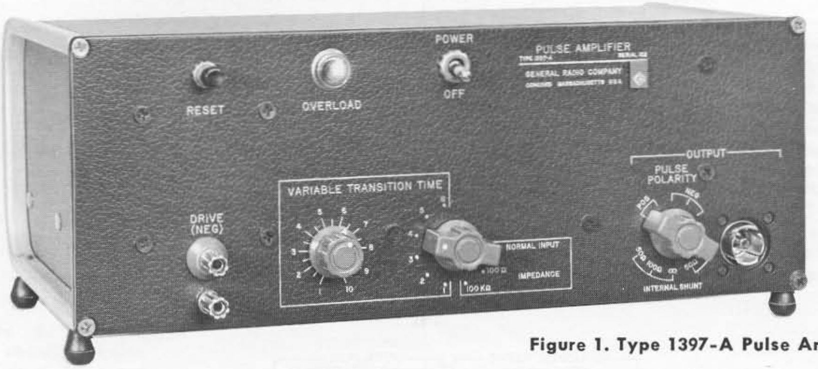


Figure 1. Type 1397-A Pulse Amplifier.

ampere can produce it continuously (another problem in economics), there must be duty-ratio restrictions, overload protection, etc.

The separation of generator and amplifier offers some important design advantages, too. The new pulse amplifier is unique, for example, as a very nearly linear amplifier capable of amplifying complex waveforms. This useful

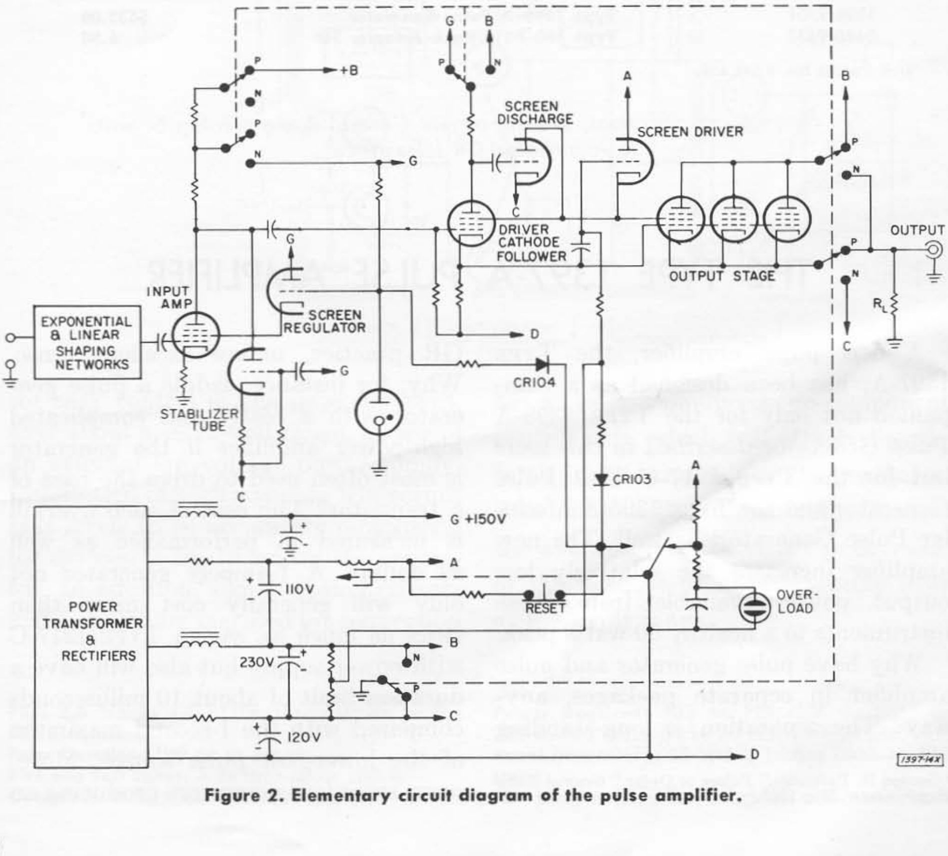


Figure 2. Elementary circuit diagram of the pulse amplifier.

characteristic would hardly be appropriate in the usual 1-ampere pulse generator, where the output stage is just a big, fast switch.

Having established the case for separate units, let us turn to the amplifier at hand. It needs only a 2-volt negative pulse at the input to produce a 1-ampere positive or negative output pulse. Rise and fall times are typically 40 nanoseconds (Figure 3), assuming the driver can get under 20 nanoseconds. Another useful feature is a variable-transition-time mode, which offers the user continuous adjustment of rise and fall times from 0.1 to 100 microseconds.

The output pulse can be either terminated in internal loads or coupled, without internal loss, to an external load. A single switch is used to select output polarity and load configuration.

Since the amplifier is substantially linear (Figure 9), it is a simple matter to control output rise time by controlling the input pulse (Figures 11, 12). An input control does this, switching in a 100-ohm or 100-kilohm resistor, for minimum rise time (normal mode), or a set of simple networks to shape the input pulse into either an exponentially or linearly rising function over a range from about 0.1 to 100 microseconds.

The maximum duty ratio for the amplifier is 1/10. If this is exceeded, internal voltages are automatically switched off and a flashing lamp on the front panel calls for a manual reset.

Circuit

The straightforward circuit is shown in Figure 2. The input shaping networks are followed by a voltage amplifier, a cathode follower, and the output stage, consisting of three parallel-connected pentodes. These are connected as cathode followers for positive output pulses, as amplifier-inverters for negative output pulses. The changes in gain

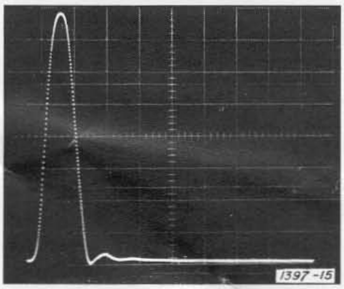
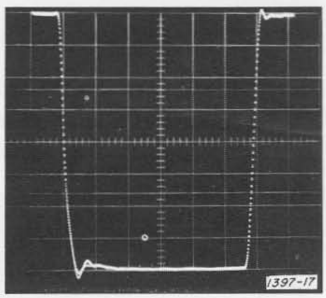


Figure 3. (*Left*) **A 0.08-μs input pulse from a Type 1217-C is amplified to 1.2 amperes in a 50-ohm system. Amplifier displays rise and fall times of about 0.04 μs and about 5% overshoot on trailing transition. Oscilloscope is Tektronix 661-A at 0.1 μs/cm.**

Figure 5. (*Right*) **The 0.6-μs pulse from the Type 1217-C is passed by the amplifier set for negative output. Note minimal ringing and clean transitions.**



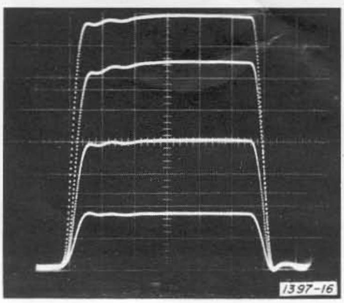


Figure 4. (*Left*) **The positive pulse duration is increased to 0.6 μs, and the driving pulse amplitude is decreased in three steps from full 1.2 amperes to about 80%, 50%, and 20% of initial level. Note slight changes in transient response. Small undershoot at positive peak flattens to small overshoot at 20% level.**

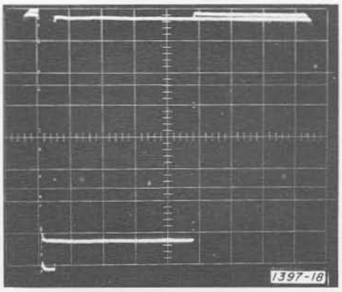


Figure 6. (*Left*) **Effects of duty ratio on output pulse. PRF is 10 kHz and duration of shorter pulse is 1 μs. When the Type 1217-C is switched to produce a 10-μs pulse, the 10% duty ratio lowers the pulse amplitude to about 92% of the initial level. Note that pulses still retain excellent shape.**

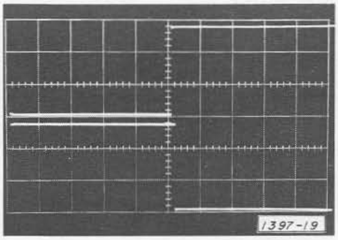


Figure 7. (*Right*) **Still longer pulses. With duration increased to 100 μs, no defects show in either positive or negative pulse. Oscilloscope is now a Tektronix 551. The Type 1217-C is still driving the amplifier in this double exposure.**

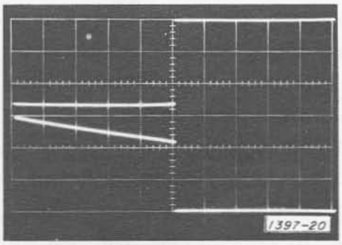


Figure 8. (*Left*) **At the slowest writing rate, 1 ms/cm, a practically flat-topped negative pulse is seen just starting to roll down at 4 ms. It will fall more rapidly from here on. The positive pulse falls linearly, is down by 30% at 5 ms.**

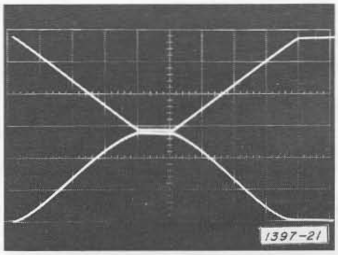


Figure 9. (*Right*) **The story on linearity. We're indebted to the Type 1395-P3 Pulse Shaper for the beautifully linear ramp on top. This is the input to the Type 1397-A. The lower trace is the output, at full 1.2-A power. (The amplifier is set to invert the negative input pulse.) Transfer function shows lower gain at very low and very high power level as pulse starts from cutoff and rises into grid-current regions of output stage.**

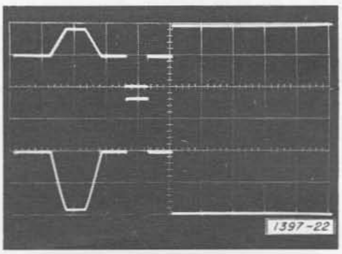


Figure 10. (*Left*) **More linearity. The Type 1395-A Modular Pulse Generator provides a distinctive pulse to amplify (top), and the Type 1397-A gives a reasonable facsimile at a 1-ampere level.**

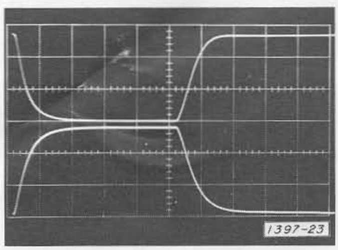


Figure 11. (*Right*) **Positive and negative 10-μs output pulses with rise-time controls set for exponentially rising pulse.**

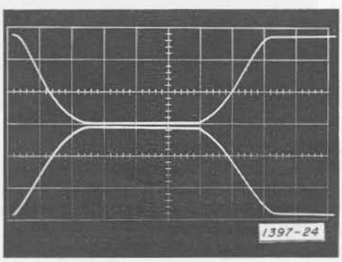


Figure 12. (*Left*) **Positive and negative 10-μs pulses with controls set for linearly rising pulse show 2-μs rise times. They again display the transfer function of Figure 9, should be useful in driving magnetic structures at slower transition rates.**

due to the different output configurations are compensated for by plate-load switching at the input amplifier.

So that the base of either positive or negative output pulse can be grounded while the pulses retain their dc component, the power supply for the entire output system is switched along with output polarity. The power supply for the input-amplifier stage is not switched, so that the grid-cathode signal voltage is always developed with respect to ground.

Performance

Oscilloscopic comparison of input and output signals is the most objective evidence of an amplifier's capabilities. The unretouched oscillograms of Figures 3 through 12 show what the Type 1397-A can do under a variety of operating conditions.

ABRIDGED SPECIFICATIONS

Mode	*Input Impedance*	*Drive Required*	*Rise and Fall Times*
NORMAL	100 Ω or 100 kΩ shunted by approx 50 pF, switch selected	−2 V, p-to-p, minimum	<50 ns (typically 30 ns) with input rise and fall times of <20 ns
VARIABLE Linear	30 kΩ, approx	−30 V, p-to-p, approx, minimum	0.1 to 100 μs, approx, linear, con-continuously adjustable
Exponential	100 Ω	−2 to −4 V, p-to-p, approx	0.1 to 100 μs, approx, exponential, continuously adjustable

OUTPUT

Rampoff: Approx 20% with 5-ms pulse duration.

Amplitude: 1.2 A, p-to-p, max (60 V into 50 Ω). 1 A, p-to-p, with 10% duty ratio. Automatic overload protector with manual reset.

Amplitude Variation: ±10% for duty-ratio changes from minimum to 10%. With ±10% line-voltage changes, positive variation is ±10%, negative output is ±5%.

Internal Shunt: Positive output, 50 Ω or open circuit; negative output, 50 Ω, 100 Ω, or open circuit.

GENERAL

Max Duty Ratio: 10%.

Power Required: 105 to 125, 195 to 235, or 210 to 250 V, 50 to 60 Hz, 100 W.

Panel Dimensions: 14 × 5⅞ inches (355 × 150 mm).

Net Weight: 18 lb (8.5 kg).

Catalog Number	*Description*	*Price in USA*
1397-9701	**Type 1397-A Pulse Amplifier**	**$495.00**
0480-9634	**Type 480-P314 Rack-Adaptor Set**	**6.00**

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

For complete specifications, see the current General Radio catalog or write to your nearest GR sales office.

Among the first shipments of instruments during the war was a number of precision air capacitors. One of these found its way to an Army laboratory in France, where Lieutenant E. H. Armstrong was experimenting on a new circuit to improve the performance of radio receivers. He appropriated the capacitor, his new circuit was a sensational success, and thus one of the Company's earliest products was incorporated in the first superheterodyne receiver in 1917.

(from *A History of the General Radio Company*)

INDEXES, BINDERS, HISTORIES OFFERED

INDEX

The 1965 *Experimenter* index is now available and will be mailed free of charge to those requesting it. The index lists all articles published during the year and is arranged by subject, author, and instrument type number.

A small quantity of indexes for other years is also available. If you would like to round out a collection, perhaps we can help you.

BINDERS

You can also obtain a binder for your *Experimenters* simply for the asking. Each binder will hold about two years' *Experimenters.*

HISTORY

The *General Radio Company 1915–1965* is the title of a 32-page monograph written by Dr. Donald B. Sinclair, GR president, and available (while the supply lasts) to *Experimenter* readers on request. The short history, presented at a meeting of The Newcomen Society, tells of the people and philosophies that helped make GR, and, in a sense, the electronics industry, grow.

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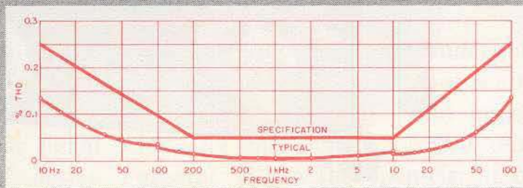


This Issue:

- All Solid-State, Low-Distortion Oscillator
- Sine-Squared Pulses

VOLUME 40 · NUMBER 3 / MARCH 1966

Figure 1.

10 Hz TO 100 kHz

ALL-SOLID-STATE, LOW-DISTORTION OSCILLATOR

SINE- OR SQUARE-WAVE

The Type 1309-A is the second in our new line of general-purpose variable-capacitance-tuned RC oscillators, recently introduced with the Type 1310-A Oscillator.[1] It continues the modern, all-solid-state design used in the Type 1308-A and Type 1311-A Audio Oscillators.

The continuing need for greater accuracy in electrical measurements, the improvements in sound-recording and reproduction equipment, and the increasingly stringent requirements on communication systems have all lowered the levels of distortion acceptable in new equipment. To design and to test this equipment, accurate distortion measurements must be made, and an essential part of any harmonic distortion measurement is a low-distortion source.

As distortion specifications have tightened, generators with lower and lower distortion levels have become available, but most of them have been quite expensive owing to a concomitant emphasis on extreme amplitude stability and a flat frequency characteristic. In lower-cost, general-purpose oscillators, on the other hand, the transition from vacuum tubes to transistors has led to higher, rather than lower, distortion levels, as a result of the techniques used to obtain these additional characteristics.

The new Type 1309-A Oscillator, an all-solid-state, capacitance-tuned oscillator, combines very low distortion with accuracy and stability ample for

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

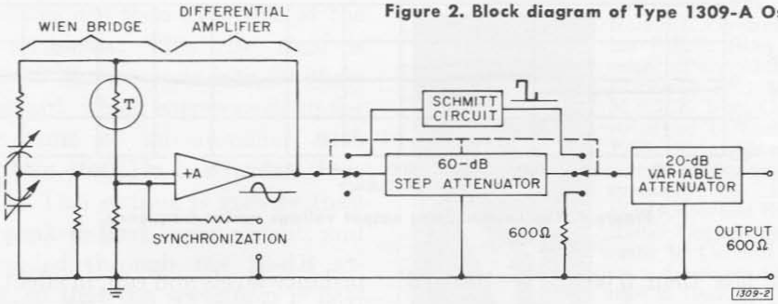


Figure 2. Block diagram of Type 1309-A Oscillator.

general laboratory use. Output is quite constant over the entire frequency range of 10 Hz to 100 kHz, and an output attenuator provides the low signal levels necessary for testing active devices. As with other General Radio RC Oscillators, there is an input-output synchronization jack. An added feature is a square-wave output for transient-response measurements. This output has a symmetrical waveform and an unusually short rise time.

Figure 1 is a panel view of the oscillator, and Figure 2 shows the three major elements: a low-distortion Wien bridge oscillator, a Schmitt squaring circuit, and an output attenuator.

DISTORTION

The low distortion is achieved through the use of a high degree of negative feedback and a thermistor of special design for amplitude control.[2] The distortion at full output is approximately constant with load impedance for any linear load of 600 ohms or greater. When the open-circuit output is one volt or less, the distortion is independent of the size of the load. The distortion is typically less than 0.01% for frequencies near 1 kHz, often below what can be conveniently measured. Note that the shape of the distortion curve (Figure 3) is typical of most audio-frequency devices, so that the margin between the source distortion and a device under test remains approximately constant with frequency.

Low levels of hum and noise are always desirable, but to be useful for broadband distortion measurements an oscillator must have noise and hum that are at least as low as the distortion. The 1-kHz output of the Type 1309-A has noise typically less than 0.005% in a bandwidth of 5 Hz to 500 kHz, and

[2] Robert E. Owen, "Solid-State RC Oscillator Design for Audio Use", *Journal of the Audio Engineering Society*, Vol 14, January, 1966.

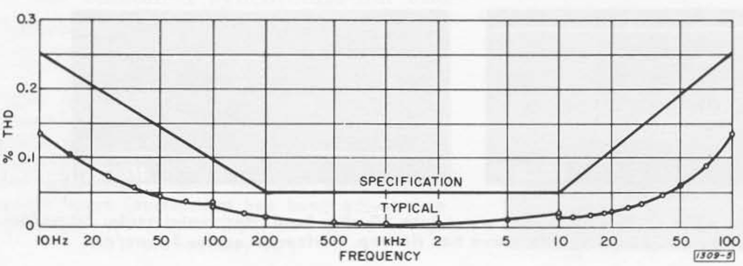


Figure 3. Oscillator distortion for 600-ohm load or open circuit.

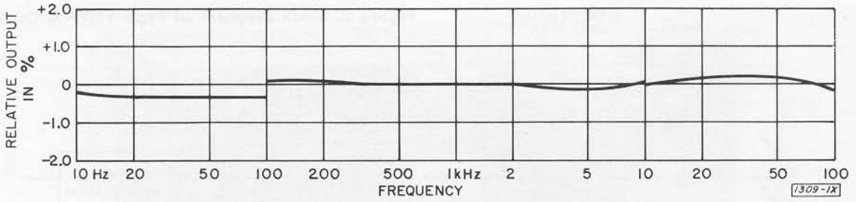


Figure 4. Typical oscillator output voltage versus frequency.

hum is less than 0.001% (−100 dB) of the full output.

ATTENUATOR

The sinusoidal open-circuit output voltage can be varied continuously between 5 volts and less than 0.5 millivolt by means of the output attenuator. An additional position, labeled zero volts, disconnects the oscillator output from the terminals yet maintains the 600-ohm output impedance. This provides a convenient transient-free means of reducing the output to zero without disturbing the continuous attenuator setting or shorting or disconnecting a carefully shielded system. Further, it aids in locating ground loops and other sources of extraneous signals when one is working with small signal levels. With the oscillator output removed, the extraneous signals are not masked, so that they are easier to measure and to eliminate. This technique offers considerable advantage over the often-used one of short-circuiting the output. Shorting drastically changes the impedance levels and can, in effect, change the whole circuit, possibly eliminating the very source that one is trying to isolate.

The variation of the output of any oscillator at different frequencies is perhaps its most noticeable departure from ideal. Because of this, and because a constant output is convenient for response measurements, most modern transistor oscillators have relatively flat output-frequency characteristics, although this property may be accompanied by moderately high distortion, which is uniform across the frequency range. The output of the Type 1309-A is constant within ±2% over its whole frequency range and is typically within ±0.5% (see Figure 4). It is stable within ±0.2% for one hour, typically, under normal laboratory conditions and after warm-up.

One position on the step attenuator connects the high-speed Schmitt circuit to the sinusoidal oscillator. Symmetrical, positive-going square waves with a rise time of less than 100 nanoseconds

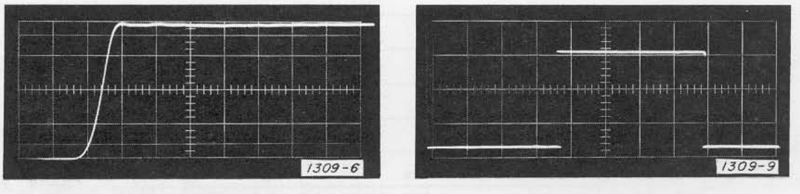


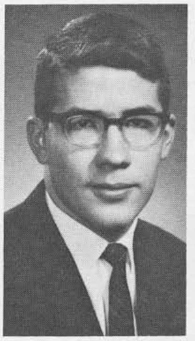
Figure 5. (Left) Leading edge of 10-kHz square wave into 50-ohm load. Horizontal scale: 50 ns/div. (Right) Direct-coupled 10-Hz square wave has flat top. Horizontal scale: 10 ms/div.

into 50 ohms are then available at the output terminals. This rise time is typically 40 nanoseconds into 50 ohms at full output, which corresponds to the response time of an amplifier with greater than 10-MHz bandwidth. (See Figure 5.) This output is greater than 5 volts, peak-to-peak, open-circuit, and is dc-coupled through the 20-dB attenuator, so that the waveform is flat-topped even at the lower frequency limit of 10 Hz.

SYNCHRONIZATION

This oscillator has a combination input/output synchronization capability similar to that described for the Type 1310-A 2-c to 2-Mc Oscillator[1], and its usefulness is enhanced by the high purity of the oscillator output. The sync output is greater than 1.5 volts, open-circuit, behind 12 kilohms and is in phase with the normal front-panel output. This output is particularly convenient for triggering counters and tone-burst generators, etc, when the attenuator output is set at a very low level. Because this output is always connected to the sinusoidal oscillator, both square-wave and sinusoidal outputs are available simultaneously. The square waves expand the uses of the synchronizing capability by providing an output waveform and amplitude that are independent of the input waveform. Figure 6 shows a further use of the combined synchronization and square-wave functions.

[1] *Ibid.*



Robert E. Owen received his B.E.E. from Rensselaer Polytechnic Institute in 1961 and his M.S.E.E. from Case Institute of Technology in 1963. During his student career, he was employed summers by Dresser Electronics and Boonton Radio Corporation. He came to General Radio as a development engineer in 1963. His field is electrical networks, both active and passive.

APPLICATIONS

The variety of uses for an oscillator with this frequency range and types of waveforms is almost unlimited. Its purity of waveform and range of available output level, however, make it particularly valuable for laboratory design and measurement use. As an example of its versatility, it can be used in audio amplifier measurements as a source:

with a wave analyzer to measure hum, noise, and harmonic distortion;

with another oscillator and a wave analyzer to measure intermodulation distortion;

with a tone-burst generator and oscilloscope to measure overload recovery and peak power output;

with an oscilloscope to measure transient response;

with a wattmeter to measure power output; and

with a voltmeter to measure frequency response.

— R. E. Owen

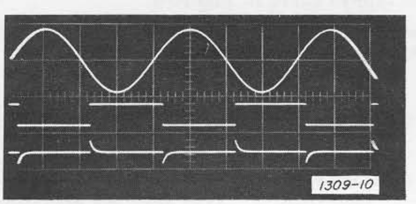


Figure 6. Using phase-shift capability of synchronized oscillator to get variable time delay pulses. (Top) sinusoidal input to oscillator synchronization jack. (Middle) Square-wave output with adjustable phase. (Bottom) Differentiated square wave (pulse) that has been adjusted to follow zero crossing of input sine wave by approximately 20°.

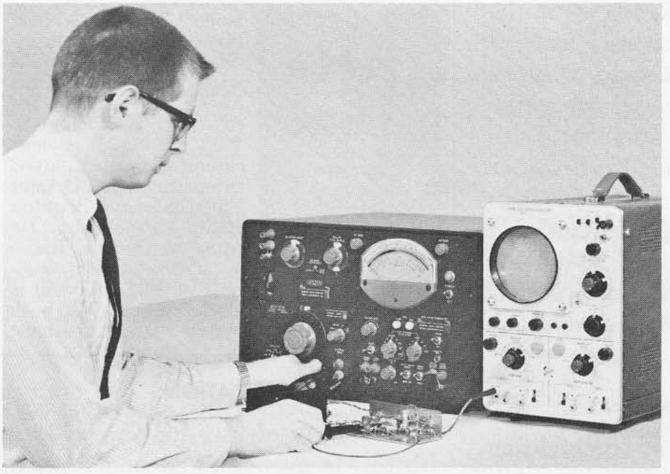


Figure 7. The Type 1309-A Oscillator and associated equipment used for testing of an audio amplifier.

SPECIFICATIONS

FREQUENCY

Range: 10 Hz to 100 kHz in four decade ranges.

Control: Continuously adjustable main dial covers range in 1 turn, vernier in 4¼ turns.

Accuracy: ±2%.

Synchronization: An external reference signal can be introduced through phone jack to phase-lock oscillator. One-volt input provides ± 3% locking range. Frequency dial can be used for phase adjustment.

OUTPUT

Sine Wave

Power: 10 mW into 600- Ω load.

Voltage: 5.0 V ±5% open circuit.

Impedance: 600 Ω. One terminal grounded.

Control: Minimum of 20-dB continuously adjustable and 60-dB step attenuator (20 ± 0.2 dB per step). Also, a zero-volts output position with 600-Ω output impedance maintained.

Distortion: Less than 0.05% from 200 Hz to 10 kHz, increasing to less than 0.25% at 10 Hz and 100 kHz open circuit or 600 Ω. See Figure 3.

Frequency Characteristic: ±2% over whole frequency range for loads of 600 Ω or greater. See typical curve in Figure 4.

Hum: Less than 50 μV regardless of attenuator setting. (0.001% of full output.)

Synchronization: High-impedance (12 kΩ), constant amplitude output of approximately 1.5 volts for use with external counter, for triggering an oscilloscope, or for synchronizing other oscillators.

Square Wave

Voltage: Greater than +5 V, peak-to-peak, open-circuit. Dc-coupled output.

Impedance: 600 Ω.

Rise Time: Under 100 ns into 50 Ω. Typically 40 ns at full output.

Control: Minimum of 20 dB continuously adjustable attenuator only.

Symmetry: ±2% over whole frequency range.

GENERAL

Terminals: Two Type 938 Binding Posts, one grounded.

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

Accessories Available: Type 1560-P95 Adaptor Cable (telephone plug to Type 274-M Double Plug) for connection to synchronizing jack, relay-rack adaptor set.

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

Mounting: Convertible-bench cabinet.

Dimensions: Width 8¼, height 6, depth 8⅛ inches (210 by 155 by 210 mm), over-all.

Net Weight: 6¾ lb (3.1 kg).

Catalog Number	*Description*	*Price in USA*
1309-9701	**Type 1309-A Oscillator, 10 Hz-100 kHz**	**$325.00**
1560-9695	**Type 1560-P95 Adaptor Cable**	**3.00**
0480-9638	**Type 480-P308 Rack-Adaptor Set**	**7.00**

GENERATION OF SINE-SQUARED PULSES WITH THE TONE-BURST GENERATOR

Sine-squared, or raised-cosine, pulses are useful in testing broadband transmission systems and, in particular, in video-bandwidth tests of television systems.[1] The sine-squared pulse, for instance, resembles very closely the electrical pulse from a television camera corresponding to a scanned white line. The spectrum envelope of a sine-squared pulse is shown in Figure 1 with that of a rectangular pulse for comparison. Above twice the fundamental, the sine-squared pulse has no components of appreciable magnitude. Below that frequency, however, its spectrum resembles very closely that of the rectangular pulse.[2]

Sine-squared pulses can easily be produced by the TYPE 1396-A Tone-Burst Generator. By means of the gating controls, the interval between pulses can be set at 1, 3, 7, 15, 31, 63, or 127 periods, or from 1 millisecond to 10 seconds with a timed control.

To generate sine-squared pulses with the Tone-Burst Generator, proceed as follows:

1. Connect a $\frac{10}{f_{HZ}}$ -μF capacitor between upper terminals of the SIGNAL INPUT and TIMING INPUT pairs.

2. Apply a signal of the desired frequency, f, to the SIGNAL INPUT terminals.

3. Set both GATE DURATION switches to 2.

4. Set CYCLE COUNTS switch to MINUS ONE. This produces one-cycle bursts similar to those of Figure 2 as seen on an oscilloscope.

5. By means of the SLOPE switch and TRIGGER LEVEL control, change the phase of the pulse so that gating occurs at peak points, as in Figure 3.

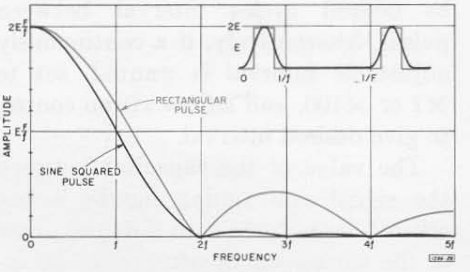
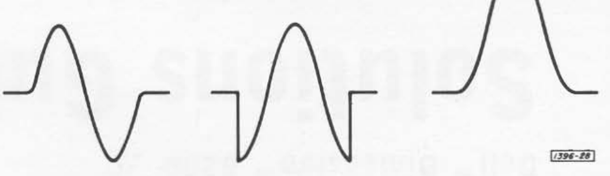


Figure 1. Spectrum envelopes of a rectangular pulse and a sine-squared pulse.

[1] Nelson, Joseph E, "Television and Sine-Squared Testing," *Tektronix Service Scope*, April 1964.
[2] Colin Cherry, *Pulses and Transients in Communication Circuits*, Chapman and Hall, Ltd., London, 1949, pp 175–181.



(Left) Figure 2. One-cycle sine-wave burst. (Center) Figure 3. One-cycle burst gated at peak. (Right) Figure 4. Sine-squared pulse.

6. Add a dc voltage source in series with the input signal and adjust this voltage to remove the steep parts of the waveform and to produce the sine-squared pulse of Figure 4.

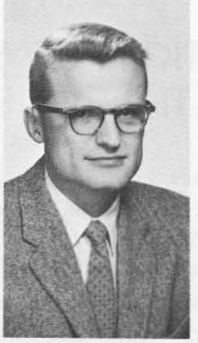
7. Set GATE DURATION-CLOSED switch to desired cycles interval between pulses. Alternatively, if a continuously adjustable interval is wanted, set to ×1 or ×100, and adjust TIMED control to give desired interval.

The value of the capacitor between the signal and timing inputs is not critical. Its purpose is to shift the phase of the timing signal relative to the input so that when switching occurs at the peak points the input circuits are working at level somewhat below peak. This ensures more reliable operation.

The output amplitude must be limited to 7 volts, peak-to-peak, because the input is dc-coupled. The output has a dc component, which can be blocked by a coupling capacitor.

Figure 5 is an oscillogram of typical sine-squared pulses generated in the manner described above. The fundamental input frequency is approximately 2.5 kHz, and the interval between pulses is 7 periods.

— J. K. SKILLING



James K. Skilling is a 1953 graduate of the University of California at Berkeley, with a BS in Electrical Engineering. He received his MS from Johns Hopkins University in 1963. He has been a junior engineer at Douglas Aircraft and an instructor in electronics at the U. S. Naval Academy. Since 1959 he has been a development engineer at General Radio, specializing in pulse techniques and circuits.

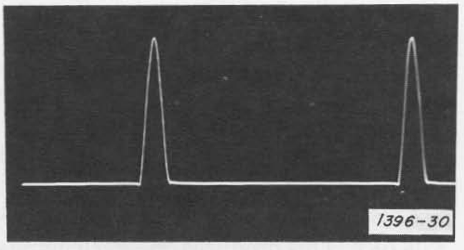


Figure 5. Typical sine-squared pulse train produced with the Tone-Burst Generator.

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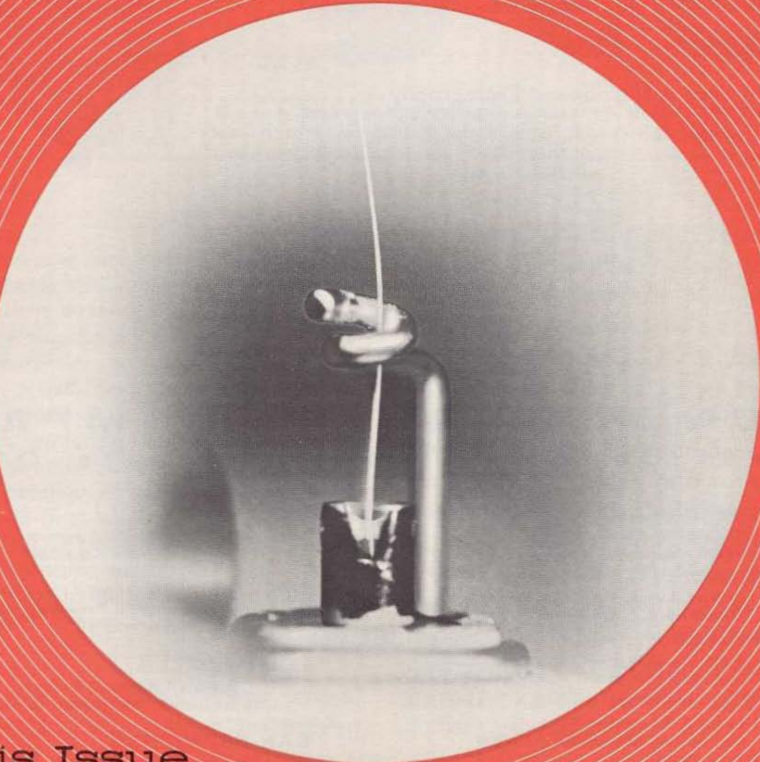
Photographed at 250,000 RPM



This Issue

New STROBOTAC® electronic stroboscope

Stroboslave · Photoelectric Pickoffs

VOLUME 40 · NUMBER 4 / APRIL 1966

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REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

Flash!...a new STROBOTAC®
electronic stroboscope

The new Strobotac and optional accessories. Next to the Strobotac, in leather case, are battery and charger. Above right are battery power cable and extension lamp. The Strobotac itself sits on the high-intensity-flash capacitor.

> The new high-speed Strobotac® electronic stroboscope described in this issue extends the speed range for stroboscopic viewing and speed measurement to over 1,000,000 rpm. In addition, the versatility of the stroboscope is enhanced by three optional accessories: a rechargeable battery for operation independent of ac power lines, a plug-in High-Intensity-Flash Capacitor for extra-bright flashes for photographic applications, and an extension lamp for access to hard-to-reach areas.

The electronic stroboscope has always been a spectacular instrument, its optical wizardry as fascinating as it is useful. The newest member of GR's STROBOTAC® family of stroboscopes follows the tradition. Its ability to flash 150,000 times a minute means that it can be used for speed measurement and stroboscopic observation of the fastest existing motors and machines, even those in the million-rpm class. Accompanying this flashing rate are two other important features: battery as well as ac operation and an accessory plug-in High-Intensity-Flash Capacitor that boosts the light output tenfold for photographic applications. The battery-power option is sure to bring cheers from thousands of veteran strobe users whose operating radii have been the lengths of their extension cords.

Shown on our cover is a Leesona false-twist spindle, used to put stretch in textile yarn. What makes the picture cover-worthy is the fact that the spindle was rotating at 250,000 rpm when it was photographed. Yet no extraordinary photographic equipment was required to make the shot; in fact, what is shown is just what an observer would have seen at the time the photo was taken. The trick, of course, is stroboscopic light, in this instance from our new Type 1538-A Strobotac®, flashing 125,000 times a minute.

Goodbye, extension cords! Textile trouble-shooter carries his own power, new Type 1538-P3 Battery and Charger, in shoulder-slung leather case.

The new Type 1538-A Strobotac is an addition to the line, not a replacement. The popular Type 1531-A Strobotac will remain available for those who do not need the extra capabilities of the Type 1538-A. The table below summarizes the difference between the two models.

APPLICATIONS

The flashing-rate limit of the new Strobotac (150,000 fpm) by no means states the upper speed limit of the instrument's usefulness. Simple harmonic relationships extend this limit to well over a million rpm. Thus high-speed dentists' drills, textile machinery (see cover), and practically anything that moves cyclically, no matter how fast, are now subject to stroboscopic observation and measurement. It's true that the use of harmonics can extend the effective range of the slower-speed Type 1531-A into the hundred-thousand-rpm area, but, as the device speed gets higher, the subharmonics come closer together on the flashing-rate dial. Then it becomes more difficult to identify them, especially if the device speed wanders. But, with the new Strobotac, even a million-rpm measurement presents no problem.

ACCESSORIES

The nickel-cadmium battery, with an automatic charger, is available as an optional accessory. A fully charged battery will power the Strobotac for about eight hours of normal operation. The battery recharges overnight from a power line. Or it can be left on charge when not used, so that it will always be ready. The Strobotac can be operated directly from an ac power line if it is more convenient.

Another important new accessory is the Type 1538-P2 Extension Lamp, a lamp-and-reflector assembly, identical to that on the Strobotac, with a six-

	Type 1531-A	*Type 1538-A*
Flashing-rate range	110-25,000 fpm	110-150,000 fpm
Speed-measurement range	to 250,000 rpm	to above 1 million rpm
Flash duration (on high to low speed ranges)	0.8, 1.2, 3 μs	0.5, 0.8, 1.2, 3 μs
Battery-operation option	no	yes
High-Intensity-Flash Capacitor option	no	yes
Extension-lamp option	no	yes
Output trigger	600-to-800-V negative pulse	6-V positive pulse
External triggering	Contact opening, 6-V, p-to-p, signal (2-V, rms, sine wave)	Contact closure, 1-V positive pulse, 0.35-V, rms, sine wave
Price	$295.00	$465.00

The Principle of the Stroboscope

A stroboscope is an instrument that permits periodic observation of a moving object in such a way as to create the optical illusion of slow or stopped motion. The electronic stroboscope is essentially a flashing light that provides periodic illumination of a cyclically moving object and thus produces the stroboscopic illusion.

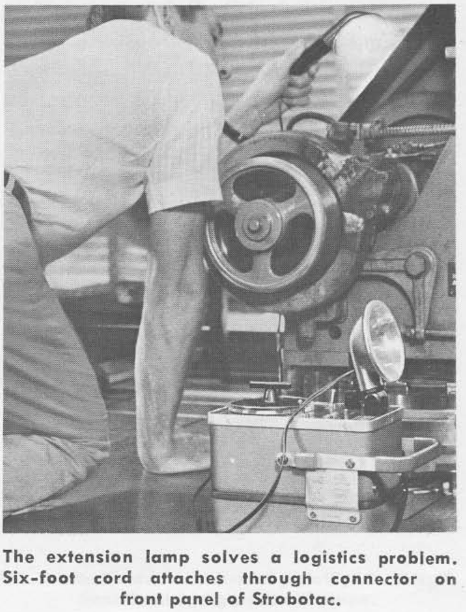
The flashing rate of the stroboscope is controlled by an electronic oscillator and is adjustable over a very wide range. If it is set to flash at, say, 1800 times a minute and if its light is used to illuminate a fan rotating at 1800 rpm, each successive flash will occur with the fan in the same position, and the fan will appear motionless. If the flashing rate is offset very slightly from the fan speed, the flashes will come at successively earlier or later parts of the fan's cycle, producing a slow-motion replica of the actual high-speed motion.

The stroboscope is also widely used as a tachometer. The flashing rate is adjusted to produce the stopped-motion effect, and the speed of the device under study is read on the dial of the flashing-rate control. Especially significant is the fact that this kind of tachometry requires no physical contact with the device. In photography, the stroboscope's microsecond flash (much faster than that of conventional speed lights) "stops" almost anything, no matter how fast it moves, for the camera.

foot cord and plug. The plug mates with a connector on the front panel of the STROBOTAC. Thus the lamp can be operated in spaces too small for the complete instrument or can be mounted in test chambers and controlled from a safe distance.

The light intensity of the STROBOTAC is more than adequate for many photographic applications. Still, there are times when, either because of the extremely high speed of the object being photographed or an unavoidably high ambient light level, a brighter flash is needed. Then the photographer can connect the new Type 1538-P4 High-Intensity-Flash Capacitor to the base of the STROBOTAC. With this accessory connected, one can produce a single flash of great brilliance (44 million beam candles) and short duration (8 microseconds).

Other accessories useful with the STROBOTAC are GR's photoelectric pickoffs (see page 11), flash-delay unit, surface-speed wheel, and two stroboscopes that can be controlled by the STROBOTAC: the STROBOLUME and the STROBOSLAVE (see page 9). These instruments and accessories constitute by far the most complete line of stroboscopic equipment available anywhere.



The extension lamp solves a logistics problem. Six-foot cord attaches through connector on front panel of Strobotac.

THE CIRCUIT OF THE NEW STROBOTAC

The sixfold increase in flashing rate of the new STROBOTAC was made possible by the development of a new strobotron tube* and of new circuits** that minimize the time required between flashes for deionization and recharging. The following is a description of the STROBOTAC circuit, with emphasis on the advances of the new model.

The strobotron flash tube comprises two main electrodes, a cathode and an anode, separated by 3/8 inch in an envelope filled with xenon gas at a pressure of one-half atmosphere. A specially designed capacitor acts as a low-impedance source to supply 800 to 1000 volts across these electrodes. The gas, however, remains nonconducting until a 5000-volt pulse is applied to trigger wires between these main electrodes. This pulse ionizes the gas, causing up to 1000 amperes to flow. The peak power of almost a million watts generates an intense flash of white light of 15 million beam candles.

* U. S. Patent No. 2,977,508.
** Patent applied for.

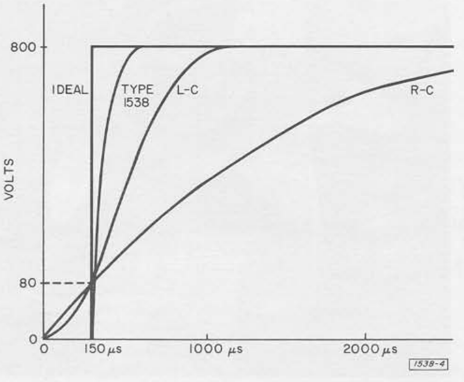


Figure 1. Voltage-vs-time characteristics of various charging circuits.

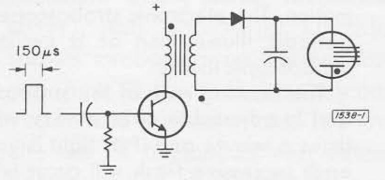


Figure 2. The charging circuit of the new Strobotac.

After the tremendous pulse of light, the tube requires about 150 microseconds to deionize. The voltage applied across the tube must remain under 80 volts during this deionization time, or continuous conduction, called "hold-over," will result. This necessary deionization period limits the maximum flashing rate of the stroboscope. Figure 1 illustrates the problem. The curves labeled *R-C* and *L-C* represent the effects of charging the capacitor through a resistor and an inductor, respectively. The slopes required to keep the voltage below the 80-volt deionization level impose delays in reaching the firing level, which in turn would restrict the maximum flashing rates to 24,000 and 54,000 flashes per minute, respectively, for the particular tube and voltages used in the TYPE 1538-A.

The answer to this problem is to hold the voltage to zero for the deionization period and then to raise it quickly to the firing level.

The new circuit shown in Figure 2 provides an almost ideal charging curve (labeled "TYPE 1538" in Figure 1). During the 150-microsecond deionization time after the strobotron has flashed, the transistor, acting as a switch, is saturated and the transformer primary current increases, storing en-

Harmonic techniques extend the usefulness of the Strobotac over the entire audio-frequency range. Engineer here is watching speaker motion through microscope while adjusting flash-delay unit to provide phase control.

ergy in the transformer core. The voltage induced in the secondary winding during this build-up is blocked by the diode rectifier, and no voltage appears across the capacitor and strobotron tube. At the end of this 150-microsecond interval, the transistor is switched off, and the primary current goes to zero. The collapsing magnetic field generates a reverse-polarity voltage in the secondary, causing the diode to conduct and the stored energy to be transferred to the capacitor. When the energy in the transformer is zero, the current again reverses and the diode opens, leaving all the stored energy in the capacitor. This transfer can be made as fast as one wishes, and the flashing rate can therefore be made to approach the theoretical maximum. The most important result of all this is an increase in flashing rate to almost the theoretical maximum. This was the main objective of the circuit development, but the fallout was almost as valuable.

The transfer of energy from the power supply to the intermediate storage inductor and then resonantly to the discharge capacitor can be made with an efficiency approaching 100%. In the conventional RC charging circuit, however, one half the available energy is dissipated in the charging resistor regardless of the value of the resistor (including zero ohms). The use of inductive charging therefore saves the power ordinarily dissipated in the charging resistor and makes battery operation practical. Moreover, the use of a transformer as the inductive element permits use of a low-voltage transistor circuit to generate the high voltage required by the strobotron tube.

A block diagram of the STROBOTAC is shown in Figure 3. A transistorized RC oscillator sets the flashing rate. Once each cycle, a transistor trigger circuit

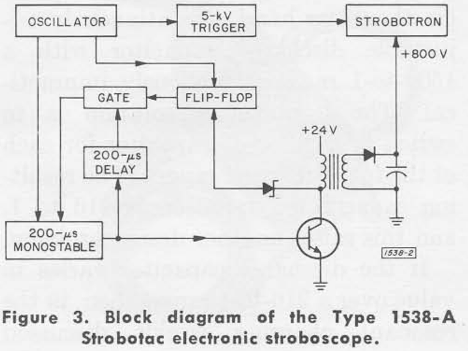


Figure 3. Block diagram of the Type 1538-A Strobotac electronic stroboscope.

M. C. Holtje received his BSEE and MSEE degrees from the Massachusetts Institute of Technology in 1950. He then joined General Radio Company as a Development Engineer and in 1960 became Leader of the Industrial Group. He is a Senior Member of the IEEE and has served on several IEEE standards committees.

generates a 5-kilovolt, 5-microsecond pulse to trigger the strobotron tube. In the time between these pulses, the main discharge capacitor must be recharged. The monostable circuit, triggered by the oscillator, generates a 200-microsecond pulse, which saturates the transistor switch, storing energy in the transformer and allowing the strobotron to deionize. At the end of the 200-microsecond pulse, enough energy has been stored to charge the capacitor resonantly to 800 volts in an additional 200 microseconds. Thus, a maximum flashing rate of 2500 per second is possible.

The average light output of a stroboscope varies directly with flashing rate and discharge capacitance. The exceptionally wide flashing-rate range of the TYPE 1538-A (1500 to 1) would mean a drastic variation in light output if only one discharge capacitor were used. On the other hand, a continuously adjustable discharge capacitor with a 1500-to-1 range is obviously impractical. The compromise solution is to switch in a different capacitor for each of the four 6:1 speed ranges. The resulting capacitance variation is 216 to 1, and this raises another design problem.

If the discharge capacitor varies in value over a 216-to-1 range, then, in the resonant charging circuit discussed earlier, either the inductance must also vary by a factor of 216 or the current must vary by a factor of $\sqrt{216}$ to supply sufficient energy per cycle. Large coils and 30-ampere currents were both unappealing, so another approach was found.

On the lower-speed ranges, where the discharge capacitance is higher, the energy stored in the transformer is insufficient to produce the 800-volt firing potential. On these ranges the 200-microsecond delay following the monostable circuit generates a trigger pulse 200 microseconds after the end of the monostable pulse to retrigger the monostable circuit. Thus, a single pulse from the oscillator starts a train of 200-microsecond pulses in the monostable circuit and its delay loop. Each of these pulses stores energy in the inductor, and this energy is repeatedly transferred to the capacitor during the time between pulses. Each pulse raises the capacitor voltage in a small step as shown in Figure 4. This process continues until the capacitor is charged to 800 volts. At each step, a voltage pulse equal to the capacitor voltage divided by the transformer turns ratio appears across the Zener diode on the transformer primary. When the capacitor

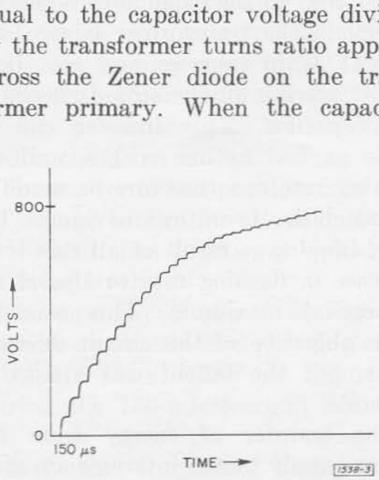


Figure 4. Curve showing step voltage buildup on charging capacitor.

reaches 800 volts, the Zener diode voltage is exceeded and the flip-flop closes the gate, breaking the feedback loop and ending the pulse train started by the oscillator. While this multiple-cycle resonant-charging technique used on the lower ranges requires more time than the single-cycle charge, a correspondingly longer time is available in which to recharge the capacitor.

— M. C. Holtje

SPECIFICATIONS

Flashing-Rate Range: 110 to 150,000 flashes per minute in four direct-reading ranges: 110 to 690, 670 to 4170, 4000 to 25,000, and 24,000 to 150,000 rpm. Speeds to over 1 million rpm can be measured.

Accuracy: $\pm 1\%$ of reading on all ranges after calibration against line frequency.

