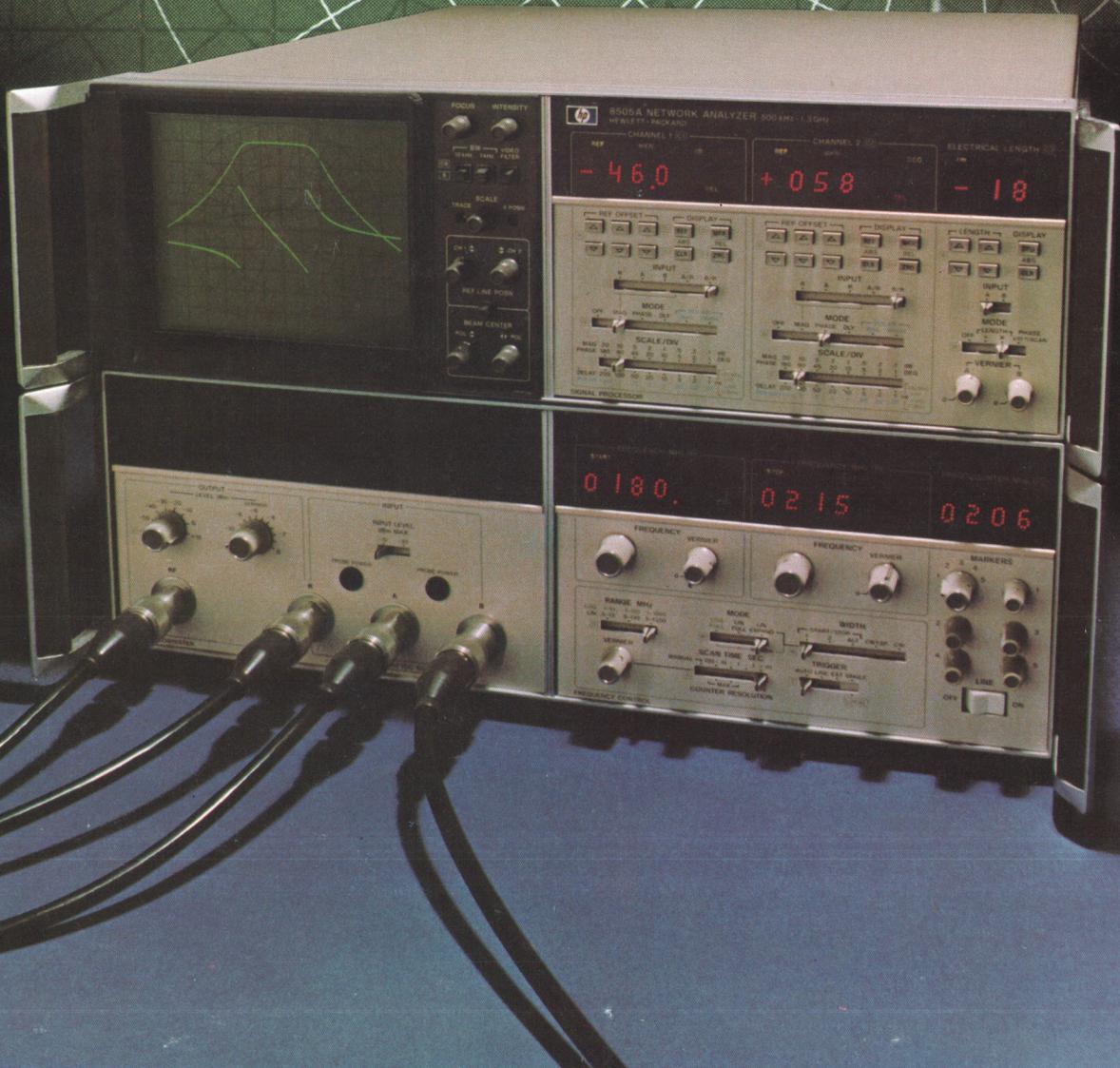
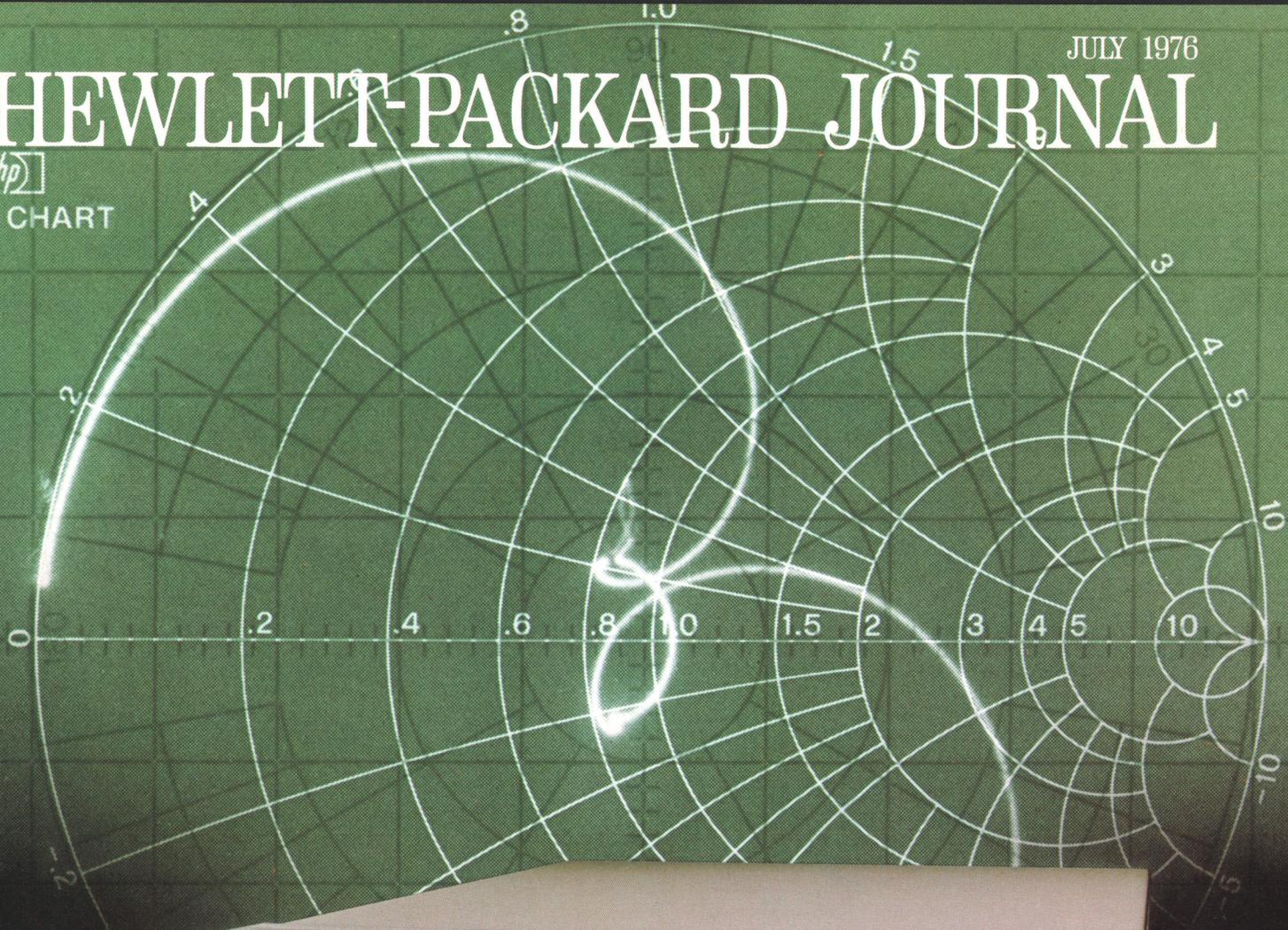


HEWLETT-PACKARD JOURNAL



SMITH CHART



A Direct-Reading Network Analyzer for the 500-kHz-to-1.3-GHz Frequency Range

Besides measurements of gain, loss, phase shift and reflection coefficient, this new all-in-one measuring set makes direct measurements of equivalent electrical length, group delay, and deviation from linear phase shift.

by Hugo Vifian

CHARACTERIZING CIRCUITS in the frequency domain is a fundamental activity in both the design and the testing of electronic circuits. Only by knowing the gain/phase-vs-frequency characteristics and the input-output impedances of each circuit can one assemble a complete device capable of meeting performance objectives.

Over the past ten years there has been a dramatic growth in the use of network analyzers to characterize component and circuit performance. By providing plots of gain or loss, phase shift, and reflection coefficient versus frequency, these instruments have given electronic engineers practical insight into circuit and component behavior, leading to more precise designs. The more precise designs have led to tighter system performance specifications—for example, closer packing of communications channels—and this in turn has led to a demand for even better measuring instruments.

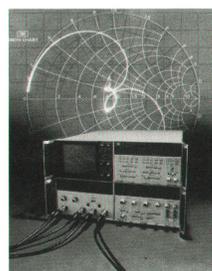
Such a new instrument is the Model 8505A Network Analyzer (Fig. 1). Intended for use in the RF, VHF and lower UHF frequency ranges, it makes traditional measurements of driving point and transfer characteristics with a greater frequency range (500 kHz—1.3 GHz), greater dynamic range (100 dB), better resolution and higher accuracy than its predecessors, and it does this with considerably more convenience. It also makes direct measurements of equivalent electrical line length, of deviations from linear phase and of group delay with much greater facility than has hitherto been possible. With these capabilities, the new network analyzer makes it possible to characterize a network quickly with a minimum of ambiguity.

All-in-One Test System

The new network analyzer is a complete measuring system that includes a precision swept-frequency

source, a three-input tracking receiver, a CRT display, a frequency counter, and an autoranging voltmeter that gives direct digital readout of gain/loss, phase, deviation from linear phase, and group delay. An internal digital processor controls the instrument's operation, simplifying the way the instrument operates and enhancing the clarity of the data presentation.

The three inputs, all of which have 100-dB dynamic range, enable simultaneous display of both the transmission and reflection coefficients of a device or



Cover: The subject of this month's issue, Model 8505A Network Analyzer presents swept-frequency measurements in either Cartesian or polar coordinates on its CRT display and its digital displays give the frequency, magnitude, phase, and delay of any point selected on the CRT trace by a movable marker.

