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An Automatic, Precision 1-MHz Digital LCR Meter

This new LCR Meter is especially designed for measurements of low-value components in integrated circuits and thin-film hybrid circuits, as well as of discrete components. It uses a four-pair measurement technique to reduce errors caused by stray capacitance, residual inductance, and mutual inductance of the test leads.

by Kohichi Maeda

IN THE MANUFACTURE of integrated circuits, good process control is essential if yield and reliability are to be maximized. Good process control requires many measurements of parameters on the chip; for example, capacitance versus voltage determines several electrical characteristics of junctions and is an important source of information about the process.

Measurement of individual components is also important in the design, testing, and production of thin-film hybrid circuits. Components of interest include resistors, capacitors, inductors, diodes, and transistor junctions.

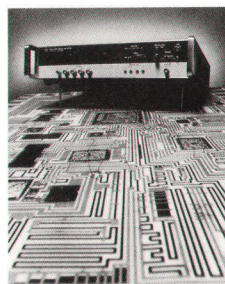
Measurements of low-value components are, of course, subject to errors caused by parasitic reactances such as stray capacitance and residual inductance. To minimize these errors, the new HP Model 4271A 1-MHz LCR Meter (Fig. 1) uses a four-pair measurement technique: the interconnections between the instrument and the circuit under test are four coaxial cables up to one meter long. Also, offset adjustments are used to cancel stray capacitance and conductance or residual inductance and resistance in the test leads between the circuit under test and the inputs to the coaxial cables.

The new LCR Meter typically makes five or more readings per second at basic accuracies of 0.1% in capacitance measurements and 0.2% in inductance, resistance, and conductance measurements. Its readout is a four-digit light-emitting-diode display that has 90% overrange capability. It displays capacitance from 0.001 picofarad to 19.000 nanofarads, inductance from 0.1 nanohenry to 1900.0 microhenries, resistance from 0.001 ohm to 19.000 kilohms (or conductance from 0.01 micromhos to 190.00 millimhos), and dissipation factors of capacitors or inductors from 0.0001 to 1.6000.

Because it is intended for semiconductor measure-

ments, the test voltages and currents used by the new LCR Meter are low. Capacitance test voltages are 20 mV and 500 mV rms. Inductance test currents are 5 mA to 2 μ A. A remotely programmable dc bias source provides 0 to 39.9 volts in 0.1V steps.

Options are available for interfacing the LCR Meter to HP calculators and computers.



Cover: Model 4271A Digital LCR Meter is well suited for production measurements of chip capacitors and other components that are used on thin-film hybrid microcircuits like the one pictured here (it's actually a large photo of a microcircuit). Also

easily measured by the new LCR Meter are varactor diodes, reed relays, delay lines, pulse transformers, and many other components, discrete and otherwise.

In this Issue:

An Automatic, Precision 1-MHz Digital LCR Meter, by Kohichi Maeda **page 2**

A Moderately Priced 20-MHz Pulse Generator with 16-Volt Output, by Günter Krauss and Rainer Eggert ... **page 10**

Laboratory Notebook—Logarithmic Amplifier Accepts 100 dB Signal Range **page 16**

Versatile VHF Signal Generator Stresses Low Cost and Portability, by Robert R. Hay **page 18**



Fig. 1. Model 4271A LCR Meter has a four-digit display with 90% overrange. Basic accuracies are $\pm 0.1\%$ in C measurements and $\pm 0.2\%$ in L, R, and G measurements. It also measures dissipation factors. All measurements are made at 1 MHz and at very low test signal levels.

How It Works

The usual way to make precise high-resolution L, C, and R measurements is the ac bridge method, one arm of the bridge being the unknown and another arm a standard. In an automatic digital ac bridge, some form of digital-to-analog converter could be used to adjust either the standard or the voltage across the standard to balance the bridge.

For a 1-MHz instrument like the 4271A, this method has disadvantages. The need to compensate for phase shift variations makes high-frequency digital-to-analog converters complicated, costly, slow, and of questionable reliability.

The method used in the 4271A is a combination of an ac bridge method and a dual-slope digital voltmeter technique. It eliminates the problem of phase errors and results in a relatively simple circuit design.

The LCR Meter consists of a bridge section and a measurement section. Fig. 2 is a block diagram of the bridge section. In a C-G measurement, if the bridge

is not balanced, an unbalance current i_d flows into the current detector, which produces an error voltage e_d . This voltage is amplified, phase-detected, and rectified to produce a dc voltage E_1 proportional to the real part of e_d and a dc voltage E_2 proportional to the imaginary part of e_d .

The dc voltages E_1 and E_2 are then integrated. The integrator outputs modulate two 1-MHz reference signals to produce two 1-MHz signals e_1 and e_2 corresponding to the real and imaginary parts of the error voltage e_d . These voltages are added in a summing amplifier and applied to the standard resistor R_s . This applied voltage, e_v , causes the current i_d to decrease.

When i_d becomes zero, e_v remains constant and the bridge, which is now balanced, automatically remains balanced. Then

$$e_v = -(j\omega C_x + G_x)e_r R_s.$$

The L-R bridge works the same way as the C-G

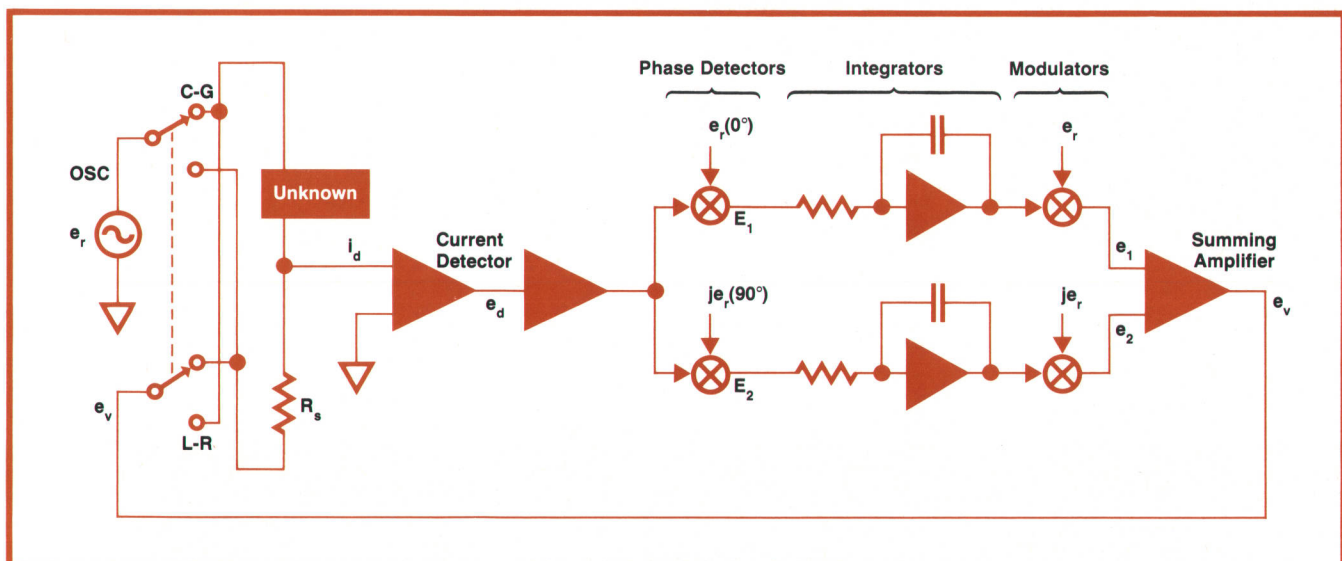


Fig. 2. Bridge section of the LCR Meter. Balancing is done automatically.

bridge. However, the oscillator voltage is now applied to the standard resistor R_s and the feedback voltage is applied to the unknown impedance. At balance,

$$e_v = -(j\omega L_x + R_x)e_r/R_s.$$

Measurement Section

In the measurement section, the dual-slope digital voltmeter technique is used to measure the vector ratio of e_v and e_r . Fig. 3 is a block diagram.

The C_s/G_s amplifier is a kind of phase shifter. When S_G is in its "on" position the voltage gain of this circuit is $k_0 G_s$ where k_0 is the current-to-voltage conversion constant of the amplifier. When S_C is in its "on" position the voltage gain is $j\omega C_s k_0$. When S_G is on, S_C is off and vice versa.

Phase detectors 1 and 2 rectify the amplifier output synchronously with 0° and 90° phase reference signals derived from e_r . The output of each phase detector is a dc voltage proportional to the component of the amplifier output that is in phase with the corresponding phase reference signal. The two dc vol-

tages then charge and discharge the two integrators according to the logic control signals as the dual-slope analog-to-digital conversion proceeds.

At the start of a **C measurement**, switch S_r is open, S_v is closed, S_G and S_{x1} are on, and S_C is off. Integrator 1 is charged for a constant period T_1 . The charge stored during this time is

$$Q_{C1} = -a_1 G_s R_s \omega C_x e_r T_1$$

where the constant a_1 is the product of k_0 , the efficiency of phase detector 1, and the integrating constant of integrator 1.

Next S_r closes, S_v opens, S_C turns on, and S_G turns off. S_{x1} remains on. Integrator 1 is discharged until its output is zero volts. If the discharge time is T_2 , the charge dissipated is

$$Q_{C2} = a_1 \omega C_s e_r T_2.$$

Then, because $Q_{C1} = -Q_{C2}$,

$$C_x = C_s T_2 / (G_s R_s T_1).$$

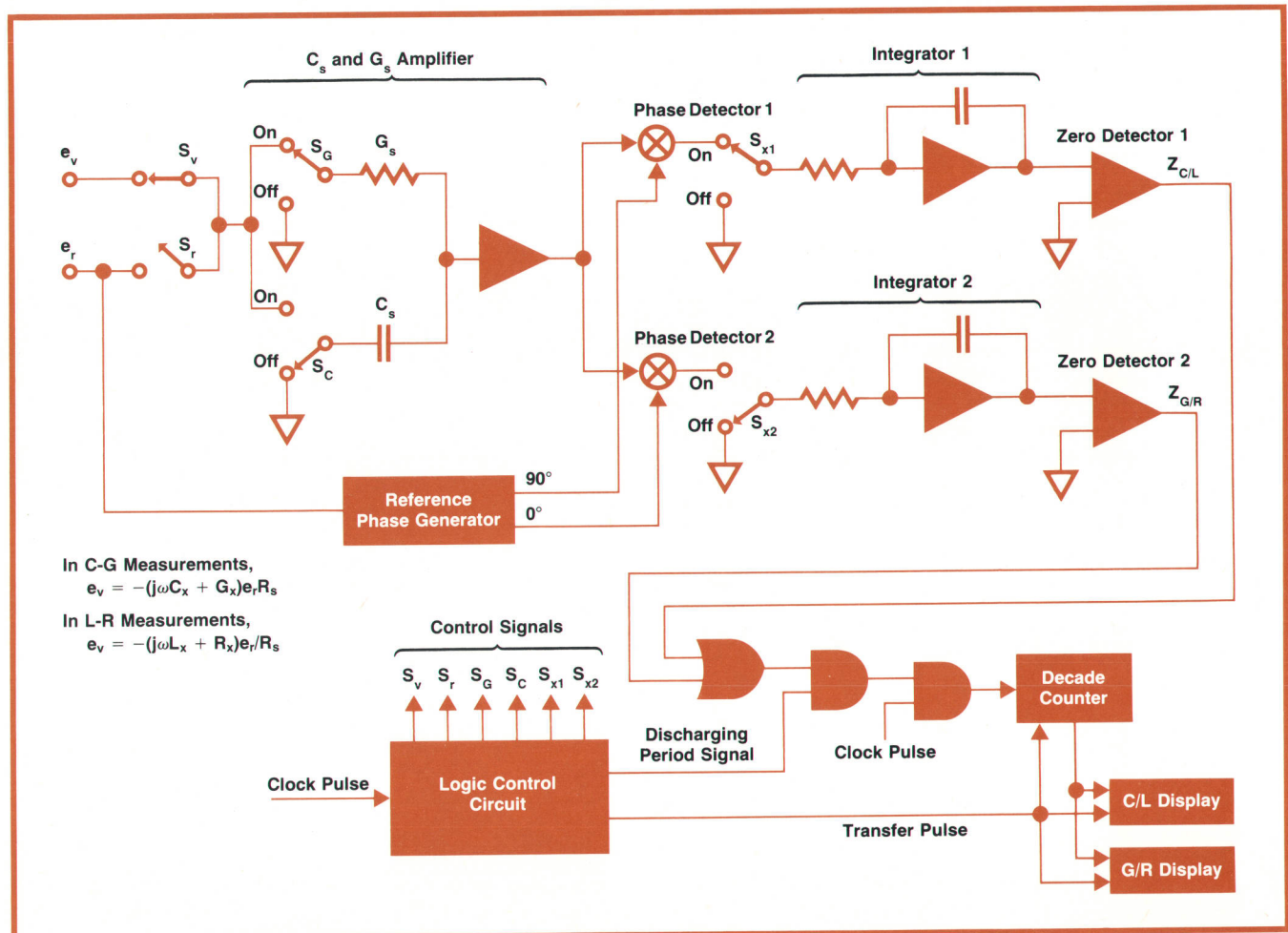


Fig. 3. Measurement section uses the dual-slope integrating digital voltmeter technique.

R_s , G_s , C_s , and T_1 are selected so the unknown capacitance can be measured directly by counting T_2 .

A **G measurement** begins with switch S_v closed, S_r open, S_G and S_{x2} on, and S_C off. The charge stored during the charging period T_1 is

$$Q_{G1} = -a_2 G_s R_s G_x e_r T_1$$

where the constant a_2 is the product of k_o , the efficiency of phase detector 2, and the integrating constant of integrator 2

During the discharge period S_r closes, S_v opens, S_G and S_{x2} are on, and S_C is off. When the output of integrator 2 reaches zero volts after a time T_3 the charge dissipated is

$$Q_{G2} = a_2 G_s e_r T_3.$$

But $Q_{G1} = -Q_{G2}$, so

$$G_x = T_3 / (R_s T_1).$$

R_s and T_1 are selected so the unknown conductance G_x can be measured directly by counting T_3 .