Flash Duration: Approximately 0.5, 0.8, 1.2, and 3 μs for high-to-low speed ranges, respectively, measured at ⅓ peak intensity; for single flashes with Type 1538-P4 High-Intensity-Flash Capacitor, 8 μs.

Peak Light Intensity: Typically 0.16, 1, 5, and 15 million beam candles (0.16, 1, 5, and 15×10^6 lux measured at 1 meter distance at the beam center) for high-to-low speed ranges, respectively; 44 million beam candles for single flash, with Type 1538-P4 High-Intensity-Flash Capacitor.

Reflector Beam Angle: 10° at half intensity points.

Output Trigger: Greater than 6-V positive pulse behind 400 Ω.

External Triggering: Either a switch closure across the input jack terminals, a 1-V, peak, positive pulse, or a 0.35-V, rms, sine wave down to 100 Hz increasing to 3.5 V, rms, at 5 Hz.

Power Required: 100 to 125 or 195 to 250 V, 50 to 400 Hz, 15 W or 20 to 30 V dc, 12 W.

Accessories Supplied: Adjustable neck strap, phone plug for input and output jacks, spare fuses.

Accessories Available: Type 1538-P2 Extension Lamp, Type 1538-P3 Battery and Charger, Type 1538-P4 High-Intensity-Flash Capacitor, Type 1531-P2 Flash Delay, Types 1536-A Photoelectric Pickoff (for use with Flash Delay), Type 1537-A Photoelectric Pickoff, and Type 1539-A Stroboslave.

Mounting: Flip-Tilt Case.

Dimensions: Width 10⅝*, height 6⅝, depth 6⅛ inches (270 by 170 by 160 mm), over-all.

Net Weight: 7¼ lb (3.3 kg).

Shipping Weight: 10 lb (4.6 kg).

* Includes handle.

Catalog No.	*Description*	*Price in USA*
1538–9701	**Type 1538-A Strobotac® electronic stroboscope**	**$465.00**
1538–9601	**Type 1538-P1 Replacement Strobotron Lamp**	**15.00**
1538–9602	**Type 1538-P2 Extension Lamp**	**55.00**
1538–9603	**Type 1538-P3 Battery and Charger, with case**	**225.00**
1538–9604	**Type 1538-P4 High-Intensity-Flash Capacitor**	**75.00**

Introducing the STROBOSLAVE

Because it is widely used as a tachometer, the conventional stroboscope includes an oscillator and associated electronic circuits necessary to adjust the flashing rate over a wide range. A purchaser who wants only to make stopped-motion observations or photographs, with flashing rate under external control, thus is forced to pay for a capability that is of no use to him. Therein lies the suggestion that a sim-

Using the Stroboslave as a diagnostic tool. The reflector has been slipped off so that the strobe lamp can probe the innards of the addressing machine. A photoelectric pickoff keeps the flashes synchronized with the machine, while a flash-delay unit allows the operator to scan the motion throughout its cycle.

ple stroboscope, designed solely for external control, is needed. Enter the STROBOSLAVE.

The TYPE 1539-A STROBOSLAVE is a small, inexpensive stroboscope, in most respects similar to the STROBOTAC. The chief difference is that the STROBOSLAVE has no internal flashing-rate control. This means that it cannot serve as a tachometer. For certain motion studies and for high-speed photography, however, the STROBOSLAVE is every bit as useful as its more sophisticated brethren.

It has, in fact, several advantages over the larger stroboscopes (in addition to the price differential). Its lamp, at the end of a five-foot cable, can be either attached to the case or maneuvered close to the cbject being observed. The case itself is small enough (2½ by 5⅜ by 4⅛ inches) to be permanently mounted on such machines as textile looms, production tools, and printing presses, where continuous stroboscopic monitoring may cut costs substantially by showing up defects in material or goods produced.

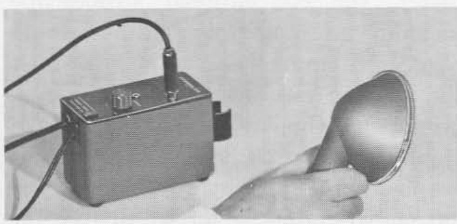
The STROBOSLAVE can be triggered by a STROBOTAC, a TYPE 1537-A Photoelectric Pickoff, a TYPE 1535-B Contactor, or any device capable of supplying a contact closure or a positive pulse of a least 2 volts peak. An extremely useful combination is the STROBOSLAVE, TYPE 1531-P2 Flash Delay, and TYPE 1536-A Photoelectric Pickoff. With such a setup, one can "stop" motion, observe it throughout its cycle, and synchronize a camera shutter with the flash.

Light duration, intensity, and flashing-rate range are all the same as for the TYPE 1531-A STROBOTAC. The STROBOSLAVE operates from standard ac power lines.

— M. C. HOLTJE

SPECIFICATIONS

Flashing-Rate Ranges: 0 to 700, 0 to 4200, 0 to 25,000 flashes per min on high-, medium-, and low-intensity ranges, respectively.
Flash Duration: Approx 0.8, 1.2, and 3 μs, measured at ⅓ peak intensity, for the low-, medium-, and high-intensity ranges, respectively.
Peak Light Intensity: Typically 0.6, 3.5, and 11 million beam candles (0.6, 3.5, and 11×10^6 lux measured at 1-m distance at the beam center), for low-, medium-, and high-intensity ranges, respectively. For single flash, 18 million beam candles.
Reflector Beam Angle: 10° at half-intensity points.

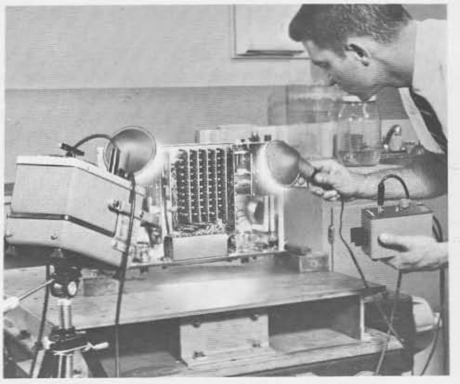


Lamp, at end of five-foot cable, can be held in hand as shown here or attached to case as shown on page 9.

External Triggering: Either a switch closure across the input jack terminals or a 2-V (peak) positive pulse.
Power Required: 100 to 125 or 195 to 250 V, 50 to 400 Hz, 16 W (max) at 115 V.
Accessories Supplied: Phone plug for input, mounting bracket.
Accessories Available: TYPE 1537-A Photoelectric Pickoff, TYPE 1531-P2 Flash Delay (with a TYPE 1536-A Photoelectric Pickoff).
Dimensions: Width 2½, height 5⅜*, depth 4⅛ inches (64 by 215 by 105 mm), over-all.
Net Weight: 2¾ lb (1.3 kg).
Shipping Weight: 8 lb (3.7 kg).

* Without lamp attached.

Catalog Number	*Description*	*Price in USA*
1539-9701	**Type 1539-A Stroboslave**	**$165.00**
1531-9604	**Type 1531-P4 Trigger Cable** (for use with Type 1531-A Strobotac)	**15.00**



Putting extra light on the subject. With an electronic frequency counter going through a vibration shake-table test, a Stroboslave is used to supplement the light from the Strobotac. Smaller strobe is triggered directly from output of Type 1538-A Strobotac.

PHOTOELECTRIC PICKOFFS

There are now two photoelectric pickoffs in the General Radio line: the TYPE 1536-A,[1] which includes a light source, and the new TYPE 1537-A, which does not. The latter is the less expensive and is the recommended pickoff for supplying a triggering pulse directly to a TYPE 1538-A STROBOTAC or a TYPE 1539-A STROBOSLAVE, assuming that a source of bright light is available. (A 1.1-watt No. 330 14-volt pilot lamp at ½ inch is adequate. If the light source cannot be placed near the object, a reflector and lens can be used to focus the light.) Where no adequate light source is available, where extra sensitivity is needed, or where the TYPE 1531-P2 Flash Delay is desired for phase control, the TYPE 1536-A Pickoff is recommended. The TYPE 1536-A, with flash-delay, is the recommended pickoff for use with the TYPE 1531-A STROBOTAC.

[1] "Using a Photocell Where It Counts," *General Radio Experimenter*, October, 1962.

The Type 1537-A Photoelectric Pickoff is identical in appearance with the Type 1536-A. Only difference is that the Type 1536-A contains a light source, the Type 1537-A does not.

Supplied with the TYPE 1537-A Pickoff, as with the TYPE 1536-A, are a C-clamp and a magnet, for easy mounting on a variety of surfaces, and two rolls of tape, one black and one silver. Pieces of this tape can be affixed to the edge of a shaft or wheel to produce alternately reflective and nonreflective areas to trigger the photocell, the choice between black and silver tape depending on whether the surface is itself reflective or nonreflective.

SPECIFICATIONS

Operating Rate: Greater than 2500 pulses/s.
Sensitivity: Effective irradiance must be at least 6.0 mW/cm² to switch on, less than 0.6 mW/cm² to switch off, at 1 micron wavelength.
Power Required: 3 to 25 V dc; 0 to 100 μA depending on operating rate. Power is supplied by instrument with which it is used.
Accessories Supplied: 10-ft roll of ⅜-in black tape, 10-ft roll of ⅜-in silver tape, carrying case.
Mounting: C-clamp (capacity 1 5/16 in, flat or round) or 1½-in magnet, both supplied.
Dimensions: Pickoff head, 11/16-in dia, 2 in long. Linkage consists of two 5/16-in diameter stainless-steel rods, 6 and 6¼ in long, and adjustable connecting clamp. Cable is 8 ft long, terminated in phone plug.
Net Weight: 1½ lb (0.7 kg).
Shipping Weight: 4½ lb (2.1 kg).

Catalog Number	*Description*	*Price in USA*
1537-9701	**Type 1537-A Photoelectric Pickoff**	**$65.00**

SEMINAR IN HIGH-SPEED PHOTOGRAPHY

A one-week seminar on the scientific and engineering uses of high-speed photography will be held at the Stroboscopic Light Laboratory of the Massachusetts Institute of Technology, July 25 to 29. Mornings will be devoted to theory and demonstrations, afternoons to laboratory practice. Subjects to be covered include pulsed stroboscopic lighting, optical high-speed cameras, Kerr cells, Faraday shutters, image converters, etc. For more information, write to:

Office of the Summer Session
MIT, Room E19-356
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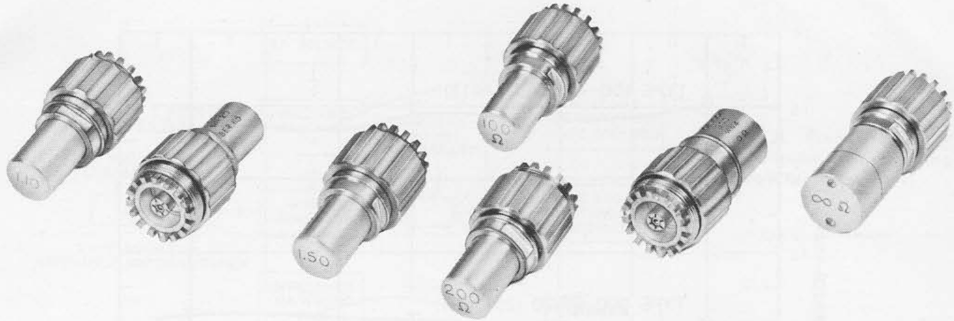


Figure 1. View of the terminations described in this article.

Calibration Standards for Precision Coaxial Lines

TYPE 900-WR STANDARD MISMATCHES

In a coaxial line, as in any uniform, distributed-parameter system, it is the terminating impedance that determines the reflection that occurs when an electromagnetic wave traveling down the line reaches the far end. From the magnitude and phase of the reflection, the nature of the terminating impedance can be deduced, and coaxial-line measurement devices are based on this principle.

As with any measuring system, standards of calibration are necessary to ensure accuracy of measurement. The standard terminations described here fill this need and are recommended for the calibration of slotted lines, bridges, impedance plotters, fixed- and swept-frequency reflectometers, and time-domain reflectometers.

These broadband mismatches are standards of VSWR, for use in the calibration of slotted-line systems, reflectometers, and other VSWR and reflection-coefficient measuring devices The TYPES 900-WR110, -WR120 and -WR150 Standard Mismatches introduce VSWR's of 1.1, 1.2, and 1.5, respectively, and each of these units exhibits nearly uniform VSWR characteristics from dc to 8.5 GHz. (See Figure 2.) Each unit comprises a 50.0-ohm GR900 Precision Coaxial Connector, a low-reflection continuous transition, and a precision cylindrical resistor. The position at which the mismatch is introduced into the 50.0-ohm system is approximately 4 cm behind the reference plane of the GR900 Connector.

The terminating elements are highly stable, deposited-metal-film resistors with dc resistances of 45.45, 41.67 and 33.33 ohms, respectively, ±0.3%. Calibration charts supplied with each unit give the measured resistance at dc and at five points in the frequency band. NBS calibration services are also available to 4 GHz with uncertainties of VSWR measurement from ±0.005 at 1 GHz to ±0.010 at 4 GHz.

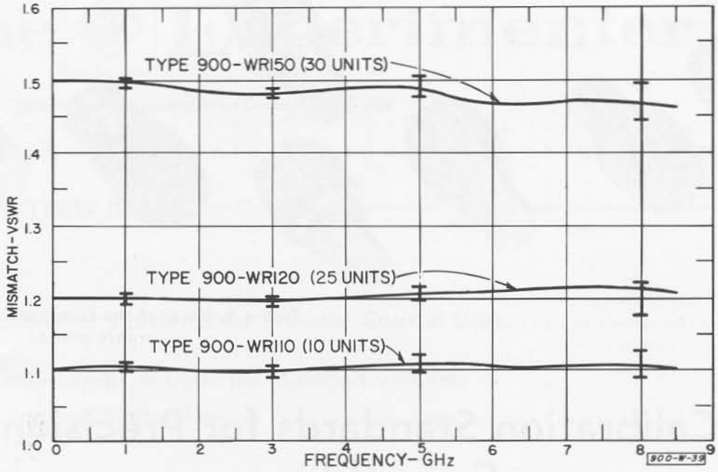


Figure 2. Average mismatch VSWR of sample lots. Spreads in the measured data are shown at 1, 3, 5, and 8 GHz. Measurement accuracy is better than ±0.003 for the Type 900-WR110, ±0.005 for the Type 900-WR120, and ±0.010 for the Type 900-WR150.

APPLICATIONS

Direct RF Calibration of Slotted-Line Systems

Many factors contribute to inaccuracy in the measurement of VSWR with a slotted-line system. Uncertainty in the detector response law, calibration accuracy of the indicating instrument, residual VSWR and probe reflections in the slotted line — all of these introduce varying effects that are dependent on the magnitude of VSWR being measured, the frequency of operation, and the nature of the instruments. The TYPE 900-WR Standard Mismatches offer a simple means of establishing directly, at the measurement frequency, the over-all system accuracy.

Figure 3 shows the standing-wave patterns of design-center mismatches at VSWR levels of 1.1 and 1.2, measured at 7 GHz with the TYPE 1640-A Slotted Line Recording System.[1]

[1] A. E. Sanderson, "A Slotted Line Recorder System," "Reference Air Lines for the GR900 Series," and "New Coaxial Tuner with Neutral Setting," *General Radio Experimenter*, January 1965.

Calibration of Frequency-Domain Reflectometers

The TYPE 900-WR Standard Mismatches are well suited for the VSWR calibration of swept-frequency reflectometers and impedance plotters based on directional couplers, hybrid junctions, magic tees, or rf bridges. Calibration through GR900 Connectors offers the greatest accuracy; however, the use of the TYPE 900-Q Adaptors makes it possible to calibrate measuring devices

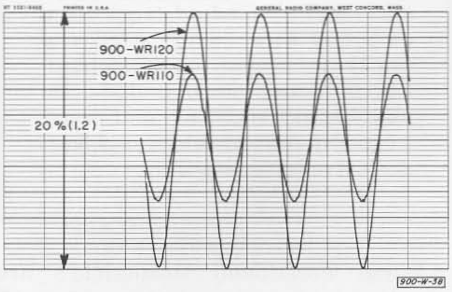


Figure 3. Standing-wave patterns of Types 900-WR110 and -WR120 Standard Mismatches as measured at 7GHz with a Type 1640-A Slotted Line Recording System.

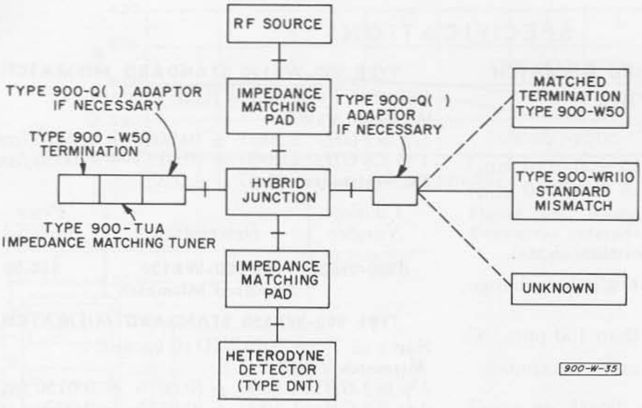


Figure 4. Fixed-frequency reflectometer for measuring VSWR's below 1.1.

equipped with many other types of connectors.

The TYPE 900-WR110 Mismatch is particularly useful for the calibration of fixed-frequency reflectometers where the residual reflection errors of the measuring instruments are tuned out. As an example, Figure 4 is the block diagram of a fixed-frequency reflectometer system built around a hybrid junction in which the error signal is proportional to the reflection coefficient of the unknown being measured. This system provides full-scale VSWR indications of 10%, 1%, and even 0.1%.*

The calibration procedure for this system is as follows:

(a) A TYPE 900-W50 Standard Termination is connected to the unknown port of the hybrid junction and the impedance-matching tuner (TYPE 900-TUA[1]) is adjusted for a null in the detected signal. This makes the instrument residual reflection equal in magnitude to the residual reflection of the termination. Thus, the accuracy of the measurement is directly dependent on the accuracy of the termination. (Highly accurate calibration techniques for determining the termination accuracy are described by Sanderson.[2])

(b) The TYPE 900-WR110 Standard Mismatch (VSWR = 10%) is connected in place of the matched termination, and the signal level is adjusted for full-scale indication on the meter of the TYPE 1216-A Unit I-F Amplifier in the TYPE DNT Detector.

The unknown to be measured is now connected in place of the standard mismatch, and its VSWR is read directly from the linear scale of the i-f amplifier meter. If the setting of the amplifier attenuator switch is reduced by 20 dB, the full-scale VSWR becomes 1%, and, for 40-dB reduction, the full-scale VSWR becomes 0.1%. The assumption that the linear scales apply is not rigorously true, since the relationship between VSWR and reflection coefficient is not a linear one. However, for VSWR's up to 10%, the error is less than 1/20th of the indicated VSWR in percent.

---

* VSWR in percent is given by (VSWR - 1) × 100.

[1] *Ibid.*

[2] Sanderson, A. E., "*Calibration Techniques for One- and Two-Port Devices Using Coaxial Reference Air Lines as Absolute Impedance Standards,*" presented at the ISA 19th Annual Conference and Exhibit, 1964; available from General Radio Company as Technical Publication B-21.

SPECIFICATIONS

TYPE 900-WR110 STANDARD MISMATCH

Frequency Range: DC to 8.5 GHz.

Mismatch VSWR

Up to 1 GHz: 1.1000 ± (0.0055 + 0.0110 f_{GHz}).
1 to 8.5 GHz: 1.1000 ± (0.0115 + 0.0050 f_{GHz}).

DC Resistance: 45.45 Ω ±0.3%.

Leakage: Better than 130 dB below signal.

Maximum Power: 1 W with negligible change; 5 W without damage.

Temperature Coefficient: Less than 150 ppm/°C.

Dimensions: Length, 2 in (51 mm); maximum diameter, 1 1/16 in (27 mm).

Net Weight: 3½ oz (100 g).

Catalog Number	*Description*	*Price in USA*
0900-9961	**Type 900-WR110 Standard Mismatch**	**$60.00**

TYPE 900-WR120 STANDARD MISMATCH

Same as Type 900-WR110 except:

Mismatch VSWR

Up to 1 GHz: 1.2000 ± (0.0060 + 0.0120 f_{GHz}).
1 to 8.5 GHz: 1.2000 ± (0.0125 + 0.0055 f_{GHz}).

DC Resistance: 41.67 Ω ±0.3%.

Catalog Number	*Description*	*Price in USA*
0900-9963	**Type 900-WR120 Standard Mismatch**	**$60.00**

TYPE 900-WR150 STANDARD MISMATCH

Same as Type 900-WR110 except:

Mismatch VSWR

Up to 1 GHz: 1.5000 ± (0.0075 + 0.0150 f_{GHz}).
1 to 8.5 GHz: 1.5000 ± (0.0155 + 0.0070 f_{GHz}).

DC Resistance: 33.33 Ω ±0.3%.

Catalog Number	*Description*	*Price in USA*
0900-9965	**Type 900-WR150 Standard Mismatch**	**$60.00**

TYPE 900-W STANDARD TERMINATIONS

These broadband resistive terminations are standards of impedance, which can be used to calibrate swept-frequency impedance-measuring systems, impedance plotters, slotted-line systems, bridges and time-domain reflectometers.

In contrast to the TYPE 900-WR Standard Mismatches, the Standard Terminations are calibrated in phase as well as magnitude; that is, the position of the standard resistance with respect to a reference point in the connector is accurately known. These terminations, therefore, find their greatest use in the calibration of impedance-measuring systems, although they are also standards of VSWR.

The TYPES 900-W100 and -W200 Standard Terminations are 100- and 200-ohm terminating resistances for a 50.0-ohm system. The resistances introduced remain very nearly equal to their dc resistances over the frequency band from dc to 8.5 GHz, as illustrated in Figure 5. These units are similar in construction to the TYPE 900-WR Standard Mismatches.

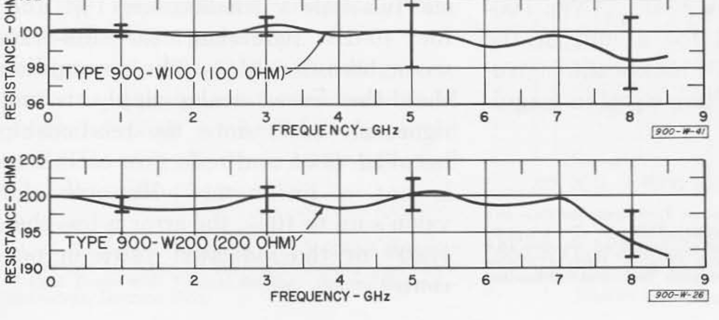


Figure 5. Average resistance of 25 units each of Type 900-W100 and Type 900-W200. Spreads in the measured data are shown at 1, 3, 5, and 8 GHz. Measurement accuracy is better than 1%.

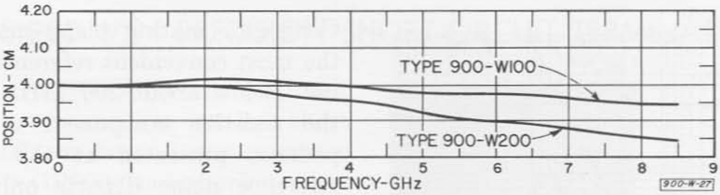


Figure 6a. Average position behind GR900 Connector reference plane at which resistance is applied for the units of Figure 5.

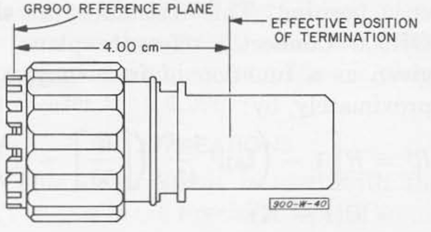


Figure 6b. Sketch showing relation of termination position and reference plane.

The reference plane at which the termination is introduced into the 50.0-ohm system is 4 cm behind the reference plane of the GR900 Connector, as shown in Figure 6. Calibration charts supplied with each unit include measured data on the position at which the resistance effectively appears in addition to the measured resistance at dc and at 5 points in the frequency band.

APPLICATIONS

Calibration of Slotted-Line and Reflectometer Systems

The Type 900-W Standard Terminations, like the Type 900-WR Standard Mismatches, are used to perform direct rf calibration of slotted-line systems. At the 100-ohm and 200-ohm levels (mismatches of 2 and 4, respectively), the errors introduced by variations in the detector-response law, uncertainties in the indicator calibration, and, most important, probe reflections in the slotted line can be appreciable. The Type 900-W Terminations permit a rapid, yet accurate, test of a system's performance, without the necessity of time-consuming check-out procedures.

Similarly, with reflectometer systems, these standard terminations provide important calibration points. Since the terminations are calibrated in both magnitude and phase, they are most useful in the calibration of complex reflection-coefficient measuring instruments such as automatic impedance plotters. Because of the phase calibration of the terminations, they can be combined with sections of precision air line to produce many known complex impedances. For example, a Type 900-W100 Termination in combination with a 6-cm air line produces (at the air-line input connector mating plane) an impedance of 40.0 − j30.0 ohms at frequencies given by

$$3\left(\frac{1+4n}{8}\right) \text{GHz and } 40.0 + j30.0 \text{ ohms}$$

at frequencies given by $3\left(\frac{3+4n}{8}\right)$GHz,

where n is zero or a positive integer. The Type 900-L Precision Air Lines and the Type 900-LZ Reference Air Lines[1] are recommended for such applications.

[1] *Ibid.*

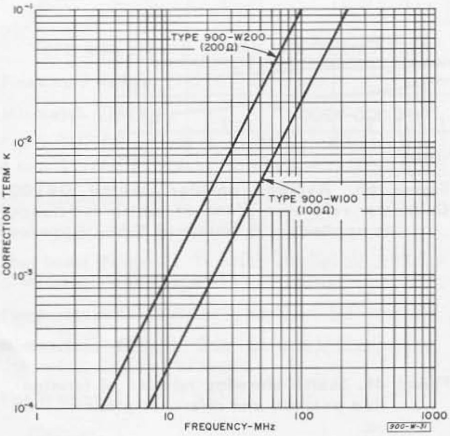


Figure 7. Correction term, K, for the 4-cm difference between the GR900 Connector reference plane and the effective position of the resistance.

Calibration of Bridges

The Type 900-W Standard Terminations are used to calibrate bridges in much the same manner as described for slotted lines and reflectometers. For some bridges, the termination reference plane 4 cm away from the GR900 Connector mating plane may not be the most convenient reference plane to use. Below about 200 MHz, however, the resistive component of the impedance presented at the connector reference plane departs only slightly from that presented at the 4-cm reference position. This resistance (at the GR900 Connector reference plane) is given as a function of frequency, approximately, by:

$$R' = R\left[1 - \left(\tan^2\frac{4\pi f}{15}\right)\left(\left[\frac{R}{50}\right]^2 - 1\right)\right]$$
$$= R(1 - \mathrm{K})$$

where R is the calibrated dc resistance of the termination in question and f is the frequency in GHz.

The correction term

$$\mathrm{K} = \left(\tan^2\frac{4\pi f}{15}\right)\left(\left[\frac{R}{50}\right]^2 - 1\right)$$

is a result of the distributed capacitance of the 4-cm length of line between the two reference planes and is plotted in Figure 7 for resistances of 100 and 200 ohms.

SPECIFICATIONS

TYPE 900-W100 100-OHM STANDARD TERMINATION

Frequency Range: DC to 8.5 GHz.
DC Resistance: 100 Ω ±0.3%.
RF Resistance
Up to 1 GHz: $100.00 \pm (0.50 + 1.00\, f_{GHz})$
1 to 8.5 GHz: $100.00 \pm (1.05 + 0.45\, f_{GHz})$
Position at Which Resistance Specification Applies
Up to 2 GHz: (4.00 ± 0.05) cm beyond the GR900 Connector reference plane.
2 to 8.5 GHz: $(4.02 - 0.01\, f_{GHz} \pm 0.05)$ cm beyond the GR900 Connector reference plane.
Leakage: Better than 130 dB below signal.
Maximum Power: 1 W with negligible change; 5 W without damage.
Temperature Coefficient: Less than 150 ppm/°C.
Dimensions: Length, 2 in (51 mm); maximum diameter, 1$\frac{1}{16}$ in (27 mm).
Net Weight: 3½ oz (100 g).

Catalog Number	*Description*	*Price in USA*
0900-9957	**Type 900-W100 100-Ohm Standard Termination**	**$60.00**

TYPE 900-W200 200-OHM STANDARD TERMINATION

Same as Type 900-W100, except:
DC Resistance: 200 Ω ± 0.3%.
RF Resistance
Up to 1 GHz: $200.00 \pm (1.00 + 2.00\, f_{GHz})$
1 to 7 GHz: $200.00 \pm (2.10 + 0.90\, f_{GHz})$
$+ 8.40$
7 to 8.5 GHz: 200.00 or
$-(8.40 + 7.20[f_{GHz} - 7])$
Position at Which Resistance Specification Applies
Up to 2 GHz: (4.00 ± 0.05) cm beyond the GR900 Connector reference plane.
2 to 8.5 GHz: $(4.04 - 0.02\, f_{GHz} \pm 0.05)$ cm beyond the GR900 Connector reference plane.

Catalog Number	*Description*	*Price in USA*
0900-9959	**Type 900-W200 200-Ohm Standard Termination**	**$60.00**

TYPE 900-WN4 PRECISION SHORT-CIRCUIT TERMINATION

The TYPE 900-WN4 Short-Circuit Termination presents a low-loss short circuit 4.00 cm beyond the reference plane of its GR900 Connector reference plane. The reflection coefficient introduced at the actual short-circuit plane is greater than 0.999, and that introduced at the connector reference plane is greater than 0.996.

APPLICATIONS

This short circuit is used with the TYPE 900-WO4 Precision Open-Circuit Termination (described below) to establish short- and open-circuit reference planes coincident within 0.02 cm over the frequency range from dc to 8.5 GHz. The reference planes so established are useful in direct impedance measurements, in loss measurements based on reflection measurements, in the calibration of reflection-coefficient measuring instruments, and, generally, in the measurement of the scattering coefficients of multiport coaxial devices.

Since its 4.00-cm reference plane coincides with those of the TYPES 900-W100 and −W200 Standard Terminations, the TYPE 900-WN4 can be used in conjunction with these terminations for the calibration of bridges, slotted-line systems, etc.

Figure 8 illustrates the calibration levels obtainable with the TYPES 900-WR, 900-W, 900-WN4 and 900-WO4 Standards. All these units are recommended for the calibration of time-domain-reflectometry systems.

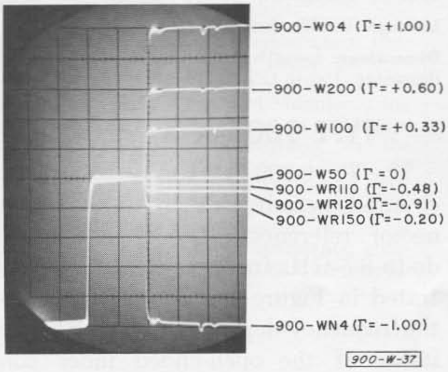


Figure 8. Multiple exposure of time-domain-reflectometer traces for the various GR900 terminations at the end of a length of 50-ohm air line.

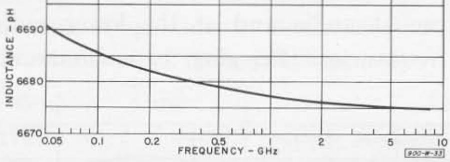


Figure 9. Inductance presented at the GR900 Connector reference plane of the Type 900-WN4 Precision Short-Circuit Termination.

Since the TYPE 900-WN4 comprises a single section of uniform transmission line with no disturbances or dielectric supports between the short circuit and the connector reference plane, it is a calculable inductance standard of high accuracy. This is particularly true at frequencies above about 50 MHz, where the current flows primarily in the silver overlays on the conductive surfaces. Figure 9 is a plot of the calculated inductance at the connector reference plane for frequencies above 50 MHz based on a conductor resistivity of 1.7 microhm-cm, which is typical for the conductors of the TYPE 900-WN4.

SPECIFICATIONS

TYPE 900-WN4 PRECISION SHORT-CIRCUIT TERMINATION

Frequency Range: DC to 8.5 GHz.
Reflection Coefficient: Greater than 0.996 at the GR900 Connector reference plane.
Location of Short Circuit: 4.00 ± 0.01 cm beyond the GR900 Connector reference plane.

Characteristic Impedance of Internal Coaxial Line: 50.0 Ω ± 0.065% at frequencies where skin effect is negligible.

Leakage: Better than 130 dB below signal.

Dimensions: Length, 2 in (51 mm); maximum diameter, 1 1/16 in (27 mm).

Net Weight: 4 oz (120 g).

Catalog Number	*Description*	*Price in USA*
0900-9975	**Type 900-WN4 Precision Short-Circuit Termination**	**$40.00**

TYPE 900-WO4 PRECISION OPEN-CIRCUIT TERMINATION

The TYPE 900-WO4 presents an open circuit 4.0 cm beyond the GR900 Connector reference plane over the full dc-to-8.5-GHz frequency range, as illustrated in Figure 10. Compensation for the frequency-dependent fringing capacitance of the open-ended inner conductor is accomplished by means of a small disk on the inner conductor tip.

APPLICATIONS

As a capacitance standard, the TYPE 900-WO4 presents a capacitance at its connector reference plane that is given approximately by

$$C = C_o \left[1 + \left(\frac{\tan \frac{4\pi f}{15}}{\frac{4\pi f}{15}} - 1 \right) \right]$$

$$= C_o \, (1 + \mathrm{K})$$

where the capacitance C_o is a result of the 4-cm length of line between the effective open-circuit reference plane and the connector reference plane and f is the frequency in GHz. The capacitance C_o has a nominal value of 2.673 picofarads. The correction term

$$\mathrm{K} = \frac{\tan \frac{4\pi f}{15}}{\frac{4\pi f}{15}} - 1$$

is a result of the distributed nature of the capacitance, which has an appreciable effect at frequencies above 70 MHz. The correction term K is plotted in Figure 11.

As an open-circuit termination for the TYPE 900-LZ Reference Air Lines, the TYPE 900-WO4 provides support for the inner conductors of the air lines. Since the effective reference plane of the TYPE 900-WO4 Open Circuit is coincident with that of the TYPE 900-WN4 Short Circuit, these two units, alone or in conjunction with the TYPE 900-LZ Reference Air Lines, form a series of accurate conjugate-reactance standards, which can be used in the calibration of impedance-measuring devices. Further, the reference plane of the TYPE 900-WO4 is coincident with those of the TYPES 900-W100 and -W200 Terminations within 0.06 cm to 2 GHz and within 0.20 cm to 8.5 GHz.

Combinations of the TYPE 900-WO4 Open-Circuit and the TYPE 900-LZ Air Lines also make an accurate series of incremental capacitance standards for use at audio and at the lower radio frequencies. Fringing capacitance at

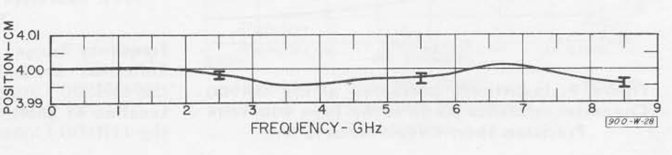
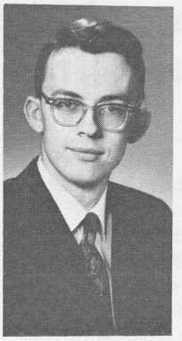


Figure 10. Average position behind GR900 Connector reference plane at which open circuit is applied. Data are based on 25 units. Spreads are shown at 2.5, 5.5, and 8.5 GHz.

the measuring-instrument terminals is eliminated when the TYPE 900-WO4 is used to establish the initial conditions. The agreement between calculated capacitance and measured capacitance at 1 kHz for a 10-picofarad (15 cm) TYPE 900-LZ15 Reference Air Line is better than 0.05%. The GR900 Connector repeatability at 1 kHz is better than 0.001 picofarad.

— T. E. MACKENZIE



Thomas E. MacKenzie received his BSEE and MS in Physics in 1958 and 1963, respectively, from Northeastern University. From 1954 to 1962, he was an engineer at Alford Manufacturing Company. As a development engineer in the Microwave Group at General Radio since 1962, his work has been in the development of microwave instruments, components, and standards.

SPECIFICATIONS

TYPE 900-WO4 PRECISION OPEN-CIRCUIT TERMINATION

Frequency Range: DC to 8.5 GHz.

Reflection Coefficient: Greater than 0.996 at the GR900 Connector Reference Plane.

Location of Open Circuit: 4.00 ± 0.01 cm beyond the GR900 Connector reference plane.

Capacitance at GR900 Connector Reference Plane: 2.673 pF ± 0.3%, dc to 70 MHz.

Characteristic Impedance of Internal Coaxial Line: 50.0 Ω ± 0.1% at frequencies where skin effect is negligible.

Leakage: Better than 130 dB below signal.

Dimensions: Length, 2 5/16 in (59 mm); maximum diameter, 1 1/16 in (27 mm).

Net Weight: 4 oz (120 g).

Catalog Number	*Description*	*Price in USA*
0900-9985	**Type 900-WO4 Precision Open-Circuit Termination**	**$40.00**

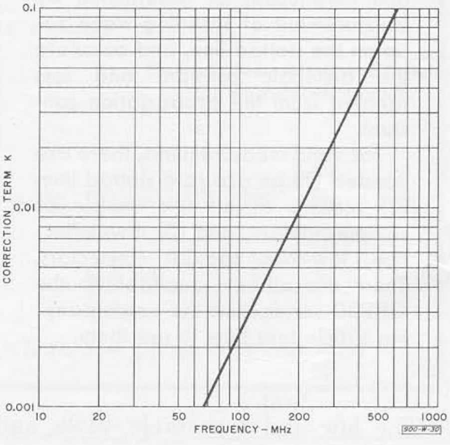


Figure 11. Correction term, K, for Type 900-WO4 Open-Circuit Termination.

Publications

PULSES — Instrument Note IN-108, "Generation and Detection of Modulated Pulses," by Dr. Gordon R. Partridge, of GR's Development Engineering Department, discusses the characteristics of pulse-position, pulse-amplitude, pulse-duration, and pulse-code modulation and describes equipment and methods for their generation and detection.

Reprints are also available of recent technical articles by GR engineers:

Number	*Author*	*Title*	*Published in*
A123	W. G. Howard	"Simple Methods of Voltage-to-Frequency Conversion"	*Frequency*, September/October, 1964
A124	H. T. McAleer	"Unique Frequency Multiplier"	*Frequency*, May/June, 1964
A126	A. E. Sanderson	"How to Measure Source Impedance with a Slotted Line"	*EEE*, November, 1965

Measurements of Dielectric Materials with the Precision Slotted Line

The slotted line has long been recognized as a fundamental tool for measuring the dielectric properties of materials at high frequencies. In principle, the measuring technique is simple: fill a section of coaxial line with dielectric material, determine the propagation constant of the filled section of line from the phase and magnitude of the reflection introduced, as determined by measurement of standing-wave ratio on the slotted line, and calculate the dielectric constant and loss tangent from the propagation constant.

For valid measurements, there are needed (1) an accurate slotted line, (2) sections of air line usable as sample holders, and (3) low-reflection, low-loss, coaxial connectors. These are all now available in the GR900 series, and the accompanying article tells how to use them.

The low and repeatable VSWR and the low loss of the GR900 Precision Coaxial Connector make possible the use of GR900 equipment for the accurate determination of dielectric constant and loss tangent. No specialized dielectric measuring apparatus is necessary.

The measuring device is the TYPE 900-LB Precision Slotted Line.[1] The combination of a TYPE 900-LZ Reference Air Line[2] and a TYPE 900-WNC Short Circuit makes a convenient sample holder for solid dielectrics. The error introduced by the inclusion of the GR900 Connector between the sample and the point of measurement is negligible for most purposes.

The dimensions for a cylindrical sample of solid dielectric are shown in Figure 1. The total length of the sample may be made up of a number of pieces and may be equal to or less than the length of the sample holder. There should be no gaps between the individual pieces. The accuracy of the measurements will depend upon the precision with which the diameters are machined. A light press fit of the sample against the inner and outer conductors is desirable, but too tight a fit may damage the TYPE 900-LZ Reference Air Line. For accurate loss-tangent measurements of a very low-loss material, the length of the sample should be selected by the procedure described below under *Effect of Contact Resistance.*

Standard lengths of TYPE 900-LZ Reference Air Lines (5 cm, 6 cm, 7.5 cm, 10 cm, 15 cm, 30 cm) will meet most needs. If other lengths are needed, they can be constructed from TYPE 0900-9508 rod, TYPE 0900-9509 tube, and TYPE 900-AP Connector Kits.

THEORY

The measurements that will be considered here are those of nonmagnetic materials in a short-circuited sample holder. Other types of measurements are described in various references.[3]

If a coaxial line containing a dielectric sample is short-circuited at its far

[1] J. Zorzy, "Precision Coaxial Equipment — The 900 Series," *General Radio Experimenter*, November 1963.
[2] A. E. Sanderson, "Reference Air Lines for the GR900 Series," *General Radio Experimenter*, January 1965.
[3] For example, A. von Hippel, *Dielectric Materials and Applications*, Technology Press of MIT, 1954.

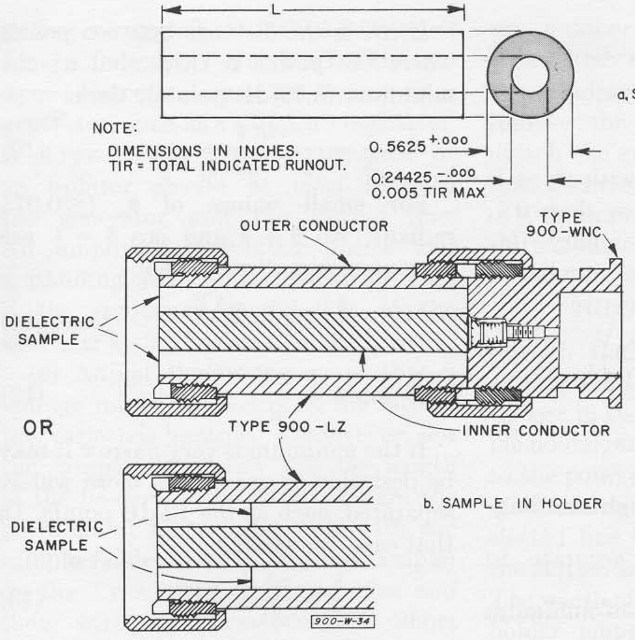


Figure 1. Dimensions and installation of dielectric sample.

end, the relationship between the propagation constant, γ, of the dielectric-filled line and the standing-wave ratio, S, and wavelength, λ_o, in an attached air-filled section of the line is:

$$\frac{\tanh \gamma d}{\gamma d} = \frac{\frac{1}{S} - j \tan \frac{2\pi X_o}{\lambda_o}}{1 - j \frac{1}{S} \tan \frac{2\pi X_o}{\lambda_o}} \frac{(-j)\lambda_o}{2\pi d} \qquad (1)^4$$

where X_o = the distance from the face of the dielectric sample to the first voltage minimum in the air-filled line,

d = the length of the sample,

λ_o = the wavelength in the air-filled line,

S = the standing-wave ratio in the air-filled line.

This equation can be separated into its real and imaginary parts and, if tan δ (the loss tangent) is less than 0.1, simplified with results accurate within $\pm 1\%$. The simplified equations are:

$$\frac{\tan \beta d}{\beta d} = \frac{-\lambda_o \tan \frac{2\pi X_o}{\lambda_o}}{2\pi d} \qquad (2)^4$$

where β = phase constant, and

$$\alpha d = \frac{\beta^2 d^2 \lambda_o}{2\pi d} \frac{1}{S} \frac{1 + \tan^2 \frac{2\pi X_o}{\lambda_o}}{\beta d (1 + \tan^2 \beta d) - \tan \beta d} \qquad (3)^4$$

where α = attenuation constant.

If the frequency and sample length are chosen so that $X_o = 0$, then tan $\beta d = 0$ and $\beta d = N_s \pi$, where N_s is the number of half wavelengths in the sample. The equation for αd then becomes:

$$\alpha d = \frac{N_s \lambda_o}{2d} \frac{1}{S}. \qquad (4)$$

From a knowledge of α and β in the dielectric-filled line, the relative dielectric constant, ϵ_r, and the loss tangent, tan δ, can be calculated. For the TEM mode in a coaxial line:

[4] T. W. Dakin and C. N. Works, "Microwave Dielectric Measurements," *Journal of Applied Physics*, September 1947, p 789.

$$\epsilon_r = \frac{\lambda^2_o}{4\pi^2}\,(\beta^2 - \alpha^2), \quad (5)$$

and $$\tan\delta = \frac{2\alpha\beta}{\beta^2 - \alpha^2}. \quad (6)$$

If α^2 is small compared with β^2, as is the case when $\tan\delta$ is less than 0.1, equations (5) and (6) simplify, for samples that are an integral number of half-wavelengths long, to

$$\epsilon_r = \frac{\lambda^2_o\,\beta^2}{4\pi^2} = \left(\frac{N_s\,\lambda_o}{2d}\right)^2, \quad (7)$$

and $$\tan\delta = \frac{2\alpha}{\beta} = \frac{\lambda_o}{\pi d}\,\frac{1}{S}. \quad (8)$$

For small values of $\frac{1}{S}$ (high standing-wave ratio) it is more accurate to determine $\frac{1}{S}$ by a width-of-minimum method rather than by direct measurement. $\frac{1}{S}$ is related to the voltage at point X, a distance $\frac{\Delta X}{2}$ from the minimum by:

$$\frac{1}{S} = \frac{\sin\theta}{\left[\left(\frac{E_x}{E_{min}}\right)^2 - \cos^2\theta\right]^{1/2}} \quad (9)$$

where $\theta = \frac{\pi\Delta X}{\lambda_o}$.

If ΔX is the distance between points where the power is twice that at the minimum (3.01-dB points), then:

$$\frac{1}{S} = \frac{\sin\theta}{(2 - \cos^2\theta)^{1/2}} \quad (10)$$

For small values of θ, (<0.075 radian), $\sin\theta = \theta$ and $\cos\theta = 1$ are close approximations. Then:

$$\frac{1}{S} = \theta = \frac{\pi\Delta X}{\lambda_o}, \quad (11)$$

and $$\tan\delta = \frac{\Delta X}{d}. \quad (12)$$

If the minimum is very narrow it may be desirable to use points more widely separated, such as the 10-dB points. In that case, for small θ

$$\frac{1}{S} = \frac{\pi\Delta X}{3\lambda_o}, \quad (13)$$

and $$\tan\delta = \frac{\Delta X}{3d}. \quad (14)$$

MEASURING PROCEDURE

(1) Insert the sample into the reference air line flush with one end and with no spaces between the pieces in the sample.

(2) Attach the Type 900-WNC Short Circuit so that it is in contact with the sample, as shown in Figure 1b.

(3) Connect the sample holder to the measuring setup (Figure 2), con-

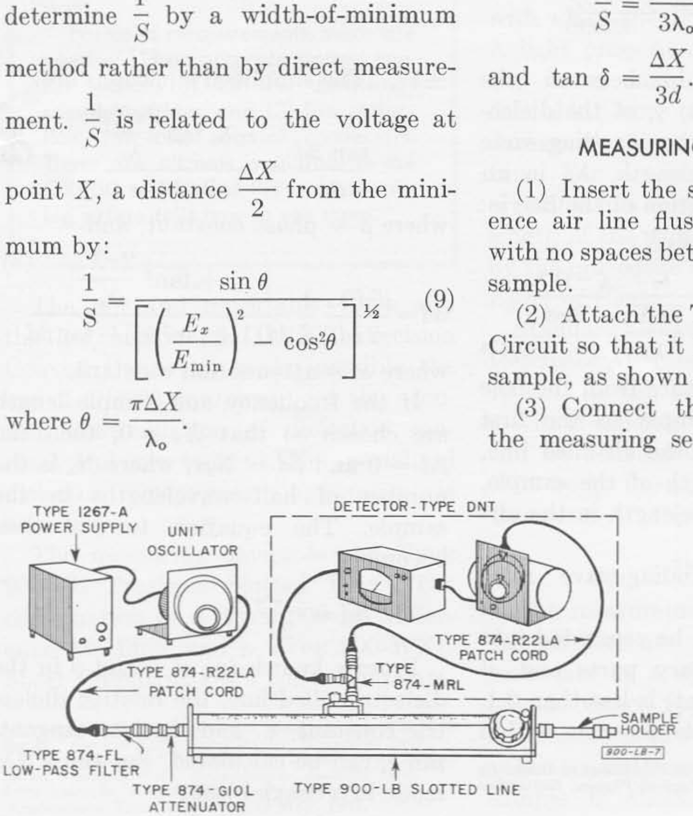


Figure 2. Setup for dielectric measurements.

sisting of the Type 900-LB Slotted Line, a Type DNT-3 or DNT-4 Heterodyne Detector, and an appropriate generator, such as a GR Unit oscillator. A Type 874-G Fixed Attenuator or an isolator should be used between the generator and the slotted line. An amplitude-modulated source and a standing-wave indicator can be used if the frequency modulation of the source is kept very small.

(4) Adjust the frequency so that a voltage minimum occurs at the face of the dielectric sample, whether or not the sample completely fills the length of the holder. To do this, compare the positions of the minima first with the sample holder (containing the sample) on the Type 900-LB Slotted Line and then with a Type 900-WN Short Circuit on the slotted line. Then adjust the frequency until the proper relation exists between the minima. For example, if the sample completely fills the Type 900-LZ Reference Air Line, the minimum position with a Type 900-WN or -WNC Short Circuit connected to the slotted line should be the same as when the sample is connected to the slotted line. If measurements must be made at a certain frequency, then it is necessary either to adjust the length of the sample or to use equations (2) and (3) with X_o not equal to zero.[5]

(5) Once the frequency is properly adjusted, proceed as follows: Record the position and width of the minimum at two places along the slotted line (one of them near the load end), preferably separated by 20 centimeters or more. Measure the width of minimum with the micrometer carriage drive. Count the number of half wavelengths between the two minima (distance between adjacent minima is $\lambda_o/2$). Then remove the sample from the holder, attach the empty sample holder to the line, and record the position and width of a minimum near the load end of the slotted line.

[5] A. von Hippel gives charts of $\frac{\tanh X}{X}$ and tables of $\frac{\tan X}{X}$ and suggests further references.