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Fig. 1. Model 8505A Network Analyzer is shown here measuring the transmission and reflection coefficients of a 50Ω filter using the Model 8502A Transmission/Reflection Test Set. With its digital readouts, in addition to a dual-mode CRT display, this new instrument can characterize circuits and components in the 0.5 - 1300-MHz frequency range more quickly, more accurately, and with greater convenience and less ambiguity than its predecessors.

its gain and phase responses. The display can be either in rectangular or polar coordinates (Fig. 2) and both a rectangular and a polar plot may be displayed at the same time. Up to five markers can be placed anywhere on the display and the frequency and amplitude or phase of the signal at any point designated by a marker can be displayed digitally (Fig. 3), giving a measurement resolution of 0.01 dB in amplitude, 0.1° in phase, and 100 Hz in frequency, an order of magnitude better resolution than that obtainable from a CRT display alone.

Electronic Delay Line

A continuously-variable delay line may be inserted in the measurement channel to match the electrical length of either the A or B input to the reference. It also enables the measurement of a device's equivalent

electrical length by substitution. The delay line, a new development, is simulated electronically, giving calibrated, finger-tip control of line length up to the equivalent of ±1500 meters (it can also add negative delay), a much greater length than would be possible with a mechanical line stretcher. It has the added advantages of not contributing mismatch errors or attenuation.

Because the line has a constant time-delay-versus-frequency characteristic at any one delay setting, the phase shift through the line varies linearly with frequency as it does through a coaxial cable. During measurements of phase shift, the line may be inserted to subtract a phase shift equivalent to a device's electrical length so the display will show the deviations from linear phase shift caused by the device as a function of frequency (Fig. 4). Interest in this measure-

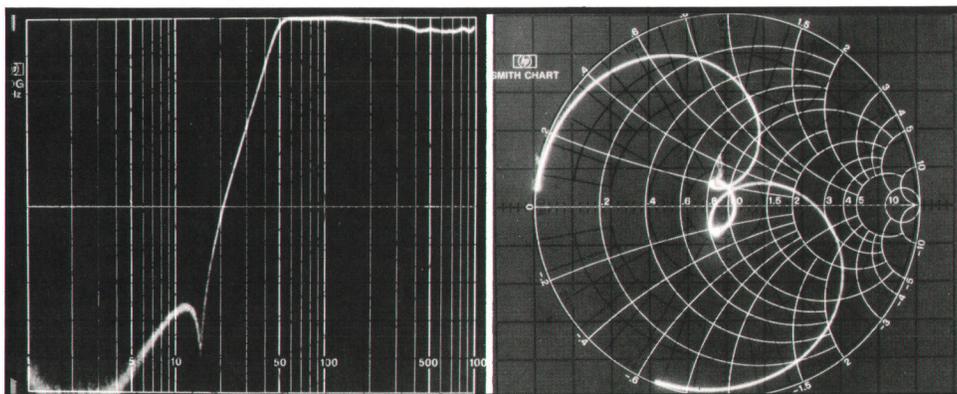


Fig. 2. Swept-frequency measurements may be displayed in either cartesian or polar coordinates. Transparent Smith-chart overlays, as shown in the photo of a filter input impedance at right, are provided for direct readout of impedance with the polar displays. Overlays with logarithmic scales are also provided for logarithmic sweeps, as in the photo at left showing the response of a 65-MHz high-pass filter over a 100-dB dynamic range, swept from 1 to 1000 MHz.

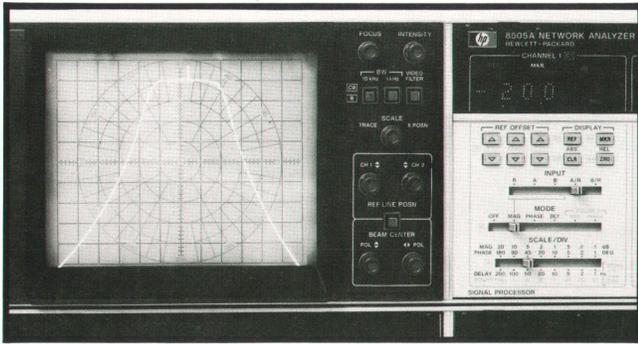


Fig. 3. Digital readouts give three-digit resolution measurements of frequency, gain/loss, phase, or delay at the point in the response curve indicated by the positive-going marker.

ment has been growing in recent years because of the necessity of preserving phase relationships in multiplexed signals and preserving the shape of complex waveforms.

The instrument can also directly display group delay, a plot whose magnitude corresponds to the slope of the phase response curve (Fig. 4). A new technique for measuring group delay directly, described on page 11, makes it possible to resolve variations in delay with a resolution of 1 ns/div.

Automated Operation

Designed to work with the HP Interface Bus (HP's implementation of IEEE Standard 488-1975), the new network analyzer can work in an automatic system for setting up measurements, reading data, and correcting for errors. When equipped with the HP-IB option, the new analyzer can be incorporated into an automatic system simply by connecting it to a controller with the appropriate cables.

In most cases, the controller will be a calculator. Considerable attention was paid to defining programming codes that would make it easy for a user who is not experienced in programming to control

the instrument remotely. Each front-panel control is labelled with the address code used by a calculator to set that particular control, so there is no need to memorize codes. For example, when programming the instrument to sweep its full frequency range, the MODE switch is set to the LIN FULL position. The program code for that setting is M2, where M stands for MODE, as indicated on the front panel, and 2 stands for the second position of the switch.

The instrument, however, provides an even easier way to program control settings, a result of the instrument's digital architecture. This is the "Learn" mode. After the appropriate part of the instrument is addressed (source or receiver), the single keystroke command "L" on the controller/calculator will cause the instrument to "read" the front-panel switch settings and make the readings available to the interface bus. Without bothering with control program codes, the user may set up test sequences on the front-panel controls, store them (e.g. on tape), and then recall them when needed.

Factory-assembled automatic network analyzers that include the Model 8505A, a 9830A Calculator with printer, an s-parameter test set, and appropriate cables and fixtures are available as Model 8507A (Fig. 5).

Accuracy Enhancement

One of the most important applications of calculator- or computer-controlled network analyzers is the removal of system errors from the measurement data. These errors arise from non-ideal test fixtures, mismatches, and other instrumentation problems. For example, the input impedance of a device in an environmental chamber is to be measured at the end of a long cable. As shown in Fig. 6, an imperfect cable severely influences measurement results, leading to results that are meaningless. An accuracy-enhancement program supplied with the calculator-controlled network analyzer can correct the data, as

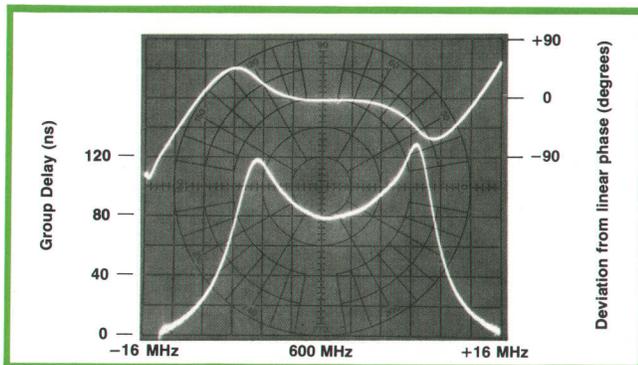


Fig. 4. Dual trace display shows swept measurement of group delay (lower trace) and deviation from linear phase shift (upper trace). Group delay is defined as the negative rate of change of the phase deviation curve.

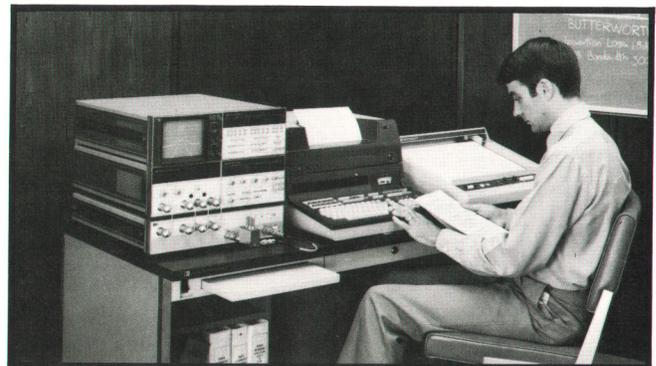


Fig. 5. Factory-assembled and tested automatic systems designed around the Model 8505A Network Analyzer are furnished with example programs and programs for accuracy enhancement, system verification, and diagnostic testing.

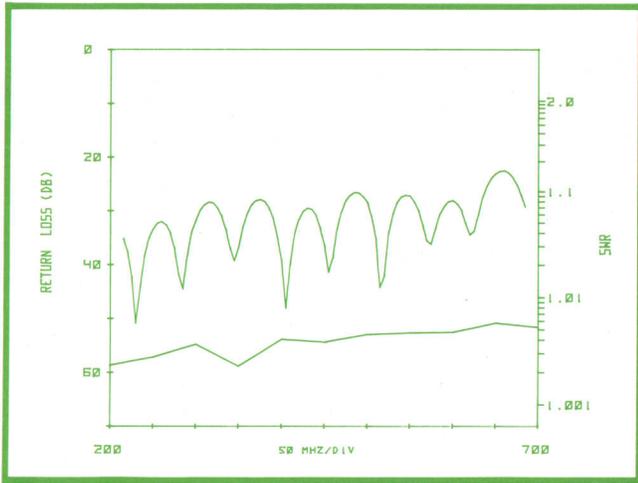


Fig. 6. Accuracy enhancement program removes cable errors from upper trace to give accurate display of device gain (lower trace).

shown by the corrected plot in Fig. 6. First, calibration measurements are made with the test device replaced by a short-circuit termination, then an open-circuit termination, and finally a 50Ω (or 75Ω) termination. With that information stored, the calculator computes the error terms and corrects the data.