L and R measurements are similar to C and G measurements. The results are

$$L_x = R_s C_s T_2 / (G_s T_1)$$

and

$$R_x = R_s T_3 / T_1.$$

Dissipation factor measurements use both integrators. To obtain the dissipation factor of a capacitor, defined as

$$D_x = G_x / (\omega C_x),$$

the sequence begins with S_r closed, S_v open, S_C and S_{x1} on, and S_G off. Integrator 1 charges for a time T_4 , storing charge

$$Q_{C4} = a_1 \omega C_s e_r T_4.$$

Integrator 1 is then discharged with S_v closed, S_r open, S_G and S_{x1} on, and S_C off, until its output reaches zero volts. S_{x2} is also on during this time, so integrator 2 charges. If the time required to discharge integrator 1 is T_5 , the charge removed from integrator 1 is

$$Q_{C5} = -a_1 G_s \omega C_x R_s e_r T_5$$

and the charge stored by integrator 2 is

$$Q_{G5} = -a_2 G_s R_s G_x e_r T_5.$$

Next, integrator 2 is discharged with S_r closed, S_G and S_{x2} on, and S_C off, until its output becomes zero after a time T_6 . The charge removed is

$$Q_{G6} = a_2 G_s e_r T_6.$$

Because $Q_{C4} + Q_{C5} = 0 = Q_{G5} + Q_{G6}$,

$$T_5 = C_s T_4 / (G_s R_s C_x)$$

and

$$T_6 = R_s G_x T_5.$$

Therefore

$$\begin{aligned} T_6 &= C_s G_x T_4 / (G_s C_x) \\ &= D_x T_4 / D_s. \end{aligned}$$

C_s , G_s , and T_4 are chosen so that D_x can be measured directly by counting T_6 .

To measure the dissipation factor of an inductor, defined as

$$D_x = R_x / (\omega L_x),$$

the procedure is the same.

Notice that in all measurements—C, G, L, R, and D—the constants a_1 and a_2 , which express the characteristics of the C_s/G_s amplifier, the phase detectors, and the integrators, have nothing to do with the final result. This is a major advantage of the dual-slope technique.

Compensation Periods Added

Fig. 4 shows the switch positions and integrator waveforms for various types of measurements. Notice that compensation periods are included in all measurements. These are for cancelling phase errors caused by variations in the phase relationships between the reference signals and the detected signals at the two phase detectors. These errors, their effects and the method of compensation are described in the box on page 6.

Four-Pair Measurement Technique

The test terminal configuration is very important in measurements of very low or very high impedances. For example, in an ordinary two-terminal measurement of a very high impedance, stray capacitance to ground can cause serious errors. These can be eliminated by surrounding the unknown with a shield or guard, thereby making the measurement a three-terminal one.

Phase Error Compensation

In the measurement section of the 4271A LCR Meter, phase detectors are used to generate dc voltages proportional to the real and imaginary parts (in-phase and quadrature components) of the signal at the inputs to the phase detectors. To do this precisely, it's necessary to maintain the precise phase relationship between the reference signal and the detected signal. If any phase error exists, the dc outputs of the phase detectors differ from their proper values.

If the input signal is $A + jB$, and if there is phase error θ_1 in the real (0°) branch of the circuit and phase error θ_2 in the imaginary (90°) branch, the actual outputs A' and B' will be related to the correct outputs A and B by

$$A' = A \cos \theta_1 + B \sin \theta_1$$

$$B' = B \cos \theta_2 + A \sin \theta_2$$

Notice that there are two types of errors, cosine and sine. Cosine error is not serious because it affects only the phase detector efficiency, which has little to do with the final result (see text, page 5).

Sine error, on the other hand, is very troublesome, especially when one of the components is much larger than the other. This error can make it difficult to make a precise measurement of the dissipation factor of a low-D inductor or capacitor, the small parallel capacitance of a resistor, or the small series inductance of a resistor. As shown in Fig. A for a C-G measurement, the error in T_2 , the C_x reading, is $D_x \theta_2$ and the error in T_3 , the G_x reading, is θ_1/D_x . Thus, for example, in a C-G measurement where $D_x = 0.1$ and $\theta_1 = \theta_2 = 0.01$ radian, the measured value of C_x would be in error by 0.1% and the

measured value of G_x would be in error by about 10%.

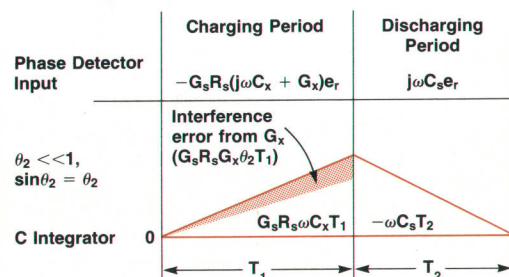
To minimize phase errors, compensation periods are added to the 4271A measurement sequence. Fig. B shows how they work in a C measurement.

During the charging period T_1 , phase error causes the charge accumulated by the C integrator to differ from the correct amount. The G integrator is also charged during T_1 , and a compensation period equal to T_3 , the time required for discharging the G integrator (integrator 2), is added to the C measurement sequence. During this time the phase error on the C integrator has approximately the same magnitude as that during T_1 , but opposite polarity. The result is that the discharging period T_2 , which follows the compensation period, is in error by only $-\theta_1 \theta_2$. The error no longer depends on D_x .

Also shown in Fig. B is a compensated G measurement sequence. The resulting error in T_3 is also $-\theta_1 \theta_2$. Thus, if $\theta_1 = \theta_2 = 0.01$ radian, the measurement errors would be only -0.01% in both C_x and G_x , instead of 0.1% and 10% as they were in our uncompensated example.

This method is also used to compensate L, R, and D measurements.

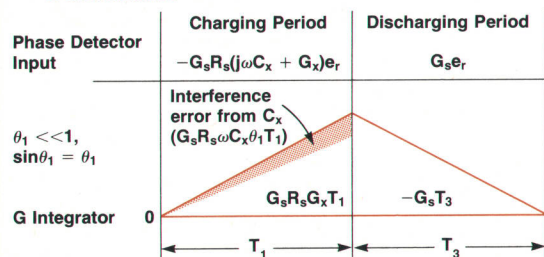
C Measurement



$$G_s R_s \omega C_x T_1 + G_s R_s G_x \theta_2 T_1 = \omega C_s T_2$$

$$\text{Therefore } T_2 = C_x (G_s R_s / C_s) (1 + D_x \theta_2) T_1$$

G Measurement

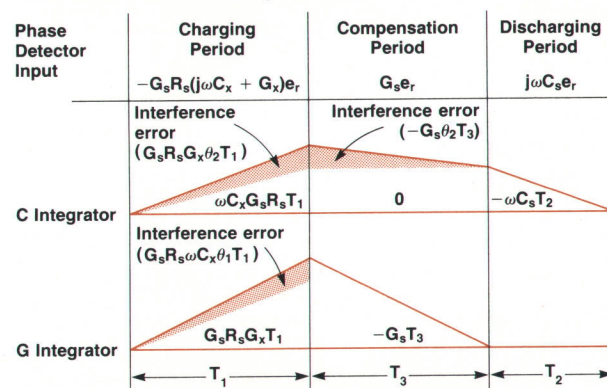


$$G_s R_s G_x T_1 + G_s R_s \omega C_x \theta_1 T_1 = G_s T_3$$

$$\text{Therefore } T_3 = G_x R_s (1 + \theta_1 / D_x) T_1$$

Fig. A. C and G measurements with uncompensated phase errors.

C Measurement with Compensation Period

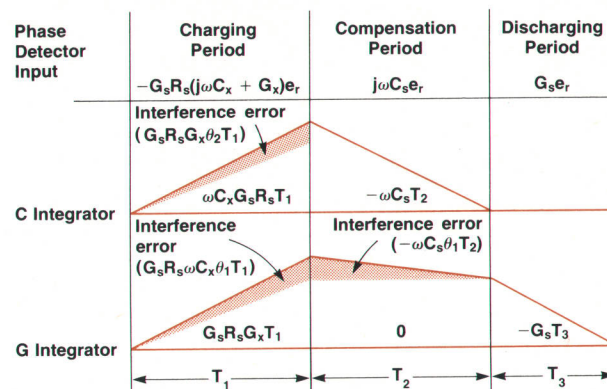


$$a: (\omega C_x G_s R_s + G_s R_s G_x \theta_2) T_1 - G_s \theta_2 T_3 - \omega C_s T_2 = 0$$

$$b: G_s R_s G_x T_1 + G_s R_s \omega C_x \theta_1 T_1 - G_s T_3 = 0$$

$$\text{From a and b, } T_2 = C_x (R_s G_s / C_s) (1 - \theta_1 \theta_2) T_1$$

G Measurement with Compensation Period



$$c: G_s R_s G_x \theta_2 T_1 + \omega C_x G_s R_s T_1 - \omega C_s T_2 = 0$$

$$d: (G_s R_s \omega C_x \theta_1 + G_s R_s G_x) T_1 - \omega C_s \theta_1 T_2 - G_s T_3 = 0$$

$$\text{From c and d, } T_3 = R_s G_x (1 - \theta_1 \theta_2) T_1$$

Fig. B. C and G measurements with phase-error compensation periods.

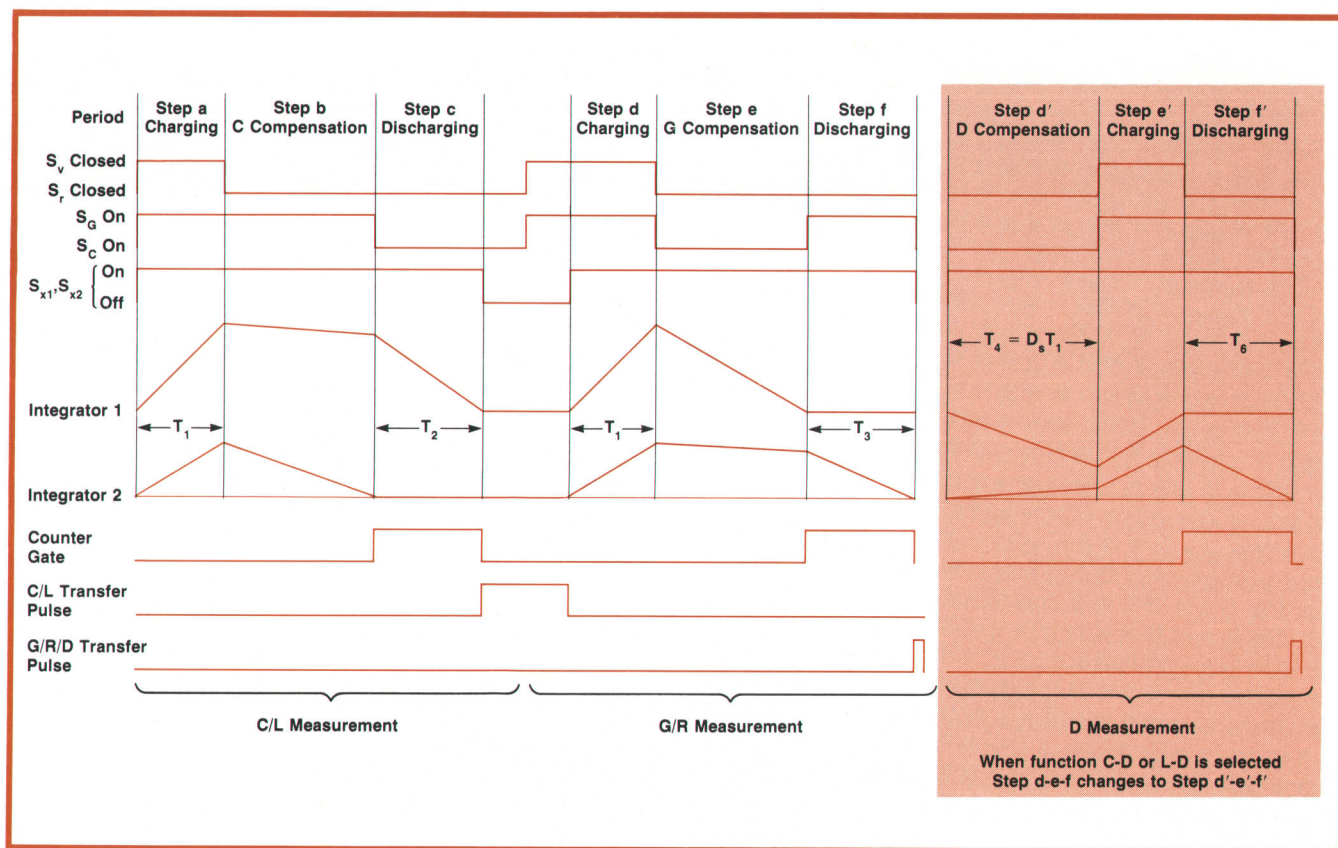


Fig. 4. Measurement sequence, showing the charge-discharge dual-slope measurement periods and compensation periods for cancelling phase errors.

In very-low-impedance measurements, lead impedances can seriously affect accuracy. To eliminate lead-impedance errors, a four-terminal technique is often used: current and voltage are measured by separate pairs of leads. To eliminate the effects of stray capacitance to ground a guard circuit may be added, making a five-terminal configuration.

There is a further problem in very-low-impedance measurements that is not solved by either the four-terminal or the five-terminal configuration, and that is the error caused by mutual inductance between the current leads and the potential leads: the field generated by current in the current leads induces an error voltage in the potential leads. This is sometimes dealt with by twisting the current leads together or maintaining the potential leads at a 90° angle from the current leads. Neither of these methods is adequate for the 4271A, which has a high resolution of 0.1 nH in inductance measurements.

The 4271A solution is to use four coaxial cables, as shown in Fig. 5. Currents in the outer and inner conductors of each cable have the same magnitude and opposite polarities. Thus there are no fields outside the cables and mutual inductance is not a problem. The new LCR Meter can tolerate cables up to one meter long without significant error.

Offset Adjustments

Some components, such as those on an IC chip, do not lend themselves to four-pair connections. In these measurements there will always be some stray capacitance or residual inductance in the probe area (see Fig. 6). To cancel these, the LCR Meter has front-panel offset adjustments for C,G,L, and R.