INTERPRETATION OF DATA

With the sample in place, the resulting width of minimum is determined by loss in the dielectric, loss in the sample holder, and loss in the slotted line up to the point of measurement. The width of minimum at a second point along the slotted line is increased by the loss in the slotted line between the two points. The width of minimum, with the sample holder empty, is determined by the losses in the sample holder and in the slotted line to the point of measurement. In order to determine the loss tangent of the dielectric, it is necessary to separate the dielectric loss from the other losses. Call ΔX_{1s} the width of minimum at position l_{1s} with the sample in place, ΔX_{2s} the width of minimum at position l_{2s}, and ΔX_{1e} the width of minimum at position l_{1e}, with the sample holder empty. Then the width of minimum due to loss in the dielectric is given by:

$$\Delta X_d = \Delta X_{1s} - \Delta X_{1e} - \frac{l_{1s} - l_{1e}}{l_{2s} - l_{1s}} \times (\Delta X_{2s} - \Delta X_{1s}). \qquad (15)$$

This width of minimum can be used in equation (12) or (14) to determine the loss tangent. The dielectric constant can be found from equation (7). If the approximate dielectric constant is unknown, then measurement at two frequencies will be necessary since N_s will not be known.

Example: A Teflon* sample 15.00 centimeters long is measured. It is found that a voltage minimum occurs at the sample face when the frequency is adjusted so that $\lambda_o = 21.34$. Then from equation (7) $\epsilon_r = \left(\frac{N_s \lambda_o}{2d}\right)^2 = \left(\frac{N_s\, 21.34}{2(15.00)}\right)^2$. The dielectric constant is known to be approximately 2. Therefore, $N_s = 2$ and $\epsilon_r = \left(\frac{2(21.34)}{2(15.00)}\right)^2 = 2.024$. Since the minimum is very narrow, the 10-dB width-of-minimum points are used. The width of minimum at $l_{1s} = 21.34$ is 0.1004 cm. The width of minimum at $l_{2s} = 42.68$ is 0.1518 cm. With the sample holder empty, the width of minimum at $l_{1e} = 17.01$ is 0.0788 cm. Then the width due to losses in the sample is found from equation (15) as

$$\Delta X_d = 0.1004 - 0.0788 - \frac{(21.34-17.01)}{(42.68-21.34)} \times (0.1518 - 0.1004) = 0.0110.$$

From equation (14), $\tan \delta = \frac{0.0110}{3(15.00)} = 0.00024.$

Note that if a lossy material is measured ($\tan \delta > 0.1$), equations (2) and (3) are no longer valid and equation (1) must be solved.[4]

FREQUENCY RANGE OF MEASUREMENT

The lowest frequency at which measurements can be made is determined by the dielectric constant of the material being measured and by the necessity that at least one minimum occur along the slotted line so that its position and width can be measured. TYPE 900-L10, -L15, and -L30 Precision Air Lines can be used between the sample holder and the slotted line to position a minimum on the slotted line at low frequencies. If these additional air lines are used they should be externally supported. Sample holders up to 66 centimeters long can be constructed for low-frequency use. The sample can be made shorter than a half-wavelength and equations (2), (3), (5), and (6) used to determine the dielectric constant and loss tangent. With these methods, measurements can be made down to 50 MHz or even lower.

The upper frequency limit for the TYPE 900-LB Slotted Line is 8.5 GHz, but special precautions should be taken at frequencies higher than $\frac{9.5 \text{ GHz}}{\sqrt{\epsilon_r}}$ as noted in the paragraph *Existence of Higher-Order Modes.*

ERRORS

Sample Fit

One of the most common sources of error in dielectric measurements by the coaxial method is the presence of air gaps between the sample and the inner and outer conductors. Correction formulas based upon a uniform distribution of the air gap can be used, but, since the actual air gap will usually not be uniformly distributed, the gaps should be avoided for maximum accuracy. The corrections for uniform air gaps for $\tan \delta < 0.1$ are

$$\epsilon_{r\,(correct)} = \epsilon_{r\,(measured)} \frac{L_2}{L_3 - \epsilon_{r\,(measured)} L_1} \quad (16)^6$$

$$\tan \delta_{(correct)} = \tan \delta_{(measured)} \left(1 + \epsilon_{r\,(correct)} \frac{L_1}{L_2}\right) \quad (17)^6$$

* DuPont trademark.
[4] *Ibid.*
[6] W. B. Westphal, "Techniques at Measuring the Permittivity and Permeability of Liquids and Solids in the Frequency Range 3 c/s to 50 kMc/s," *Technical Report No. 36*, Laboratory for Insulation Research, M.I.T., July 1950. (Out of print)

where $L_1 = \mathrm{Log}\,\frac{D_2}{D_1} + \mathrm{Log}\,\frac{D_4}{D_3}$,

$L_2 = \mathrm{Log}\,\frac{D_3}{D_2}$,

$L_3 = \mathrm{Log}\,\frac{D_4}{D_1}$,

$D_1 = 0.24425$,

D_2 = inside diameter of sample,

D_3 = outside diameter of sample, and

$D_4 = 0.5625$.

Meter Errors

If a 3-dB width of minimum is used, meter indications on the GR Type 1216-A Unit I-F Amplifier will, in general, cause negligible error when the upper part of the scale is used and when care is taken to tune the local-oscillator frequency exactly for maximum output. A 10-dB width-of-minimum measurement may require that the i-f amplifier calibration be checked with a precision attenuator for greatest accuracy. As an example of the errors in loss-tangent measurements caused by poor i-f amplifier calibration, an error of 0.1 dB in a typical 3-dB width-of-minimum measurement will cause an error of 1.9% in tan δ. An error of 0.3 dB in a typical 10-dB width-of-minimum measurement will result in a 3.9% error in tan δ.

Effect of Contact Resistance

Although the connector contact resistance is typically less than half a milliohm, a small part of the measured loss is due to this resistance. The magnitude of the error caused by this loss depends upon the relative current through the contact for each measurement and is significant only when very low-loss dielectrics are measured. If the current is the same when the sample is measured as when the empty sample holder is measured, the contact loss will have no effect on the accuracy of the tan δ measurement. If the currents differ, there may be an error in tan δ as large as 0.0001. The amount of loss due to the finite contact resistance in a given measurement is

$$\text{Loss} \approx \frac{\cos 2\phi + 1}{2} \times \text{maximum loss}, \qquad (18)$$

where $\phi = \frac{l}{\lambda} 360°$ and l is the distance from a voltage minimum to the contact. Maximum loss occurs when a voltage minimum occurs at the contact. It is difficult to evaluate the maximum loss exactly because of its small value. The condition that the current be the same for both measurements (with and without sample) may be met by appropriate choice of length and frequency for a sample with a given dielectric constant. If the dielectric constant is unknown, it may be necessary first to measure dielectric constant and then to trim the sample to the proper length for accurate determination of loss. This is necessary only for very accurate measurements of the loss tangent of low-loss dielectrics. For low-loss materials, the current through the contacts will be of approximately the same magnitude with and without the sample in the holder when the frequency and length are so chosen that sample

$$\text{length } d = \frac{2\,N_s b}{(N+1)\sqrt{\epsilon} + N_s}, \qquad (19)$$

$$\text{and } \lambda_o = \frac{2\sqrt{\epsilon}}{N_s}\,d, \qquad (20)$$

where N_s and N are integers and b is the length of the sample holder. Lengths

John F. Gilmore received his BSEE degree in 1961 and his MSEE in 1963, from Northeastern University. He was employed as a cooperative student at Alford Manufacturing Company and as an engineer with that company from 1961 to 1963. He joined the Microwave Group at General Radio as a development engineer in 1963 and is currently engaged in microwave circuit design.

that satisfy the relationship and the corresponding values of N_s can be determined from Figure 3 for a 15-cm sample holder and from Figure 4 for a 30-cm sample holder. Figure 4 shows only the most useful curve of a very large family of curves. As an example of the use of these curves, suppose that the loss tangent of a low-loss material with a dielectric constant of 2 is to be measured. If a sample 12.43 cm long is used in a 15-cm sample holder, tan δ can be measured with maximum accuracy at λ_o = 17.60 cm, 11.73 cm, 8.80 cm, 7.05 cm, 5.87 cm, and 5.03 cm, corresponding to N_s = 2, 3, 4, 5, 6, 7. If, instead, a sample 13.54 cm long were chosen, tan δ could be measured with maximum accuracy only at λ_o = 5.48 cm, N_s = 7.

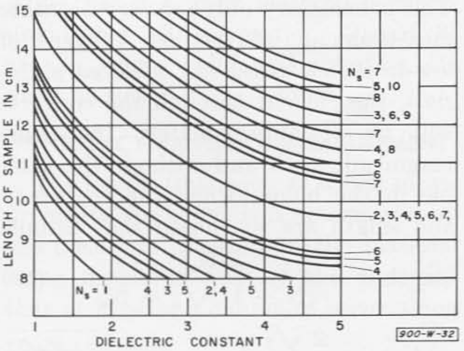


Figure 3. Lengths of samples for a Type 900-LZ15 Reference Air Line for minimum tan δ error in low-loss dielectric measurements. (See equations 18 and 19.)

Existence of Higher-Order Modes

At frequencies higher than $\frac{9.5}{\sqrt{\epsilon}}$ GHz, higher-order modes, particularly the TE_{11} mode, can be excited by axial dissymmetries in the dielectric material. While the air-filled section of line between the sample and the point of measurement acts as a filter for these higher-order modes, in some instances coupling between the TEM and TE modes may be great enough to produce an error in measurement. Measurements above this frequency, therefore, should be made at small (say 10%) frequency increments and compared with measurements below $\frac{9.5}{\sqrt{\epsilon}}$ GHz, in order that anomalous results can be detected.

— J. F. Gilmore

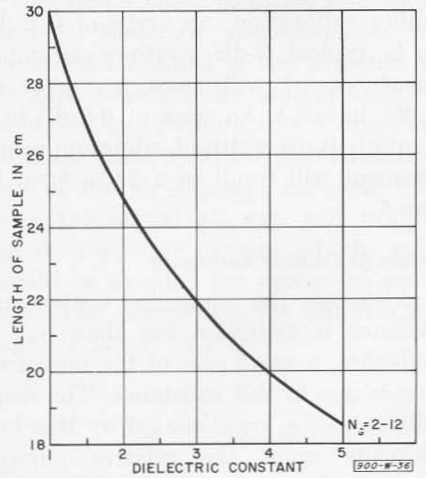


Figure 4. Lengths of samples for a Type 900-LZ30 Reference Air Line for minimum tan δ error in low-loss dielectric measurements.

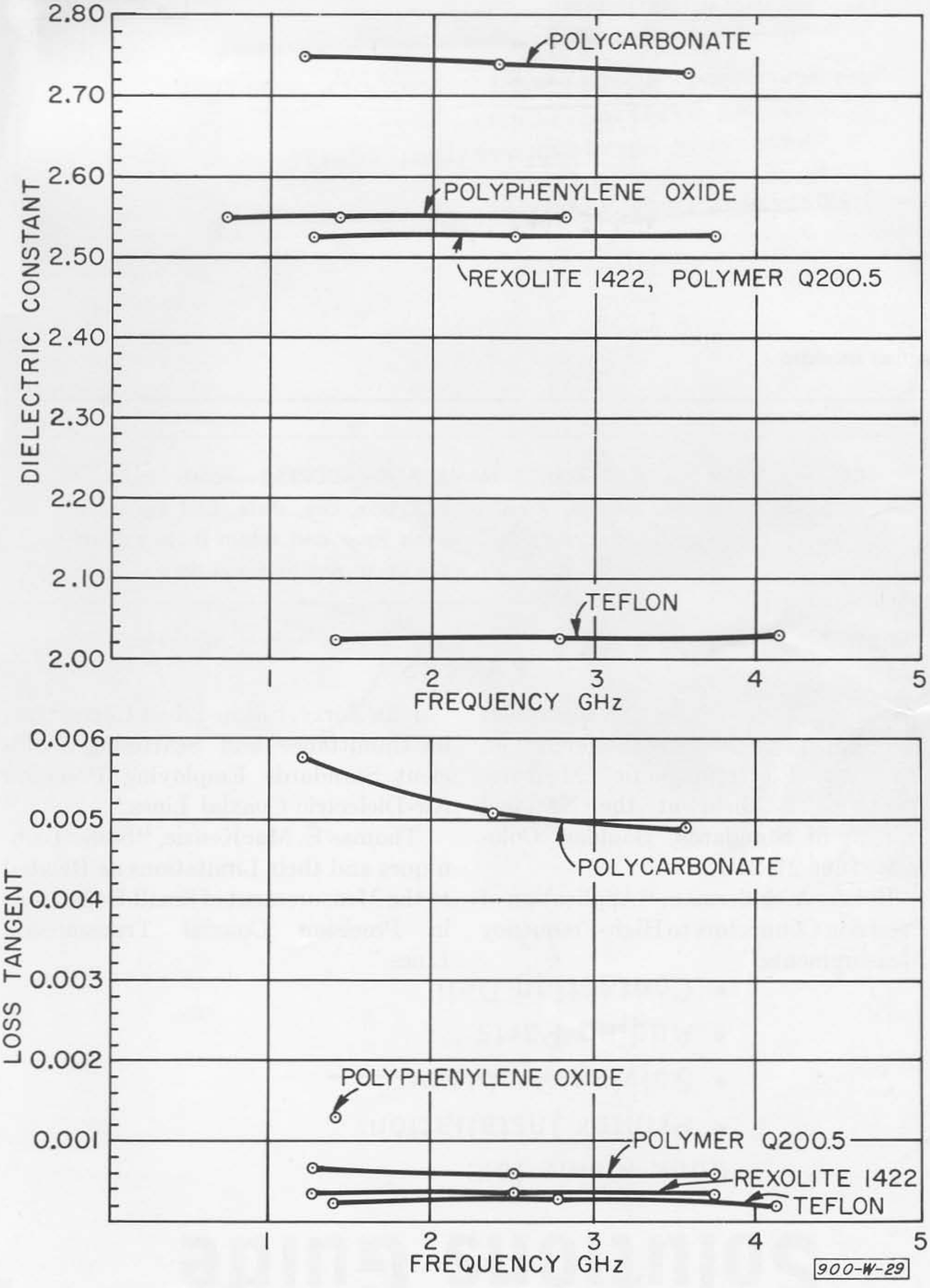


Figure 5. Dielectric constant and loss tangent of typical materials as measured on the Type 900-LB Precision Slotted Line.

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PAPERS

General Radio engineers will present three papers at the 1966 Conference on Precision Electromagnetic Measurements to be held at the National Bureau of Standards, Boulder, Colorado, June 21–24:

Robert A. Soderman, "Application of Precision Connectors to High-Frequency Measurements."

John Zorzy, "Skin-Effect Corrections in Immittance and Scattering-Coefficient Standards Employing Precision Air-Dielectric Coaxial Lines."

Thomas E. MacKenzie, "Some Techniques and their Limitations as Related to the Measurement of Small Reflections in Precision Coaxial Transmission Lines."

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This Issue:

- The Measurement of Electrolytic Capacitors

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REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

Figure 1.

THE MEASUREMENT OF ELECTROLYTIC CAPACITORS

The measurement of electrolytic capacitors requires specialized measuring instruments. High capacitance values, high dissipation factors, limited voltage ratings, and the need for dc bias are all factors, that influence the design of the capacitance bridge. Moreover, the residual impedances of the leads can be comparable with the impedances to be measured, and their effects must be eliminated if accurate measurements are to be made.

This article discusses the measurement conditions and the design of a new capacitance bridge to satisfy them.

Perhaps the largest capacitances that most of us have had to deal with were those in the problems given in circuit-theory textbooks[1] where all circuit elements were in ohms, henrys, or farads. While actual capacitors in the farad range didn't seem very practical, we realized it was the principle involved that was important and not the magnitudes. However, students these days may accept such grotesque circuit values without question and, in fact, may have seen capacitors of this value advertised or displayed at the recent IEEE Show demonstrating our new Type 1617 Capacitance Bridge (Figure

[1] For instance, E. A. Guillemin, *Introduction to Circuit Theory*, John Wiley and Sons, New York, 1953.

1). One-farad capacitors may not become very common for some time yet, but capacitors of 0.1 farad and up are being made by several manufacturers, and capacitors of over 0.01F (10,000 μF) are not uncommon in low-voltage power supplies.

To avoid large errors from the impedances of the connecting leads, one should measure these large capacitors on a four-terminal bridge. One farad has a reactance of only 1.3 milliohms at 120 Hz, the standard test frequency, and the resistance of test leads may be many times larger. Even for capacitance values well below 1000 μF, two-terminal measurements of *D* may be subject to errors. For larger values, capacitance-measurement errors may be caused by lead inductances, even at frequencies as low as 120 Hz.

The new TYPE 1617 Capacitance Bridge is designed especially for measuring these large-valued capacitors, as well as other electrolytic types, most of which require the special measurement conditions prescribed by MIL or EIA specifications (see Table 1). The internal test frequency of 120 Hz is the fundamental of the ripple signal that would be applied to filter capacitors used in full-wave power supplies. While this is far from the only application for electrolytic capacitors, 120 Hz has been the generally accepted measurement frequency for all except the nonpolarized types used for motor starting and other special 60-Hz applications. The TYPE 1617 can be used with a suitable external generator* down to 40 Hz and up to 1 kHz over most of its

* GR TYPE 1311-A or TYPE 1310-A.

TABLE I

Summary of EIA and MIL Specifications on Testing Electrolytic Capacitors

Specification and Capacitor Type	*Frequency*	*AC Level*	*Accuracy* C	*Accuracy* *Loss*	*DC Polarizing Voltage*
MIL C—3965 C Tantalum Foil and Sintered Slug Capacitors	120±5 Hz	Less than 30% of DCWV or 1 V, whichever is smaller	2%	*R* or P.F. 2%	C—Sufficient for no reversal of polarity. *D*—"Polarized Capacitance Bridge" Sum of ac and dc shall not exceed DCWV.
MIL C—26655-B Solid Tantalum Capacitors	120±5 Hz	Limited to 1V, rms	2%	*D*, 10%	C—Max bias 2.2 V. *D*—"Polarized Bridge", 2.2-V dc max.
RS 228 Tantalum Electrolytic Capacitors	120 Hz	Small enough not to change value	±2½%	*D*, 5%	Optional
MIL C-62 B Polarized Aluminum Capacitors	120±5 Hz	Limited to 30% of DCWV or 4 V, whichever is smaller	2%	*D*, 2%	No bias required if ac voltage less than 1 V. However, if bias causes differences, measurements with bias shall govern.
RS 154 B Dry Aluminum Electrolytic Capacitors	120 Hz	Small enough not to change value	±2½%	*R* or RC	Optional, but if substantial difference occurs, rated dc should be used.
RS 205 Electrolytic Capacitors for use in Electronic Instruments	120 Hz	Small enough not to change value	±2½%	*D*	Optional

range with no degradation in capacitance-measurement accuracy.

Test-Signal Level

The ac signal level is particularly important for tantalum types, which tolerate only a very low reverse voltage. The generator level switch has three positions: 2 V, 0.5 V, and 0.2 V, rms, maximum. This is the applied bridge voltage; the actual voltage applied to the unknown capacitor is somewhat less. The 0.5-volt maximum signal level has become an informal industry standard. The 2-volt level was included for increased sensitivity on the lowest capacitance range (0 pF – 1000 pF), and the 0.2-V position was added to allow a check to ensure that the level was "small enough not to change value," in the words of some specifications.

DC Bias

While dc bias is not specifically required by the MIL and EIA standards for capacitance measurements (see Table 1), it is specified for MIL dissipation-factor measurements of tantalum units. Also, the test conditions used by capacitor manufacturers, which vary considerably, usually incorporate dc bias. For tantalum capacitors, the bias specified is such that, when the ac signal is added, the sum will never go negative and will never exceed the rated voltage. A common practice is to use 1.5-volt dc bias with less than 0.5-volt ac signal. However, several manufacturers require that almost the full dc rated voltage be applied. To cope with such a large variation in the dc-bias requirements, the bridge includes an adjustable internal supply, with six ranges up to 600 volts. This covers the ratings of almost all electrolytic capacitors as well as those of most other types. External bias up to 800 volts may be applied.

The applied bias voltage is indicated on the panel meter, which also serves as the null indicator and as a leakage-current ammeter with full-scale ranges from 60 μA to 20 mA. The ultimate resolution of about 0.5 μA is adequate for most aluminum capacitors and some tantalum types. For others an external microammeter can be used.

Safety Features

Measurements on biased capacitors may be inherently dangerous, because lethal energy can frequently be stored in the capacitor being measured. Moreover, in order to get good ac sensitivity, a large bypass capacitor is required in the internal dc supply, which may also store dangerous energy in the instrument itself. The user and the instrument are protected by several features, including two warning lights and discharge circuitry. If bias is not required, three switches must all be improperly set in order to get a dangerous voltage.

D Range

Not only the wide capacitance range (1 pF to 1.1 F) but also a wide dissipation-factor range is needed to measure large electrolytic capacitors, whose *D* value is often well over 1 (100%) at 120 Hz. The *D* range of the TYPE 1617, which extends up to 10, would be awkward to use above unity were it not for the inclusion of the ORTHONULL® balancing mechanism, which is also a feature of our popular TYPE 1650 Impedance Bridge.[2] This

[2] H. P. Hall, "Orthonull — A Mechanical Device to Improve Bridge-Balance Convergency," *General Radio Experimenter*, April, 1959.

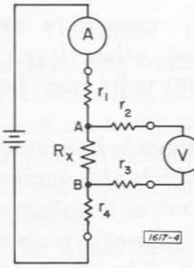


Figure 2. Voltmeter-ammeter measurement of a four-terminal resistor, R_x. The terminal resistances r_1 and r_4 have no effect, nor do r_2 and r_3 if the voltmeter impedance is infinite.

nonreciprocal ganging mechanism converts interacting bridge adjustments into independent ones so that the real and imaginary parts of the impedance can be balanced without a time-consuming sequence of adjustments and readjustments.

Brief Theory of Four-Terminal Measurements

Very low impedances must be measured by a four-terminal method when a high degree of accuracy is to be realized. This is the only way in which errors caused by the impedance of the connecting leads, terminals, and contacts (which may be much larger than the unknown impedance being measured) can be avoided. The simplest four-terminal measurement circuit is the voltmeter-ammeter method shown in Figure 2. This is a *transfer*-impedance measurement, because the impedance, Z_X, is the ratio of output voltage to input current, Z_{21}. From an inspection of this circuit, it is obvious why two terminals of a four-terminal standard are called the "current" terminals and two called the "potential" terminals. Although these terminal pairs are theoretically interchangeable because $Z_{21} = Z_{12}$ for a linear, passive, bilateral network, the current ratings of the two pairs of terminals may be quite different, as may other practical aspects of the design. Note that the unknown, R_X, is defined as the resistance between the junction of the upper two leads (point A) and the junction of the lower two leads (point B). The location of these junctions should be permanent in a low-impedance standard, independent of the position of connecting leads, and, therefore, such a standard must have four separate terminals.

More accurate impedance measurements can be made with a bridge, Figure 3, to avoid meter calibration errors. However, here the residual resistances, r_2 and r_4, appear in other bridge arms, where they can cause errors unless the resistances of these arms are high enough to make the residuals negligible by comparison. Note that r_1 and r_3 are in series with the source and detector and thus cause no error.

For precision measurements on low-valued resistors for example, the Kelvin double bridge is used (Figure 4). As shown, the unknown is compared with another four-terminal resistor, R_S. If R_A and R_B are sufficiently large, the main source of error is the branch containing r_4 and r_5 and results from the voltage drop, E_y, which may be appreciable because r_4 and r_5 are in series with low-impedance R_X and R_S.

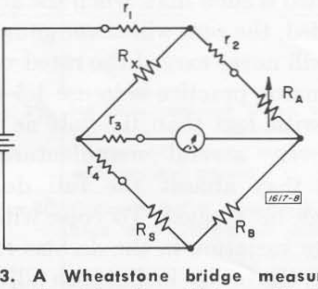


Figure 3. A Wheatstone bridge measuring a four-terminal resistor, R_x, with its residual terminal impedances shown in adjacent circuit branches.

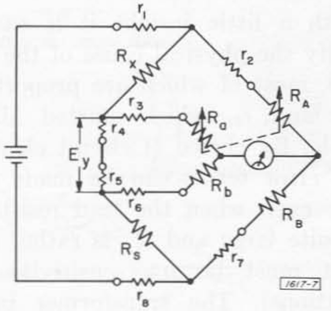


Figure 4. A Kelvin double bridge comparing two four-terminal resistors. The second set of ratio arms, R_a and R_b, divide the error voltage, E_y, proportionately.

The Kelvin bridge uses auxiliary arms, R_a and R_b, to divide this error voltage proportionately between the two halves of the bridge so that its effect is balanced and cancelled.

The excellence with which this second set of ratio arms (R_a and R_b) balances out the error voltage depends on the difference between their ratio, $\frac{R_a}{R_b}$, and that of the main arms, $\frac{R_A}{R_B}$. The balance expression is:

$$R_X = R_S \frac{R_A}{R_B} + \frac{R_b(r_4 + r_5)}{R_a + R_b + r_4 + r_5}\left(\frac{R_A}{R_B} - \frac{R_a}{R_b}\right)$$

As an example, if the unknown and standard resistances and the stray resistance, r_4 plus r_5, are all equal, and if R_A/R_B and R_a/R_b differ by 1%, the error will be ½%. In precision Kelvin bridges, the adjustable arms, R_A and R_a, track to much closer tolerance than this.

An obvious way to make a capacitance bridge that will measure very high capacitances is to adapt the Kelvin bridge principle to a conventional capacitance bridge, as shown in Figure 5. Any ac bridge for measuring complex impedances must have two adjustments (here C and D); to keep the ratio of the secondary arms equal to that of the main arms, two pairs of tracking adjustments are therefore necessary.

What kind of lead errors would we get with such a bridge? A farad has a reactance of 1.3 milliohms at 120 Hz, and the resistance of the connecting leads could easily be 20 milliohms (two feet of #20 wire). If we assume that R_A = 10 milliohms* and that R_N and R_n can track to 1%, the resulting capacitance error is 2%. The error would be of this magnitude over most of the top range (if $D_X = 0$). This design might be acceptable but is somewhat marginal.

A New Circuit for AC Four-Terminal Measurements

Another way to compensate for the error voltage, E_y in Figure 5, is to introduce, by some means, an equal voltage in the corresponding place in

* This is the value of the ratio arm on the top range of the TYPE 1617. It must be very small to obtain reasonably good sensitivity.

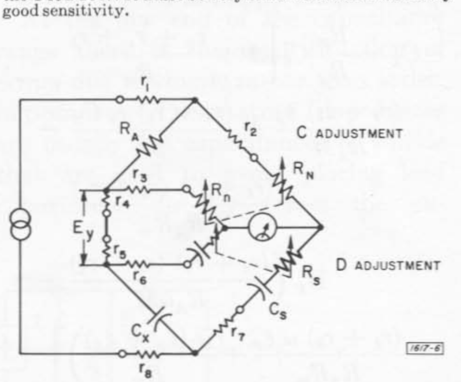


Figure 5. A four-terminal capacitance bridge using the Kelvin double bridge principle. For ac measurements on a complex impedance, two ganged adjustments are necessary.

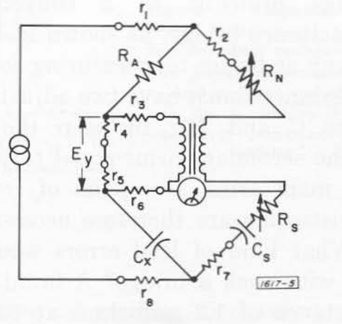


Figure 6. The basic circuit of the 1617, where a voltage equal (almost) to the error voltage, E_y, is placed in the opposite side of the bridge by a tightly coupled transformer.

the opposite side of the bridge. If this is done, no error results, because this extraneous voltage can be considered to be effectively in series with the generator and, thus, will change only the level of the voltage applied to the bridge.

An ideal transformer of 1:1 ratio would do this job perfectly if the R_N and C_S arms were of infinitely high impedance; it is surprising how well a small, practical transformer with practical values for R_N and C_S will actually do the job. For the circuit of Figure 6, the balance equation is:

$$C_X = \frac{R_N}{R_A} C_s \left[1 + \frac{r_p + r_s + r_3}{R_N} - \frac{(r_4 + r_5)\,\ell_p}{R_A M} + \frac{(r_4 + r_5)\,r_6\,C_X}{M} - \frac{(r_4 + r_5)\,(r_3 + r_p)}{R_A R_m} - D_X \left(\frac{(r_4 + r_5)\,(r_3 + r_6)}{R_A \omega M} - \frac{(r_4 + r_5)\,\omega\,\ell_p}{R_A R_m} + \frac{\omega\,(\ell_p + \ell_S)}{R_N} \right) \right]$$

where r_s, r_p, ℓ_s, ℓ_p, R_m and M* are defined by the transformer equivalent circuit of Figure 7.

With a little insight it is easy to identify the physical cause of the error terms, most of which are proportional to r_4 and r_5, which started all the trouble. By choice of circuit elements, these error terms can be made quite small, even when the lead resistances are quite large and R_A is rather small (as it must be for sensitivity considerations). The transformer in the TYPE 1617 uses Mumetal laminations to get a large mutual inductance and a bifilar winding that makes the leakage-inductance terms negligible. (Actually, the measured ratio $\frac{\ell\dagger}{M}$ is $< 3 \times 10^{-6}$, which may seem unbelievable but which is actually not at all difficult to achieve in bifilar transformers.) The winding resistance and the other lead resistance, r_6, appear in the terms that cause the largest errors.

How well does this scheme work? For the circuit values used in the previous example (R_A = 10 mΩ, $r_4 + r_5 = r_6 = 20$ mΩ, C_X = 1 F), the error is approximately 0.1% (if D is small). The specifications allow less than 1% additional error in C or 0.01 in D for resistances of 0.1 ohm in any or all leads on the highest range or 1 ohm in any or all leads on the lower ranges. These values permit the use of connecting leads of considerable length, so that large ca-

* Actually, a capacitor is placed across the transformer to resonate the mutual inductance, thus reducing the error terms containing M.
† $\ell = \ell_s = \ell_p$.

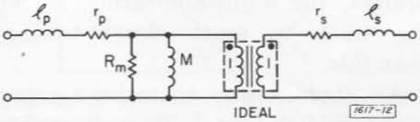


**Figure 7.
Equivalent circuit of the transformer.**

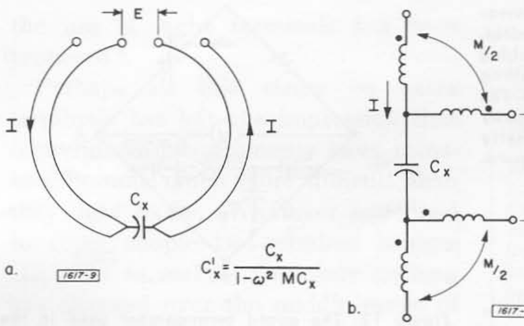


Figure 8. When the "current" and "potential" leads form concentric loops (8a), the resulting mutual inductance (8b) affects the value of the capacitance measured.

pacitors can be measured while remotely located.

A somewhat similar problem with a very similar solution has been reported by Foord, Langlands and Binnie.[3] Therefore, we cannot claim the principle as new, but we can say that we haven't seen a circuit exactly like ours before.

An Interesting Error—And Its Cure

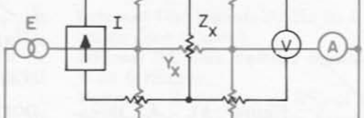
While four-terminal measurements remove the error caused by reasonable amounts of resistance and self-inductance in the connecting leads, mutual inductance between the "current" leads and the "potential" leads can cause appreciable error when the capacitance being measured approaches a farad. If concentric loops are formed by one-foot leads, as shown in Figure 8, the mutual inductance can be as much as 0.3 μH, which causes a 15% error, in a measurement of one farad at 120 Hz. This is a series-resonance effect, and it increases the measured value of capacitance, although the mutual inductance may be negative if one loop is reversed with respect to the other.

The cure is simple; just twist together either the current leads or the potential leads to reduce the mutual inductance. This removes the error almost entirely, leaving only the mutual coupling of the bridge terminals, which is almost completely negligible, even for measurements of one farad.

Three-Terminal Measurements, Too

At the low end of the capacitance range there is the dual situation of errors due to shunt, rather than series, impedances. These stray impedances are usually the capacitances of shields that are used to avoid placing lead capacitance directly across the un-

[3] T. R. Foord, R. C. Langlands and A. J. Binnie, "Transformer-Ratio Bridge Network with Precise Lead Compensation," *Proceedings of the IEE* (Eng), Vol 110, #9, p 86, September 1963.



$$Z_X = \frac{\text{OPEN-CIRCUIT VOLTAGE}}{\text{INPUT CURRENT}} \qquad Y_X = \frac{\text{SHORT-CIRCUIT CURRENT}}{\text{INPUT VOLTAGE}}$$

Figure 9. Superposed diagrams of the measurement of a four-terminal impedance, black, and a three-terminal impedance, red, showing duality. Note that only three circuit branches could define the three-terminal impedance, because two are redundant as shown.

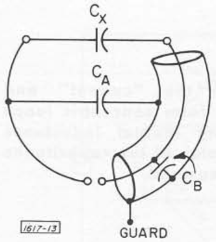


Figure 10. Measurement of a capacitor, C_x, with a shielded lead and the resulting stray capacitances. The shield prevents stray capacitance directly across the unknown.

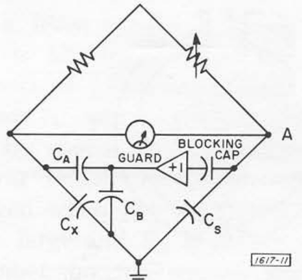


Figure 12. The guard arrangement used in the 1617. No voltage appears across C_A at null (ideally), and C_B just loads the amplifier.

known (Figure 10). With a shield added, there are capacitances from both terminals of the unknown to the shield, forming a three-terminal network (Figure 11). Many standard capacitors of low capacitance value are constructed as three-terminal devices to avoid lead errors[4]. To eliminate the effect of stray capacitances, various types of guard have been used in capacitance bridges.

In the TYPE 1617, the standard capacitor is large enough (0.5 μF) to allow considerable shunt capacitance without appreciable error, so that point A could have been used for a guard. In this case C_A would shunt the detector (no error), and C_B would shunt the standard. However, this point and both the unknown and the standard capacitors get full bias, and we do not encourage our customers to put 600 volts on the shields of their connections. Therefore, an isolating, 3-stage, unity-gain amplifier is used, which blocks the dc and still gives the necessary guard signal to reduce the effect of C_A (by a factor of about 1000) and to give a low output impedance that can tolerate a considerable load (C_B).

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

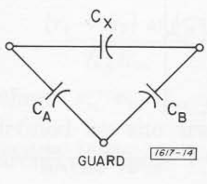


Figure 11. A three-terminal capacitor, C_x, with its "stray" capacitances, C_A and C_B.

Would You Believe 5 Terminals?

The TYPE 1617-A Capacitance Bridge has five terminals and is one of the world's few five-terminal bridges. One cannot help wondering if there is a measurement situation where all five terminals would be useful simultaneously. Suppose, for example, that one wanted to measure a one-microfarad capacitor that was 1000 feet away from the bridge. The leads should be shielded to avoid pickup, and a guard would be necessary to avoid a capacitance error unless a very low-capacitance cable were used. Likewise, if an accurate D value is desired, a four-terminal measurement should be made, because the lead resistance will probably be several ohms. Actually, we haven't tried this, but the TYPE 1617 ought to do it. If anyone tries it, we'd like to hear about it, particularly if it works.

While this example may not be a common problem, there is a very important problem in extremely high precision capacitance measurements where even five-terminals are not enough, and the use of eight terminals has been proposed [5].

Perhaps all this stress on extra terminals has left the impression that capacitance measurements have somehow become much more difficult than they used to be. (Whatever happened to those simple two-terminal bridges GR used to make?) Certainly nothing has changed over the middle range of capacitance values, and only two terminals are required, as always. The TYPE 1617 is just as easy to use for capacitors of these values, particularly if one uses the two-lead, shielded cable harness, which connects the proper terminals together at the bridge. Also supplied is a four-lead harness for the measurement of those larger capacitors, for which other bridges are not satisfactory.

— HENRY P. HALL

[5] R. D. Cutkosky, "Four-Terminal Pair Networks as Precision Standards," *IEEE Transactions on Communications and Electronics*, #70, January, 1964, p 19.

Henry P. Hall, Group Leader of the Impedance Measurement Group in GR's Engineering Department, attended both Williams College and M.I.T., receiving in 1952 his A.B. from the former and his S.B. and S.M. in Electrical Engineering from the latter. He came to General Radio first as a cooperative student in 1949, becoming a full-time development engineer in 1952, Section Leader in 1963, and Group Leader in 1964. A member of IEEE, he has served on various committees of that and other professional and industry organizations. He is a member of Phi Beta Kappa, Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

SPECIFICATIONS

Quantity	*Frequency*	*Range*	*Accuracy* *
Capacitance	120 Hz internal	0 to 0.11 F	$\pm 1\% \pm 1$ pF, smallest division 2 pF; resident ("zero") capacitance approximately 4 pF
		0.11 F to 1.1 F	$\pm 2\%$
	40 Hz to 120 Hz external (useful down to 20 Hz with reduced accuracy)	0 to 1.1 F	Same as above with suitable generator
	120 Hz to 1 kHz external	0 to 1 F $\left(\frac{100}{f_{Hz}}\right)^2$	$\pm 1\% \pm 1$ pF with suitable generator and precautions
Dissipation Factor	120 Hz internal or 40 Hz to 120 Hz	0 to 10 $\frac{f_{Hz}}{120}$	$\pm 0.001 \pm 0.01\, C \pm 2\%$
	120 Hz to 1 kHz	0 to 10	$\left[\pm 0.001 \pm 0.01\, C\right] \frac{f_{Hz}}{120} \pm 2\%$

* C is expressed in farads.

Lead-Resistance Error (4-terminal connection): Additional capacitance error of less than 1% and D error of 0.01 for a resistance of 1Ω in each lead on all but the highest range, or 0.1Ω on the highest range.

Internal Test Signal: 120 Hz (synchronized to line) for 60-Hz model; 100 Hz for 50-Hz model. Selectable amplitude less than 0.2 V, 0.5 V, or 2 V. Phase reversible.

External Test Signal: 20 Hz to 1 kHz with limited range (see above).

Internal DC Bias Voltage and Voltmeter: 0 to 600 V in 6 ranges.

Voltmeter Accuracy: $\pm 3\%$ of full scale.

Internal DC Bias Current: Approximately 15 mA maximum.

Ammeter Range: 0 to 20 mA in 6 ranges. Can detect ½-μA leakage.

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SPECIFICATIONS *(Cont'd)*

Ammeter Accuracy: ±3% of full scale.
External Bias: 800 V maximum.
Power Required: 105 V to 125 V or 210 V to 250 V, 60 Hz, 18 W maximum. Models available for 50-Hz operation.
Accessories Supplied: Four-lead and shielded two-lead cable assemblies.
Accessories Required: None for 120-Hz measurements. The Type 1311-A Oscillator is recommended for measurement at spot frequencies, the Type 1310-A Oscillator for continuous frequency coverage.

Cabinet: Flip-Tilt; relay-rack model also is available.
Dimensions: Portable model — width 16¼, height 15, depth 9 inches (415, 385, 230 mm); rack model — width 19, height 14, depth behind panel 6⅛ inches (485, 355, 160 mm), over-all.
Net Weight: Portable model, 26 lb (12 kg); rack model, 28 lb (13 kg).
Shipping Weight: Portable model, 34 lb (15.5 kg); rack model, 43 lb (20 kg).

Catalog Number	*Description*	*Price in USA*
1617-9701	**Type 1617 Capacitance Bridge,** Portable Model (115 V, 60 Hz)	**$1195.00**
1617-9286	**Type 1617 Capacitance Bridge,** Portable Model (230 V, 60 Hz)	**1195.00**
1617-9206	**Type 1617 Capacitance Bridge,** Portable Model (115 V, 50 Hz)	**on request**
1617-9266	**Type 1617 Capacitance Bridge,** Portable Model (230 V, 50 Hz)	**on request**
1617-9820	**Type 1617 Capacitance Bridge,** Rack Model (115 V, 60 Hz)	**1195.00**
1617-9296	**Type 1617 Capacitance Bridge,** Rack Model (230 V, 60 Hz)	**1195.00**
1617-9216	**Type 1617 Capacitance Bridge,** Rack Model (115 V, 50 Hz)	**on request**
1617-9276	**Type 1617 Capacitance Bridge,** Rack Model (230 V, 50 Hz)	**on request**

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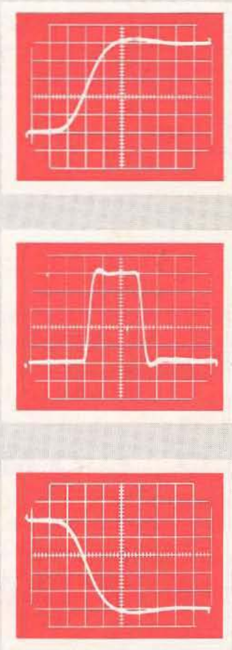

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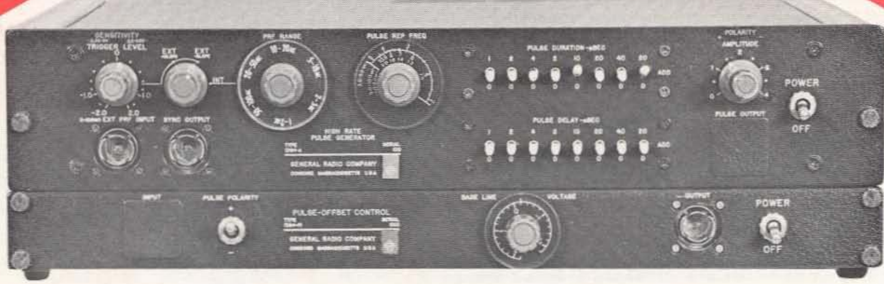

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REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

Figure 1. Panel view of the Type 1394-Z High-Rate Pulse Generator, consisting of *(top)* the Type 1394-A High-Rate Pulse Generator and *(bottom)* Type 1394-P1 Pulse-Offset Control.

VHF PULSE GENERATOR

NEW CIRCUITS YIELD HIGH PERFORMANCE AT MODERATE COST

The continuing rapid expansion of digital techniques into the vhf region has created a demand for generators of high-prf, fast-rise-time pulses. General Radio's new vhf pulse generator makes maximum use of standard, economical components in straightforward yet novel circuits, resulting in state-of-the-art performance at a moderate price.

The new Type 1394-A High-Rate Pulse Generator meets the growing requirements for test sources for high-speed computers and data-transmission and processing systems. Its important features include:

1. High repetition rate; 1- to 100-MHz range internally generated; dc to 100 MHz with external drive.
2. Fast rise time: 2 nanoseconds.
3. Duty ratios up to 96%.
4. Internal prf generator with excellent frequency stability.
5. Controls for precise synchronization with external clock signals.
6. Calibrated controls for pulse repetition frequency, amplitude, delay, and duration.

Performance of this order is essential in the design and test of high-speed digital systems. In addition to its use in computer development, this pulse generator has many applications in such fields as data transmission, modern radar systems, nuclear instrumentation, and component testing.

Through the development of new circuits, the above features are made available in a pulse generator of comparatively modest price. Both the prf oscillator and the bistable output circuit embody new ideas; the delay functions are performed by lengths of coaxial cable.

A companion instrument, the Type 1394-P1 Pulse-Offset Control (page 7), is available for those applications where dc output coupling is required. The combination is the Type 1394-Z.

Figure 2 is a simplified block diagram of the generator. A pulse train from the prf oscillator is applied to

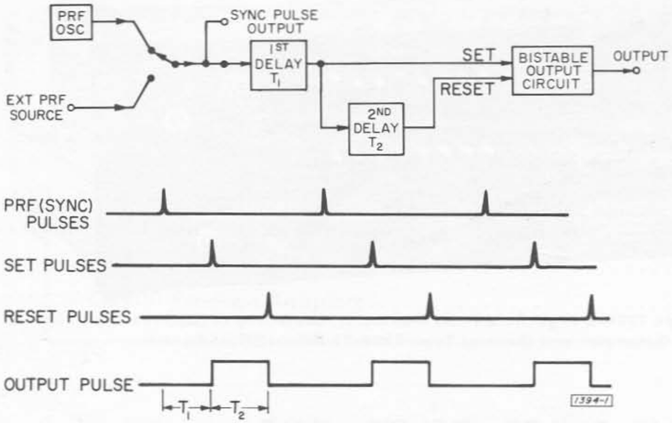


Figure 2. Elementary block diagram with waveforms.

the first delay circuit, which delays the pulses by time T_1. The delayed pulses are applied to the bistable output circuit as set pulses and also to the second delay circuit, which delays the train further by an amount T_2 to form reset pulses. The first delay is thus the delay between the sync pulse and the leading edge of the output pulse, and the second delay is the duration of the output pulse. On page 5 is a discussion of the individual circuits, illustrating their unique aspects and pointing out their advantages to the user.

APPLICATIONS

One important application of the Type 1394-A High-Rate Pulse Generator is in testing of complementary flip-flops or scalers. Tests of maximum input pulse frequency as a function of pulse duration, amplitude, supply voltage, etc are easily made with this generator. Double- or triple-pulse testing at low repetition rates, on the other hand, although it gives an indication of the maximum frequency at which a digital circuit may operate properly, is not completely adequate. Only sustained operation at the maximum frequency can show the effects of self-biasing due to ac coupling and nonlinearities and also the effects of power dissipation due to rapid switching, both of which may be significant factors in circuit performance.

Another application of the instrument is as a clock-pulse generator for a digital system. Both the precision of setting and the stability of the prf oscillator are quite important in this application.

Many digital devices have the properties of a threshold detector; when a pulse exceeds a certain voltage level, the circuit acts. Threshold circuits can be conveniently tested with the Type 1394-A/P1, since the combination of a stepped pulse attenuator and a smooth, precise offset control allows continuous adjustment of pulse level through a range of +6 to −6 volts.

The input circuits include a calibrated trigger-level control and a slope-polarity switch. Sensitivity is better than 0.4 volt, peak-to-peak. Consequently, when the instrument is triggered by an external signal, optimum operation can be obtained over wide ranges of input waveform and voltage level. This capability is important when the instrument is used as a regenerator in a system.

The circuit configurations were chosen to provide the high performance necessary to meet today's vhf pulse needs and to do so without pushing conventional techniques to the limits, which is never satisfactory from either the cost or the reliability standpoint. These circuits provide two additional advantages that improve on present practice.

The first is freedom from the duty-ratio limits usually imposed by delay-circuit recovery or output-circuit power-dissipation limits. This instrument is very convenient to operate, and it is satisfying to know that the output circuit will operate in any environment and with any combination of control settings without being on the edge of a dissipation limit.

The second advantage is that the instrument has accurately calibrated controls and is stable in operation. Because the settings of the panel controls tell the operator what the generator is doing, he can devote his attention to the system under test with a minimum of attention to the pulse source, and without the necessity of tying up an expensive scope to monitor the pulses.

CIRCUIT DESCRIPTION

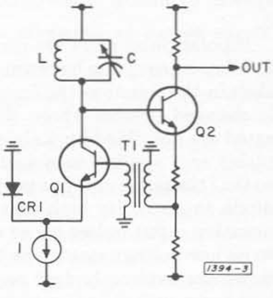


Figure 3. PRF Circuit, simplified circuit diagram.

PRF Oscillator*

The circuit of the prf oscillator is shown in elementary form in Figure 3.

A constant-current source, *I*, is switched between diode *CR1* and transistor *Q1* by a large sinusoidal signal on the base of *Q1*. The resulting square wave of collector current drives the LC tuned circuit at its resonant frequency, producing a high-amplitude sinusoidal voltage at the base of *Q2*. The collector of *Q2* supplies the sinusoidal output voltage. The feedback loop is closed by the application of a fraction of *Q2's* emitter voltage to the base of *Q1*, via inverting transformer *T1*, which is a wideband one-to-one transformer of the type described by Ruthroff[1].

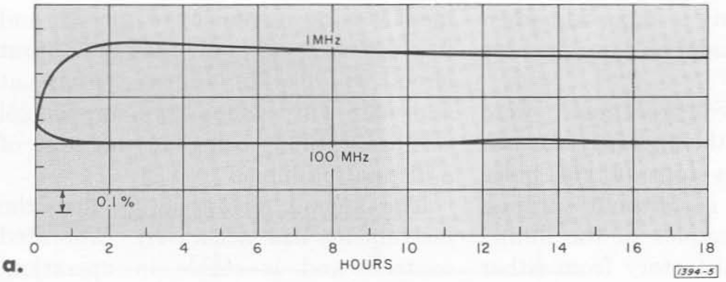
The advantages of this circuit configuration are that the simplest resonant circuit is used for frequency determination and that the amplitude is proportional to the current from the constant-generator current, which is the collector of a transistor used in the automatic amplitude control. Figure 4 shows the variation of oscillator frequency with warmup time and with line voltage.

The Delay Function

Pulses propagating through polyethylene-dielectric coaxial cable are delayed by one nanosecond for approximately every 20 cm (7.8 in) of cable length. Since the delays required by this generator are less than 100 nanoseconds, they can be provided by reasonable lengths of cable. The delay circuits consist of lengths of coaxial cable cut for 1, 2, 4, 2, 10, 20, 40, and 20 nanoseconds. These can be switched in or out of the signal path to change the delay in one-nanosecond increments from zero to 99 nanoseconds.

Cable-delay circuits have the advantages of economy, high duty ratio, high prf, and accuracy as compared to the usual lumped-constant delay circuit. The conventional lumped circuit charges a reactance to produce a time delay. The time taken to discharge the reactance before the next delay period can start places a limit on attainable duty ratios. At high prf's the rates of change of energy to and from the reactance become large and make it desirable to use very small inductances or capacitances, introducing inaccuracies due to uncontrolled stray reactances. Attempts to improve prf or

* Patient Applied For.
[1] C. L. Ruthroff, "Some Broad-Band Transformers," *Proceedings of the IRE*, August 1959, p 1337 ff.



