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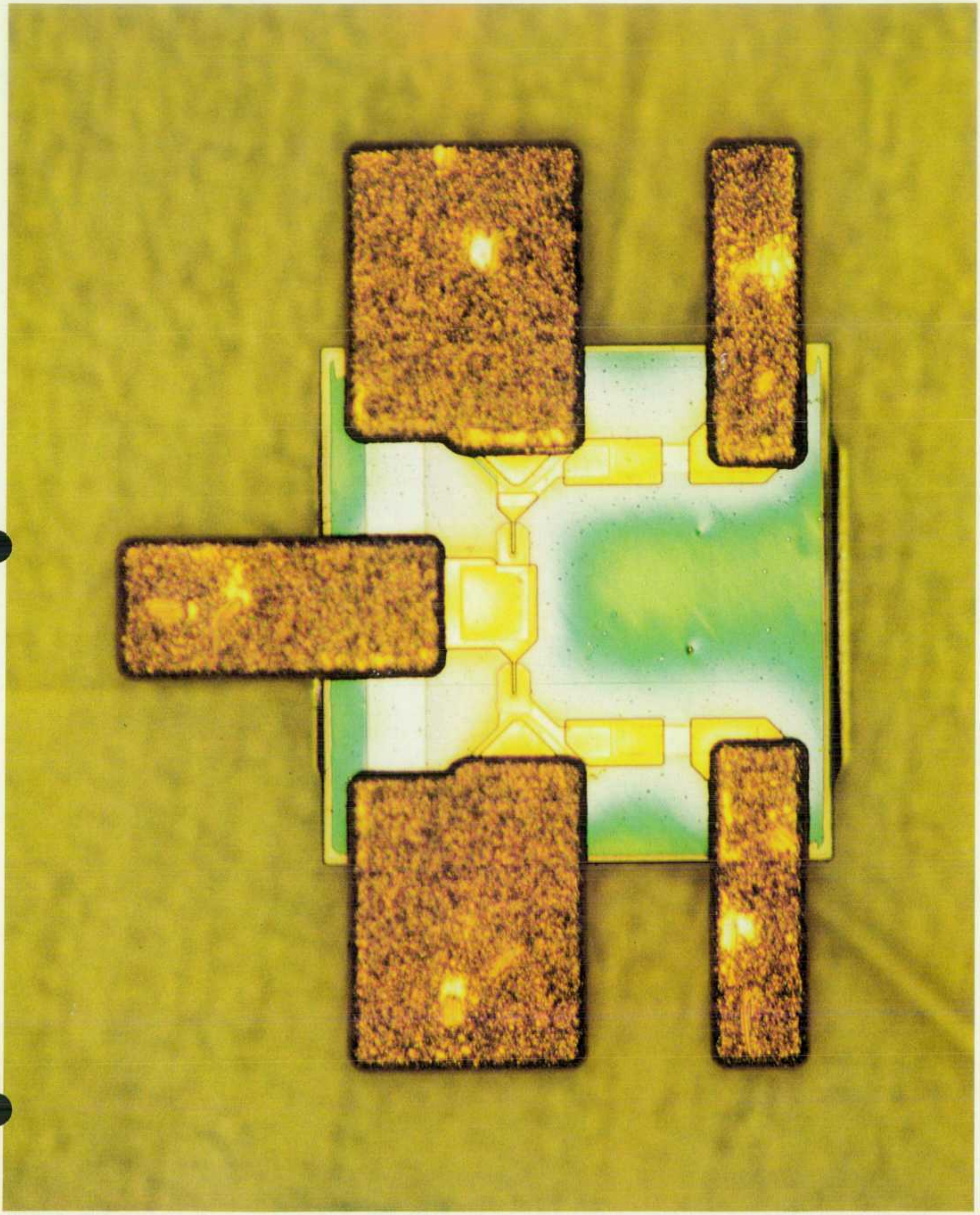

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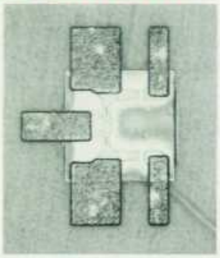

The product designs described in this issue represent the present state of the art in three of HP's oldest product lines—frequency counters, microwave instrumentation, and voltmeters. On the cover is a photomicrograph of a gallium arsenide sampler chip, the keystone of the designs of the HP 5350 Series of microwave counters. The three counters in this series are capable of counting frequencies as high as 20, 26.5, and 40 gigahertz. The GaAs sampler and all of the other microwave components are combined in a single hybrid circuit, so that all of the high-frequency circuitry is contained in just one hybrid package. To quote Scott Gibson, author of the article on page 4, "Besides the GaAs IC and hybrid circuit, engineering contributions in the new counters include halving the number of boards used in previous designs, lowering internal operating temperatures by 80%, achieving a production hybrid yield of almost 100% by improving matching between the harmonic generator and the sampler, extending calibration intervals to as long as five years, simplifying the block diagram, designing friendlier software and diagnostics, and streamlining manufacturing to double the speed of production." On page 11, Luiz Peregrino tells us how the optimum combination of local oscillator frequencies and intermediate frequency bandwidth was found for these heterodyne counters.

The assignment for the designers of the HP 3457A Digital Multimeter (page 15) was to develop a new state-of-the-art, top-of-the-line systems multimeter. That meant the highest possible performance and the maximum measurement versatility consistent with a competitive price. The HP 3457A is the first DMM from HP to offer 3½-to-6½-digit resolution, seven measurement functions, extended resolution to 7½ digits, and optional built-in scanners. Designed for both automatic systems and lab bench use, it can take more than 1300 3½-digit readings per second for high-speed measurement bursts, or less than one high-resolution reading per second for accurate lab measurements. It's a typical smart instrument, offering the user data processing as well as measurement capabilities.

The first swept network analyzers revolutionized the process of characterizing microwave components by measuring and displaying the amplitude and phase of transmitted, reflected, and absorbed power over a wide range of frequencies almost instantly. Later, scalar analyzers provided a lower-cost amplitude-only solution to the same problem. HP's latest scalar network analyzer, the HP 8757A, updates this product line to meet the ever-present need for more accuracy and versatility generated by continuing advances in RF and microwave component and system designs. A contribution of the HP 8757A that's sure to be welcomed by many engineers is its built-in limit lines. No longer do test specifications have to be drawn on the display with a crayon and observed visually. Other contributions include wider dynamic range, new detectors, more inputs and display channels, and better software and firmware. The HP 8757A's designers tell their story beginning on page 24.

-R.P. Dolan

What's Ahead

Most of the articles in next month's issue will be papers that were presented at the 1985 Hewlett-Packard Software Productivity Conference. Most of the papers deal with tools and methods for software development that are used internally at HP Divisions. One paper describes HP's artificial intelligence workstation technology. Also included is an article on a new data acquisition ROM for the HP-71B Handheld Computer.

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Gallium Arsenide Lowers Cost and Improves Performance of Microwave Counters

A proprietary GaAs sampling integrated circuit is the basis for a new family of microwave counters that operate up to 40 GHz.

by Scott R. Gibson

INTEGRATION of several microwave counter components in a single hybrid gallium arsenide (GaAs) circuit, along with other advances, has yielded a new family of microwave frequency counters that offer comprehensive feature sets, high-speed data transfer, optional low-aging-rate oscillators, extended mean time between failures (MTBF), and low cost. The family members include the HP 5350A, which measures from 10 Hz to 20 GHz, the HP 5351A, which measures from 10 Hz to 26.5 GHz, and the HP 5352A (Fig. 1), which measures from 10 Hz to 40 GHz.

The new counters have a single input to provide resolution of 1 Hz for input frequencies from 500 MHz to their upper frequency limits. A second input measures frequencies from 10 Hz to 525 MHz with resolution as fine as 0.001 Hz. A single-synthesizer design and a new measurement algorithm allow outputs of 80 measurements per second over the HP-IB (IEEE 488, IEC 625) and tolerate FM deviation on the incoming signal as high as 20 MHz in the HP 5350A and HP 5351A and 12 MHz in the HP 5352A. The new counters' sensitivity varies from −25 dBm to −15 dBm, depending on the input frequency.

Besides the GaAs IC and hybrid circuit, engineering contributions in the new counters include halving the number of boards used in previous designs, lowering internal operating temperatures by 80%, achieving a production hybrid yield of almost 100% by improving matching between the harmonic generator and the sampler, extending calibration intervals to as long as five years, simplifying the block diagram, designing friendlier software and diagnostics, and streamlining manufacturing to double the speed of production.

Microwave Counter Basics

The fundamental measurement task of a frequency counter is to indicate the frequency of a component in the spectrum of a signal. Modulation, noise, or interference frequently mask the component to be measured. Several "filtering" methods can reduce the counter's sensitivity to this distortion, but the desired component frequency must still be measured.

The speed limitations of digital counting circuitry, presently about 3 GHz, require designers to employ a frequency translation technique to count most frequencies in the microwave region (see Fig. 2). The frequency translation

(continued on page 6)

Fig. 1. *The HP 5352A Microwave Frequency Counter measures frequencies from 10 Hz to 40 GHz. Other members of the same counter family are the HP 5350A (10 Hz to 20 GHz) and the HP 5351A (10 Hz to 26.5 GHz).*

Creating Useful Diagnostics

The HP 5350A, HP 5351A, and HP 5352A Microwave Counters are designed for user friendliness. One user-friendly feature is the diagnostics. The engineering challenge was to create diagnostics that many different groups could use, including R&D, manufacturing, and service.

The diagnostics use the kernel technique, that is, each diagnostic test relies on not more than one untested assembly to perform the test. The diagnostics are organized by assembly. The counter is tested in the following order: 1) microprocessor, 2) power supply and time base, 3) multiple register counter (MRC), 4) synthesizer, and 5) input channels.

The counter has a 24-character alphanumeric liquid-crystal display (LCD). Previous instruments have been severely limited in the types of messages they could show. One example is PASS 10. The operator must look up the meaning of diagnostic 10. The HP 5350/51/52A can display a diagnostic like PASS TIMEBASE A1 D10, which means that the time base on the A1 assembly, diagnostic 10, has passed its test. Another feature of the alphanumeric display is that, unlike 7-segment LEDs, the LCD can form all letters that counters use, such as Q, Z, and M.

The diagnostics include auxiliary function tests, which do not specifically test any one assembly, but merely provide a way of examining the components of the result. Examples are:

Diagnostic 2	Display IF
Diagnostic 4	Display LO Frequency
Diagnostic 5	Display Harmonic Number and Sideband

These tests proved extremely useful during the development phase of the counter for debugging software and hardware.

The power-up and self-check diagnostics are used to test each functional piece of hardware, as well as each input, at least once. These tests also check the memory completely to find any RAM or ROM faults. The power-up and self-check tests are identical; the self-check test allows the operator to check the instrument without having to recycle power.

One of the features of the power-up sequence is that it will flag failing (or missing) boards, but will still allow the user to continue the power-up sequence and make some measurements even if some hardware is failing. Previous instruments usually refused to continue once some power-up failure was found. In the HP 5350/51/52A, for example, if a failure occurs on the low-frequency board (Input 2), there is no reason why the operator cannot proceed with measurements on Input 1. The HP 5350/51/52A allows the operator to do just that.

The numbering scheme for the diagnostics uses the first digit to identify the faulty assembly and the second digit to identify the test. The simple PASS/FAIL tests precede more complex tests within each assembly set. The inference chart shown in Fig. 1 gives the operator an outline of the hierarchy of the diagnostic set. The chart shows from left to right the order in which the counter operates and is tested. If a diagnostic has found a failure, the operator can use this chart to eliminate certain assemblies.

Acknowledgments

The HP 5350/51/52A diagnostics were the creation of a team of design and service engineers—Brian Beasley, Cathrin Callas, David Clark, Bruce Greenwood, Lisa Stambaugh, and Mark Wechsler. I would like to extend my thanks to each of them.

Sally Martin
Service Engineer
Santa Clara Division

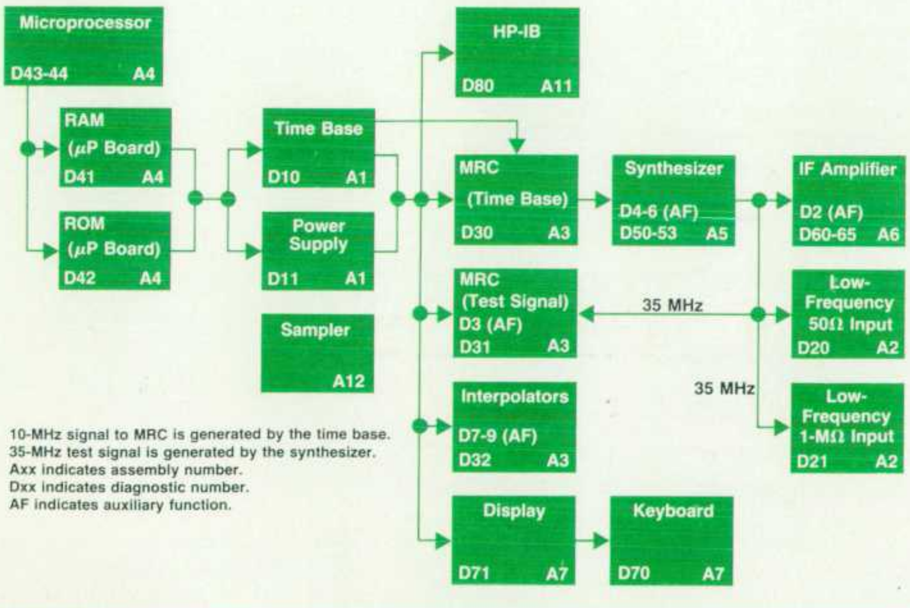

Fig. 1. *Inference chart gives the operator an outline of the hierarchy of the HP 5350/51/52A diagnostics.*

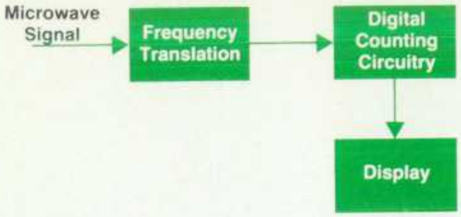

Fig. 2. *Frequency translation of microwave signals for counting.*

technique must be deterministic so that counter can use the relationship between the microwave signal and the translated signal to calculate the microwave frequency from the translated frequency. A translation technique that is particularly well-suited to microwave counters is signal sampling. Fig. 3 shows a sampling down-conversion block diagram. In this system, a local oscillator signal, synthesized from a high-stability time base to preserve frequency accuracy, drives the sampling circuitry. The sampling produces an intermediate frequency (IF) signal, which the IF section filters, amplifies, and detects before it is counted by the digital counting circuitry. The IF and the microwave signal frequency are related by:

$$f_x = Nf_{LO} \pm f_{IF} \tag{1}$$

where f_x is the microwave signal frequency, N is an integer close to f_x/f_{LO}, f_{LO} is the local oscillator frequency, and f_{IF} is the IF signal frequency.

The microprocessor steps the local oscillator through a range of frequencies until the IF section detects an IF in the selected band. The digital counting circuitry then measures f_{IF}. Thus the microprocessor knows the values of f_{LO} and f_{IF} and need only determine N and the sign to calculate f_x, the frequency of the microwave signal.

To determine N, the microprocessor shifts the LO frequency slightly (by Δf) while maintaining the signal in the IF passband. In this case N is constant and can be computed by

$$N = \left| \frac{f_{IF1} - f_{IF2}}{\Delta f} \right| \tag{2}$$

where f_{IF1} and f_{IF2} are the two IF signal frequencies. The processor determines the sign by noting whether the IF frequency shift direction is the same as that of the LO

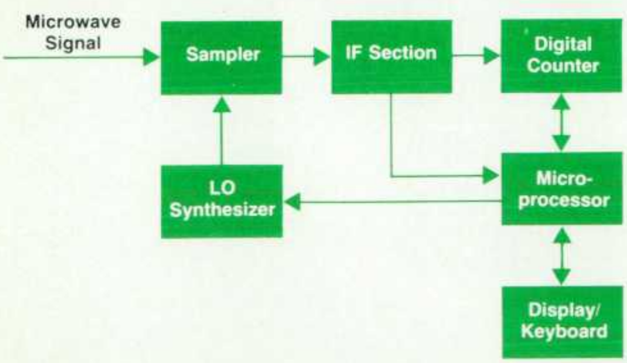

Fig. 3. *Sampling microwave counter block diagram.*

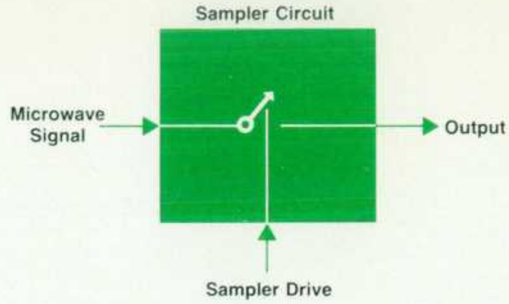

Fig. 4. *Sampling by gating.*

frequency shift, or the opposite (same = −, opposite = +).

In the new counter family, a single programmable synthesizer provides the LO signals used in down-conversion and for N and sideband determination. This system dramatically reduces the complexity and part count over previous designs.[1] The use of the 114th harmonic of this synthesizer to resolve 1 Hz in one second at 40 GHz attests to its excellent phase purity.

Understanding Sampling

In a sampling down-converter (see basic block diagram, Fig. 4), a transmission line delivers the unknown microwave signal, which is injected into a sampling or gating circuit. The designer can control the length of time the sampling gate is closed (τ) and the time between closings (T).

We can model the gating operation as:

$$y(t) = x(t)g(t) \tag{3}$$

where x(t) is the microwave signal, g(t) is the conductance of the sampler gate, and y(t) is the output of the sampling circuit. For periodic signals, Fig. 5 shows the waveforms graphically.

If we assume that g(t) alternates between zero and K, then the Fourier series representing g(t) in the frequency domain is:[2]

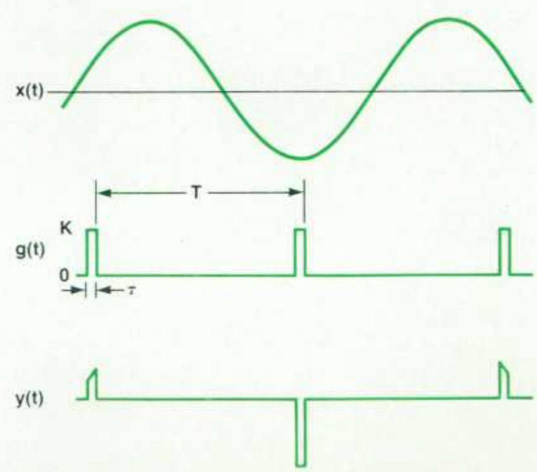

Fig. 5. *Sampling action. x(t) is the microwave signal, g(t) is the conductance of the sampler gate, and y(t) is the output of the sampling circuit.*

Manufacturing Advances

The manufacturing of the HP 5350/51/52A is faster and easier than that of any previous microwave counter. The key factors are fewer parts and greater integration of components. The hybrid, for example, is over 70% more integrated than earlier designs. This has greatly reduced the number of bonds. Locating the step-recovery diode inside the hybrid has eliminated the need to adjust this critical interface. Another bonus is that none of the printed circuit assemblies requires any special processing on the subassembly line, nor is much soldering done on the final assembly line.

Minimum hardware was a manufacturing goal. A good example involves the five TO-220 packages on the motherboard. They are held in place during machine soldering by a fixture. Then, during final assembly, these five devices are clamped by only three pieces of hardware to the card cage. The card cage provides both heat sinking and structural support, in addition to its main function as card cage for four printed circuit assemblies.

The electrolytic capacitors presented still another opportunity for pruning. The large electrolytic capacitors on the motherboard for the power supply have studs that can be soldered directly onto the board during machine soldering, rather than having to be installed on the final production line with screws. This saves over 16 pieces of hardware, not to mention time and labor.

The liquid-crystal display (LCD) mounting structure is very easy to build with the assistance of an assembly fixture. The fixture and the display are designed to minimize LCD assembly errors. Using the fixture ensures proper alignment and uniform pressure around the entire display. Once the display is fitted together, it can be tested before any screws are in place. After testing, the unit is easily fastened together with the fixture providing screw alignment. Power drivers provide a fast, accurate way of maintaining the specified torque level.

Another factor that oftens boosts the cost of manufacturing is work in process (WIP). On the final assembly line, three to five days of WIP are created by running the units in a heated chamber. To reduce WIP, manufacturing coordinated with quality assurance and decided to test the units during the heat run. During this time, the counters are continually cycled through one of the self-test diagnostics. The instruments are controlled and monitored over the HP-IB (IEEE 488). If a unit fails, the time into the test is logged. The data will be analyzed with a view to minimizing the heat run period and thereby reducing WIP.

To eliminate the possibility of labels being applied incorrectly, the label information is silkscreened on the rear of the instrument. The serial number label is the only one applied to the exterior. At the other end of the package, the keyboard has 17 keys with only two different keycap styles. These keycaps do not have any markings on them. This ensures a very simple, easy-to- assemble, low-cost keyboard.

Finally, leverage in design and volume was obtained by using the package of a previous product.

As a result of these manufacturing methods, production time of the HP 5350/51/52A is half that of previous products of similar complexity.

Acknowledgments

Much of the early manufacturing plan was created with help from Charlie Martin. Tireless efforts by Phil Mindigo, Bob Shearer, and Darryl Scroggins produced an excellent test system. Contributions from Kathy Clayton improved the assembly portion of the manufacturing plan. Finally, having a top-notch R&D team closely coordinated with manufacturing was invaluable.

Tom Beckman
Manufacturing Engineer
Santa Clara Division

$$G(f) = \sum_{N=-\infty}^{\infty} K\tau\, f_{LO}\, \mathrm{sinc}\,(Nf_{LO}\tau)\delta(f - Nf_{LO}) \qquad (4)$$

where f_{LO} is the sampling frequency (1/T). This has the familiar graph shown in Fig. 6a. Now if x(t) is a single-frequency microwave signal of amplitude A, then it can be represented in the frequency domain as shown in Fig. 6b.

Since the sampling action was modeled as multiplication in the time domain, it is modeled as convolution in the frequency domain:

$$Y(f) = G(f)*X(f) \qquad (5)$$

where * denotes the convolution function

$$L(f)*M(f) = \int_{-\infty}^{\infty} L(r)M(f-r)dr \qquad (6)$$

Carrying out this convolution graphically we get the resultant output frequency function Y(f) as shown in Fig. 7.

Since we want to translate the microwave signal to a frequency in the range of the digital counting circuitry, we select a sampler output frequency component with a low-pass IF filter. We also want to do the translation with minimum signal loss (maximum conversion efficiency), so we op-

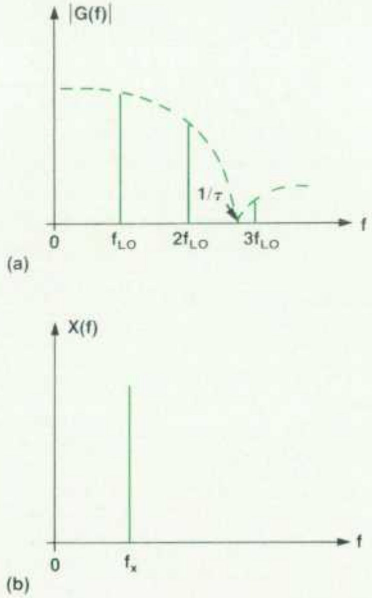

Fig. 6. *(a) Frequency-domain representation G(f) of the sampling function g(t). (b) The microwave signal in the frequency domain.*

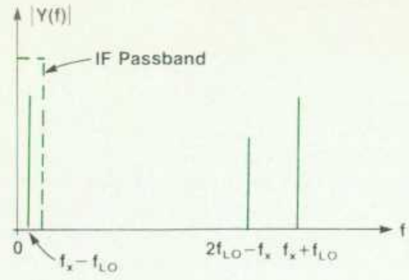

Fig. 7. *The sampled microwave signal in the frequency domain.*

timize the gating function to maximize the amplitudes of the spectral components in the frequency range we want down-converted.