Offset adjustments of residual inductance and resistance are essential in measurements of low impedances such as high value capacitors, low-value inductors, and low-value resistors. Similarly, offset adjustments of stray capacitance and conductance are essential in measurements of high impedances such as low-value capacitors, high-value inductors, and high-value resistors.

A special advantage of the multifunction 4271A is the ability to make L-R offset adjustments in C-G measurements and C-G offset adjustments in L-R measurements. Fig. 6 shows how much influence residual inductance and resistance and stray capacitance and conductance have on C-G and L-R measurements, respectively. These errors can be minimized by offset adjustments.

For example, in a high-capacitance measurement, an L-R measurement can first be made on a short circuit and the offset controls adjusted so the L and R

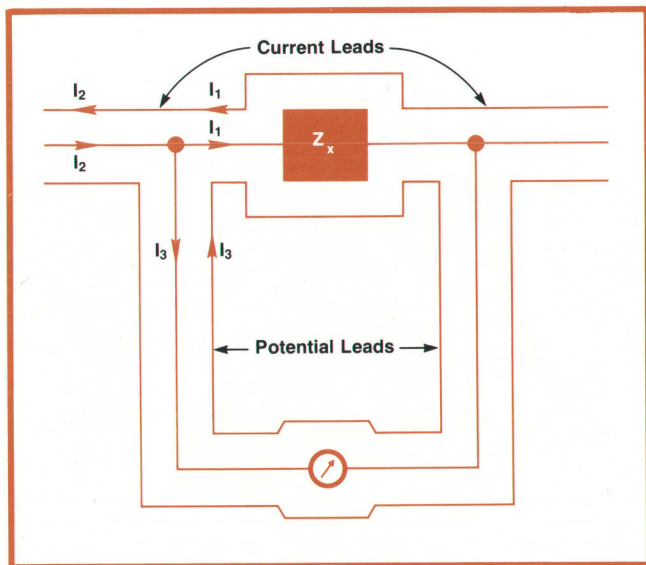


Fig. 5. Four-pair measurement technique uses coaxial cables for current and potential leads, eliminating mutual inductance that causes errors.

displays read zero. Then the unknown capacitor can be measured accurately in a C-G measurement.

Inductance Standard Not Needed

Another advantage of the multifunction 4271A is that it can be calibrated to make precise inductance measurements using precision standard capacitors, which are considerably better than available inductance standards. As Fig. 2 shows, the only difference between a C-G measurement and an L-R measurement is the way the unknown and the standard resistor R_s are connected. Thus if R_s and the measurement section are calibrated using standard capacitors, L-R measurements are also calibrated except for the error caused by changing the connections of R_s and the unknown. This is a constant systematic error that can be determined for a given instrument and corrected for, thereby making L-R measurements as precise as C-G measurements.

Accessories

Several accessories are available for the Model 4271A Digital LCR Meter.

Model 16022A Test Fixture is a general-purpose fixture for connecting unknowns to the 4271A. The fixture is supplied with inserts for both vertical and axial lead connections.

Model 16021A Test Fixture is used to calibrate the 4271A with a standard capacitor or a standard resistor that has a GR-900 connector.

Model 16032A Test Leads (with BNC connectors) are used for connecting the 4271A to a four-terminal device or user-manufactured test fixture that has BNC connectors.

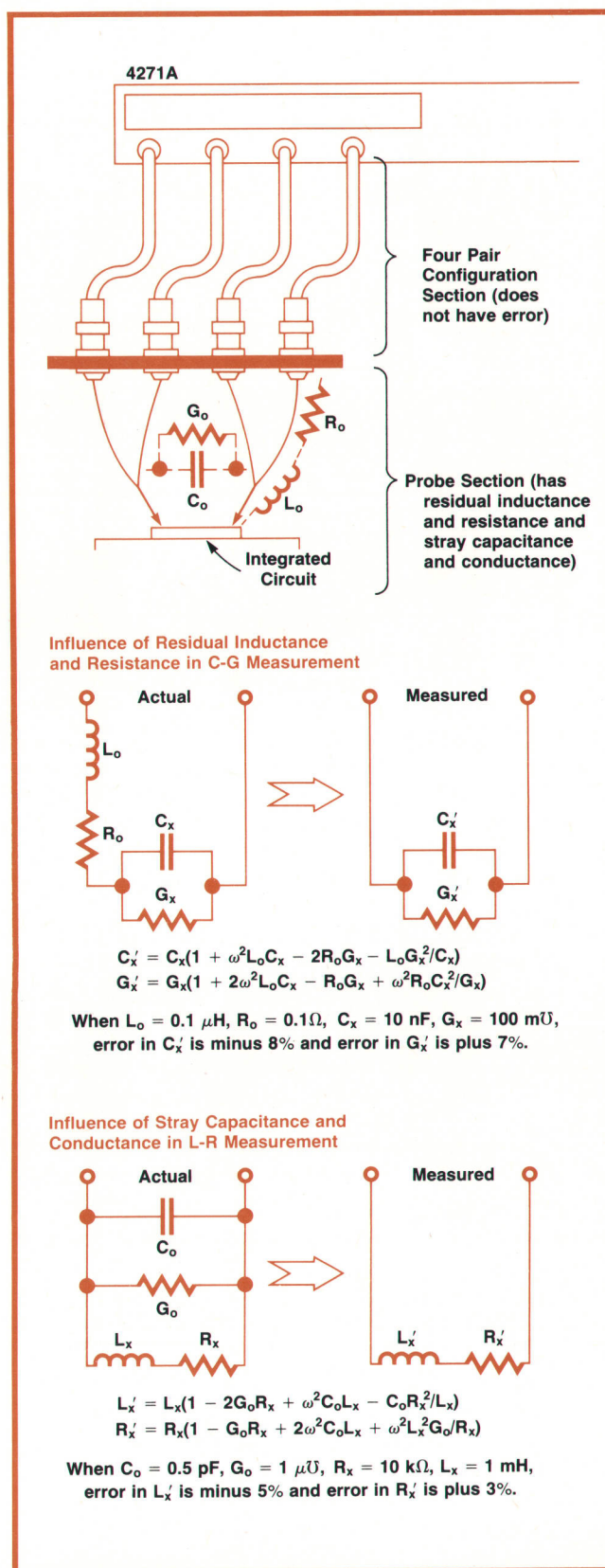


Fig. 6. When probes are used to measure IC components, parasitic L and C in the probe section influence the measurements. The 4271A LCR Meter has offset adjustments to minimize these errors.

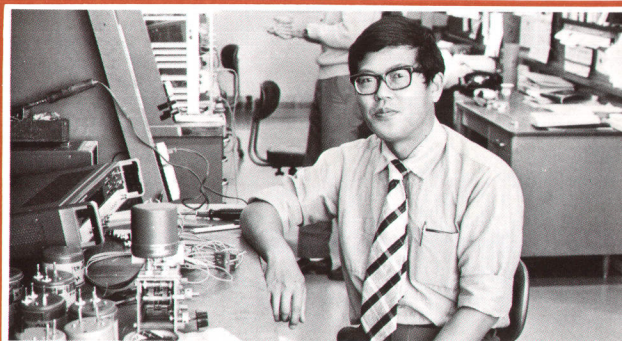
Model 16033A Test Leads (with small coaxial connectors) are used for connecting the 4271A unknown terminals to a sample to be measured.

Model 16023A DC Bias Controller is used to control the internal bias source provided by Model 4271A Option 001.

Acknowledgments

The project required a real team effort. The team members included Hitoshi Noguchi, who directed the project, Takeshi Kyo and Satoru Hashimoto, who designed the logic circuits, Toshio Tamamura and Nori Sugihara, who designed the bridge section, Toshio Manabe and Kiyohiko Hagino, who did the mechanical design, and Kazu Shibata, who handled the industrial design.

I especially want to thank Art Fong for recommending this project and outlining the requirements.



Kohichi Maeda

Kohichi Maeda received his BSEE from Waseda University, Tokyo, in 1963. After a year of working at Yokogawa Electric Works as an R&D engineer, he joined Yokogawa-Hewlett-Packard in 1964. At YHP he worked on the design of the Model 4204A Digital Oscillator and the Model 4270A Automatic Capacitance Bridge, then joined the 4271A Digital LCR Meter project. He developed the measurement-section circuitry for the 4271A. In his spare time Kohichi likes to play tennis.

BASIC SPECIFICATIONS

HP Model 4271A LCR Meter

Capacitance Measurement

FULL SCALE RANGES:

| | Range | Capacitance | Conductance | Dissipation Factor* |
|--------------------|-------|-------------|-----------------------|---------------------|
| Full Scale Display | 1 | 10.000 pF | 100.00 μS | 1.0000 |
| | 2 | 100.00 pF | 10000.0 μS | |
| | 3 | 1000.0 pF | 10.000 mS | |
| | 4 | 10.000 nF | 100.00 mS | |
| Over-range | 1-4 | 90% | 90% | 60% |

*When reading of C is more than 1500 counts.

OFFSET RANGE:

CAPACITANCE: 1 pF

CONDUCTANCE: 1 μS

FREQUENCY: 1 MHz $\pm 0.01\%$.

CAPACITANCE ACCURACY:

| Range | Test Sig Level High $\pm(\% \text{ of reading} + \text{counts})$ | Test Sig Level Low $\pm(\% \text{ of reading} + \text{counts})$ |
|-------|---|--|
| 1 | 0.1+7 | 0.2+8 |
| 2 | 0.1+3 | 0.2+4 |
| 3 | 0.1+3 | 0.2+3 |
| 4 | 0.4+3 | 0.4+3 |

CONDUCTANCE ACCURACY:

| Range | Test Sig Level High $\pm(\% \text{ of reading} + \text{counts})$ | Test Sig Level Low $\pm(\% \text{ of reading} + \text{counts})$ |
|-------|---|--|
| 1 | 0.2+(7+NC/1000) | 0.3+(7+2NC/1000) |
| 2 | 0.2+(3+NC/1000) | 0.3+(3+2NC/1000) |
| 3,4 | 1.2+(2+2NC/1000) | 1.2+(2+2NC/1000) |

DISSIPATION FACTOR ACCURACY:

| Range | Test Sig Level High $\pm(\% \text{ of reading} + \text{counts})$ | Test Sig Level Low $\pm(\% \text{ of reading} + \text{counts})$ |
|-------|---|--|
| 1 | 1.0+(10+20,000/NL) | 1.0+(15+30,000/NL) |
| 2,3 | 1.0+(10+10,000/NL) | 1.0+(15+20,000/NL) |
| 4 | 1.0+(15+30,000/NL) | 1.0+(15+30,000/NL) |

Accuracy specifications apply 23°C $\pm 5^\circ\text{C}$.

On Range 4, Test Sig Level is Low only.

NC is capacitance readout in counts.

MEASURING SPEED

FIXED RANGE: 100 ms to 250 ms in C-G and L-R measurements. 160 ms to 400 ms in C-D and L-D measurements.

AUTORANGE: 100 ms/range step added to above values.

DC BIAS VOLTAGE

INTERNAL: optionally provided, variable from 00.0 V to 39.9 V in steps of 0.1 V, controlled by HP calculators, HP computers or HP 16023A DC Bias Controller.

EXTERNAL: input connector is provided on the rear panel, ± 200 V maximum.

MONITOR: terminal on rear panel.

Inductance Measurement

FULL SCALE RANGES:

| | Range | Inductance | Resistance | Dissipation Factor* |
|--------------------|-------|----------------------|-------------------|---------------------|
| Full Scale Display | 1 | 1000.0 nH | 10.000 Ω | 1.0000 |
| | 2 | 10.000 μH | 100.00 Ω | |
| | 3 | 100.00 μH | 1000.0 Ω | |
| | 4 | 1000.0 μH | 10.000 k Ω | |
| Over-range | 1-4 | 90% | 90% | 60% |

*When reading of L is more than 1500 counts.

OFFSET RANGE:

INDUCTANCE: 100 nH

RESISTANCE: 100 m Ω

FREQUENCY: 1 MHz $\pm 0.01\%$.

INDUCTANCE ACCURACY:

| Range | Test Sig Level High $\pm(\% \text{ of reading} + \text{counts})$ | Test Sig Level Low $\pm(\% \text{ of reading} + \text{counts})$ |
|-------|---|--|
| 1 | 1.0+15 | 1.0+15 |
| 2 | 0.6+4 | 0.6+6 |
| 3,4 | 0.2+4 | 0.3+6 |

RESISTANCE ACCURACY:

| Range | Test Sig Level High $\pm(\% \text{ of reading} + \text{counts})$ | Test Sig Level Low $\pm(\% \text{ of reading} + \text{counts})$ |
|-------|---|--|
| 1 | 1.2+(8+2NL/1000) | 1.2+(8+2NL/1000) |
| 2 | 1.2+(2+2NL/1000) | 1.2+(2+2NL/1000) |
| 3,4 | 0.2+(2+2NL/1000) | 0.3+(2+2NL/1000) |

DISSIPATION FACTOR ACCURACY:

| Range | Test Sig Level High $\pm(\% \text{ of reading} + \text{counts})$ | Test Sig Level Low $\pm(\% \text{ of reading} + \text{counts})$ |
|-------|---|--|
| 1 | 1.0+(20+30,000/NL) | 1.0+(20+30,000/NL) |
| 2,3 | 1.0+(15+10,000/NL) | 1.0+(20+20,000/NL) |
| 4 | 1.0+(15+20,000/NL) | 1.0+(15+30,000/NL) |

Accuracy specifications apply 23°C $\pm 5^\circ\text{C}$.

On Range 1, Test Sig Level is Low only.

NL is inductance readout in counts.

General

DATA OUTPUT AND REMOTE PROGRAMMING INPUT: see data sheet.

OPERATING TEMPERATURE: 0°C to 50°C.

POWER: 100/120/220/240 V $\pm 10\%$, 48-66 Hz, 80 VA max.

DIMENSIONS: 3-15/32" \times 16 3/4" \times 19-9/16" (89 \times 425 \times 498 mm)

WEIGHT: net, 22 lbs (10 kg).