Instrument Organization

A simplified block diagram of the network analyzer is shown in Fig. 7. A sweep-frequency test signal anywhere within a 0.5-1300 MHz range is generated by mixing a YIG-tuned 4.2105-5.5100-GHz sweep signal with the output of a fixed 4.2100-GHz oscillator.

A second sweep-frequency signal, offset from the RF output by 100 kHz, is generated by mixing the YIG-tuned signal with the output of a 4.2099-GHz oscillator. This signal is used as a local-oscillator (LO) signal to convert the test signals to 100-kHz IF signals for use in the signal-processing section of the instrument.

The test-signal frequency is measured at any point selected by a marker during the sweep by a counter (the counter actually measures the LO signal and then corrects the reading for the offset). The counter uses a technique that measures the frequency at a single point indicated by a marker without stopping the sweep (see box, next page).

The magnitude and phase information impressed on the test signals by the device under test is translated through the down-converters and detected by the analyzer's amplitude and phase detectors. This information is presented on the CRT display either in cartesian coordinates (Bode plot) or polar coordinates (Nyquist plot). A sample-and-hold circuit retains the

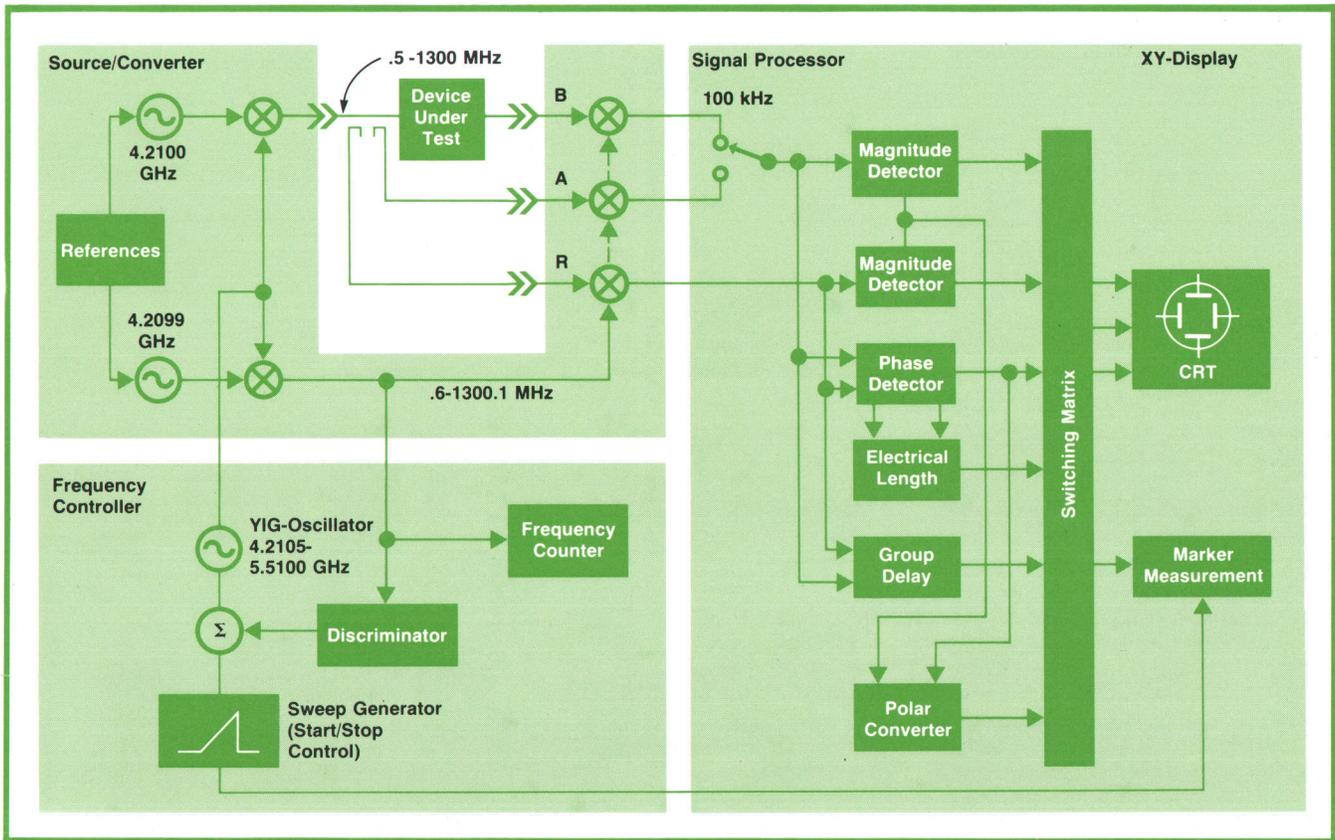


Fig. 7. Organization of the Model 8505A Network Analyzer.

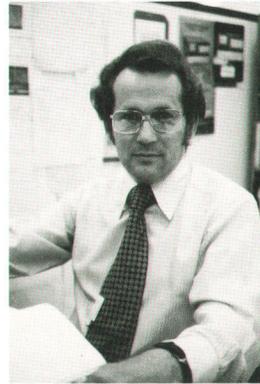
displayed magnitude or phase level at any selected point in the sweep for measurement and display by the built-in digital voltmeter.

Details of the operation of all these circuits are described in the articles that follow.

Acknowledgments

Many talented people, whom the author would like to thank in the name of the design team, made significant contributions to the Model 8505A Network Analyzer. Special thanks go to Dave Eng and Roy Church for the industrial design, and to Doug Rytting, Rit Keiter and Cyril Yansouni for their leadership, support and encouragement during the development of the Model 8505A. 

Hugo Vifian



A Swiss native, Hugo Vifian attended college on a scholarship after working eight years in the production, design and test of telecommunications equipment, earning a Diplom Ingenieur from the Swiss Federal Institute of Technology in 1967. He later earned a PhD from the same school (1973) with Stanford University coordinating his thesis work. He joined HP in 1969, was principal designer for the 8755A Frequency Response Test Set and is now program manager

for the 8505A Network Analyzer instrument family. Spare time activities include canoeing with his wife and twin sons.

ABRIDGED SPECIFICATIONS HP Model 8505A Network Analyzer

Source

FREQUENCY RANGE: 500 kHz to 1.3 GHz
SWEEP FREQUENCY ACCURACY: $\pm 1\%$ of range (linear sweep)
FREQUENCY COUNTER: measures frequency at continuously variable marker position without interrupting frequency sweep.
COUNTER ACCURACY: ± 2 counts \pm time-base accuracy
TIME-BASE ACCURACY: ± 5 ppm ± 1 ppm/°C ± 3 ppm/90 days
FREQUENCY STABILITY: $< \pm 0.01\%$ of reading $\pm 0.01\%$ of range
POWER OUTPUT RANGE: $+10$ dBm to -72 dBm
LEVELING: ± 0.5 dB from 500 kHz to 1.3 GHz
SOURCE IMPEDANCE: 50 Ω ; ≥ 16 dB return loss at -10 dBm output level
RESIDUAL FM: 20 Hz rms, 500 kHz - 13 MHz range (1 kHz bandwidth)
 200 Hz rms, 500 kHz - 130 MHz range (1 kHz bandwidth)
 2 kHz rms, 500 kHz - 1300 MHz range (10 kHz bandwidth)
HARMONICS: > 25 dB below main signal at $+10$ dBm output level.
SUBHARMONICS AND SPURIOUS: below -50 dBm at $+10$ dBm output level.
SWEEP MODES: linear, log, start/stop 1, start/stop 2, alt, CW $\pm \Delta F$.
SWEEP TIMES: 10 ms to 100 sec.
TRIGGER MODES: auto, line sync, external sync, single scan.
OUTPUT CONNECTOR: Type N female.

Receiver

INPUT CHARACTERISTICS (three channels, R, A, and B)
IMPEDANCE: 50 Ω ; ≥ 20 dB return loss (typically ≥ 26 dB).
DAMAGE LEVEL: $+20$ dBm or ≥ 50 V dc.

INPUT CONNECTORS: Type N female.
MAGNITUDE CHARACTERISTICS:
RANGE: -10 dBm to -110 dBm
ABSOLUTE FREQUENCY RESPONSE: ± 1.5 dB
RATIO FREQUENCY RESPONSE (A/R, B/R): ± 1.5 dB
DYNAMIC ACCURACY: from ± 0.01 dB/dB between -20 and -40 dBm to ± 4.0 dB below -100 dBm.
REFERENCE OFFSET ACCURACY: ± 0.02 dB ± 0.003 dB/dB of offset
PHASE CHARACTERISTICS:
FREQUENCY RESPONSE: $\pm 3^\circ$ from 0.5 to 750 MHz; $\pm 5^\circ$ from 750 to 1300 MHz.
RANGE: $\pm 180^\circ$
ACCURACY: $\pm 0.01^\circ$ /degree to $\pm 170^\circ$; $\pm 0.01^\circ$ /degree $\pm 0.5^\circ$ for $\pm 180^\circ$.
DYNAMIC ACCURACY: from $\pm 0.02^\circ$ /dB between -20 to -40 dBm to $\pm 3^\circ$ between -70 and -90 dBm.
REFERENCE OFFSET ACCURACY: $\pm 0.5\%$ of offset
POLAR CHARACTERISTICS: Dynamic response and reference offset specifications same as magnitude and phase characteristics.
FULL-SCALE MAGNITUDE RANGE: 0:01 to 1 in 1-2-5 sequence.
DELAY CHARACTERISTICS:
FREQUENCY RESPONSE: ± 1 ns from 500 kHz to 1.3 GHz.
DELAY ACCURACY: $\pm 3\%$ of reading ± 3 units (unit = 1 ns for 0.5-1300-MHz range, 10 ns for 0.5-130-MHz range, 100 ns for 0.5-13-MHz range).
RANGE: 0 to 80 μ s (0.5-13 MHz), to 8 μ s (0.5-130 MHz), to 800 ns (0.5-1300 MHz).
ELECTRICAL LENGTH/REFERENCE PLANE EXTENSION CHARACTERISTICS:

RANGE:	0.5-13 MHz	0.5-130 MHz	0.5-1300 MHz
X1:	± 19.9 m	± 1.99 m	± 19.9 cm
X10:	± 100 m	± 10 m	± 1 m

ACCURACY: $\pm 3\%$ of reading $\pm 1\%$ of range.
LINEARITY: $\Delta\delta = 0.5\% \times 1.21$ (MHz) $\times 1$ (meters)
LINEAR PHASE SUBSTITUTION (degrees/scan)
RANGE: $\pm 1700^\circ$ per scan with 0° offset ± 1.4 km/scan width (MHz) or $\pm 4.75^\circ$ scan width (MHz)
ACCURACY: $\pm 3\%$ of reading $\pm 10^\circ$ /scan.