If we pick the IF filter bandwidth to be half the sampling frequency $f_{LO} = 1/T$, then the filter will pass only the output component generated by convolving the microwave signal with the nearest sampler gating function component.

To optimize the conversion efficiency, we first look at the envelope of the sampler gating function spectrum:

$$|G_e(f)| = K\tau f_{LO}|\mathrm{sinc}(f\tau)| \tag{7}$$

At low frequency, the envelope amplitude is

$$|G_e(0)| = K\tau f_{LO} \tag{8}$$

and the first amplitude zero is at $f = 1/\tau$. Suppose the amplitude of the translated signal must be greater than some value A′ to maintain an adequate signal-to-noise ratio for proper frequency counting. If we know the gate's on conductance K and the gating frequency f_{LO}, we can vary the gate time τ to vary the envelope of G(f) and the conversion efficiency, as shown in Fig. 8. Using this technique, we can optimize signal sensitivity for a particular microwave counter bandwidth.

Down-Conversion with GaAs

A distinguishing feature of the HP 5350/51/52A down-conversion system is the use of a proprietary GaAs sampling IC developed at HP's Microwave Technology Division. This circuit is shown in Fig. 9.

In this circuit, the LO sampling pulse is split and opposite phases are applied to the GaAs IC. Assuming no microwave

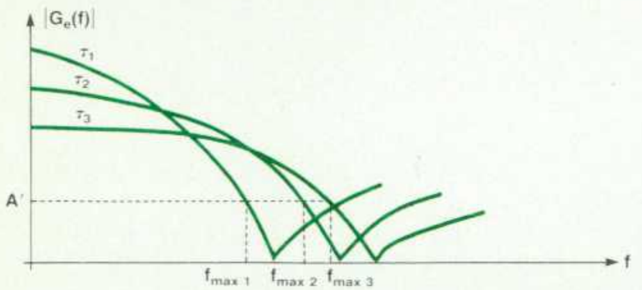

Fig. 8. *The effect of gate time τ on the sampler frequency response. $|G_e(f)|$ is the envelope of the sampler gating function spectrum.*

input signal, the rising edge of the pulse charges capacitors C1 and C2 to the peak pulse voltage through the sampling diodes D1 and D2. Then, as the LO voltage returns to ground potential, the stored capacitor potential falls across the diodes, reverse-biasing them. Resistors R3, R4, and R5 serve to discharge the capacitors slowly by bleeding current around the reverse-biased diodes. Now, when a second set of LO pulses arrives, the diodes again become forward-biased, provided that the pulse peak voltage is greater than the negative self-bias on the diodes. So, with a triangular LO pulse shape, one way to control the sampling time (the diode conduction time) is to adjust the diode reverse bias with resistors R3, R4, and R5. Resistors R1 and R2 simply terminate the microwave input transmission line in its characteristic impedance.

Suppose, then, that a positive voltage is present at the

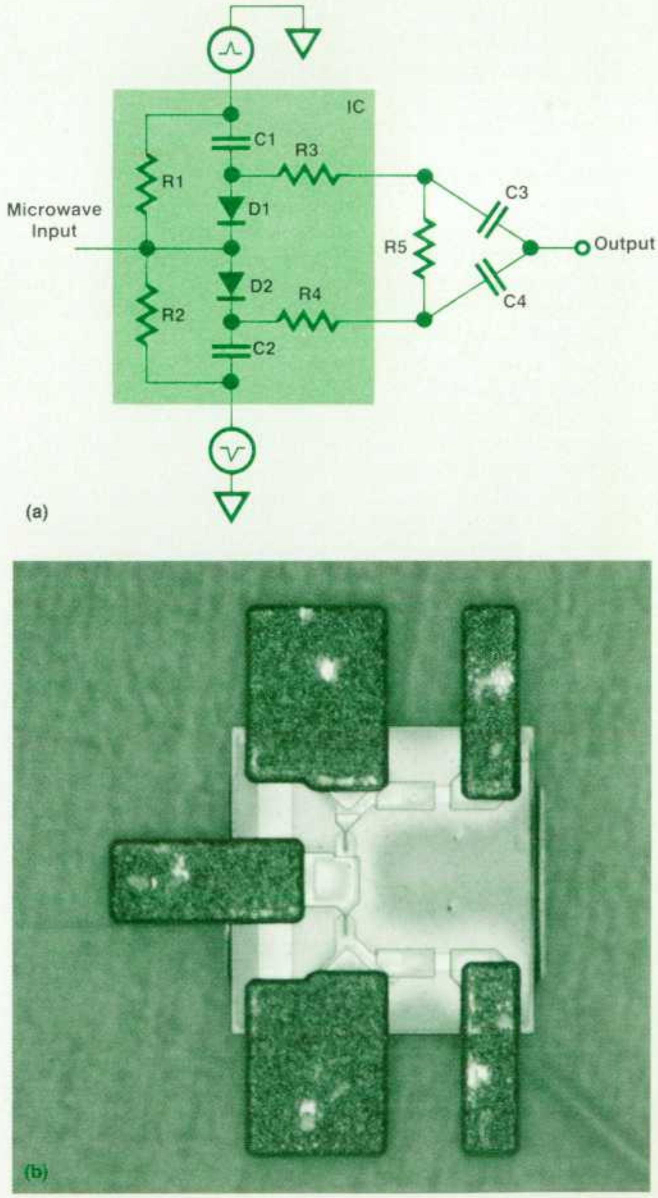

Fig. 9. *(a) Sampler circuit with pulse generators. (b) GaAs sampler.*

A New Power Transformer

The HP 5350A/51A/52A family of microwave counters incorporates a new HP-developed power transformer that employs a modular bobbin and detachable leads. While the configuration of the windings and the laminated E-I core is similar to those of traditional transformers, the lead connection scheme and bobbin have been redesigned to improve manufacturability. The salient features of the new design are variable-height dual concentric bobbins and a piggyback printed circuit board. The primary and secondary bobbins (see Fig. 1) consist of top and bottom end caps and side pieces that can be varied to provide exactly the required stack height for a given lamination size (1¼ inches in the new counters). The end caps can be common to all transformers of the same lamination size, which leverages the tooling.

The printed circuit board mounted to the top of the transformer (see Fig. 2) connects the leads to the windings. The ends of the windings are brought out during the winding process and left hanging. Then, after the core is built up, all the loose ends are wave-soldered to the board along with connectors for the primary and secondary leads. The board is common to all transformers of the same lamination size.

No soldering is needed within the windings. Not only does this improve quality and speed assembly, it also eliminates the major cause of field failure. In addition, the amount of taping necessary during winding is greatly reduced. The leads are separable from the transformer, which eliminates nicked and mislabeled leads as a cause of scrapped transformers. The dual bobbins allow the primaries and secondaries to be wound separately, reducing the number and complexity of the winding machine setups.

These innovations result in a lower scrap rate, higher quality, and a transformer cost reduction of 45%.

Acknowledgements

Many thanks are due to the HP Santa Rosa transformer design team of Jeff Argentine, Loren Eggleston, Don Maddox, and Jim McGoldrick. Their excellent design work and ability to keep pace with the counter project schedule were important contributions.

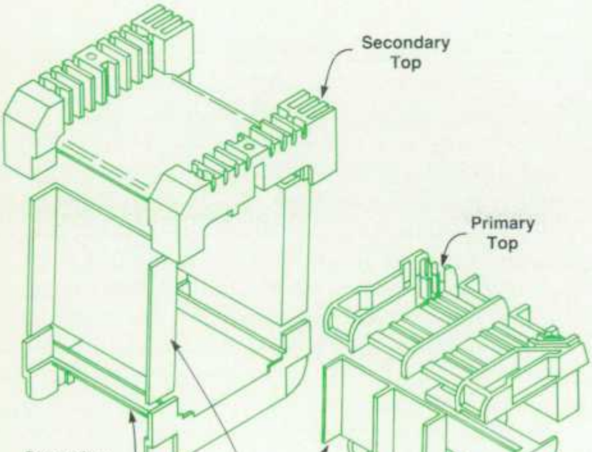

Fig. 1. *Power transformer bobbin construction.*

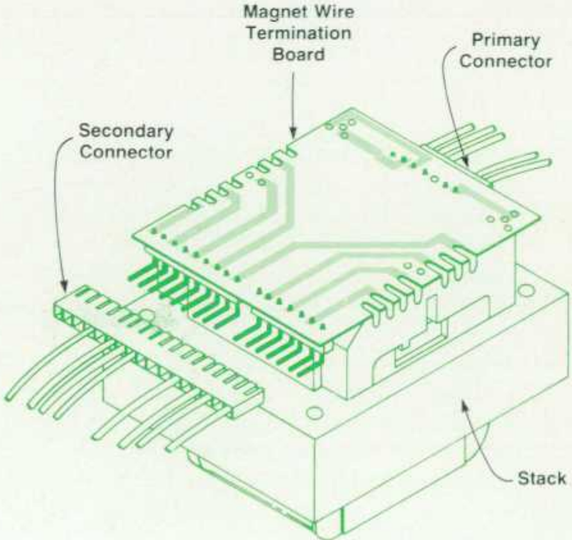

Fig. 2. *The printed circuit board mounted to the top of the transformer connects the leads to the windings.*

Bo Garrison
Development Engineer
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microwave input to the IC. When the LO pulse is present, the total current flowing through D2 is slightly more, and the total current flowing through D1 slightly less, than if no voltage were present at the microwave input. Therefore, capacitor C2 is charged slightly more and capacitor C1 is charged slightly less than if no voltage were present at the input. So a signal voltage appears across the capacitors and causes currents to flow through resistors R3 and R4 and capacitors C3 and C4 to the input of the IF preamplifier. If the input signal is negative, similar operation causes the opposite-polarity signal to appear at the output.

If the microwave input signal frequency is an even harmonic of the sampling rate, the same voltage appears at the input each sampling period, and the output of the circuit is a constant voltage. If the microwave frequency is different from an even harmonic of the sampling frequency by an offset f_{IF}, then phase accumulation causes the voltage sampled to vary with time, producing the IF signal equal to f_{IF}.

Harmonic Generator

Producing a narrow pulse for driving the sampling circuit

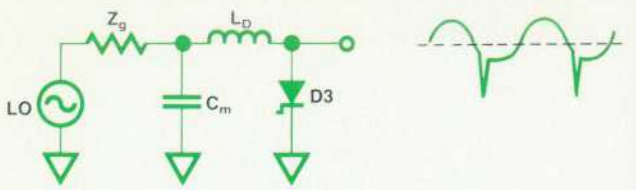

Fig. 10. *Pulse generating circuit based on a step-recovery diode.*

is central to the proper operation of the down-conversion system. The width of the pulse (sampling time) is directly related to the down-conversion frequency response: the narrower the pulse, the flatter the frequency response, but the greater the conversion loss (see equation 7). A step-recovery diode pulse generator fulfills the requirements well. The circuit of Fig. 10, driven by the local oscillator signal, produces large negative-going pulses of very short duration at the LO frequency. The simplicity of the circuit allows the generator to be integrated inside the sampling hybrid along with the GaAs IC. This eliminates the difficulty and cost of passing a second microwave signal into the hybrid.

Physical Configuration

The overall goal of the down-conversion system is maximum reliability and performance over the bandwidth of interest, along with minimum cost. A single hybrid package contains all the microwave circuitry, so there is only a single microwave connector for the microwave input signal. The package is designed to have as few tight machine tolerances as possible for ease of manufacture. One can quickly remove the package cover and all components are easily accessible for repair.

Both Si and GaAs designs were evaluated during the development of the down-conversion system. GaAs gave better performance for a number of reasons:

- Lower parasitic capacitance in the sampling diodes reduces the allowable sampling time and increases high-frequency performance.
- Beam-lead device integration reduces the number and complexity of electrical bonds for lower assembly cost, higher reliability, and easier repair.
- Integration of resistors and capacitors along with the semiconductor devices reduces substrate complexity and cost.
- The high-barrier GaAs diodes increase the dynamic range of the sampler over Si medium-barrier diodes.

Inside the package (see Fig. 11), a thin-film substrate provides support for the beam-lead diodes of the GaAs sampling circuit and three types of transmission lines. The microwave signal travels through the microwave connector into the hybrid package. Once inside, a transition to coplanar waveguide guides the signal to the input of the IC. The LO signal comes through the package wall on a simple feedthrough pin. From there a short bond wire carries it past matching capacitor C_m to drive inductor L_D, which is a long bonding ribbon. On the end of the inductor wire is a microstrip transmission line which is also connected to the step-recovery diode D3 with a small bonding wire. The narrow pulse generated by this circuit travels along the microstrip line past a tightly coupled slot transmission line to the terminating resistor. The slot line conveniently carries the pulse voltage wave to the IC for sampling. The two unfiltered outputs of the sampling circuit leave the hybrid package through two feedthrough pins connected by a pair of bond ribbons.

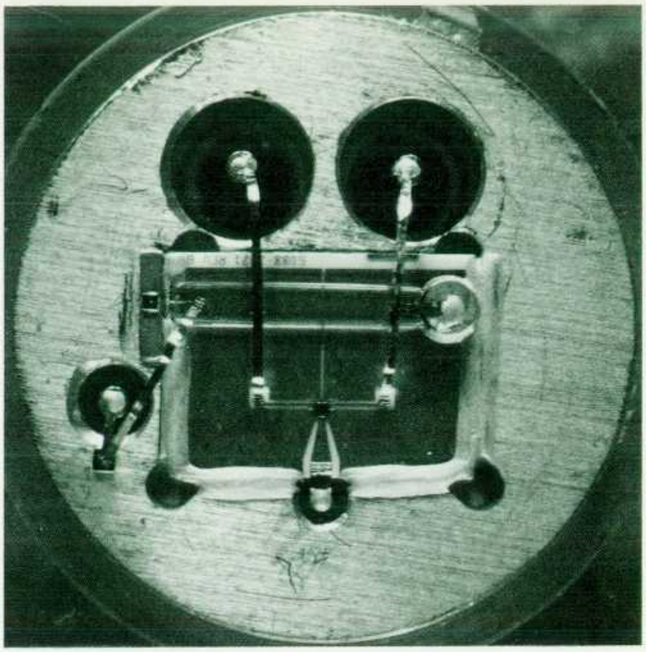

Fig. 11. *Packaged sampler with cover removed.*

Outside the hybrid, the signals add in an IF preamplifier, which increases the signal level to reduce its sensitivity to noise. The IF assembly establishes the noise bandwidth with a very sharp-cutoff elliptical filter and amplifies and limits the IF signal to bring it to the required level for the digital counting hardware. The IF assembly also detects the presence of the IF signal in the required frequency band to notify the microprocessor that an IF exists to count.

Acknowledgments

Considerable commitment from people too numerous to list was necessary to make this product concept a reality. Suffice it to say that without the strong team effort of each of these contributors, led by project manager Bob Rehner, these products would not have met their goals. I would like especially to acknowledge the invaluable support and guidance of Al Barber and the work of Luiz Peregrino on the system design, Jim McCarthy on the sampler design, and Al Tarbutton on the LO driver/IF preamplifier design. George Lee helped greatly with suggestions on this manuscript.

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Optimum Solution for IF Bandwidth and LO Frequencies in a Microwave Counter

by Luiz Peregrino

A MICROWAVE COUNTER consists of a microwave receiver and a low-frequency counter to measure the receiver IF. The microwave receiver must have enough range to cover the input signal frequency range, and an IF amplifier wide enough to preserve any modulation present in the input signal. Furthermore, the harmonic number and sideband must be determined to compute the frequency of the input signal. This paper presents an optimum solution to the problem of determining the IF bandwidth and the local oscillator frequency range for the microwave receiver. The optimum solution was obtained by reducing the problem to linear programming.

Heterodyne Microwave Receiver Model

In a heterodyne microwave receiver counter, we determine the harmonic number and sidebands by applying a small frequency deviation, Δf, to the local oscillator frequency.

Assuming that for the two values of the local oscillator frequencies, f_{LO} and $f_{LO}+\Delta f$, the intermediate frequencies remain within the IF amplifier bandwidth and result from the same harmonic number and sideband, we can derive two sets of two equations, one for each sideband (plus or minus sign):

$$f_x = Nf_{LO} \pm f_{IF}$$
$$f_x = N(f_{LO}+\Delta f) \pm (f_{IF}+\Delta f_{IF})$$

Solving, we get the harmonic number N, the sideband (+ or −), and the unknown frequency f_x. Because the harmonic number N is a positive integer, we can eliminate part of the measurement errors and get:

$$H = \mp N = \mathrm{ROUND}(\Delta f_{IF}/\Delta f)$$

where ROUND means round to the nearest integer.

Using H as defined above, we get:

$$f_x = |Hf_{LO} - f_{IF}|$$

The problem is to determine the proper local oscillator frequencies, the local oscillator frequency variation Δf, and the IF bandwidth that ensure that the previous assumptions are satisfied.

A simplified block diagram of the receiver is shown in Fig. 1. In the simplified block diagram, it is assumed that the low-pass filter following the harmonic mixer is fixed, with its cutoff frequency set at half the maximum local oscillator frequency. This ensures that, for any value of the input frequency f_x, there will be a signal at the input of the limiter. This guarantees that the counter will measure the strongest signal when several are present.[1]

The IF amplifier frequency response is shown in Fig. 2. The bottom and top parts of the IF amplifier frequency range are allocated to preserve any modulation present in the input signal. We will denote the width of the part allocated to the modulation as f_G, for frequency guard band. We will consider a signal to be in the IF amplifier band if the carrier frequency (center frequency) is between the IF frequencies indicated as $\min(f_{IF})$ and $\max(f_{IF})$ in Fig. 2.

System Equations

The next step is to derive the relations that the local oscillator frequency f_{LO}, the input frequency f_x, the IF frequency f_{IF}, and Δf must obey. For this purpose we will proceed as though there were two receivers—one with local oscillator frequency f_{LO} and the other with local oscillator frequency $f_{LO}+\Delta f$. Furthermore, a synthesizer generates both local oscillator frequencies, so they can be set only to integer multiples of a quantization frequency f_Q, resulting in:

$$f_{LO} = Mf_Q$$
$$\Delta f = Kf_Q$$

where M and K are positive integers.

To simplify our notation, we will use a ˆ over a variable name to denote the maximum of the variable and a ˇ over a variable name to indicate the minimum of the variable. For example, $\max(f) = \hat{f}$ and $\min(f) = \check{f}$.

In reality, there is only one microwave receiver, and the local oscillator frequency is switched between f_{LO} and $f_{LO}+\Delta f$. This could cause errors in the determination of the harmonic number and sideband because the IF frequencies are not determined at the same time. For a solution to this problem, see references 2, 3, and 4.

Assuming for the time being that the local oscillator frequency is continuous, we can determine f_x as a function of the local oscillator frequency using harmonic numbers and sidebands as parameters as shown in Fig. 3 for two

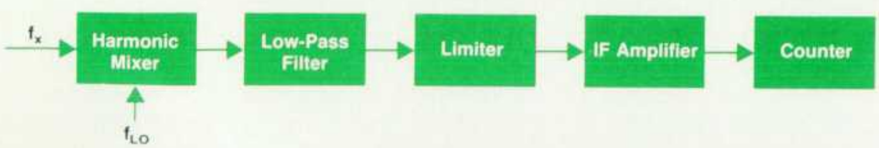

Fig. 1. *Heterodyne microwave receiver counter. f_x is the unknown input frequency to be measured and f_{LO} is the local oscillator frequency.*

harmonic numbers, N and N + 1. The two sets of parallelograms in Fig. 3 indicate the ranges covered by the receivers for different sidebands and harmonic numbers. Because the signal must be received by both receivers, the actual range covered is the intersection of the two ranges, shown shaded in Fig. 3. From this graph we can determine most of the conditions that must be satisfied by computing the edges of the parallelograms:

1. Minimum input frequency $\check{f}_x$:

$$\check{f}_x = \check{N}(\check{f}_{LO} + \Delta f) - \hat{f}_{IF} \quad (1)$$

2. No gap in frequency coverage while changing from lower to upper sidebands:

$$\check{N}\hat{f}_{LO} - \check{f}_{IF} \geq \check{N}(\check{f}_{LO} + \Delta f) + \check{f}_{IF} \quad (2)$$

This inequality must be true for all harmonic numbers N, but it is easy to verify that if it is true for the minimum harmonic number $\check{N}$, it will be true for all N.

3. No gap in frequency coverage while changing from harmonic number N to N + 1:

$$\check{N}\hat{f}_{LO} + \hat{f}_{IF} \geq (\check{N} + 1)(\check{f}_{LO} + \Delta f) - \hat{f}_{IF} \quad (3)$$

Again, it is easy to verify that if the above inequality is true for the minimum harmonic number $\check{N}$, it will be true for all harmonic numbers.

4. No overlap in coverage between the lower and upper sidebands for the two different local oscillator frequencies, f_{LO} and $f_{LO} + \Delta f$. This will guarantee that in both receivers the same sideband generates the IF. This is ensured by:

$$N\check{f}_{LO} + \check{f}_{IF} > N(\hat{f}_{LO} + \Delta f) - \check{f}_{IF}$$

We can reduce this inequality further, resulting in an upper limit for the harmonic number:

$$\hat{N} < 2\check{f}_{IF}/\Delta f \quad (4)$$

5. No overlap in coverage for different harmonic numbers with the two different local oscillator frequencies, f_{LO} and $f_{LO} + \Delta f$. This will guarantee that for the two different local oscillator frequencies, the same harmonic number generates the IF. This gives the inequality:

$$(N + 1)\check{f}_{LO} - \hat{f}_{IF} > N(\check{f}_{LO} + \Delta f) + \hat{f}_{IF}$$

This inequality imposes one more limit on the maximum harmonic number:

$$\hat{N} < \frac{\check{f}_{LO} - 2\hat{f}_{IF}}{\Delta f} \quad (5)$$

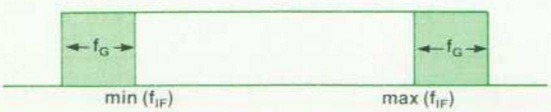

Fig. 2. *IF amplifier frequency response. The guard band f_G is allocated to accommodate modulation on the input signal.*

Now, let us examine the conditions that must be satisfied so that the intermediate frequencies for both values of local oscillator frequencies remain in the specified IF bandwidth, that is, in the range $\check{f}_{IF}$ to $\hat{f}_{IF}$. For this purpose, let us assume that the two intermediate frequencies generated by the local oscillator frequencies f_{LO} and $f_{LO} + \Delta f$ are as shown in Fig. 4.