PRICE IN U.S.A.: 4271A 1 MHz Digital LCR Meter, \$4500.

MANUFACTURING DIVISION: YOKOGAWA HEWLETT-PACKARD LTD.

9-1 Takakura-cho
Hachioji-shi
Tokyo 192, Japan

A Moderately Priced 20-MHz Pulse Generator with 16-Volt Output

Suitable for general-purpose use, this instrument also has the pulse amplitude needed for driving the newer CMOS circuits.

by Günter Krauss and Rainer Eggert

THE PULSE GENERATOR and oscilloscope have long been an effective team for studying the behavior of circuits in the time domain, comparable to the way sweep oscillators and spectrum analyzers are now teamed for analyses in the frequency domain. Much information can be gleaned from the effects that circuits under test have on pulse shape.

Nowadays, the pulse generator assumes increasing importance among the basic test instruments as the use of digital circuits spreads to an ever-broadening range of applications. The pulse generator is useful as a source of variable-rate clock pulses and gating signals for testing these circuits.

It was with the thought of designing an inexpensive pulse generator with output amplitudes suitable for driving the increasingly popular CMOS circuits that we embarked on the design of the new Model 8011A Pulse Generator (Fig. 1). In meeting the requirements of CMOS circuits, it would also span the amplitude range needed for TTL, HTL, and HNIL circuits. It soon became apparent, however, that this instrument's moderate price would make it attractive as a general-purpose pulse generator for a broad range of users, including technical schools and service shops as well as the design lab.

The new pulse generator has a maximum pulse out-



Fig. 1. Model 8011A Pulse Generator produces 16-volt pulses at repetition rates up to 20 MHz. It has convenient pushbutton operation.

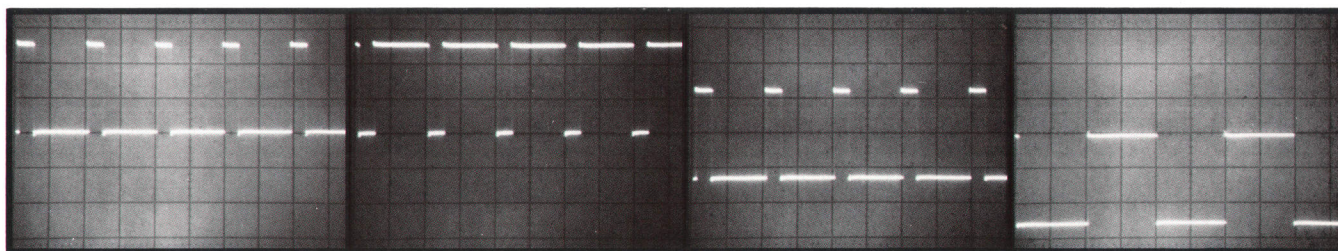


Fig. 2. Oscillograms show various modes of operation of Model 8011A Pulse Generator. From the left these are (a) positive-going, (b) positive-complementary, (c) symmetrical, and (d) negative square wave. The center CRT graticule line represents ground potential in all photos. (Oscilloscope sweep time is 2 μ s/cm and vertical deflection factor is 5 volts/cm.)

put of 16 volts, and this can be a positive-going pulse train, a negative-going pulse train or symmetrical pulses (baseline at $-8V$, pulse top at $+8V$, or vice versa). Having a range of pulse amplitudes from $\pm 0.25 V$ up to $\pm 16 V$, this instrument can be used to test designs involving CMOS circuits as well as TTL, HTL, and HNIL logic families that use either positive or negative power supplies. The relatively large pulse amplitude is also useful for testing operational amplifiers and other analog circuits.

Amplitude with Speed

Pulses can be obtained at any repetition rate up to 20 MHz, enabling the broad range of digital circuits to be driven at their operational speed. The instrument has an internal rate generator that has a range of 0.1 Hz to 20 MHz. It can also work with external triggers. A front-panel pushbutton or the external trigger

input can be used for single pulses.

The 16-volt amplitude can be obtained across a 50-ohm load, the instrument being capable of supplying up to 320 mA into any load of 50 ohms or less, even a short circuit, from a high source impedance.

A switchable 50-ohm termination is provided for those cases where pulse energy reflected back into the instrument would cause pulse distortion unless absorbed by the internal termination. This arrangement gives a maximum pulse amplitude of 8 volts when driving a 50-ohm load, or up to 16 volts with a high-impedance load. Whenever the instrument is switched to amplitude ranges below 4 volts, the 50-ohm termination is automatically switched in.

Pulse transition times are fixed at less than 10 ns in all modes of operation. The pulse baseline is at ground, except in the symmetrical mode (Fig. 2). The instrument can also operate in the pulse complement



Fig. 3. Model 8011A Pulse Generator fitted with counted-burst option.

mode; the pulse is then inverted but amplitude and polarity remain the same. This permits simple conversion from a positive logic convention to a negative logic convention or vice-versa.

The range of pulse widths is from 25 ns to 100 ms. Wider pulse widths can be achieved by using the pulse complement mode or the squarewave mode, which gives a pulse width of one half the pulse period at all repetition rates up to 10 MHz.

Provisions for pulse delay, variable dc offset, and gated operation were omitted to keep the price as low as possible. However, available as an option is "counted burst" operation, similar to that available with the Model 8015A Pulse Generator.¹ This enables the Model 8011A to deliver a defined number of pulses to the load upon receipt of a trigger, the number of pulses—up to 9999—being selected by a 4-decade thumbwheel switch (Fig. 3). The burst may also be initiated by pressing the manual pushbutton. Unlike the more familiar gated mode of operation, the number of pulses delivered is always the selected number regardless of any changes in pulse repetition rate, whether deliberate or incurred as a result of period jitter.

One of the useful characteristics of the counted-burst mode is that following the burst, additional single pulses can be generated with the front-panel pushbutton (Fig. 4). With this arrangement, a digital processor can be clocked rapidly to a particular step with the counted burst, and then stepped along one clock at a time with the pushbutton, allowing the designer time to check the states of the processor.

Inner Workings

A block diagram of the new pulse generator is shown in Fig. 5. It follows traditional practice in that a rate generator that is either self-triggered or externally triggered supplies pulses to a width generator that in turn drives an output stage.

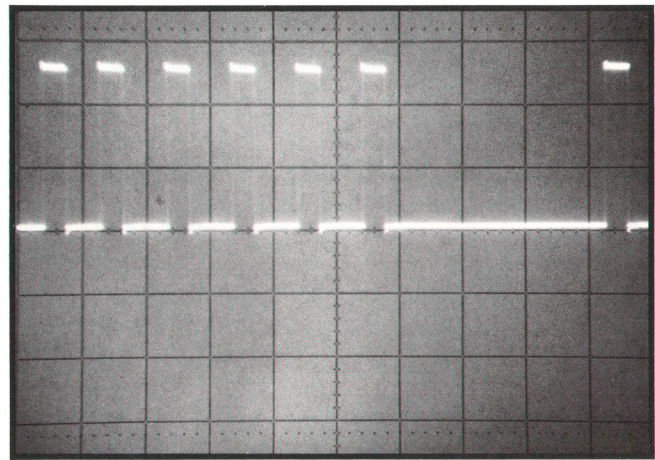


Fig. 4. With counted-burst option, a predetermined number of pulses is generated. Additional single pulses can be added under pushbutton control.

If the instrument is equipped and operating with the optional burst mode, the rate generator is inhibited by the burst-control circuit. An input trigger initiates a burst by causing the burst control to enable the rate generator. The burst control has a counter that is preset to the number selected on the front-panel and decremented by the output of the rate generator. When its count reaches zero, it disables the rate generator terminating the burst.

The width generator is the basic pulse-forming circuit. Its output is inverted for negative-going pulses or for the complement of positive pulses. In the squarewave mode, the width generator is used as the source of trigger pulses for a flip-flop (divider) that generates square waves at one-half the rate generator frequency. The selector, actually a group of gates, selects the appropriate signal for the output amplifier and also for the output trigger generator.

The output trigger generator supplies constant amplitude, constant width, positive-going pulses (1 volt high, 20 ns wide) to external circuits. It pro-

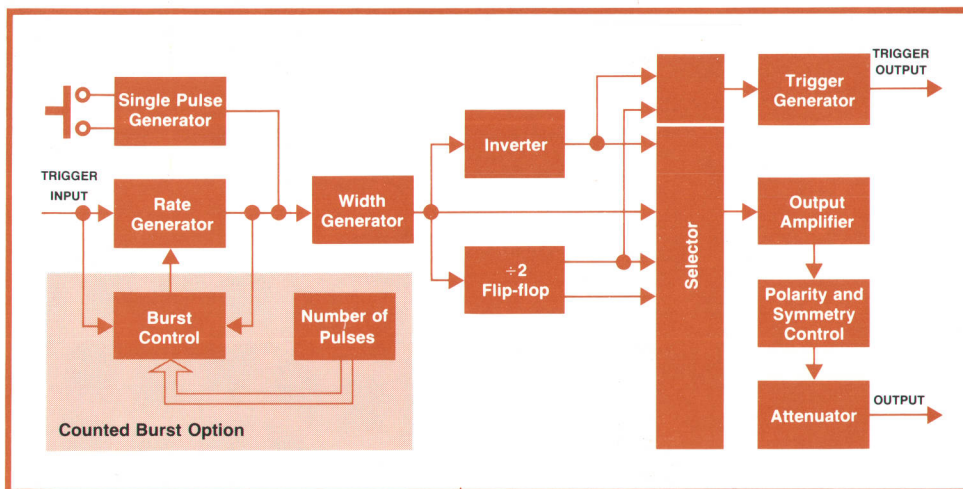


Fig. 5. Simplified block diagram of Model 8011A Pulse Generator.

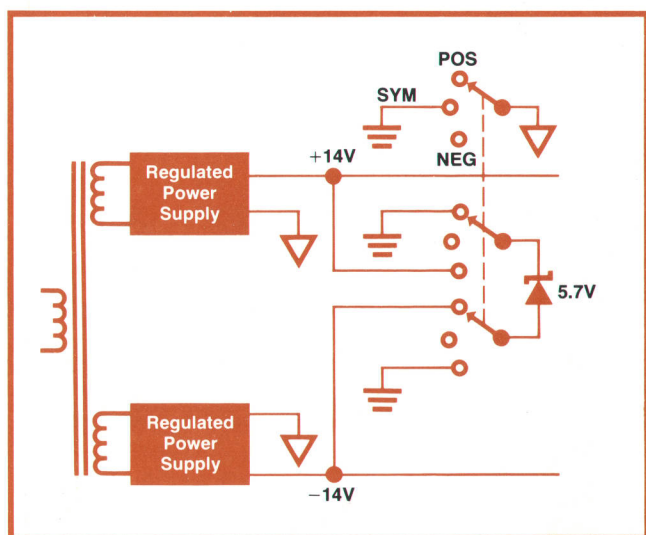


Fig. 6. Floating circuit ground is biased with respect to chassis ground, enabling the use of low voltage circuits to generate large pulses of either polarity.

duces one trigger for each pulse generated by the width circuit, except in the squarewave mode when it generates one trigger for every complete squarewave cycle (one trigger pulse for every two rate generator pulses).

Switched Grounds

A key factor in making possible the generation of either positive or negative 16-volt pulses with minimum cost was the use of a floating ground for all circuits except those concerned with the front-panel connectors. The circuits were designed to operate with +14V and -14V power supplies and the output stage has a push-pull configuration that swings the output waveform between -8V and +8V with respect to the floating ground.

As shown in Fig. 6, circuit ground connects directly to chassis ground when the instrument is in the symmetrical mode but it is offset +8.3V in the positive mode or -8.3V in the negative mode. In either of these modes, offset current is added to the output to fix the pulse baseline precisely at chassis ground. It is thus possible to derive +16V or -16V pulses using a $\pm 14V$ power supply, reducing overall power dissipation and, in keeping with the design philosophy, allowing lower cost, low-voltage transistors to be used.

The use of the floating power supply dictated the choice of circuits associated with the front-panel connectors. External triggers are fed to a Schmitt trigger that has its transistor bases referenced to chassis ground, but the circuit output is taken from a collector connected to the +14V supply through a resistive load. Switching the floating ground with respect to the chassis ground affects only the collector-base

bias, and this has relatively minor effect on circuit performance.

Similarly, the trigger output is taken from a cascode circuit that has its load resistor returned to chassis ground while the remaining elements of the circuit are referenced to the floating ground. Here again, switching the floating ground affects only the collector-base bias of the output transistor, with little effect on circuit performance. Normally, the circuit is cut off, so the trigger output rests at chassis ground between pulses.

Wide-Range Rate Generator

The rate generator is a ramp generator that uses a constant-current source to charge a capacitor. A current switch discharges the capacitor when it reaches a particular amplitude, starting a new cycle. The discharge is initiated by a Schmitt trigger circuit that responds to the ramp amplitude. The pulse period range is selected by switching capacitors. Vernier control is obtained by adjusting the magnitude of the constant current.

Low repetition rates, useful for subsonic work or for incremental stepping of a digital circuit, present problems to this type of rate generator. The leakage current of a capacitor large enough for a 10-second pulse period would be an appreciable part of the charging current. This would cause ramp nonlinearities and timing instabilities. Reducing the charging current to permit use of a smaller capacitor is not a satisfactory solution because other leakage currents would then be comparable to the charging current.

This problem was surmounted by using a "capacitance multiplying" technique* when the instrument

*Developed for the HP Model 3310A Function Generator by Stephen B. Venzke of HP's Loveland Division.