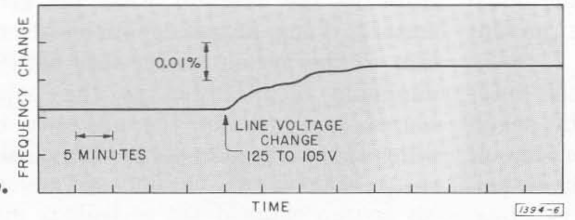


Figure 4. Stability records. (a) Frequency variation on 1- and 100-MHz ranges from cold start. (b) Frequency variation on 100-MHz range due to line-voltage change.

duty ratio by a decrease in the voltage or current swing on the reactance also result in decreased accuracy, because the error in detection of the charged voltage or current level does not decrease in proportion to the signal swing.

Cable-delay circuits operate on a much different principle from that of the conventional delay circuit, and duty-ratio restrictions are not applicable in the same sense as they are to the latter. The delay of a pulse train in cables may even exceed its period. Accuracy of cable-produced delay depends upon the stability of the cable length and upon the cable dielectric constant. Cable-produced delays are accurately known and are much more stable than those obtainable by conventional circuits operating at very high prf's.

Pulse Regeneration

When cables are used for delay, the pulse shape deteriorates as the pulse is propagated along the line and must be restored. Pulse shapes are regenerated at several points in the system by means of the circuit shown in Figure 5.

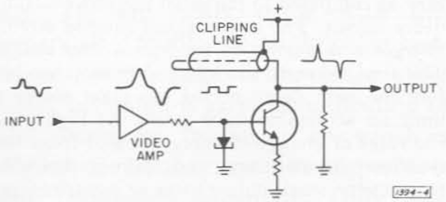


Figure 5. Pulse-Regeneration Circuit.

Bipolar pulse pairs are used in this generator to allow ac coupling between stages without a dc shift in the signal as the frequency of the signal is changed. Pulses whose shapes have deteriorated are amplified by a class-A broadband amplifier and applied to a bistable tunnel diode, so that the positive input pulses cause the tunnel diode to go to its high-voltage state, and the negative input pulses cause the diode to return to its low-voltage state. The tunnel diode transitions are extremely fast and, as a result, the higher-frequency components of the signal, which were lost in the delay line, are restored. The fast pulse from the tunnel diode is applied to a clipping line, which differentiates the signal to reproduce the desired bipolar pulse.

The Bistable Output Circuit

Figure 6 is a simplified schematic diagram of the bistable output circuit. The set and reset pulses are applied to the tunnel diode through 100-ohm coaxial cable, and, since the tunnel diode appears as a very low impedance (approximately 5 ohms except for the extremely short time that it is switching between states), the pulse voltages appear almost equally on both bases. The differential amplifier *Q3-Q4* amplifies only the signal difference between its bases and not a voltage applied to both bases. Thus, although the tunnel diode voltage is amplified, the set and reset pulses are not and do not appear in the output. This type of connection

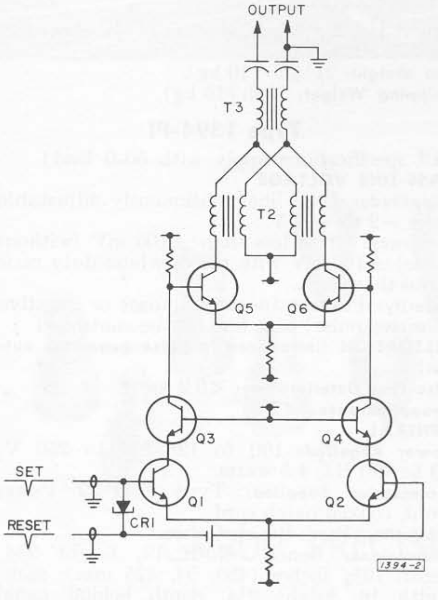


Figure 6. Simplified schematic of output circuit.

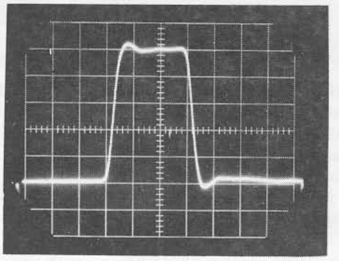


Figure 8. Output-pulse waveform. A 15-ns pulse at 20-MHz repetition rate. Horizontal scale, 5 ns per major division.

greatly reduces the sensitivity of output wave shape to variations in the set or reset pulses.

Q5 and *Q6* are grounded-base stages that allow *Q3* and *Q4* to operate into very low collector impedances, for maximum bandwidth. The grounded-base stages drive a second differential amplifier, *Q7* and *Q8*, which is coupled to the output by transformers *T2* and *T3*.

Transformer *T2* is a 2:1 balanced transformer constructed along the lines suggested by Ruthroff.[1] It has an extremely wide bandwidth, from less than 100 kHz to a few hundred MHz. Figure 7 is a photograph of *T2* with its protective cover removed. *T3* is a balanced-to-unbalanced transformer.[2] Figure 8 shows the output-pulse waveform. The output circuit includes a precision 50-ohm attenuator, which drops the output from a maximum of four volts to zero in half-volt steps.

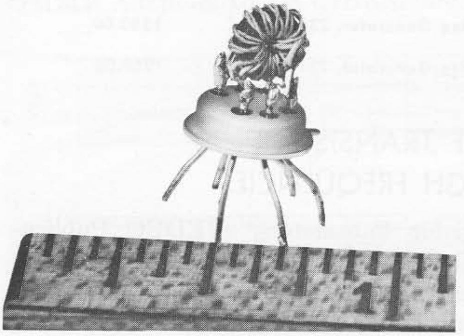


Figure 7. Photograph of output transformer, T2. Size is indicated by scale on the ruler.

Pulse-Offset Control

The Type 1394-P1 Pulse-Offset Control is a companion instrument to the Type 1394-A High-Rate Pulse Generator (see Figure 1). Interconnections are made at the rear of the instruments since both are designed so that all front-panel connectors are easily transferred to the rear panel. The Pulse-Offset Control consists of a peak-voltage detector, reference voltage generator, and a high-gain control-amplifier. It inserts a dc component in the pulse output of the pulse generator so that the base line of the pulse is regulated to be equal to the reference voltage. This combination of instruments meets the needs of those applications that require dc coupling. The pulse-offset voltage is continuously adjustable from −2 to +2 volts.

— J. K. Skilling

[1] *Ibid.*
[2] Lewis and Wells, *Millimicrosecond Pulse Techniques*, 2nd Edition, Pergamon Press, 1959, p 104 ff.

Note: A brief biography of James K. Skilling, author of the foregoing article, appeared in the March 1966 issue of the *Experimenter*. — Editor

SPECIFICATIONS

Type 1394-A

PULSE REPETITION FREQUENCY

Internally Generated: 1.0 MHz to 100 MHz; six ranges in 1-2, 2-5, 5-10 sequence. Continuous coverage, ±5% of setting. Jitter, <0.1 ns, peak.

Externally Controlled: dc to 100 MHz, 0.4 to 4.0 V, p-to-p, amplitude range plus 10 to 1 attenuator,

(*Continued*)

SPECIFICATIONS (Cont'd)

1 W max. 50 Ω, choice of ± slope, and trigger level from −2 to +2 V.

SYNCHRONIZING-PULSE CHARACTERISTICS
Description: Bipolar pulses, leading edge of positive pulse is reference.
Duration: 4 ns, typical.
Amplitude: Approx 250 mV, p-to-p, into 50 Ω.
Delay (between sync pulse and leading edge of output pulse): 0 to 99 ns in 1-ns steps, ±2.5% ± 1-ns accuracy. No restriction on ratio delay period. Jitter, <0.1 ns, peak.
Residual Delay: 35 ns, typically.

OUTPUT-PULSE CHARACTERISTICS (all specifications apply to 50-Ω load)
Duration: 4 to 99 ns in 1-ns steps, ±2.5% ± 1-ns accuracy. Jitter, <0.1 ns, peak.
Rise and Fall Times: 2.0 ns ±20%.
Voltage: Ac coupled. 0 to 4 V in calibrated ½-volt steps. Plus or minus polarity.
Duty Ratio: Limited only by rise-plus-fall time.
Overshoot: 12% typically.
Droop: <±10% at maximum duration.

GENERAL
Power Required: 100 to 125/200 to 250 V; 50 to 400 Hz; 24 W.
Accessories Supplied: Type CAP-22 Power Cord; spare fuses.
Mounting: Rack-Bench Cabinet.
Dimensions: Bench, width 19, height 3 15/16, depth 16¾ inches (485, 100, 425 mm); rack, width 19, height 3½, depth behind panel 14½ inches (485, 89, 370 mm), over-all.
Net Weight: 21½ lb (10 kg).
Shipping Weight: 34 lb (15 kg).

Type 1394-PI

(All specifications apply with 50-Ω load)

BASE-LINE VOLTAGE
Amplitude: Base line continuously adjustable from −2 to + 2 V.
Accuracy: Error less than ±100 mV (without pulse) ±100 mV with pulses whose duty ratio is less than 90%.
Polarity: Positive (negative pulse) or negative (positive pulse) base line can be controlled.

DISTORTION (introduced in pulse-generator output)
Rise-Time Deterioration: <0.2 ns.
Droop Increase: <2%.

GENERAL
Power Required: 100 to 125/200 to 250 V; 50 to 400 Hz; 4.5 watts.
Accessories Supplied: Type CAP-22 Power Cord, coaxial patch cord.
Mounting: Rack-Bench Cabinet.
Dimensions: Bench, width 19, height 2⅛, depth 16¾ inches (485, 54, 425 mm); rack, width 19, height 2⅛, depth behind panel 14½ inches (485, 54, 370 mm), over-all.
Net Weight: 12¼ lb (6 kg).
Shipping Weight: 17 lb (8 kg).

Catalog Number	*Description*	*Price in USA*
1394-9801	**Type 1394-A High-Rate Pulse Generator,** Bench Model	$995.00
1394-9811	**Type 1394-A High-Rate Pulse Generator,** Rack Model	995.00
1394-9611	**Type 1394-P1 Pulse-Offset Control,** Bench Model	255.00
1394-9621	**Type 1394-P1 Pulse-Offset Control,** Rack Model	255.00
1394-9911	**Type 1394-Z High-Rate Pulse Generator,** 115 volts Bench Model	1250.00
1394-9912	**Type 1394-Z High-Rate Pulse Generator,** 115 volts Rack Model	1250.00
1394-9913	**Type 1394-Z High-Rate Pulse Generator,** 230 volts Bench Model	1250.00
1394-9914	**Type 1394-Z High-Rate Pulse Generator,** 230 volts Rack Model	1250.00

MEASUREMENT OF TRANSISTOR PARAMETERS AT HIGH FREQUENCIES

Users of the Type 1607-A Transfer-Function and Immittance Bridge for transistor measurements will be interested in a recent JEDEC standard entitled "A Method for the Measurement of Small-Signal High-Frequency Transistor Parameters," JEDEC Publication No. 55, March 1966, $1.10. Copies can be obtained from EIA, Engineering Department, 2001 Eye Street, N.W., Washington, D.C. 20006.

COAXIAL MICROWAVE NEWS

Figure 1. GR900 Precision Adaptors.

GR900 PRECISION ADAPTORS

The TYPES 900-QSCJ* and -QSCP* Adaptors to Type SC are the latest additions to the GR900 series of precision adaptors to military-type connectors. Adaptors previously announced[1] include the TYPES 900-QNJ, -QNP Adaptors to Type N, the TYPES 900-QBJ, -QBP Adaptors to Type BNC, the TYPES 900-QTNJ, -QTNP Adaptors to Type TNC, and the TYPES 900-QCJ, -QCP Adaptors to Type C.

The new TYPES 900-QMMJ and -QMMP Adaptors to OSM/BRM are for use with the nonmilitary miniature connectors called group A by Brinton[2]. These connectors include those specified as ASM, BRM, ESCAM, MICRO, MOB-50, NPM, OSM, SRM, and STM.

Each adaptor contains a GR900 Precision Coaxial Connector, a specially designed continuous transition between

* The suffix J indicates that the adaptor is female (contains a jack) and the suffix P indicates that the adaptor is male (contains a plug).

[1] *General Radio Experimenter*, November 1963 and January 1965.

[2] Brinton, J. B., Jr., "Miniature Coaxial Components," *MicroWaves*, February 1965, p 32.

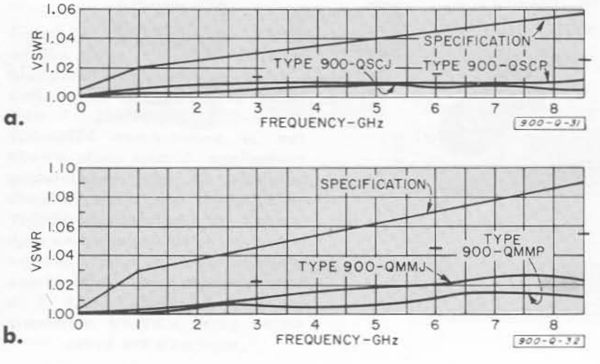


Figure 2. Measured data on sample lots of (a) Types 900-QSCJ and -QSCP Adaptors to SC and (b) Types 900-QMMJ and -QMMP Adaptors to OSM/BRM. Solid curves are averages; bars at 3, 6 and 8.5 GHz are maximums. Measurement error is less than ¼th of the specifications.

In measurement and standards laboratories, many different types of coaxial connectors are encountered. For maximum utility, a basic line of measuring equipment based on precision connectors must include adaptors to other commonly used types. Recognizing this, General Radio has provided several adaptors from the GR900 Precision Coaxial Connector. We now add two new types, SC and OSM/BRM, described in the following article. We have also included a discussion of the effect of mating dimensions and gaps, which, we hope, will help to answer some of the questions that have been asked about VSWR errors from these sources.

line sizes, and a low-VSWR version of the applicable (jack or plug) SC or OSM/BRM connector. The VSWR performance of these adaptors is shown in Figure 2.

PRECISION ADAPTOR APPLICATIONS

By means of precision adaptors, the excellent performance of GR900 coaxial standards and instruments can be extended to measurements on devices equipped with other types of connectors.

Slotted-Line Measurements

With a set of GR900 Precision Adaptors in combination with a TYPE 900-LB Precision Slotted Line one can make accurate impedance measurements through 14 different coaxial connectors as shown in Figure 3: the military types N, BNC, TNC, C, and SC, jacks and plugs; the OSM/BRM types or equivalents, jack and plug; the general-purpose GR874; and the precision GR900. Figure 4a shows the specified performance of the various adaptor-slotted-line combinations; typical VSWR is about half that specified, as illustrated in Figure 4b for a Type N combination.

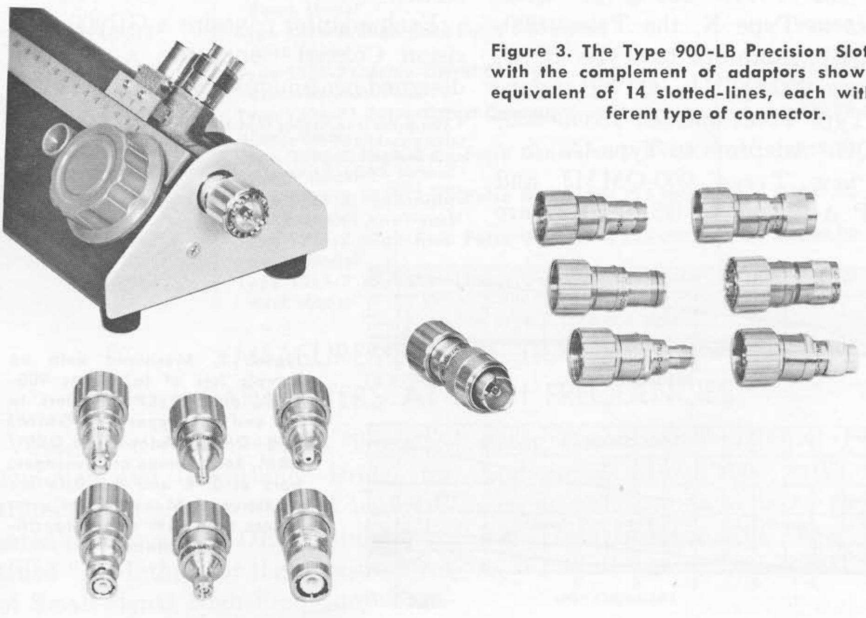
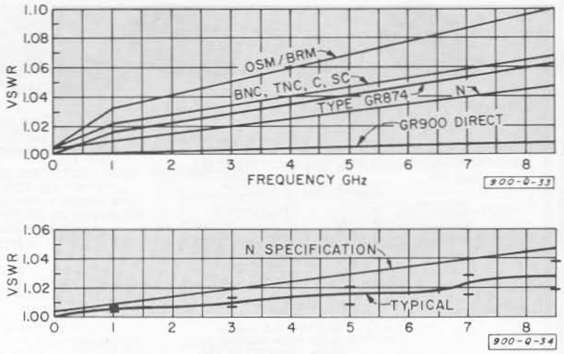


Figure 3. The Type 900-LB Precision Slotted Line with the complement of adaptors shown is the equivalent of 14 slotted-lines, each with a different type of connector.



a.

b.

Figure 4. (a) Specified residual VSWR of the Type 900-LB Precision Slotted Line in combination with various GR900 Precision Adaptors. (b) Typical data on a sample lot of the slotted line, Type-N-jack-adaptor combination. Bars above curve at 1, 3, 5, 7 and 8.5 GHz are maximum VSWR's; bars below curve are averages.

Matched Terminations

Similarly, the precision adaptors in combination with a Type 900-W50 50-ohm Standard Termination provide low-vswr terminations for the various connector types. Figure 5 shows the specified performance.

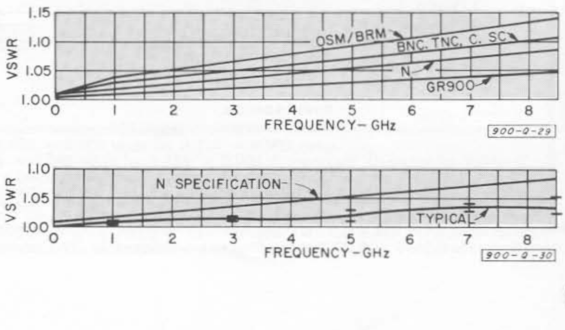
Advantages

There are two important advantages to utilizing precision adaptors as described above: accuracy and economy. The accuracy of measurement through each connector type is usually better than that provided by slotted lines or terminations designed specifically for the connector type of interest. This is because (1) the Type 900-LB Slotted Line and the Type 900-W50 Termination exhibit very low residual vswr's, which can be accurately calibrated at the GR900 Connector reference planes; (2) the continuous transitions in the adaptors are nearly reflectionless, and (3) connectors of the series being adapted to are optimally designed.

The advantage of economy is obvious: the cost of one slotted line, one termination and several adaptors is far less than the cost of individual slotted lines and terminations for each connector type.

MATING DIMENSIONS

When two GR900 Precision Connectors are mated, the vswr introduced by one of the connectors is not influenced by variations in the dimensions of the second connector. On the other hand, when two Type N connectors are mated, the vswr introduced by the jack connector is directly dependent on the diameter of the pin of the plug-connector inner conductor against which the jack inner-conductor fingers rest. Similarly, the vswr introduced by the

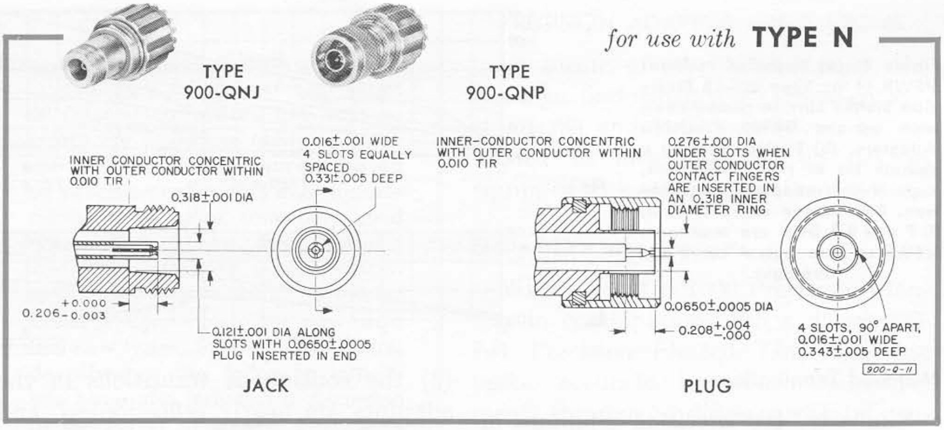


a.

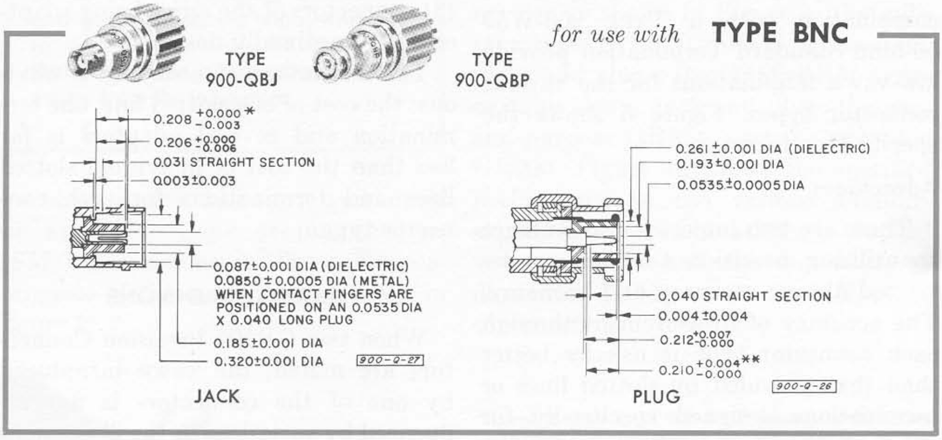
b.

Figure 5. (a) Specified VSWR of the Type 900-W50 50-ohm Standard Termination in combination with various GR900 Precision Adaptors. (The 900-W50/900-Q874 combination is not shown since nearly equivalent performance can be obtained directly with the Type 874-W50BL Termination.) (b) Typical data on a sample lot of the termination, Type N plug adaptor combination. Bars above curve at 1, 3, 5, 7, and 8.5 GHz are maximum VSWR's; bars below curve are averages.

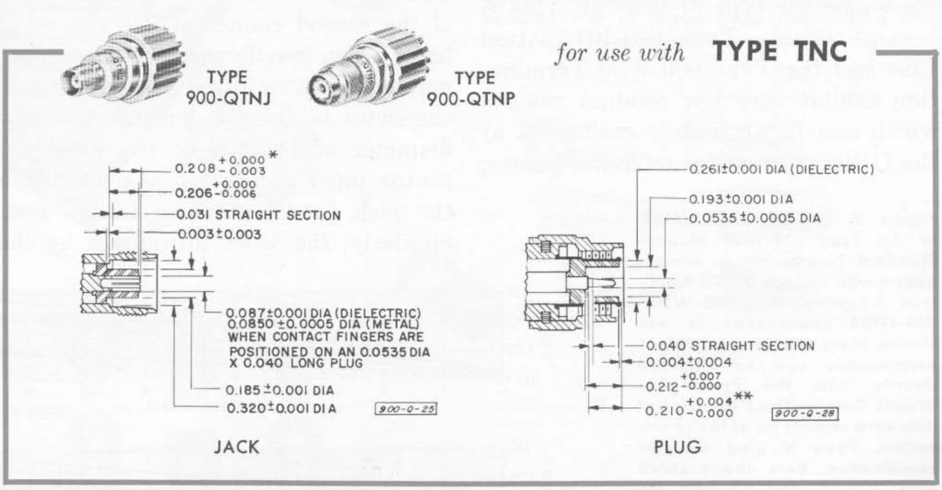
for use with **TYPE N**



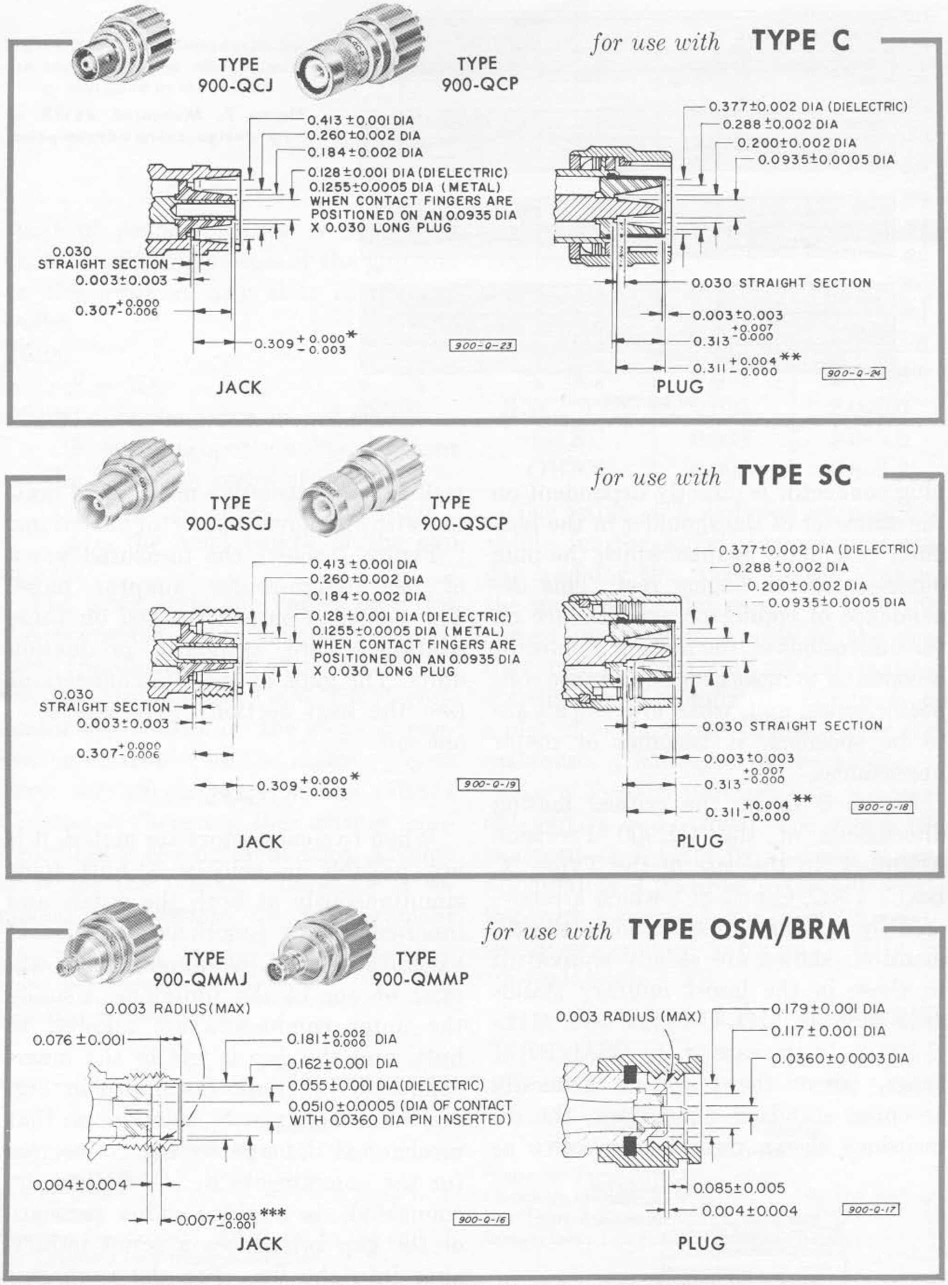
for use with **TYPE BNC**



for use with **TYPE TNC**



* Inner conductor has 4 equally spaced slots 0.008 ±0.001 wide by 0.187 ± 0.005 deep.
** Outer conductor has 6 slots 60° apart, 0.015 ± 0.001 wide by 0.235 ± 0.003 deep; inner diameter in region of contact-fingers is 0.2650 ± 0.0005 when fingers are inserted in a 0.3200 inner-diameter ring.



* Inner conductor has 4 slots, equally spaced, 0.012 ± 0.001 wide by 0.210 ± 0.005 deep.
** Outer conductor has 6 slots, 60° apart, 0.016 ± 0.001 wide by 0.255 ± 0.005 deep; inner diameter in region of contact-fingers is 0.3820 ± 0.0005 when fingers are inserted in a 0.413 inner-diameter ring.
*** Inner conductor has 4 equally spaced slots 0.008 ± 0.001 wide by 0.078 ± 0.005 deep.

Figure 6. Critical mating dimensions of low-VSWR connectors used with the GR900 Connector on the precision adaptors. (All dimensions are in inches.)

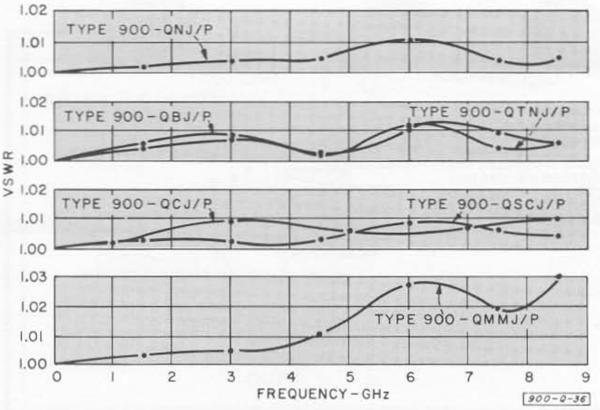


Figure 7. Measured VSWR of near-design-center adaptor pairs.

plug connector is directly dependent on the diameter of the shoulder in the jack outer conductor against which the plug outer-conductor fingers rest. This dependence of connector performance on the dimensions of the mating connector is common to most general-purpose connector series, and, when low VSWR's are to be specified, it becomes of major importance.

Figure 6 shows the critical mating dimensions of the GR900 Precision Adaptors. In the case of the Types N, BNC, TNC, C and SC, which are covered by military specifications, the dimensions shown are closely equivalent to those in the latest military standards such as MIL-C-23329 and MIL-C-39012. In the case of the OSM/BRM types, where there are no generally accepted standard dimensions, the dimensions shown provide low VSWR as well as nondestructive mechanical mating with all known connector variations.

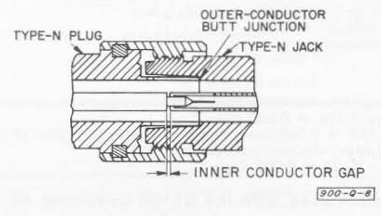


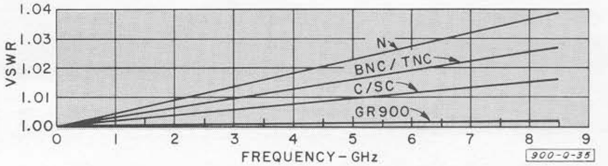
Figure 8. Gap in Type N connector junction.

Figure 7 shows the measured VSWR of near-design-center adaptor pairs. The GR900 Connectors used on these adaptors were standard production units. The gaps in the UG connections (see the next section) were less than one mil.

GAPS

When two connectors are mated, it is not possible to achieve a butt joint simultaneously at both the outer- and inner-conductor junctions. Because of axial mechanical tolerances, a gap will exist at one of the junctions. Usually the outer conductors are allowed to butt, and the gap is left in the inner-conductor junctions (as shown in Figure 8 for the Type N junction) so that mechanical damage to the connectors (or the components to which they are connected) is avoided. This presence of the gap introduces a series inductance into the line. For the connector types of interest, the VSWR resulting from this residual inductance is a linear function of frequency and of the axial dimension of the gap. The con-

Figure 9. VSWR's introduced by the maximum gap dimensions, g, indicated in the table.



stant of proportionality is dependent on the radial dimension of the gap and on the width of any slots in the gap walls[3].

Thus,

$S = Kfg$

where: S is the VSWR in per cent,

K is the proportionality constant for the connector series of interest,

f is the frequency in GHz, and

g is the axial length of the gap in mils.

The table gives the values of K for the connector types covered by military specifications. Values of K are not included for the OSM/BRM connector types, because the steps in conductor diameters at the mating planes alter the effect, nor for the GR874 Connector, because the mating configuration is of a different kind. A value of K for the GR900 Connector is included to illustrate the improvement gained through the use of precision connectors.

Connector Type	*K*	*Spread in g (mils)*
N	0.051	2.0–9.0
BNC, TNC	0.035	2.0–9.0
C, SC	0.021	2.0–9.0
GR900	0.008	0.8–3.2

The value of K is based on nominal values for the radial dimension of the gap and for the width of the slots in the gap walls. The spread in g is the spread in the axial length of the gap that results from the critical mating dimensions of Figure 6. Figure 9 is a plot of VSWR versus frequency for the maximum g indicated in the table.

For a Type N junction with a nominal gap (g = 5.5 mils), the VSWR introduced at 7 GHz by the gap is approximately 1.02. For a GR900 junction with a nominal gap (g = 2.0 mils), the corresponding VSWR at 7 GHz is approximately 1.001.

— T. E. MacKenzie

[3] MacKenzie, T. E., and Sanderson, A. E., "Some Fundamental Design Principles For the Development of Precision Coaxial Standards and Components," *IEEE Transactions on Microwave Theory and Techniques*, Vol MTT-14, No 1, January 1966, p 29–39.

SPECIFICATIONS

Type 900-QSCJ Adaptor (contains SC jack)

Frequency Range: DC to 8.5 GHz.
VSWR: Less than $1.005 + 0.015 \times f_{GHz}$ to 1 GHz; $1.015 + 0.005 \times f_{GHz}$, 1 to 8.5 GHz.
Electrical Length: 5.03 ±0.05 cm to the end of the Type SC jack inner conductor.
Voltage: 1000 V peak.
Power (Average): 7 kW up to 1 MHz; 7 kW/ $\sqrt{f_{GHz}}$ above 1 MHz.
Dimensions: Length, 2 inches (51 mm); maximum diameter, 1 1/16 inches (27 mm).
Net Weight: 3 1/2 ounces (100 grams).

Type 900-QSCP Adaptor (contains SC plug)

Same as Type 900-QSCJ except:
Electrical Length: 5.60 ±0.05 cm to the end of the Type SC plug outer conductor.
Dimensions: Length, 2 1/8 inches (54 mm).

Type 900-QMMJ Adaptor (contains OSM/BRM jack)

Frequency Range: DC to 8.5 GHz.
VSWR: Less than $1.005 + 0.025 \times f_{GHz}$ to 1 GHz; $1.022 + 0.008 \times f_{GHz}$, 1 to 8.5 GHz.

(*Continued*)

the Experimenter

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SPECIFICATIONS (cont'd)

Electrical Length: 4.67 ±0.05 cm to the outer conductor junction.

Dimensions: Length, 1⅞ inches (48 mm); maximum diameter 1$\frac{1}{16}$ inches (27 mm).

Net Weight: 2½ ounces (70 grams).

Type 900-QMMP Adaptor (contains OSM/BRM plug)

Same as Type 900-QMMJ except:

Electrical Length: 4.78 ±0.05 cm to the outer conductor junction.

Catalog Number	*Description*	*Price in USA*
0900-9713	**Type 900-QSCJ Adaptor**	**$75.00**
0900-9813	**Type 900-QSCP Adaptor**	**85.00**
0900-9723	**Type 900-QMMJ Adaptor**	**75.00**
0900-9823	**Type 900-QMMP Adaptor**	**80.00**

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THE GENERAL RADIO

Experimenter

IMPROVED GR874 SLOTTED LINE

PRECISION COAXIAL TUNERS

TE_{11}-MODE RESONANCE IN TYPE 900-BT CONNECTOR

HIGH-RESOLUTION WAVE-ANALYSIS RECORDING

NEW DECADE CAPACITOR

10-pF REFERENCE STANDARD

VOLUME 40 · NUMBER 8 / AUGUST 1966

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August 1966

the GR Experimenter

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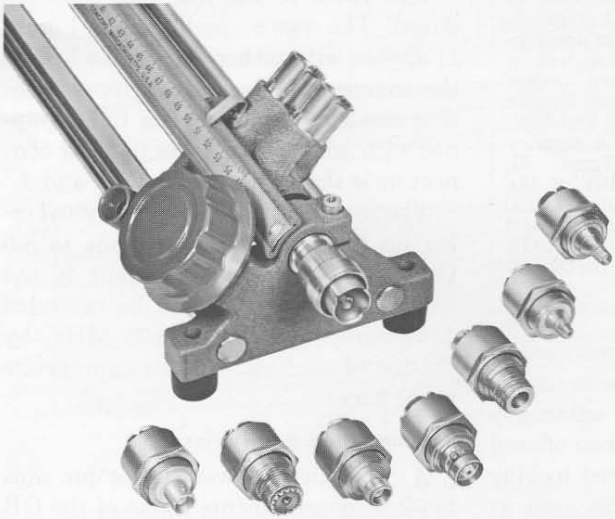
COAXIAL MICROWAVE NEWS



THE IMPROVED GR874 SLOTTED LINE

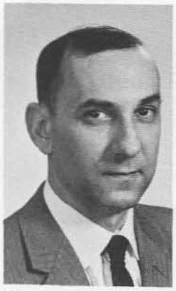
The slotted line is the traditional workhorse of the microwave measurement industry. It can be used to measure a great number of parameters, over an extremely broad band of frequencies and with good accuracy. The slotted line has its own internal impedance reference, an accurately constructed coaxial line, which is essentially free from aging effects. It is used to measure VSWR, reflection coefficients, transmission coefficients, impedance, and admittance (resistance, inductance, and capacitance) of both passive and active networks (diodes, tubes, transistors); electrical length of two-ports; phase delay; insertion loss or attenuation of networks, cables for example; and dielectric constants of materials. It can also be used as a precision phase shifter and as a wavemeter.

The slotted line can also be employed for accurate sweep-frequency measurements of VSWR[1], performing the function of a reflectometer over a wide frequency range.

GR manufactures two slotted lines, one based on the GR874 connector, the other on the GR900 precision connector. The GR874 line, while not so accurate as the precision type, is satisfactory for most everyday measurements, and thousands are now in use throughout the world. It is particularly well suited to the student laboratory, where its modest price is a boon to the budget.

The GR874 Slotted Line has recently been redesigned to reduce the residual VSWR, to increase the high-frequency range, and to improve the constancy of probe coupling. The model number has been changed to TYPE 874-LBB. Over

[1] R. H. Behle, L. J. Smith, "Studies of Cold-Test Procedures Used in the Development of the L-4061 Crossed Field Amplifier," BTL Internal Report, No MM-63-2843-7, p 9; Bell Telephone Laboratories, Laureldale, Pa.



John Zorzy received his B.S. in Physics from George Washington University in 1948 and his M.S. from Tufts College in 1950. From 1950 to 1960 he was engaged in the design and development of radar, antennas, and microwave devices at Trans-Sonics, Hughes Aircraft, Avco, and Raytheon. He is a member of IEEE, Sigma Xi, Sigma Pi Sigma, and a member of the Precision Connector Subcommittee of IEEE and Committee JS-9 of JEDEC. He joined the Development Engineering Staff at GR in 1963 and is Section Leader in the Microwave Group.

the past several years many mechanical improvements have also been incorporated, and, recently, an externally adjustable probe tuner has been offered as an accessory. The improved locking connector, Type 874-BBL, is used at both ends of the slotted line.

These improvements result in an extremely versatile instrument with more than adequate performance for a general-purpose line. This line can be converted to use any of the popular UG connectors in a matter of seconds, through a Type 874-Q low-VSWR adaptor. These adaptors are available, both plug and jack types, for Types BNC, C, HN, LC, LT, Microdot, N, OSM, SC, TNC, ⅞″-UHF line, GR900, and Amphenol 7-mm Precision Connectors.

Performance

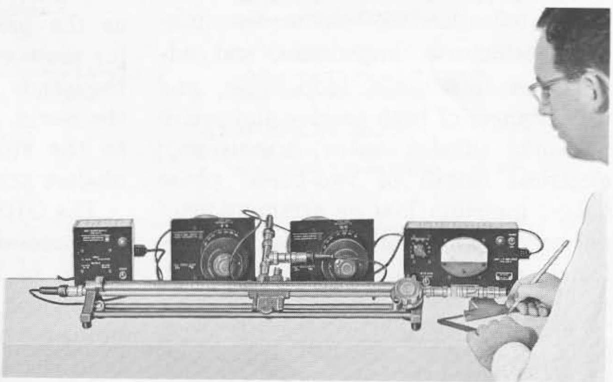
The VSWR of the line has been reduced. The VSWR specification (Figure 1) applies with either end of the line as the source end. A series of representative residual VSWR's of the line equipped with adaptors to other popular connectors is shown in Figures 2, 3, and 4.

The frequency range, formerly covering up to 5 GHz, now extends to 8.5 GHz. The low-frequency limit is not fixed at 300 MHz; it can be extended downward, typically to 150 MHz, by the use of air lines and the appropriate probe tuner.

Recommended Accessories

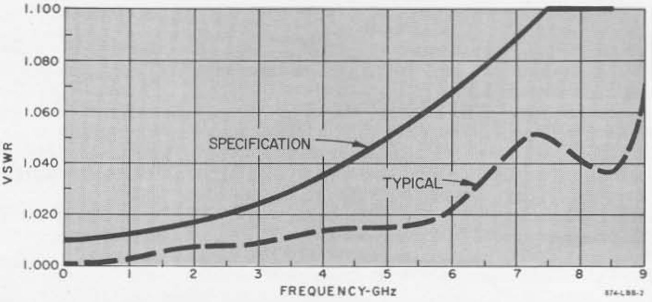
A convenient power source for slotted-line measurements is one of the GR Unit oscillators, which are available in several models and offer a wide choice of frequency range and power supply. The detector can be a conventional standing-wave indicator or one of the GR Type DNT Heterodyne Detectors.

Probe tuning can be accomplished with the Type 900-DP Probe Tuner or, if the convenience of external probe-depth adjustment is not needed, with a Type 874-D20L or -D50L Stub.



A complete measuring assembly, consisting of the improved slotted line, a Unit high-frequency oscillator with power supply, and a Type DNT Detector.

Figure 1. Type 874-LBB VSWR.



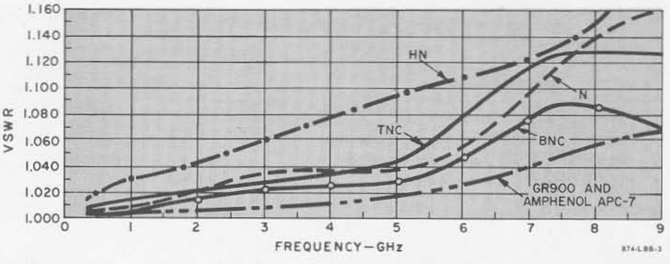
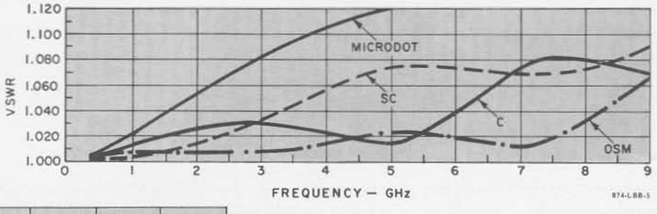


Figure 2.

Figure 3.



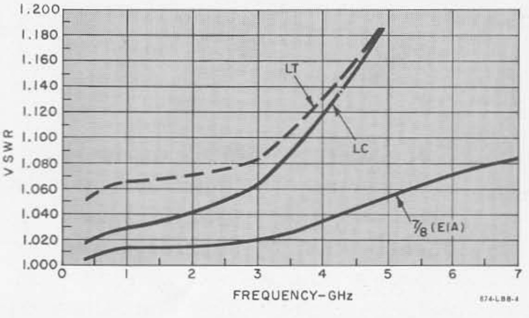


Figure 4.

Figures 2, 3 and 4. Representative residual VSWR of the Type 874-LBB equipped with adaptors to other coaxial connector types.

Other useful accessories: the Type 874-LV Micrometer Vernier, which allows a precise adjustment of probe-carriage position and is particularly useful in the width-of-minimum method of VSWR measurement and in electrical length measurements; the Type 874-ML Component Mount for measurements on lumped elements.

Last, but by no means least in importance, are the Type 874-Q Adaptors mentioned above*. With the Type 874-LBB Slotted Line and a complete set of these adaptors, the equivalent of 23 slotted lines can be obtained at a total price of $771.25.

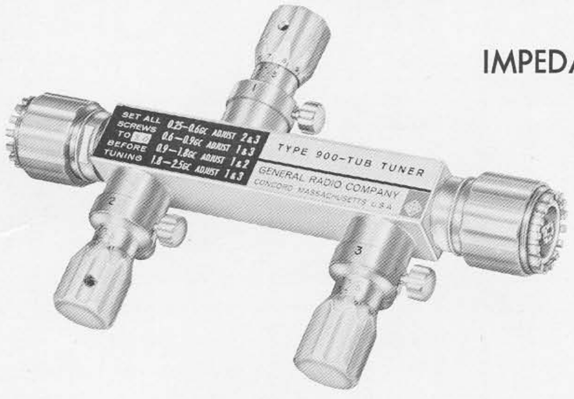
J. Zorzy

* See the GR catalog or ask for a complete listing.

SPECIFICATIONS

Characteristic Impedance: 50 Ω ±0.5%.
Probe Travel: 50 cm. Scale in centimeters; each division is 1 mm.
Scale Accuracy: ±(0.1 mm + 0.05%).
Frequency Range: 300 MHz to 8.5 GHz (usable to 9 GHz). At 300 MHz, the slotted line covers a half wavelength. Operation below 300 MHz is possible by use of lengths of Type 874 Air Lines.
Constancy of Probe Pickup: ±1.25%.
Residual VSWR: Less than 1.01 + 0.0016 f^2_{GHz} to 7.5 GHz; less than 1.10 from 7.5 to 8.5 GHz; see also Figure 1.
Accessories Supplied: Storage box, rf probe, and 2 microwave diodes.
Accessories Required: Type 900-DP Probe Tuner, recommended, or Adjustable Stub (Type 874-D20L) for tuning the crystal rectifier when audio-frequency detector or microammeter is used; suitable detector and generator; one each, Type 874-R22LA and Type 874-R22A Patch Cords, for generator and detector connections (patch cords are supplied with Type DNT Detector and GR Unit oscillators).
Dimensions: Width 26, height 4½, depth 3½ in (660, 115, 89 mm).
Net Weight: 8½ lb (3.9 kg).
Shipping Weight: 23 lb (10.5 kg).

Catalog Number	*Description*	*Price in USA*
0874-9651	**Type 874-LBB Slotted Line**	**$395.00**
0874-9652	**Type 874-LV Micrometer Vernier**	**41.00**
0874-9511	**Type 874-D2OL Adjustable Stub**	**20.00**
0900-9654	**Type 900-DP Probe Tuner**	**75.00**



IMPEDANCE-MATCHING TUNERS FOR PRECISION COAXIAL-MEASURING SYSTEMS

The Type 900-TUB Tuner, for the 0.25- to 2.5-GHz frequency range, complements the previously announced Type 900-TUA Tuner[1], which covers the 1- to 8.5-GHz frequency range. The two tuners are similar in design and construction and, in addition to their wide bandwidths, have the following desirable features:

1. A unique neutral position, from which rapid convergence to match can always be achieved

2. A fineness of control, so that vswr's as low as 1.0002 can be tuned out with ease

3. Stability — when a residual standing-wave ratio is tuned out, it stays tuned out; if the setting is changed, the original tuning is duplicated when the original setting is restored; if the connection between line and tuner is broken and then restored, the tuning is unchanged.

Each tuner has three tuning screws. In operation, two of the three screws are adjusted for match (which two depends on the frequency), while the unused screw is set to the neutral posi-

[1] "Coaxial Tuner with Neutral Setting," *General Radio Experimenter*, January 1965.

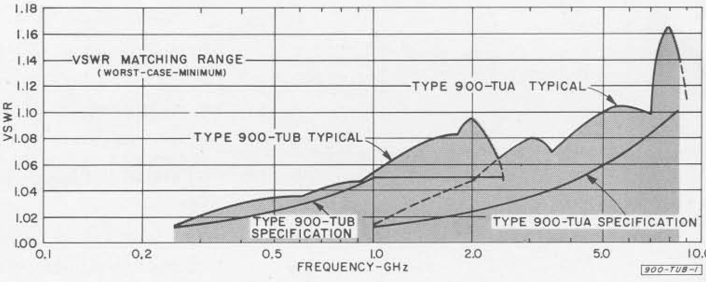


Figure 1. VSWR matching range of the Types 900-TUB and -TUA Tuners. Specifications and data shown are under the most restrictive phase conditions of the reflection to be matched out.

tion. Each screw has a scale, with vernier, and can be locked at any setting.

The VSWR matching ranges of the TYPES 900-TUB and -TUA Tuners (see Figure 1), while they are high enough for most applications, have been kept sufficiently low that extremely fine matches can be achieved with ease and speed.

APPLICATIONS

Matching to a Standard of Impedance

With the GR900 Tuners one can reduce the residual reflections introduced into a coaxial system by terminations, measuring instruments (such as slotted lines, rf bridges and directional couplers), adaptors between line sizes, and connectors. These residual reflections must always be considered with respect to some standard of impedance. The standard may be part of a measuring instrument, it may be a termination, or it may be a section of precision air line. The tuner is used, therefore, to match the impedance of the device in question to that of the standard.

The Measuring Instrument as an Impedance Standard

The TYPE 900-LB Precision Slotted Line is an excellent impedance standard. It covers the 0.3- to 8.5-GHz frequency range, and, at 2 GHz, for example, the residual impedance error (expressed in VSWR) is less than 1.003. If a composite termination consisting of a TYPE 900-W50 Standard Termination and a TYPE 900-TUA or -TUB Tuner is assembled, as shown in Figure 2, and the tuner is adjusted so that the amplitude of the standing-wave pattern observed on the slotted line is reduced to zero, then the residual VSWR of the composite termination is made equal to the residual VSWR of the slotted line. The improvement in VSWR can be as

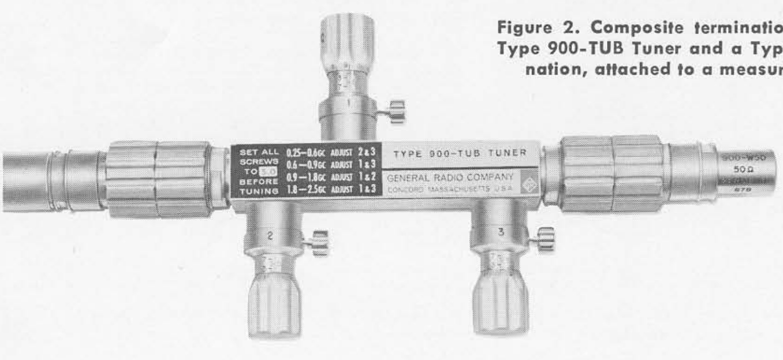


Figure 2. Composite termination consisting of a Type 900-TUB Tuner and a Type 900-W50 Termination, attached to a measuring instrument.