General

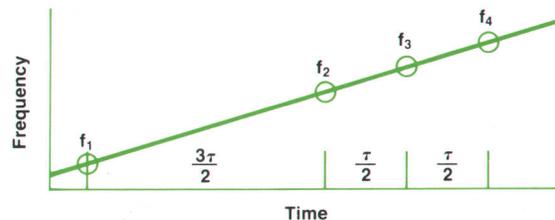
AUXILIARY OUTPUTS: Channels 1 and 2 and sweep (0.25V/div), pen lift.
POWER: 100, 120, 220, or 240V; $\pm 5\%$, -10% , 50 to 60 Hz, 275 W.
WEIGHT: 36 kg (86 lb)
DIMENSIONS: 426 mm W \times 279 mm H \times 553 mm D (16 $\frac{1}{4}$ \times 11 \times 21 $\frac{1}{4}$ in)
OPTIONS: 001, HP Interface Bus
ACCESSORIES: Model 8502A/B Transmission/Reflection Test Set (50 or 75 Ω)
 Model 8503A/B S-Parameter Test Set (50 or 75 Ω)
 Model 11850A/B 3-Way Power Splitter (50 or 75 Ω)
PRICES IN U.S.A.: Model 8505A, \$22,500; Opt. 001, add \$2950.
 Model 8502A/B, \$1850; Model 8503A/B, \$3700.
 Model 11850A/B, \$450.
MANUFACTURING DIVISION: Santa Rosa Division
 3273 Airway Drive
 Santa Rosa, California 95401, U.S.A.

Measuring a Linearly Changing Frequency

A common problem in swept-frequency measurements is how to measure the frequency of any selected point in the sweep while the sweep is progressing. In early instruments, a frequency reading was derived by reading the CRT graticule in conjunction with the start/stop frequency settings. Crystal-controlled frequency markers improved accuracy but it still took time to interpolate the readings and arrive at an answer.

Later instruments used conventional frequency counters but had to stop the sweep long enough for the counter measurement. The difficulty of stopping the sweep quickly and precisely limited the precision of this method and the act of stopping the sweep often introduced transients into the measurement process. Other techniques involved stopping the sweep for measurement during the sweep flyback, but hysteresis in the sweep return affected the accuracy of doing it this way.

A new approach that uses an up-down counter sidesteps these problems. When a marker pulse occurs, the counter starts counting up on the RF frequency and continues counting for a selected gate time (f_1 to f_2 in the diagram). The RF fre-



quency increases linearly during the counter gate, however, so the total count is higher than wanted. The counter pauses, then counts down on the RF for another gate time (f_3 to f_4).

Analysis shows that if the gate times shown in the diagram are used, the number of cycles counted during up-down counting is the same as the number counted at a constant frequency f_1 for a gate time equal to τ . Thus, the instantaneous frequency at a point indicated by a marker can be measured without stopping or otherwise interfering with the sweep.

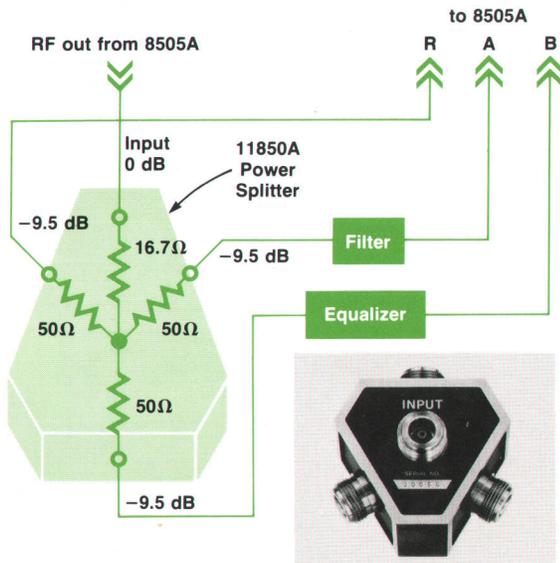
--Frederick Woodhull

A Family of Test Sets for the 500 kHz-1.3 GHz Range

Whether or not good data is obtained from a measurement at high frequencies may be determined by the way the device under test is connected to the measuring instrument. Impedance mismatches, loose connections, and phase shifts through unequal cable lengths can degrade a measurement significantly. To minimize these problems, and thus to assure consistent accuracy in measurements made by the new Model 8505A Network Analyzer as well as for other applications in the frequency range between 500 kHz and 1.3 GHz, a new family of test sets has been developed.

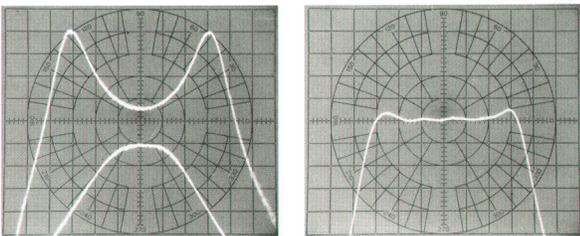
Three-Way Power Splitter

For making comparison measurements such as matching networks, as described in the measurement shown below,



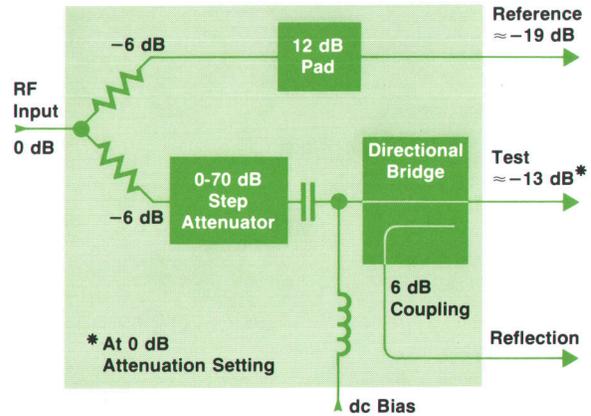
a three-way power splitter is an important tool. The Model 11850A/B Power Splitter was designed to match signal tracking between any two ports within 0.1 dB in magnitude and 1.5° in phase. It is available in both 50Ω and 75Ω versions.

The results of the typical application shown in the diagram are seen in the CRT photos. The photo at left shows the group delays through a filter and a compensating equalizer. The use of the three-way power splitter enables adjustments to be made on the individual networks, and the results observed, before the networks are cascaded for final verification, as in the photo at right.



Transmission and Reflection Test Set

Much time can be saved by making transmission and reflection measurements simultaneously. This is especially true in applications where the device under test must meet certain return loss specifications as well as having a given transmission characteristic, since the two almost always interact. The Model 8502A/B Transmission/Reflection Test Set enables



this to be done with a minimum of external cabling.

As shown in the diagram below, the test set has a two-way power splitter for splitting off a reference signal and it has a bridge circuit that separates the reflected signal from the incident signal. The power level at the test port can be reduced by the built-in step attenuator without affecting the reference signal in applications, for example such as making measurements on sensitive amplifiers. Dc bias up to 30V and 200 mA can be applied to the test port through the bias input for measurements on the parameters of active components.

The 8502A/B Test Set is available in both 50Ω and 75Ω versions. Sets of cables matched for electrical length within 2° at 1.3 GHz are available for making the necessary interconnections for the 8502A/B test set, and also for the three-way power splitter.

The S-Parameter Test Set

All four s-parameters of a device can be measured quickly with the aid of the Model 8503A S-Parameter Test Set. Basically it is like the 8502A Transmission/Reflection Test Set but it includes a second directional bridge for the output port of the device under test and a relay that alternately switches each bridge to the power source and to a power sink for reversing the direction of RF power flow during measurements of s-parameters. It is fully programmable through the HP Interface Bus for use in automatic systems, as well as being operable manually.

Acknowledgments

Steve Lund and Brent Palmer did initial design work on the RF bridge and Jim Jones contributed circuit design. Product design was by Bill Misson and Oleg Volhontseff. Dave Eng and Roy Church did the industrial design.

--Julius Botka

Processing Wide-Range Network Analyzer Signals for Analog and Digital Display

by William S. Lawson and David D. Sharrit

TEST SIGNALS, AFTER PASSING through or being reflected from a device under test, are down-converted to a constant 100-kHz IF in the Model 8505A Network Analyzer and then sent to the instrument's signal processor for measurement.

The signal processor detects the amplitudes, phase relationships, and delays of the test signals, and conditions the results for display on the CRT and for measurement by the built-in voltmeter. Display conditioning includes magnification and offsetting, conversion to polar coordinates if desired, and the switching required for displaying two quantities simultaneously.

The instrument has two identical processing channels, permitting the simultaneous measurement and display of two independent quantities. Two channels also make possible measurements of the ratio of two input signals and the delay and phase relationships between them. One of the processing channels is time-shared by the A and B measurement input ports, enabling concurrent display of two ratio measure-

ments such as simultaneous display of a device's transmission and reflection characteristics. This channel has the electronic line stretcher that introduces up to $\pm 1700^\circ$ of electrical length.

Processor operation is under control of a ROM-based digital processor. It stores independent offsets for magnitude and delay for both measurement channels, and independent electrical length offsets for the A and B inputs. It also has calibration storage registers that enable the 0-dB reference line on the display to correspond to any input power level.