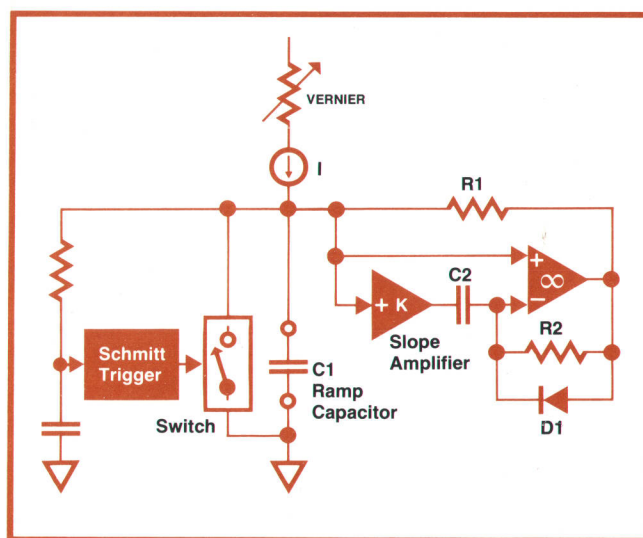


Fig. 7. Rate generator uses a capacitance-multiplying technique to extend the range of the slowest repetition rate.

is switched to the longest pulse period range. This circuit, shown in the diagram of Fig. 7, withdraws 98% of the charging current but in so doing it reduces proportionally the effects of any leakage currents.

With reference to Fig. 7, current is drawn off through resistor R1, which is connected around the operational amplifier in a bootstrap configuration that maintains a constant voltage across R1. The voltage maintained, however, is adjusted by the differentiating action of capacitor C2 and resistor R2. Because of the differentiating action, the voltage at the amplifier's inverting input is affected by the slope of the ramp, and this adjusts the voltage across R1 to maintain ramp linearity.

It can be shown (see appendix) that the current ratio is:

$$\frac{i_2}{i_1} = (K - 1) \frac{C2}{C1} \times \frac{R2}{R1}$$

where K is the gain of the slope amplifier. In the Model 8011A, circuit constants were chosen to make $i_2 = 54i_1$, so the charging current i_1 is 1/55 of I, the constant current. Thus, with respect to the constant-current source, capacitor C1 seems to be 55 times as large as it actually is and any leakage currents that add to or subtract from I are reduced by a factor of 55. Capacitor C2 charges during the ramp so it must be discharged at the time the ramp resets. During reset, a low-impedance discharge path is provided by diode D1.

When the instrument is operated with externally supplied triggers, the ramp charging current is disconnected. The Schmitt trigger circuit then serves to generate uniform triggers in response to positive-going input signals.

The width generator is also a ramp generator but it uses a flip-flop to control the discharge switch. A trigger sets the flip-flop, initiating a new ramp. When the ramp voltage reaches the turn-off level, it resets the flip-flop, terminating the ramp. The rectangular output of the flip-flop is used to drive the output stage.

The Shape of Things

Total instrument power dissipation was kept below 70 W by the economical use of the internal power supplies. Thus, sufficient radiating surface, to keep internal heat at a reasonable level, can be obtained with a half-rack module. The package size provides room for all circuits to be on one double-sided board, reducing test and assembly time (Fig. 8). Hand wiring is needed only to connect the front-panel potentiometers and the ac input line. The board can be tested by automatic means before final assembly.

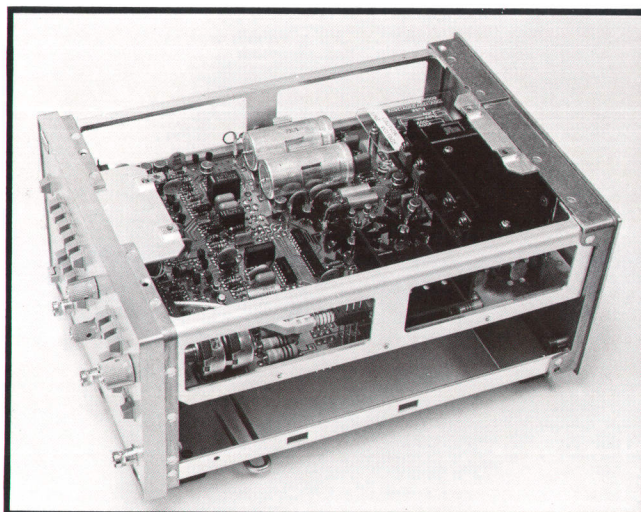


Fig. 8. All components, including the pushbutton switches, power transformer, fuses, and ac voltage selector switches, are contained on a single printed-circuit board.

Acknowledgments

The authors wish to express appreciation to our production engineer, Roland Koch, to Margaret Ulmann and Henriette Werner for the circuit board layout, and to Group Leader Gunther Zieher for advice and counsel. We also wish to acknowledge the contributions of Gerd Globas and Theo Papatheodorou to the design of the counted burst option.

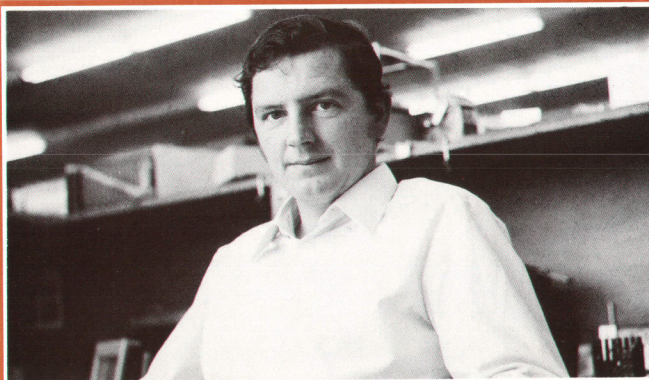
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Günter Krauss

At twenty-three, Günter Krauss is possibly the youngest project leader at HP, a responsibility he undertook after completing some investigative projects. Günter joined HP in early 1972 after qualifying as an Ingenieur Graduate in Electronics at the Fachschule in Furtswangen. In his spare time he enjoys skiing, tennis, jazz, and karting, training in a friend's go-kart with the hope of racing competitively during the 1974 season (fuel permitting). Günter is also a part-time member of the HP GmbH R&D soccer team.



Rainer Eggert

Rainer Eggert joined HP eight years ago to work in tool design but he later switched to product design, contributing to the mechanical design of several products, including the 8002A, 8003A, and 8008A Pulse Generators. Prior to HP, he had been an apprentice at a lighting equipment firm before attending the Staatliche Ingenieurschule Gauss in Berlin where he qualified as a Feinwerktechnik Ingenieur Graduate (mechanical engineer). Married, and the father of two pre-school youngsters, Rainer is also interested in coin-collecting, having some 300 coins dating as far back as 1597. He also enjoys cinephotography, skiing, and ice-skating.

SPECIFICATIONS HP Model 8011A Pulse Generator

Pulse Characteristics

TRANSITION TIMES: <10ns fixed.

OVERSHOOT, RINGING AND PRESHOOT: <±5% of pulse amplitude. May increase to 10% at counter-clockwise position of amplitude vernier.

PULSE WIDTH: 25ns to 100ms in four ranges. Vernier provides continuous adjustment within each range.

WIDTH JITTER: <0.1% + 50ps of any width setting.

MAXIMUM DUTY CYCLE: >50% (100% using pulse complement).

MAXIMUM OUTPUT: 8V with 50Ω source and load impedances. 16V, with internal 50 ohms and external high impedance or, with internal high impedance and external 50Ω.

ATTENUATOR: three-step attenuator provides ranges 0.25V - 1V - 4V - 16V. Vernier provides continuous adjustment within each range.

SOURCE IMPEDANCE: 0.25V - 1V - 4V ranges, 50 ohms ± 10% shunted by 30pF. 4V - 16V range, 50 ohms or high impedance switch selectable.

POLARITY: positive, negative or symmetrical, switch selectable.

FORMAT: normal or complement, switch selectable.

Repetition Rate and Trigger

REPETITION RATE: 0.1 Hz to 20 MHz in 5 ranges. Vernier provides continuous adjustment within each range.

PERIOD JITTER: < 0.1% + 50ps of any period setting.

SQUARE WAVE: 0.05 Hz to 10 MHz.

TRIGGER OUTPUT: dc-coupled 50-ohm (typ) source delivering ≥ +1V across 50-ohm load.

TRIGGER PULSE WIDTH: 20ns ± 10ns.

Externally Controlled Operation

EXTERNAL INPUT

TRIGGER POLARITY: positive.

SENSITIVITY: 1V

MAXIMUM INPUT: ±5V.

INPUT IMPEDANCE: 50 ohms.

MANUAL: front panel pushbutton for generating single pulse.

EXTERNAL TRIGGERING

REPETITION RATE: 0 to 20 MHz. In square wave, output frequency is half of input frequency.

TRIGGER SOURCE: manual or external signal. Minimum external signal width, 10ns.

PULSE BURST MODE (Option 001): preselected number of pulses generated on receipt of trigger.

BURST TRIGGER SOURCE: external signal or manual. Minimum external signal width, 25ns.

General

OPERATING TEMPERATURE RANGE: 0°C to 55°C.

POWER: 100V, 120V, 220V or 240V, +5%, -10% 48 Hz to 440 Hz, 70 VA max.

WEIGHT: 9 lbs (4 kg).

DIMENSIONS: 7.9 in. W × 5.6 in. H × 11.8 in. D (200 × 142 × 300mm).

PRICES IN U.S.A.:-

8011A, Pulse Generator, \$435.

Pulse Burst Option (001), add \$300.

MANUFACTURING DIVISION: Hewlett-Packard GmbH
Herrenberger Strasse 110
D-7030 Boblingen, Wurtemberg
Germany

Appendix A Capacitance Multiplier

Current I from a constant-current source is partitioned into currents i_1 and i_2 , as indicated in the diagram.

It will be shown here that the ratio of i_2 to i_1 is a function of the circuit constants C_1 , C_2 , R_1 , R_2 and K .

$$\text{First of all, } i_1 = C_1 \frac{dV_1}{dt}.$$

The voltage V_2 at the output of the slope amplifier is KV_1 , so

$$\frac{dV_2}{dt} = K \frac{dV_1}{dt}$$

Now, because feedback through R_2 causes the negative input of the operational amplifier to track the positive input, the slope of the voltage across capacitor C_2 is:

$$\frac{dV_2}{dt} - \frac{dV_1}{dt} \text{ or } (K - 1) \frac{dV_1}{dt}$$

The current i_3 into capacitor C_2 thus equals $C_2 (K - 1) \frac{dV_1}{dt}$. By inspection, it is seen that the operational amplifier output voltage, V_3 , equals $V_1 - i_3 R_2$ or,

$$V_3 = V_1 - C_2 (K - 1) \frac{dV_1}{dt} R_2.$$

We can now determine the current i_2 , which is $(V_1 - V_3)/R_1$ or,

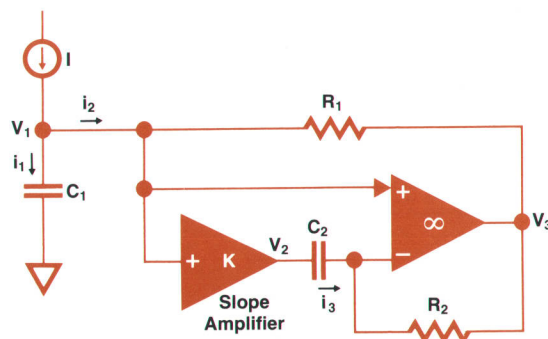
$$i_2 = \frac{V_1 - V_1 + C_2 (K - 1) \frac{dV_1}{dt} R_2}{R_1} = \frac{C_2 (K - 1) \frac{dV_1}{dt} R_2}{R_1}$$

Thus, the ratio of i_2 to i_1 is:

$$\frac{i_2}{i_1} = \frac{C_2 (K - 1) \frac{dV_1}{dt} R_2}{C_1 \frac{dV_1}{dt}} = (K - 1) \frac{C_2}{C_1} \times \frac{R_2}{R_1}$$

(In the Model 8011A Pulse Generator, R_2 is in parallel with a diode that becomes forward-biased during ramp reset, reducing the effective magnitude of R_2 .)

— Stephen B. Venzke



Laboratory Notebook

Logarithmic Amplifier Accepts 100-dB Signal Range.

In numerous measurement situations—swept-frequency tests of filter response, for example—the signal level may change rather abruptly. To accommodate a wide amplitude range when making plots of frequency response, the detector often works with a logarithmic amplifier so low amplitude regions can be expanded to disclose detail. At the same time, high-amplitude levels are compressed to prevent the related recorder or display device from being driven off scale.

Even so, the amplitude range that can be accommodated by such an arrangement has commonly been 40-60dB, with 80dB being something of a special achievement. But now, with the growing sophistication of hybrid thin-film and monolithic integrated circuit technologies, it is possible to design a logarithmic amplifier with an input range of over 100 dB, and build it at modest cost.

This amplifier makes use of the exponential voltage-current relationship of a transistor, as shown in Fig. A. With the input voltage plotted on a logarithmic (dB) scale, the transfer curve is reasonably linear over an input amplitude range of -35 to -25 dBV.

