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HEWLETT-PACKARD JOURNAL

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Microwave spectrum analyzers were first recognized as practical measurement tools during World War II when they were used to evaluate how effectively radar transmitters concentrated their radiated power into spectral bandwidths detectable by the receivers. The narrowest resolution bandwidth of these first analyzers was a wide 100 kHz, the maximum frequency sweep was merely 30 MHz, and results were only qualitative. Nevertheless, the kind of information they provided—separation and simultaneous display of the frequency components that make up a signal—proved to be invaluable in analyses of complex signals. Further developments came slowly, but by the 1960s, bandwidth had narrowed to 1 kHz, 2-GHz sweeps were possible and spectrum analyzers were beginning to find wide use on the lab bench for evaluating the performance of mixers, oscillators, amplifiers, and other high-frequency circuits as well as providing quantitative measurements of electromagnetic interference and radio spectrum activity. By the 1970s, resolution had sharpened to 100 Hz and it had become possible to measure amplitude and frequency directly from the display. Now for the 1980s, the new microprocessor-controlled Model 8566A is capable of 10-Hz resolution, 20-GHz sweeps, and precision digital readout of amplitude and frequency. (The 8566A on the front cover is displaying a 20-GHz sweep—the steps on the noise floor occur where the sweeping local oscillator automatically steps back in frequency and then continues the sweep upwards using the next higher harmonic.)

Usually the design of a digital system controller and external hardware is parcelled out to several engineers, each of whom faces a basic problem: how to test each module thoroughly before integrating the modules into the system. The 8170A Logic Pattern Generator, described on page 20, addresses this problem by being able to simulate real-time data-bus traffic, complete with handshake signals when needed. Thus, it can be used to test modules for bus compatibility so that hardware, firmware, and software development can proceed independently.

As the range of applications for HP-interface-bus-controlled instrument systems has expanded, there has been a growing demand for some means of allowing the bus to operate with units separated by more than the 20 metres that the standard bus circuitry allows. This limitation on distance is now removed by the development of the 37201A HP-IB Bus Extender, described on page 26. This device transfers the HP-IB data through a modem onto telephone lines and sends it reliably any distance the telephone system allows—half way around the world if need be—to another extender that converts it back to HP-IB format for the units at the distant location.

-H. L. Roberts

New Performance Standards in Microwave Spectrum Analysis

Low-level microwave signals not previously identifiable with spectrum analyzers can be measured up to 22 GHz with the aid of this new analyzer's low phase noise, 10-Hz bandwidth, and high sensitivity.

by Siegfried H. Linkwitz

ADVANCING TECHNOLOGY NOW makes it possible for a spectrum analyzer designed for the microwave frequency range (above 1 GHz) to achieve the same frequency and amplitude accuracies as the best of those designed for lower frequencies.

In the past, spectrum analyzers designed for the microwave range had limited accuracy for the measurement of an unknown signal's frequency, and were unable to resolve closely spaced spectral components. These limitations existed because of the difficulty of determining the exact frequency of the wideband voltage-tuned microwave local oscillator used for frequency conversion and because of the local oscillator's inherent instability.

A new microwave spectrum analyzer, Model 8566A (Fig. 1), incorporates new solutions to these problems, bringing to the microwave region the performance formerly associated only with high-grade spectrum analyzers designed for lower frequencies. The performance of the new Model 8566A is such that it is possible, for example, to measure the frequency of a 20-GHz signal with ±32-Hz accuracy. Furthermore, the new analyzer's stability and low phase noise allow the use of a 10-Hz resolution bandwidth throughout its extremely wide 100-Hz-to-22-GHz input range, enabling for the first time the resolution of close-in, power-line related sidebands on a microwave signal (Fig. 2).

To achieve frequency accuracy in this instrument, frequency synthesis techniques are used to establish the start frequency of a sweep very accurately, and a "lock-and-roll" technique then allows smooth continuous tuning across the sweep. A self-calibrating discriminator-stabilized swept oscillator technique, described in the article beginning on page 13, obtains the low phase noise and residual FM required to meet the stringent requirements of narrowband frequency sweeping (spans as narrow as 100 Hz) while also allowing the instrument to make very wide sweeps (up to 20

Fig. 1. *The Hewlett-Packard Model 8566A Spectrum Analyzer has a frequency range of 100 Hz to 22 GHz and an amplitude range of −134 dBm to +30 dBm. The 10-Hz minimum resolution bandwidth is useful to 22 GHz. For the majority of measurements, only the dark-colored keys need be used to select the frequency span and amplitude reference level and to set the tunable markers for frequency and amplitude readout. The other parameters are selected automatically and all control settings are displayed on the CRT.*