Internal Details

A simplified block diagram of the analog signal processing circuits is shown in Fig. 1. The signals fed to the input connectors first go to the down-converters (not shown) and the resulting 100-kHz IF signals go to the signal processor where they are buffered and bandpass filtered before being sent to the magnitude and phase detectors. The gain of the buffer amplifier is changed by 20 dB in accordance with

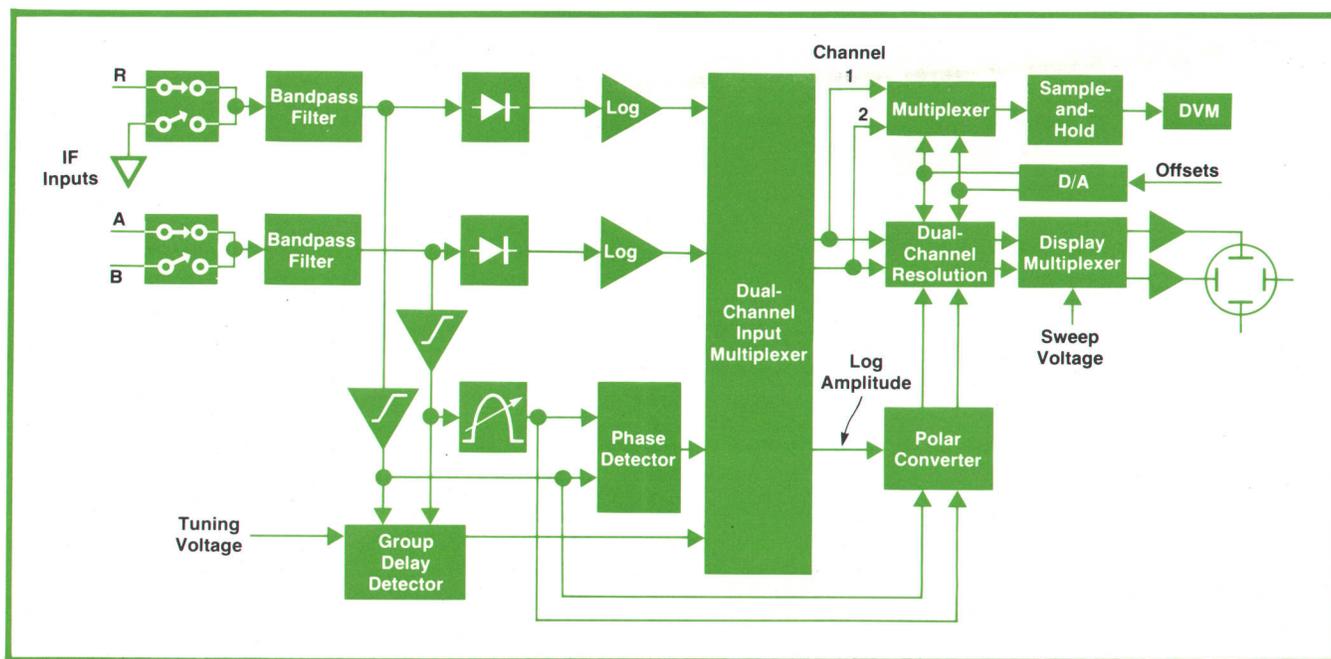


Fig. 1. Simplified block diagram of the signal processor in the Model 8505A Network Analyzer.

A Wide-Range, Continuously-Variable Line Stretcher

A line-stretcher is an indispensable tool for high-frequency network analysis because it enables the electrical reference point of a phase measurement to be placed physically anywhere required by the measurement. Usually, the reference point is placed at the input to a device to enable measurement of the phase shift through the device. Knowing the phase shift as a function of frequency, the equivalent electrical length of the device can be derived. By inserting into the reference path an additional length equivalent to that of the device, the deviations from linear phase shift through the device can then be determined.

Telescoping sections of coaxial line plus fixed sections of line of various lengths have been used for adjusting the electrical length of an RF path for these measurements. Besides the obvious inconvenience of doing it this way, this arrangement also contributes to measurement errors because of the imperfect impedance matches that connectors inevitably introduce into the test signal path.

The electronic line-stretcher used in the Model 8505A Network Analyzer makes it much easier and faster to shift the phase reference point over an extremely wide range, and it introduces no mismatch errors. Basically, it is a voltage-controlled phase shifter that introduces a phase shift proportional to the RF test-signal frequency, thus simulating a fixed time delay.

The basic scheme is to up-convert the incoming IF signal (100 kHz) by mixing it with a local oscillator signal and removing all but the first-order upper sideband mixing products with a narrowband filter. The signal is then down-converted to 100 kHz by a second local oscillator that is phase-locked to the first. A phase offset proportional to the RF frequency is introduced into the phase-lock loop so the phase shift of the IF output varies with respect to the IF input in proportion to the RF frequency.

The theory behind this operation is as follows. The output of the reference (first) oscillator is $\cos(\omega_m t + \phi_{ref})$. When the reference frequency is mixed with the incoming IF frequency, ω_{IF} , the mixer output has many components: $N\omega_m \pm M\omega_{IF}$. The narrowband filter removes all but the first-order upper sideband, namely $\omega_m + \omega_{IF}$. This is then mixed with the voltage-controlled-oscillator signal, $\cos(\omega_m t + \phi_{VCO})$.

The output of the second mixer is

$$V = A \cos[\omega_m t + \omega_{IF} t + \phi_{ref}] \times \cos[\omega_m t + \phi_{VCO}]$$

The trigonometric terms when expanded become

$$\begin{aligned} & \frac{1}{2} [\cos(\omega_{IF} t) \times \cos(\phi_{ref} - \phi_{VCO}) + \\ & \frac{1}{2} \cos(2\omega_m + \omega_{IF}) t \times \cos(\phi_{ref} + \phi_{VCO}) + \\ & \frac{1}{2} \sin \omega_{IF} t \times \sin(\phi_{ref} - \phi_{VCO}) - \frac{1}{2} \sin(2\omega_m + \omega_{IF}) t \sin(\phi_{ref} + \phi_{VCO}) \end{aligned}$$

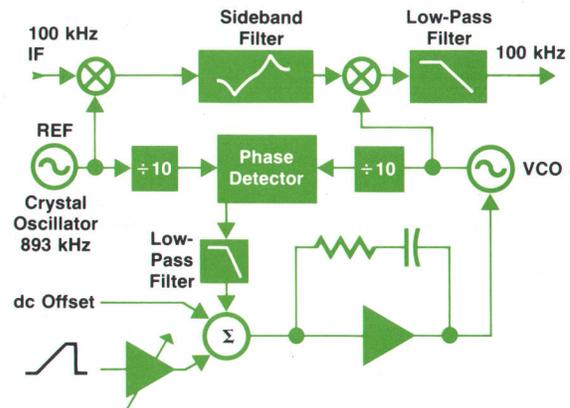
Low pass filtering removes the higher order terms. Substituting ϕ_{OFS} for $\phi_{ref} - \phi_{VCO}$, the mixer output then becomes:

$$V = A (\cos \omega_{IF} t \times \cos \phi_{OFS} + \frac{1}{2} \sin \omega_{IF} t \sin \phi_{OFS})$$

which yields: $V = A \cos(\omega_{IF} t + \phi_{OFS})$

Thus, the original IF input signal is recovered but with a phase offset equal to the offset between the two oscillators.

A block diagram of the electronic line stretcher is shown below. A voltage ramp derived from the RF source sweep-control voltage is added to the phase detector output to give a phase offset proportional to RF frequency. The ramp is scaled according to the effective line length desired. A dc offset derived from the instrument's offset storage register is added to permit control of the steady-state phase shift.



The phase detector is a conventional 0-360° detector. To extend the offset range, the oscillator outputs are divided by a factor of 10 before the phase comparison, allowing the VCO to be offset almost as much as 3600° from the reference. By suitable choice of a voltage level to place the zero offset condition at the center of the phase detector's range, the offset operates over a range of $\pm 1700^\circ$.

It is possible to obtain a wider phase-shift range by dividing the oscillators' outputs by factors greater than 10 but noise performance would then suffer and the filtering required to keep feedthrough in the phase detector out of the control loop would be more difficult.

Although it would be preferable to use a higher oscillator frequency for reasons of noise and loop stability, the sideband filtering requirements establish a practical limit. The lower sideband must be suppressed more than 70 dB below the upper sideband to keep "sideband" noise below visible threshold on the CRT trace.

--William Lawson

the setting of the MAX INPUT LEVEL switch (-10 dBm, -30 dBm), which enables improved accuracy when measuring low-level signals.

The A and B inputs time-share the measurement channel by means of the electronic switches shown at the IF inputs. To eliminate interference between the channels, the switches were designed for 130-dB "off" isolation; careful shielding and layout keeps

crosstalk between them more than 120 dB down.

The magnitude detector is similar to conventional average-responding meter-rectifier circuits. The resulting dc is processed through a logarithmic amplifier and buffered before going to the input multiplex switch. The two magnitude detectors can be calibrated to track within 0.1 dB over a range of input levels greater than 60 dB.

Phase Detection

For phase detection, IC line receivers perform a limiting function on the incoming sinusoids to achieve constant-level signals. The test signal then goes through the electronic line stretcher (see box, page 9) while the reference signal goes through a compensating bandpass filter. Both signals are squared in Schmitt triggers and then used to drive a J-K flip-flop that performs the phase comparison. The resulting rectangular pulses, whose widths are proportional to the phase difference between the two signals, are integrated in a low-pass filter to derive a dc level proportional to phase difference. This signal goes to the input multiplex switch.

With the limiting provided, this phase detector can work over a 100-dB range of input signal levels. Measurement uncertainty is shown in Fig. 2 as a function of signal level. At levels greater than -60 dBm, detection accuracy is better than $0.01^\circ/\text{degree}$.

Following the input multiplexer, offsets are added to the detected signals (magnitude and delay) and the results are amplified according to the settings of the SCALE/DIV switches. Video filtering can be introduced here if the displayed signal is particularly noisy. Finally, the signals enter the display multiplex where they are switched to the display according to the display mode selected. The marker waveforms are added to the displayed signals at this point.

Markers are generated in response to triggers derived from comparators that respond to the level of the sweep tuning ramp. The combination of a single unipolar triangle () for the vertical deflection and a bipolar dual triangle () for the horizontal generates a diamond-shaped marker that is easily discerned regardless of the slope of the displayed response curve.