Twelve differential pairs of the type shown in Fig. A are used with their outputs connected to a common output but with their inputs driven from separate taps on an attenuator-amplifier ladder, as shown in Fig. B. At any signal level, one differential pair is in its linear range while those at higher level

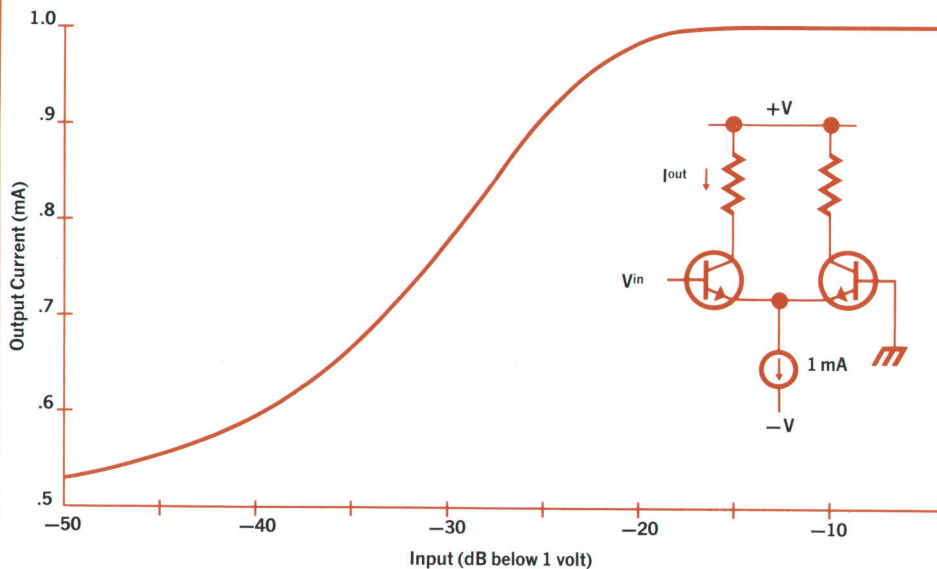
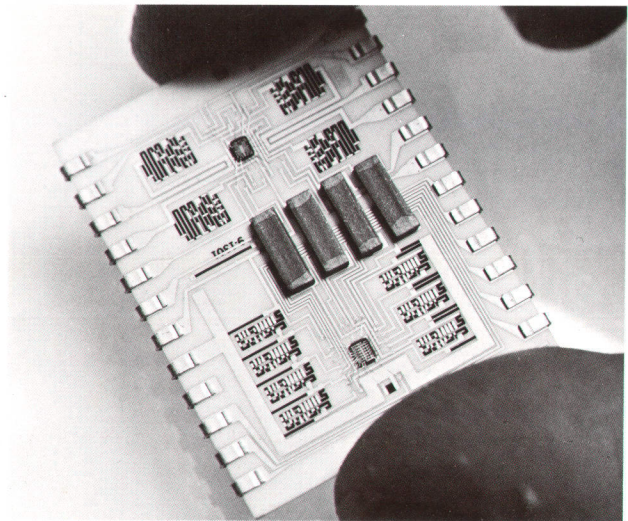


Fig. A. Voltage-current transfer curve of a single differential pair amplifier stage is reasonably linear over a 10-dB range when the input is plotted in dB units.

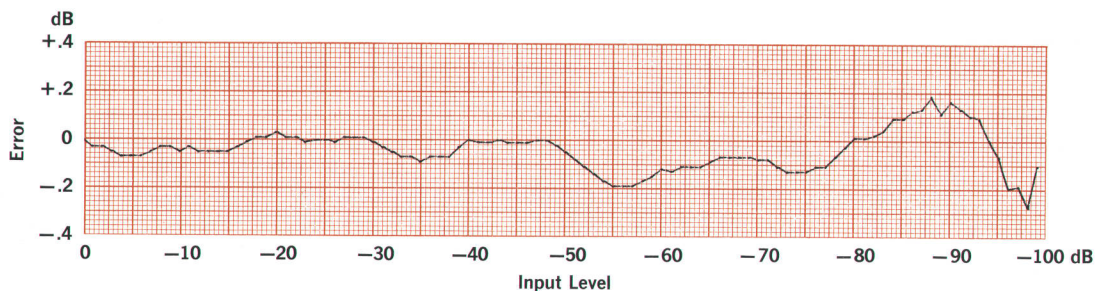


Fig. C.

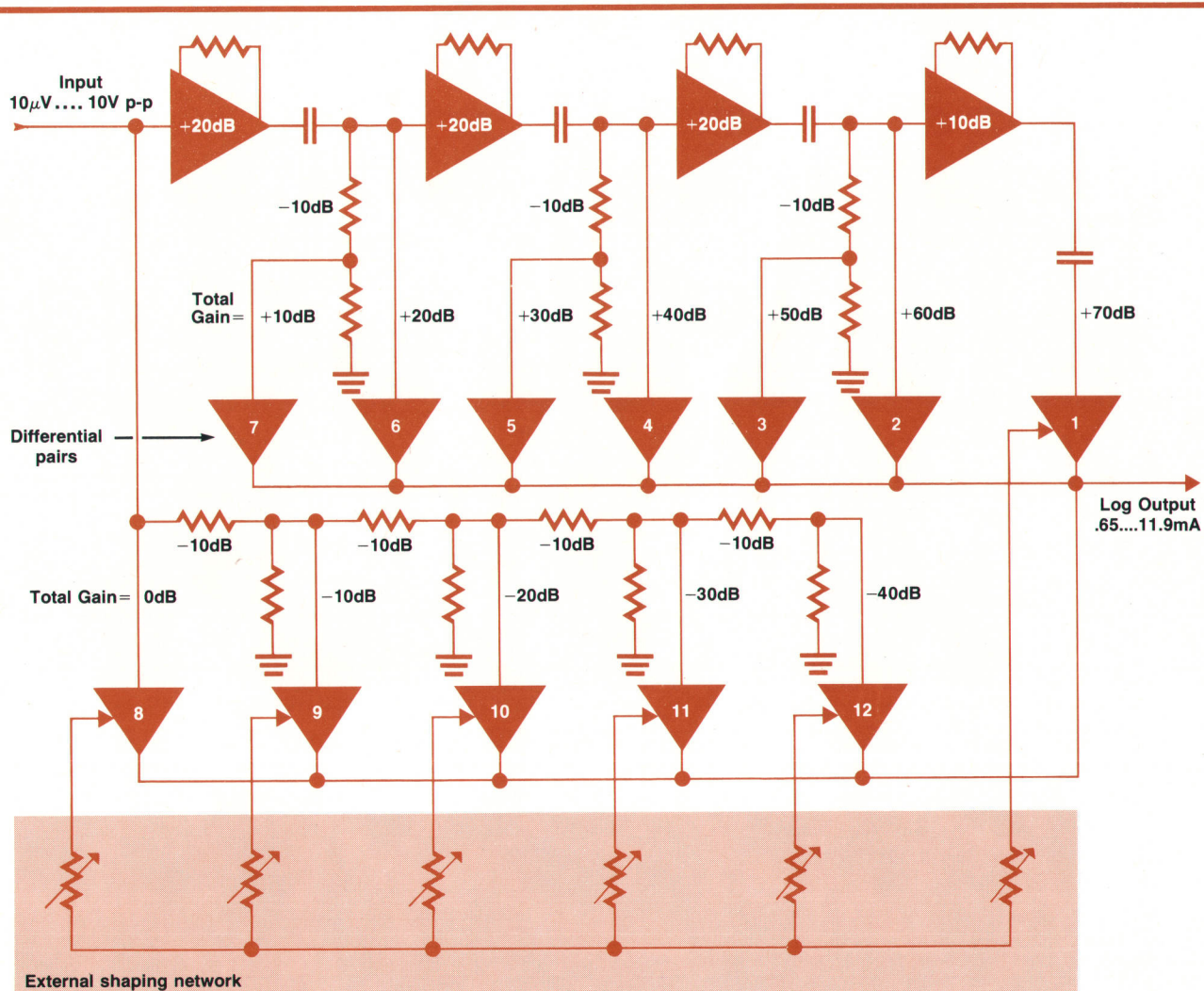


Fig. B.

taps are saturated and those at lower level taps are for all practical purposes cut off. As the signal grows larger the differential pair on the next lower tap moves into its linear range as the pair on the currently active tap saturates.

Although the basic concept is not new, it required the new technologies to obtain accuracy over such a wide amplitude range. All the differential pairs are fabricated on one monolithic chip, assuring good stage matching. A computer-controlled laser trims the thin-film resistors on the hybrid substrate to assure accurate 10-dB steps in the attenuator-amplifier ladder.

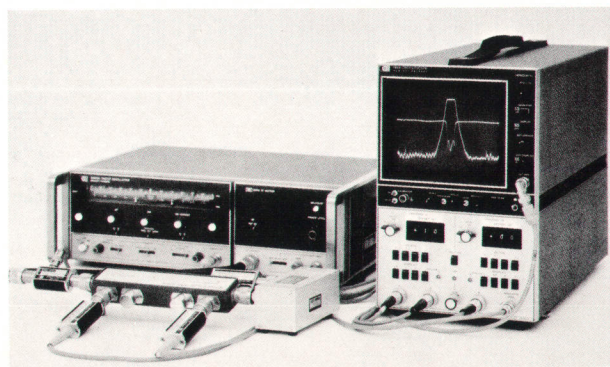
Typical linearity is shown in Fig. C. Compensation for the non-linear characteristics of a detector or other device can be obtained by adding a shaping network, also shown in Fig. B, to alter the transfer characteristics of individual stages.

This log amplifier is used in the Model 3570A Network Analyzer¹ and the Model 8755A Frequency-Response Measuring set.² Much helpful advice was provided by Hugo Vifian of HP's Santa Rosa Division, project leader for the 8755A, and Paul Thomas of the Loveland Instrument Division, analog circuit designer for the Model 3570A.

—Robert Jeremiasen
Loveland Instrument Division

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2. G. Nelson, P. Thomas, R. Atchley, "Faster Gain-Phase Measurements with New Automatic 50 Hz-to-13 MHz Network Analyzers," *Hewlett-Packard Journal*, October 1972.



Model 8755L Frequency Response Test Set uses the logarithmic amplifier described here.

Versatile VHF Signal Generator Stresses Low Cost and Portability

This high-quality, reliable instrument fits the budget and the performance needs of a variety of users.

by Robert R. Hay

SIGNAL GENERATORS ARE AMONG the electronics industry's most fundamental and widely used instruments. So varied are their applications that no single instrument can meet the needs of all users. For example, to characterize linear networks a signal generator must have high power, good flatness, and rapid tuning. To test narrowband receivers it must have a well calibrated, extremely low-level output signal (often less than 10^{-15} watts), along with low RF leakage, good stability, high-resolution tuning, and versatile modulation capabilities. A signal generator used as a local oscillator in a hetero-

dyne system must deliver moderate power (typically ten milliwatts) and have good settability, freedom from spurious signals, and good stability. If the same instrument is used for more than one application it needs some combination of these characteristics.

To meet the requirements of as many users as possible, HP has developed a line of general-purpose signal generators that have varying degrees of precision, stability, modulation capability, and other important characteristics. These are offered at prices commensurate with their abilities. Two of these generators, Models 8660 and 8640, have been de-



Fig. 1. Model 8654A is a compact, low-cost, solid-state signal generator that provides automatically leveled power from +10 to -130 dBm over a frequency range of 10 to 520 MHz. Amplitude and frequency modulation are independent and can be internal or external.

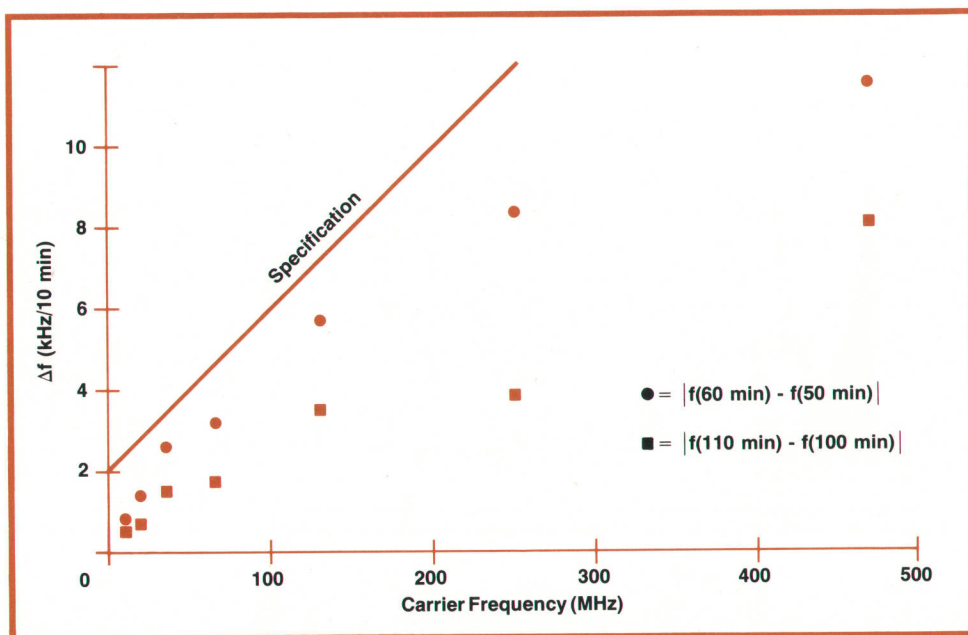


Fig. 2. Worst-case drift of ten typical 8654A production instruments.

scribed previously in these pages.^{1,2,3,4}

Model 8654A (Fig. 1), the latest in this line of signal generators, is the lowest-priced instrument in the series. Whereas the other generators were designed for applications that call for extreme precision, performance, and versatility, Model 8654A is designed with emphasis on lower cost and greater portability. Thus it is expected that the 8654A will be most useful for service and production applications, whereas the other generators may be better suited for applications in the laboratory and in automated and precision production testing.

Model 8654A provides calibrated output and versatile modulation capabilities for testing receivers, amplifiers, antennas, and filter networks over a car-

rier frequency range of 10 to 520 MHz. Effective RF shielding and wide output range permit receiver sensitivity measurements to be made down to levels of $0.1 \mu\text{V}$.

Internal oscillators provide either amplitude modulation or frequency modulation at 400 Hz or 1000 Hz. An external modulation signal from a conventional audio oscillator may also be used. The front-panel meter indicates amplitude modulation percentage.

Carrier output power is automatically leveled and is typically accurate within $\pm 1.5 \text{ dB}$ over the entire frequency range.* The power level is variable over a

*Level accuracy error consists of allowances for meter accuracy, detector linearity, temperature, flatness, attenuator accuracy, and twice the measurement error. The figure given here is typical. Specifications are more conservative (see page 24).