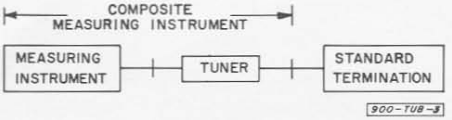


Figure. 3. Standard termination attached to a composite measuring instrument.

much as five-fold over the direct residual VSWR of the termination alone.

Termination as an Impedance Standard

A well-matched termination, such as the Type 900-W50 Standard Termination or, even better, the composite termination described above, can also be used as an impedance standard. For example, if the measuring instrument is a directional coupler or a hybrid junction, the instrument residual may be much greater than that of an available standard termination. In these cases a composite measuring instrument, consisting of the basic measuring instrument and a Type 900-TUA or -TUB Tuner, can be formed, as shown in Figure 3, and the tuner adjusted so that a null is observed with the measuring instrument. The residual VSWR of the composite instrument is thus made equal to that of the standard termination.

Air Line as an Impedance Standard

The most accurate impedance standard is the characteristic impedance of a section of precision air-dielectric coaxial line, such as a Type 900-LZ Reference Air Line. The residual VSWR of these air lines at 2 GHz is less than 1.0009. Both a composite measuring instrument and a composite termination, as shown in Figure 4, can be independently and simultaneously matched to the characteristic impedance of the air-line standard, at frequencies where the air-line length is an odd multiple of a quarter wavelength[2]. The matching is accomplished by alternate adjustment of the tuners, I and II, until no reflection is observed by the measuring instrument (1) when the composite termination is connected through the air-line standard to the composite measuring instrument and (2) when the composite termination is connected directly to the composite measuring instrument (that is, with the air line out of the system).

Simplifying Substitution Measurements

Substitution techniques, such as those described by Sanderson[3] and Zorzy[4], are used to obtain accurate measurements of small reflections in the presence of comparable residual reflec-

[2] MacKenzie, Thomas E., "Some Techniques and Their Limitations as Related to the Measurements of Small Reflections in Precision Coaxial Transmission Lines," presented at the 1966 Conference on Precision Electromagnetic Measurements, June 21–24, NBS, Boulder, Colorado. (Publication scheduled in the *IEEE Transactions on Instrumentation and Measurement*, December 1966.)

[3] Sanderson, A. E., "A New High-Precision Method for the Measurement of the VSWR of Coaxial Connectors," *IRE Transactions on Microwave Theory and Techniques*, Vol MTT-9, No 6, November, 1961, p 524–528.

[4] Zorzy, J., "Precise Impedance Measurements with Emphasis on Connector VSWR Measurements," 18th Annual ISA Conference and Exhibit, Chicago. Preprint No 47.4.63, 1963. (Available as General Radio Reprint No B20.)

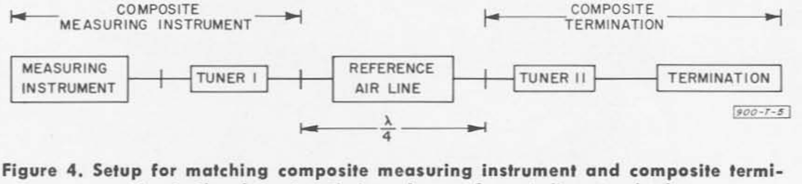


Figure 4. Setup for matching composite measuring instrument and composite termination to the characteristic impedance of an air-line standard.

tions in the measuring systems. In these techniques, two measurements are required, and the desired quantity is dependent on the vector difference of the two measured quantities.

When an impedance-matching tuner is used to make one of the measured reflection coefficients equal to zero, the second measurement alone provides the answer[2,5]. This means that it is not necessary for one to perform the vector subtraction or to plot measurements on a Smith Chart and to make tedious constructions. Also, if only the magnitude of the answer is required, it can be obtained directly from just one magnitude measurement.

Quarter-Wavelength Substitution to Measure Termination

The simplification resulting from the use of the tuner is illustrated by the example of the substitution technique that employs a quarter-wavelength reference air line to determine the reflection of a termination in the presence of the residual reflection of the measuring instrument.

Without the tuner, two measurements are required, one with and one without the reference air line in the system, as illustrated in Figure 5. (It is assumed that the reference air line is reflectionless and lossless and that the reflection coefficients of interest are small.) Thus

$$\Gamma_1 = \Gamma_m + \Gamma_t \qquad (1)$$

$$\Gamma_2 = \Gamma_m + \Gamma_t e^{-j\pi} = \Gamma_m - \Gamma_t \qquad (2)$$

where

Γ_1 is the measured reflection coefficient without the reference air line inserted,

Γ_2 is the measured reflection coefficient with the reference air line inserted,

Γ_t is the reflection coefficient of the termination, and

Γ_m is the residual reflection coefficient of the measuring instrument.

The termination reflection coefficient is given from the difference of equations (1) and (2) by the vector relation

$$\Gamma_t = \frac{\Gamma_1 - \Gamma_2}{2} \qquad (3)$$

The reflection coefficients Γ_1 and Γ_2 are plotted on the Smith Chart of Figure 6 with the construction required to obtain Γ_t.

Now, with the tuner forming a composite measuring instrument as shown

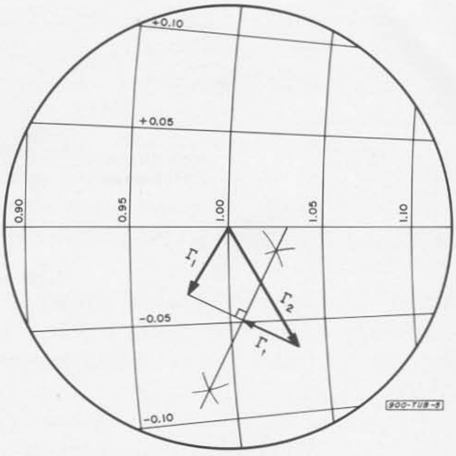


Figure 6. Smith Chart construction to determine Γ_t.

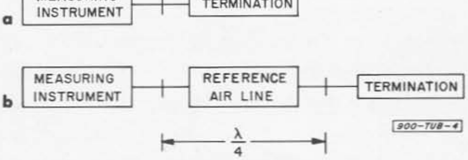


Figure 5. a) Setup to measure Γ_1. b) Setup to measure Γ_2.

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[2] *Ibid.*
[5] Sanderson, A. E., "Calibration Techniques for One- and Two-Port Devices Using Coaxial Reference Air Lines as Absolute Impedance Standards," *19th Annual ISA Conference and Exhibit*, New York, Reprint No. 21, 6-8-64. (Available as Reprint No B21 from General Radio Company, West Concord, Massachusetts.)

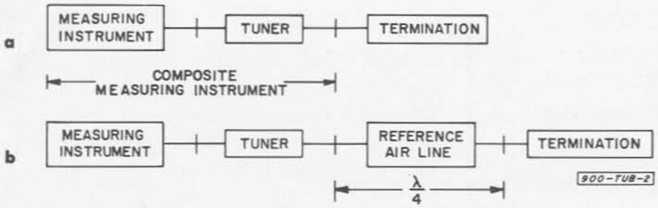


Figure 7. a) Setup to adjust Γ_1 to zero. b) Setup to measure Γ_2 and to determine Γ_t directly.

in Figure 7, Γ_1 can be tuned to zero.

$$\Gamma_1 = \Gamma'_m + \Gamma_t = 0 \qquad (4)$$

where Γ'_m is the residual reflection coefficient of the composite measuring instrument under the condition of (4). With

$$\Gamma_2 = \Gamma'_m - \Gamma_t, \qquad (5)$$

the termination reflection coefficient is given directly by

$$\Gamma_t = -\frac{\Gamma_2}{2}. \qquad (6)$$

No vector subtraction, and therefore no constructions on the Smith Chart, is required.

Similar simplifications can be realized in other substitution techniques through the application of the TYPES 900-TUA and -TUB Tuners.

— T. E. MACKENZIE

Note: A brief biography of Thomas E. MacKenzie, author of the foregoing article, appeared in the May 1966 issue of the *Experimenter*. — Editor

SPECIFICATIONS

	900-TUA	900-TUB
Frequency Range	1 to 8.5 GHz	0.25 to 2.5 GHz
Characteristic Impedance	50 Ω	50 Ω
VSWR Matching Range (worst-case minimum)*	$1.00 + 0.012 f_{GHz}$	$1.00 + 0.05 f_{GHz}$ to 1 GHz 1.05 from 1 to 2.5 GHz
VSWR Resettability	$<1.0005 + 0.0003 f_{GHz}$	$<1.0005 + 0.0003 f_{GHz}$
Residual VSWR (all controls at neutral)	<1.03 to 5 GHz <1.05 from 5 to 7 GHz	<1.03 to 1.5 GHz
Insertion Loss	<0.1 dB to 4 GHz <0.3 dB to 8.5 GHz	<0.1 dB
Repeatability of Connection	0.05%	0.05%
Electrical Length	12.0 cm	18.5 cm
Dimensions	4½ × 3½ × 1 in (115, 88, 25 mm)	6½ × 4¾ × 1 in (165, 120, 25 mm)
Net Weight	1 lb (0.5 kg)	1¼ lb (0.6 kg)
Shipping Weight	3 lb (1.4 kg)	4 lb (1.9 kg)
Catalog Number	0900-9635	0900-9637
Price	**$180.00**	**$265.00**

* Range is wider under most conditions

TE_{11}-MODE RESONANCES IN PRECISION COAXIAL CONNECTORS

It has been common practice to specify the upper frequency limit of a precision coaxial system as the calculated frequency at which the next higher mode above the TEM mode (TE_{11}) could propagate in the air-dielectric section of the line. It has, of course, been recognized that this mode can propagate in dielectric support beads below this frequency, but the interaction between the TEM and TE_{11} modes is insignificant if the effective electrical length of the bead is so short that a TE_{11}-mode resonance does not occur.

> Very sharp higher-order-mode resonances can exist in dielectric beads in coaxial lines. These resonances can be excited by asymmetry in the line but are often so small in magnitude that they are overlooked. Recently a resonance of this type was observed in Type 900-BT Connectors at 8.70 GHz, its presence was confirmed by a thorough experimental investigation, and its existence was explained by the analysis described in this article. Since the resonance causes an increase in both VSWR and insertion loss exceeding specifications, we have lowered the upper frequency specification for Type 900-BT Connectors to 8.5 GHz.

However, the effective electrical length of the bead for the TE_{11} mode is increased by the reactive loads presented to the TE_{11} mode at each end of the bead by the air-dielectric line sections, which are below cutoff for this mode, and the resonant frequency may therefore occur at a lower frequency than would be predicted by simple theory.

In a coaxial system, either a single bead or a pair of beads that are separated by a short section of air-filled line can resonate. The TE_{11}-mode resonance can then cause an increase in both VSWR and insertion loss; if the beads are part of connectors, these increases can vary in magnitude as the relative connector orientation is changed. The effect of the resonance is noticeable over only a narrow frequency band. The TE_{11} mode may be initiated by asymmetries such as slots or probes, as in a slotted line, or by eccentricities or irregularities that may occur in transitions between lines of different sizes.

Calculations of resonant frequency can be made on an impedance basis through the use of conventional transmission-line equations. Best agreement between calculated and measured data is obtained, in the particular case investigated, when the short undercut lengths,† such as 1-2 (see Figure 1), are treated as sections of air-dielectric line. The condition for resonance of a single bead (Figure 1a) is that $Z_3 = Z_2^*$, where Z_2^* is the conjugate of Z_2. The input impedance, Z_1, to the air-dielectric line is given by

$$Z_1 = Z_o \frac{Z_L + Z_o \tanh \gamma l}{Z_o + Z_L \tanh \gamma l} \qquad (1)$$

where

Z_o = characteristic impedance of air-dielectric line

Z_L = terminating impedance of air-dielectric line

† These undercuts compensate for the discontinuity introduced by the abrupt changes in the conductor diameters at the bead.

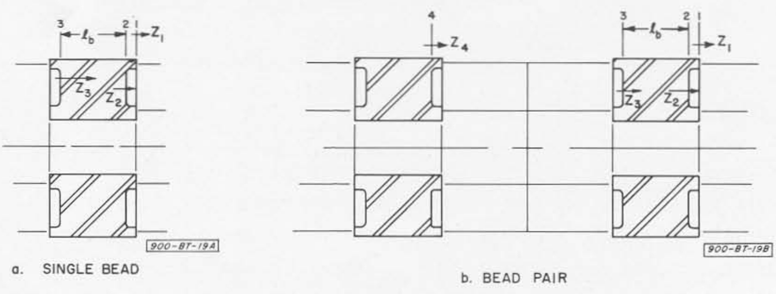


Figure. 1. Cross-sections of GR900 dielectric support beads.

γ = propagation constant of air-filled line

l = length of air-filled line.

For TE_{11} mode:

$$Z_o = j\eta \frac{f}{f_c} \frac{1}{\sqrt{1 - \left(\frac{f}{f_c}\right)^2}} \text{ for } f < f_c \quad (2)$$

where

f_c = cutoff frequency for TE_{11} mode in air-dielectric line

$\eta = 376.7$.

And

$$\gamma = \frac{2\pi}{\lambda_c} \sqrt{1 - \left(\frac{f}{f_c}\right)^2} \quad (3)$$

where λ_c = cutoff wavelength for the mode in the line.

In most practical cases, the air-dielectric section of line is long enough so that $0.99 < \tanh \gamma l \leq 1.00$. Then

$Z_1 \approx Z_o \frac{Z_L + Z_o}{Z_o + Z_L} = Z_o$. Thus the input impedance, Z_1, is an inductive reactance below the cutoff frequency of the air-dielectric line and is given by equation (2). The characteristic impedance and propagation constant in the bead are given by

$$Z_{o_b} = \frac{\eta_b}{\sqrt{1 - \left(\frac{f_c}{f}\right)^2}} \text{ for } f > f_c \quad (4)$$

and $\gamma_b = jk \sqrt{1 - \left(\frac{f_c}{f}\right)^2}$ (5)

where

$$k = \frac{2\pi \sqrt{\epsilon_r}}{\lambda}$$

$$\eta_b = \frac{376.7}{\sqrt{\epsilon_r}}$$

f_c = cutoff frequency for TE_{11} mode in dielectric bead[1,2]

ϵ_r = dielectric constant of bead.

If we treat region 1–2 as a section of air-dielectric line and let $Z_2 = Z_1$, then Z_3 can be calculated from

$$Z_3 = Z_{o_b} \frac{Z_2 + Z_{o_b} \tanh \gamma_b l_b}{Z_{o_b} + Z_2 \tanh \gamma_b l_b} \quad (6)$$

where l_b = length of bead.

One method of solving for f is to assume a value of f, solve for Z_o ($=Z_2$) in equation (2), Z_{o_b} in equation (4) and γ_o in equation (5), and then find Z_3 from equation (6). At resonant frequency, $Z_3 = Z_2^*$.

The resonant frequency of the bead pair (Figure 1b) may be calculated in a similar manner. Two resonant frequencies exist, one below and one above the resonant frequency of a similar single bead. Determine Z_1 from equations (1) and (2), Z_3 from equation (6), and Z_4 from equation (1), using Z_3 for Z_L and calculating γ from equation (3). The short undercut regions are again treated as part of the air line. At the resonant frequencies, $Z_4 = Z_3^*$. The two resonant frequencies can be found by successive trials.

Calculations on this basis were made for the GR900 connector. This connector uses a Teflon† support bead, in which the TE_{11}-mode cutoff frequency[1,2] is 7.03 GHz. The cutoff frequency in the adjacent air-dielectric line is 9.49 GHz. The bead is much shorter than a half wavelength (which might be expected to be the resonant length) for the TE_{11} mode, even at 9 GHz, but the loading effect of the reactive impedance of the air-dielectric lines for this mode causes the reso-

† DuPont trademark

1 J. Dimitrios, "Exact Cutoff Frequencies of Precision Coax," *MicroWaves*, June 1965.

2 N. Marcuvitz, *Waveguide Handbook*, MIT, Radiation Laboratory Series, Vol. 10, McGraw Hill, New York, 1951, p 77.

nance of a single bead at 8.86 GHz. Typically, the VSWR increases by 0.10 at this frequency, and the insertion loss by 0.35 dB. A pair of beads spaced as in a Type 900-BT Connector pair ex-

however, the mechanical stability and ruggedness would be impaired if the bead length were shortened by almost 1/3, the amount necessary to raise the resonant frequency of the connector

Figure 1.

NEW DECADE CAPACITOR – WIDE RANGE, HIGH RESOLUTION

The engineer who has occasion to use variable capacitors over a considerable frequency range will welcome the new TYPE 1412-BC Decade Capacitor. It has four polystyrene-dielectric decades, covering a range of 1.111 microfarads, as well as a continuously variable air capacitor, permitting very fine capacitance adjustment. In experimental circuits, it can be used not only at low frequencies where circuit capacitances are large but also, by virtue of its air capacitor and low internal inductance, at the lower radio frequencies.

The wide capacitance range and high resolution of this decade capacitance box make it useful in both laboratory and test shop. Owing to its fine adjustability of capacitance, it is a convenient variable capacitor to use with the Type 1605-A Impedance Comparator.

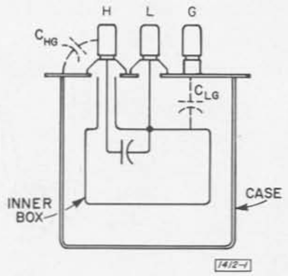


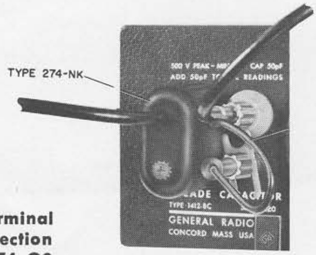
Figure 2. The double shielding used in the Type 1412-BC Decade Capacitor keeps C_{HG} very small. This capacitance is the difference between the three-terminal and two-terminal capacitance of the box; C_{LG} is approximately 125 pF.

Construction

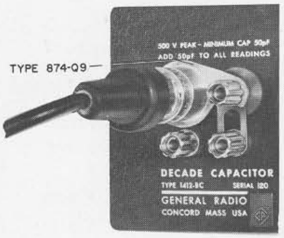
Ceramic-insulated switches, with solid-silver-alloy contacts, select parallel combinations of capacitors having values in the ratio of 1:2:2:5. The capacitors are of extended foil construction for minimum inductance and low series resistance. Polystyrene dielectric is used for stability of capacitance, low dielectric losses, and high insulation resistance.

The capacitors are housed in a double shield, consisting of an inner box and the outside case, so that the difference between the capacitances for two-terminal and three-terminal connections is very small — only one picofarad, the capacitance (C_{HG}) of the binding post H to the case. It is particularly desirable that this capacitance be as small as possible when the capacitor is used at low capacitance settings with the Type 1605-A Impedance Comparator.

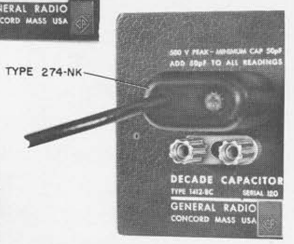
Figure 3.



***(Below)* Two-terminal (grounded) connection using a Type 874-Q9 Adaptor.**



***(Above)* Three-terminal connection with both shields connected to the case.**



***(Above)* Type 274-NK Shielded Double Plug used for two-terminal (floating ground) connection on Type 1412-BC. *(Left)* Two-terminal (grounded) connection using bus wire.**

Figure 2 shows the arrangement of the shields, and Figure 3 typical methods of connection to the panel terminals. Hardware is supplied for installing the assembly in a standard rack, and terminals are brought out at the rear for convenient connection (Figure 4).

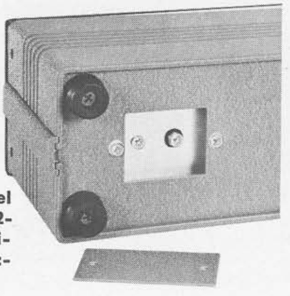


Figure 4. Rear panel view of Type 1412-BC, showing terminals for rear connections.

Readout

The four decades have clear, easy-to-read dials with numbered steps from 0 to X (X = 10). The air-capacitor dial has ten 10-pF divisions, plus additional readout to 1 pF per graduation. To read the dial, simply add the number of graduations (counting from 0) on the fixed vernier scale to the corresponding numbered division on the dial. For an increase in capacitance, the knob turns clockwise and the small graduations read clockwise. Sample settings are illustrated in Figure 5.

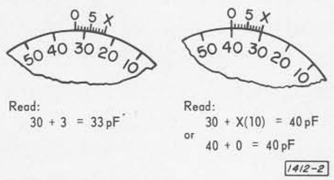


Figure 5.

Dissipation Factor

The dissipation factor of the polystyrene dielectric is quite low and relatively constant over the frequency range ordinarily encountered in most applications. At the extremes of the capacitance range, minor increases can be expected.

At the lower capacitance settings, the dissipation factor is increased by losses in the switch insulation and other materials outside the capacitors. The effect of these losses increases as the frequency is lowered.

At higher capacitance settings, the dissipation factor is increased by the series resistance of the wiring. This effect increases with frequency.

Capacitance

Capacitance changes with changes in frequency are principally a function

Robert W. Orr received his B.S. in E.E. from Texas A and M in 1928. After two years as a student engineer at General Electric Company, he held engineering and administrative posts in the field of capacitor engineering with RCA; Erie Resistor Corp; AMP, Inc.; and Aerovox Corp. He joined GR's Development Engineering Department in 1964. He is chairman of ASTM Committee D-9 on Electrical Insulating Materials, a member of Committee D-27 on Electrical Insulating Liquids and Gases, a member of the NRC Conference on Electrical Insulation, and a member of IEEE.

of the dielectric material below 1 kHz and a function of the amount of series inductance above 1 kHz. Polystyrene dielectric ensures negligible variations of capacitance below 1 kHz, and extended foil construction keeps the inductance of the capacitor itself low.

Most of the inductance is in the wiring. When the operating frequency (f) is well below the resonant frequency (f_r), the approximate increase in effective capacitance (ΔC) over the zero-frequency capacitance (C_o) is given by the expression:

$$\frac{\Delta C}{C_o} \approx \left(\frac{f}{f_r}\right)^2.$$

Typical values of the resonant frequency are:

Decade Capacitance	*Resonant Frequency*	
	Front Terminals	*Rear Terminals*
1.11115 μF	430 kHz	310 kHz
1.0 μF	440 kHz	320 kHz
0.1 μF	1.25 MHz	1.2 MHz
0.01 μF	3.5 MHz	4.3 MHz
1050 pF	10 MHz	17 MHz
150 pF	27 MHz	70 MHz

At frequencies up to 30 kHz, the effective capacitance at any setting will be less than 1% higher than the value of capacitance at 1 kHz. At most settings, the error will be much smaller.

— R. W. Orr

SPECIFICATIONS

Capacitance: 50 pF to 1.11115 μF in steps of 100 pF with a 0- to 100-pF variable air capacitor providing continuous adjustment with 1-pF divisions. Capacitance for 2- and 3-terminal connections differs by about 1 pF.
Dielectric: Polystyrene for decade steps.
Accuracy: ±(1.0% + 5 pF) at 1 kHz.
Temperature Coefficient: -140 ppm/°C (nominal).
Frequency Characteristics: $\frac{C_{dc}}{C_{1kHz}} < 1.001$. At higher frequencies the increase is approximately $\Delta C/C = (f/f_r)^2$. See table above for typical values of f_r.
Maximum Operating Temperature: 65°C.
Dielectric Absorption (Voltage Recovery): 0.1% maximum.
Dissipation Factor: 150 to 1000 pF, 0.001 max, at 1 kHz; over 1000 pF, 0.0002, max, at 1 kHz.
Insulation Resistance: 10^{12} ohms, minimum, at 500 V, dc.
Maximum Voltage: 500 V peak up to 35 kHz.
Terminals: Four Type 938 Binding Posts with grounding link are provided on the panel. Two of the binding posts are connected to the case and located for convenient use with patch cords in 3-terminal applications. Access is also provided to rear terminals for relay-rack applications.
Dimensions: Width 17 5/16, height 3½, depth 6 inches (440, 89, 155 mm), over-all.
Net Weight: 8½ lb (3.9 kg).
Shipping Weight: 10 lb (4.6 kg).

Catalog Number	*Description*	*Price in USA*
1412-9410	**Type 1412-BC Decade Capacitor**	**$190.00**

A 10-pF REFERENCE STANDARD CAPACITOR

To supplement the highly stable TYPE 1404-A (1000 pF) and TYPE 1404-B (100 pF) Reference Standard Capacitors[1], we now have available a 10-pF model, the TYPE 1404-C.

Standard capacitors in the 1404 series are highly stable units, hermetically sealed in an inert gas, and closely adjusted to their nominal values.

All critical parts of the plate assembly are made of Invar for stability and low temperature coefficient. After heat cycling and adjustment, the assembly is mounted in a heavy brass container, which, after evacuation, is filled with dry nitrogen under a pressure slightly above atmospheric and sealed. The container is mounted on an aluminum panel and protected by an outer aluminum case. Each capacitor is subjected to a series of temperature cycles to determine hysteresis and temperature coefficients and to stabilize the capacitance.

Two locking GR874 coaxial connectors are used as terminals. The outer shell of one is connected to the case, but the outer shell of the other is left unconnected to permit the capacitor to be used with an external resistor as a dissipation-factor standard.

[1] John F. Hersh, "A Highly Stable Reference Standard Capacitor," *General Radio Experimenter*, August 1963.

SPECIFICATIONS

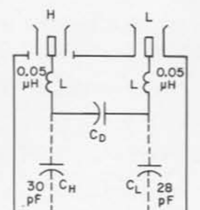
Calibration: A certificate of calibration is supplied with each capacitor, giving the measured direct capacitance at 1 kHz and at 23° ± 1°C. The measured value is obtained by a comparison to a precision better than ±1 ppm with working standards whose absolute values are known to an accuracy of ±20 ppm, determined and maintained in terms of reference standards periodically measured by the National Bureau of Standards.

Adjustment Accuracy: The capacitance is adjusted before calibration with an accuracy of ±5 ppm to a capacitance about 5 ppm above the nominal value relative to the capacitance unit maintained by the General Radio reference standards.

Stability: Long-term drift is less than 20 parts per million per year. Maximum change with orientation is 10 ppm and is completely reversible.

Temperature Coefficient of Capacitance: 2 ± 2 ppm/°C for TYPES 1404-A and -B, 5 ± 2 ppm/°C for TYPE 1404-C, from −20°C to +65°C. A measured value with an accuracy of ± 1 ppm/°C is given on the certificate.

Equivalent circuit showing direct capacitance, C_d, and average values of residual inductance, *L*, and terminal capacitances, C_a and C_b. C_d = 1000 pF for Type 1404-A, 100 pF for Type 1404-B, and 10 pF for Type 1404-C.



Temperature Cycling: For temperature cycling over range from −20°C to +65°C, hysteresis (retraceable) is less than 20 ppm at 23°C.

Dissipation Factor: Less than 10^{-5} at 1 kHz.

Residual Impedances: See equivalent circuit for typical values of internal series inductances and terminal capacitances.

Maximum Voltage: 750 V.
Terminals: Two locking GR874 coaxial connectors; easily convertible to other types of connectors by attachment of locking adaptors. Outer shell of one connector is ungrounded to permit capacitor to be used with external resistor as a dissipation-factor standard.

Accessories Required: For connection to TYPE 1615-A Capacitance Bridge, 2 TYPE 874-R20A or TYPE 874-R22LA Patch Cords.
Dimensions: Width 6¾, height 6⅝, depth 8 in (175, 170, 205 mm), over-all.
Net Weight: 8½ lb (3.9 kg).
Shipping Weight: 14 lb (6.5 kg).

Catalog Number	*Description*	*Price in USA*
1404-9701	**Type 1404-A Reference Standard Capacitor,** 1000 pF	**$225.00**
1404-9702	**Type 1404-B Reference Standard Capacitor,** 100 pF	**225.00**
1404-9703	**Type 1404-C Reference Standard Capacitor,** 10 pF	**225.00**

U.S. Patent Number 2,548,457.

INCREASED FREQUENCY RESOLUTION FOR WAVE-ANALYZER RECORDINGS OF VIBRATION, ACOUSTIC, AND ELECTRICAL SIGNALS

The TYPE 1900-A Wave Analyzer[1] is widely used for low-frequency spectrum analysis, because it has three bandwidths, 3, 10, and 50 Hz, and an 80-dB dynamic range for recording. The 3-Hz bandwidth is particularly popular, because of its excellent resolution. In order to take full advantage of that resolution, we are now making available a link unit that will permit recording with an expanded frequency scale on the TYPE 1521-B Graphic Level Recorder. This TYPE 1900-P3 Link Unit is shown in Figure 1 installed on the wave analyzer and recorder.

With this new link unit the frequency scale of a recording is spread out to 2 inches for 100 Hz, so that a frequency difference of 1 Hz can be noticed. An additional frequency scale of 2 inches for 1000 Hz, identical with that of the TYPE 1900-P1 Link Unit, is also provided. A neutral position simplifies the setting of the frequency control to the desired starting point.

Figure 1. Type 1900-P3 Link Unit installed on the wave analyzer and graphic level recorder.

The same chart paper, TYPE 1521-9464, is used for both speeds.

The recording reproduced in Figure 2 shows evidence of the smoothness of the frequency-control drive with this link unit. The components displayed are spaced 10 Hz apart in the vicinity of 50,000 Hz, and the absence of significant jitter in this expanded display demon-

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

strates that the new drive can be used to advantage over the full frequency range of the wave analyzer. Of course, the applied signal must also be sufficiently stable that the 3-Hz bandwidth can be used. For the pulse signal analyzed in Figure 2, the pulse repetition frequency of a Type 1217-C Unit Pulse Generator was controlled at 10 Hz by a highly stable, crystal-controlled, time-mark generator so that the harmonics up to and beyond the 5000th would not show any appreciable jitter.

The resolution of the 3-Hz bandwidth and the convenient display of the analyzed spectrum on the recorder make the system well suited to the analysis of certain types of electrical, acoustic, and vibration signals, including for example, the acoustic noise and vibration produced by such rotating machinery as gear trains, electrical motors, and turbines.

Arnold P. G. Peterson received his B. Eng degree from the University of Toledo in 1934 and his S.M. and Sc.D. degrees from M.I.T. in 1937 and 1941, respectively. He was a research assistant at M.I.T. from 1936 to 1940. He came to General Radio in 1940 as a Development Engineer and became Group Leader of the Audio Group in 1947. He is a Fellow of IEEE and of the Acoustical Society of America, of which he was Vice President in 1958–59. He is a member of AAAS, AAPT, and AGU, and of several standards committees in the general field of acoustics.

Figures 3, 4, and 5 illustrate the detail that can now be obtained in a practical case. These recordings are analyses of the vibration of the paper-drive frame of the Type 1520-A Sampling Recorder. A scan of the frequency range of the analyzer in the normal drive position shows (Figure 3)

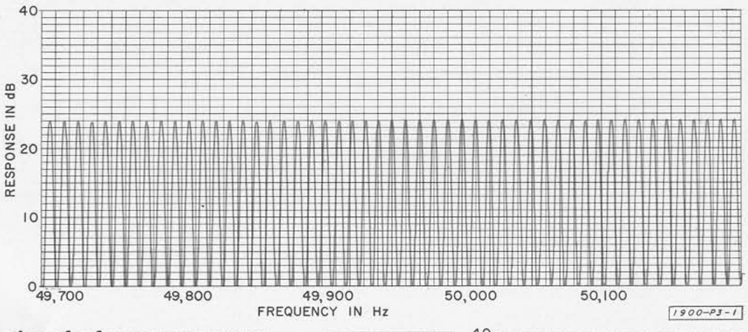


Figure 2. Section of a frequency spectrum in the vicinity of 50,000 Hz of a 30-μs pulse repeated every 0.1 s. The expanded frequency scale is used to show the individual components every 10 Hz.

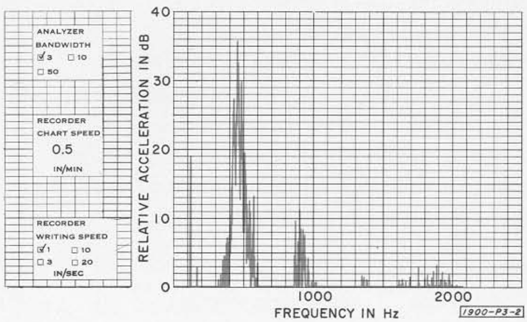


Figure 3. Analysis of the vibration of a frame holding a gear-belt drive. The frequency scale is sufficiently compressed to show the general character of the spectrum.

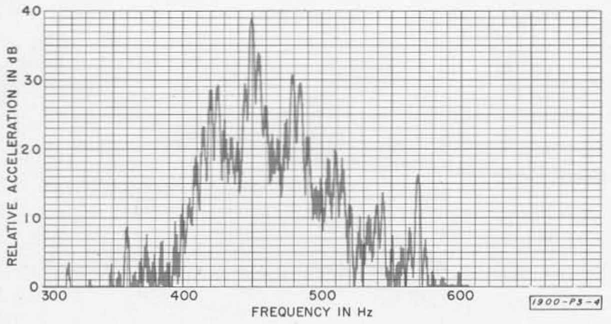


Figure 4. Analysis of the same vibration signal of Figure 3, but with the expanded frequency scale.

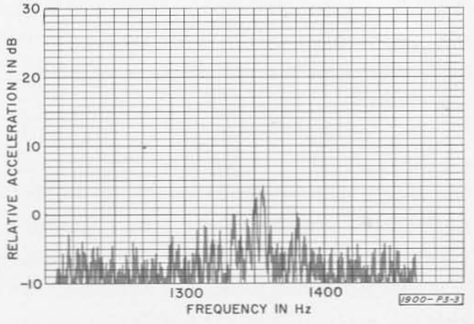


Figure 5. The expanded frequency scale is used here to show the components in the vicinity of 1350 Hz.

that the significant components of vibration are at 120 Hz, and in the vicinity of 450 Hz, 900 Hz, and 1350 Hz.

A detailed recording in the vicinity of 450 Hz with the expanded scale (Figure 4) shows that the belt-gear-drive tooth-contact rate of 450 impacts per second determines the frequency of the dominant component, and the gear belt with its speed of 5 r/s introduces a host of components spaced about the main component by multiples of 5 Hz. The torque pulsations from the 1800-rpm synchronous motor and the 120-Hz magnetically driven vibration in the motor also influence the spectrum.

The large number of components is a result of complex interactions of the various impacts and forces.[2] The 3-Hz bandwidth and the expanded scale make it possible to display these many components and to determine their actual frequencies.

A significant amount of random motion in the mechanism being measured obscures some of the weaker components. This effect is even more important at higher frequencies. As shown in Figure 5, a few components in the immediate vicinity of 1350 Hz are displayed clearly, but the existence of many others is obviously probable by reason of the spacing of fluctuating peaks at 5-Hz intervals.

This new accessory makes the recording wave analyzer an even more versatile tool than before for the analysis of stable, complex signals.

— Arnold Peterson

[2] L. S. Wirt, "An Amplitude Modulation Theory for Gear-Induced Vibrations," Chapter 17 of *Measurement Engineering* by P. K. Stein, Tempe, Arizona, 1962.

Catalog Number	*Description*	*Net Wt*	*Ship Wt*	*Price in USA*
1900-9603	**Type 1900-P3 Link Unit**	1 lb (0.5 kg)	4 lb (1.9 kg)	**$55.00**

LOCATING SUBMARINE CABLES

Crossing San Francisco Bay are a great many submarine cables, which the Pacific Telephone and Telegraph Company uses for telephone services. An important factor in the maintenance of these cables is the ability to locate them accurately, not only in case of cable failure but also when contractors or government agencies are working in the area.

Some years ago, the Bay Area Chief Engineer's Department of Pacific Telephone developed a magnetic-induction system to locate submarine cables. A high-current oscillator sends an audio-frequency tone into one end of the cable with ground return. Receiving equipment is installed in a motor launch. The field produced by the cable current induces a voltage in a coil, which is amplified to operate a meter and a loudspeaker.

Recently a new receiver and transmitter were assembled to replace older and less reliable equipment. It was desirable to increase the range, which formerly was about 50 feet, and to package the receiver into one small transistorized unit. The new system is shown in Figures 1 and 2.

There are several problems associated with putting a tone on working telephone cables at the necessary high level and then in detecting it from a distance:

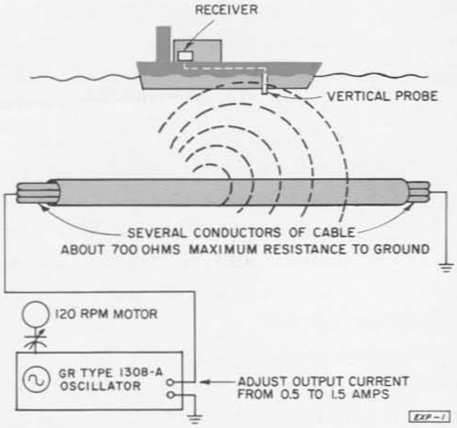


Figure 1. A signal is applied to the cable as shown, generating a magnetic field, which is detected by a probe and receiving equipment in the launch.

1. Crosstalk enters working pairs of the cable.
2. Harmonics interfere with carrier systems in the cable.
3. The receiving equipment picks up 60-Hz fields.
4. Harmonics of 60 Hz interfere with the received signal.
5. The sensitive receiver picks up noise from working circuits in cables.
6. Cable capacity shunts the transmitted tone to ground.

Tests have showed that a frequency

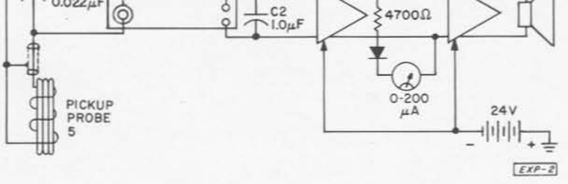


Figure 2. Elementary schematic of the receiving equipment.

of 150 Hz produces the best results. If the transmitted power is reasonably free of distortion and the receiver sharply tuned to 150 Hz, the above problems are, for all practical purposes, overcome.

The 150-Hz transmitter must supply a considerable amount of power in order to produce a magnetic field strong enough to be detected at some distance. The General Radio Type 1308-A Audio Oscillator and Power Amplifier proved to be ideal for this application. It is capable of 200 watts output and has meters to monitor the output voltage and current.

Figure 1 shows how the signal is applied to the cable. Several conductors of the cable are connected together at both ends, and the conductors are grounded at the far end. Maximum resistance is about 700 ohms. The 150-Hz power is applied between the multipled conductors and ground, and the output current is adjusted to be between 0.5 to 1.5 amperes.

To ensure that the received signal is the one of interest and does not come from a power cable or some other

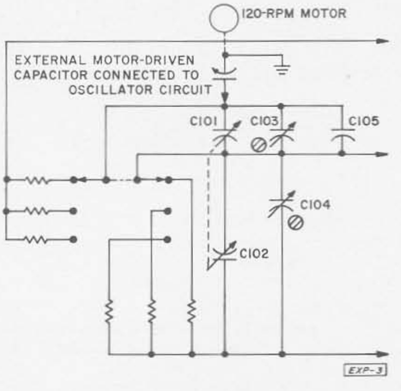


Figure 3. To produce a warble tone, a motor-driven capacitor is connected to the oscillator-tuned circuit.

NOTE

This interesting description of the use of electronic instruments to solve an important communications-system problem is published through the courtesy of the Pacific Telephone and Telegraph Company. Those interested in further details of the apparatus are referred to an article by L. W. Gunn and H. E. Bomar, entitled "Submarine Cable Detection," published in the December 1965 issue of *Electrical Construction and Maintenance.*

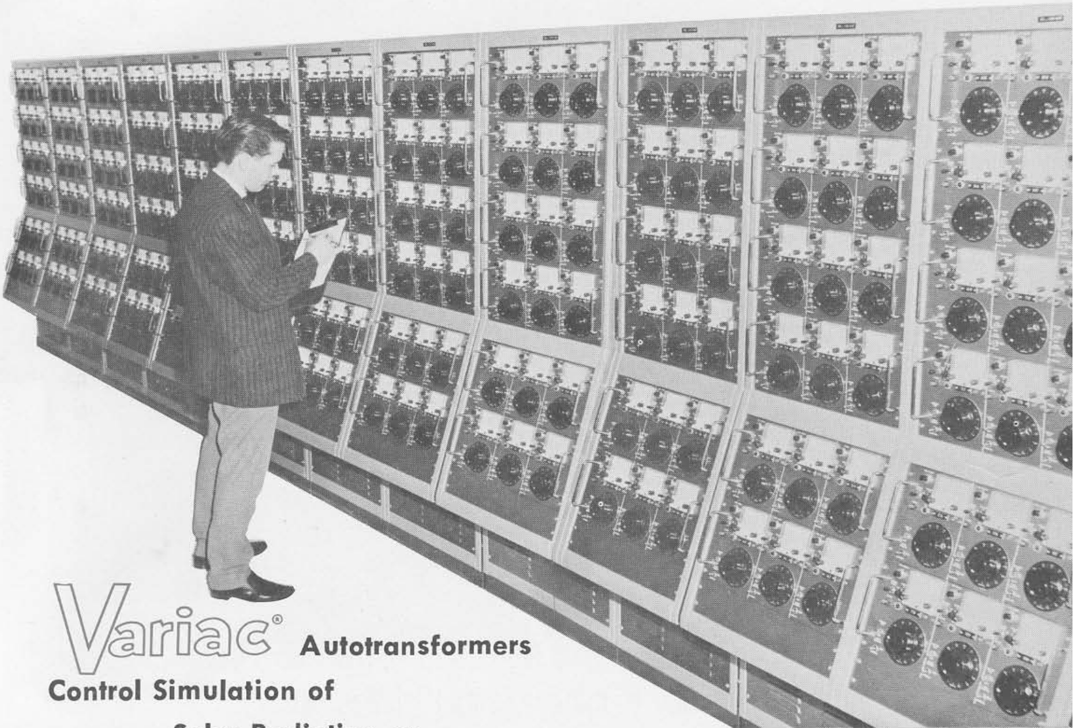
source, a distinctive tone is produced. A small variable capacitor connected to the oscillator circuit is driven by a 120-rpm clock-type synchronous motor to shift the frequency ±2 Hz (see Figure 3).

The detecting probe consists of a 5-henry winding on a laminated iron core, tuned to 150 Hz with a shunt 0.22-μF capacitor. It is mounted vertically near the bow and below the deck of the launch. The probe connects to the input of a General Radio Type 1232-A Tuned Amplifier and Null Detector, tuned to 150 Hz. The output drives a transistorized power amplifier and speaker (see Figure 2).

When driven at 150 Hz, a trumpet-type speaker was found to produce a marked third-harmonic output, which is easy to hear. As the launch approaches the cable, the operator hears the tone on the speaker and sees the relative level on the meter. The tone becomes louder until the probe is directly over the cable, at which time a null is produced. The meter is used to locate the exact spot. The tone is heard again as the probe moves past the cable.

To locate a cable break, the tone is followed until it disappears. The sea water grounds the severed end of the cable to complete the circuit.

With 0.5 ampere of transmitter current and the cable in 80 to 100 feet of water, the receiver gives an excellent indication about 100 feet away from the cable. If additional conductors of the cable are available, they can be multipled to reduce the resistance. To increase the range, the current can be increased and still not exceed the 200 watts output. During tests with 1.5 amperes of transmitter current, the launch received a signal, somewhat weak, but usable, in over 200 feet of water.



Variac® Autotransformers Control Simulation of Solar Radiation on Lunar Excursion Module

Photo courtesy of Grumman Aircraft Engineering Corporation, Bethpage, L. I., New York.

Shown here are 12 of 14 bays in the thermal-test console designed by Grumman Aircraft Engineering Corporation to check out a Lunar Excursion Module mock-up under simulated thermal conditions of outer space. Each of the 252 motor-driven Variac® autotransformers accurately controls the voltage applied to a resistive heating element affixed to the skin of the LEM mock-up. A single pushbutton energizes the motors of all 252 Variac transformers, and their output voltages all rise simultaneously from zero to preset voltages established by the needle positions of ammeters in conjunction with optical meter relays. If further fine adjustment of any particular Variac is required, it is done manually with the front-panel knob.

Variac autotransformers were selected for this installation for several important reasons. First, their proven reliability, simple design, and 1000%-overload capability guarantee dependable operation and minimum maintenance. Second, a Variac does not destroy waveform purity by chopping or reshaping; if the input voltage is a sine wave, the output voltage is a sine wave. Consequently, no RFI is produced. Third, motor-driven Variac autotransformers can be programmed, which eliminates the need for manual adjustment of the initial voltage applied to each heating element.

the GR Experimenter

GENERAL RADIO COMPANY
WEST CONCORD, MASSACH

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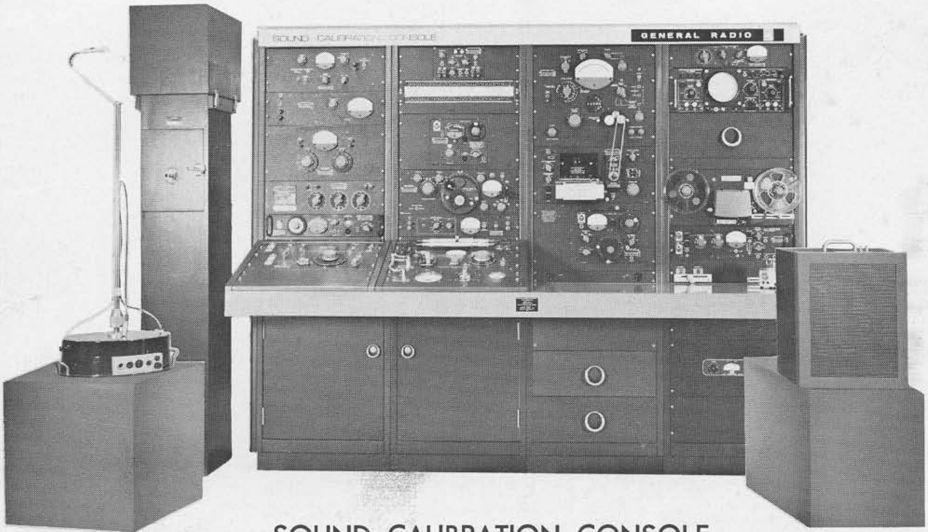

SOUND CALIBRATION CONSOLE

This Sound Calibration Console was supplied to the Calibration and Metrology Division, Newark Air Force Station, Newark, Ohio, for use as a laboratory standard of acoustical calibrations for the U. S. Air Force. Among the measurement capabilities of the console are the following:

Microphone Calibration—*Reciprocity Method*

Microphone Calibration—*Comparison Method*

Directional Calibration of Microphones

Frequency Response of Microphones

Frequency Analysis—*Narrow Band*

Frequency Analysis—*⅓-Octave Band*

Frequency Analysis—*Octave Band*

Characteristics of Anechoic Rooms and Chambers

Reverberation Measurements—*Bands of Noise*

Reverberation Measurements—*Warble Tones*

Frequency Response of Amplifiers

Frequency Response of Tape Recorders

Tape Recording of Signals

Measurement and Analysis of Tape-Recorded Signals

▶ ***Inquiries are invited for acoustic-measurement systems.***

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THE GENERAL RADIO

Experimenter

New 70-MHz Solid-State Frequency Synthesizer

Narrow-Band Wave Analyzer

1% Bandwidth

VOLUME 40 · NUMBER 9 / SEPTEMBER 1966

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September 1966

the GR Experimenter

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THIS ISSUE

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70-MHz SYNTHESIZER JOINS THE FAMILY

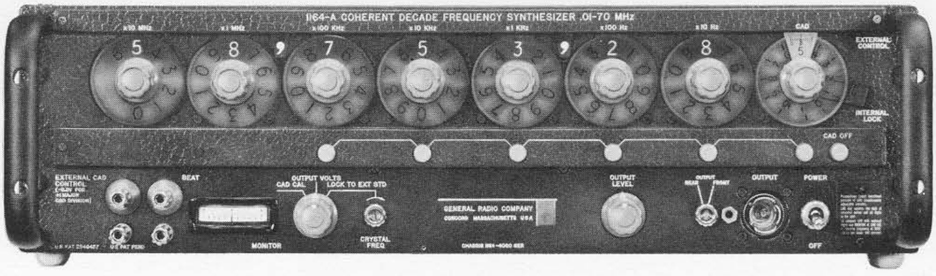


Figure 1. Type 1164-A7C.

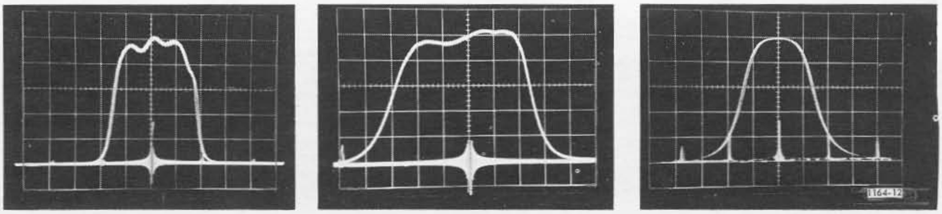
The newest addition to the GR synthesizer line, the TYPE 1164-A Coherent Decade Frequency Synthesizer (Figure 1), offers greatly expanded frequency coverage, extending well beyond the normal high-frequency, radio-communications bands and including the popular intermediate frequencies. All the features of the low-frequency models[1,2] — modular construction, in-line readout, provision for sweeping, continuously adjustable decade (CAD), programmability,[3] ac and battery operation — have been retained, and a few new ones added.

The frequency synthesizer, a relatively recent development, is nonetheless well established as an important instrument in the laboratory and in countless frequency-control applications. Also well established is GR's 1160 family of coherent decade frequency synthesizers, which have until now supplied frequencies up to 12 MHz. Now a new member—the 1164—extends coverage to 70 MHz and thus brings the considerable advantages of this type of instrument to many more users.