Digital Readout

As mentioned previously, in addition to the real-time CRT display of a network's swept response, Model 8505A provides digital readout of the network's gain, phase or delay at the frequency desig-

nated by a marker. This is done by sampling the response of the selected detector wherever the particular marker occurs during a sweep. A sample-and-hold circuit retains the detected value for conversion by a 2000-count dual-slope A-to-D converter, which sends the results to the digital processor where the counts are averaged. The processor drives the appropriate digital readout on demand.

The sample-and-hold circuit is preceded by a low-pass filter and an amplifier that has switchable $\times 1$ or $\times 10$ gain. The $\times 10$ gain is switched in to retain 3-digit resolution whenever the measured quantity is less than 10 dB or 100° .

Besides reading the value of the quantity displayed on the CRT trace at the frequency of the selected marker, the digital readout can also show the offset corresponding to the measurement mode selected, i.e. magnitude or delay. The choice of readout is made by the pushbuttons labeled MKR (for marker) and REF (for reference)

While being displayed digitally, the offset can be modified by the three pairs of UP/DOWN increment buttons, one pair for each digit in the display. The display can be returned quickly to its calibrated position by pressing the CLR button momentarily, which clears the corresponding offset register.

Once set, an offset is retained in a storage register even though the measurement mode may be changed. The offsets in storage are sequentially loaded into a pulse-width modulator that drives each of four low-pass filters in turn, giving dc values for the four offsets used in the analog signal processing.

In addition to storage registers for offsets, the instrument has storage registers for display calibration. For example, it is often convenient to establish the 0-dB reference level at the signal level within the passband of a filter regardless of the filter's insertion loss. Finding the 3-dB points or other points of interest is then straightforward, with no need to account for insertion loss.

There are two ways the operator can change the display calibration. One is to switch to the REFERENCE mode, use the UP/DOWN buttons to move the desired part of the trace to the reference line, then press the ZRO pushbutton. This adds the contents of the appropriate offset register to the corresponding calibration register and clears the contents of the offset register. The digital display goes to 0.00 and the REL indicator turns on to show that a non-zero number is in the calibration register.

The second way is to use the MARKER mode, position the selected marker at the desired point on the trace, then press the ZRO button. The digital processor compares the stored offset to the value of the measured signal at the point designated by the marker, and increments the offset until this difference is zero.

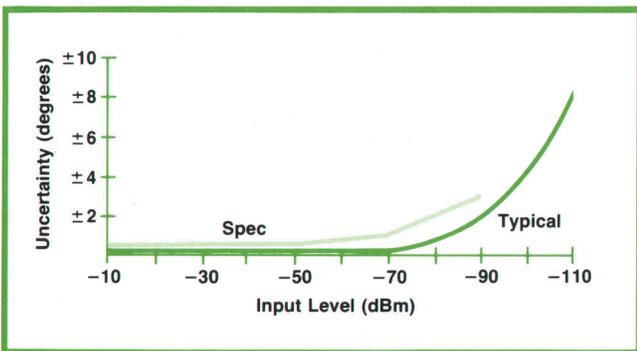


Fig. 2. Accuracy of phase measurement over a 100-dB range of input levels.

A Wide-Range, High-Resolution Group Delay Detector

Measurement of the group delay of a network is important for the evaluation of the distortion characteristics of a transmission device. A distortionless device introduces a phase shift that is a linear function of frequency so it has the same time delay for all frequencies. When a complex signal passes through a device that delays some frequency components with respect to the others, the device distorts the signal even though the relative amplitudes may remain the same.

Group delay (t_D), a term originating in the telephone industry, is defined as the negative rate of change of phase with frequency: $t_D = -d\phi/d\omega$. A plot of group delay versus frequency in a distortionless device is then a straight horizontal line. With typical devices the phase shift does not vary linearly so the phase slope, and hence the group delay, varies as a function of frequency.

The linear sweep-frequency characteristics of the Model 8505A Network Analyzer enable direct measurement of group delay. The sweep-frequency test signal supplied to the device under test can be represented as:

$$V_{in} = A \cos(\omega_0 + \frac{1}{2}\alpha t)$$

where A is the signal peak amplitude, ω_0 is the frequency at $t = 0$ and α is the rate of change of frequency.

The output of the device, assuming a quasi-stationary solution, can then be expressed as:

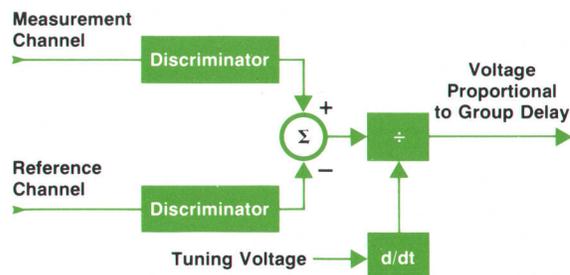
$$V_{out} = G \times A \cos\{[\omega_0 + \frac{1}{2}\alpha(t - t_D)](t - t_D) + \phi\}$$

where G , t_D and ϕ are the gain, delay, and residual phase of the device at ω_0 .

The shift in frequency between input and output is $\Delta\omega = -\alpha t_D$; so the delay is $t_D = -\Delta\omega/\alpha$. In the Model 8505A, the delay at frequency ω_0 is measured by determining the frequency difference between input and output of the device under test during a sweep, and dividing by the rate of change of frequency. For example when sweeping a 100-MHz band in a sweep time of 10 ms, a frequency shift of 10 Hz is equivalent to a 1-ns delay.

At slow sweep rates or with narrow sweeps, the frequency difference is small and the signal-to-noise ratio of the measure-

ment is reduced, as it is in other methods of delay measurement. With slow and/or narrow frequency sweeps, the Model 8505A obtains calibrated group delay measurements with acceptable signal-to-noise ratio by superimposing a small linear 1-kHz sawtooth waveform on the sweep tuning ramp and making the delay measurement coincident with the peak of the sawtooth. The frequency "aperture" may then be kept small so ripples in the delay-versus-frequency curve are not smoothed out.



The block diagram for the circuits performing the group delay measurement is shown above. The two discriminators obtain voltages proportional to the frequencies at the input and output of the device, and a voltage proportional to the frequency difference $\Delta\omega$ is obtained by subtracting one from the other. The tuning ramp from the frequency sweep control is differentiated to obtain α , the rate of change of frequency, and used to divide $\Delta\omega$ to obtain t_D .

For slow and/or narrow sweeps, the sawtooth voltage is added to the frequency control ramp. The output of the differentiator is then filtered by a synchronous low-pass filter to obtain a dc voltage proportional to the sawtooth up-slope. The output of the divider is gated so the measurement is made at the peak of the sawtooth waveform.

--David Sharrit

Acknowledgments

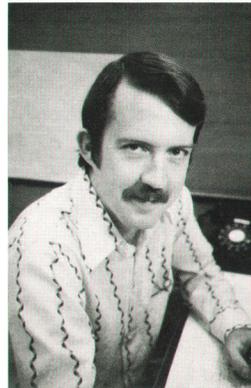
Roger Oblad designed the detectors, Jim Jones the

CRT display, and John Barr the HP-IB interface. Product design was by Oleg Volhontseff. 



William S. Lawson

Project leader for the signal processor part of Model 8505A, Bill Lawson previously worked on the Model 8600A Digital Marker and the 86220A 10-1300-MHz RF Drawer. Prior to joining HP in 1969, he worked two years on an electron paramagnetic resonance spectrometer at another firm. He has BSEE and MSEE degrees from Stanford University. From southern California, he now lives in Santa Rosa with his wife and 5-week-old son. Outside of work he golfs, plays tennis and gardens.



David D. Sharrit

Graduating with a BSEE degree from Arizona State University in 1970, Dave Sharrit designed radar signal processors before joining HP in 1973 to work on the Model 8505A Network analyzer. Dave is pursuing an MSEE degree in the HP Honors Co-op program, using the video link between Stanford University and HP's Santa Rosa facility. For recreation, Dave enjoys outdoor activities, primarily bicycling.

A Precision RF Source and Down-Converter for the Model 8505A Network Analyzer

by Rolf Dalichow and Daniel R. Harkins

WHILE SUPPLYING A CLEAN TEST signal with well-controlled amplitude over a multi-decade frequency range, the RF system of the Model 8505A Network Analyzer has to supply a tracking signal that is precisely offset in frequency. It is used as an LO signal by the input RF down-converters in the receiver.

A key factor in the analyzer's performance is the design of the down-converters. To be meaningful, network measurements require that the input ports have equal phase and magnitude characteristics. This makes accurate ratio measurements possible. It is also desirable that each input have a wide dynamic range, as this broadens the range of applications.

Each of the three input ports of the Model 8505A has a 100-dB dynamic range and tracks the other two in phase and magnitude over the 0.5-1300-MHz frequency range, as indicated by the ratio response plotted in Fig. 1. Contributing to this performance is the use of only one frequency-conversion step, made possible by the choice of an IF frequency that is below the RF frequency range of the instrument.

The basic organization of the RF system was indicated in the diagram on page 5. The RF test signal is generated by mixing a 4.2105-5.510-GHz swept microwave signal with a fixed 4.210-GHz signal and supplying the amplified and leveled difference frequency to the output port. The use of these high frequencies results in a multi-decade stimulus signal that is quite free of higher-order mixing products.

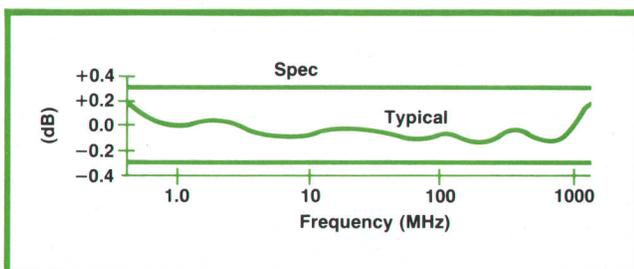


Fig. 1. Frequency response of the Model 8505A Network Analyzer in making amplitude ratio measurements (A/R or B/R). Close tracking of the input mixers is required to obtain this performance.