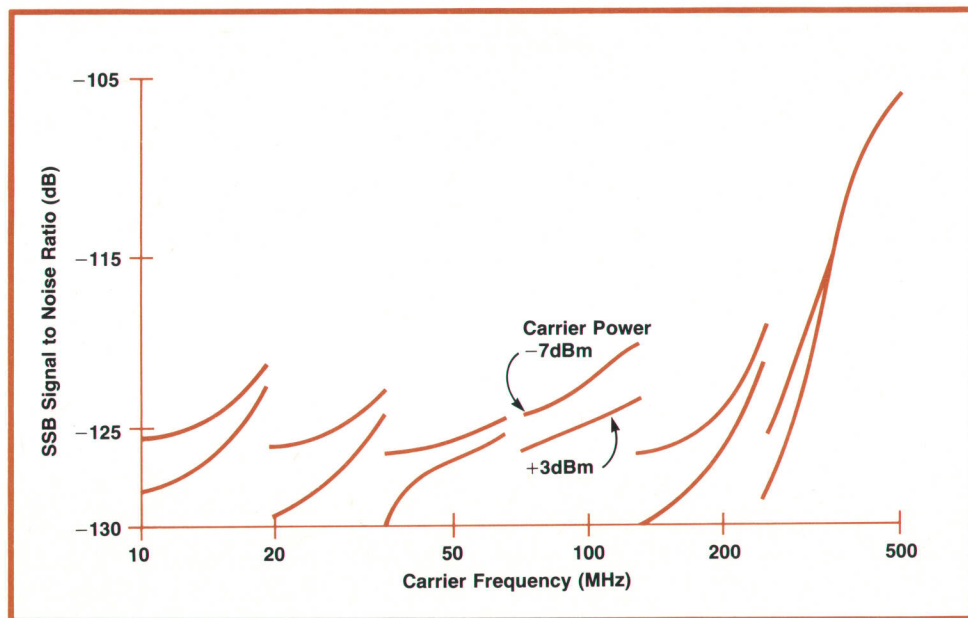


Fig. 3. Typical 8654A SSB signal-to-noise ratio. Total noise power relative to carrier power in a 1-Hz bandwidth 20 kHz from the carrier.

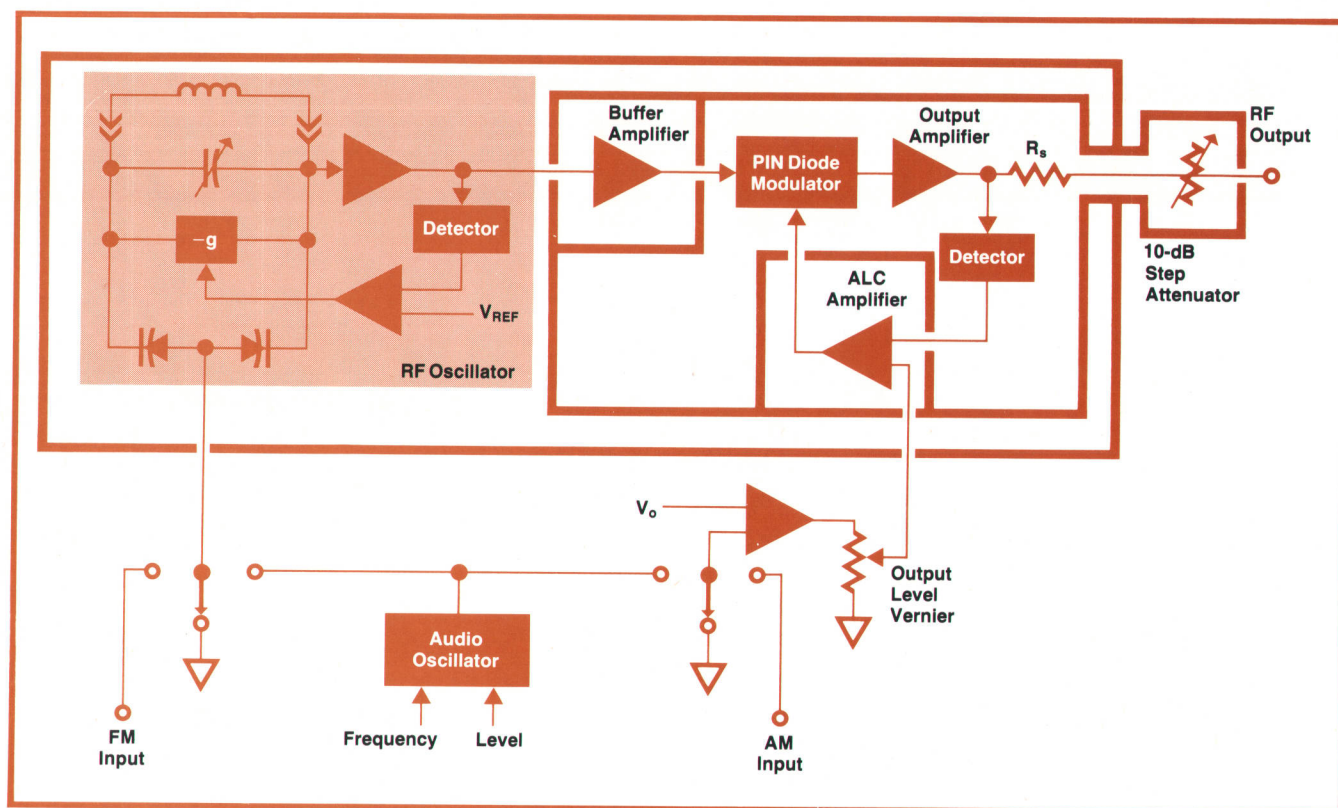


Fig. 4. 8654A simplified block diagram. Heavy lines show the major shielding required to assure low RF leakage and low incidental FM. The oscillator is tuned over its 10-to-520-MHz range by means of a variable capacitor and six fixed bandswitched inductors.

range of 140 dB. The front-panel meter displays the output power in dBm and volts.

An auxiliary uncalibrated RF output is also available at the rear panel for use with a counter or other external equipment. Auxiliary output level is typically greater than -7 dBm.

The 8654A has a specified stability of 1 kHz plus 0.002% of carrier frequency per 5-minute operating period after a two-hour warmup. It will typically recover specified stability within 5 minutes following a frequency change. Typical drift performance of production instruments is shown in Fig. 2.

Harmonic distortion is more than 20 dB below the carrier. Residual AM is more than 70 dB below the carrier and residual FM is less than 0.5 ppm, both in a post-detection bandwidth of 0.3 to 3 kHz.

Important in receiver measurements is the signal-to-noise power ratio in a 1-Hz bandwidth 20 kHz away from the carrier. This is shown in Fig. 3 for a typical instrument. (See page 21 for a discussion of noise specifications and their significance.)

Design Approach

Designing a low-cost instrument requires a number of decisions regarding the value of various features to the user. Inevitably, such an instrument will not be able to fulfill all needs, but it must be a

high-quality, reliable instrument that will provide high value to the user who is willing to sacrifice some degree of precision and performance in exchange for lower cost and greater portability.

To assure high reliability, the 8654's components are all solid-state. To provide economy and good performance, liberal use is made of Hewlett-Packard microwave transistors packaged for mounting on economical standard printed circuit boards.

The block diagram of the 8654A is shown in Fig. 4. The carrier signal is generated by the RF oscillator. The buffer amplifier, which has nearly unity gain, serves to isolate the oscillator from the modulator. A feedback loop consisting of the PIN diode modulator, the output amplifier, and the ALC amplifier controls the output level. The 10 dB step attenuator provides attenuation over a 120-dB range in 12 steps. To control the output level between the 10-dB steps, the ALC loop provides for more than 10 dB of continuously adjustable level variation.

Frequency modulation capability is provided by varactor diodes in parallel with the main tuning capacitor. These may be driven either by the internal audio oscillator or by an external input.

Amplitude modulation is achieved by varying the voltage supplied to the output-level vernier potentiometer. In this way the audio modulation signal is

Specifying Signal Generator Noise

Total carrier noise power is the sum of angular (phase or frequency) noise modulation and amplitude noise modulation. These two types of noise tend to be independent, and their effects on the system are generally different. For many signal generators, phase noise varies with the carrier frequency and amplitude noise varies with the carrier level.

We can classify noise as either amplitude fluctuations or phase fluctuations and according to the rate at which it occurs. Phase fluctuations can be characterized as either phase noise or frequency noise. In most cases it is easy to convert from one characterization to the other.

Drift

Noise at frequencies below the lowest information rate is generally characterized in the time domain as drift rate, rather than in the frequency domain as power spectral density. Frequency-domain measurements at these subaudio frequencies are rarely of much value and, because of the high resolution required, are difficult to make accurately. Drift rate, on the other hand, is a meaningful characteristic, for example to a user who is making a measurement for a short period of time on a narrow-band receiver. Drift is generally specified in Hertz or parts per million (or a combination of both, as for the 8654A) per 5 or 10 minutes. For some applications, drift may also be specified for a period of one day or longer. Fig. 2, page 19, shows typical 8654A drift performance.

Residual AM and FM

Noise within the information bandwidth is added to the information when the carrier is demodulated. It must be characterized as noise power from a demodulator equivalent to that being used in the communication system being tested. Consequently, it is important to specify both the amplitude noise and the angular noise that occurs at rates within the information bandwidth. For many applications, this is best done by describing the power spectral density of the residual amplitude noise and the residual frequency noise (residual AM and FM). 8654A residual AM and FM are specified as maximum noise power (relative to carrier power) in two post-detection bandwidths: 0.3 to 3 kHz and 50 Hz to 15 kHz. The narrower bandwidth coincides with that of many voice communications systems and the wider bandwidth with the audio range.

While this frequency-domain description is probably the most useful one, there are applications in which time-domain

characterization of the noise is more appropriate. This is true for systems that perform time-domain analyses of the signals they receive. Such systems include Doppler radar systems and pulse-code-modulation and time-division-multiplexed systems. These systems tend to be very specialized, so it is all but impossible to provide any time-domain statistical noise characterization that meets the needs of all users. Fortunately, because of the predictability of the noise mechanism of most oscillators, there is a correlation between the frequency-domain noise power spectral density and most time-domain noise statistics.

SSB Signal-to-Noise Ratio

Noise at rates higher than the information bandwidth is an important problem because it delivers power to the adjacent bands of a communication system. When signal generators are used to measure a receiver's selectivity, intermodulation, and cross-modulation distortion, erroneous results will be obtained if the generator, when tuned to a nearby out-of-band channel, also delivers noise power to the channel to which the receiver is tuned.

Because only one sideband of the noise falls into the channel being measured, the type of noise (i.e., AM or FM) is irrelevant. It is impossible to determine whether noise is caused by amplitude fluctuations or phase fluctuations when only one sideband is being analyzed. Therefore, for noise rates greater than the highest information rate, it is generally not necessary or meaningful to separate the amplitude and phase noise in the characterization. The important parameter is the total noise power added to the adjacent-channel signal. 8654A SSB signal-to-noise ratio in a 1-Hz bandwidth 20 kHz away from the carrier frequency is shown in Fig. 3, page 19 as a function of carrier frequency.

Note that for single sideband receivers, where only one sideband is observed, the noise within the information bandwidth does not have to be specified by type, because the receiver cannot separate amplitude from phase noise. Thus, for SSB applications, only the total noise is important.

Broadband Noise

Noise at rates greater than 20 kHz tends to be additive, consisting of equal parts of amplitude noise and angular noise. Its primary source is thermal noise in the output amplifier. It is relatively independent of frequency. In the 8654A its level is approximately $-140 \text{ dBm}/\sqrt{\text{Hz}}$.

maintained at a constant percentage of the dc reference level at the wiper of the potentiometer, so the depth of the amplitude modulation remains fixed as the vernier is varied.

The oscillator is a standard LC-tuned oscillator, using a push-pull transistor pair as its active elements. Six adjoining frequency ranges, each covering nominally a 2:1 range, are obtained by switching the inductive element of the oscillator tank circuit. As shown in Fig. 5, the transistor pair acts as a negative conductance ($-g$) in parallel with the tank circuit. An AGC loop internal to the oscillator circuit (separate from the output ALC loop) controls this negative conductance so its magnitude is exactly equal to the positive conductance of the tank circuit. The result is

that the oscillator oscillates at a fixed level determined by the AGC reference level. In practice, the negative conductance is slightly nonlinear, a phenomenon that contributes to the stability of the loop⁵. Because the internal AGC loop of the oscillator holds the output level essentially constant, the four-stage buffer amplifier that follows the oscillator can be designed to operate at a nearly constant level.

Incidental FM

The buffer amplifier, while it provides essentially no gain, serves to isolate the oscillator from the modulator. This is an extremely important function. With any oscillator, a change in the load impedance causes a change in the oscillator frequency. Also, the

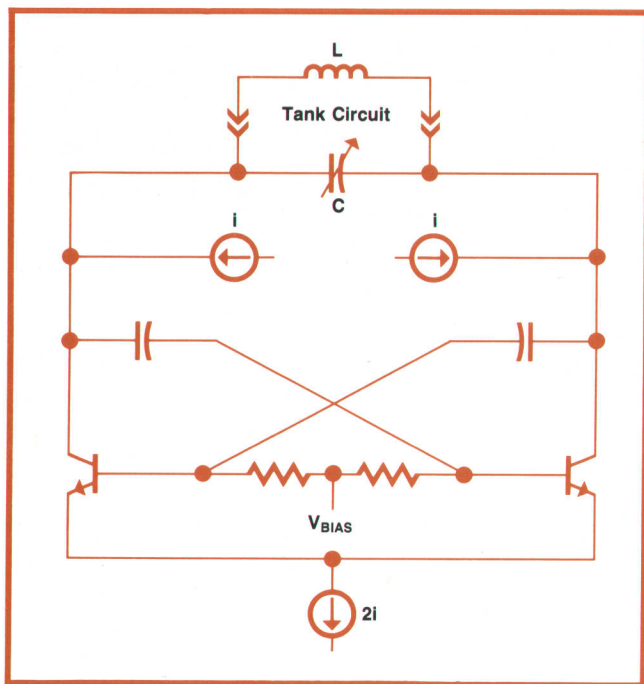


Fig. 5. 8654A oscillator active circuit has negative input conductance. The current i can be used to control the magnitude of the negative conductance. When a parallel-tuned resonant circuit is placed across the input terminals of this circuit, the system oscillates when the negative conductance of the active circuit is sufficient to overcome the positive conductance of the resonant circuit and the load.

input impedance to any modulator varies somewhat with its degree of attenuation. Combining these two facts, one can see that a modulator used to amplitude modulate an oscillator will always cause some frequency modulation. This is known as incidental frequency modulation.

Incidental frequency modulation causes problems for users who rely on a carrier frequency that remains constant while its level is varied. Examples are users who are measuring AM rejection of FM receivers, distortion and other parameters in narrow-band AM receivers, and receiver sensitivity. While a certain amount of incidental FM is inevitable, careful system design can reduce it to a point where it is insignificant and indeed nearly unmeasurable.