The first obvious condition is that the IF frequency response must have a lower cutoff frequency greater than zero, giving the relation:

$$\check{f}_{IF} - f_G > 0 \quad (6)$$

Now, let us assume that the input signal frequency varies, and that to maintain it in the desired bandwidth we change the local oscillator frequency by the minimum allowed value f_Q. This causes a change in both IFs in one direction or the other of Nf_Q. The worst case, when one of the IF signals is on the edge of the specified limits, gives the relation:

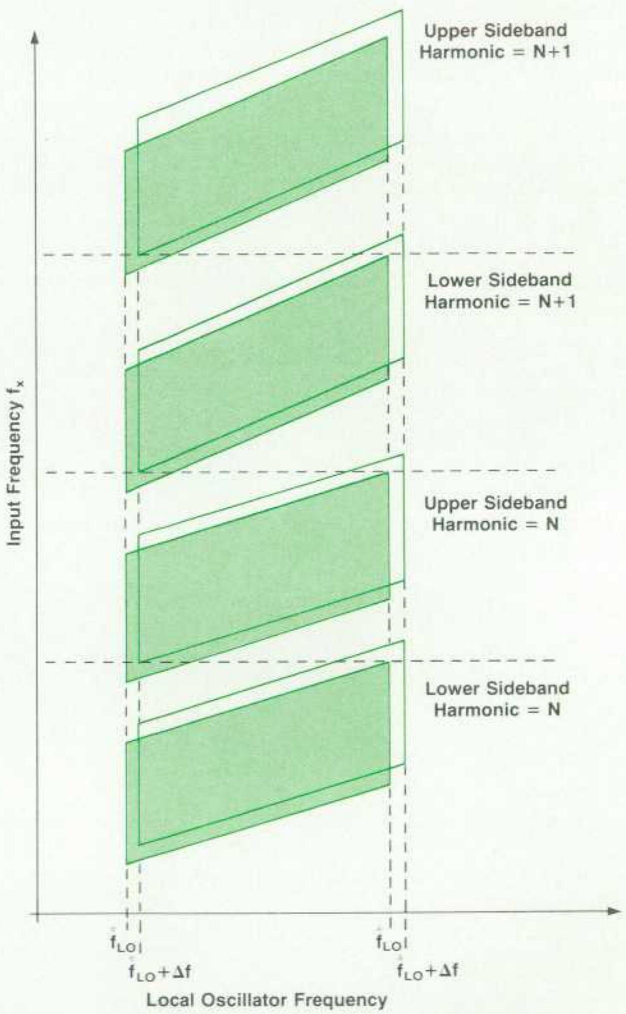

Fig. 3. *Ranges of input frequencies covered by local oscillator frequencies f_{LO} and $f_{LO} + \Delta f$ for two harmonic numbers, N and N + 1. The shaded area shows the ranges covered by both LO frequencies.*

$$N\Delta f + Nf_Q < \hat{f}_{IF} - \check{f}_{IF}$$

This imposes one more limit on the maximum harmonic number:

$$\hat{N} < \frac{\hat{f}_{IF} - \check{f}_{IF}}{\Delta f + f_Q} \qquad (7)$$

Next, let us derive a relation that guarantees no intermodulation in the limiter caused by IFs generated by the different harmonic numbers. This can be achieved by forcing the undesired IF to be higher than the cutoff frequency of the low-pass filter following the sampler. In this case we must include any frequency spread because of frequency modulation present in the input signal, that is, we must consider the full span of the IF amplifier including the guard band. The worst case is for the minimum local oscillator frequency and maximum IF and is given by:

$$\check{f}_{LO} - (\hat{f}_{IF} + f_G) > \hat{f}_{LO}/2 \qquad (8)$$

To ensure that we always receive the maximum-amplitude signal, we must have a signal at the input of the limiter for at least one of the local oscillator frequencies and for all possible values of the input signal frequency. The worst case occurs when the input frequency f_x is close to one of the harmonics of the local oscillator, specifically halfway between Nf_{LO} and $N(f_{LO}+\Delta f)$, resulting in the relation:

$$f_c < \check{N}\Delta f/2 \qquad (9)$$

where f_c is the low-frequency cutoff of the limiter amplifier (that is, the limiter amplifier doesn't need to go all the way down to dc).

Linear Programming and Analytical Solutions

All the inequalities derived so far can be considered as the constraints in a linear programing problem.[5] To solve the equations subject to the constraints, a computer program called LINPRO was used. This program is available in the business section of the HP Timesharing Library.

To gain insight, a simplified analytical solution to the problem was also derived. With proper choice of parameters, the problem reduces to two dimensions, which can be very easy visualized. It will be assumed that the minimum unknown frequency, the guard band f_G, and the offset frequency Δf are given, and we are solving for the other parameters.

To obtain the analytical solution, inequalities 1, 2, 3, and 8 derived above were used and the IF bandwidth was maximized. Then, the maximum harmonic number $\hat{N}$ was determined as function of $\Delta f + f_Q$ and all the other relations were checked.

Determining $\hat{f}_{IF}$ in the equation for $\check{f}_x$ (equation 1), and inserting the result in the next three inequalities, we get:

$$\hat{f}_{LO} \geq \check{f}_{LO} + \Delta f + 2\check{f}_{IF}/\check{N}$$

$$\hat{f}_{LO} \geq \frac{1-\check{N}}{\check{N}}(\check{f}_{LO} + \Delta f) + \frac{2\check{f}_x}{\check{N}}$$

$$\hat{f}_{LO} \leq 2(1-\check{N})\check{f}_{LO} + 2(\check{f}_x - f_G - \check{N}\Delta f)$$

In the coordinate system determined by the variables $\check{f}_{LO}$ and $\hat{f}_{LO}$, each of these inequalities divides the space into two regions—one in which they are satisfied and one in which they are not. The intersection of all the regions where the inequalities are satisfied is the region of all possible solutions and is denoted as the feasible region, that is any point in this region could be used as a solution giving a specific value for $\check{f}_{LO}$ and $\hat{f}_{LO}$. In Fig. 5 we show all these regions, with the feasible region shown shaded.

At this point, it should be emphasized that there are no feasible solutions for $\check{N}<2$.

To obtain the maximum IF bandwidth solution, we plot the line equation for constant IF bandwidth and select any point on this line inside the feasible region. Then, we move the constant bandwidth line, in the direction of increasing bandwidth, up to the point where there would no longer be a feasible solution for any point on this line. The last point of contact of the constant bandwidth line and the feasible region determines a feasible solution with maximum bandwidth.

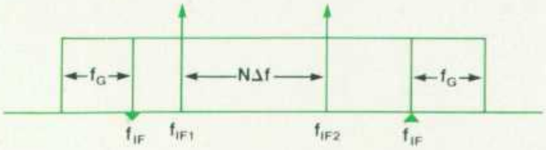

Fig. 4. *IF response with two signals representing the mixed-down signals for the two LO frequencies f_{LO} and $f_{LO}+\Delta f$.*

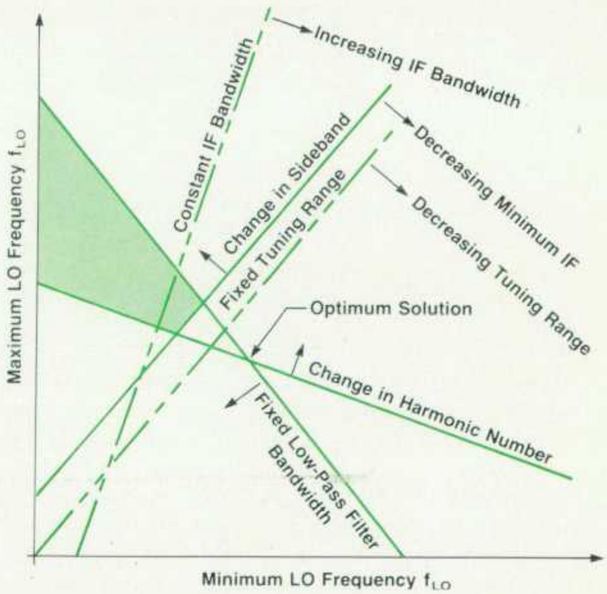

Fig. 5. *The shaded region is the region where all of the linear programming inequalities are satisfied. Any point in this region can be used as a solution, giving values for $\check{f}_{LO}$ and $\hat{f}_{LO}$. The optimum solution is found by manipulating the variables shown.*

A line through the point ($\check{f}_{LO}=0$, $\hat{f}_{LO}=0$) represents a constant tuning range for the local oscillator. To obtain a minimum tuning range, we rotate this line in the direction of reducing slope. The last point of contact between this line and the feasible region determines a feasible solution with minimum tuning range.

Decreasing the minimum IF frequency $\check{f}_{IF}$, we can further reduce the tuning range of the local oscillator and increase the IF bandwidth. The optimum solution obtained gives the minimum local oscillator tuning range and the maximum IF bandwidth, which is equivalent to maximizing $\hat{N}$ for the given Δf and f_Q.

Solving for the optimum solution we get:

$$\check{f}_{IF} = \frac{\check{N}}{\check{N}-1}\,(f_G + \Delta f)$$

$$\check{f}_{LO} = \frac{2}{2\check{N}-1}\,(\check{f}_x - \check{f}_{IF}) - \Delta f$$

$$\hat{f}_{LO} = \check{f}_{LO} + \frac{2}{\check{N}}\,\check{f}_{IF} + \Delta f$$

$$\hat{f}_{IF} = \check{N}(\check{f}_{LO} + \Delta f) - \check{f}_x$$

This system of equations gives an optimum solution as a function of the minimum frequency of the input signal f_x, the desired amount of guard band f_G, and the local oscillator offset Δf. The local oscillator frequencies can be set only at multiples of the quantization frequency f_Q, and the only valid answers are the quantized feasible solutions. This is a harder problem, but we can perturb the solution obtained above to get a quantized solution. This was done by solving the set of equations to obtain:

$$\check{f}_{IF} = \check{f}_x - \frac{2\check{N}-1}{2}\,(\check{f}_{LO} + \Delta f)$$

$$\hat{f}_{LO} = \check{f}_x - \frac{2\check{N}-3}{2}\,(\check{f}_{LO} + \Delta f)$$

$$\hat{f}_{IF} = \check{N}(\check{f}_{LO} + \Delta f) - \check{f}_x$$

$$f_G = \frac{\check{N}-1}{\check{N}}\,\check{f}_{IF} - \Delta f$$

To find a quantized solution for this system of equations, all that we have to do is perturb $\check{f}_{LO}$ and $\check{f}_x$ in the desired direction to obtain multiples of f_Q, since Δf is a multiple of f_Q.

After determining the local oscillator frequencies, one final refinement was made. Since, depending on harmonic number, the LO frequency can change in steps larger than f_Q, we can eliminate the unnecessary frequencies. This was done by deriving formulas for coverage for each specific LO frequency and maintaining only the essential ones.

Acknowledgments

I want to acknowledge the many very informative discussions with Dick Schneider, Art Bloedorn, and Darwin Throne which led me to this approach and also to George Lee for the manuscript criticisms.

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Seven-Function Systems Multimeter Offers Extended Resolution and Scanner Capabilities

This new 3½-to-6½-digit DMM measures frequency and period as well as dc and ac voltage, dc and ac current, and resistance. Extended resolution provides an extra digit.

by Scott D. Stever, Joseph E. Mueller, Thomas G. Rodine, Douglas W. Olsen, and Ronald K. Tuttle

A NEW DIGITAL MULTIMETER (DMM), the HP 3457A Multimeter (Fig. 1), is designed to provide both bench and systems users with the highest possible performance and the maximum measurement versatility consistent with a very competitive price. System features such as an easy-to-use multimeter language, programmable front and rear input terminals, and front-panel access to virtually all of the HP-IB (IEEE 488/IEC 625) commands for convenient test program debugging, help improve throughput in automatic test system applications. Reading rate, an important parameter for a systems multimeter, can be traded off with measurement resolution.

In the HP 3457A, more than 1300 readings per second can be converted to ±3000 counts, or more than 1200 readings per second can be converted to ±30,000 counts with a 100-μs sample aperture. For measurements where rejection of 50 or 60-Hz noise is critical, 53 readings per second can be converted with 0.33-ppm resolution. More noise attenuation can be achieved by further slowing the conversion rate, still maintaining greater than 160 dB of effective common mode noise rejection.

For applications requiring precise measurement pacing, the HP 3457A lets the user set the interval between multiple readings taken from a single trigger. This internal timer capability is particularly useful for waveform sampling and digitization. The timer is settable to 0.02% accuracy in 1-μs increments. This function is performed by the system microprocessor, and jitter on the timer interval is limited to a negligible 2 ns.

The ability to reprogram the multimeter rapidly is an important consideration for a systems instrument. In situations where the multimeter is used with a scanner, it is common to require a different function and range for each channel. The HP 3457A has a special programming mode for these applications, enabling it to change function and range and take a reading up to 30 times per second.

The HP 3457A offers seven functions: dc and ac voltage, dc and ac current, resistance, frequency, and period. The 24-hr dc accuracy on the 3V range is ±5.5 ppm. To broaden the range of applications, several functions have a greater measurement capability than is normally found in similar instruments. For example, resistance measurements can be made to 3 GΩ and useful dc current measurements can be made down to 100 pA. The ac voltage function has full-scale ranges as low as 30 mV and is specified to 1 MHz.

The HP 3457A's 2000-byte internal memory can be partitioned between reading, subprogram, and state storage. In addition, up to 11 complete setups can be stored in a nonvolatile state memory for easy reconfiguration.

Statistical functions such as mean and standard deviation are part of the built-in math routines. Other routines include thermistor linearization, pass-fail limit testing, dB, dBm, scale, offset, and single-pole digital filters. The rms math function aids low-frequency ac voltage measurements.

The HP 3457A is programmable via the HP-IB. A **VOLTMETER COMPLETE** output and an **EXTERNAL TRIGGER** input are useful for synchronizing other test instrumentation with the DMM. The HP 3457A also features a set of new easy-to-use program commands that will allow today's software to be used with tomorrow's DMMs. The commands form a core language called HP-ML (HP Multimeter Language).

Autocal

Two HP 3457A functions, ac volts and the 3-GΩ resistance ranges, use an autocalibration feature to improve their accuracy under changing environmental conditions. This feature, which requires no external standards, can be ini-

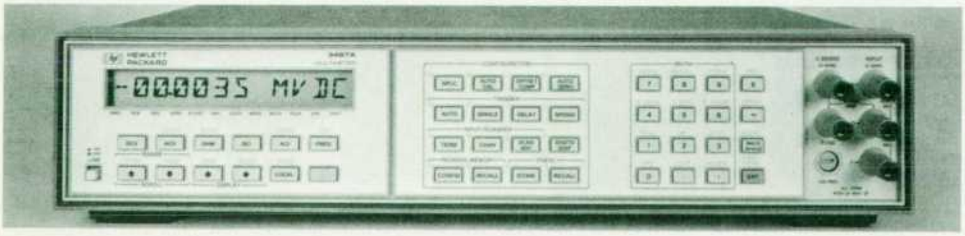

Fig. 1. *The HP 3457A Multimeter is designed for both bench and systems use. Reading rate can be traded off with measurement resolution. Seven functions provide versatility.*

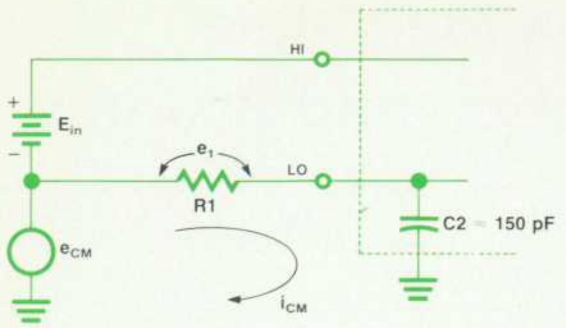

Fig. 2. *Common mode current flow in a floating voltmeter such as the HP 3457A.*

tiated from the front panel and is also programmable. Autocal corrects both gain and offset errors for the extended ohms function. In the case of the ac functions, Autocal recompensates the input attenuator for flat frequency response and nulls out dc offsets in the input amplifier for better dc accuracy. However, it does not correct for mid-band gain changes. This must be done using the electronic calibration process along with an external standard.

Noise Rejection

An important consideration in making precision measurements is the effect of power-line-related noise on the measurement. An unwanted noise voltage, shown as e_1 in Fig. 2, can result from a 50-Hz or 60-Hz voltage that is common to both input leads. This occurs because of a voltage divider action between the impedance from instrument **LO** to ground (C2) and the resistance in the **LO** input lead, R1. Guarding is one way to reduce the effect of the **LO**-to-ground impedance.[1] In a guarded voltmeter, an additional sheet-metal structure is placed between the analog circuitry and chassis ground. It is almost unavoidable that the capacitance between this shield and the chassis is very large, on the order of 1000 to 2000 pF. When the guard terminal is used properly (Fig. 3a), this large capacitance is not a serious problem since the noise current does not flow through the measurement loop. However, the most common arrangement is to leave the guard connected directly to the **LO** terminal. For example, this may be necessary when scanning several signals that do not have a common guard point. This situation is commonly encountered in data acquisition applications. Unfortunately, connecting the guard to **LO** causes the large guard-to-ground capacitance to appear between ground and **LO**. Fig. 3b depicts this situation; the **LO**-to-ground capacitance is approximately ten times larger than it would have been without the guard. In this case, the noise voltage e_1 is also ten times larger than it would have been without the guard.

In the HP 3457A, considerable thought was given to minimizing the stray capacitance from the isolated analog measurement section to chassis ground. The result is a typical capacitance of 150 pF. Referring to Fig. 2 and using C2 = 150 pF and R1 = 1 kΩ, the effects of e_{CM} are reduced by about 86 dB at 60 Hz. Thus, low capacitance to ground allows the HP 3457A to have very good rejection of ac common mode signals even though it is not fully guarded. The inherent normal mode rejection of an integrating voltmeter further improves this performance. For example, with a 10-PLC (power line cycle) integration time, the effective common mode rejection ratio (the sum of ac common and normal mode rejection) is 156 dB. This means that a 50Vac, 50 or 60-Hz common mode voltage will only result in 1 μV of noise when integrating over ten PLCs.

High-Resolution Mode

While the front-panel can display 6½ digits (± 3,000,000 counts), a seventh digit is available. This extra digit is continuously written into a math register when using the longer integration times (10 or 100 PLC). To access the seventh digit over the HP-IB, the user can recall the high-resolution math register (HIRES). Since this register is already scaled, it can be added directly to the current reading to get a 7½-digit measurement. Fig. 4 shows typical 3V-range noise for 7½-digit readings.

Scanner Options

To add versatility, the HP 3457A has provisions for field-installable input multiplex options. Two different cards are available, an armature relay multiplex card and a reed relay multiplex card. The HP 44491A armature relay multiplex card (Fig. 5) offers eight two-wire channels that can be configured from the front panel or under program control as four four-wire channels (for resistance measurements) or any combination of four-wire and two-wire inputs. The input channels have a maximum switching and measurement speed of 33 channels per second. Two other channels

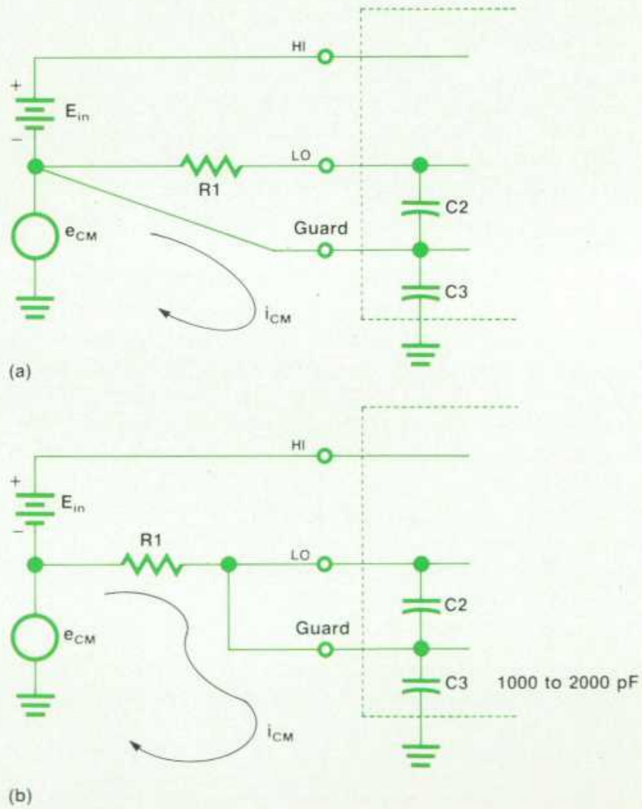

Fig. 3. *(a) The correct way to connect the guard terminal to shunt the common mode current away from the* **LO** *lead. (b) The most commonly used guard connection.*

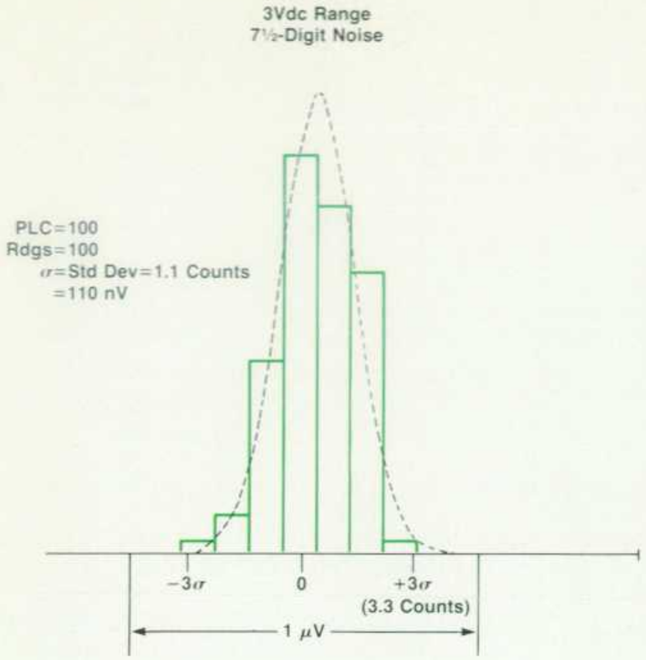

Fig. 4. *Histogram showing typical HP 3457A seventh-digit noise performance.*

are available that can be configured as either ac or dc current inputs or as actuators able to switch 1.5A at 250Vac or 1.5A at 30Vdc.

The HP 44492A reed relay multiplex assembly offers ten two-wire channels able to switch up to 125V peak at a rate of 300 channels per second (see Fig. 6).

Analog-to-Digital Converter

System DMMs must meet stringent measurement requirements and provide great flexibility. HP's Multi-Slope II conversion technique[1] is unequaled in its ability to respond to these needs. Speed, resolution, and noise rejection can be traded off as the measurement situation dictates.

Development of the analog-to-digital converter (ADC), and indeed of the entire instrument, relied upon the use of available, proven technology that could be adapted or enhanced to meet our needs. Multi-Slope II conversion is implemented using the hybrid converter hardware developed for the HP 3468A and 3478A Multimeters.[2] All of the fundamental integrator reference (slope) currents are provided by the hybrid. MOS switches for current source selection and some control logic are also included in the hybrid. Enhancements to the analog and digital hardware along with refinements to the measurement algorithms account for most of the HP 3457A's increased measurement speed and resolution.

Fast decision times are critical to increasing both conversion rate and measurement resolution. An 8051 microcomputer controls both the measurement setup and the ADC. Clocked at 12 MHz, the 8051 allows the ADC hardware to run at 2 MHz. The 0.5-μs cycle times and rich instruction set of the 8051 allowed us to realize a significant improvement in decision times. The 8×8 hardware multiply allowed the measurement calibrations to be overlapped with the conversion, yielding further speed improvements.

Although the HP 3457A uses the same ADC hybrid as the HP 3478A, the similarities stop there. The converter is actually closer to that of the HP 3456A[3] with some hardware enhancements. To improve decision times further, logic external to the microcomputer was added. Part of the counting and Multi-Slope II run-down slope control functions were offloaded from the microcomputer. To maximize conversion speed it is critical to spend a minimum of time in run-down. Run-down is the interval after the input voltage has been integrated (see Fig. 7). To minimize run-down time, it is important not to permit the integrator to overrun a zero crossing too far. Every microsecond of overshoot will require 10 μs at the next smaller slope (1/10 the previous slope) to return to zero. To involve the microcomputer in these zero-crossing decisions would quickly add tens of microseconds to the run-down conversion time.