To review the chief characteristics of the 1160 line briefly: Each synthesizer can contain up to seven plug-in digit modules, each controlled by a front-panel rotary switch, plus a continuously adjustable decade (CAD). By the push of a button, the CAD can be electrically substituted for one or more of the step-digit modules. The output level is adjustable up to 2 volts and is monitored by a panel meter. The synthesizer can be locked to an external frequency standard, can be swept, and, with appropriate plug-in modules, can be programmed.

Because of the modular construction, the 1160 series can be provided in a great many variations to suit customer requirements. One can, for instance, buy a synthesizer with as few as three of the maximum seven digits, adding more later as more resolution is needed. The programmable option further increases the number of available models. For the

[1] A. Noyes, Jr., "Coherent Decade Frequency Synthesizers," *General Radio Experimenter*, September 1964.
[2] A. Noyes, Jr., "12-Mc Coherent Decade Frequency Synthesizer," *General Radio Experimenter*, November-December 1965.
[3] G. H. Lohrer, "Remote Programming for GR Synthesizers," *General Radio Experimenter*, May 1965.



Communications receiver i-f with 455-kHz center frequency and 1.5-kHz mechanical filter. Horizontal scale: 500 Hz/cm. Marker frequency: 455.00 kHz. The CAD was in the 1-kHz position and was swept $\pm 2\frac{1}{2}$ major divisions, or $\pm 2\frac{1}{2}$ kHz.

Response of 50-MHz crystal filter. Horizontal scale: 2 kHz/cm. Marker frequency: 50.00000 MHz. The CAD was in the 10-kHz position and was swept ± 1 major division, or ± 10 kHz.

Response of GR Type 1900 Wave Analyzer tuned to 10 kHz and set for 10-Hz bandwidth. Horizontal scale: 5 Hz/cm. Marker frequency: 10,000.00 Hz. The CAD was in the 10-Hz position and was swept $\pm 2\frac{1}{2}$ major divisions, or ± 25 Hz.

Figure 2.
Oscillograms showing frequency response of various filters to swept output of synthesizer CAD.

1164 alone, twenty "standard" combinations are listed.

Applications

One of the most common uses for synthesizers is frequency control of communications transmitters and receivers. Used with a transmitter, a complete 70-MHz synthesizer is roughly equivalent to a bank of 7 million crystals that can easily be switched to change frequency. (There are 700 million readable frequencies when the CAD is used.) In a receiver, the synthesizer can serve as a highly stable local oscillator.

The 70-MHz upper frequency limit of the new synthesizer covers not only most of the commercial broadcasting spectrum but also intermediate frequencies of many vhf and uhf systems. Frequency-translation techniques can be applied, moreover, to provide useful frequencies for devices operating in the gigahertz region.

In a telemetry receiver, for example, a synthesizer can be used as a second- or third-conversion oscillator, automatically compensated for Doppler shift. The Doppler shift is sensed by an external phase detector, which in turn is used to control the synthesizer's CAD frequency. Or, in a nuclear magnetic resonance (NMR) study, the synthesizer frequency can be added to the output of a stable microwave source to produce a 70-MHz-wide microwave band under synthesizer control. Conversely, the NMR band can be beat against a stable microwave frequency and the beat frequency compared with the synthesizer output.

Of course, the synthesizer output itself can be multiplied, but, the higher the multiplier, the more noticeable any instability.

Sometimes it is desirable to divide the synthesizer frequency. In one application, for instance, an aerospace laboratory wished to measure the rotational speed of a satellite to eight significant digits, then to supply a corresponding frequency to the satellite. With the satellite rotating only 100 revolutions a minute, it was necessary to divide the synthesizer output by means of a 10^4 scaler; the synthesizer then served admirably to measure and to reproduce the satellite rotational frequency.

The sweep capability is all-important in many applications and is especially so in the measurement of the characteristics of crystal and other mechanical filters (see Figure 2). The CAD can be swept electrically over ± 5 major dial divisions, corresponding to a frequency span as wide as 1 MHz or as narrow as 10 Hz. The sweep can be centered on any point of the 12-division manual dial except on the 100-kHz functional position, where the limits for the frequency excursion are − 100 and + 1100 kHz.

Perhaps the single most important capability of a frequency synthesizer is programmability. In the TYPE 1164 synthesizer, output level as well as frequency can be programmed. This means, for instance, that in a frequency-response measurement, frequency and level can be simultaneously programmed to effect 3-dB, 6-dB, 10-dB, etc, level changes at selected frequencies.

The maximum programmable bandwidth of the TYPE 1164 synthesizer is at present 1 MHz. In other words, one can electrically control output frequency over any 1-MHz (or smaller) range, up to the 70-MHz frequency limit of the synthesizer. Frequency switching time is less than two milliseconds.

Constructional Features

The synthesizer consists of plug-in modules in a bench or rack frame with a panel only 5¼ inches high. Five of the seven digit modules in the TYPE 1164 are identical and interchangeable with one another and with corresponding modules in other GR synthesizers. The advantages of such modular construction in servicing are obvious. Less obvious, perhaps, but equally important are the manufacturing economies of this approach, which are translated into low prices.

All etched boards in the instrument are made of fiberglass, and the improved power supply uses all silicon transistors. The rear panel is an engineer's delight: in addition to the primary output (also available at the front panel), the following are available: 100 kHz, 1 MHz, 5 MHz, 5-5.1 MHz, 30 MHz, 42 MHz, 40-49 MHz, 50-51 MHz, 90 MHz, and +18V dc.

HOW IT WORKS

Figure 3 is a complete block diagram of the synthesizer.

The DI-1, CAD, and AFS-1 modules have been described in an earlier article.[1] The new modules are briefly described below.

DI-2, DI-3 Digit-Insertion Units

Figure 4 is a block diagram of the DI-2 unit. The frequency of the digit oscillator goes from 40 to 49 MHz in 1-MHz steps as its dial rotates from 0 to 9. The digit-oscillator frequency is rough-tuned to the proper frequency by a step switch controlled by the dial, and a voltage-control servo establishes phase lock with a multiple of the reference frequency. A gated or sampling-type phase detector is used here. The gate opens only during every 40th to 49th cycle (depending on the frequency selected) of the output frequency of the DI-2 unit (every 30th to 24th cycle of the DI-3) and stays open for only a small fraction of a period of the output frequency. If the phase of the output frequency is crossing zero while the gate is open, no voltage is placed on the holding capacitor, and therefore no correction voltage is applied to the digit oscillator. If the phase of the output frequency has passed zero in a negative direction when the gate opens, a negative voltage is placed on the holding capacitor; if the phase has not reached zero, a positive voltage is placed on the capacitor. With the proper phasing in the frequency-control loop of the oscillator, the phase error between the reference-frequency-controlled gate and the output frequency is minimized. The latter is thus locked at an exact multiple of the reference frequency.

[1] *Ibid.*

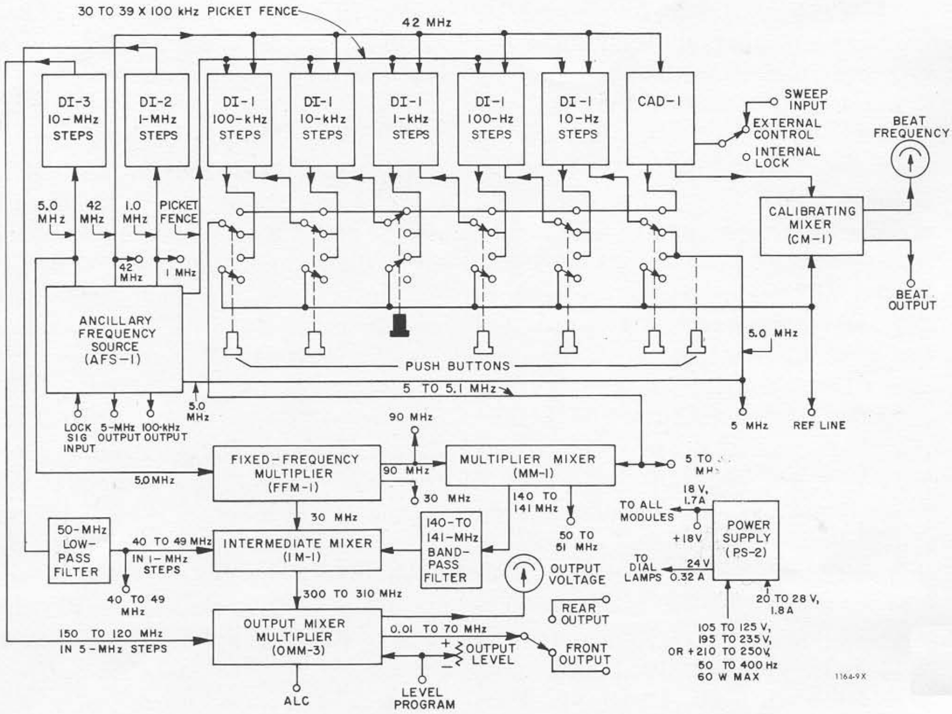


Figure 3. Block diagram of the synthesizer.

Any change in the output frequency is immediately sensed as a phase error and corrected.

Proper phase condition for phase lock exists whenever the digit-oscillator frequency is an exact multiple of the reference frequency. Rough-tuning to the approximate frequency by the digit dial determines which multiple is chosen. A dial-light warning system indicates failure to achieve stable phase lock. As in the DI-1's, the presence of any appreciable ac signal in the phase-control loop is sensed, causing the dial light of the affected unit to go out.

The DI-3 unit is identical to the DI-2 except that the frequency from the ancillary frequency source is 5 MHz, and the step-tuned oscillator goes from 150 MHz to 120 MHz in 5-MHz steps as its dial setting is changed from 0 to 6.

Fixed-Frequency Multiplier (FFM-1)

The fixed-frequency multiplier is shown in Figure 5. Undesired frequencies are rejected by frequency-selective amplifiers and by the connection of pairs of diodes in push-push for doubling and push-pull for tripling. Levels are kept relatively high to minimize noise.

Mixers MM-1 and IM-1

Figures 6 and 7 are block diagrams of the multiplier mixer (MM-1) and intermediate mixer (IM-1), respectively. In each, bandpass amplifiers reject frequencies outside the desired ranges. Each uses transistor frequency multipliers and double-diode balanced mixers (one in the MM-1, two in the IM-1).

Output Multiplier Mixer OMM-3

Figure 8 shows the output multiplier mixer in block form. The 150- to 120-MHz output of the DI-3 unit is amplified in a two-stage bandpass amplifier, doubled in frequency by a pair of diodes in a full-wave doubler, and filtered by a six-pole bandpass filter to provide the input to the final mixer, which produces the 10-MHz steps in the output. A four-diode double-balanced mixer is used to subtract this frequency from the 300- to 310-MHz output from the IM-1 unit to produce the final output frequency.

Signal levels at this mixer are kept low to minimize undesired mixing products within the output-frequency passband. A six-stage broad-

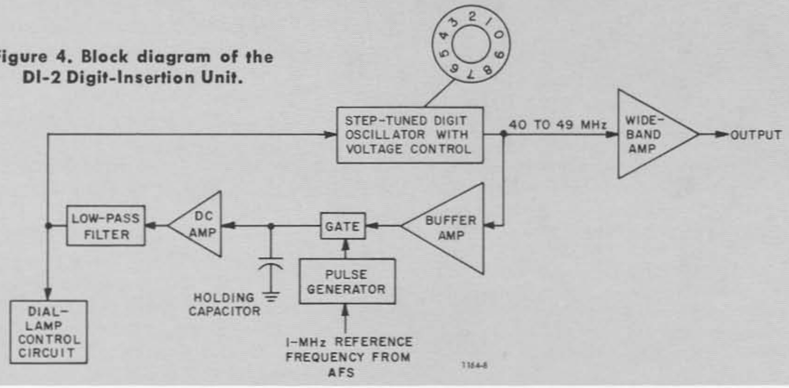


Figure 4. Block diagram of the DI-2 Digit-Insertion Unit.

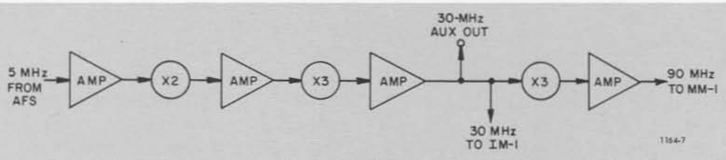


Figure 5. Block diagram of the Fixed Frequency Multiplier (FFM-1).

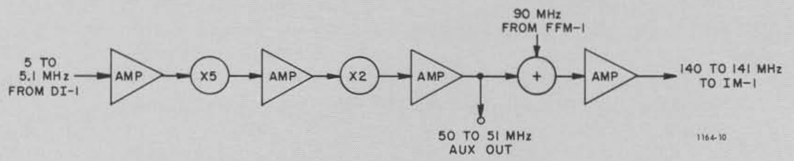


Figure 6. Block diagram of the Multiplier Mixer (MM-1).

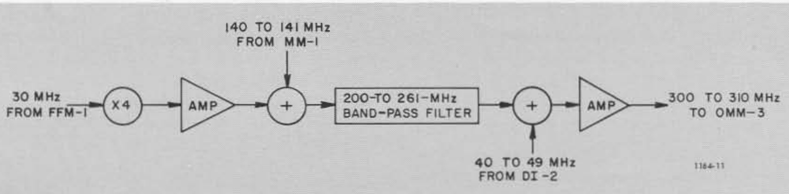


Figure 7. Block diagram of the Intermediate Mixer (IM-1).

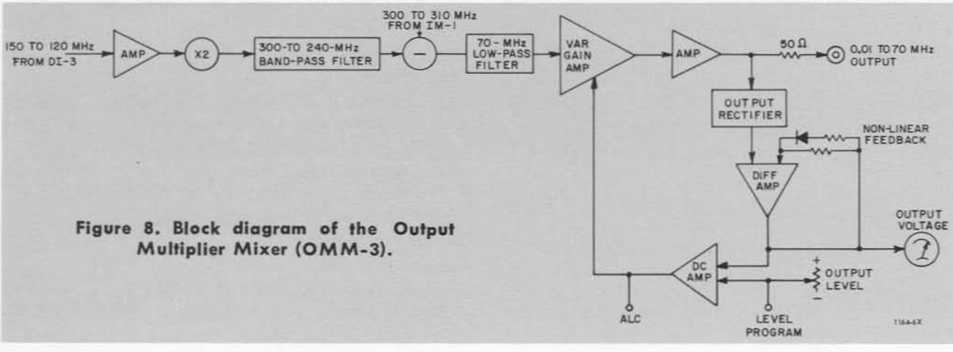
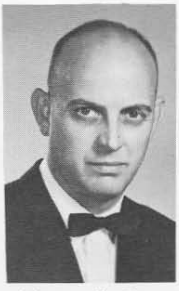


Figure 8. Block diagram of the Output Multiplier Mixer (OMM-3).



William F. Byers was graduated with high honors from Ohio University in 1943 with the degree of BSEE. He was a visiting lecturer in electrical engineering at Ohio University prior to coming to General Radio as a development engineer in late 1943. He is at present a Section Leader in the development group concerned with radio-frequency circuits, including standard-signal generators, oscillators, and frequency synthesizers. He is a member of the IEEE.

band amplifier covering the frequency range of 0.01 to 70 MHz follows the output filter of the mixer, increasing the level to a maximum of 2 volts behind 50 ohms. This level is maintained even with the output connector short-circuited.

Two of the earlier stages are variable in gain by diode-controlled emitter degeneration, permitting automatic output leveling. The last two stages are operated in push-pull, using complementary transistor pairs. Emitter, shunt, and feedback compensation are all used to achieve a reasonably flat response. The full-wave, peak-responding output rectifier senses voltage ahead of an accurate 50-ohm resistor to provide a 50-ohm source.

Following the output rectifier is a dc amplifier with a nonlinear feedback network, which makes the dc output a linear function of the ac signal applied to the rectifier. That is, it linearizes the rectifier characteristic over an ac-voltage range of 0.4 to 2 volts. This amplifier drives the level-control circuits and output meter. The linear scale on the output meter thus gives a true account of output level, and the voltage available for automatic level control is linear over this range. The automatic leveling circuit is completed by a difference amplifier, which compares the indicated output voltage with an adjustable voltage set by the output control or applied at the level program connector, and which delivers a control current to the variable-gain amplifier to minimize the difference. Enough gain variation is available so that any output level from 0.2 to 2.0 volts can be set. This level is then automatically maintained to within 0.3 dB for all load and frequency variations within the range of the instrument. An external detector can easily be connected to the automatic-level-control bus (internal control is disabled when the panel level control is turned to the off position).

Power Supply PS-2

The power supply is an improved version of the PS-1 used in earlier synthesizers. It can supply more current at 18 volts (regulated) to meet the demands of the 1164-A with a 200-mA reserve for accessories. The new, all-silicon-transistor power supply is completely short-circuit proof and current-limited.

A toroidal power transformer is enclosed in an A-metal case to minimize stray fields. A special input jack permits operation of the synthesizer from a battery, with the internal series regulator functioning to maintain normal operation with battery voltages from 20 to 28 volts.

— W. F. Byers

CREDITS

The present four GR Synthesizers are the result of the combined efforts of Atherton Noyes, Jr., Group Leader, G. H. Lohrer, C. C. Evans, and the author.

In the Type 1164-A, the primary responsibility for the new modules, IM-1 and PS-2, was that of G. H. Lohrer; for the new DI-2 and DI-3 modules, that of C. C. Evans. The primary responsibility for the remaining new modules, the FFM-1, MM-1, OMM-3, and the new chassis, as well as coordination of the effort, was that of the author.

SPECIFICATIONS

Frequency Range: 0.01 to 70 MHz.

Smallest Digital Increment: 10 Hz on step-digit controls; 0.1-Hz divisions on CAD dial (−A7C model).

Maximum Bandwidth Controllable by CAD: 1.2 MHz.

Frequency Accuracy: Same as that of 5-MHz driving signal. Internal oscillator, which can be readily phase-locked to an external standard, has temperature dependence of approx $2 \times 10^{-7}/°C$ when operated without external lock, and can be adjusted, unlocked, approx $\pm 5 \times 10^{-6}$ by front-panel screwdriver control.

Spurious-Frequency Levels: Harmonic, at maximum output with 50-Ω load, −30 dB from 0.1 to 70 MHz, −25 dB from 0.01 to 0.1 MHz; discrete nonharmonic, −60 dB.

Output Impedance: 50 Ω.

Output Level: 0.2 to 2.0 V behind 50 Ω.

Levelling Characteristics: With 50-Ω load, ± 3% ± 0.02 V from 0.1 to 70 MHz, ± 5% ± 0.02 V from 0.01 to 0.1 MHz. At single frequency, level is held constant ± 0.02 V as load resistance varies from 0 to ∞.

Output Metering Accuracy: ± 5% from 0.1 to 70 MHz, +5 −10% from 0.01 to 0.1 MHz.

Digital Programming Characteristics (with remotely programmable units, RDI):

Method: 10-line circuit closure for each controlled digit.

Programmable Bandwidth: Any integral megahertz band (e.g., 49–50, 68–69 MHz), in steps as small as 10 Hz (−AR7 model).

Switching Time: <2 ms.

Relay Life: Over 5 × 10^7 operations.

CAD Sweeping and Programming Characteristics

Frequency Range: In 10-kHz or lower functional position, ± 5 major CAD dial divisions. The sweep can be centered on any manual dial setting except in the 100-kHz functional position, where the swept frequency limits are −100 and +1100 kHz.

Sweep Rate: Up to 1 kHz.

Control Voltage Required: −0.3 V per major CAD division.

Beat Output (difference between CAD frequency and that of replaced digit selectors, available at BEAT terminals; panel meter indicates zero beat)

Frequency: 10 kHz per major CAD division.

Frequency Range: 0 to 110 kHz.

Voltage: Greater than 0.5 V behind 3 kΩ.

Level Programming Characteristics

Level Range: 0.2 to 2 V, rms, behind 50 Ω.

Level Response Time: 10 ns for change 95% complete.

Level Programming Control: Either a 5- to 25-kΩ resistance or 6 to 10 V into 5 to 6 kΩ.

External-Frequency-Standard Requirements

Frequency: 5 MHz or any submultiple down to 100 kHz.

Minimum Level: 0.25 V, rms.

Input Impedance at Lock Input Connector: Approx 1 kΩ at minimum level, 50 Ω at 2 V or higher (max allowable level, 5 V).

Rear-Panel Connections

At locking GR874 connectors: 100 kHz and 5 MHz outputs.

At Subminiature Connectors: 1, 5, 5/5.1 and 5/5.1 reference, 30, 42, 40–49, 50/51, and 90 MHz; +18 V dc, outputs.

At BNC connectors: Main output, beat output, CAD control, level program.

Accessories Supplied: TYPE 874-R22LA Coaxial Patch Cord, bridging unit (substitute for DI-1 during maintenance) with panel insert, TYPE CAP-22 3-wire power cord, spare dial lamps and fuses.

Power Required: 105 to 125, 195 to 235, or 210 to 250 V, 50 to 400 Hz, 60 W, max; or 20 to 28 V dc, 1.8 A.

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

Dimensions: Bench model, width 19, height 5¼, depth 19¼ inches (485, 135, 490 mm); rack model, width 19, height 5¼, depth behind panel 17 inches (485, 135, 432 mm), over-all.

Net Weight: 45 lb (20.5 kg).

Shipping Weight: 52 lb (24 kg).

Catalog Number	*Type*	*Units Included*	*Smallest Step (DI only)*	*Price in USA*
1164-9597	**1164-A7C**	7DI + CAD	10 Hz	**$7065.00**
1164-9596	**1164-A6C**	6DI + CAD	100 Hz	**6620.00**
1164-9595	**1164-A5C**	5DI + CAD	1 kHz	**6175.00**
1164-9594	**1164-A4C**	4DI + CAD	10 kHz	**5730.00**
1164-9593	**1164-A3C**	3DI + CAD	100 kHz	**5285.00**
1164-9417	**1164-A7**	7DI	10 Hz	**6525.00**
1164-9416	**1164-A6**	6DI	100 Hz	**6080.00**
1164-9415	**1164-A5**	5DI	1 kHz	**5635.00**
1164-9414	**1164-A4**	4DI	10 kHz	**5190.00**
1164-9413	**1164-A3**	3DI	100 kHz	**4745.00**

PROGRAMMABLE MODELS*

Catalog Number	*Type*	*Smallest Programmable Increment*	*Price in USA*
1164-9527	**1164-AR7C**	10 Hz	**$7515.00**
1164-9526	**1164-AR6C**	100 Hz	**6980.00**
1164-9525	**1164-AR5C**	1 kHz	**6445.00**
1164-9524	**1164-AR4C**	10 kHz	**5910.00**
1164-9523	**1164-AR3C**	100 kHz	**5375.00**
1164-9507	**1164-AR7**	10 Hz	**6975.00**
1164-9506	**1164-AR6**	100 Hz	**6440.00**
1164-9505	**1164-AR5**	1 kHz	**5905.00**
1164-9504	**1164-AR4**	10 kHz	**5370.00**
1164-9503	**1164-AR3**	100 kHz	**4835.00**

*The X10 MHz and X1 MHz decade units are not programmable. However, these two decades can be programmed between whole 1-MHz steps (e.g., between 11 and 12 MHz, 50 and 51 MHz, etc).
US Patent No. 2,548,457. Patents applied for.

A ONE-PERCENT-BANDWIDTH WAVE ANALYZER

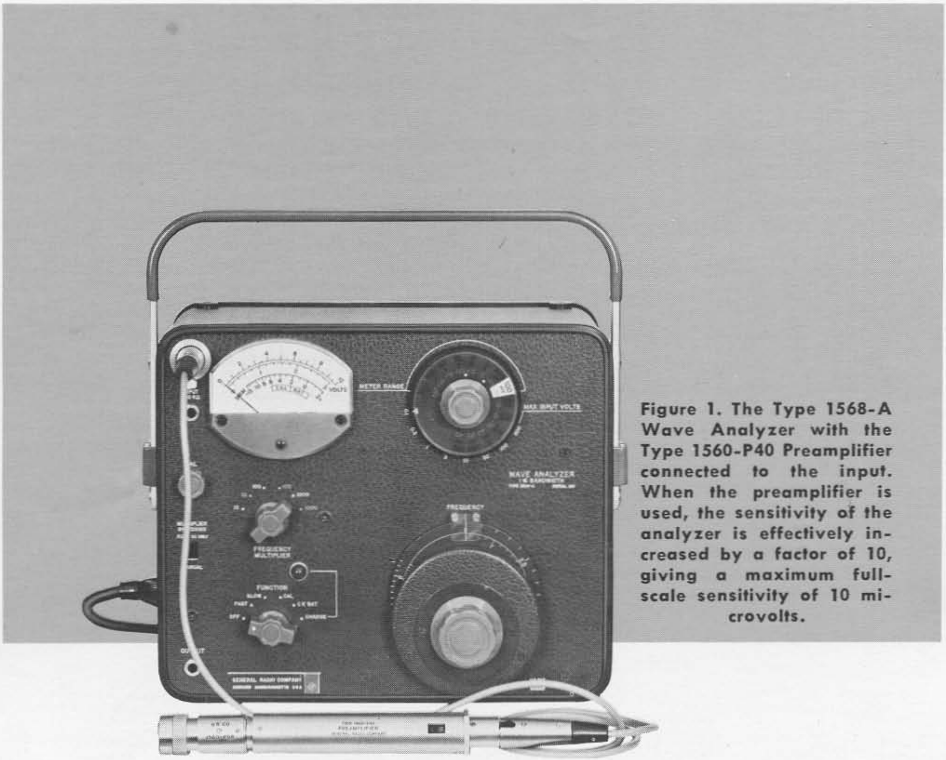


Figure 1. The Type 1568-A Wave Analyzer with the Type 1560-P40 Preamplifier connected to the input. When the preamplifier is used, the sensitivity of the analyzer is effectively increased by a factor of 10, giving a maximum full-scale sensitivity of 10 microvolts.

To be classed as a wave analyzer, an instrument must have a very narrow bandwidth filter to make possible the separation of closely spaced discrete frequency components. The filter must also have high initial-cutoff rate and high ultimate attenuation, so that it can resolve small frequency components in the presence of larger ones. It should be capable of "seeing" at least 70 dB into a spectrum. Further, its self-generated distortion should be so low that the analyzer itself, under normal conditions, cannot detect it. The new Type 1568-A Wave Analyzer, as well as most fixed-bandwidth-type analyzers, meets these requirements.

Narrow-band, constant-percentage-bandwidth analyzers are not generally classed as wave analyzers. They have traditionally been characterized by poor filter shape, inadequate stop-band attenuation, and limited dynamic range. Low ultimate filter attenuation, resulting from the use of twin-T filter circuits, has often limited analyzing range to no more than 30 or 35 dB.

The new TYPE 1568 Wave Analyzer, like the TYPE 1564,* (Figure 1) incorporates many recent advances in circuit

* W. R. Kundert, "New Performance, New Convenience with the New Sound and Vibration Analyzer," *General Radio Experimenter*, September-October 1963.

design to overcome these shortcomings. Chief among these advances is a new filter design. This and other performance features of the new analyzer are described below.

PERFORMANCE FEATURES

Filter Response (see Figure 2)

The bandwidth is narrow, 1% of the selected frequency. The filter, comprised of two resonant sections, has an initial attenuation rate of about 600 dB per octave, and attenuation at twice and at one half the selected frequency is at least 75 dB. The filter characteristic does not flatten in the stop band; rather, the attenuation rate approaches 12 dB per octave far from the center frequency, and attenuation is still increasing at the noise level of the instrument.

Sensitivity and Analyzing Range

Other features of the new analyzer include full-scale meter ranges from 100 microvolts to 300 volts (in 10-dB steps) and extremely low input-circuit distortion, so low that it cannot be detected by the analyzer itself, which can detect harmonic-distortion products as low as 0.01%. For manual analysis, the analyzing range is never limited by input-circuit distortion and rarely by lack of sensitivity. Sensitivity can be increased to 10 microvolts, full scale, at an input impedance equivalent to 6 pF shunted by 500 megohms, by use of a Type 1560-P40 Preamplifier. The preamplifier, shown in Figure 1, is designed to drive long connecting cables, permitting the analyzer to be operated remote from the signal source. Power for the preamplifier is supplied by the analyzer.

Calibration

A built-in, feedback-type calibrator can check amplitude calibration at any frequency. When greatest accuracy is required, the instrument can be calibrated at the frequency of each component in the spectrum being measured.

Dynamic Range—Automatic Recording

The Type 1568-A has a wide dynamic range, which is necessary for a wide analyzing range in automatic recording. Dynamic range, an often-misunderstood term, means the range between overload and noise level at any one setting of the attenuator controls

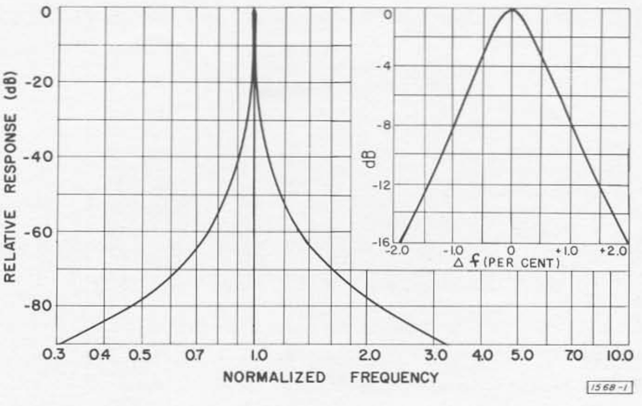


Figure 2. Attenuation characteristic of the filter.

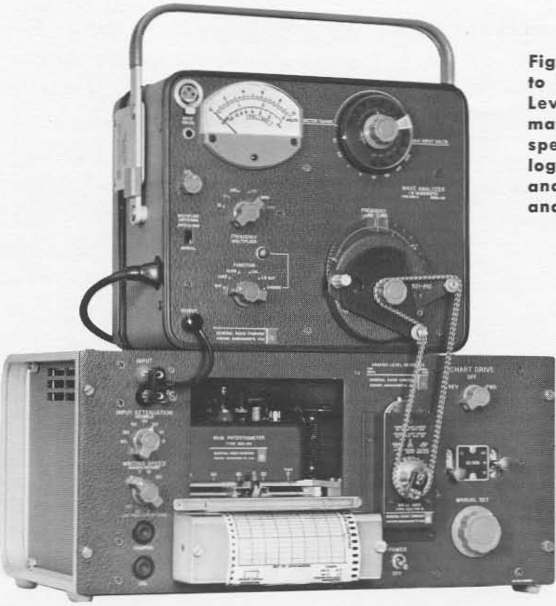


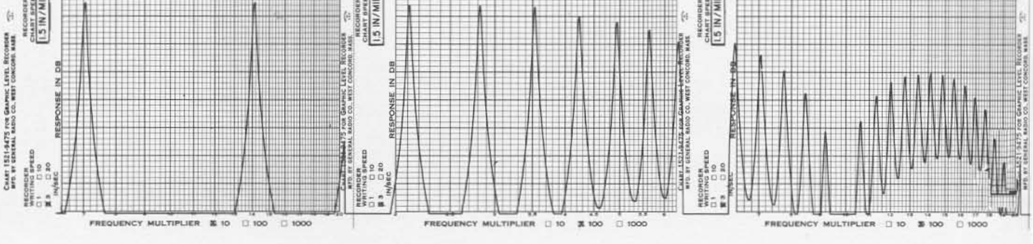
Figure 3. The analyzer coupled to the Type 1521-B Graphic Level Recorder for the automatic plotting of frequency spectra. The chart paper has a logarithmic frequency scale, and frequency ranges on the analyzer are changed automatically.

The dynamic range of the Type 1568-A varies somewhat with frequency, but it is sufficient to yield a recording range that approaches 80 dB when the overall signal applied to the analyzer is 30 millivolts or greater. The Type 1568-A has been designed for convenient automatic recording with the Type 1521-B Graphic Level Recorder[1] (combination shown in Figure 3). The frequency range changes automatically when the main frequency control is driven. The chart produced has a frequency scale of 10 inches per decade and a 4-inch vertical scale of 20, 40, or 80 dB, depending on the recorder potentiometer used.

[1] Martin W. Basch, "New Talents for the Graphic Level Recorder," *General Radio Experimenter*, September 1964.

APPLICATIONS

The following discussion of applications necessarily includes a comparison of the Type 1568-A and fixed-bandwidth wave analyzers. It should be emphasized that the Type 1568-A's advantages of low price, portability, and ease of use will, for some applications, outweigh some of the features of the complex and more expensive fixed-bandwidth analyzers.



Measurement of Harmonic Distortion

Harmonic distortion is readily measured with either type of analyzer, though some fixed-bandwidth models do not have sufficient filter attenuation at frequencies below 100 Hz. The second harmonic, being closest to the fundamental, is the most difficult to resolve. When the analyzer is tuned to the second harmonic, attenuation of the fundamental must be sufficient to reduce its level to less than that of the harmonic. Half-frequency attenuation in the TYPE 1568-A Wave Analyzer is at least 75 dB, independent of the frequency to which it is tuned. When the third harmonic is selected, the fundamental is attenuated by more than 85 dB.

Harmonic Analysis

The TYPE 1568-A will separate about 50 harmonics, a sufficient number for almost all applications. Above this number, it will display the envelope of the spectrum, which usually contains sufficient information. Figure 4 illustrates this effect.

When more than 50 components of a simple periodic signal are to be resolved, there may be some advantage in a fixed-bandwidth analyzer and a linear frequency scale for recording. The separate harmonics are spaced an equal number of hertz apart, and so, if a fixed-bandwidth analyzer has sufficient attenuation to resolve a few harmonics (it may not at low frequency), it will resolve them all within its amplitude and frequency limits.

Measurements on Modulated Signals

A periodic signal modulated with a simple periodic signal also has equally spaced components. Component spacing is equal to the fundamental frequency of the modulating signal.

Whether the TYPE 1568-A can resolve a carrier and sidebands depends on the ratio of the carrier frequency to the lowest frequency component of the modulating signal and also on the relative amplitude of these components. As a rule of thumb, the frequency ratio must be less than 50.

Measurement of Discrete Components at Low Frequencies

The bandwidth of the TYPE 1568-A is reduced to 0.2 Hz at its low-frequency limit, much narrower than that of even the narrowest fixed-bandwidth analyzers. At low frequencies it can separate components spaced only a fraction of one hertz apart. Furthermore, in this range, it does not have the annoying frequency drift associated with some heterodyne instruments, and its frequency accuracy and dial resolution are far superior. The filter curves for the TYPE 1568-A Wave Analyzer

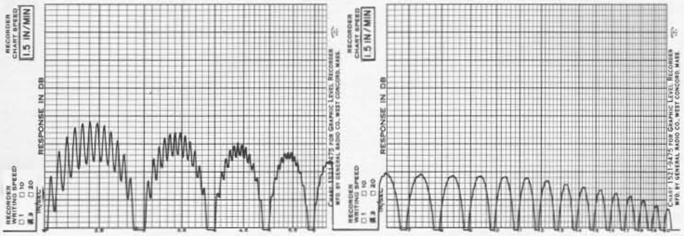


Figure 4. Frequency spectrum analysis of a 1.0-ms pulse at a 70-Hz repetition rate. The 1% bandwidth yields high resolution at low frequencies, shows the envelope at high frequencies.

and the Type 1900-A Wave Analyzer, [2] a constant-bandwidth instrument, are compared in Figure 5 at center frequencies of 20 and 60 Hz. The bandwidth of the Type 1900-A is set to 3 Hz, its narrowest.

Measurements of Discrete Components of Other Types of Signal

In a data-transmission system many carriers may be used for information transmission. It is often desirable to monitor each carrier individually for amplitude and frequency. The Type 1568 and a frequency meter make an excellent combination for this purpose.

In an Instrument Landing System, distortion products must be low to ensure an accurate, course-direction indication in the cabin of a landing aircraft. The analyzer can be used to check distortion by measuring the spectrum of the sideband energy produced by the ILS Localizer Transmitter.

Random-Noise Measurements

The detail required in the analysis of noise spectra containing no important discrete components does not often warrant the use of a bandwidth as narrow as that of the Type 1568-A. The one-third- or one-tenth-octave bandwidths of the Type 1564-A Sound and Vibration Analyzer are better suited to this application. The Type 1568-A is generally used when a recorded analysis is desired. However, the analyzer can be used manually and a slow meter speed is included to facilitate noise measurements.

Analysis of Machinery Noise and Vibration

The reduction of the sound and vibration generated by machines and appliances is becoming increasingly important,[3] because of their effect on man and because they result in mechanical failure and excessive wear of machine

[2] Arnold Peterson, "New Wave Analyzer has 3 Bandwidths, 80-dB Dynamic Range," *General Radio Experimenter*, April 1964.
[3] Arnold Peterson, "Vibration: Problems, Measurements, and Control," General Radio Preprint B-22.

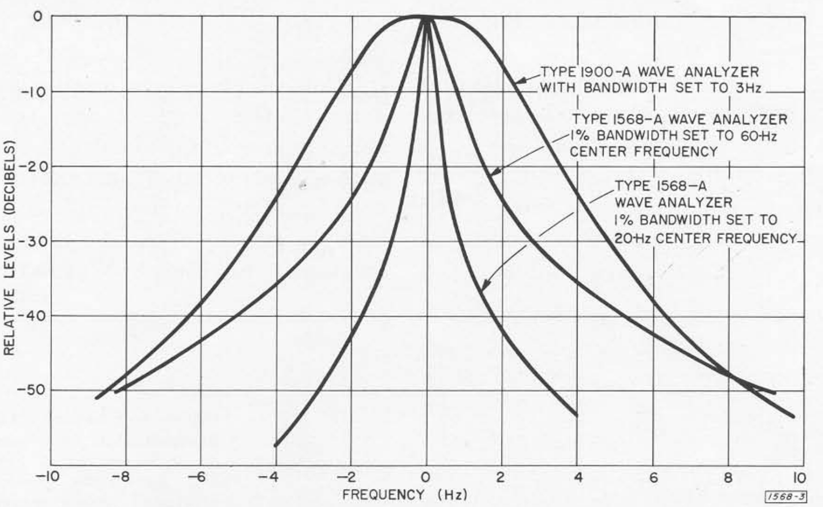
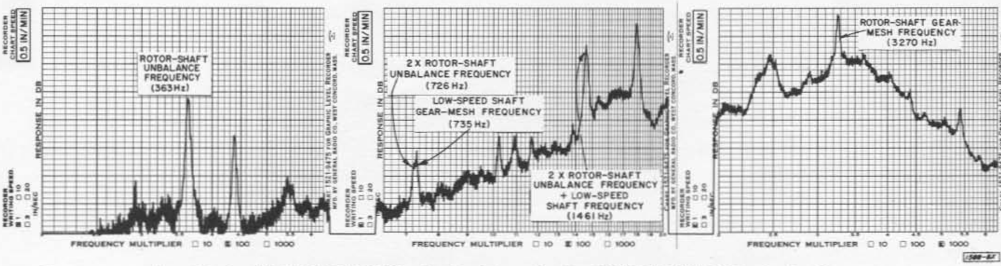


Figure 5. Comparison of the filter characteristics of the Type 1568-A Wave Analyzer at 20 Hz and 60 Hz with the 3-Hz bandwidth of the Type 1900-A Wave Analyzer.



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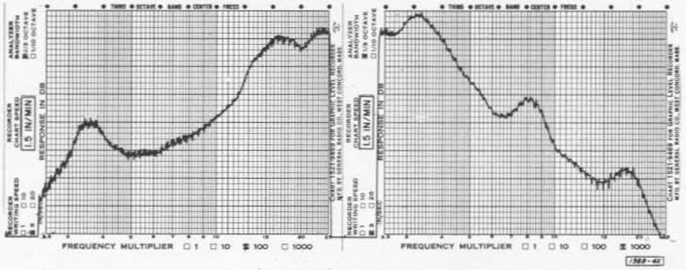

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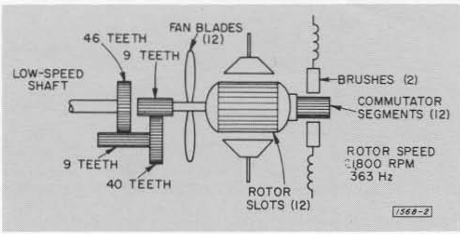


Figure 6. Chart records of the vibration acceleration spectrum of a motor and gear-train assembly (see sketch). (a) A 1%-bandwidth analysis, taken with the Type 1568-A Wave Analyzer and (b) a ⅓-octave analysis. For diagnostic measurements of this kind, the detail of the spectrum provided by the 1% bandwidth is by far the better.

elements. The source of the noise within the machine must first be identified, then eliminated or reduced. A fine-spectrum analysis of the sound or vibration (or both) with the Type 1568-A Wave Analyzer will show various discrete frequencies and resonances, which can be related to motion within the machine. Broader-band octave or one-third-octave measurements are of little use in such diagnostic applications.

Figure 6 shows the acceleration spectrum measured on the housing of a machine employing a high-speed universal motor and gear train. For comparison, a one-third-octave spectrum is also shown. The various sources of vibration that are identified include frequencies caused by motor-armature unbalance and gear-teeth meshing. The large component at 3270 Hz is critical, from the standpoint of its effect on man, since it is in the frequency range where human hearing is most acute. A reduction in the level of this component through improved gear design at this point would markedly reduce the loudness and speech-interference level of the sound. Alternatively, to reduce loudness, the frequency of this component might be shifted by a change in design. Various other components will be found critical when other aspects of the vibration problem are considered. For example, bearing wear could be reduced by improved armature balance.

Mil-Std-740B

Mil-Standard-740B (SHIPS) (Airborne and Structureborne Noise Measurements and Acceptance Criteria of Shipboard Equipment) allows a narrow-band vibration analysis in the frequency range below 500 Hz when requirements cannot be met by a one-third-octave analysis. The acceptance limits are the same in either case. A narrow band is defined as ". . . a band whose width is not less than one percent or more than eight percent of the band center frequency." This specification should make the TYPE 1568-A an important tool for acceptance testing in accordance with the standard.

DESCRIPTION

Input

The block diagram of Figure 7 shows the various sections that make up the analyzer. The INPUT drives a three-section attenuator. The first section is operated by the dial of the coaxial METER RANGE switch and is set in accordance with the over-all level of the signal being measured. The second block in the diagram represents part of the analyzing attenuator, which is adjusted by the knob of the same switch. A third attenuator, which is continuously adjustable, is used in amplitude calibration. The attenuators present a nearly constant impedance of 100 kΩ at the input connectors.

[4] W. R. Kundert, "The RC Amplifier-Type Active Filter: A Design Method for Optimum Stability," *IEEE Transactions on Audio*, Vol. AU-12, No. 4, July-August 1964, p 66–71. (General Radio Reprint A-113).

Filter

The filter used in the TYPE 1568-A is unique.[4] Active RC filters often require an unreasonable degree of stability in one or more filter components to maintain stable transmission with tuning, time, temperature, and other variables. However, it is possible by careful design to trade off component sensitivities and thus to reduce the filter sensitivity to a given disturbing effect. In the TYPE 1568-A, the filter has been designed for low sensitivity to tracking errors in the tuning potentiometer. This design also results in relatively low sensitivity of the filter to drift in all other passive components. The price paid is a high sensitivity factor for amplifier gain, but, surprisingly, an amplifier can readily be constructed with much better stability than that of readily available passive components.

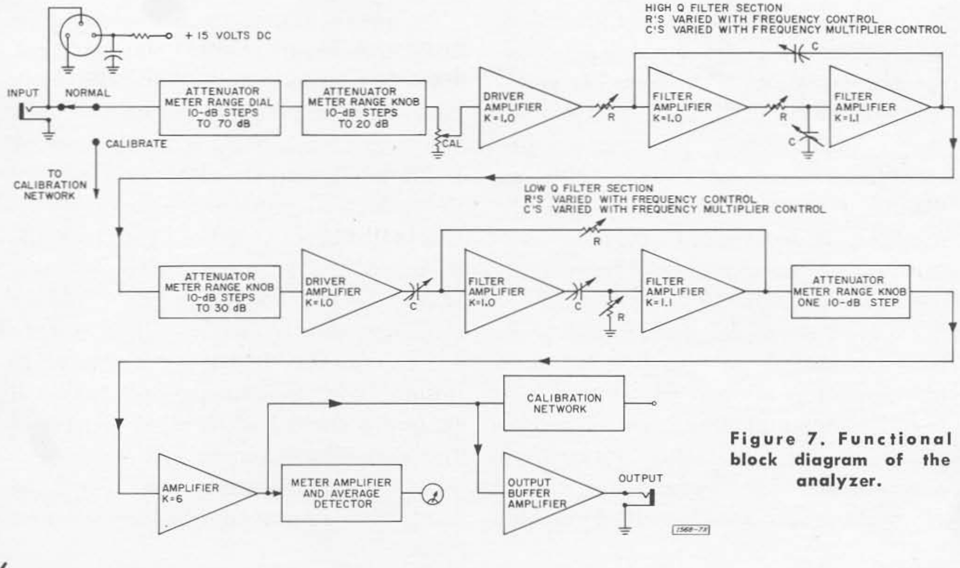


Figure 7. Functional block diagram of the analyzer.

The filter is synthesized as an isolated cascade of two active, RC resonant sections. These sections differ in Q to reduce further the effect of potentiometer tracking error. The audio range from 20 Hz to 20 kHz is covered in six bands, each spanning a half decade (range of 3.16 to 1). Within any one range the instrument is tuned by means of a four-gang potentiometer, whose resistance-versus-angle characteristic produces a logarithmic frequency scale on the FREQUENCY control. Close-tolerance polystyrene capacitors are switched to change ranges. Manual or automatic range changing is selected with a panel switch. The automatic mode is used in recording with the Graphic Level Recorder. Three-transistor filter amplifiers provide the necessary high ratio of input to output impedance and stable voltage gain. As indicated by the block diagram, sections of the analyzing attenuator are included between the first and second filter stages and after the second stage. Through this separation of the analyzing attenuator into three sections, wide dynamic range can be maintained over a wide range of input signal levels.

Output

A final amplifier supplies a signal for the meter circuit and output buffer. The output is isolated, so that performance is not affected by the load. The 6000-ohm output is intended primarily for driving a recorder or counter, although any load can be connected. The meter, driven by an average-type detector, has three full-scale ranges: 0 to 3 and 0 to 10 volts and −15 to +2 dB for reading in dB referred to one milliwatt in 600 ohms.

Calibrator

The feedback-type calibrator used to standardize the gain is similar to that used in other GR analyzers. To amplitude-calibrate the instrument, a signal from the output amplifier is fed to the input through a limiter and a calibrated attenuator. When the gain of the analyzer is adjusted, by means of the CAL control, to equal the loss in the feedback path, the system oscillates, signifying that the instrument is calibrated as a voltmeter. The frequency of oscillation is determined by the center frequency of the filter, and this calibration can be made at any selected frequency. For the measurement of levels in dB or percent, with an arbitrary reference, a special attenuator dial is supplied, and the instrument is calibrated to suit a reference level.

Power Supply

The power supply permits either line or battery operation. The battery supplied is a rechargeable nickel-cadmium unit, which also

Warren R. Kundert received his BSEE degree in 1958 and his MSEE in 1961 from Northeastern University. He came to General Radio as a development engineer in the audio group in 1959, where he is engaged in the design of audio-frequency circuits, filters, and other networks, and in the development of instruments for measurements at audio frequencies. He is a member of the IEEE, Acoustical Society of America, Audio Engineering Society, and Eta Kappa Nu.

serves as a ripple filter for line operation. The built-in charger operates from the ac line. A fully charged battery provides about 20 hours of operation and requires 16 hours for recharging. Though the instrument should normally be operated using line power when it is available, because of grounding considerations it is sometimes desirable to operate the instrument isolated from the line.

Cabinet

The analyzer is packaged in the General Radio Flip-Tilt case, whose protective cover serves as a mounting base when the instrument is in use and allows its panel to be adjusted to a convenient angle for operation. Alternatively, the analyzer can be supplied adapted for rack mounting.

The Type 1568-A Wave Analyzer is an important new instrument for both electrical and sound-and-vibration analysis when high resolution is needed. For sound-and-vibration work, it fills a gap between the Type 1564-A Sound and Vibration Analyzer (1/3 octave and 1/10 octave) and the Type 1900-A Wave Analyzer. It combines the desirable features of a narrow, constant-percentage-bandwidth analyzer with an excellent filter shape. Its low cost, portability, and simplicity of operation, and the option of line or battery operation are features seldom encountered in other wave analyzers.

— W. R. Kundert

SPECIFICATIONS

FREQUENCY

Range: 20 Hz to 20 kHz in six half-decade bands.
Dial Calibration: Logarithmic.
Accuracy of Frequency Calibration: 1%.
Filter Characteristics: Bandwidth between 3-dB points on selectivity curve (see Figure 2) is one percent of selected frequency.

Attenuation at 20% above and at 20% below selected frequency is greater than 50 dB referred to the level at the selected frequency. Attenuation at twice and at one-half the selected frequency is at least 75 dB referred to the level at the selected frequency. Ultimate attenuation is greater than 85 dB.

Uniformity of filter peak response with tuning is ±1 dB from 20 Hz to 6.3 kHz and ±2 dB from 20 Hz to 20 kHz.

INPUT

Impedance: 100 kΩ.
Voltage Range: 100 μV to 300 V, full scale, in 3–10 series steps. Power is supplied at input socket for the TYPE 1560-P40 Preamplifier, which extends the sensitivity to 10 μV, full scale, and increases the input impedance to more than 500 MΩ.
Distortion: Input-circuit distortion is lower than −80 dB relative to input-signal level.

OUTPUT

Impedance: 6000 Ω. Any load can be connected.
Voltage: At least one volt open circuit when meter reads full scale.
Crest-Factor Capacity: Greater than 13 dB.
Output Meter: In addition to normal-speed mode, meter has slow-speed mode for manual measurements of noise.