In a signal generator, incidental FM caused by the varying reflection coefficient of the modulator can be reduced by using a buffer amplifier that has very high reverse isolation, or equivalently, a very small S_{12} . However, careful packaging is also necessary, to assure that the RF signal from the modulator and output amplifier does not leak back to the oscillator. If it can leak back, the amplified signal from the power amplifier will cause the oscillator frequency to change, because it will cause a change in the phase-versus-frequency characteristics of the resonator. As the level of this reverse leakage signal varies, so does

its effect. The shielding required to minimize this effect is indicated in the block diagram, Fig. 4.

Attention must also be given to careful design of the power supplies to be certain that small voltage variations caused by varying modulator currents do not cause a change in the oscillator frequency.

The modulator uses PIN diodes that have a long carrier lifetime. These diodes work well over a wide frequency range (10-520 MHz), providing both good dynamic range at the high frequencies and low distortion at low frequencies. This combination is normally rather difficult to achieve economically. Although the modulator does perform well in this regard, the long carrier lifetime of the PIN diodes forces sacrifice of the wider modulation bandwidths often provided in more expensive signal generators. However, the 8654A modulator performs well over the audio range.

For the same reasons that incidental frequency modulation is a problem for many users, incidental phase modulation is also a problem. This phenomenon is caused by a modulator whose phase shift is a function of its modulation. The low junction capacitance of the PIN diodes, along with some compensation circuitry, results in a low index of incidental phase modulation.

Output and Shielding

The modulator of the 8654A is followed by a power amplifier. This amplifier, shown in Fig. 6, uses HP transistors, and its design was assisted by an in-house computer-aided-design program. It uses a quasi-push-pull design to provide high power with low distortion.

The RF signal leaving the amplifier is fed into an at-

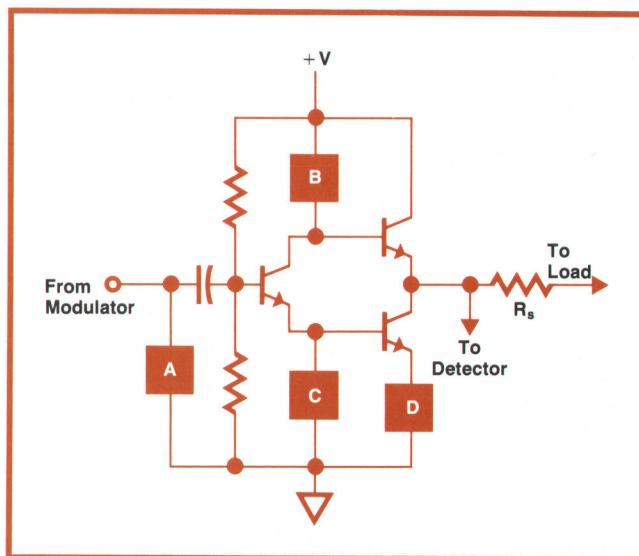


Fig. 6. Broadband 8654A output amplifier is designed to provide 10 mW into 50Ω . Because the output is leveled ahead of R_s , R_s is the effective output impedance. Impedances A, B, C, and D were chosen with the aid of an in-house computer program. HP transistors are used.

tenuator, which attenuates the signal in 10 dB steps.

It is important that the leakage from the instrument be low enough that it will not induce a significant amount of power into the user's system when he is making a measurement that requires a known amount of very low power. To maintain the required low leakage, the oscillator, buffer amplifier, modulator, and power amplifier are all contained in a well-shielded enclosure. The RF signal from this enclosure is carried to the attenuator through a semi-rigid coaxial cable, which further assures low leakage. The attenuator is enclosed within a tightly sealed cast aluminum housing.


FM Sensitivity

Although the 8654A was not designed for applications where FM deviation sensitivity as a function of carrier frequency must remain constant, it does possess a predictable sensitivity characteristic. As can be seen in Fig. 4, a pair of varactor diodes comprise part of the oscillator tank circuit. Varying the bias voltage across these diodes causes a change in their capacitance and consequently a change in the oscillator frequency.

The deviation of frequency caused by the change of the varactor capacitance, $d\omega/dC$, varies in a predictable manner over each frequency band. Using the

fundamental relationship $\omega = 1/\sqrt{LC}$, we find that $d\omega/dC = -L\omega^3/2$ for a fixed L. Thus the FM sensitivity of the 8654A, as shown in Fig. 7, varies by approximately a factor of 8 within each octave band (where the value of L is fixed). In many cases, knowledge of this sensitivity characteristic can be used to control an externally applied modulating signal to yield the required FM deviation at various frequencies.

Acknowledgments

Charles Cook and Bill Powell contributed the mechanical design. Russ Riley designed a major portion of the RF and audio oscillators. Bob Schwehr helped design the ALC loop. Jim Harmon assisted greatly in transferring the instrument to production. Wally Rasmussen provided many suggestions and ideas in the early stages of the project. 

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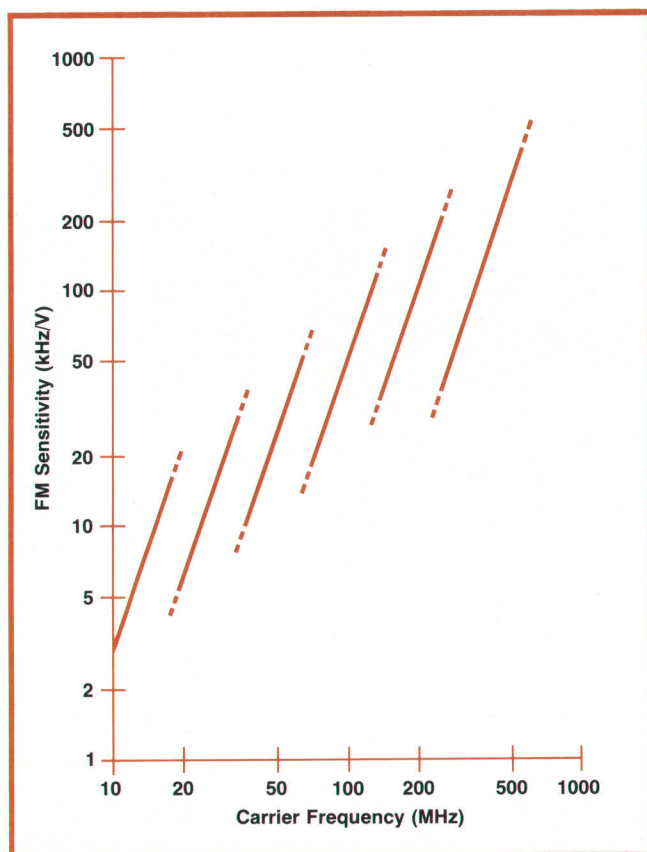
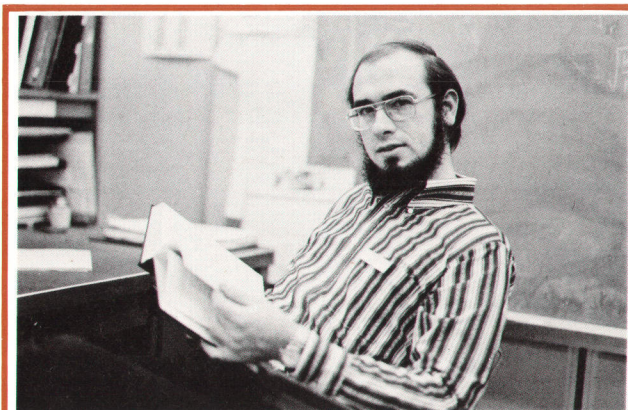


Fig. 7. FM sensitivity of Model 8654A Signal Generator.



Robert R Hay

8654 project manager Bob Hay holds BSEE (1967) and MSEE (1968) degrees from Carnegie Mellon University and an MBA degree (1972) from the University of Santa Clara. After joining HP in 1968, he helped design the 8443A Tracking Generator before taking on the 8654A. He holds two patents and is a member of IEEE. A native of Pittsburgh, Pennsylvania, Bob now lives in Los Altos Hills, California. He is an officer of a local Sierra Club section and often leads club backpacking and hiking trips. He also enjoys ski touring and bicycling. Last summer he toured Europe by bicycle, taking several weeks and visiting six countries.

SPECIFICATIONS

HP Model 8654A Signal Generator

Specifications apply from 10-520 MHz for output power $\leq +10$ dBm and over the top 10 dB of output level vernier range unless otherwise specified.

Frequency Characteristics

RANGE: 10 to 520 MHz in 6 bands

10-18.6 MHz 35-66 MHz 130-250 MHz

18.6-35 MHz 66-130 MHz 250-520 MHz

ACCURACY: $\pm 2\%$ after 1-hour warmup.

SETTABILITY: Setttable to within 5 ppm of the desired frequency with an external indicator after 1 hour warmup.

STABILITY:

TIME: < 1 kHz plus 20 ppm/5 min. (after 2 hr warmup and 15 min. after frequency change.)

LINE VOLTAGE: < 1 ppm for $+5\%$ to -10% change from nominal line voltage.

LEVEL CHANGE: < 1 ppm for a 10 dB output level change.

LOAD: < 1 ppm with any passive load change for power settings $\leq +3$ dBm.

Spectral Purity

HARMONIC DISTORTION: > 20 dB below carrier; > 25 dB below carrier

18.6-130 MHz.

SUBHARMONICS AND NON-HARMONIC SPURIOUS: (excluding line related) none measurable.

RESIDUAL AM: (averaged rms) > 70 dB below carrier in a 0.3-3 kHz post-detection noise bandwidth; > 60 dB below carrier in a 50 Hz to 15 kHz post-detection noise bandwidth.

RESIDUAL FM: (averaged rms deviation) < 0.5 ppm in a 0.3-3 kHz post-detection noise bandwidth; < 1 ppm in a 50 Hz to 15 kHz post-detection noise bandwidth.

Output Characteristics

RANGE: 10 dB steps and a 13 dB vernier provide power settings from $+10$ dBm to -130 dBm (0.7 V to 0.07 μ V) into 50 Ω .

LEVEL ACCURACY:

| | Using Top 10 dB of Vernier Range | | | | Using Full Vernier Range |
|---|----------------------------------|-----------|------------|-------------|--------------------------|
| Output Level (dBm) | +10 to -7 | -7 to -57 | -57 to -97 | -97 to -127 | +10 to -130 |
| Total Accuracy as Indicated on Level Meter (dB) | ± 1.5 | ± 2.0 | ± 2.5 | ± 3.0 | Add ± 0.5 |

NOTE: Level Accuracy error consists of allowances for: meter accuracy, detector linearity, temperature, flatness, attenuator accuracy, and twice the measurement error. All but the attenuator accuracy and the measurement error can be calibrated out with a power meter at a fixed frequency and a fixed vernier setting.

LEVEL FLATNESS: ± 1 dB referenced to the output at 50 MHz for output levels > -7 dBm.

IMPEDANCE: 50 Ω ac coupled (75 V dc maximum), VSWR $< 1.3:1$ on 0.1 range or lower.

AUXILIARY RF OUTPUT: nominally -7 dBm (100 mV).

LEAKAGE: (with all RF outputs terminated properly) Leakage limits are below those specified in MIL-I-6181D. Furthermore, with an output level < 0.01 V, less than 0.5 μ V is induced in a 2-turn, 1-inch diameter loop 1 inch away from any surface and measured into a 50 Ω receiver.

Modulation Characteristics

AMPLITUDE MODULATION: Specifications apply for output power $\leq +3$ dBm. (AM is possible above $+3$ dBm as long as the combination of the AM depth plus carrier output level does not exceed $+9$ dBm.)

DEPTH: 0 to 90%.

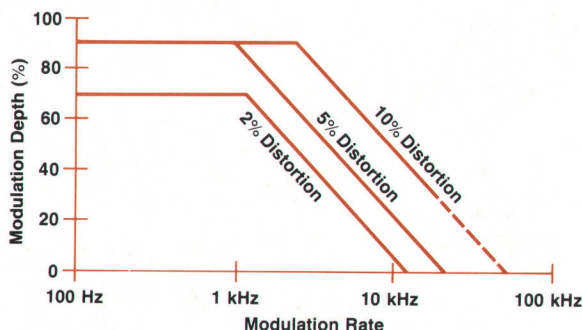
MODULATION RATE: Internal, 400 & 1000 Hz $\pm 10\%$; External 3 dB bandwidth, dc-coupled to > 20 kHz.

EXTERNAL AM SENSITIVITY: (0.1 ± 0.01) % AM/mVpk into 600 Ω at 400 and 1000 Hz rates.

INDICATED AM ACCURACY: $\pm (5\%$ of reading $+5\%$ of full scale) for modulation rates of 400 and 1000 Hz.

PEAK INCIDENTAL FREQUENCY DEVIATION: (30% AM), less than 100 Hz plus 0.1 times modulation rate.

ENVELOPE DISTORTION:



FREQUENCY MODULATION:

DEVIATION: $> 0.1\%$ of carrier frequency maximum.

MODULATION RATE: Internal, 400 & 1000 Hz $\pm 10\%$. External, 3 dB bandwidth dc-coupled to > 25 kHz driven from 600 ohms or less.

EXTERNAL FM SENSITIVITY: 10 Vpk into 600 Ω yields $> 0.1\%$ deviation (± 15 volts max).

General

POWER: 100, 120, 220 or 240 Volts $\pm 5\%$, -10% , 48 to 440 Hz; 15 VA maximum. 7 1/2 ft (2.29m) power cable furnished with mains plug to match destination requirements.

WEIGHT: Net, 16 lb 4 oz (7.4 kg). Shipping 22 lb (10 kg).

DIMENSIONS: 10 1/2" W \times 7" H \times 12" D (266 mm \times 178 mm \times 305 mm).

PRICE IN U.S.A.: \$1275.

MANUFACTURING DIVISION: STANFORD PARK DIVISION
1501 Page Mill Road
Palo Alto, California 94304 U.S.A.

Hewlett-Packard Company, 1501 Page Mill Road, Palo Alto, California 94304

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