The input voltage is integrated during the run-up phase of the conversion cycle (see Fig. 8). Unlike dual-slope conversion, Multi-Slope II actually starts converting the answer in the run-up phase. For the run-up interval, the 8051 microcomputer has active control of slope decisions. Slopes in run-up are quantized in fixed increments of time and the 8051 keeps track of the total time spent in each slope direction. During run-down, however, the slope control is performed by synchronous logic clocked by the same 2-MHz signal as the time interval counter (see "Frequency

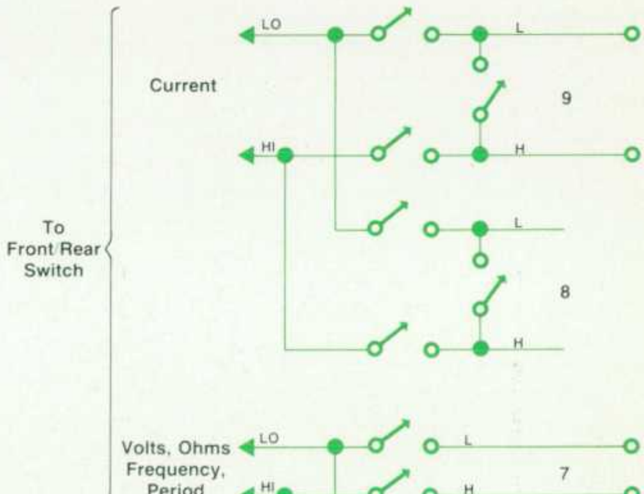

Fig. 5. *HP 44491A Armature Relay Multiplexer schematic diagram.*

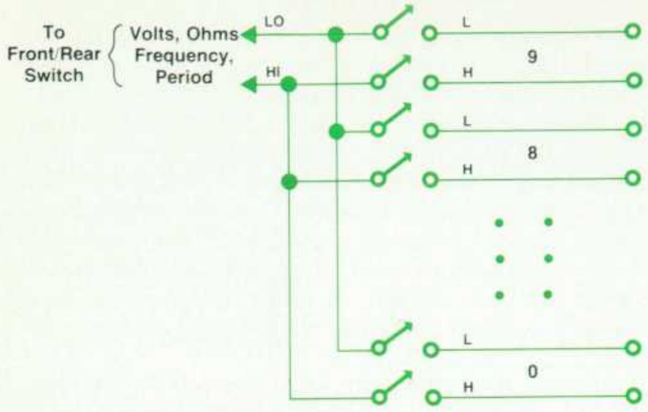

Fig. 6. *HP 44492A Reed Relay Multiplexer schematic diagram.*

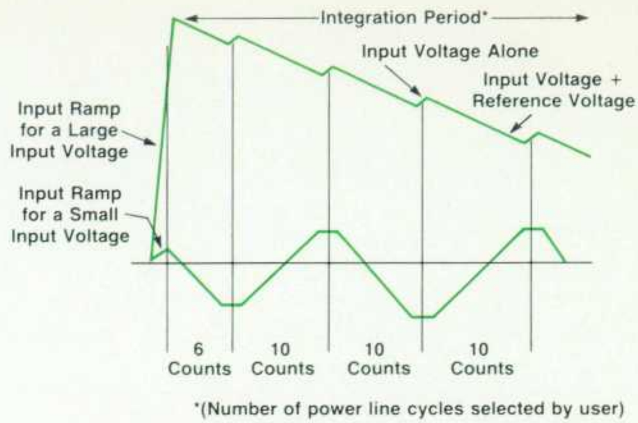

Fig. 8. *During the integration time used in the Multi-Slope II method, the reference voltage ramp is applied periodically at a fixed rate independent of the unknown input voltage level. The polarity of the reference ramp during each period is dependent on the value of the sum of the previous ramp and the input voltage just before the previous ramp was turned off. This approach reduces the voltage level at the end of the integration time, thus reducing the demand on the integrator capacitor for low dielectric absorption. The number of positive reference ramps used during the integration time minus the number of negative ramps determines the most-significant digits of the reading.*

Counter Technique," page 19). Slope times are accumulated in a 20-bit counter enabled selectively from several sources as needed. The time interval counter consists of one of the microcomputer's internal 16-bit counters prescaled by a microcomputer-readable external 4-bit counter clocked by a phase-stable 2-MHz signal. Thus, the run-down slope times are quantized and measured in 0.5-μs increments, and a minimum of slope overshoot is maintained. Initiation of a run-down slope and the selection of the slope magnitude are still performed through microcomputer intervention. After completion of each run-down slope, the external counter is read by the 8051 and the result is accumulated with the appropriate scale factor. Run-down converts the three least-significant internal ADC digits with a total measurement overhead of only 150 μs.

As previously mentioned, the HP 3457A's ADC is very similar to that of the HP 3456A. It was necessary to increase the speed of the analog circuitry to support operation at the higher conversion rates. Less time is allowed for circuit settling, so error sources had to be minimized. A primary source of ADC dynamic error is dielectric absorption from the integrator capacitor. This can be a limiting factor in ADC linearity. The Multi-Slope II algorithm is intrinsically a zero-seeking process, but it still yields an average integrator voltage proportional to the input. Although Multi-Slope II is much less sensitive to such effects than other conversion techniques, dielectric absorption can still introduce linearity errors. The HP 3457A incorporates a new technique to minimize this error. A compensation circuit forces the average voltage on the integrator to approach zero more closely for all inputs. Dielectric absorption then becomes of only secondary importance. A total measurement linearity error (ADC and input conditioning) less than 1 ppm of range is typical. Remaining nonlinearities are caused by resistor self-heating effects, which are tightly controlled through a fine-line resistor process (see next section). Precision current sources are produced using extremely stable HP resistor networks. All ADC performance is achieved without the use of mechanical adjustments or selected component values. Thus linearity of the DMM is virtually unchanged by exposure to any specified environmental conditions.

Fig. 7. *The Multi-Slope II A-to-D technique used in the HP 3457A DMM uses four slopes during the rundown period to determine, in succession, the four least-significant digits of the final reading.*

Long-Term Stability

Long-term accuracy of a DMM is fundamentally limited by three sources of error: the ratio stability of critical resistors, the stability of calibration adjustments, and the long-term drift of the instrument's voltage reference. In a real environment, significant errors may be produced through exposure to vibration, temperature cycling, or humidity changes, conditions prevalent in many system environments.

The HP 3457A uses an HP fine-line resistor process which exhibits extremely low long-term drift. Resistors from this process are used in all critical applications. The resistors of the ADC and input signal conditioning sections that can contribute to long-term drift are examples. Another dominant factor contributing to the long-term stability of an instrument is the ability to hold a calibration adjustment. Environmental factors are of major importance here. The HP 3457A has entirely electronic calibration. Electronic correction cannot account for future changes to mechanical settings, so electronic calibration means that there are no mechanical adjustments in the circuits that can affect calibration integrity. The third factor in DMM long-term stability is the instrument voltage reference. In an instrument with electronic calibration like the HP 3457A's, the voltage reference is by far the dominant term. However, if this were not the case, the calibration adjustment stability could be of similar or greater significance.

The HP 3457A uses a pretested voltage reference assembly to guarantee the long-term stability characteristics of the instrument. Each reference assembly is monitored for a two-month period in a temperature stabilized environment. Individual reference drift rates are determined and compared against acceptable limits. Reference boards that drift beyond these limits are rejected. Much effort was placed on fine-tuning our ability to test and characterize reference assemblies in a production environment so that a reliably low-drift voltage reference could be achieved. Fig. 9 shows the results obtained. The mean long-term drift of the references for one year is characterized to be approximately 3.5 ppm. Also shown is the sample mean plus three standard deviations. This data shows an expected maximum drift of approximately 12 ppm in one year, a significant improvement. The earlier HP 3456A has taken advantage of these results also and an amended data sheet has been issued to reflect this enhancement. Reduced instrument drift translates directly to a lower cost of ownership. Costly system downtime resulting from frequent recalibration cycles can be reduced.

Frequency Counter Technique

The HP 3457A Multimeter can make floating frequency measurements on signals ranging from 10 Hz to 1.5 MHz at levels from 10 mV to 300V. The DMM automatically acquires the input signal, performs the necessary range selection, and measures the input frequency.

This process does not require any user intervention to set up a trigger level control. All signal conditioning is performed by the ac voltage front end and is designed to ensure that the proper internal levels are presented to frequency comparator U3 (Fig. 10).

As mentioned previously, the ADC has external digital hardware to support the measurement of run-down slope times in 0.5-μs increments. This same hardware is used in the frequency measurement mode to accumulate the actual measurement gate time. Since by definition, frequency is cycles or zero crossings (A) per unit of time (B), the job becomes one of measuring these parameters and computing the result (A/B). To do this, the DMM measures the number of zero crossings during the gate period.

Referring again to Fig. 10, one can see that the 8051 microcomputer has two 16-bit counters that are prescaled by 4-bit counters U5 and U6. The 8051 outputs a measurement gate of approximately 0.5 second (the actual length is unimportant). The frequency comparator generates a logic-level signal at the measured signal's zero-crossing rate, which is used to clock flip-flop U4 and the zero-crossing counter U6. U4 provides an enable signal to the time interval (U5) and zero-crossing (U6) counters as long as the 8051 measure gate signal is present. After the enable signal, both counters are read by the 8051, including the external prescalers, and the input frequency is computed as A/B hertz. The actual measurement gate is always guaranteed to be at least one period of the input frequency,

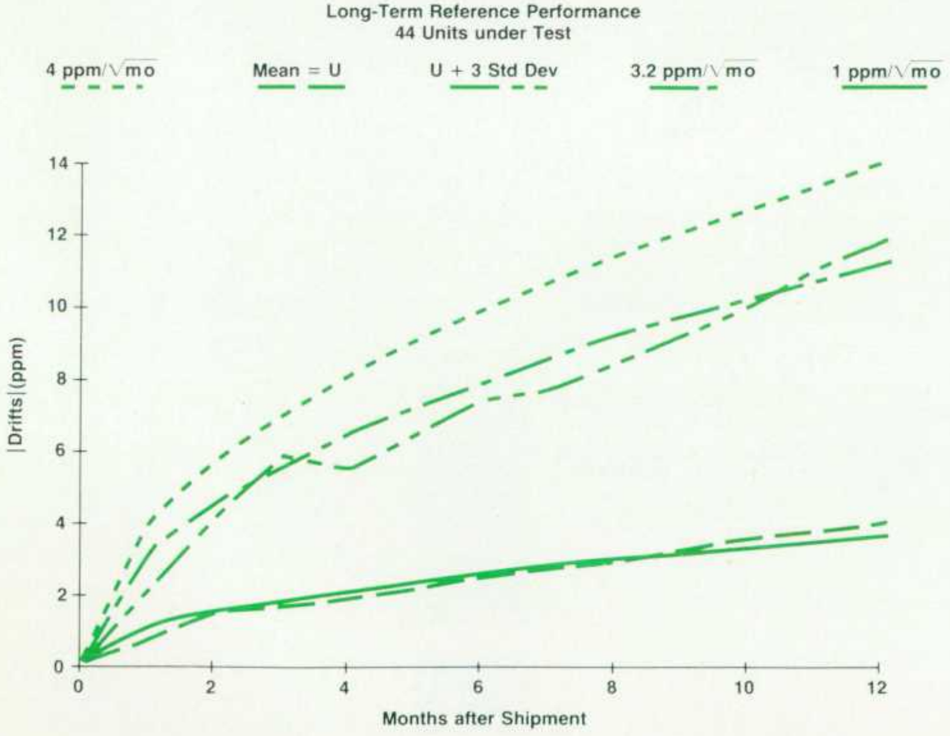

Fig. 9. *Characterized voltage reference drift mean and mean plus 3σ (σ = standard deviation). The 4 ppm/$\sqrt{month}$ represents the reference drift contribution included in the HP 3457A's specifications.*

thanks to the operation of U4. This technique yields a constant measurement resolution of 1 ppm independent of the input frequency two times per second. The frequency is then electronically gain-corrected to yield accurate results.

The benefits of this counter technique are:

- Floating frequency measurements 450V peak from earth ground
- Autoranged input leveling for easy measurements
- Selectable frequency measurements on ac voltage, ac + dc voltage, ac current, or ac + dc current functions
- Constant 1-ppm resolution at any frequency within range.

AC Section

The ac measurement functions of the HP 3457A are voltage, current, frequency, and period. They can be used in either an ac coupled mode or a dc coupled mode. For ac current measurements, the input signal is routed through the same set of shunt resistors used for dc current measurements and the ac voltage drop across those resistors is measured. The current ranges provided are 30 mA, 300 mA, and 3A rms. However, the high input current limit is 1A. Five voltage ranges are provided in decade steps from 30 mV rms to 300V rms full scale. The 30-mV full scale range allows good measurement accuracy of small signal inputs. The ability to measure the frequency of these small signals is another useful feature of the HP 3457A.

The circuitry of the ac section, Fig. 11, consists of two compensated attenuators, two $\times 10$ gain stages, an rms converter, a filter, and various control circuits. Many voltmeters make ac measurements at frequencies up to 100 kHz, but it was desired to have the HP 3457A operate to 1 MHz. To achieve this, compensated RC attenuators are used. At low frequencies the attenuators are resistive dividers, while at high frequencies they act as capacitive dividers. To equalize the high-frequency and low-frequency responses of such an attenuator, a manual adjustment of one of the capacitance elements is often used. In the HP 3457A, that adjustment is performed electronically with the use of the Autocal function. By adding a third capacitor to the attenuator and driving it with a scaled value of the input signal, the capacitive attenuator can be controlled so that a flat frequency response is produced. The signal fed back to the capacitor is controlled by a DAC (digital-to-analog converter) operated as a variable attenuator, as shown in Fig. 12.

The response of the RC attenuator to a pulse input is directly related to the accuracy of its compensation. Therefore, during Autocal, a pulse is internally generated and applied to the input of the ac board. Its time response is measured by the ADC and the change needed in the feedback signal is calculated. The feedback is adjusted and the pulse is remeasured. This process, shown in Fig. 12, is repeated until the desired response is achieved. With this technique the flatness of the attenuator can be set quite rapidly without the need for external equipment.

Range switching is accomplished with the use of FET switches on the input attenuators and following the first gain stage. An rms converter follows the second gain stage and a filter is included in the rms section to reduce the output ripple of the converter. The filter has two bandwidth settings controlled by the ACBAND function; this provides some flexibility in reading speed. Either setting can be used to 1 MHz, but if frequencies below 400 Hz are to be measured or maximum filtering is desired, a number between 1 and 399 is used in the ACBAND function, e.g., ACBAND 20. For higher frequencies or faster response, the user can set the ACBAND function to 400 or greater. The output of the second stage also drives a comparator whose output is used by the frequency counter.

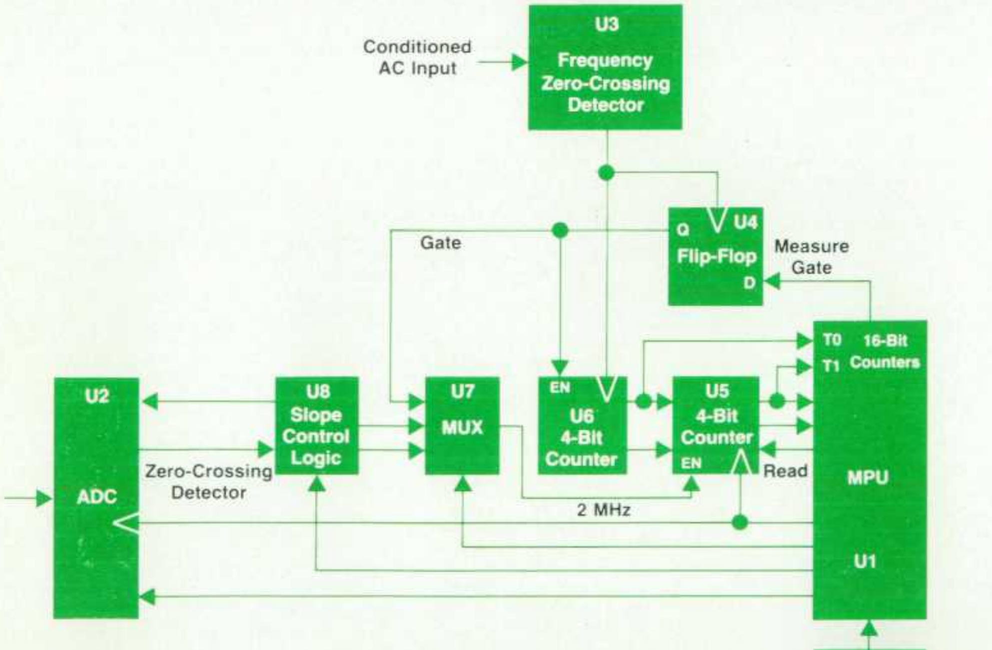

Fig. 10. *Simplified schematic diagram of the analog-to-digital converter and frequency counter in the HP 3457A DMM.*

Calibration RAM Protection

In recent years, software calibration algorithms have become common. By doing the instrument calibration in software, it is possible to eliminate all hand adjustments, and thereby highly automate the calibration process. Typically, such an algorithm requires the user to place some known signal at the inputs to the instrument. The instrument then uses this value to generate appropriate calibration constants. One drawback of this practice is that these constants must be stored in read/write memory. Should this memory be contaminated, the instrument would no longer make accurate measurements. Contamination may be the result of a processor run amok after an electrostatic discharge or some other transient condition that would not normally have serious repercussions.

A frequent solution to this problem is to provide some sort of mechanical device that prevents the processor from writing to calibration memory. This mechanical device might be a simple switch, or a key in a lock. In the case of a key, the calibration is further protected by limiting access to the key. The disadvantages of this approach are that calibration then requires operator intervention, and that the instrument remains vulnerable if the device is left in the wrong position.

The HP 3457A uses a special circuit that prevents the processor from writing to calibration memory unless it can ensure that the processor is executing in a well-behaved manner. This circuit detects that the calibration algorithm has been invoked and completed without error. Special checks are made on various processor status lines to ensure that the processor has not run amok. The instrument software also provides the security of a key by allowing the user to specify a password that must be entered before calibration is allowed. The HP 3457A allows an entire calibration to be done under the control of a computer, with no operator intervention.

The calibration RAM protection circuit checks several events to guarantee that writes to RAM should be allowed. The most fundamental portion of the algorithm is that a processor reset must occur, followed after exactly four milliseconds by a pulse from an output port. Although other things are monitored, these events contain the gist of the algorithm. The requirement that the processor be reset does two things. First, it guarantees that the processor is executing instructions normally. Second, it synchronizes the hardware with the software. The reason for requiring a short pulse exactly four milliseconds after the reset is to ensure that the software is requesting a calibration write enable. This sequence is virtually impossible to generate randomly.

The software procedure is as follows. First, the calibration constants are generated. Then, a reset is generated. The reset routine then determines if the reset was actually a calibration request. This is done by checking numerous strategic variables. If these checks all pass, the remainder of the four-millisecond wait is generated, followed by the pulse and the writes to calibration RAM.

Since the software tests are done immediately after reset, and the hardware requires the reset to enable calibration RAM writes, we have guaranteed that the software tests are all done and passed before calibration RAM is enabled. Herein lies the key to the technique. We have guaranteed that the software tests have been executed and passed instead of merely moving the point of vulnerability to just after the tests have been completed. The hardware is necessary to ensure that these checks are executed and passed. It is then the responsibility of the software to see that sufficiently robust checks are included. If the hardware were not present, the possibility of a randomly executing processor entering the code right after the software checks and writing to calibration RAM exists. But the hardware scheme prevents this by tying the reset event, which initiates the software checks, to the enabling of the calibration RAM.

To ensure the integrity of the calibration memory, we subjected the instrument to large magnetic fields and 15-kV static discharges, and ran tests in which the processor was intentionally traumatized thousands of times. In all of these tests, the calibration protection circuit prevented any loss of calibration constants.

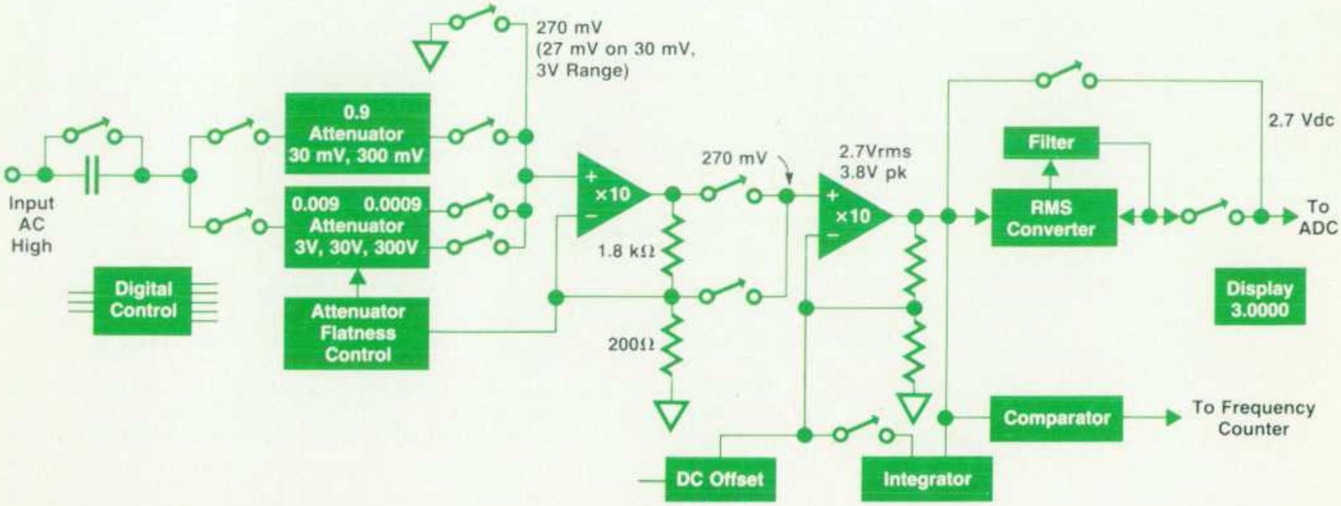

Fig. 11. *HP 3457A ac input section.*

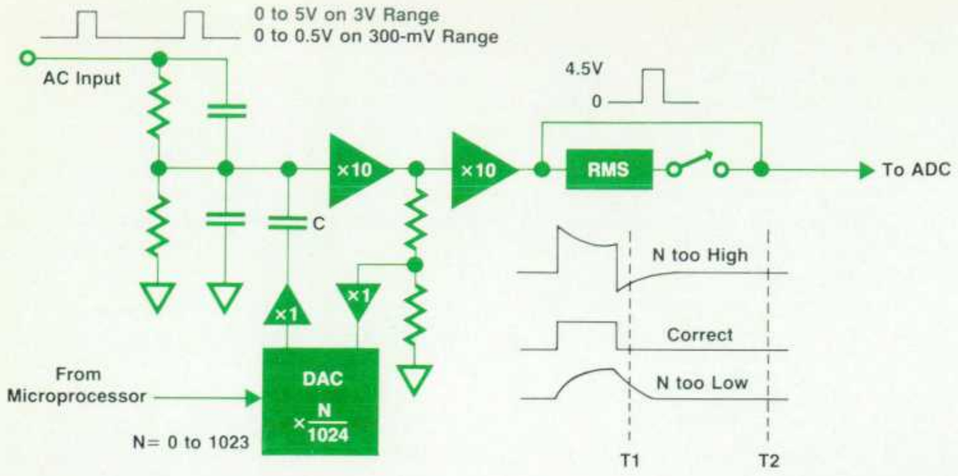

Fig. 12. *Attenuator compensation adjustment technique. When the feedback is correctly adjusted, the voltage readings at T1 and T2 will be equal.*

Command Language

The HP 3457A has many different measurement modes and makes a wide variety of measurements. One of the challenges in the design of the HP 3457A was to provide the user with a convenient technique to access this functionality.