The offset LO signal is generated by mixing the swept microwave signal with a fixed 4.2099-GHz signal, resulting in a 0.6-1300.1-MHz signal that is always 100 kHz above the RF output signal. If good tracking characteristics are to be achieved, it is necessary for each down-converter to receive this LO signal at the same level and with the same phase as the others. Uniform drive characteristics are assured by mounting the three mixers in close proximity, an arrangement that on the other hand, had interesting implications in achieving 100-dB isolation between input channels. The solution to this is discussed in the box on the next page.

Stabilized 4.2-GHz Oscillators

Although operating at about 4.2 GHz, the two fixed oscillators of the RF generating system are offset only 100 kHz from each other. The response characteristics of the receiver circuits require that this frequency difference be maintained within ± 50 Hz over the entire tuning range of the instrument. With this and the requirement for frequency accuracy, some sort of phase-locking system is the obvious design choice. Phase-locking also offers improved noise sidebands, necessary for achieving the goal of 100-dB dynamic range.

Of the various ways of phase-locking two oscillators, an approach was selected that promised to be the most efficient in terms of circuit complexity. It is diagrammed in Fig. 2. The two microwave signals are generated in identical thin-film microcircuit oscillators and each is phase-locked to a separate crystal reference. The references, however, operate at much lower frequencies where the 100-kHz difference is more easily maintained. The crystals are matched for temperature coefficient so the 100-kHz difference varies less than 25 Hz over a temperature range of 25-40°C, eliminating the need for a temperature stabilizing system.

Samplers perform the down-conversion for the phase comparison at the lower frequencies. The sampling rate is 100 MHz so the low-pass filtered output of the sampler is the 10-MHz difference between 4.210 GHz and the 42d harmonic of 100 MHz. At this

Well-Matched Down-Converters

Input mixers of the Model 8505A Network Analyzer were to have a good input impedance match, less than 0.1-dB compression at a -10 -dBm input level, and a noise floor that would permit a 100-dB dynamic range. The mixers were to track closely in amplitude and phase and cross talk between them was to be more than 100 dB down.

To meet these requirements, the mixer design diagrammed below was developed. Constructing the mixers as thin-film, hybrid microcircuits on a sapphire substrate assured uniformity.

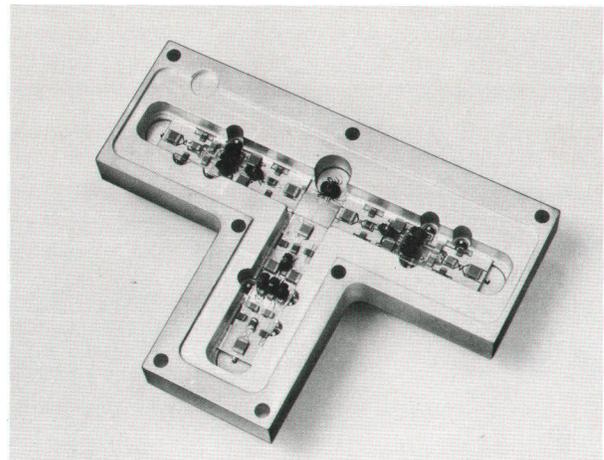
As shown in the diagram, both the signal and LO inputs to the mixer are through common-base amplifiers. This configuration was chosen for three reasons. First, the phase and amplitude characteristics are consistent from transistor to transistor of different lots. Second, the input impedance is only 2 to 4 ohms throughout the 1300-MHz frequency range. Third, the configuration has high reverse isolation, necessary for achieving the 100-dB crosstalk specification.

The low input impedance allows the overall input impedance to be set by a series input resistor, obtaining a typical return loss of 24 dB with little signal loss. The noise figure of the input mixer is about the same with the amplifier as it would be without, but without the amplifier an input attenuator would be needed to obtain a good return loss figure, and this would reduce the dynamic range.

A coplanar waveguide structure rather than the usual microstrip transmission-line structure is used for the microcircuits to keep the ground currents on the top surface. Chip connections to ground can thus be made short for low inductance. With the resulting very short base leads, the reverse isolation

of each amplifier is about the same as the transistor chip itself (typically 36 dB at 1300 MHz). The coplanar structure also allows variations in the spacing of the ground conductor for better control of the inductance of the input capacitor and resistor.

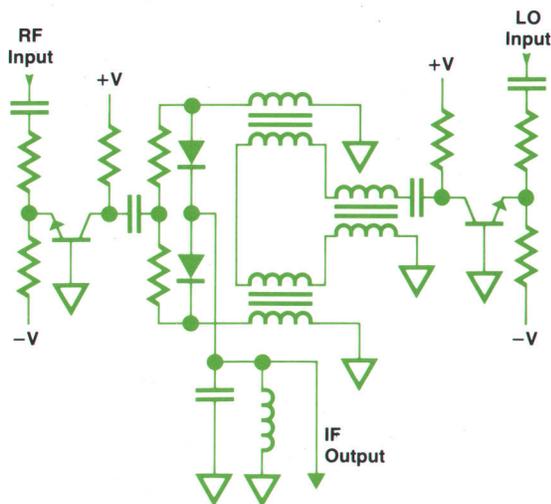
The RF signal drives the diodes of the single-balanced mixer by way of a resistive power divider. The LO drive is through a balun structure. A balun is required to balance the LO drive signal to ground, isolating the LO signal from the RF and IF paths. Three separate baluns are used. The first "lifts" the input to the other two so the parasitic capacitance-to-ground of their inputs is balanced, tending to cancel any RF and IF feedthrough to the LO channel.



The 100-kHz tank circuit in the IF output improves the signal-to-noise ratio but its impedance in the RF range is higher than wanted. The solution to this problem was to use an IF input amplifier that has an input impedance of less than 3 ohms. The IF input amplifier is not part of the microcircuit but it is mounted close to the down converter for minimum line length. It is followed by a 300-kHz low-pass filter for further attenuation of the LO and RF signals.

As shown by the photo of the microcircuit above, the three mixers are arranged symmetrically with respect to the LO drive input to establish an identical LO drive path for each mixer. A single inductive loop in each branch of the LO power divider, along with a chip capacitor, forms a low-pass filter to ground for image feedthrough into the LO drive. Overall, the down-converters achieve the design goals with respect to input return loss, dynamic range and crosstalk. Amplitude tracking between mixers is better than $\pm 3^\circ$ up to 750 MHz, and within $\pm 5^\circ$ to 1300 MHz.

--Wayne Frederick



point, the sampler noise floor is some 10 dB below that of the oscillator, allowing for some noise improvement within the bandwidth of the phase-lock loop.

Dual-Tracking Sources

The 4.2-GHz oscillator outputs are translated down to the 0.5-1300-MHz frequency range by the arrangement shown in Fig. 3. The swept microwave source is supplied through a hybrid power splitter as the high-

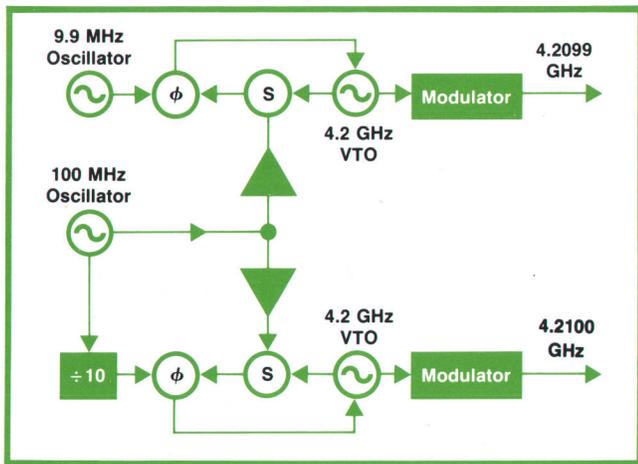


Fig. 2. Source oscillators are phase-locked to crystal references to obtain 100-kHz frequency offset between oscillators.

level (+10 dBm) LO signal to the double-balanced mixers. The stabilized 4.2-GHz oscillator outputs are the low-level inputs.

The principal design effort with these tracking sources was directed towards maintaining greater than 90-dB isolation between the two output signals to prevent spurious responses in the measurements. The techniques used in obtaining this high isolation are described in the box on the next page. Very clean spectrums for the two tracking sources were obtained as a result of the high isolation, plus the fact that the lowest ordered spurious mixing product falling within the 0.5-1300-MHz band is a seventh-order product.

Frequency Control

The 4.2105-5.5100-GHz swept microwave signal is generated by a YIG-tuned oscillator (YTO) tuned by a voltage ramp originating in a Miller-integrator type sawtooth generator. The ramp also drives the CRT horizontal deflection.

Sweep oscillators like this have been designed and built many times before. However, the principal con-

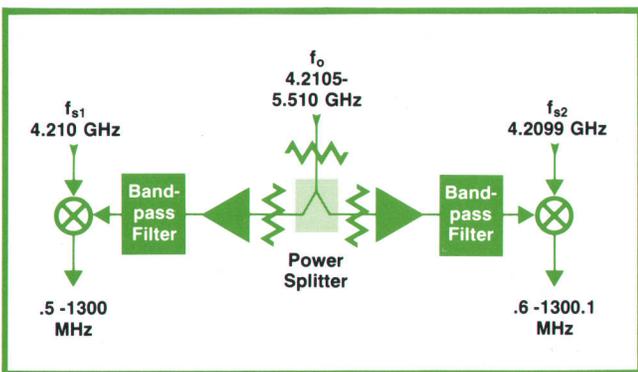


Fig. 3. Stabilized oscillator outputs are mixed with the YIG-tuned sweep-frequency signal to obtain the network analyzer's RF output and the offset tracking LO signal.

straint on this design, aside from the need for very good stability, was the need for digital programmability.

Fig. 4 diagrams the technique for digitally setting the stop and start frequencies of what is essentially an analog circuit. The sweep voltage is inverted and offset in the inverting amplifier. Both the non-inverted and inverted ramps are attenuated by digitally-controlled attenuators and then summed in the ramp output amplifier. DCA1 attenuates the inverted ramp according to the selected start frequency (the non-inverted ramp is always zero at the start) and DCA2 attenuates the non-inverted ramp according to the desired stop frequency (the inverted ramp is always zero at the end of the ramp).