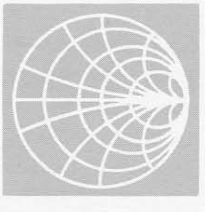
GENERAL

Analyzing Range: 80 dB. Components of an input signal that differ in amplitude by as much as 80 dB can be measured.
Automatic Recording: Automatic range switching is provided to allow convenient, continuous spectrum plotting when the TYPE 1521 Graphic Level Recorder is used. Medium-speed motor is recommended. Chart paper is Catalog No. 1521–9475. Frequency scale is logarithmic, 10 inches per decade; vertical scale is 4 inches for 20, 40, or 80 dB, depending on the potentiometer used in the recorder.
Amplitude Calibrator: A built-in, feedback-type calibration system permits amplitude calibration at any frequency.
Accessories Supplied: TYPE CAP-22 Power Cord; TYPE 1568–2090 Detented Knob and Dial Assembly, used to facilitate measuring the components of an input signal as a percentage or in decibels with an arbitrary voltage reference.
Power Supply: 100 to 125 or 200 to 250 V, 50 to 60 Hz. 2 W for normal operation, 3.5 W for battery operation. A rechargeable nickel-cadmium battery is also supplied. Battery provides about 20 hours of operation when fully charged and requires 16 hours for charging. Internal charger operates from the power line.
Mounting: Flip-Tilt case.
Dimensions: Portable model, with case closed, width 13¼, height 13, depth 8¼ inches, (340, 330, 210 mm), over-all, including handle. Rack model, 19-inch (485 mm) rack panel, 12¼ inches (35 mm) high; depth behind panel 5 inches (130 mm).
Net Weight: 21½ lb (10.0 kg).
Shipping Weight: 27 lb (12.5 kg).

Catalog Number	*Description*	*Price in USA*
1568-9701	**Type 1568-A Wave Analyzer,** Portable Model	**$1350.00**
1568-9820	**Type 1568-A Wave Analyzer,** Rack Model	**1350.00**
1560-9510	**Type 1560-P40J Preamplifier and Adaptor Set** *	**184.00**

U. S. Patent Nos. 3,012,197; D187,740; 2,966,257.
* For complete details see *General Radio Experimenter*, June 1965.

New BIG Smith Charts



Occasional inquiries for Smith Charts larger than the standard 8½" x 11" size have induced us to introduce a new giant-size chart, which is 22½" x 35". These charts, direct enlargements of the regular normalized Smith Chart, are printed on relatively heavy paper in red ink. We assume their greatest use would be in classrooms or conferences where several people are involved. However, some presbyopic engineers may also find them useful. The new charts are Catalog No. 5301-7563 NX, and pads of about 75 sheets are $6.00.

ADDITION — The publication date of the report given in Footnote 1, page 3 of the August *Experimenter* was August 19, 1963.

VIBRATION ANALYSIS KEEPS 'COPTERS FLYING

Boeing "Chinook" helicopters in Viet Nam are providing better service at less expense to Uncle Sam, thanks in part to GR's Type 1564-A Sound and Vibration Analyzer. Field service engineers of Boeing's Vertol Division reporting from the combat zone tell of consistent success with the analyzer in spotting potential problems in the helicopters' engines, transmissions, drive shafts, and auxiliary equipment, as part of normal preventive maintenance procedures. Early diagnosis of impending trouble helps keep these modern-day pack mules in service when they are so badly needed and saves countless tax dollars in repair costs.

For use with the Sound and Vibration Analyzer, Boeing specifies the Type 1560-P52 Vibration Pickup, the Type 1560-P35 Permanent-Magnet Clamp, and a 25-foot extension cable for the pickup. The magnetic clamp is used to secure the pickup to aluminum bolts (which contain steel) on the helicopter structure. For locations where magnetic clamping is impossible, Boeing's engineers use a vice-grip clamp, to which has been brazed a bracket with a ¼-28 threaded stud that accepts the pickup.

Trouble-shooting usually consists of measuring the frequency of a vibration and then tracking down its source by relating the vibration frequency to the characteristic speeds of various components.

This vibration-measuring system has been praised for sensitivity (better than 0.001 G), frequency accuracy (±2%), and frequency discrimination (7% bandwidth). Vibration components at frequencies from 3.8 to 250 Hz

the Experimenter

GENERAL RADIO COMPANY
WEST CONCORD

Photo courtesy of The Boeing Company, Vertol Division

Engineer measures the frequencies of the vibration components in a helicopter transmission. He can then track down the vibration sources by relating the measured frequencies to the characteristic speeds of the various transmission parts.

are measured with separations as small as 1.5 Hz at low frequencies.

The portability of the 1564 is all-important in this application. In-flight measurements can be made with ease. The rechargeable batteries can be charged overnight from any field generator to be ready at a moment's notice.

— C. W. ALSEN

THE GENERAL RADIO

Experimenter

A BETTER WAY TO ACOUSTICAL MEASUREMENTS: TAPE

ALSO IN THIS ISSUE:

EARPHONE COUPLERS FOR AUDIOMETER CALIBRATION

A-WEIGHTING ADDED TO OCTAVE-BAND ANALYZER

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Figure 1. The Type 1525-A Data Recorder.

A MAGNETIC TAPE RECORDER FOR ACOUSTICAL, VIBRATION, AND OTHER AUDIO-FREQUENCY MEASUREMENTS

Magnetic tape recorders are widely used in science and engineering for storing signals from transducers in a reproducible form. Although a remarkable variety of tape recorders is available, only one, the new Type 1525-A Data Recorder, is designed specifically for the important field of acoustic-noise measurement. It is an unusually useful tool in noise-control programs and a basic instrument for many acoustical measurements. The use of this recorder is, of course, not limited to recording signals from microphones only; electrical signals from other transducers, such as those from vibration pickups, can be measured, recorded, and reproduced, as can any signals in the audio-frequency range.

APPLICATIONS OF RECORDERS IN ACOUSTICAL MEASUREMENTS

Owing to the widespread use of tape recording, the general principles of operation are well known, and some applications to acoustic-noise studies are obvious. But in order to discuss the features that make the Type 1525-A Data Recorder particularly well suited for acoustic-noise measurements, it will be helpful first to review a number of these applications.

Non-Steady-State Sounds

Tape-recording a noise is sometimes an obvious first step in getting the information needed to solve a noise problem. For example, if the sound is of relatively short duration, as in a rocket blastoff, many machining operations, a single note from a musical instrument, and speech syllables, a recording for later playback and analysis is often almost essential. One can then form a section of the recorded tape into a loop for continuous playback to facilitate the frequency analysis.

Similarly, impact or explosive sounds, such as those produced by blasting, gun shots, drop forges, and typewriters, can be easily studied after they are made repetitive by means of tape loops.

The analysis of intermittent sounds or signals can be a helpful step in tracking down the sound sources. By means of tape recording, one can monitor the noise from a device for long periods to catch these intermittent sounds, which can then be separated out for analysis.

Some sounds vary significantly in level and character with time. Appliances that go through a cycle of different operations (dishwashers and clothes washers, for example) produce such sounds. Aircraft flyby is another example. Although an appliance can be programmed to stay in the same phase of the cycle for long periods, it is usually more convenient to make a recording of each phase. Sections can then be separated out for detailed analysis.

Some devices, for example a gas engine, drift slowly but significantly in speed. As a result, the basic noise pattern changes, and the drift is often serious enough to preclude direct, detailed analysis of the noise spectrum at a variety of speeds with the usual slow-scan techniques. One can, however, run the engine for a reasonable period at each of a number of selected nominal speeds and record short samples, say two seconds, at each of these speeds to form a series of tape loops. Each loop is played back and is analyzed by a wave analyzer. Since the inertia of the rotating system is often so large that serious fluctuations in speed do not occur in the short interval of the tape loop, the engineer obtains a series of frequency spectra that can be related to shaft speed. He may then be able

Figure 2. Recorder, showing loop mechanism.

to deduce much about the noise producing mechanisms from the relations between amplitude, frequency and shaft speed.

On-Site Recording

The complex instruments used in studying a noise can be installed and operated most conveniently in an acoustical laboratory set up for that purpose. If the noisy machine cannot readily be brought to the laboratory for study, the noise can be recorded at the site and later analyzed under laboratory conditions.

When it is expensive, impractical or inconvenient to run a noisy device for long periods or repeatedly for short periods, tape-recording simplifies the problem of studying the sound. Jets, rockets, turbines, compressors, and other large machines are expensive to run, and tape recordings of the noise they generate will save money.

An experienced engineer can often decide quickly just what information he needs to produce an economical solution to a noise problem. Occasionally, however, he may find that he needs more information than he at first expected. If he still has the original source to test, he can proceed to get the added information. If he has gone to some distant place to make the measurement, however, he may not have brought with him the equipment that he needs for a more detailed test. If he makes a tape recording of the noise, he can later make as detailed a study as he wishes at his home base.

Avoiding Ambient Noise Interference

Background noise may make noise studies of machines impractical when acoustically isolated rooms are not available. Often, however, the background noise is much less during lunch periods or outside normal working hours, particularly early in the morning, and measurements may then be practical. Even during such periods a complete study of the noise may be awkward or inconvenient, but, if tape recordings can be made during the quiet periods, the recorded signal can be analyzed at any convenient time.

Frequency and Time Scaling

Tape recording has also been used for frequency translation to bring a sound into the optimum operating range of particular instruments used for the analysis. Although dramatic changes of scale are not possible in one step, the TYPE 1525-A Data Recorder does make possible a two-to-one shift. Both the frequency and time scales are changed.

When recording and playback are at different tape speeds (this is discussed under Time and Frequency Translation, page 12), the frequency shift for wave analyzers of the constant-hertz-bandwidth type results in an apparent change in effective bandwidth, which may also be useful. The time-scale shift can be particularly useful in effectively doubling the speed of graphic recorders.

Impact sounds have a very rapid rise in sound level following the impact, but the subsequent decay of the sound is much slower. If one desires to record graphically the peak level, the task is less demanding of the instrumentation when the time scale of such a transient is reversed. The tape recorder makes this inversion possible. Incidentally, one can dramatically illustrate the significance of attack, or initial transient, on the subjective response to a sound by playing backwards a tape recording of a piano selection.

Comparisons of Recordings

Tape recorders are used to keep reproducible records of progressive changes in a sound. These changes may be a result of the application of successive noise-control procedures, for example. Such a series of records may be particularly useful for demonstrating to a consultant's client or to the management of a plant what has been accomplished in reducing noise. From an engineering viewpoint, these recordings are valuable when a change in plans requires a change in analysis procedure.

The recordings of noisy devices constitute a useful library of reproducible sounds. They can be a convenient means of demonstrating comparisons between different makes of the same product or of demonstrating the probable effects of contemplated noise-control procedures.

The tape recorder permits one to make subjective comparisons of the noise from different devices when the devices themselves may not be available at the same time or place, or when they cannot be conveniently operated in such a fashion that easy and psychologically valid comparisons can be made.

The Second Channel

By virtue of its two channels, the TYPE 1525-A has a number of additional advantages and uses.

Source and Detector

The recorder can function as a measurement system with one channel used to play a recorded test signal and the other channel used to record the results of the application of the signal to the device under test. The recorded source signals may be a series of third-octave or octave bands of noise, a pure tone swept over the audio range, or any of a wide variety of audio signals. Reverberation measurements can be facilitated by this technique.

A recorded signal can also be played back and measured on the main amplifier to obtain an A-weighted level, for example.

Narration

The second channel is often used as a "talk-in" channel to describe the conditions of the test, the settings of the controls, any changes that occur during the test, and other pertinent details. This procedure, when prompted by a check list, helps one to avoid mistakes in the original recording procedure and also ensures that a recording does not become useless if the written summary of the original test conditions is mislaid.

Timing

The signal recorded on the second channel can be a timing signal that is related to the information recorded on the first channel. For example, in studies of the vibration of rotating machinery, one may wish to relate the vibration at a particular instant to the angular position of a rotating shaft. A signal controlled by a contactor or a photoelectric pickoff can be recorded on the second channel for this purpose.

GENERAL DESCRIPTION OF THE INSTRUMENT

The TYPE 1525-A Data Recorder operates on the same principles as other magnetic-tape direct-recording systems

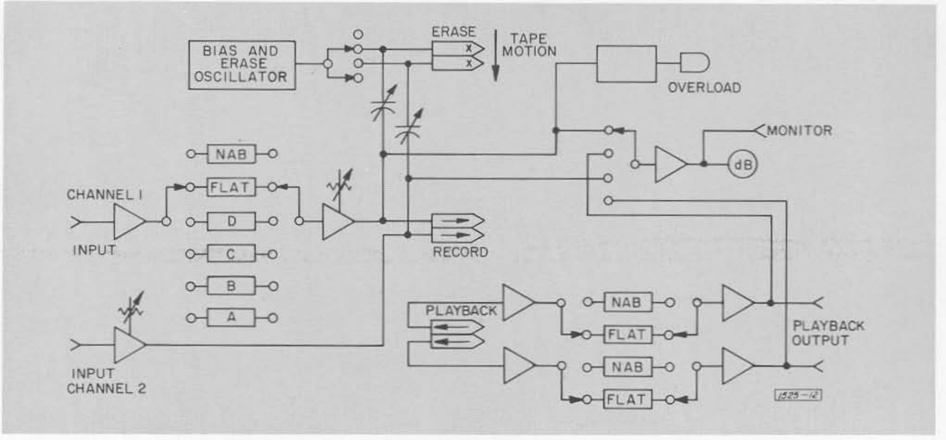

Figure 3. Block diagram of the recorder.

already described in detail in other publications, and a review of these principles is therefore unnecessary here. It does, however, differ from other tape recorders in a number of details that make it better suited for acoustic noise measurement and noise-control applications. These details will be discussed after a brief summary of the elements that make up the recorder.

Tape Transport

The TYPE 1525-A Data Recorder uses an Ampex PR-10-2 tape-transport mechanism, a widely used, professional-quality unit. This transport has two tape speeds, 15 and 7½ in/s. The tape speed is exceptionally constant, with low flutter and wow.

Electronic System

The various elements of the electronic system are shown in the block diagram of Figure 3. The main amplifier, called channel 1, has a field-effect input stage, followed by a series of weighting networks and a high-gain amplifier, whose gain is controlled by an accurate step attenuator and a continuous control. This amplifier supplies a signal to one of the windings on the record head and to an overload indicator. The signals to the record heads can be disconnected by a selector switch.

The second record amplifier is a simple two-stage transistor unit that supplies a signal to the other winding on the record head. This channel can be used in combination with the first to provide binaural recording.

When additional gain is needed, the TYPE 1560-P40 Preamplifier[1] (20 dB) or an auxiliary TYPE 1551-C Sound-Level Meter[2] (100 dB) is recommended.

When the instrument is recording, the bias-and-erase oscillator at about 95 kHz is energized, and it supplies an erase signal and a bias signal for the erase and record heads, respectively, for the selected channel or channels.

The outputs of both windings of the playback head are amplified by two identical amplifiers, and the outputs

[1] C. H. Woodward, "A New Low-Noise Preamplifier," *General Radio Experimenter*, June 1965.
[2] E. E. Gross, "TYPE 1551-C Sound-Level Meter," *General Radio Experimenter*, August 1961.

Figure 4. The Type 1560–P40K Preamplifier and Microphone Set, with an additional Type 1560–P5 Microphone, is a convenient accessory for two-channel operation of the recorder.

of the amplifiers are always available at jacks on the front panel, even during recording. Each amplifier includes the equalization circuits for NAB recording and for constant-current recording at 7½ and 15 in/s.

A monitoring amplifier that drives an output jack and an indicating instrument with the characteristics specified for sound-level-meter instruments can be switched to monitor the signals either at the record head or at the outputs of the playback amplifiers.

IMPORTANT FEATURES

The field-effect input stage for the main amplifier has so high an input impedance that a TYPE 1560-P5 piezoelectric ceramic measurement microphone or a vibration pickup can be used directly at the input. Sufficient gain is provided to make possible a setting of the calibration control for direct reading of sound level over a range of 34 to 140 dB (re 0.0002 μbar), even with 25 feet of cable between the microphone and the data recorder. These circuits, with the TYPE 1560-P5 Microphone, meet the requirements for sound-level meters as specified in American Standard S1.4-1961 and the International Electrotechnical Commission 123-1961.

If the best signal-to-noise ratio is desired at low-to-moderate sound levels, use of the TYPE 1560-P40 Preamplifier is recommended. Very long cables can be used between this preamplifier and the recorder without significant loss in signal. The recorder supplies the dc power required by preamplifiers for both channels.

The frequency response of the weighting networks provided in the main amplifier is shown in Figure 5. The A-, B-, and C-weighting characteristics of the sound-level meter permit one to make the usual weighted-level readings. The A-weighting network is particularly popular as a means for obtaining a single number for rating a noise. The D-weighting network has a decreasing response above 1 kHz. It has seen some use, but it is not specified by current standards. It permits one to make a quick estimate of the relative energy above 1 kHz by a comparison of the relative levels of the C-weighted response and the D-weighted response.

The standard sound-level-meter characteristics provided here make possible the measurement of a noise by approved techniques before it is recorded. This cannot be done directly on other tape recorders.

In addition to these sound-level-meter weighting characteristics, a flat, or uniform, response and NAB-weighting characteristics are provided. The NAB responses are those specified by the National Association of Broadcasters for tape recorders, and they have been designed for good results with speech and music signals. The responses, as shown in Figure 5, have rising characteristics at both high and low frequencies. Although such responses are generally helpful in obtaining a better dynamic range for speech and music, it is not usually desirable for recording industrial noises. As an alternate, therefore, the flat response, which is often called "constant-current" in magnetic-tape recording, has also been provided. It is this characteristic that should be used for recording most noise signals. This constant-current recording is not available on tape recorders designed for speech and music but only on instrumentation-type recorders.

On high-fidelity and professional types of magnetic-tape recorders designed for speech and music recording, the metering system monitors the signal level before any modification of the response characteristic is made. This modification, usually called pre-emphasis, affects a signal in a manner that depends on the spectrum of the signal. One cannot therefore be certain, without knowing much about the signal, what happens to the recorded signal. On such recorders it is possible for some types of signals to have the recording approach saturation of the tape even

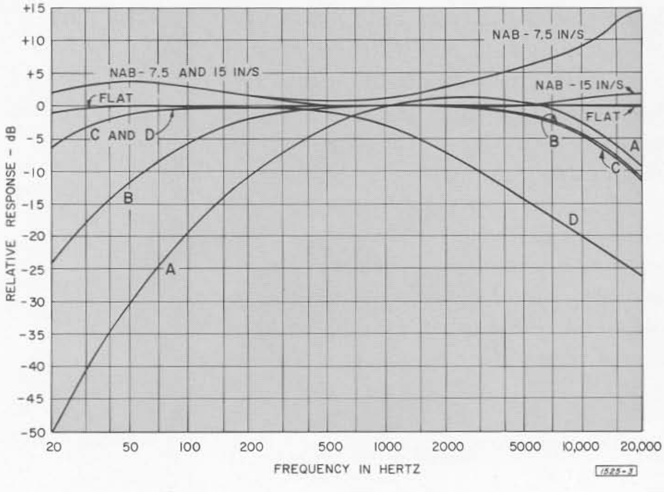

Figure 5. Frequency response of weighting networks.

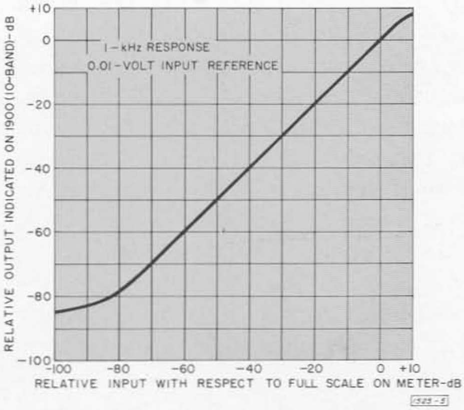

Figure 6. Amplitude-response characteristic of the complete instrument at 1000 Hz.

when the monitoring meter stays within the normal operating range. The Type 1525-A Data Recorder avoids this possibility, because its metering circuit is placed after all the weighting networks and just before the final stage driving the recording head.

In addition to supplying the meter monitoring circuit, the output of the main amplifier is applied to a peak-responding detector. When the output of this detector exceeds a preset value, which is near the point at which the recording-head current would saturate the magnetic tape, the output triggers a circuit that causes a light to turn on. It stays on until a reset button is pressed. This light functions as a monitor for impact-type sounds or vibrations, whose peak levels are not satisfactorily indicated by the relatively slow meter movement. It makes possible the selection of the proper amplifier-gain setting for such sounds without danger of overloaded and spoiled recordings.

Linearity

Linearity of response is an important characteristic when a tape recorder is used as a measurement device. The linearity at high levels is limited by magnetic saturation of the tape, and the apparent linearity at low levels is limited by tape noise. The results of some measurements of this linearity are shown in Figure 6. In this measurement the output of the playback signal was analyzed by the Type 1900-A Wave Analyzer,[3] set for a 10-Hz bandwidth. This reduced the effects of noise and eliminated the false improvement in apparent linearity that would result from including harmonics in the measured signal at high levels. The excellent linearity over a range of more than 70 dB is obvious from the plotted results.

Low-Frequency Response

The low-frequency response of the usual tape recorder designed for speech and music is not good below 40 or 50 Hz. In contrast with this performance, the response of this new tape recorder at the 7½-in/s, flat-recording condition is remarkably uniform down to 20 Hz with only a few dB drop in response at 15 Hz. This performance is important for many industrial noises that have maximum energy at these lower frequencies, and it increases the usefulness of the recorder for many applications in the field of vibration control, particularly where the ultimate goal is acoustic-noise control. Typical over-all response characteristics are shown in Figure 7.

The generally good response of this recorder can be illustrated in another way. The analyses of a test signal applied to the recorder and of the signal

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

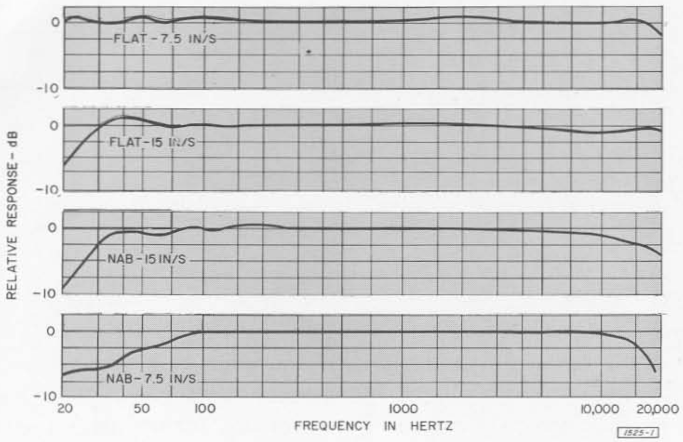


Figure 7. Typical overall frequency response characteristics.

played back from the recorded tape are shown in Figures 8 and 9. This test signal was an 8-kHz tone burst of 4 cycles duration repeated every 2.5 milliseconds. This test tone has a wide range of component frequencies as shown by the analyses, and the good frequency response is shown by the close similarity of the before-and-after analyses.

Attenuator

The accurate step attenuator in the main amplifier has several advantages. It makes possible recordings over a wide range of levels with a good dynamic range. This feature is not provided on most tape recorders, and it is essential for handling the wide range of signal levels encountered in noise-control work. It also makes possible the use of a calibrating signal (such as that from a Type 1552-B Sound-Level Calibrator driving a measurement microphone or a Type 1557-A Vibration Calibrator driving a pickup) that is widely different in level from the signal to be studied. The difference in the settings of the step attenuator for the calibrating and the unknown signals is then used to relate the playback levels of the two signals. Expressed

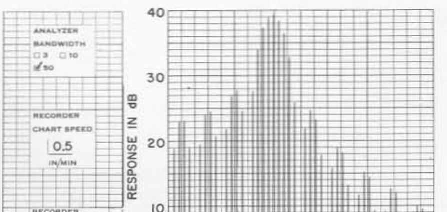


Figure 8. Analysis of an 8-kHz tone burst applied to the tape recorder. The tone burst was of 4 cycles duration repeated every 2.5 milliseconds.

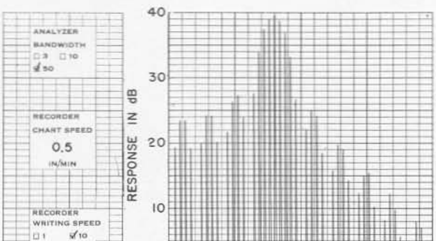


Figure 9. Analysis upon playback of the tone burst of Figure 8.

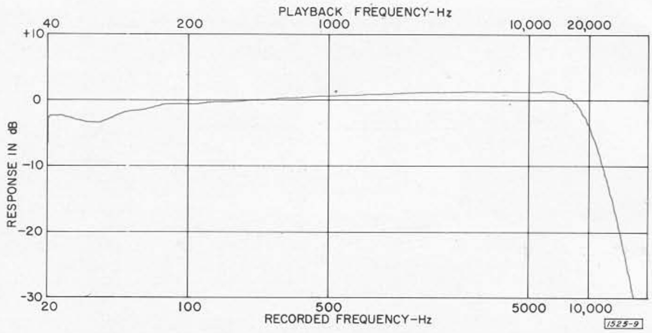


Figure 10. Over-all frequency response of the recorder when the signal is recorded at 7.5 in/s and played back at 15 in/s.

in another way, the accurate step attenuator is used to shift a calibrating signal to any convenient level before recording. This procedure is an essential element for accurate measurements on a recorded signal.

Tape Loops

In many applications of a recorder in noise-control studies, a section of the tape is cut out and formed into a loop for continuous playback and analysis by a sweeping analyzer. The TYPE 1525-A Data Recorder includes tape loop guides that simplify the playback of these loops. (See Figure 2.)

Because of the type of noise usually encountered, the bandwidths used for analysis, and the averaging times employed, a sample duration of 2 to 3 seconds is convenient for acoustic and vibration studies. A length of loop that gives this duration is readily used on the TYPE 1525-A Data Recorder.

Time and Frequency Translation

The response characteristics given above apply for recording and playback at the same speed. If the frequency and time scales are modified by a change in speed between recording and playback, the apparent frequency response is somewhat different from that encountered in normal use, particularly at the frequency extremes.

Graphs of measured responses are shown in Figures 10 and 11. They illus-

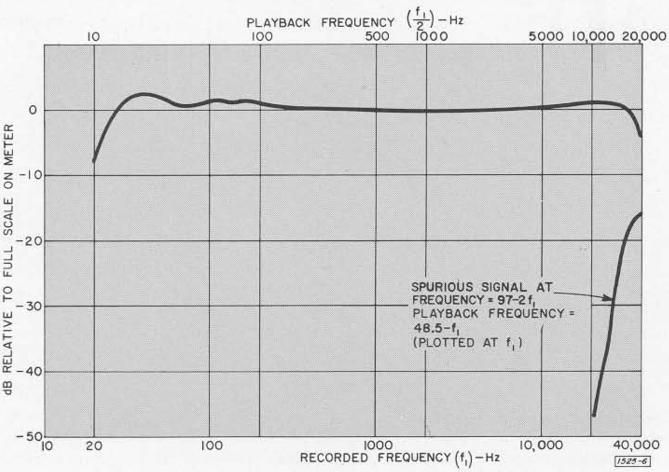


Figure 11. Over-all frequency response of the recorder when the signal is recorded at 15 in/s and played back at 7.5 in/s.

trate the fact that the playback characteristic is the dominating one. Thus, for example, a 15-kHz tone recorded at 7.5 in/s becomes at 15 in/s a 30-kHz tone, which is beyond the normal playback range, and the apparent output is low at that frequency.

The inverse effect also occurs when the process is reversed. A tone at 30 kHz recorded at 15 in/s becomes at 7.5 in/s a 15-kHz tone, which is within the reproducing range of the recorder, and satisfactory response is obtained. By this technique the apparent frequency range of the recorder for some measurement purposes can be doubled.

The extension of the frequency range is accompanied by another effect, however, which limits its usability. In the magnetic material on the tape, the mixing of the bias field and that from the incoming signal results in some spurious signals. The most important of these is at a frequency of f_B-$2f_1$, where f_B is the bias frequency (about 95 kHz for the TYPE 1525-A) and f_1 is the frequency of the signal being recorded. For frequencies up to 20 kHz this spurious component is at 55 kHz or higher, and no significant playback output of the spurious component results. When a 40-kHz tone is recorded, the spurious frequency is at 15 kHz, and a significant amount of the spurious signal is present as shown on the graph. The relative amount of this spurious signal drops rapidly with signal level. The value shown was measured at the normal maximum input level. When the input level is lowered 10 dB, the spurious signal drops about 20 dB. In addition, as shown on the graph, the level of the spurious signal drops rapidly as the frequency is reduced, becoming negligible at a recording frequency of 20 kHz.

This behavior shows the main reason for the use of a relatively high-frequency bias. For normal recording techniques, the bias frequency should be significantly higher than three times the highest signal frequency to be used if these spurious signals are to be avoided. For two-to-one frequency-scaling, the ratio of bias frequency to signal frequency should be greater than 4:1. But if this limitation of spurious signals is recognized, one can use this recorder for some applications at frequencies well above 20 kHz.

Incidentally, a number of hi-fi recorders use a bias-oscillator frequency in the range of 35 to 50 kHz. If an attempt is made to use such recorders at frequencies above 10 kHz, the spurious responses may be troublesome. Any attempt at frequency scaling downward would be even more subject to difficulty with spurious signals.

Modulation Noise

One limitation on the performance of a tape recorder is a form of noise known as modulation noise, which appears when a signal is recorded. The modulation-noise energy is concentrated in the frequency region immediately adjacent to that of the applied signal frequency components. The effect can be measured by an analysis of the playback of a recorded pure tone. The results of such a measurement on a TYPE 1525-A Data Recorder are shown in Figure 12. The level of the recorded 1-kHz tone is displayed as 0 dB, and the accompanying noise measured in the 10-Hz passband of the TYPE 1900-A Wave Analyzer is shown to peak in the vicinity of 1 kHz and to be more than 60 dB lower. A similar

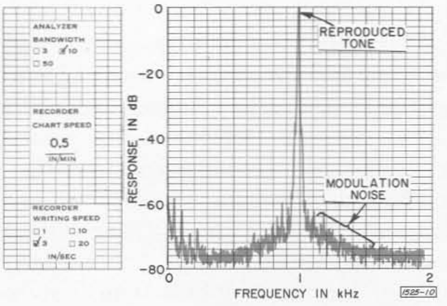


Figure 12. An analysis of the playback signal of a recorded 1000-Hz tone. The analysis on the Type 1900-A Wave Analyzer with a 10-Hz bandwidth shows the relative levels of the tone and the noise added in the recording and playback process. Note that the ordinate scale covers a range of 80 dB.

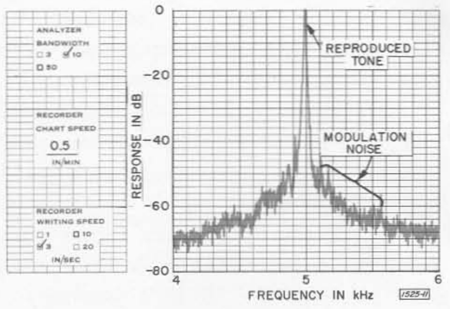


Figure 13. The same as Figure 12 except that the recorded signal was a 5000-Hz tone.

measurement for a 5-kHz tone is shown in Figure 13. The modulation noise measured here is significantly lower than that obtained on many other recorders, and it is so low that it does not ordinarily limit the performance of the recorder for acoustical and vibration measurements.

Modulation noise is dependent on the tape used, the transport tape drive and support mechanism, and the condition of the tape surface and heads. Use of high-quality tape, such as Ampex Type 641, careful control to avoid accumulation of dirt particles on the tape surface, and good maintenance of the transport mechanism are essential for obtaining this low level of modulation noise. Fortunately, for most applications in acoustical and vibration measurements, reasonable care in use, storage, and handling is adequate for satisfactory results.

SUMMARY

Provision of sound-level-meter characteristics, a variety of weighting networks, an accurate step attenuator, an amplifier with high gain and high input impedance, a transient overload indicator, two-channel recording, simultaneous playback on recording, good low-frequency response, and tape loop guides make the versatile TYPE 1525-A Data Recorder particularly well suited for acoustic-noise and vibration measurements.

— ARNOLD PETERSON

The TYPE 1525-A Data Recorder is based on one developed for the General Radio Company by G. W. Kamperman, D. Noiseux, and R. V. Jones of the firm of Bolt Beranek and Newman Inc.

Note: A brief biography of Arnold Peterson, author of the foregoing article, appeared in the August 1966 issue of the *Experimenter*. — Editor

SPECIFICATIONS

Channels: 2 channels with separate record and playback amplifiers and separate channel erase.

Measurement Range (Input Level): 10 μV to 1 V on channel #1 (40 to 140 dB sound-pressure level for microphone sensitivity of -66 dB re 1 V/μbar). About 0.7 V on channel #2 for full-scale meter indication. (For high sensitivity, channel #2 can be driven by the output of a separate sound-level meter.)

Frequency Response:

At 15 in/s (38.1 cm/s)

Constant Current: ±2 dB, 50 to 15,000 Hz. $+2$, -4 dB, 30 to 18,000 Hz.

NAB equalization: ±2 dB, 50 to 15,000 Hz.

At 7½ in/s (19.05 cm/s)

Constant Current: ±2 dB, 20 to 10,000 Hz. $+2$, -4 dB, 15 to 16,000 Hz.

NAB equalization: +2, −4 dB, 50 to 15,000 Hz.

Signal-to-Noise Ratio:
NAB equalization: Over 54 dB below 2% distortion point as measured according to NAB standard (A weighting).
Constant Current: Over 45 dB below 2% distortion point for noise band from 20 to 15,000 Hz (over 65 dB for octave band at 1 kHz) with input to channel #1 of greater than 10 mV.

Input Impedance: *Channel #1:* approx 20 pF shunted by 400 MΩ.
Channel #2: >100 kΩ.

Weighting Characteristics: NAB and constant current for both playback amplifiers; NAB constant current, and A, B, and C weighting (standard sound-level-meter characteristic) for record channel #2.

Monitoring: Electronic voltmeter with 16-dB range and sound-level-meter ballistic characteristics, switchable to monitor record or playback level on either channel. Peak monitor on record channel #1.

Tape Speeds: 15 in/s (38.1 cm/s).
7½ in/s (19.05 cm/s).

Tape: ¼ inch, professional quality.

Reel Size: 7-in reel (maximum).

Flutter and Wow: Below 0.2% rms.

Bias and Erase Frequency: 95 kHz nominal; separate erase for each channel.

Power Required: 105 to 125 V, 60 Hz, 135 W.

Accessories Supplied: Guides for tape loop; Type 1560-P99 Adaptor Cable; CAP-22 Power Cord; reel of tape; rack-mount hardware; maintenance kit.

Accessories Available: Type 1560-P5 Microphone and Type 1560-P34 Tripod and Extension Cable for sound measurements and recording. Type 1560-P40K Preamplifier and Microphone Set for sound measurements and recording at levels below 50 dB where the best signal-to-noise ratio must be maintained (the recorder supplies the necessary power to operate a Type 1560-P40 Preamplifier). For sound and noise analysis, Type 1900-A Wave Analyzer, Type 1564-A Sound and Vibration Analyzer, Type 1568-A Wave Analyzer, Type 1558 Octave-Band Noise Analyzers.

Dimensions: Portable model — width 21, height 16, depth 9 inches (540, 410, 230 mm); rack model — width 19, height 14, depth behind panel 7 inches (485, 355, 180 mm), over-all.

Net Weight: Portable model, 53 lb (25 kg); rack model, 50 lb (23 kg).

Shipping Weight: Portable model, 60 lb (28 kg); rack model, 57 lb (26 kg).

Catalog Number	*Description*	*Price in USA*
1525-9701	**Type 1525-A Data Recorder**	**$1995.00**

A STANDARD EARPHONE COUPLER FOR FIELD CALIBRATION OF AUDIOMETERS

There is much evidence that frequent calibration of audiometers in common use for industrial audiometry is a practical necessity. Cudworth[1] has emphasized the need for maintaining proper calibration of audiometers and has proposed a field calibration and audiometer evaluation system using the sound-level meter. Reports on audiometer performance presented at various hearing and noise symposiums of industrial hygiene and safety organizations have urged periodic calibration.

A published study [2] reports on the performance of five audiometers over a period of 42 months. At the start of the test period all were adjusted by the manufacturer to be within specifications. They were calibrated three times a month at first and then once a month unless the operator suspected trouble. A record was kept of calibration results, and a note made each time an audiometer performed outside its specification. The results are significant:

Defect in Performance	*Number of Occurrences*
Sound pressure at or beyond tolerance limits	63
Faulty earphone performance	11
Frequency outside limits	10
Excessive harmonic distortion	13
Extraneous instrument noise	19

References are listed on page 20.

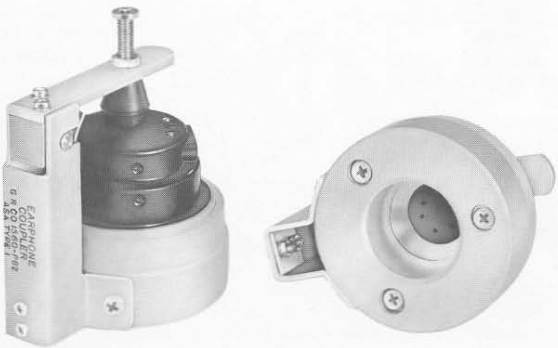


Figure 1. (Right) The Type 1560–P82 Earphone Coupler; (center) the coupler with earphone installed; (left) coupler with earphone, installed on sound-level meter.

Since incorrect sound pressure is the chief offender, it is apparent that a readily available, simple, convenient, and reliable acoustical calibration system for audiometers is needed. Since many industrial hygienists are responsible for noise surveys as well as for industrial audiometry, the sound-level meter or other sound-measuring instrument is usually readily at hand. The missing link has been a readily available standard earphone coupler.

Standard Earphone Coupler

An earphone coupler is a device for coupling an earphone to a standard-pressure microphone for the purpose of determining the acoustic characteristics of the earphone. It provides a tightly sealed chamber of specified volume to simulate the compliance of the human ear. An earphone coupler and a sound-level meter are all that is needed to evaluate the over-all performance of an audiometer.

There are two types of standard earphone coupler in use today: (1) the NBS Type 9A Coupler, specified in ASA Z24.5–1951 [3] and in ASA Z24.12–1952 [4]; and (2) the ASA Type 1 Coupler, specified in ASA Z24.9–1949 [5] and in ASA Z24.13–1953.[6]

The Type 1 Coupler is the newer design of the two and, because of its shape, has been found easier to use than the 9A for field calibrations [1]. According to the standards [3,4], there is but slight measurable difference between results obtained with the two couplers. This is confirmed by measurements reported by Morrical et al [7].

The Type 1560-P82 and Type 1560-P81 Earphone Couplers

Since the Type 1, specified by ASA Standard, is easier to use, we have chosen to manufacture a version of this coupler.

This new coupler is available in two models: (1) The Type 1560-P82 for use on 15/16-inch-diameter, piezoelectric, ceramic, sound-level-meter microphones, such as the GR1560-P5 and 1560-P6 Microphones (it will also fit Type L Laboratory Standard Microphones like the WE Type 640AA); and

(2) the TYPE 1560-P81 for use on the older 1⅛-inch-diameter, piezoelectric, ceramic, sound-level-meter microphones, like the GR1560-P3 or 1560-P4.

The TYPE 1560-P82 Earphone Coupler, shown in Figure 1, is a precision-machined, aluminum device designed to provide the proper acoustic coupling medium between a microphone and an earphone. The coupler also includes a toggle clamp to apply the prescribed force of 500 grams to the earphone under test. The clamp has an adjustable pressure foot to accommodate earphones of various heights and to maintain the lever arm in the horizontal position necessary for the proper applied force. The top part of each coupler model is designed to accept the Telephonics TDH-39 Earphone, which seems to be most commonly used on American audiometers. The Permoflux PDR series and Western Electric 705-A earphones will also fit these couplers.

In appearance, the TYPE 1560-P81 Earphone Coupler is the same as the TYPE 1560-P82. The cross-section views (Figure 2) show how the two GR couplers differ from each other and from the prototype ASA Type 1 Standard Earphone Coupler. The TYPE 1560-P82 Coupler has a small step added to provide positive and accurate location of the microphone.

The TYPE 1560-P81 Coupler, in addition to the small locating step, has a larger diameter recess to accept the 1⅛-inch-diameter microphones, and the height of the acoustic coupling cavity is reduced to maintain the required 6-cm^3 coupling volume with the large microphone and its larger effective volume.

These modifications add measurably to the convenience of use but in no way impair the acoustical performance.

New Coupler Performance

Measurements made in the General Radio Laboratories verify that there is only a slight difference in the response of a TDH-39 Earphone as measured with an NBS 9A Coupler and with the ASA Type 1 Coupler. Our measurements also show that the differences are negligible when the earphone response is measured in the NBS 9A Coupler and the TYPE 1560-P82 or TYPE 1560-P81 Earphone Coupler. Figure 3 shows the responses of two Telephonics TDH-39 Earphones, each

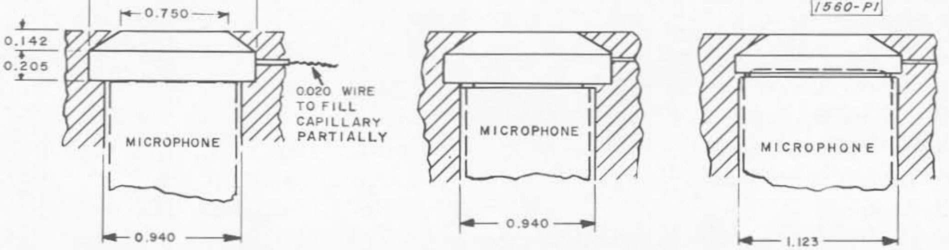
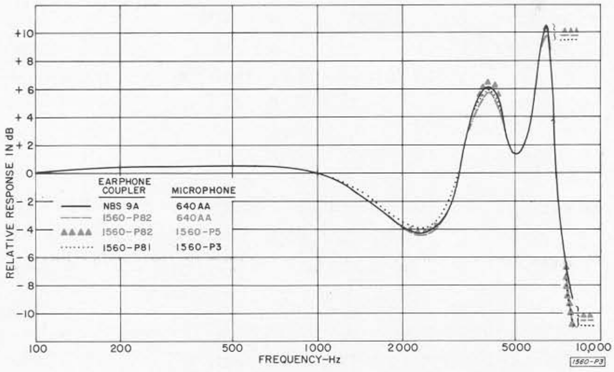


Figure 2. Standard Earphone Couplers. (a) ASA Type 1 (Z 24.9–1949); (b) Type 1560–P82 ASA Type 1 (modified to accept 15/16-inch diameter, piezoelectric, ceramic, sound-level-meter microphone); (c) Type 1560-P81 ASA Type 1 (modified to accept 1⅛-inch diameter, piezoelectric, ceramic, sound-level-meter microphone). Note that the General Radio couplers make possible accurate and repeatable positioning of the microphone.



(a)

(b)

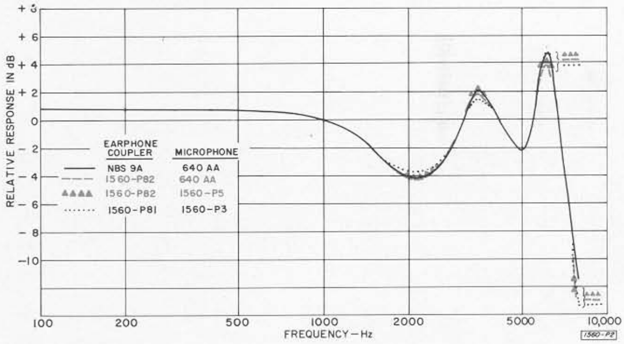


Figure 3. Frequency response. (a) Telephonics TDH-39 Earphone No. 3; (b) Telephonics TDH-39 Earphone No. 101.

measured in the NBS 9A Coupler with a Western Electric 640AA Microphone, then in the Type 1560-P82 Coupler with the 640AA and with a Type 1560-P5 Microphone, and finally in the Type 1560-P81 Coupler with a Type 1560-P3 Microphone. Earphone No. 3 has had intermittent use over four or five years; earphone No. 101 is a new unit. We have measured a number of other earphones in this manner, and, in each case, there has been excellent agreement between the responses as measured in the various couplers.

Application

To use a coupler to evaluate an audiometer frequency characteristic, one measures the sound-pressure level developed in the coupler by the audiometer earphone. This is usually done at a hearing-level dial setting of 60 dB. If the audiometer is calibrated in accordance with Z24.5–1951 [3], the sound-pressure levels that should be developed in the coupler are derived from Table 2 in that standard, which lists the threshold (0 dB hearing level) sound-pressure levels at each audiometric frequency for the Western Electric Type 705-A Earphone. The standard then states that, for any other type or configuration of earphone, the threshold levels must be determined by loudness balance tests against a Type 705-A Earphone. This work has been done and the data published [8] for the Telephonics TDH-39 Earphone with MX41/AR cushion. The data in lines 1 and 3 of Table I, below, are derived from the data of Reference 8.

TABLE I

Sound-pressure levels (dB re 2 × 10^{-4} μbar) at audiometric frequencies in the Type 1560-P82 or Type 1560-P81 Earphone Coupler corresponding to a 60-dB hearing level for a TDH-39 Earphone with MX41/AR cushion as read on typical sound-level meters.

Coupler Type No.	*Sound-Level Meter*		*Audiometer Reference Threshold*	*Frequency-Hz*									
	Type	*Weighting Position*		*125*	*250*	*500*	*1000*	*1500*	*2000*	*3000*	*4000*	*6000*	*8000*
1560-P82	1551-C*	C	ASA 1951	112	99.5	84	77	78	78	74.5	71	75	83 **dB**
1560-P82	1551-C*	C	ISO 1964	105.5	85	71	67	67	68.5	69.5	67	70	69 **dB**
1560-P81	1551-B	20 kc	ASA 1951	112	99.5	84	77	78.5	79	77.5	76	79	80 **dB**
1560-P81	1551-B	20 kc	ISO 1964	105.5	85	71	67	67.5	70	72	72	74	66 **dB**

* Also Type 1565-A Sound-Level Meter

New audiometers may be calibrated to the new international audiometric zero.[9,10] Data for the TDH-39 Earphone, though not in the international standard at present, are available[11], and, apparently, some efforts are being made to add the TDH-39 data to the international standard.[12,13] The data for lines 2 and 4 of Table I are derived from the data of Reference 12.

Table I lists the sound-level-meter readings that should be produced by an audiometer set to the 60-dB hearing level and coupled to a typical sound-level meter by a Type 1560-P82 or Type 1560-P81 Earphone Coupler.

In place of the sound-level meters listed in the table, the earlier Type 1551-A Sound-Level Meter or one of the Type 1558 Octave-Band Analyzers can be used, when equipped with the appropriate microphone.

Most industrial hygienists and others concerned with noise measurement as well as audiometry have a sound-level meter or an octave-band analyzer; one of these couplers is the only additional equipment necessary to check the acoustic output of an audiometer.

For those who wish to make complete acoustical and electrical tests on an audiometer, the Type 1564-A Sound and Vibration Analyzer, with a Type 1560-P40 Preamplifier, a Type 1560-P5 Microphone, a Type 1560-P82 Earphone Coupler, and an audio oscillator, is recommended.

— E. E. Gross

Figure 4. Type 1560-P82 Earphone Coupler used on Type 1565-A Sound-Level Meter.

Ervin E. Gross received his B.S.E.E. degree from Northeastern University in 1936 and has been with General Radio ever since. As a development engineer in General Radio's Audio Group, he has specialized in the design of sound and vibration measuring instruments. He is a Senior Member of the IEEE and a member of the Acoustical Society of America and has been active in standards work for both organizations.

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10. ISO Technical Communication 43—Acoustic, No. 554, "A Standard Reference Zero for the Calibration of Pure Tone Audiometers," (1964). (Note: See Reference 9).
11. TABLE — "Recommended Reference Equivalent Threshold Sound Pressure Levels in the 9A Coupler," Proposed for the International Organization of Standards by the National Bureau of Standards, Washington, USA, 1963.
12. C. G. Rice and R. R. A. Coles, "Normal Threshold of Hearing for Pure Tones by Earphone Listening with a Self-Recording Audiometric Technique," *Journal of Acoustical Society of America*, 39, 1185–1187 (L) (1966).
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16. A. Glorig, *Noise and Your Ear*, Grone and Stratton, New York (1958).

SPECIFICATIONS

Type Coupler: ASA Type 1.

Volume: 6 cm³ including equivalent volume of microphone (Type 1560-P5 Microphone for Type 1560-P82 Coupler; Type 1560-P3 Microphone for the Type 1560-P81 Coupler).

Axial Holding Force: 500 grams.

Frequency Range: 100 Hz to 8000 Hz; ±1 dB from 100 Hz to 6000 Hz, increasing to ±3 dB at 8000 Hz (with corrections for pressure response of microphone).

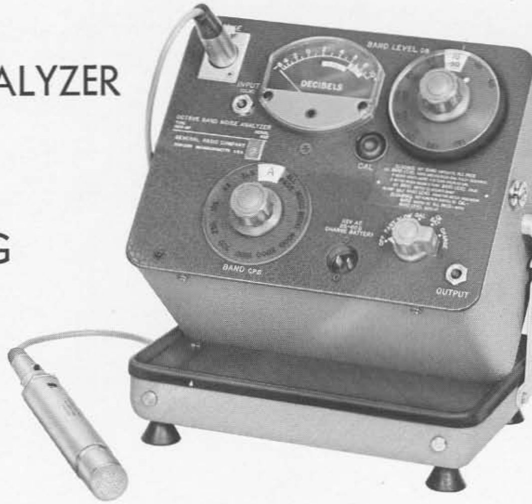
Dimensions: Coupler, diameter 2¼ by height 1$\frac{1}{16}$ inches (5.7, 2.5 mm); over-all, width 2¼, length 3, height 3 inches (5.7, 7.7, 7.7 mm).

Net Weight: 8 oz (230 g).

Shipping Weight: 2 lb (1 kg).

Catalog Number	*Description*	*Price in USA*
1560-9682	**Type 1560-P82 Earphone Coupler**	**$30.00**
1560-9681	**Type 1560-P81 Earphone Coupler**	**30.00**

OCTAVE-BAND ANALYZER WITH A-WEIGHTING



An A-weighted sound level is now often used to rate similar types of noise, because a good correlation has been found for many noises between A-weighted sound levels and loudness, speech interference, and "noisiness."