An ideal language lets the user communicate to the instrument the measurement requirements in the terms in which the user perceives the measurement. The instrument must translate these parameters into those required by the hardware. Another important aspect of an instrumentation language is the ability to apply the same language to configure other instruments. By designing a language that describes the measurement parameters instead of the instrument parameters, this is also realized.

The HP 3457A language is made up of about 100 mnemonic instructions. Of these instructions, only about a dozen are required to meet most measurement needs. Each instruction can have several parameters. For convenience, defaults are available for each parameter except in rare circumstances where the value is extremely critical, e.g., calibration.

An example of a measurement-oriented instruction is the function command. A typical function command might be ACV 17,0.001. This indicates that an ac voltage measurement should be made on a range capable of accepting at least 17 volts, and that the measurement should be made to a resolution of at least 0.001%. Notice that this instruction could be used to configure any instrument capable of making this measurement since no part of the instruction is specific to the HP 3457A. Future HP multimeter products will use this same language, which is called HP-ML (HP Multimeter Language).

Reading Storage

A feature of the HP 3457A is its ability to store over 1000 readings internally. This can provide significant speed advantages in systems because many measurements can be made and the overhead of reading from the instrument need only be encountered once. This capability is also important when the measurements need to be made faster than they can be sent to a controller. The results can be extracted from memory either by reading from the instrument, in which case the reading memory behaves like a buffer between the instrument and the computer, or by explicitly specifying a group of readings to be recalled.

Readings can be placed in memory and read out in any one of four different formats: 16-bit integer, 32-bit integer, 32-bit real (IEEE 754), or ASCII (the number is converted to an ASCII string). Since the output format is independent of the storage format, it is possible to optimize the timing for a particular measurement. For instance, in a situation where computer time is of concern, the instrument can be configured to store readings in the same format that will be output to the computer. When the computer is ready, the formatted readings can be sent out at high speed. When the reading rate is the critical factor, the instrument should be configured to store readings in the format most convenient to the voltmeter, thus optimizing the reading rate.

When the readings are recalled from memory they can be converted to the format most convenient to the computer.

Mechanical Design

Product design plays a key role in the manufacturing cost of an instrument. To minimize factory cost, a major HP 3457A objective was to minimize part count, part cost, and assembly labor. Other objectives were to keep the mechanical design one phase ahead of the electrical design, to strive for simplicity and to use existing technologies.

It was decided early in the project to use an existing plastic package. This package was originally designed for the HP 3421A Data Acquisition/Control Unit and has since been adapted to the HP 3488A Switch Control Unit. Because this package is plastic, shielding and internal temperature rise were concerns, more so than with traditional metal packages. In addition, environmental requirements such as shock, vibration, and drop tests, as well as safety issues had to be carefully considered. It was desirable for manufacturing to build only one configuration. This required that the customer have access to line voltage selector switches and be able to install the optional multiplexer card options.

To achieve these objectives and requirements, the initial layout was closely evaluated using cardboard mockups and design reviews. As the design evolved it became obvious that hardware such as screws and fasteners could easily be eliminated by using snap-together parts. These included the shield around the isolated analog section, the card guide for the optional plug-in multiplexer cards, standoffs for the printed circuit boards, and a snap-in support between the isolated and nonisolated sections.

System voltmeters have isolated and nonisolated sections. In addition to good shielding, the isolated section requires low capacitance and high impedance to ground or the nonisolated section. Traditional shields are fabricated of aluminum held together with screws to provide multiple electrical contact. Fig. 13 shows the HP 3457A snap-together assembly made of two mating U-shaped sheet-metal shields, which snap together using tabs combined with sheet-metal deformation. In general, the more contact points and the smaller the separation between shields the better the effectiveness will be. This snap-together design creates ten points of contact approximately every 75 mm with minimal separation. The deformation occurs to the top shield; a double bow is created when it is snapped into the bottom shield's two tabs. The top shield's vertical folds combined with a high middle contact point create three high points and the two tabs create the low points, as shown in Fig. 14. Depending on the three fold lengths and the hole-edge-to-hole-edge tolerance between the two tabs and the high middle contact point, the amount of bow varies. The contacting points of the bottom shield between folds are designed so that they are always in tension. This snap-together design incorporates standard sheet-metal tolerances for ease of fabrication.

In addition to the electrical shielding and isolation requirements, this section requires good mechanical isolation from externally induced shock or vibration. A high-impedance polycarbonate snap-in support attaches this isolated section to the firm nonisolated section. The card guides used to hold the optional multiplexer cards are also molded from polycarbonate. The card guides snap into cutouts punched into the supporting sheet metal. The ac board has a shield separating it from the main analog board. Snap-in spacers hold this inner shield and make point contact with the upper and lower boards. A single screw holds the ac board and shield in place. Rotational motion is eliminated by keying both board and shield in a cutout in the vertical side sheet metal.

The front panel incorporates a silicone elastomer keyboard, a concept borrowed from the HP 3468A/3478A Multimeters.[2] The entire keyboard is contained on three

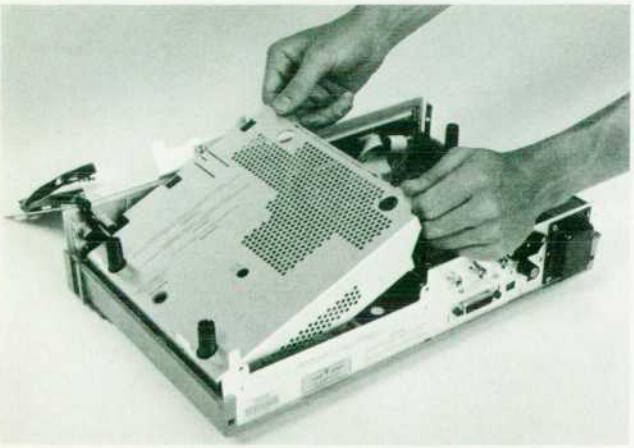

Fig. 13. *HP 3457A snap-together assembly.*

Fig. 14. *Side view of the isolated section's top shield showing the intentional double bow. The top shield snaps into a mating U-shaped shield without the use of a tool.*

keypads containing either twelve or sixteen keys. The elastomer keyboard is sandwiched between the plastic molded front panel and a printed circuit board by heat staking. The elastomer key concept has proved to be a highly reliable, simple, low-cost design. The front panel also contains a low-power 12-character alphanumeric liquid-crystal display. Input terminals are traditional binding posts, which are made of copper to minimize any thermally induced voltages.

Inherent in this package is the ability to stack parts. The six long case screws that clamp the two case halves together also capture the sheet-metal isolated and nonisolated sections via spacers. The printed circuit boards are screwed to the sheet metal for good electrical contact. The front and rear panels slide into recesses in the upper and lower case halves and therefore do not require any hardware.

Other design features keep the assembly time low. Connectors and push-on lugs minimize the number of hand solder joints. All unnecessary parts were eliminated to keep material costs low and to minimize assembly time. For example, nomenclature on the rear input terminals is molded into the plastic insulator that holds them. Hardware such as screws, when used, serves multiple purposes in many cases. A bar code is used for the serial number to speed paperwork and minimize the chance for errors in processing.

A special multifunction fixture is used for assembly. This fixture is used to hold the front and rear panels during preassembly. It also aids in the installation of the plastic isolation support in the digital section, and holds the bottom clamshell to install the case screws.

Acknowledgments

In addition to the authors, the team that developed the HP 3457A included Jim Vyduna (ADC firmware and hardware), Gerry Raak (option cards, environmental and abuse testing), and Les Hammer (digital hardware). Special appreciation is expressed for the guidance and support of Joe Marriott, who was engineering section manager during the development of the HP 3457A.

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Advanced Scalar Analyzer System Improves Precision and Productivity in R&D and Production Testing

This "voltmeter for the microwave engineer" measures insertion loss and gain, return loss, and power quickly and accurately.

by Jacob H. Egbert, Keith F. Anderson, Frederic W. Woodhull II, Joseph Rowell, Jr., Douglas C. Bender, Kenneth A. Richter, and John C. Faick

CONTINUING ADVANCES in RF and microwave component and system designs constantly generate a need for more accurate and versatile test equipment. When the HP 8755A Scalar Analyzer[1] was introduced, its ability to measure insertion and return loss simultaneously in a manual test system satisfied a large portion of the microwave design engineer's test equipment needs. With advances in technology, however, has come the need to make more accurate measurements over a wider dynamic range. Measuring the rejection response of a filter, for example, may require a wide dynamic range along with real-time response so the effects of adjustments on the filter characteristics can be observed as they are made. In addition, many devices require simultaneous viewing of three or more measurements and not just the traditional two of the past.

The requirements of microwave component and system manufacturers for automated test equipment in production areas have also increased significantly. There is a need for flexible measurement systems that are easily configured to test products on production lines. Automated testing can ensure that each device is tested completely and exactly the same way. Another need of the ATE environment is the ability to build a production history based on test results. This allows manufacturers to monitor their production processes closely and improve efficiency while at the same time increasing the quality of their products.

The requirements placed on a scalar analyzer system by today's RF and microwave component and systems manufacturers have led to the development of the HP 8757A Scalar Analyzer System (Fig. 1). The HP 8757A is able to function as a manual test system or a fully programmable ATE system that can execute the automated testing and data logging required in the production environment. The HP 8757A Scalar Analyzer, together with the HP 11664 and HP 85025 Detector Families and the HP 85027 Direc-

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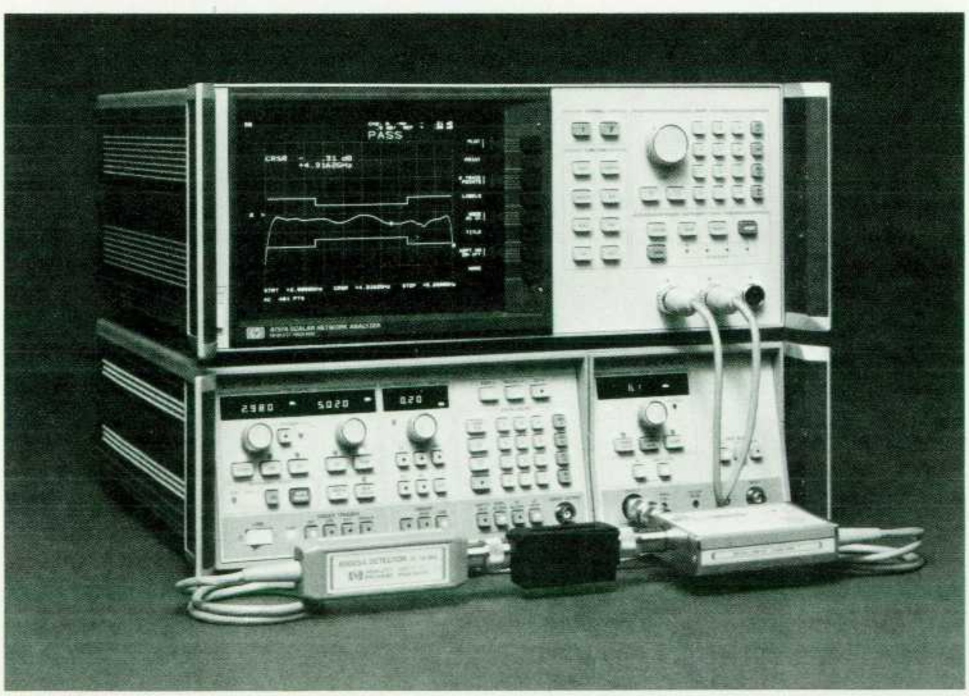

Fig. 1. *The HP 8757A Scalar Network Analyzer (shown here with the HP 8350B Sweep Oscillator and an HP 85025 Series Detector) measures insertion loss and gain, return loss, and power quickly and accurately to characterize microwave components and systems over a frequency range of 10 MHz to 40 GHz. It has four independent display channels and three or four detector inputs.*

Filter Measurement with the Scalar Network Analyzer

The measurement or design of filters can provide excellent examples of how the HP 8757A Scalar Network Analyzer can make device characterization easy. In the production environment, where throughput is critical, the HP 8757A can quickly verify performance specifications. In the development lab, the analyzer can measure and display broadband and narrowband interactions simultaneously.

Fig. 1 shows a display of a bandpass filter measurement. Two channels are used. Channel 1 displays the measurement of the filter over its passband while Channel 2 displays a broadband measurement of the same device. The measurements are made as the source alternately sweeps over a narrow band on one sweep and a broad band on the succeeding sweep. This alternate sweep mode can be entered via the **ALT** key on the source front panel if the source's receiver interface is connected to the analyzer's system interface. Each state of the alternate sweep mode can be saved or recalled for quick setups.

During the filter design phase, it is common that parameters that alter passband performance also influence shape factor, bandwidth, or out-of-band rejection. Alternate sweep mode provides a continuously updated display of the in-band and out-of-band performance so that the effects of each adjustment are displayed almost in real time.

Design limits can be entered into the HP 8757A with the built-in limit lines (point, slope, or flat line). These limits adjust with scale, reference, and frequency changes and are more than a simple display "grease pencil." When the device performance is within the entered limits, a large **PASS** appears in the channel annotation area; otherwise, **FAIL** appears. This simplifies production testing for either manual or automated setups.

All HP 8757A features, including error indications such as limit line pass/fail, are accessible over the HP-IB. The cursor and marker functions further simplify filter measurements. Fig. 1 shows seven marker symbols on trace 1 and two symbols on trace 2. The HP 8757A provides cursor-to-maximum/minimum and cursor-to-reference functions, a cursor delta mode, and cursor search and bandwidth functions. In addition to the cursors, the analyzer displays up to five source markers. In Fig. 1, cursor-to-maximum and cursor-to-reference operations have the point of least insertion loss on trace 1 at the reference position at the top of the screen. Entering cursor delta mode and performing a cursor-to-minimum operation then causes the analyzer to show directly the peak-to-peak amplitude variation within the passband of this filter. Shape factors (60-dB bandwidth/6-dB bandwidth is one example) can be determined quickly using the **BANDWIDTH** function under the **CURSOR** menu.

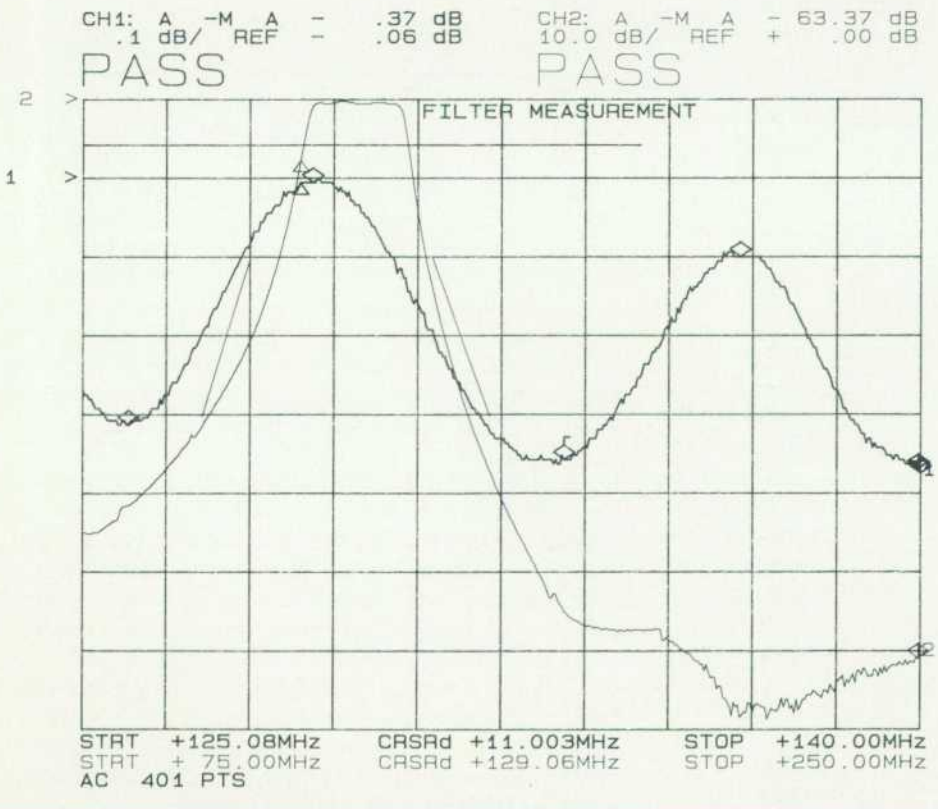

Fig. 1. *HP 8757A cursor and marker functions simplify filter measurements like this one.*

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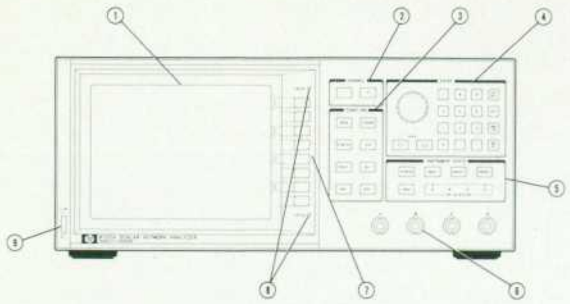

Fig. 2. *User areas on the HP 8757A front panel. 1. CRT and softkeys area. 2.* **CHANNEL** *area. 3.* **FUNCTION** *area, 4.* **ENTRY** *area. 5.* **INSTRUMENT STATE** *area. 6. Detector inputs. 7. Softkeys. 8. CRT Controls. 9. Line switch.*

tional Bridge family, offers new levels of flexibility and performance for RF and microwave component and system measurements.

Instrument Features

The key features of the HP 8757A system are:

- 10 MHz to 60 GHz frequency range.
- Five inputs. The **A**, **B**, **C** (optional), and **R** inputs have −60-dBm sensitivity and 76-dB dynamic range. Each is capable of making measurements of either modulated or unmodulated RF. The auxiliary rear-panel **ADC INPUT** can be used to measure voltages between −10V and +10V.
- Four display channels, each capable of displaying any of the selected measurements. The large 9-inch vector display shows each channel fully annotated.
- Modulated (ac) and unmodulated (dc) RF measurement capabilities. With the HP 85025 Detectors and the HP 85027 Directional Bridges, measurements can be made using either of these detection methods. This allows the user to choose the detection method that best suits the measurement.
- Limit lines. These allow the user to enter test limits through the front panel and have the analyzer do pass/fail testing without an external controller.
- Direct plotter and printer output. The graphics presentation on the CRT may be output directly to an HP-IB plotter or printer without an external controller.
- Extended firmware features. Enhancements have been added to increase the efficiency of the user and make the user interface friendlier. Some of these are a bandwidth function, cursor search, autoscale, and adaptive normalization.
- Save/recall capabilities. This function allows the user to configure a test setup including calibrations and store it away in nonvolatile memory. This configuration may then be recalled when the same measurement is to be made later.
- System software packages. Software packages make it easy for the ATE test engineer to configure the test software to fit production needs. Each package offers a friendly user interface and provides the flexibility to configure the tests to the test device criteria.
- Fully HP-IB (IEEE 488/IEC 625) programmable.
- Test firmware. A comprehensive set of test firmware aids in the verification, testing, and troubleshooting of the instrument.

User Interface

In an effort to make the instrument as friendly and simple as possible to operate, a menu and softkey approach is used for the front-panel operation. The front panel of the HP 8757A is divided into five distinct areas (Fig. 2). These five areas are arranged to assist the user in making measurements quickly.

The CRT display is a large 9-inch vector graphics display. Channel information, the active entry area, softkey labels, frequency labels, a graticule, and measurement traces are displayed here. The channel information is annotated for each of up to four channels, since each channel can be set up independently. The channel annotation includes the measurement being made, the cursor amplitude (or active marker amplitude value if the cursor is off), the scale factor per division, the reference level value, and other symbols indicating the enabling of certain functions. The active entry area shows the currently active function, such as the reference level or the scale factor per division.

The **CHANNEL** area allows for the selection of Channel 1 or 2. Selection of either of these keys also brings up the channel selection menu to the left of the softkeys. Control of the channels and the selection of Channels 3 and 4 is provided by this menu.

The **FUNCTION** area provides access to the measurement parameters. These include the measurement (A, R, B/A, etc.), the display (measurement, memory, or measurement minus memory), the scale factor per division, the reference values, the cursor, averaging, calibration, and special functions.

The **ENTRY** area is used for data entry by virtually all functions through the use of the keypad, step keys, or knob.

The **INSTRUMENT STATE** area provides user control of the save/recall registers, HP-IB addresses (with the **LOCAL** key), and other system-level functions (affecting all channels).

Each of the front-panel keys, except for the **PRESET** key and the **ENTRY** area keys, always brings up a menu of softkey labels, which appear on the CRT display to the left of the softkeys. The eight softkeys are integrated into the display bezel, thus coupling the action of the keys with the functions described by the softkey labels. To maintain user friendliness, nesting of the menus has been kept to one level as much as possible. Exceptions to this are calibration routines in which the operator is led through the procedure by user prompts. The softkey menu structure for the HP 8757A is shown in Fig. 3.