The digital data bus controlling the attenuators also drives the digital readouts for the start and stop frequencies. By eliminating the dials, scales, pointers and dial cords usually used for frequency indication, the digital system simplified the mechanical design of the instrument to a great extent. It also made it possible to store and recall different frequency settings, useful for alternate display of response curves over two different sweep frequency ranges.

For the display of Bode plots, a logarithmic sweep, obtained by passing the linear ramp through an exponential amplifier, is provided in three fixed ranges (1-10 MHz, 1-100 MHz, 1-1000 MHz).

Expanded Sweep

Any two frequencies within the range of the instrument may be selected for the start/stop end points of the frequency sweep by the two front-panel FREQUENCY controls. When the MODE switch is set to LIN FULL or LOG FULL, causing the instrument to sweep its full range, the selected end points are indicated by two markers on the CRT trace. The operator can then use the FREQUENCY controls to bracket a frequency range of interest, and then switch to the LIN EXPAND mode to expand the selected range to full display width.

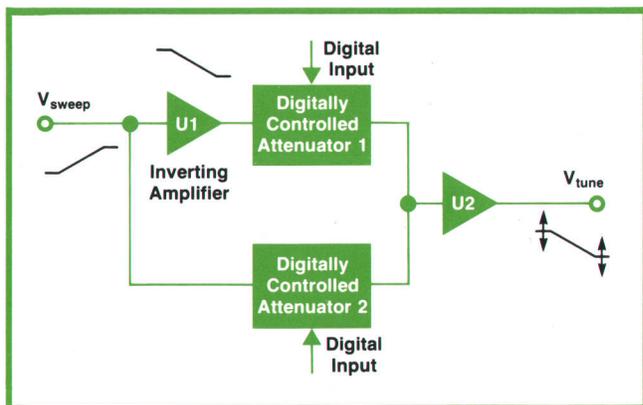


Fig. 4. Start and stop levels of the YIG-tuning ramp are controlled digitally by the arrangement shown here.

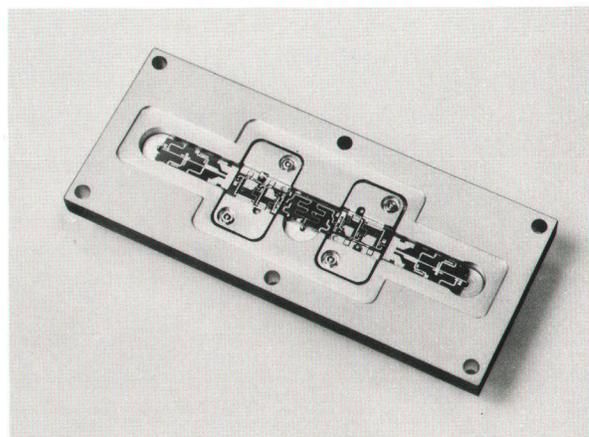
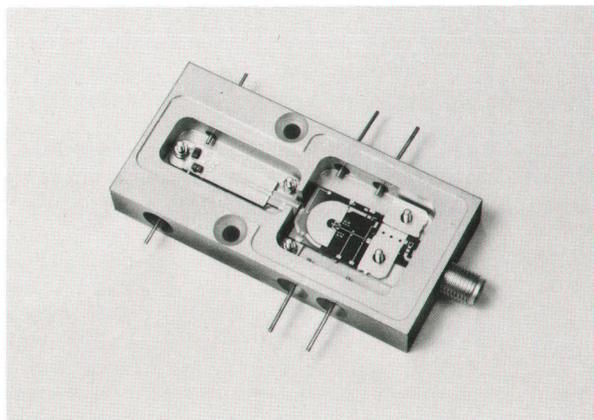
Generating Well-Isolated Tracking Source Signals

The photo below shows one of the two identical hybrid thin-film 4.2-GHz oscillators used in the RF section of the Model 8505A Network Analyzer. Each oscillator is a push-pull, transmission-line-coupled type similar to one previously described for the HP Model 86602A Synthesizer RF Section.¹ The phase-lock circuits drive a varactor diode coupled to one end of the resonator, giving a tuning range of about 15 MHz. A second tuning input is provided by the collector circuit which allows a 35-MHz variation in response to collector voltage variations. This input is used for coarse tuning and for temperature compensation. Because of the constant-current characteristics of the collector circuit, a simple thermistor network in series with this input holds temperature-induced frequency variations to less than 3 MHz over a 60°C temperature range.

Two outputs are coupled out of the oscillator. One goes to a reflective modulator that has greater than 40-dB dynamic range. It provides leveling and vernier control of the RF output. The other goes to the sampler, shown in the left half of the microcircuit in the photo.

The sampler operates similarly to others previously used in HP instruments.² By integrating each oscillator and its associated sampler into a hybrid microcircuit, all of the microwave operations can be performed within the package, simplifying shielding and reducing the number of RF connectors to one.

The basic circuitry for converting the oscillator outputs to the 1.3-GHz frequency range is diagrammed in Fig. 3 on the opposite page. Control of three kinds of leakage was crucial in attaining high isolation between the two output signals. First was the leakage of the signal frequency (f_s in Fig. 3) through the first mixer across the LO amplifier and power splitter into



the f_o port of the second mixer, where it would be down-converted and introduced into the IF output of the second channel.

Second was the third-order mixing product ($2f_o - f_s$) generated in the first mixer leaking out the f_o port across the power splitter into the second mixer. Third was the leakage of the sum frequencies, $f_o + f_s$, coupling from mixer to mixer across the common LO path. This proved to be the most troublesome.

The first two sources of leakage depend on mixer balance between f_s and f_o ports, forward and reverse gain of the amplifiers, splitter isolation, and the two-tone conversion efficiency of the mixer. Mixer characteristics were well-controlled by the use of a thin-film double-balanced mixer. The isolation in the LO drive path was improved significantly by the use of a two-section hybrid power splitter with 3-dB pads at its port.

Leakage of the sum frequencies $f_o + f_s$ was further attenuated by insertion of a 4-6-GHz bandpass filter between each amplifier and its associated mixer. However, surface-wave modes and the TE_{10} waveguide mode were potential problems. Therefore the package dimensions for the hybrid splitter, amplifiers and filters, shown in the photo above, were chosen in such a way that any moding occurs above the frequency range of interest.

--Mark Roos
--Philip Chen

References

1. R. Hassun, M. Humphreys, D. Scherer, Y.D. Kim, B. Stribling, and C. Cook, "Synthesized Signal Generation to 1.3 Ghz," Hewlett-Packard Journal, March 1973.
2. J. Merkelo, "A dc-to-20-GHz Thin-Film Signal Sampler for Microwave Instrumentation," Hewlett-Packard Journal, April 1973.

The sweep WIDTH control has positions labelled START/STOP 1 and START/STOP 2 allowing two sets of sweep end points to be selected and stored in RAM. When the control is moved, say, from position 1 to position 2, the sweep end points selected while the switch was in position 1 are stored. The FREQUENCY controls may then be used to select a new pair of end points. Moving the WIDTH control back to 1 stores the new pair of points and recalls the previously stored pair.

When the WIDTH control is set to ALT, the stored

end points are used alternately to control the sweep, causing simultaneous display of two frequency sweeps. This provides a means, for example, of displaying a high resolution view of a filter's passband at the same time that the total filter response, including the stop bands, is shown. The effect that adjustments made on the passband have on the stop band are readily seen as the adjustments are made.

When the WIDTH control is set to $CW \pm \Delta F$, the left-hand FREQUENCY control selects the center frequency of a sweep and the right-hand control selects

the sweep width.

Markers

Up to five markers can be positioned on the displayed trace by the MARKER controls. Markers are initiated by comparators that generate a pulse when the frequency-control ramp reaches the dc level of a voltage set by the corresponding MARKER control. A switch selects the number of markers displayed (these are in addition to the two sweep end point markers displayed in the full sweep modes). The highest numbered marker initializes data taking by the digital measurement circuits. This marker points up above the displayed trace while the others point downwards.

YIG Oscillator Stabilization

For consistent measurements of Q over a wide frequency range, drift and residual FM have to be more tightly controlled at the low end of the frequency range. Therefore, to allow full-scale display of low-end narrowband sweeps, the network analyzer provides a choice of three frequency ranges with two low-end bands (0.5-13 MHz, 0.5-130 MHz) stabilized.

A sample of the RF output is applied directly to a low-noise discriminator and the output of the discriminator goes to the control-loop summing point to correct the YIG tuning voltage for any drift or residual FM in the RF output. With this control system, residual FM is only 20 Hz rms on the low range and 200 Hz rms on the middle range.

Acknowledgments

In addition to those credited in the boxes inserted in this article, there were many contributors to the project. Weldon Jackson, George Henry, Pete Planting, and Floyd Bishop provided assistance with the hybrid technology and Billy Knorpp and Rich Swain contributed to the microelectronic development. Jerry Ainsworth and Les Brubaker were involved in the initial design and Rich Pope contributed several circuit designs, including the YIG stabilization loop. Product design was by Al Knack. Dennis Handlon and Ron Zimmerman guided the instrument into production. 



Rolf Dalichow

Project leader for the RF frequency control section of the Model 8505A, Rolf Dalichow came to HP in 1973 after 12 years experience in hybrid RF circuits and communications transmitter design both in the United States and in West Berlin, Germany. Rolf obtained an Ing. Grad degree in 1961 from the Staatliche Ingenieur Schule, Giessen, Germany. When time permits, Rolf, his wife, and their two children, 4 and 7, enjoy travel and camping in their camper.



Daniel R. Harkins

Dan Harkins joined HP in 1969, working initially on investigative projects and subsequently becoming project leader for the Model 8505A Network Analyzer's RF source/down-converter. He earned a BSEE degree from Seattle University (1969) and later obtained an MSEE degree from Stanford University in the HP Honors Co-op program. A native of Snohomish, Washington, Dan now lives in Santa Rosa, playing tennis and going hiking.

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