A-weighted sound level is often measured in lieu of or as a supplement to an octave-band analysis. To make both measurements, the noise sleuth has needed two instruments: a sound-level meter and an octave-band analyzer.

The TYPE 1558-BP Octave-Band Analyzer provides both capabilities in a single instrument. It is identical to the Type 1558-AP Octave-Band Noise Analyzer[1] except for the inclusion of an A-weighting network. A-weighting is selected by the BAND switch, which also selects ten octave bands at frequencies specified by ASA S1.6–1960 Preferred Frequencies for Acoustical Measurements. Both this instrument and the TYPE 1558-AP Octave-Band Noise Analyzer conform with the requirements of the new *American Standard Specification for Octave, Half-Octave, and Third-Octave-Band Filter Sets S1.11–1966 for Type E, Class II octave-band filters.*

The new analyzer is shown in Figure 1 together with the TYPE 1560-P40 Preamplifier and Microphone. This combination measures levels ranging from 24 dB to 150 dB, the same range as the TYPE 1551-C Sound-Level Meter.

An existing TYPE 1558-AP Octave-Band Noise Analyzer can be converted to a −BP model. For details, write to our Service Department.

[1] W. R. Kundert, "New, Compact, Octave-Band Analyzer," *General Radio Experimenter*, October 1962.

Catalog Number	*Description*	*Price in USA*
1558-9890	**Type 1558-BP Octave-Band Noise Analyzer,** Portable Model	**$850.00**
1558-9848	**Type 1558-BP Octave-Band Noise Analyzer,** Rack-Adapted Model	**850.00**
1560-9520	**Type 1560-P40K Preamplifier and Microphone Set**	**251.00**

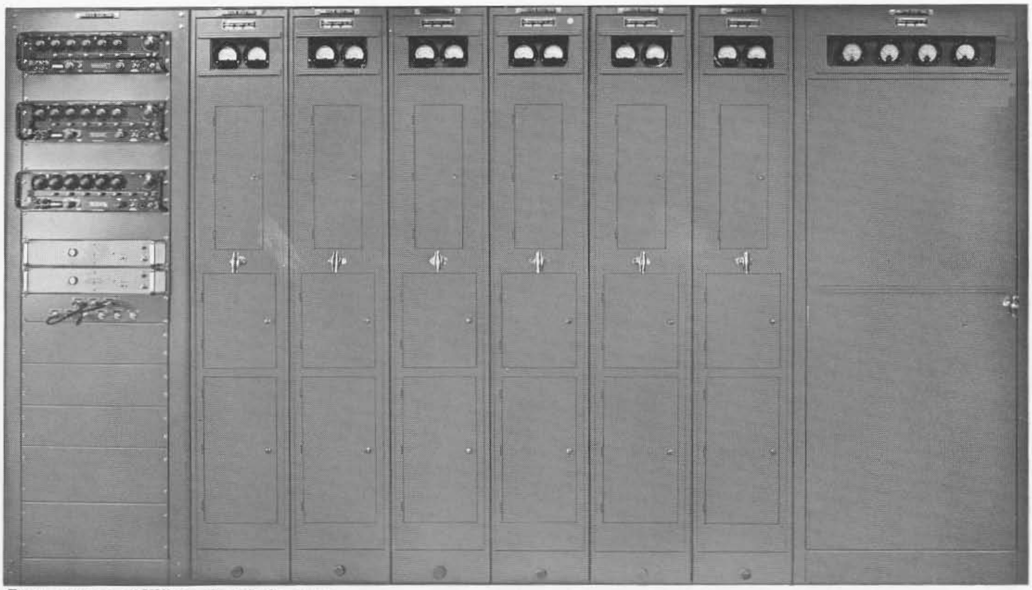


Photo courtesy of Wilcox Electric Company

Wilcox Type 96 transmitter, with GR frequency synthesizers installed in far-left rack.

VERSATILE TRANSMITTER USES GR SYNTHESIZERS

Those whose dark business it is to jam radio transmissions are not likely to welcome the increasing use of frequency synthesizers to control transmitter frequencies. The engineer at the transmitter has, in a typical synthesizer, the equivalent of several million crystals and can jump from one hertz to the next with the abandon of a grasshopper.

The Wilcox Electric transmitter system shown above was designed and built by Wilcox Electric for a government anxious to maintain communications in the face of interference, intentional or otherwise. Each of six transmitters in the system can be fired up instantly on any one or two of 32 preset channels. Primary frequency control is by crystal oscillator, but three GR frequency synthesizers stand ready to take over if a crystal fails or if the operator wants to set up new frequencies without grinding new crystals. The total number of channels that can be set by means of the synthesizer is high enough to make any would-be jammers take up another profession.

A SIMPLE WAY TO SYNCHRONIZE MAGNETIC TAPE WITH OSCILLOSCOPE TRACE

During some recent investigations of transient sound pulses in our Engineering Department, it was necessary to synchronize the starting of an oscilloscope trace with the position of the pulse on a magnetic recording tape. The TYPE 1536-A Photoelectric Pickoff [1] proved very well suited to this task.

[1] "Using a Photocell Where it Counts," *General Radio Experimenter*, October 1962.

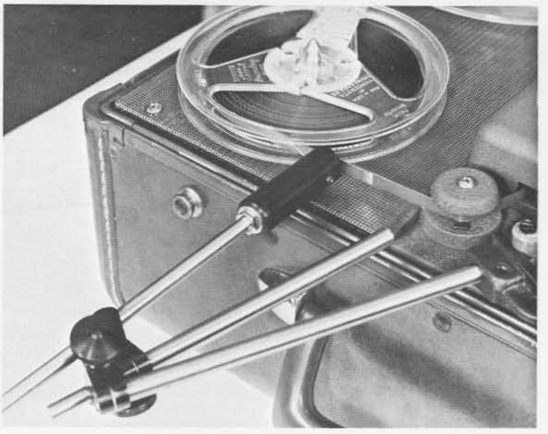


Figure 1. Photoelectric pickoff shown mounted in tape recorder. Note piece of reflective foil attached to tape.

The pulse was located on the magnetic tape by trial-and-error to within an inch or two, and a bit of aluminum foil was fastened to the back of the tape at this spot by cellophane tape. The photoelectric pickoff was clamped to the recorder and positioned so that the light beam shone across the width of the tape. The passage of the aluminum foil through the light beam initiated the oscilloscope trace.

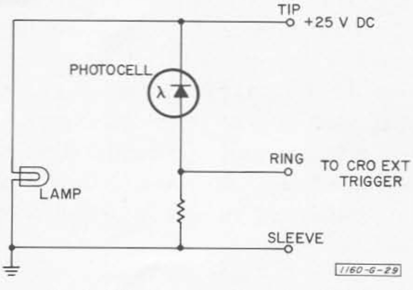


Figure 2. Electrical connections to pickoff plug.

Figure 1 shows the photoelectric pickoff mounted on the tape recorder. A piece of aluminum tubing was screwed to the chassis to provide a convenient gripping surface for the mounting clamp. For final positioning of the pickoff, the mounting rods were adjusted until the sound pulse occurred coincident with the start of the trace. If greater precision were required, the photoelectric pickoff could have been mounted on a micrometer carriage.

Electrical connections are shown in Figure 2. It is necessary only to supply dc voltage of approximately +25 volts for the exciter lamp and to connect the ring of the TYPE 1536-A plug to the oscilloscope external trigger terminal. The scope itself, of course, should be adjusted for single-sweep operation.

— GORDON R. PARTRIDGE

NEW NBS LABORATORIES TO BE DEDICATED

The new laboratory complex of the U. S. Department of Commerce's National Bureau of Standards, located in Gaithersburg, Maryland, will be dedicated on November 15. Secretary of Commerce John T. Connor will head the list of dignitaries participating in the ceremonies. *(Cont'd)*

the Experimenter

GENERAL RADIO COMPANY
WEST CONCORD,

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NBS LABORATORIES *(Continued)*

In conjunction with the dedication, the Secretary is sponsoring a two-day Symposium on Technology and World Trade, to be held November 16 and 17 on the NBS grounds. About 500 distinguished experts from all over the world have been invited to participate.

Among the important new facilities at Gaithersburg are a nuclear research reactor and a powerful linear electron accelerator (linac). In the field of nuclear physics, in the nuclear generation of electric power, and in the rapidly expanding use of radiation to process materials and products in industry, one of the most urgent needs has been for measurements and standards. The reactor and linac will enable NBS to fulfill its important functions in this area.

GENERAL RADIO COMPANY
WEST CONCORD, MASSACHUSETTS 01781

THE GENERAL RADIO

Experimenter

REVOLUTION IN CAPACITOR TESTING

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REPRESENTATIVES IN PRINCIPAL OVERSEAS COUNTRIES

AUTOMATIC CAPACITOR-TESTING SYSTEMS

The inspection bottleneck was broken by GR's 1680-A Automatic Capacitance Bridge Assembly, announced in 1964. Many users who formerly tested capacitors at a rate of ten a minute at best are now testing at rates better than 120 a minute! In this issue we describe some of the complex automatic testing systems designed around this bridge and two new accessories, the 1781 Digital Limit Comparator and the 1770 Scanner System.

The increasingly demanding requirements for extreme reliability in electronic equipment, both military and commercial, have heightened the need for instruments for the rapid and accurate testing of the components that make up this equipment. General Radio has for many years manufactured instruments for component testing — impedance bridges, capacitance and inductance bridges, impedance comparators, etc — most of which were designed for use by skilled operators in a laboratory. For on-line testing of high-volume production lots, manually operated instruments are no longer adequate; automation is a must.

General Radio entered the field of automatic component testing in 1964 with the introduction of the 1680-A Automatic Capacitance Bridge Assembly.[1] This instrument measures the capacitance and loss of a component in a half second or less at the push of a button (or at electrical command) and presents the answer in both a digital display and an electrical output. The widespread acceptance of this device has led to a demand for complete measurement systems built around the bridge and for additional accessory instruments, some to simplify the automatic connection of components to the bridge and others to make use of the prodigious amounts of data that can now be obtained so rapidly. In addition to providing many accessory instruments, we are now prepared to provide complete measurement systems.

MEASUREMENT SYSTEM

"System" is an often-used word whose meaning varies with context. At General Radio we define a system as a collection of instruments — of which some may be standard, some modified, some specially designed, some purchased — assembled to solve a specific measurement problem.

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

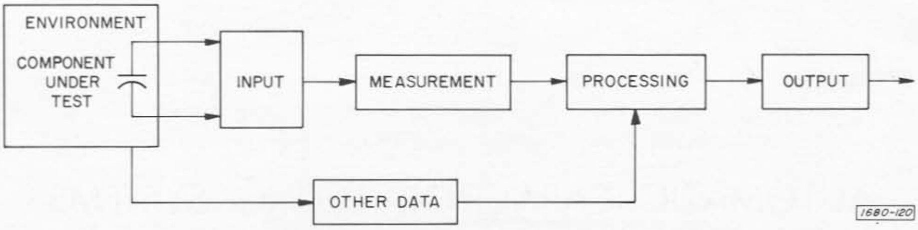

Figure 1. Block diagram of a component-testing system.

In this definition we recognize that a system is more than a group of our standard catalog instruments and a few patch cords. The grouping must be carefully thought out, both from an electrical and a mechanical standpoint. Interface problems must be identified and solved. Some instruments may have to be adapted to the requirements of the system. Others may have to be purchased or designed. Special cables, racks, and consoles may be necessary, along with special operating and maintenance instructions. In short, the system must be thoroughly engineered; the whole must be greater than the sum of the parts.

Since most of these system-design problems are beyond the inclination of many of our customers, we volunteer to do the job for them by offering custom-engineered measurement systems. Before we look at some examples of these systems, let's discuss a component-testing system in general.

GENERAL BLOCK DIAGRAM

Figure 1 shows a block diagram of an idealized component-testing system. The component or components are conditioned by the desired *environment* equipment, which may take the form of a temperature chamber, a vibration table, voltage-soaking equipment, etc. The next block includes component handlers, positioners, scanners, and other *input* devices. The heart (or brains) of the system is the *measurement* equipment — in this discussion the 1680-A Capacitance Bridge. *Processing* equipment operates on the information received from the measurement equipment to put it in a more useful form. Examples of processing equipment are digital-to-analog converters, digital limit comparators, and parallel-to-serial converters. *Output* equipment includes analog recorders, digital printers, and tape and card punches. The remaining block, *other data*, includes serial-number generators, time-code generators, digital thermometers, and other devices that generate supplementary information. Such information is usually fed into the processing equipment along with the measurement data.

FILLING IN THE BLOCKS

Measurement

The measurement instrument central to the systems discussed in this article is the 1680-A Automatic Capacitance Bridge Assembly.[1] When a capacitor is connected to the bridge, the instrument automatically selects the proper range, achieves balance, and presents the

[1] *Ibid.*

measured capacitance and either dissipation factor or conductance on an in-line digital readout, complete with decimal points and units. All this information is also presented in binary-coded decimal form (1-2-4-2 BCD) for use by printers or other data-handling equipment. The entire balance operation consumes a half second or less.

Three switch-selected generator frequencies are available: 120, 400, and 1000 Hz. Capacitance range is 100 pF (full-scale) to 100 μF at 400 and 1000 Hz (resolution is 0.01 pF) and 1 μF (full-scale) to 1000 μF with a 120-Hz signal. Dissipation-factor range is 0.0001 to 1.0, and the bridge will measure parallel conductance from 0.1 nanomho to 1 mho at 400 and 1000 Hz, from 1 micromho to 1.0 mho at 120 Hz.

Basic accuracy of capacitance measurement is $\pm 0.1\%$ of reading $\pm 0.01\%$ of full scale. Accuracy of frequencies supplied by the oscillator is $\pm 1\%$.

The bridge features several operating modes to accommodate a wide range of possible applications. In the TRACK CONT mode, for example, it continuously follows variations in a capacitor under test, permitting automatic recording of the effects of temperature or other environmental conditions. In the TRACK SAMPLED mode, the bridge follows variations but yields data only on command.

In addition to measuring capacitors, this instrument can be used to measure any parameter that can be expressed in terms of an equivalent parallel capacitance and conductance. Thus parameters of dielectric materials, cables, thin-film circuits, inductors (negative capacitors), and resistors can be measured.

Output

There are many other ways of using the measurement results besides looking at them or writing them down. Figure 2 shows the bridge feeding data to the 1137-A Data Printer. This is the simplest means of obtaining a permanent printed record of the measurements.

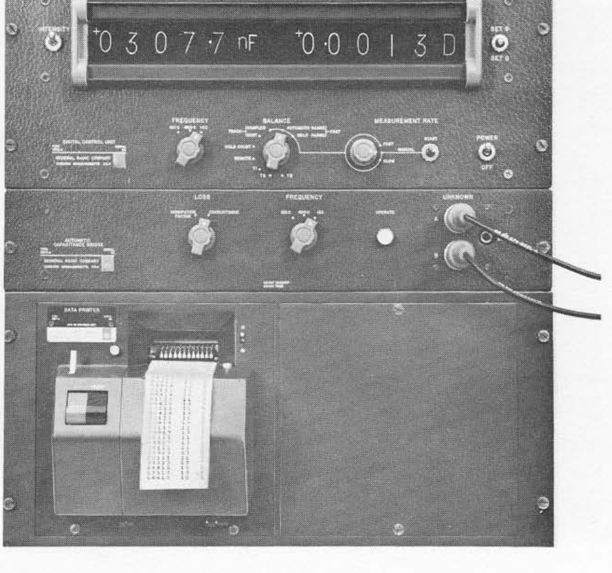

Figure 2. Automatic capacitance bridge and data printer.

The 1510-A Digital-to-Graphic Recording Assembly, shown with the bridge in Figure 3, is convenient for plotting changes in the value of a given component. The digital-data output of the bridge, after conversion to an analog current (or voltage) by a high-speed digital-to-analog converter, is plotted by a graphic recorder. By selection of the proper digits for conversion, a precise zero suppression can be obtained so that very small changes in the value of a component can be easily plotted. This combination is used extensively in environmental tests on capacitors, as, for example, in the measurement of temperature coefficients.

One of the most popular means of storing data for further statistical analysis is the punched card. Through a parallel-to-serial converter the bridge can be connected to a card punch for this purpose.

Other forms of data recording for future processing by a computer include punched tape and magnetic tape. Inquiries are invited for systems including equipment for tape recording.

Another class of output equipment operates on the component itself rather than on the measurement data. This equipment can be combined with the input equipment to sort the components into tolerance bins, for example. While General Radio provides the proper processing equipment (see below), we do not supply the mechanical handling equipment, which is readily obtainable from several manufacturers.

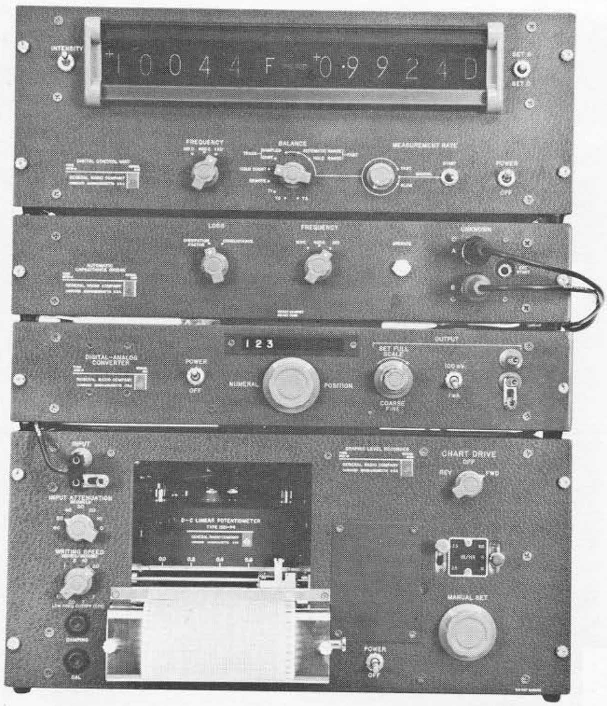

Figure 3. Automatic capacitance bridge and digital-to-graphic recording assembly.

Processing

Processing equipment takes several forms. The 1136-A Digital-to-Analog Converter mentioned above is one example. Another, described more fully in the following article, is the 1781 Digital Limit Comparator, which compares the bridge reading with preset limits to determine whether a component is in or out of tolerance. Other processing (interface) instruments convert output signals from one instrument into the form required at the input of another instrument. For example, the parallel-to-serial converter mentioned above converts the parallel (all-at-once) data from the bridge into the serial (one-at-a-time) data required by a card punch. Other interface instruments change the voltage and impedance levels or the logical coding of signals from one form to another.

Input

Input equipment is used to perform the actual connection of the terminals of the component to the terminals of the bridge. Such equipment can range all the way from the 1680-P1 Test Fixture, in which components are manually inserted, to the more sophisticated reel-type handlers and vibratory-hopper feeders.

These devices connect components to the bridge one at a time. Some applications require that components be connected in a prescribed automatic sequence and then recycled, or scanned, in the same or a different sequence. For these applications the 1770 Scanner System described on page 13 is ideal.

Other Data

Information in addition to the measured value is often required, especially when automatic output equipment is used. Identifying data, such as serial number, or data on conditions of measurement, such as temperature, time of day, or elapsed time, can be obtained from additional instruments connected into the processing equipment (through interface equipment if necessary) along with the primary measurement data.

Figure 4. An automatic capacitor-test system produced by General Radio.

TYPICAL SYSTEMS

Figure 4 is a photograph of an automatic capacitor-testing system produced by General Radio for a large aerospace manufacturer. It is used for incoming inspection and quality-control testing of a broad variety of capacitors. The following components are included:

1) Component carrier and automatic indexer,
2) Bias-voltage power supply,
3) 1680-A Automatic Capacitance Bridge Assembly,
4) 1781 Digital Limit Comparator,
5) Serial-number generator,

Figure 5. Automatic capacitor-test system produced by General Radio for Corning Glass Works.

Photo courtesy of Corning Glass Works.

6) Two 1137-A Data Printers,
7) Interface panels,
8) Console.

The capacitors to be tested are loaded into the 50-component carrier stick, which is then inserted in the automatic indexer. The power supply applies bias voltage to each capacitor, and the capacitance is measured by the automatic bridge, which indicates the value and supplies the measurement data to the 1781 comparator. The comparator determines whether each capacitor is in or out of tolerance and feeds its data, along with a three-digit number from the serial-number generator, to the two printers. One printer prints the serial number and capacitance of each capacitor, the other the serial number and dissipation factor. The print-out is black for a good capacitor and red for a bad one, and it includes a digit to indicate the reason for the acceptance or rejection of the capacitor — GO, HIGH C, LOW C, or HIGH D.

Three modes of operation are provided. The MANUAL mode allows the capacitors to be tested one at a time by means of a push button on the indexer. The AUTOMATIC mode allows all 50 capacitors to be tested. In the STOP-NO-GO mode, the capacitors are tested and stepped along automatically until an out-of-tolerance value occurs. At this point the system stops, and it stays stopped until it is manually restarted.

Figure 5 shows a system supplied to Electronic Products Division, Corning Glass Works. This system includes a stick indexer and a card punch. To satisfy the requirements of the system, a 16-character data printer and a specially designed card-punch coupler were used.

This system is used by quality-control personnel to monitor the production of glass-dielectric capacitors. The capacitors are loaded into sticks and measured by the system. A tape printer and a card punch simultaneously record serial number, capacitance, and dissipation factor.

OTHER APPLICATIONS

These automatic systems can be used in the design of components for the evaluation of experimental units or for the testing and sorting of the output of a production line. They can also be employed to test purchased components in incoming-inspection or quality-assurance programs.

Capacitor testing is only one of the bridge's many applications. Atomic Energy of Canada, Limited, uses a system including a 1680-A bridge, 1136-A D/A converter, and 1521-B recorder to plot the effects of immersion in an electrolyte on the impedance of thin-film components. Another large laboratory uses similar systems in the design of inductors for telephone applications. At General Radio we use a 1680-A bridge system to check the turns ratio of the toroidal transformers used in the bridge itself!

ECONOMY

Although the cost-saving advantages of automation seem obvious, the results of the installation of these automatic systems are sometimes startling. Cost analyses indicate a saving of up to 80% or more on the per-unit cost of component inspection, in spite of the higher cost of the automatic equipment with respect to manually operated equipment. Customers have told us that a 1680-A bridge can pay for itself in six weeks!

SUMMARY

Those who wish to automate capacitance-measuring operations can now obtain from General Radio entire systems, custom-engineered for specific applications. This systems capability is graphically summarized in Figure 6.

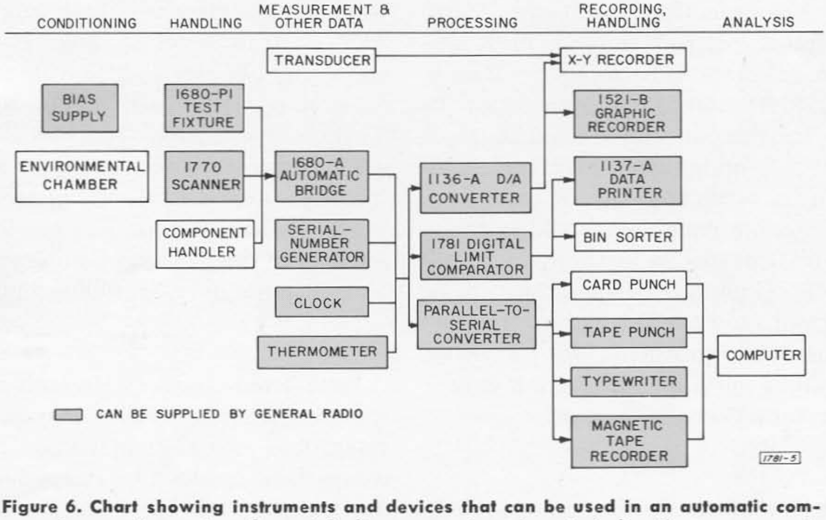

Figure 6. Chart showing instruments and devices that can be used in an automatic component-measuring system. General Radio can supply systems including those components indicated by tinted blocks.

Figure 1. GR 1781 Digital Limit Comparator shown with automatic bridge.

DIGITAL LIMIT COMPARATOR

As mentioned in the previous article, accessory instruments are required to exploit fully the capabilities of the 1680-A Automatic Capacitance Bridge Assembly. Figure 1 shows a most useful member of this accessory family, the 1781 Digital Limit Comparator. In this instrument, measurement data from the bridge are compared with manually set limits, and the results are presented on panel lamps and in relay-contact closures at a rear-panel connector. Thus the comparator can be used not only for testing and manual sorting of components to prescribed tolerances but also, with output equipment, for automatic sorting.

INPUT

The input to the comparator is in the form of 11 four-line BCD digits, five for loss value (dissipation factor or conductance), five for capacitance value, and one for range code. Positive true logic is required, with the logic "1," or true, level at least 10 volts above the "0" level. Although 1-2-4-2 data are normally used, a 1-2-4-8 modification is available. Two 5-digit limits, one upper and one lower, are set for the capacitance data, an upper limit is set for the loss data, and one limit is set for the range digit. Complete comparison takes only 2.5 milliseconds.

OUTPUT

Four front-panel lamps and corresponding internal SPDT relays provide visual and electrical indication of the comparison result. The lamps indicate GO, HIGH C, LOW C, or HIGH D/G. The relay contacts are isolated from the instrument and from each other and are rated at 115 V, 0.1 A. A data-output connector is also provided for connecting a data printer or an additional limit comparator.

SORTING

A 1680-A bridge and a 1781 comparator form a complete testing system by means of which an operator can manually sort capacitors into the four categories indicated by the comparator. This process can be speeded up by use of the output relay contacts to operate automatic bin sorters.

Several comparators can be connected together to sort components into several classifications. Figure 2, for example, shows three comparators connected to sort components into four tolerance groups: 1 percent, 5 percent, 10 percent, and greater than 10 percent. The measurement data digits are connected to all three comparators. After the first comparator has completed its calculation, it indicates the result and starts the second comparator. This sequence continues until one comparator indicates GO or until all indicate NO GO. The final decision will apply power to the appropriate output circuit through the relay contacts.

A 1137-A Data Printer can be connected to the comparator to obtain a record of the results. Three printer modes are provided:

1) Print all measurements;

2) Print in-tolerance measurements only;

3) Print out-of-tolerance measurements only.

In addition, a color-control signal is provided to print GO measurements in black and NO GO measurements in red. A digit representing the result of the comparison is also printed.

CIRCUIT DESCRIPTION

The 1781 Digital Limit Comparator is basically a sequential digital calculator. Comparison of the input data with preset limits is accomplished by a

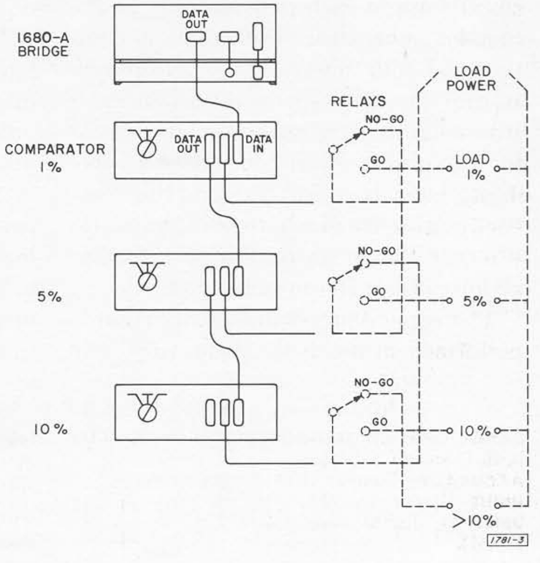

Figure 2. Diagram showing interconnection of three comparators for multiple-limit sorting. (GO-NO-GO relays shown are incorporated in comparators.)

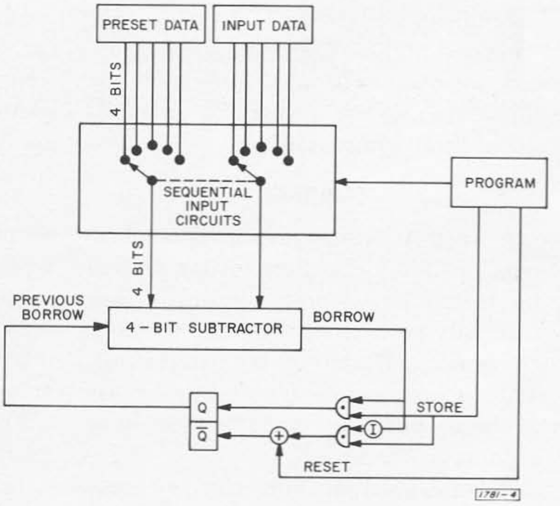

Figure 3. Simplified logic diagram illustrating comparison technique.

sequential technique in which one four-line binary-coded decimal digit is subtracted from another in a four-bit parallel subtraction circuit. In the loss comparison, for example, the measured value is subtracted from the preset limit value digit by digit, starting with the least significant digit. The simplified block diagram of Figure 3 illustrates the principle. The sequential input circuits apply each pair of digits to the four-bit subtractor (each digit is composed of four binary bits). After each subtraction, any borrow bit generated is stored and entered as a previous borrow for the next subtraction. After all five digits have been subtracted, the presence or absence of a borrow bit indicates whether the measured value is higher or lower than the preset value.

The capacitance-value comparison is performed in much the same way. The range digit is used as the most significant digit of capacitance value. At the completion of comparison, the final borrow information is used to select the proper output relays and lamps.

This technique involves a minimum of critical circuitry and requires only 2.5 milliseconds for a complete comparison.

CONCLUSION

Many of the advantages of an automatic bridge can be negated by an operator who misreads the indicated values or who occasionally translates measured data into incorrect decisions in handling or processing. The limit comparator, by substituting digital logic for human decision, advances the measurement system an important step nearer to elimination of human error in routine jobs.

SPECIFICATIONS

RANGE OF LIMIT SETTINGS: 00000 to 99999 for both C and D.
ACCURACY: Same as that of data source.
INPUT
Data: 11 digits, BCD, 1–2–4–2 (1–2–4–8 optional).
Logic Levels:
"1" $= V_{REF}$ (or V_{REF} to $V_{REF}-2$ volts)
"0" $= V_{REF}-10$ (or $V_{REF}-8$ to $V_{REF}-50$ volts).
Input Resistance: >47 kΩ (connected to V_{REF}).
Maximum Source Resistance: 100 kΩ.

Reference Voltage: ±50 V with respect to chassis ground. Maximum source resistance, 1 kΩ.

Compare Command: Logic 1 to logic 0 transition. Minimum duration, 2.5 ms; input resistance, >50 kΩ. Maximum source resistance, 20 kΩ.

OUTPUT

Data: Identical to input.

Comparison Result: BCD digit, behind 10 kΩ.

Print Command: Logic 1 to logic 0 transition, behind 2.2 kΩ.

Relay Contacts: 4 SPDT contacts, 115 V, rms, 0.1 A, rms, maximum.

GENERAL

Accessories Supplied: CAP-22 Power Cord, spare fuses, 4205-1000 Signal Cable to connect comparator to measuring instrument.

Accessories Required: If sorting equipment is used, 4205-1010 cable is also needed. See price table below.

Power Required: 105 to 125 or 210 to 250 V. 50 to 60 Hz, 10 W. (195 to 235 also available.)

Cabinet: Rack-bench.

Dimensions: Bench model, width 19, height 4, depth 16½ in (485, 105, 420 mm) over-all; rack model, width 19, height 3½, depth behind panel 16 in (485, 89, 410 mm).

Net Weight: 20 lb (9.5 kg).

Shipping Weight: 30 lb (14.0 kg).

Catalog Number	*Description*	*Price in USA*
1781-9801	**Type 1781 Digital Limit Comparator,** Bench Model	**$1625.00**
1781-9811	**Type 1781 Digital Limit Comparator,** Rack Model	**1625.00**
4205-1010	**Accessory Cable** (from sorting equipment to comparator)	**65.00**

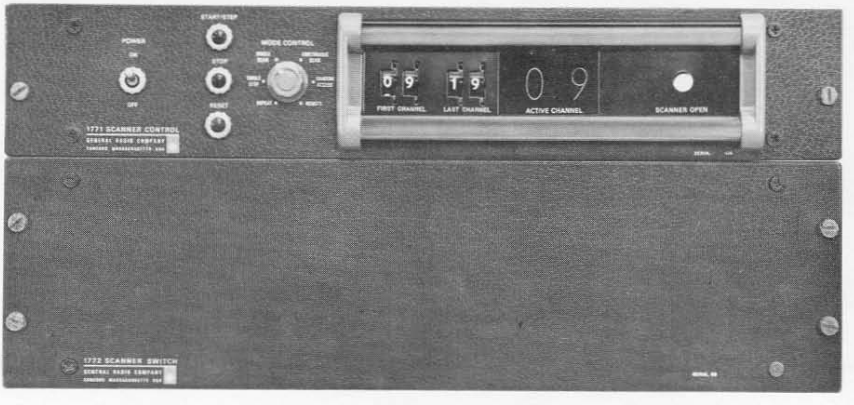

Figure 1. GR 1770 Scanner System.

AUTOMATIC SCANNER SYSTEM

The 1770 Scanner System shown in Figure 1 is an automatic instrument for the sequential connection of many pairs of terminals to a single pair. Its primary purpose is to connect unknown capacitors to the 1680-A Automatic Capacitance Bridge Assembly. This bridge can detect changes in capacitance as small as 0.01 pF and, because of its three-terminal configuration, it will ignore stray capacitance from each component terminal to ground. With other scanners, this capability cannot be used to advantage. In the 1770 Scanner System a guard arrangement connects stray capacitance (caused by long cables, for instance) from the component terminals to ground, leaving less than 0.01 pF across the component itself.

DESCRIPTION

The Scanner System is composed of two main units, the 1771 Scanner Control and the 1772 Scanner Switch. The components under test are connected to the switch unit by long cables, each consisting of 10 wires in a single shielded bundle. The switch unit is similarly connected to the control unit by a long cable. This arrangement is extremely adaptable, since the components can be separated from the switch unit, which can in turn be separated from the control unit. In environmental testing, for example, components in a test chamber can be connected to the switch unit outside the chamber by the cable bundle. The arrangement also allows expansion of the channel capacity, since several switch units can be connected to a single control unit.

SCANNER SWITCH UNIT

The 1772 Scanner Switch accepts up to 10 plug-in reed-switch modules. Two types of switch modules are available: the 1772-P1 SPST Scanner Module and the 1772-P2 SPDT Scanner Module. Each module can accept 10 input lines and connect them sequentially to one output line. The 1772-P1 SPST module connects the input lines, one at a time, to the output line and leaves the unused lines open. The 1772-P2 SPDT module operates in the same manner except that the inactive input lines are connected to a common (ground) terminal. Figure 2 shows the use of these modules in the guarded scanning of two-terminal components. Only three of the 10 input channels are shown. Note that this connection places the stray capacitance of the inactive channels between the low output terminal and ground rather than across the output terminals. This connection is shown more clearly in Figure 3.

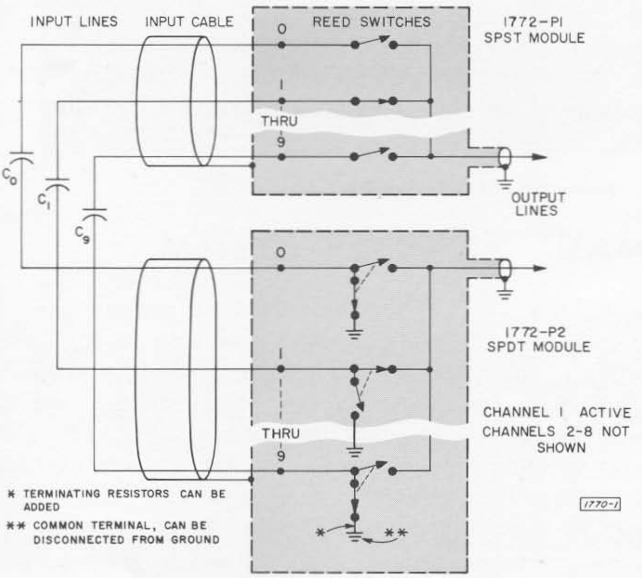

Figure 2. Simplified schematic diagram showing connections for guarded scanning of two-terminal components.

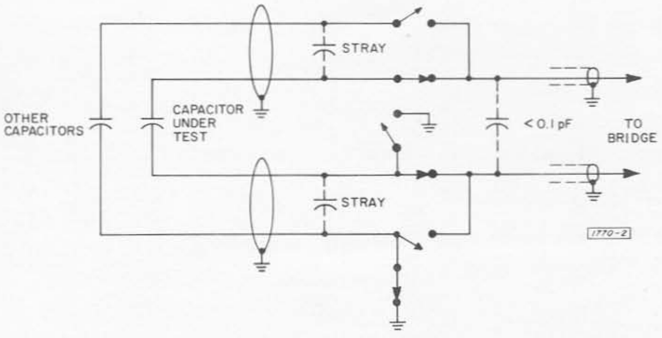

Figure 3. Simplified schematic diagram of scanner switching circuit.

Each module uses long-life reed switches capable of over 25 million operations. Total channel resistance is less than 200 milliohms.

The 1772 Scanner Switch can accept 10 modules of either type and can provide, for example, a scanning capacity of 50 two-terminal components or 100 single-terminal signal lines.

Physical Description

Figure 4 shows a 1770 Scanner System used with a 1680-A Automatic Capacitance Bridge Assembly. The 1772 Scanner Switch is shown at the left, housed in a 5½-inch-high relay-rack cabinet. All connections are made at the rear of the unit. The 1772-P1 and -P2 Scanner Modules plug in as shown. Each module includes a 14-pin connector for the input lines and two parallel BNC connectors for the output lines. One BNC output connector can be used for connection to the external measuring instrument and the second to other modules. The main chassis has two connectors, one for connection to the 1771 Scanner Control and one for connection to other Scanner Switches. Two 50-foot rolls of 10-conductor shielded cable are supplied for connection to the input lines.

SCANNER CONTROL UNIT

The 1771 Scanner Control Unit is an all-solid-state instrument that can control up to four 1772 Scanner Switches. Its maximum capacity is 100 single or

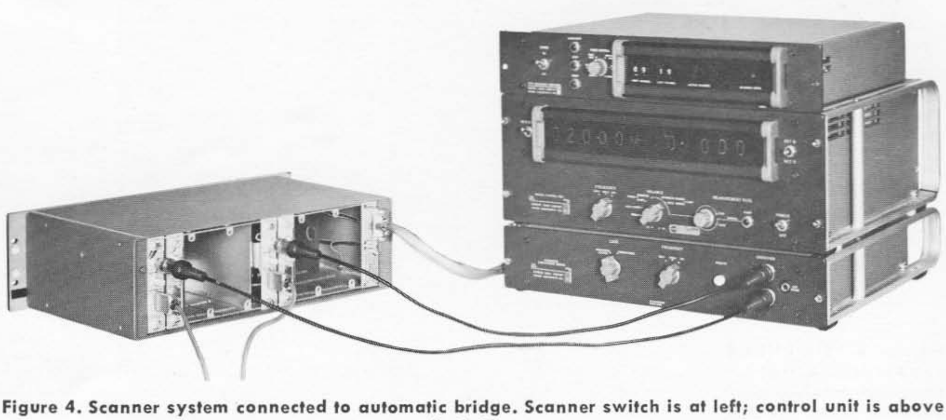

Figure 4. Scanner system connected to automatic bridge. Scanner switch is at left; control unit is above the bridge.

multiple channels. The instrument has thumb-wheel switches for channel selection and a bright in-line numerical readout of the active channel. An electrical readout is also provided, along with various other electrical inputs and outputs.

OPERATING MODES

The thumb-wheel switches are set to the first and last channels of a desired sequence. Six modes of operation can be selected by the front-panel MODE CONTROL switch:

(1) SINGLE STEP, in which channel advance is initiated by manual operation of a front-panel pushbutton or by an electrical signal applied to a rear connector:

(2) SINGLE SCAN, in which a channel group is automatically scanned once from the first to the last channel. At the end of the sequence the first channel is ready to be activated:

(3) CONTINUOUS SCAN, in which a channel group is repetitively scanned;

(4) REPEAT, in which the scanner remains on the first channel. A pushbutton or an external input is used to generate an output "measure" signal;

(5) RANDOM ACCESS, in which any channel can be selected by BCD contact closures applied to a rear connector;

(6) REMOTE, in which any mode of operation can be selected by contact closures applied to a rear connector.

TIMING

A scanning sequence is initiated by a signal from the automatic capacitance bridge. An output delay interval, adjustable from 50 milliseconds to 2.5 seconds, is generated to allow proper operation of other equipment that may be activated by the same signal. After this delay the scanner advances one channel, and a second interval, of 10 milleseconds, allows for switch bounce time and data settling. At the end

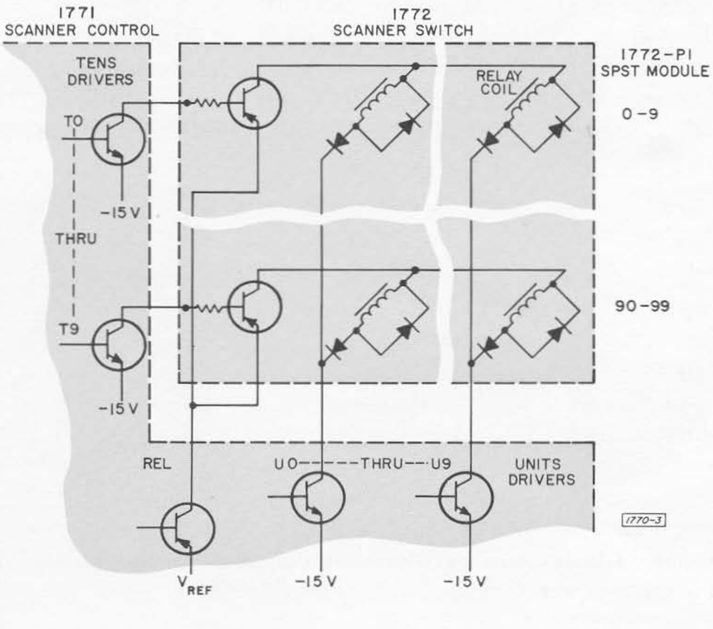

Figure 5. Simplified schematic diagram showing interconnections between control unit and switch unit.

Figure 6. Bridge-scanner system used in environmental tests on dipped-mica capacitors.

Photo courtesy of the Electro Motive Manufacturing Co., Inc.

of this interval an output signal is generated to operate the bridge. At the same time, a third interval, adjustable from 100 milliseconds to 25 seconds, is initiated. In the SINGLE SCAN or CONTINUOUS SCAN mode the scanner advances to the next channel at the end of this interval unless a signal from the bridge has caused it to advance sooner.

Front-panel terminals allow the operator to monitor the delay signals on an oscilloscope.

SCANNER CONTROL – SCANNER SWITCH INTERCONNECTIONS

Figure 5 shows a simplified schematic diagram of some of the interconnections between the Scanner Control Unit and the Scanner Switch Unit. Two decade counters in the Scanner Control Unit determine the active channel by means of the matrix ("crosspoint") connection shown. For example, if the tens decade is set to 0 and the units decade to 9, tens-driver transistor *TO* and units-driver transistor *U9* will be held in saturation, connecting the coil of reed switch *09* between V_{REF} and the −15-volt supply. The diodes in series with the reed-switch coils allow the matrix to operate, while those in parallel with the coils prevent the surge transient from damaging the driver transistors.

The coil-return line, *REL*, is connected through a normally saturated transistor to V_{REF}. This transistor is cut off for the reset operation, disabling all channels.

The channel-driver transistors can supply up to 80 mA. The reed-switch coils in the 1772-P1 SPST Scanner Module require 15 mA and those in the 1772-P2 SPDT Scanner Module require 30 mA. Thus one Scanner Control Unit can drive, for example, four 100-channel single-wire Scanner Switch Units (10 1772-P1 SPST modules in each) or one 100-channel guarded-two-terminal scanner consisting of two Scanner Switch Units (10 1772-P1 SPST modules in one and 10 1772-P2 SPDT modules in the other).

Harold T. McAleer received the B.S.E.E. and M.S.E.E. degrees from the Massachusetts Institute of Technology in 1953. He was employed as a cooperative student at GR on the design of high-frequency measuring and recording instruments. After two years' service as an engineer with the U. S. Army Signal Corps at Fort Monmouth, he returned to GR as a development engineer in GR's Frequency Group, where he designed frequency counters, and associated instruments. Since early 1966 he has been responsible for the design of systems at GR. He is a registered professional-engineer and a member of the IEEE, Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

Richard F. Sette, designer of the 1781 Digital Limit Comparator and 1770 Scanner System, received the B.S.E.E. and M.S.E.E. degrees from Northeastern University in 1960 and 1962, respectively. He performed graduate research in solid-state physics at the Air Force Cambridge Research Laboratory and later served in the U. S. Army Electronics Laboratory at Fort Monmouth, where he was concerned with the testing and evaluation of integrated circuits. Since joining GR in 1964 he has been responsible for the design and development of data-processing equipment for automatic testing systems. He is a member of the IEEE and Eta Kappa Nu.

APPLICATIONS

Figure 6 shows the 1770 Scanner System used with a 1680-A Automatic Capacitance Bridge Assembly at Electro Motive Manufacturing Company in Willimantic, Connecticut, to perform environmental-performance inspection of high-reliability dipped-mica capacitors. The capacitors are checked in an environmental chamber over a range of −55 to +150°C.

In another custom-designed measurement system, a 1770 Scanner System is used in the testing of touch-dial-telephone inductors. By use of the random-access mode, external programming equipment is used to connect a prescribed group of inductors, first for demagnetizing and then, after a known time interval, for measurement. The flexibility of the scanner system makes it ideal for this application.

The 1770 Scanner System can be used for signal-line scanning with either type of switch module. The 1772-P2 SPDT Scanner Module is constructed so that terminating resistors can be automatically connected to inactive channels (see Figure 2).

The scanner system can also be used in reverse as an output scanner, to connect a single input signal to many output terminals in sequence. In one such application, for example, two scanner switch units are operated from one control unit. One of the switch units connects input signal lines one at a time to a measuring instrument, while the other switch unit connects the output of the instrument to a multiple-channel recorder.

— Harold T. McAleer
Richard F. Sette

SPECIFICATIONS

1771 SCANNER CONTROL

INPUT

Logic Levels ($V_{REF} = V_{chassis} \pm 50$ V):
"1" = V_{REF} (or V_{REF} to $V_{REF} - 2$ volts)
"0" = $V_{REF} - 10$ (or $V_{REF} - 8$ to $V_{REF} -$ 15 volts).

Channel Advance: Logic-level transition, either polarity; input resistance = 50 kΩ; maximum source resistance = 20 kΩ.

Or, contact closure to V_{REF}; input resistance = 1.8 kΩ; open-circuit voltage = $V_{REF} - 15$ volts.

Remote Control: Start, stop, reset, skip, inhibit, random-access channel selection and mode; contact closure to V_{REF}; input resistance = 1.8 kΩ; open-circuit voltage = $V_{REF} - 15$ volts.

OUTPUT

Channel Identification: BCD 1–2–4–2 (1–2–4–8 optional); output resistance = 22 kΩ.

Measure Command: Logic 0 to logic 1 transition behind 0.001 μF.
Channel-Switch Drive Current: 80 mA.
End of Scan: Logic-level transition, either polarity, behind 10 kΩ.
TIMING DELAYS
Settle Delay: 10 ms.
Safety Delay: Adjustable 100 ms — 25 s.
Output Delay: Adjustable 50 ms — 2.5 s (can be inhibited).

1772 SCANNER SWITCH
Capacity: 10 scanner modules.
1772-P1 SPST SCANNER MODULE
Complement: 10 SPST switch contacts, rated at 25 V, 0.1 A, maximum. Channel-drive requirement: 15 V at 15 mA.
1772-P2 SPDT SCANNER MODULE
Complement: 10 SPDT switch contacts, rated at 25 V, 0.1 A, maximum. Channel-drive requirement: 15 V at 30 mA.

ORDERING INFORMATION — *The 1770 Scanner System is not presently available as an "off the shelf" item. Inquiries are invited for scanner systems as described above. The price for a typical 50-channel guarded scanner is approximately $3500.*

EXPERIMENTER on Microfilm

Libraries and others faced with the problems of periodical storage can now obtain any or all volumes of the *Experimenter*, from Volume 1 (1926) to the present, on microfilm. Each volume is in the form of positive microfilm, on metal reels suitably labeled. Enlarged copies can be printed on request. All inquiries regarding this service should be directed to University Microfilms, 313 North First Street, Ann Arbor, Michigan 48106.

INSTRUCTION MANUALS FOR THE CLASSROOM

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ADDENDUM

In the specifications for the TYPE 1394 High-Rate Pulse Generator (July 1966 issue, page 8), the type designations 1394-A and 1394-Z appear without explanation of the distinction between them. As mentioned early in the issue, the High-Rate Pulse Generator alone is designated TYPE 1394-A; the designation 1394-Z identifies the combination of TYPE 1394-A High-Rate Pulse Generator and TYPE 1394-P1 Pulse-Offset Control.

In the specifications for the 1525-A Data Recorder (October issue), a line of type was omitted from the *Weighting Characteristics*. The complete specification follows:

Weighting Characteristics: NAB and constant current for both playback amplifiers; NAB constant current, and A, B, and C weighting (standard sound-level-meter characteristic) for record channel #1. Constant current for record channel #2.

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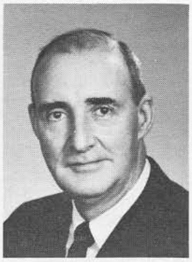

With this issue the *Experimenter* bids farewell to Charles E. Worthen, who for over 30 years has been its Editor.

Charles Worthen joined GR in 1928, after graduating from M.I.T., and, working with J. K. Clapp, helped pioneer the first commercial primary frequency standards. In 1934 he turned his full attention to the written word; since then, as Publicity Manager and as Director of the Sales Promotion Department, he has shepherded the *Experimenter*, catalogs, instruction manuals, handbooks, ads, and general publicity. His editorial judgment was always sound; his writing was always characterized by good taste, a style both readable and correct, and a talent for uncomplicating the complex. Perhaps more than any other individual, he has set standards for the literature of the electronics industry.

Now he retires, the better to enjoy his family, his books, some travel, and a well-earned freedom from deadlines. Our best wishes go with this man whose words were always so well chosen.

F. Van Veen, Editor

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