This layout guides the user to set up the measurement logically. First, the active channel is selected using the **CHANNEL** area. This active channel is then highlighted on the CRT display by its higher screen intensity. Next, the **FUNCTION** area is used to make changes to this channel's settings. For example, pressing **CHANNEL 1** followed by the **MEAS** key activates the measurement function and brings up the **MEAS** menu. The user can then change the measurement by choosing the desired power or ratio measurement with the softkeys. The display desired, either measurement, memory, or measurement minus memory, can be selected using the **DISPLAY** menu. Also, the user can enter data into

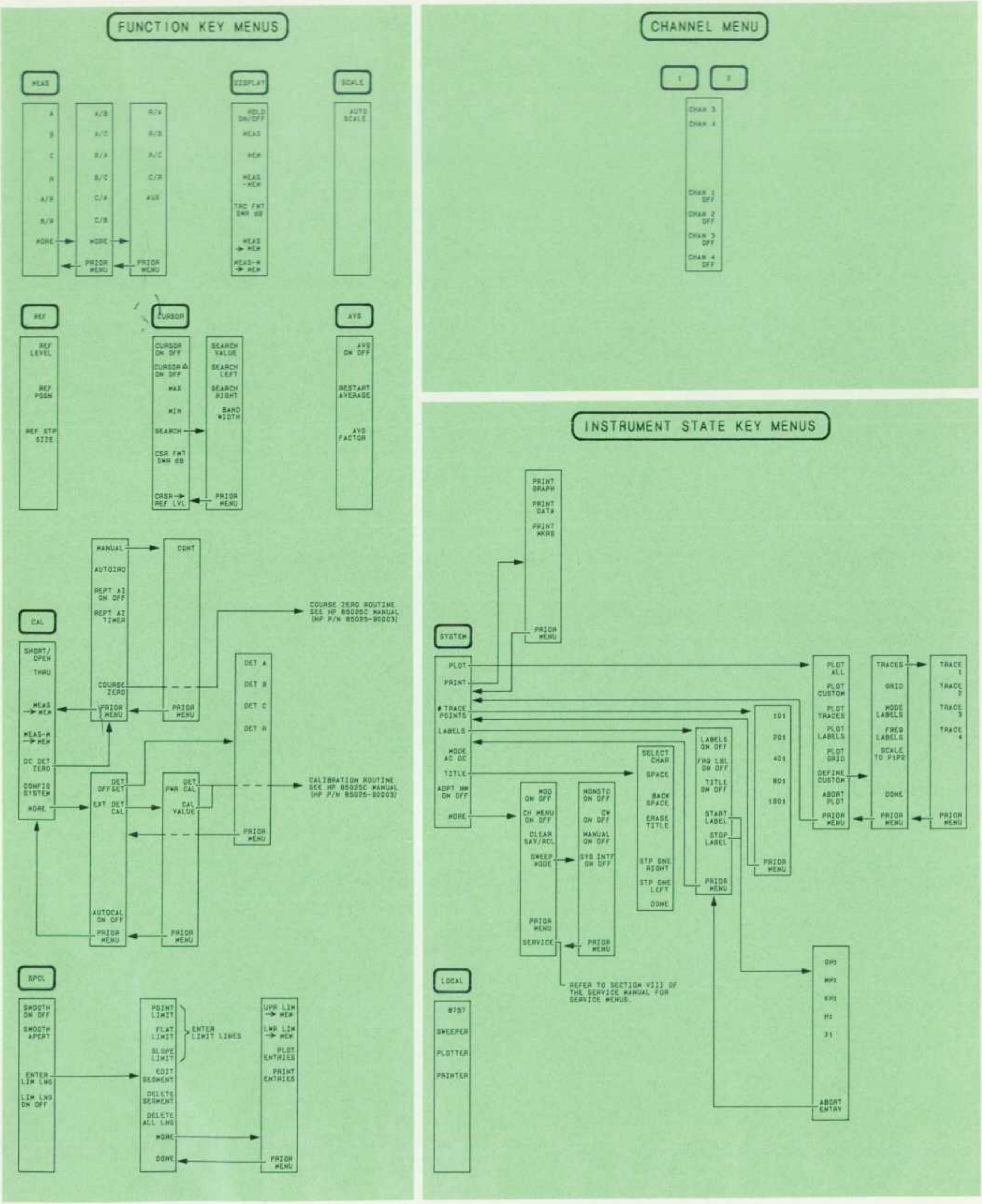

Fig. 3. *HP 8757A softkey menu structure.*

memory using this menu.

Other useful functions that enable the user to set up and begin measurements quickly are: the **SCALE** factor per division, the **REF** menu, which allows changes of the reference line position and level, and the AUTOSCALE function under the **SCALE** menu, which automatically adjusts the scale factor per division and the reference level so that the trace appears fully on-screen at the highest resolution. AUTOSCALE makes it easier for the user to observe a measurement quickly without having to adjust the reference level.

One of the advantages of a menu-driven instrument is that, with careful design, the operation of the instrument can be easy to understand. For example, to control the labeling functions, the user selects the **SYSTEM** menu and looks for the softkey LABELS. After selecting this softkey, the possible functions are to turn on or off all of the labels, the frequency labels, or the title.

The user can also be prompted to take certain actions before continuing, or can be warned that the action attempted will result in a measurement error or is not allowed. An example of this prompting action occurs when the user selects SHORT/OPEN under the **CAL** menu. This function leads the user in performing a short/open calibration for a reflection (return loss) measurement. The prompts are displayed in the active entry area in the upper left portion of the graticule area. When SHORT/OPEN is pressed, the message CONNECT SHORT...STORE WHEN READY appears in the active entry area. The user connects a short circuit to the test port of the directional bridge/coupler and presses the STORE SHORT softkey. A new prompt, CONNECT OPEN...STORE WHEN READY, then appears in the active entry area. When this action has been completed, the user presses the STORE OPEN softkey. This causes the message SHORT/OPEN CAL SAVED IN CHx MEM (where x is the channel number) to appear, notifying the user that the calibration has been saved.

Limit Testing

A necessary procedure in the design and production of RF and microwave components, as well as in other application areas, is the comparison of device performance with specifications. With scalar network analyzers, this typically was done by setting the reference level and reference position equal to the specification. This line was used to verify visually that the device met its design or manufacturing specification. This method allowed only one specification to be checked, so a user would often draw lines on the CRT with a grease pencil for other specifications. Furthermore, since the comparison was done visually, the verification was subject to error.

With the HP 8757A, a user can enter up to 12 limit tests, called segments, for both Channel 1 and Channel 2. Each segment can be a point, a flat line, or a sloping line with upper and lower bounds. Entry is accomplished by selecting the **SPCL** (special) menu and pressing the ENTER LIM LNS softkey. The user chooses the type of limit (point, flat, or slope), and through a series of prompts, enters the frequency and the upper and lower limit values. Limit testing is turned on or off using a softkey under the **SPCL** menu.

Limit test results (pass or fail) are shown in large letters below the channel for which the limit tests were entered. The limit test compares trace data with each limit segment with 0.01-dB resolution regardless of the scale factor or reference level settings, eliminating the inaccuracies of a visual comparison. Fig. 4 shows a filter response with one set of flat limits and two sets of slope limits. The large PASS indicates that the device meets its specifications.

When the HP 8757A system interface is not connected to a compatible sweep oscillator or is turned off, the user may enter start/stop frequencies using the **SYSTEM** menu, selecting the LABELS softkey. The LABELS menu has softkeys for the start and stop frequency entries. These label values are then used for limit line testing, if it is enabled.

Auxiliary Input

The HP 8657A provides the ability to measure a voltage between −10V and 10V at its rear-panel **ADC INPUT** for display of other than the normal **A**, **B**, **C**, and **R** inputs. This input can be used to measure phase detector outputs, control voltages, or other device outputs and display them simultaneously with the usual transmission or reflection

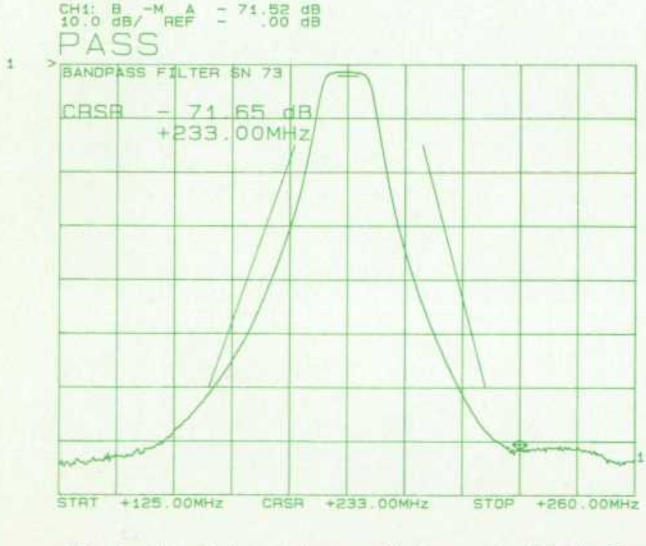

Fig. 4. *Filter limit test display with three sets of limits. The filter passes the test.*

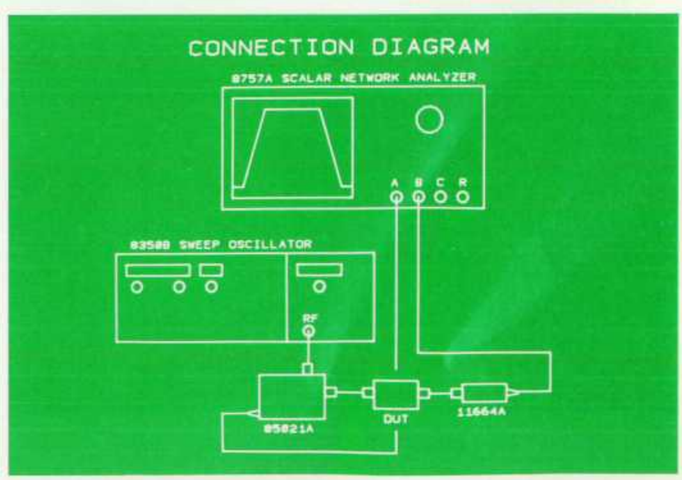

Fig. 5. *Graphics capability makes it possible to display connection diagrams along with instructions to the operator.*

Scalar Analyzer System Error Correction

For making a power measurement in the HP 8757A, the basic signal flow is shown in Fig. 1. The microwave signal is first detected and converted into a 27.78-kHz IF signal. The log amplifier then produces an output proportional to the logarithm of the IF signal, and that output is read by the CPU via the ADC. The ADC readings are related to power input, but contain the effects of two potential sources of error: the input-versus-output relationships of the power detector and the log amplifier. The detector's response curve of voltage output as a function of power input is shown in Fig. 2a. It follows a square-law relationship below −30 dBm (voltage output is proportional to power input), a linear relationship above about +10 dBm (voltage output squared is proportional to power input), and a smooth transition between. Although all detectors follow this general curve, there can be variations from detector to detector in sensitivity (offset), curve conformance in the transition region, and saturation for high input power. Also, the response curve for a given detector may vary with temperature.

The log amplifier, which produces an output proportional to the logarithm of the input signal, is the second potential source of error (see Fig. 2b). The log amplifier's response exhibits ripples and curvature, resulting in a maximum 3-dB deviation from the ideal response. The log amplifier's gain is inversely proportional to temperature, causing a 0.028-dB/°C offset at the output when the temperature changes.

Fig. 1. *Basic signal flow for a power measurement.*

Since the two major sources of error are unrelated, error correction is performed in two separate steps. The log amplifiers are initially calibrated using a special step attenuator to remove their errors (see box, page 33). For each detector connected to the input, a response curve of power input versus voltage output is generated using the specific detector's characteristics. The log amplifier's response curve is combined with the detector response curve to find the system response of detector power input versus ADC readings (see Fig. 2c).

Each log amplifier is calibrated by applying a logarithmically decreasing 27.78-kHz signal and recording the resulting ADC readings in an EEPROM table for that input. A temperature sensor on the log amplifier board is used by the CPU to correct for temperature drift. The log amplifier, its calibration tables, and the temperature correction routine are combined so that the HP 8757A can be considered a narrowband temperature-compensated 27.78-kHz logging voltmeter with flat response over a 140-

(continued on next page)

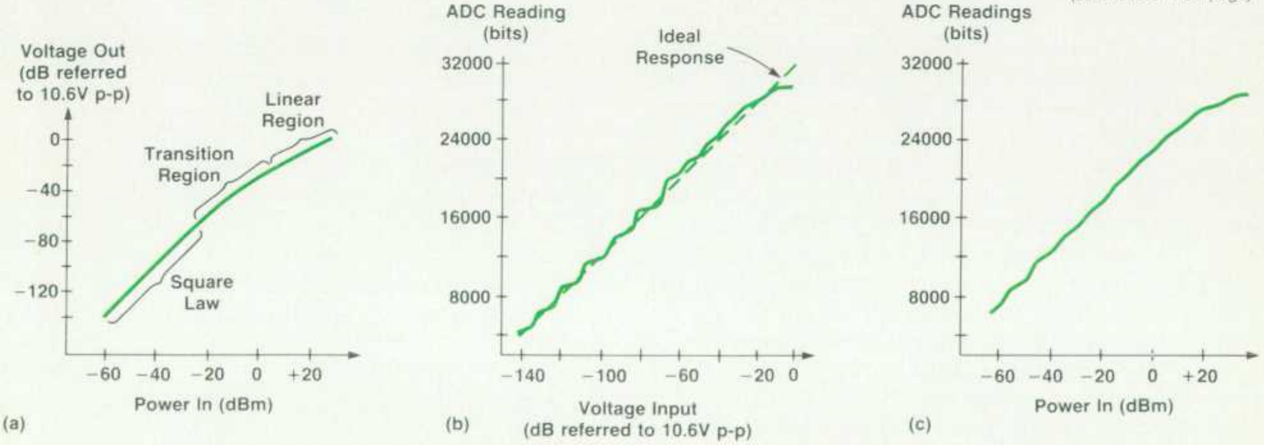

Fig. 2. *(a) Detector response, (b) Logarithmic amplifier response (errors exaggerated). (c) HP 8757A system response (errors exaggerated).*

measurements.

The auxiliary input display is annotated in units of volts, including reference level, scale factor per division, limit lines, and cursor readouts.

Averaging and Smoothing

For some measurements, displayed noise peaks can be reduced by trace-to-trace digital averaging or smoothed by the trace smoothing function. The HP 8757A performs stable exponential averaging.[2,3] Each data point is an average of the current trace with the previous N−1 traces. N, the averaging factor, is user-selectable via the keypad and the knob and must be a power of two. Averaging can be restarted by the RESTART AVERAGE softkey. The HP 8757A automatically restarts averaging in response to changes of the measurement channel and to changes in sweeper settings if the sweeper source/receiver interface is connected. This results in quick response to measurement changes with a minimum of user keystrokes.

Trace smoothing is similar to low-pass video filtering. It provides an easy way to view noisy traces, and can be used to eliminate effects of source mismatch. When the measured data changes more slowly than the source match variations, the smoothing aperture can be set to smooth away measurement ripples caused by the mismatch without otherwise affecting the displayed measurement. The dB signal range.

For detector correction, the HP 8757A uses the control line in the detector cable to sense a resistor that identifies the detector type. In the ac/dc detectors and bridges, this line is also used to measure a resistor representing detector sensitivity and a thermistor that senses the detector temperature. (The ac detectors and bridges are all adjusted to have the same sensitivity and are internally temperature compensated so that a thermistor and a sensitivity resistor are unnecessary.)

Once the detector type, sensitivity, and temperature are known, an equation is created relating the detector output voltage to its input power. This detector equation is combined with the log amplifier EEPROM table to create a power table relating ADC readings to the detector input power. Each HP 8757A input port has its own power table stored in RAM. In the event of an EEPROM failure, the system will use the table of an adjacent log amplifier for default data. If all EEPROM tables have failed, a default table will be generated in firmware.

Of course, if the temperature of the log amplifier or the ac/dc detector changes, or if a different detector is connected to a given input port, the power table stored in RAM for that input must be updated, which takes about 1.5 seconds per channel. To provide real-time temperature correction, the CPU computes an approximate offset correction factor over a narrow temperature range for each input's power table. If a significant change in temperature occurs, the power tables are completely regenerated.

The main advantage of correcting the log amplifier errors and detector variations separately is flexibility. For a given log amplifier, widely different detectors may be used by modeling the detector in firmware with a few simple equations and combining these equations with the table characterizing the log amplifier for the overall system response. Compared with the older HP 8755A/56A, the system calibration has been simplified in that the HP 8757A does not depend on an accurate RF level and HP 11664 Detector for accuracy. Instead, a 27.78-kHz signal is applied directly to the HP 8757A input, although it must be accurately controlled over a 140-dB attenuation range. Finally, temperature correction is easily implemented, since it is handled in firmware for the log amplifier and the detector separately before they are combined for the total response.

aperture is settable from 0.1% of span to 20% of span via the **ENTRY** area. Each smoothed data point is an average over the aperture, centered about the current position. Edge points are smoothed averages over a smaller aperture determined by the number of measured data points available from the point to the beginning or end of the trace.

Normalization

The HP 8757A normalization functions provide a simple means of canceling source and receiver accuracy and flatness errors and test fixture and cable frequency response errors. A softkey sequence can be used to store SHORT/OPEN averages or THRU calibrations with step-to-step display prompts. The data-into-memory (MEAS→MEM) and data-minus-memory-into-memory (MEAS – M→MEM) softkeys can also be used to store the calibration. Measurements can then be made relative to the stored reference with improved accuracy. Each calibration stored is valid over a specific frequency range. Recalibration is required for any change in source frequency settings. The HP 8757A can automatically perform the recalibration if in adaptive normalize mode. This mode is available through the **SYSTEM** menus and requires the user to perform one broadband calibration. As source frequency settings are altered, the HP 8757A automatically regenerates an interpolated calibration memory corresponding to the new frequency settings of the compatible source connected to the HP 8757A **SYSTEM INTERFACE** port. When frequency settings extend outside the original stored broadband calibration frequency limits, the HP 8757A extends the value of the originally stored endpoints to cover the out-of-band region. A U uncalibrated indication appears on the display to indicate out-of-band calibration memory. With adaptive normalization, the user does only the one-time calibration. The need for time-consuming recalibrations and extra device handling as frequency settings are altered is minimized.

Trace Points

The HP 8757A allows a variable number of trace points. For applications requiring real-time feedback for the operator while adjusting the DUT, sweep speed and refresh rate are critical. The HP 8757A allows as few as 101 trace points to obtain the maximum sweep rate. For a broadband measurement or for measurements of high-Q devices, the HP 8757A provides maximum resolution with up to 1601 trace points per sweep. The number of trace points is selectable using the softkeys. The HP 8757A automatically limits source sweep rates via its **SYSTEM INTERFACE**. This allows the operator to be concerned only with DUT response time errors as sweep speeds are changed.

Sweep Modes

During normal operation, the HP 8757A requires an input sweep ramp of 0 to 10V. A tracking digital-to-analog converter (DAC) causes measurements to occur at evenly spaced intervals over the 10V range. Zero volts corresponds to the start frequency and 10V corresponds to the stop frequency. Not all measurement setups supply a 0-to-10V ramp. The HP 8757A is capable of normal operation in these nonstandard setups.

The HP 8757A can make measurements with input sweep ramps less than 0 to 10V when the nonstandard sweep mode is selected via softkeys. Restrictions are that $V_{start} \geq 0V$, $V_{stop} \leq 10V$, and $V_{stop} - V_{start} \geq 2V$. On entering this mode, the HP 8757A searches for the minimum and maximum sweep voltages to determine the sweep endpoints. This information is used for setting the tracking DAC start point and increments as the DAC tracks the input sweep voltage. Measurements can be made easily on voltage controlled devices whose inputs are provided by a ramp generator with variable-voltage outputs.

For measurement setups that do not provide sweep inputs, the HP 8757A can be set to a nonswept CW mode. In this mode, a sweep ramp is generated internally. Measurements are made continuously with rapid display updates, making power-meter-type measurements quick and easy with no loss in functional capability.

The HP 8757A can also make measurements in a manual sweep mode. This mode is entered via the **SYSTEM** menu or via the **MANUAL** key on the source. In this mode, the source is tuned to a fixed frequency. The HP 8757A continuously performs measurements and updates the cursor indicator as the frequency is tuned manually and the measurement changes. With a frequency counter, the exact frequency of measurement points of interest is easily determined.

dB or SWR Data

Cursor information can be displayed in units of dB/dBm or standing wave ratio (SWR). The user can also display the entire trace in SWR format and have complete control over the scale per division and the reference level. Limit line tests can be entered and the traces can be plotted and/or printed in SWR. This feature is particularly useful for devices manufactured or specified in terms of SWR. The HP 8757A eliminates the tedious manual conversion of dB (return loss) data to SWR by using an internal lookup table for fast trace updating.

Interpolated Cursors

The HP 8757A introduces cursor search and bandwidth functions. The bandwidth softkey can be used for a quick reading of the N-dB bandwidth (e.g., the 3-dB bandwidth). The search and bandwidth functions interpolate between data points to find the exact N-dB values. The value N is set by the knob or the keypad.

Without the search and bandwidth functions, it may not be possible to place the cursor exactly on the 3-dB frequency point. Magnitude values are digitized with 0.003-dBm resolution for power measurements and 0.006-dB resolution for ratio or normalized measurements. In all cases, the magnitude value is rounded to 0.01-dB resolution for display. For a frequency sweep of 1000 MHz digitized at 401 points, the frequency resolution is 1000/401, or about 2.5 MHz per increment. Therefore, moving the cursor one increment results in a frequency change of about 2.5 MHz. The search and bandwidth functions interpolate between these points to locate the exact 3-dB frequency.

Hard-Copy Output

The ability to generate a hard-copy representation of the graphics on the CRT is a feature that most microwave design and test engineers require in their scalar analyzer systems. The HP 8757A supports this via data transfers to HP-IB plotters and printers over its system interface. The user first presses the **SYSTEM** key, which brings up the softkey menu defined in Fig. 3. The user can then select the desired format for the hard-copy output. By selecting the PLOT and DEFINE CUSTOM softkeys, the user can define the plot format, and then obtain hard-copy output in the customized format by pressing the softkey labeled PLOT CUSTOM.

Save/Recall

For some components, many different measurement setups and tests may be required. It is highly desirable for an instrument to be able to remember the test configurations so that they do not have to be reentered each time a new component is to be tested. The ability of the HP 8757A to store its state in nonvolatile memory allows the user to enter test configurations one time and then recall them as needed. The HP 8757A stores not only its state but also the calibration memories and limit lines, thus saving the user from doing a calibration or reentering the test limits each time a measurement configuration is recalled. In addition, when the HP 8757A is configured with an HP 8350 or HP 8340 Sweep Oscillator, the state of the source is also stored.

HP-IB

Remote operation of the HP 8757A Scalar Network Analyzer is by means of the Hewlett-Packard Interface Bus (HP-IB), Hewlett-Packard's implementation of IEEE Standard 488 and IEC 625-1. The HP 8757A accepts specific programming commands (listen mode) for selecting virtually all of the instrument's functions, including special HP-IB-only functions. The HP 8757A also outputs data (talk mode) from a designated channel (measurement, memory, or measurement minus memory) in a format specified by

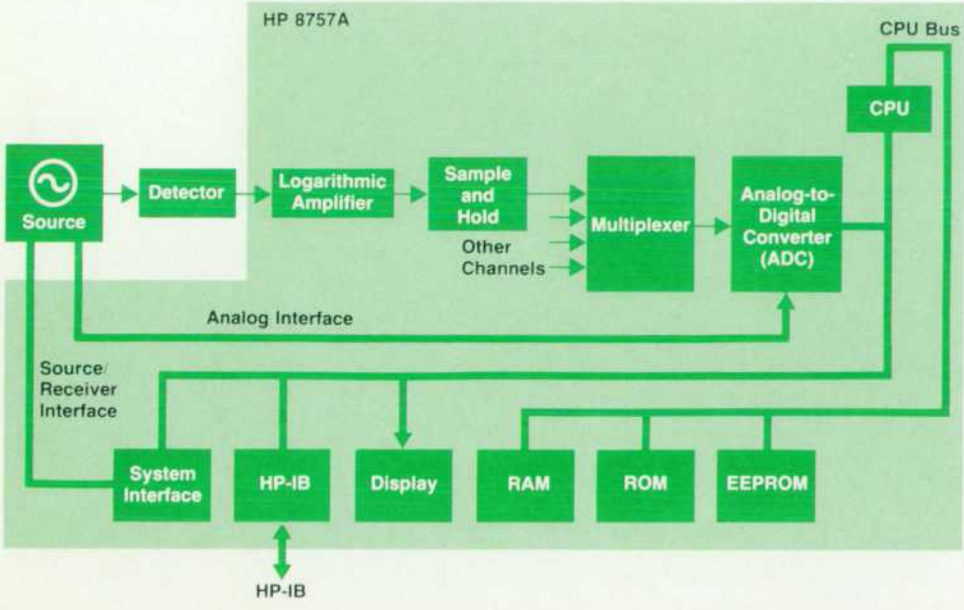

Fig. 6. *HP 8757A Scalar Network Analyzer functional block diagram.*

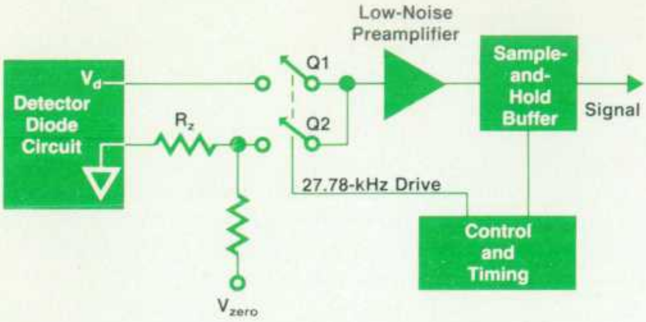

Fig. 7. *HP 85025 Series Detector block diagram.*

the user. Data can be transferred either as ASCII strings or as 16-bit integers. Readings can be taken of a single value (CW mode) or of an entire trace, and the cursor position and value can be output. The HP 8757A can pass commands through to any of the compatible instruments and/or peripherals connected to its system interface.

CRT Graphics

CRT graphics with the HP 8757A are enabled through the system interface. The graphics commands are mostly a subset of the Hewlett-Packard Graphics Language (HP-GL), allowing the HP 8757A to understand most HP desktop computer plotter output commands. Implementation of these commands required a translation of the HP-GL command desired into the corresponding function on the display, which is controlled by 16-bit commands.

This graphics capability allows the user to draw test connections or test procedures directly on the display. Instructions to the operator can be clearly spelled out and supplemented with connection diagrams that help eliminate any confusion (see Fig. 5, page 28). This eliminates the need for multiple copies of test procedures and manuals. Using the HP 8757A's graphics capability and softkey functions, the user can execute the test procedure by interacting with the display and softkeys instead of moving back and forth between the measurement system and the computer. This also allows control of multiple measurement stations with one central computer.

Hardware Design

Fig. 6 shows the functional blocks that make up the HP 8757A. The instrument itself is a wide-dynamic-range 27.78-kHz receiver and requires the use of at least one of a family of detectors to convert RF, microwave, and millimeter-wave frequencies to 27.78 kHz. The HP 85025 Detectors allow conversion with either modulated or unmodulated signals, while the HP 11664 Detectors require the RF to be modulated. The logarithmic amplifier conditions the 27.78-kHz signal to provide a dc signal to be digitized by the analog-to-digital converter (ADC). The ADC, which is synchronized with the sweeper via the sweep and blanking signals, digitizes the log amplifier output. Detector control on the ADC board determines the detector type, sensitivity, and temperature, sets the operating mode (ac or dc), and zeros the detector amplifier. An internal modulator drive provides an accurate 27.78-kHz signal to drive either external modulators or sweepers without that internal capability. The processor provides the user interface, performs the measurement corrections, controls the measurement hardware, communicates with external instruments via the two HP-IB ports, and updates the data on the display.

Detector

The HP 85025 Detectors provide a new capability to scalar analysis: detection of both modulated and unmodulated signals. An advantage of using a modulated (ac) system is rejection of spurious unmodulated signals and broadband noise. The ability to detect unmodulated (dc) RF and microwave signals is desirable where it is difficult to modulate the device being measured, such as in output power measurements of swept oscillators.[4] In addition, these detectors provide the following characteristics that are important for accurate scalar measurements:

- High return loss (low reflection coefficient)
- Low flatness variation with frequency
- Large dynamic range
- Low thermal drift.

The block diagram of the HP 85025 Detectors is shown in Fig. 7. The essential microwave component is the detector diode and its associated impedance-matching circuit. This circuit determines the critical RF performance characteristics of return loss and flatness and is a critical factor in determining the sensitivity or minimum detectable power level of the system. A microwave signal applied to the RF connector of the detector is down-converted to a dc voltage by the detector diode. The voltage depends on the magnitude of the RF power level applied. The output voltage from the diode is processed by a chopper preamplifier circuit, which provides the 27.78-kHz square-wave signal to the input of the HP 8757A.

High-accuracy measurements can be made with external diode detectors by using the HP 85025C Detector Adapter. Either modulated or unmodulated RF detection can be selected. The user needs to perform an external calibration sequence, which can only be enabled after the detector is connected and is recognized as an adapter by the HP 8757A. A coarse zero is the only manual adjustment necessary for a particular detector input. The normal zeroing process is then performed by the analyzer. The user, on selecting EXT DET CAL, is prompted for a high power level, which is set by the user and then entered in the HP 8757A **ENTRY** area. After a second prompt for a lower power level, the analyzer

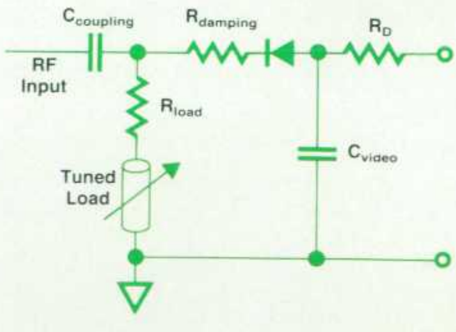

Fig. 8. *Microwave detector diode circuit of the HP 85025 Series Detectors.*

then calculates a new internal calibration table based on the two power level inputs. Through the use of this detector adapter, the range of the HP 8757A can be extended to over 110 GHz.

Detector Circuit Design

The microwave detector circuit (Fig. 8) is designed around a beam-lead zero-bias Schottky-barrier diode that has the necessary junction impedance (R_v) and RF voltage conversion efficiency to accommodate the signal processing circuitry, and has the low junction capacitance (C_j) and low series inductance (L_s) required for good broadband detection. A 0.25-mm-thick sapphire microstrip circuit that employs input resistor peaking and minimum diode circuit path inductance[5] is housed in a hermetic Kovar™ cylindrical package. The diode package has a dual output-pin glass-to-metal seal to minimize the thermoelectric effects of dissimilar metals in the detector signal path.

Careful part control and assembly technique are used so that no tuning of the circuit is required for RF match (return loss) from 10 MHz through 26.5 GHz. Small batch-to-batch variations in diode junction capacitance can affect the RF conversion efficiency across the band. These variations are compensated by the removal of discrete bonding wires that change the input resistor peaking transmission line impedance.

The output from the detector diode ranges from approximately 1.5 volts at the maximum input level of +16 dBm to approximately 4 microvolts at the specified −50-dBm sensitivity level in the dc mode. This voltage must be pro-

Calibrator Accessory

The HP 8757A log amplifiers must be calibrated over a 140-dB dynamic range for accurate power measurements (see box, page 29). For this purpose, the HP 11613A Calibrator (Fig. 1) will supply an accurate 27.78-kHz square-wave signal over a 154.5-dB range in 0.5-dB steps. This self-contained accessory interfaces to the HP 8757A with a single detector cable. Power, amplitude control, and the calibration signal are all transmitted over the detector cable. An HP 9000 Series 200 or 300 Computer connected to the HP 8757A HP-IB port controls the calibration process.

A 27.78-kHz crystal-controlled square-wave oscillator provides the maximum signal needed for calibration. Attenuation is achieved by a series of six resistive divider networks separated by high-input-impedance buffers and selected using analog multiplexers (see Fig. 2). By combining these six networks, any attenuation between 0 dB and 154.5 dB in 0.5-dB increments can be chosen. The signal then passes through a low-noise discrete output buffer and is sent to the HP 8757A log amplifier over the interface cable.

The control line from the HP 8757A supplies a serial stream of pulses to the HP 11613A control circuitry, which synchronously decodes the data into 16-bit words. These 16 bits are used to set the attenuator and to turn on the oscillator.

The HP 11613A must produce a relatively accurate signal attenuated over a 140-dB range to calibrate the HP 8757A. With a maximum output signal of about 10V p-p, this means that an accurate 1-μV p-p minimum signal is desired. There is always a 10V p-p signal present somewhere in the circuit, so inaccuracies caused by internal crosstalk can result from electromagnetic radiation, circulating ground currents, and power supply pickup. Careful circuit design and layout in addition to internal shielding avoid most of these problems, while the high common mode rejection ratio of the HP 8757A log amplifiers ignores the rest.

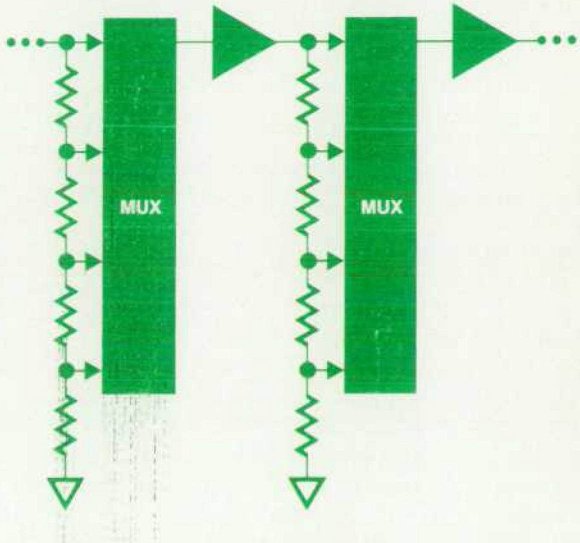

Fig. 2. *Two stages of the HP 11613A attenuator circuit.*

A complete system for calibrating the HP 8757A consists of an HP 11613A, a Series 200 or 300 Computer, and the calibrator software. The HP 11613A is plugged into the detector port being calibrated and the computer is connected to the HP 8757A via the HP-IB. The software sets the HP 11613A to a series of output levels, reads data from the HP 8757A, creates a calibration table, and stores the table in the HP 8757A's EEPROM, provided that two test points are jumpered. The accuracy of the existing EEPROM tables can also be checked using this system.

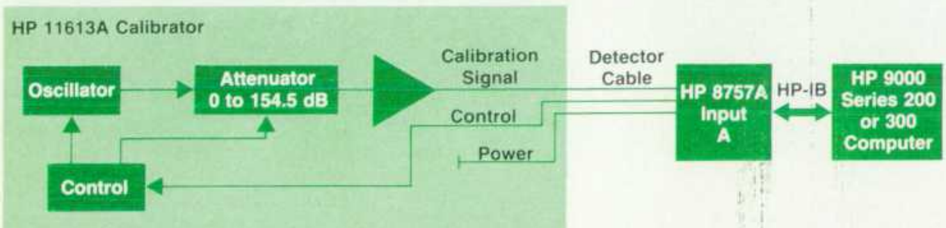

Fig. 1. *Setup for calibrating HP 8757A input* **A**.

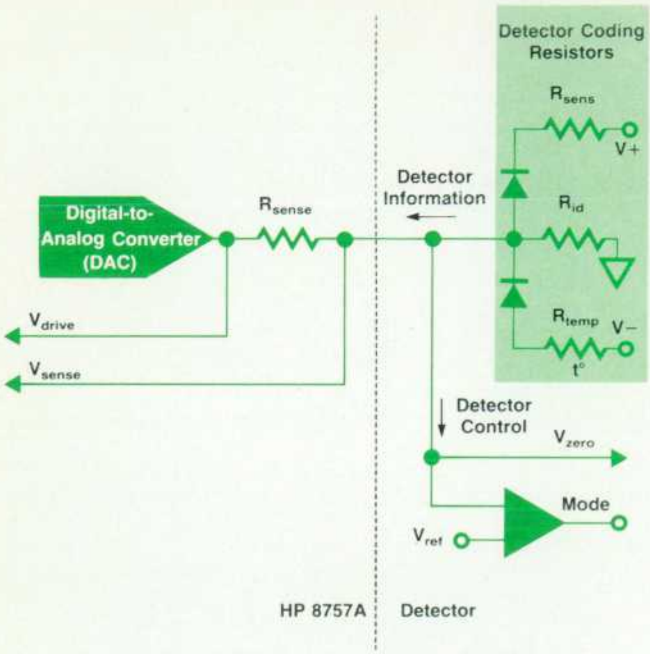

Fig. 9. *The detector control line (one per channel) is used to control the detector functions and send information to the HP 8757A.*

cessed with an error of less than ±0.3 microvolts for accurate power measurements at low levels in dc mode.

Signal Processing

In either mode of operation, ac or dc, the log amplifiers process a 27.78-kHz square-wave signal. In ac mode, the modulation is recovered by the detector and sent directly to the HP 8757A. In dc mode, the dc voltage from the detector diode must be converted into a 27.78-kHz square wave for the HP 8757A to process. The diode signal is chopped into a square wave by series-shunt chopper field-effect transistors Q1 and Q2 (Fig. 7). The chopper drive is generated by a crystal-controlled oscillator. The resulting square wave is amplified by the low-noise preamplifier stage. A side effect of the chopping process is the injection of the drive waveform into the signal path. At low signal levels, this 27.78-kHz component is larger than the actual signal from the detector. If not removed, this would degrade the sensitivity of the detector because the chopped signal would be masked by the chopper feedthrough transients. The chopper transients at the output of the preamplifier stage settle in a short time compared with the period of the 27.78-kHz square wave. The transients are removed by the process of synchronously sampling the signal at a point in the 27.78-kHz waveform where the feedthrough at the preamplifier output has settled to a negligible level. The sampling process improves the sensitivity by approximately 20 dB.

Detector Control and Accuracy Enhancement

Any small dc offset voltages in the signal path from the detector diode through the chopper stage must be compensated by a nulling process. The offsets are primarily caused by thermoelectric effects. The diode itself can generate a dc offset voltage if there is a thermal gradient across it. Thermal gradients across dissimilar-metal junctions between the detector diode and the chopper FETs will also introduce errors. These offsets are nulled by introducing a compensating voltage across R_z (Fig. 7).

With the RF power turned off, an automatic process in the HP 8757A adjusts the detector interface DAC (see Fig. 9) to null the offset and maintain good power level accuracy. A detector connected to any input port of the HP 8757A is controlled by that port's digital-to-analog converter, which drives a control line at each detector input connector (Fig. 9). The control line is used to control the detector's functions and to communicate information back to the HP 8757A. The functions controlled in the detector are the mode (ac or dc detection) and dc zeroing. The control voltage is applied to the input of a comparator in the detector which converts this analog signal to a logic level which is applied to the clock oscillator to enable either ac or dc operation.

The information the HP 8757A can read from the detector consists of the detector's type code, its sensitivity, and its temperature. By placing voltages on the control line and sensing the current in R_{sens}, an HP 8757A algorithm can measure the individual resistances R_{id}, R_{sens}, and R_{temp}. R_{sens} is a factory-set value that establishes the correction required for the individual detector diode's RF sensitivity, and R_{temp} is a precision thermistor used to modify the correction constants in the HP 8757A to correct for temperature effects in the detector. Firmware in the HP 8757A accounts for switching-diode and parallel-resistance effects in this circuit, then automatically converts the resistor readings into parameters used by the microprocessor to generate the detector correction curves.

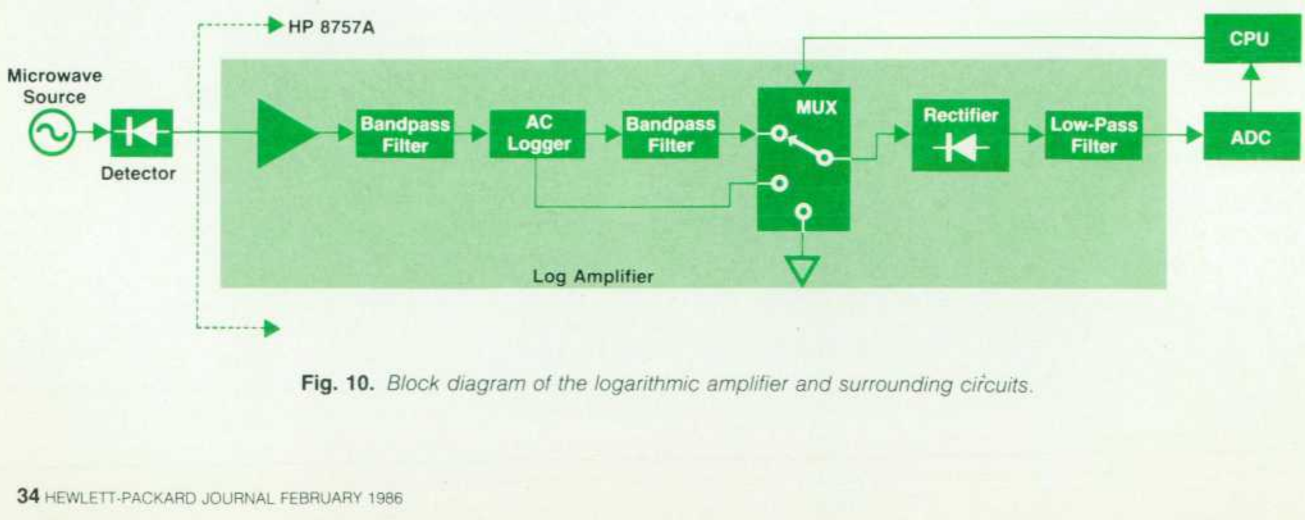

Fig. 10. *Block diagram of the logarithmic amplifier and surrounding circuits.*

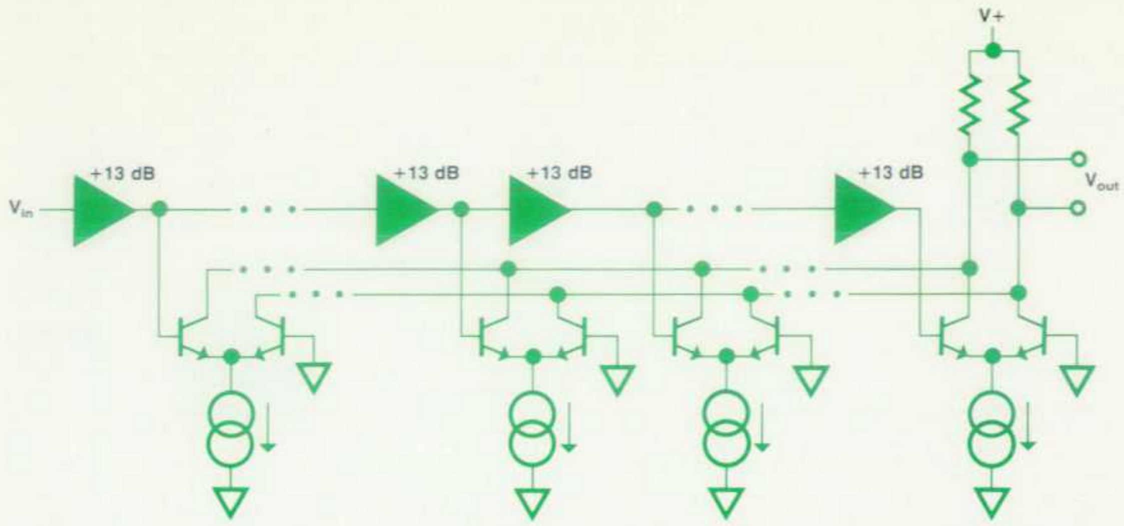

Fig. 11. *Simplified diagram of the ac logger.*

Logarithmic Amplifier

Fig. 10 shows a block diagram of the log amplifier and the surrounding circuits. The detector provides the log amplifier with a 27.78-kHz square-wave signal proportional to the detector's input power. The log amplifier produces a dc output voltage proportional to the logarithm of the 27.78-kHz input signal over a 140-dB dynamic range (equivalent to about 80 dB in detected power).[1] This amplifier is similar to that in the HP 8755A and 8756A Analyzers, although component and circuit changes have been introduced to increase the dynamic range more than 16 dB in detected power and to improve overall accuracy and drift.

The detector output is fed directly into a differential amplifier which provides impedance matching and high common mode rejection. Next, a 2-kHz-wide bandpass filter converts the square wave into a sine wave and reduces out-of-band noise components. The log amplifier's improved dynamic range compared to the HP 8755A and HP 8756A versions is a result of this narrow filter reducing the noise floor. Transient response of the circuit depends primarily on the bandpass filter's response.

After filtering, the signal is compressed by the ac logger, which produces an ac output proportional to the logarithm of the input signal magnitude. Fig. 11 is a simplified diagram of the ac logger. The logger consists of twelve 13-dB amplifier stages connected in series, with each amplifier driving one of twelve transistor pairs. Each pair acts like a nonlinear current switch and its output current is summed with those of the other transistor pairs through the output load resistor. For each transistor pair, a large input will cause the output to saturate at the bias current, while for a small input the pair is nearly off. Since the amplifiers are connected in series, whenever the ac logger input increases by 13 dB, one more transistor pair saturates, causing the output signal to increase by the bias current of one transistor pair. This establishes an approximately logarithmic relationship between the input signal and the output

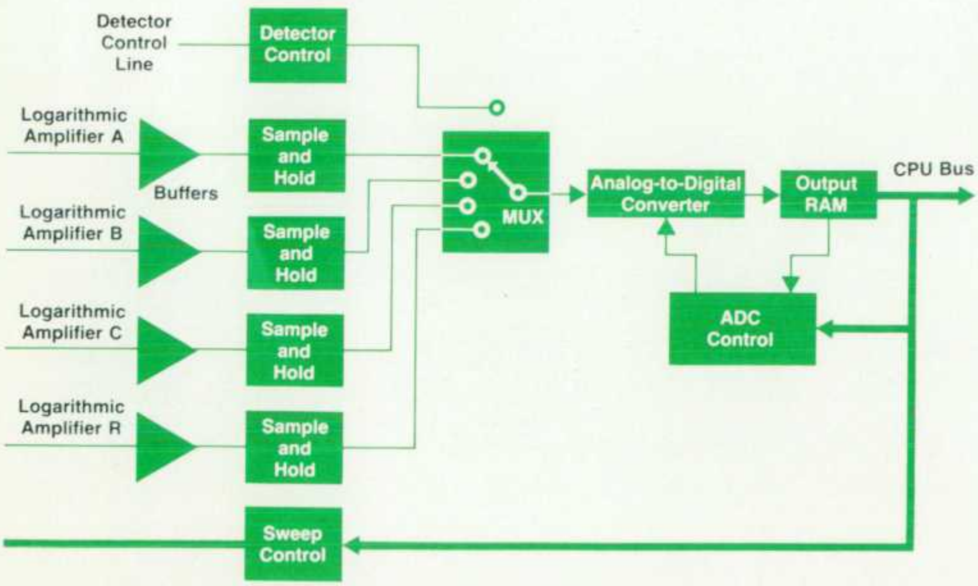

Fig. 12. *Analog-to-digital converter (ADC) system block diagram.*

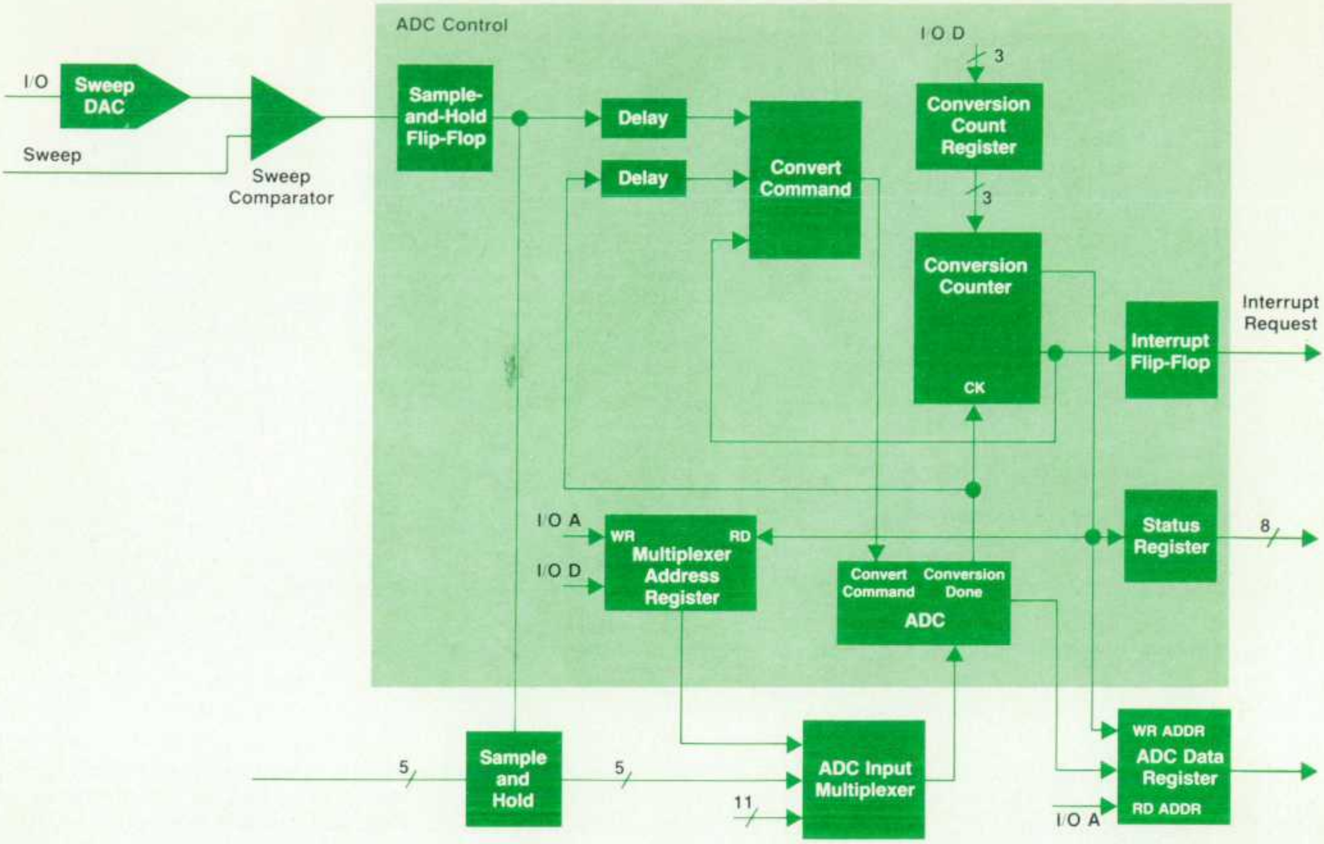

Fig. 13. *ADC system schematic diagram. A simple state machine reduces interaction with the main processor.*

current. The basic HP 8757A ac logger operation is the same as in the HP 8755A and HP 8756A Analyzers, the major differences being an increase in amplifier spacing from 10 dB to 13 dB for increased dynamic range and closed-loop control of the transistor pair bias currents for lower temperature sensitivity.

The ac logger is followed by a bandpass filter which reduces the signal noise bandwidth and shapes the nearly square logger output into a sinusoid for the rectifier stage.

Next, the signal passes through a multiplexer controlled by the CPU. From this point on, the signal is dc coupled. The multiplexer selects either the logged ac signal, a dc voltage from the temperature sensor in the ac logger for temperature drift compensation, or the offset of the dc circuitry.

The dc section consists of an active full-wave rectifier with low first-harmonic feedthrough, followed by a low-pass filter. The filter removes all noticeable ripple from the rectifier output without significantly degrading the log amplifier's overall transient response. The resulting dc signal is fed to the ADC board for digitizing.

ADC

The block diagram, Fig. 12, shows the overall functions of the ADC system, which are:

- Sample and hold
- Analog multiplexer
- ADC
- ADC control
- Sweep control.

The main functions of the ADC circuitry are to digitize the analog outputs of the log amplifiers, to provide the interface to the sweep oscillator's sweep voltage, blanking, and stop sweep signals to allow synchronization of the data-taking process, and to provide an interface to the external detectors. The circuitry that digitizes the analog signal uses a monolithic sample-and-hold circuit, a 16-input multiplexer, and a 16-bit hybrid ADC which is short cycled to 15 bits and converts in 54 μs. A simple state machine (Fig. 13) is included to reduce interaction with the main

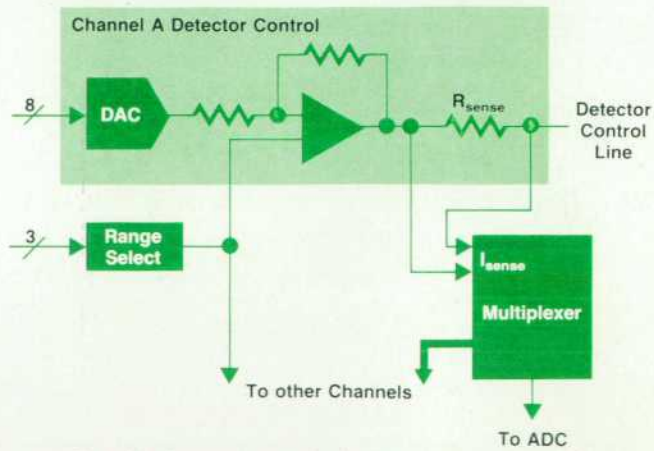

Fig. 14. *Detector control circuitry lets the HP 8757A determine the type of detector.*

Voltage-Controlled Device Measurements

The HP 8757A Scalar Network Analyzer incorporates a number of features that make the measurement of voltage-controlled devices easier and more accurate. These devices include voltage-controlled or voltage-tuned oscillators (VCOs or VTOs) and attenuators (VCAs).

One helpful feature of the HP 8757A is its ability to accept a sweep input that does not conform to the standard sweep oscillator output of 0 to 10 volts (see "Sweep Modes," page 30).

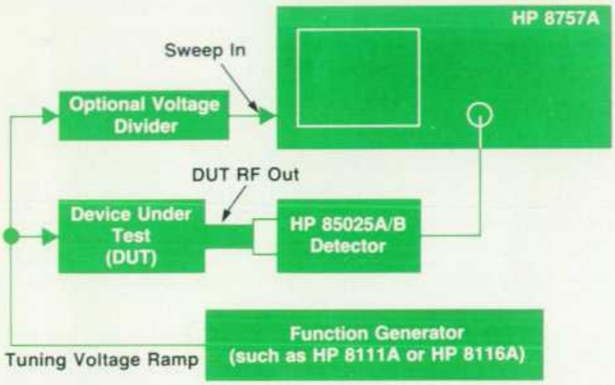

Fig. 1. *HP 8757A configured for voltage-controlled device measurements.*

The HP 8757A can measure RF power by detecting either modulated RF (ac detection) or unmodulated RF (dc detection) using the HP 85025A or HP 85025B Detectors. The choice of ac detection offers the best dynamic measurement range and high immunity to noise and temperature offsets. Since a voltage-controlled oscillator may be difficult to modulate with the required 27.78-kHz square wave for ac detection, the user can easily switch to dc detection by means of a softkey under the **SYSTEM** menu. A generalized measurement setup for making absolute power or normalized measurements is shown in Fig. 1.

Another useful feature is the HP 8757A's limit testing capability. After entering a limit line, the user can put that limit specification (either the upper or the lower limit) into the active channel's memory. This is particularly important in a voltage-controlled attenuator measurement, where the linearity of the attenuator's attenuation-vs-voltage characteristic is a critical specification. The user first enters on the HP 8757A the start and stop labels for the left and right graticule lines, corresponding to the start and stop voltages from the function generator (see Fig. 1). Then the user enters a sloped limit line over the voltage range of interest, representing the attenuator's ideal linear characteristic, and puts this limit into memory. When the measurement minus memory display is selected, the display will show the deviation from the ideal linear sloped characteristic.

Figs. 2 and 3 show two plots of a voltage-controlled attenuator measurement. In Fig. 2, the attenuation characteristic is measured relative to the control voltage, which is entered on the plot as 2 to 7 volts. Although the attenuator appears to be fairly linear, a high-resolution (0.5 dB/division) plot of the deviation from the sloped limit line clearly shows the nonlinearity (see Fig. 3).

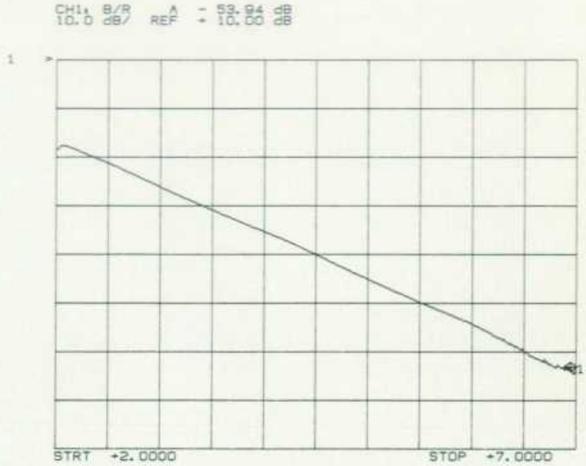

Fig. 2. *Voltage-controlled attenuator measurement of attenuation as a function of control voltage. This measurement was performed at 3 GHz. The frequency response error was removed by entering detector offsets.*

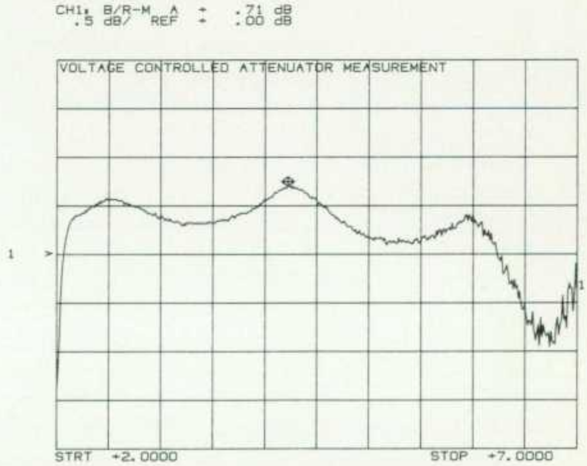

Fig. 3. *Voltage-controlled attenuator linearity measurement. This is the same measurement as in Fig. 2, except that here the linear slope characteristic was removed by placing a sloped limit line into memory and selecting the measurement minus memory function.*

processor. In setting up the instrument state, the CPU sets the multiplexer select and conversion count registers to the states required to make the desired measurement(s). Conversions are enabled by writing to the sweep DAC the value of the sweep voltage for the next data point. This loads the conversion counter with the value of the conversion count register, which allows the sample-and-hold flip-flop to be set by the output of the sweep comparator. This initiates a process that selects the desired input, waits for settling, triggers a conversion, and stores the data in an output register. This is repeated until the programmed number of conversions have been made. An interrupt is then generated for the CPU to read the data and set the next sweep DAC value.

The detector control circuitry (Fig. 14) lets the HP 8757A determine the type of detector connected. If the detector is from the HP 85025 Series, a number of additional capabilities are provided to enhance the measurement accuracy. The circuitry for each channel consists of one quarter of a monolithic quad 8-bit voltage-output DAC with built-in latches, a summing amplifier which combines the DAC output and one of three range voltages, and a current sense resistor to monitor the output current. The voltage on either side of the sense resistor can be measured by the ADC. The three output bias ranges allow different impedances to be biased on within the detector. Within a range, two different bias voltages are applied to the detector and the resulting currents measured. The effective load impedance can then be computed. In the case of the detector identity and sensitivity resistors discussed earlier, these values are used as an entry to a lookup table to determine the detector ID, capabilities, input power response constants, and sensitivity.

Range 2 allows measurement of the detector's ID resistor, which always remains in the circuit, and in the case of the HP 85025 Detector, selects the dc mode and provides the autozero feedback to null the offset voltage of the chopper amplifier. For the HP 85025, range 1 biases the detector sensitivity resistor and range 3 enables the ac mode and biases on the detector's temperature sensing thermistor.

CPU

The digital processing capabilities of the HP 8757A are provided by an MC68000 microprocessor operating at 10 MHz. I/O is memory mapped and access timing is controlled by the addressed memory block's selecting the appropriate output of the digital wait state generator for a given memory block, except in the case of the display, which requires handshaking because of its variable access time. In this case, a maximum of 11 wait states are allowed before the default acknowledge occurs. The memory is 96K words of ROM and 16K words of static RAM. The RAM, which has two-week battery backup, stores the current instrument state, nine instrument setups, and four sets of measurement calibrations. There are also 2K words of EEPROM to store the instrument calibration that compensates for the log amplifiers, sample-and-hold circuits, multiplexer amplifier, and ADC.

An RC timer generates periodic interrupts to service user requests, perform calibrations, and if selected, perform a dc detector autozero. Status LEDs are used to indicate failures that occur during the instrument verification and cannot be displayed on the CRT, such as ROM, RAM, I/O bus, CRT interface, and interrupt failures. Also, test points are included that are scanned immediately after a reset; these can be used to direct the processor to execute a tight loop to exercise a specific group of digital lines for troubleshooting purposes.

A removable jumper causes the microprocessor to execute a MOVEQ instruction continuously. This allows a signature analyzer to verify the address and data lines and the ROM contents.

System Interface

The **SYSTEM INTERFACE** port provides HP-IB control of a compatible plotter, printer, and sweep oscillator. Hard-copy output is initiated by making the desired selections under the **SYSTEM** menu, as explained earlier.

This interface provides the user with important sweep oscillator information on the CRT display of the HP 8757A, yet allows independent control of the sweep oscillator. With a compatible HP-IB sweep oscillator (HP 8350B, HP 8340A, or HP 8341A), the system interface provides for combined **SAVE/RECALL** of both the sweep oscillator and the analyzer instrument states. Annotation appears on the CRT display corresponding to the start and stop frequencies. The active marker frequency and the amplitude measurement are also displayed if enabled (and the **CURSOR** function disabled). The alternate sweep function of the sweep oscillator (allowing two different frequency ranges and/or power levels to be swept alternately) is only possible with the system interface connected.

Acknowledgments

As is the case with any large project, the list of people who contributed to the success of the HP 8757A Scalar Analyzer System is quite long. Dick Barg was the product designer for the HP 8757A. He and Jim Minor did the product design of the HP 85025 Detectors and the HP 85027 Directional Bridges. Dennis McCarthy designed the HP 85025C Detector and developed its calibration method. Fred Rawson was responsible for the design of the power supply. Judi Cowell defined and wrote the diagnostics and self-tests. Julie Chapman contributed the firmware that enables hard-copy representation of the CRT display on a printer. Dave Blackham wrote the original production test software for the HP 85025 Detectors and HP 85027 Bridges, and helped with their transition into production. Chuck Compton, Frank Hamlin, Mike Heinzelman, and Jim Young all contributed to the transition of the HP 8757A into production. Roy Church was responsible for the industrial design of the HP 8757A Scalar Analyzer, the HP 85025 Detectors, and the HP 85027 Directional Bridges. The HP 8757A Scalar Analyzer System was developed and introduced into production with the support of R&D manager Irv Hawley and section manager Hugo Vifian. Many people at other HP Divisions have contributed. We thank Loveland Technology Center and Santa Clara Technology Center for the log amplifier integrated circuits, Colorado Springs Technology Center for the HP 85025 Series Detectors' thick-film hybrid circuits, and Microwave Technology Division for the thin-film circuit. Thanks also to Colorado Springs Division for the development of the large-screen graphics display. We would sincerely like to thank all of those mentioned above, plus many others we were unable to mention here. Without their many contributions and support, the successful development and introduction of this product would not have been possible.

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Authors

February 1986

4 — Microwave Counters

Scott R. Gibson

A development engineer at HP's Santa Clara Division, Scott Gibson has been with HP since 1980. He worked on a number of instrumentation products in his first assignments in production engineering and later, as an R&D engineer, contributed to the design of the microwave module for the HP 5350A and HP 5351A Microwave Counters. He was also a project leader for the HP 5352A Microwave Counter. Scott was born in Arcadia, California, received his BS degree in engineering from Harvey Mudd College in 1980, and is currently working toward an MSEE degree at the University of Santa Clara. He is the author of an article that recently appeared in *Microwaves & RF*. A resident of Santa Clara, California, he likes hiking, running, photography, and bicycling. He recently returned from an extended trip to western Canada and boasts that he has *no* home computer.

11 — IF and LO Parameters

Luiz Peregrino

Lou Peregrino received the degree of Engenheiro de Eletronica in 1959 from the Instituto Tecnologico de Aeronautica in Brazil. He also did graduate-level work at Stanford University. After joining HP in 1960 he worked on a variety of instruments, including systems for s-parameter measurement, synthesizers, and sweepers. He has also developed an architecture for microwave counters and a random modulation technique for harmonic number determination. Most recently he developed measurement algorithms for the HP 5350 family of microwave counters. Now a resident of Cupertino, California, Lou enjoys skiing, sailing, and windsurfing.

15 — Multimeter

Thomas G. Rodine

Tom Rodine has worked on circuit design and mechanical design for a number of HP products since joining HP in 1966, as well as serving as production engineer for two of the products. He has contributed to the HP 3571A Spectrum Analyzer, the HP 3586A Selective Level Meter, the HP 3488A Control Unit, and the HP 3457A Digital Multimeter, among other products. He is named inventor on a patent on a method for preventing phase reading errors and is a licensed professional engineer in the state of Colorado. He was born in Des Moines, Iowa and attended Iowa State University (BSEE 1966) and Colorado State University (MSEE 1969). Tom lives in Loveland, Colorado, is married, and has three teen-age children. He is active in his church and enjoys family activities, hunting, and fishing.

Ronald K. Tuttle

Born in Altadena, California, Ron Tuttle attended the University of California at Berkeley and received a BS degree in chemical engineering in 1959 and an MSEE degree in 1961. After joining HP in 1961 he contributed to the development of the HP 3406A Voltmeter. Later he was project manager or coproject manager for a number of products, including the HP 204A Oscillator, the HP 3320A Synthesizer, the HP 3478A Multimeter, the HP 3585A Spectrum Analyzer, and the HP 3457A Multimeter. He is presently a production engineering manager. Ron lives in Loveland, Colorado, is married, and has three children. When not working on his mountain cabin, he enjoys backpacking and cross-country skiing.

Douglas W. Olsen

Doug Olsen was born in Portland, Oregon and is a 1979 graduate of Oregon State University (BSME). With HP since 1979, he is an R&D engineer at the Loveland Instrument Division and is currently working on a cost reduction project in production engineering. He has contributed to the development of the HP 3054A Data Acquisition/Control System, the HP 3478A Multimeter, and the HP 3457A Digital Multimeter. He is the coauthor of an article on the HP 3457A published in a trade magazine and is a registered professional engineer in the state of Colorado. Doug's outside interests include hiking, skiing, bicycling, and auto mechanics.

Joseph E. Mueller

Joe Mueller has designed hardware and firmware for HP voltmeters since joining the company in 1979 and most recently has worked on HP-IB peripheral drivers. His work on the RAM protection scheme for the HP 3457A Multimeter is the subject of a pending patent. Born in Champaign, Illinois, Joe was awarded a BSEE degree from the University of Illinois in 1979. He and his wife live in Loveland, Colorado, are the parents of one child, and are expecting an addition to their family soon. He takes an active part in his church and is interested in working with wood and leather, reading science fiction, and building and flying radio-controlled gliders.

Scott D. Stever

Scott Stever was born in Corpus Christi, Texas and studied electrical engineering at the Georgia Institute of Technology. He came to HP in 1979, the same year he completed work for his BSEE degree. He investigated precision ac measurements before working on the dc sections and reference for the HP 3457A Digital Multimeter. He is named coinventor on a patent application related to the ac autocalibration technique used in the HP 3457A. Scott, his wife, and their son are residents of Loveland, Colorado. During leisure hours he enjoys flying, skiing, tennis, and reading.

24 Scalar Analyzer System

Douglas C. Bender

Doug Bender has worked on several network analyzers since joining HP in 1980. He was a product marketing engineer for the HP 8756A and the HP 8510A Network Analyzers and was a software development engineer for the HP 8757A Network Analyzer. Born in Lincoln, Nebraska, he attended the U.S. Military Academy and completed work for his BS degree in 1973. He served in the U.S. Army as an electronics communications officer and continued his studies at Stanford University, receiving an MSEE degree in 1980. Doug and his wife live in Santa Rosa, California. His outside interests include bicycling and amateur radio (WA0JRB).

Jacob H. Egbert

A graduate of the University of Oklahoma, Jake Egbert completed course work for a BSEE degree in 1969 and for an MSEE degree in 1970. He also served in the U.S. Air Force and designed bus systems for another computer firm before joining HP in 1971. He was a design engineer for the HP 8500 Graphics System, the HP 8542 Network Analyzer, and the HP 8501A Storage-Normalizer. He has also managed the development of a number of network analyzer products, including the HP 8757A Network Analyzer, the HP 8756A Network Analyzer, and the HP 85025 AC/DC Detectors. He is currently an R&D section manager for the HP 70000 family of modular spectrum analyzers. Jake was born in Toledo, Ohio and now lives in Santa Rosa, California with his wife and three children. He coaches youth soccer and softball, plays soccer himself, and likes golfing, fly-fishing, and skiing.

Kenneth A. Richter

Born in Brooklyn, New York, Kenneth Richter studied electrical engineering at Drexel University, completing work for his BSEE degree in 1967. After working as a microwave design engineer in the aerospace industry, he came to HP in 1973. He has worked on the HP 85020A/B RF Bridges and was a project leader for the HP 85025A/B/C Detectors and the HP 85027A/B Bridges. He is named inventor on a patent related to an ac/dc integrated detector and is the author of a number of technical papers. Kenneth, his wife, and three sons live in Santa Rosa, California. He coaches youth soccer and basketball and enjoys a change of pace from work by racing a car that he built himself.

Frederic W. Woodhull, II

Fred Woodhull studied electrical engineering at Oregon State University before joining HP in 1968. He has been a service technician and production line leader and worked as a design engineer on the HP 8505A and the HP 8754A Network Analyzers. He was the project leader for the HP 8757A. He is named coinventor on a patent related to a method for measuring the frequency of a sweeping signal. Fred was born in Ross, California and served in the U.S. Army. He now lives in Santa Rosa, California with his son and likes waterskiing, camping, gardening, and working with wood and metal.

John C. Faick

Born in Silver City, New Mexico, John Faick studied electrical engineering at the University of Arizona and received his BSEE degree in 1972 and his MSEE degree in 1974. He has been with HP since 1974 and has contributed to the development of various scalar network analyzers, including the HP 8757A. He has also worked on the HP 8568A and HP 8565A Spectrum Analyzers. He is named coinventor on a patent application related to the HP 85025A AC/DC Detector and is interested in digital and analog signal processing. John is married, has a daughter, and lives in Santa Rosa, California. He's putting the finishing touches on a house he and his wife built and spends the rest of his free time on home electronics projects.

Keith F. Anderson

With HP since 1981, Keith Anderson contributed to the development of the HP 11613A Calibrator and the HP 8757A Network Analyzer. He is interested in analog circuit design and is a California registered professional engineer and a member of the IEEE. Born in Seattle, Washington, Keith received a BSEE degree from the University of Washington in 1980 and an MSEE degree from Stanford University in 1981. He lives in Santa Rosa, California and likes bicycling, skiing, and hiking.

Joseph Rowell, Jr.

With HP since 1973, Joseph Rowell has contributed to the design of a number of analyzer products. He did the analog/digital design for the display sections of the HP 8568A and HP 8566A Spectrum Analyzers, contributed to the digital design and firmware for the HP 8756A Network Analyzer, and wrote firmware for the HP 8757A. A patent has resulted from his work on a video processor for a spectrum analyzer. Born in Tuskegee, Alabama, Joseph received a BSEE degree from the University of California at Berkeley in 1973 and an MSEE degree from Stanford University in 1976. He lives in Santa Rosa, California and supplements his HP income by investing in real estate. He enjoys collecting and working on motorcycles and likes backpacking and racquetball.

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