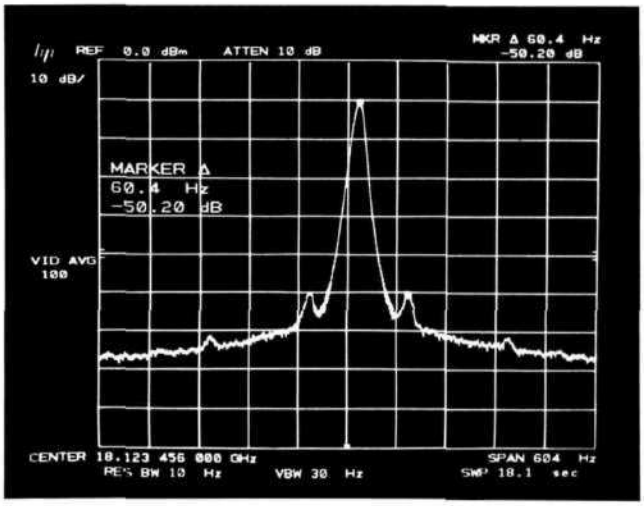


Fig. 2. *Spectrum display of an 18-GHz signal made by the Model 8566A Spectrum Analyzer with a 10-Hz resolution bandwidth across a frequency scan of 600 Hz. The Δ marker (rightmost brightened dot) is positioned on a 60-Hz line-related sideband that is 50 dB below the carrier. Until now, this high degree of resolution, made possible by the low phase noise and 10-Hz resolution bandwidth of the 8566A, had not been available in microwave spectrum analyzers.*

GHz wide).

Unprecedented flatness in the new analyzer's frequency response was made possible by extensive use of microcircuit technology. Ideally, front-end circuit elements should be small compared to the wavelength of the input signal so the elements will behave as lumped circuit constants even at the highest frequencies encountered. The front end of a microwave spectrum analyzer, however, has to process signals of extremely short wavelengths, e.g., 15 mm at 20 GHz. Consequently, the input preselector filter of the new analyzer was integrated with the input mixer to achieve the response flatness desired, as will be described in the article beginning on page 8.

Microprocessor Control

Similar in organization to the 100-Hz-to-1500-MHz Model 8568A Spectrum Analyzer,[1] the new Model 8566A uses microprocessors for instrument control, for display data processing, and for operation on the HP interface bus (HP-IB).* A block diagram is shown in Fig. 3. As is commonly done in spectrum analyzers, the input signal goes through a series of frequency conversions to a fixed IF frequency (21.4 MHz) where the resolution bandwidth filtering takes place. The first local oscillator is swept so the individual frequency components of the input signal are heterodyned one by one into the 21.4 MHz slot for subsequent detection and display on the frequency scale of the CRT display.

The detected video signal is sampled and stored digitally with 1000×1000-point resolution for repetitive read-out by the display circuits. The video detector includes a "rosenfell" detector[2] that enables the display circuits to present a more accurate reconstruction of the video signal than is usually achieved with digital storage of spectrum analyzer signal traces.

Microprocessor control led to the implementation of operating features that are especially useful for a microwave analyzer. For example, the instrument can sweep over its entire 100-Hz-to-22-GHz frequency range without requiring the operator to be concerned with the four bandswitching points. For all frequency spans, the manual tuning control requires the same number of turns (1½) to move a signal across the display, an especially convenient feature in view of the wide range of frequency spans offered, from 100 Hz to 20 GHz. A signal track mode maintains a drifting signal identified by a marker at center screen by automatically retuning the analyzer. The frequency of the drifting signal is also displayed continuously.

Digital storage of a spectrum enables flicker-free viewing even though the instrument may sweep slowly. Digital storage also allows comparisons of two spectra and other data manipulations, such as normalizing a trace (subtracting errors stored during a calibration sweep). With the analyzer's HP-IB port connected to a desktop computer, the stored data can be reformatted in the computer and then displayed in the new format on the analyzer's CRT.

Under computer control, the analyzer can be used for complicated or time-consuming measurement routines with minimum operator involvement. It can also interact through the HP-IB with other instrumentation such as plotters and signal generators. It is thus ideally suited for automatic component test of amplifiers, mixers, and oscillators, as well as for such tasks as spectrum searches for RFI or unknown, intermittent signals. Every front-panel function can be programmed through the HP-IB and additional functions are provided to simplify the data handling between a controller and the instrument.

Quiet, Accurate, Local Oscillator

The capability of resolving low-level signals that lie close to a large signal is a function of the phase noise characteristics of the local oscillator as well as of the analyzer's filter bandwidth and shape factor. The local oscillator's phase-noise characteristics will be impressed on any signal on the spectrum display and could mask the smaller of two signals. To minimize this effect, the synthesized local oscillator of the 8566A was designed for outstanding spectral purity. For example, noise sidebands for input signals up to 5.8 GHz are more than 80 dB below the signal carrier in a 10-Hz bandwidth at 320 Hz offset from the carrier.

Like other microwave oscillators, the new analyzer's local oscillator has a limited tuning range (2-6 GHz) whereas frequency spans of several gigahertz are desirable for microwave spectrum analysis. In addition, the YIG-tuned oscillator commonly used as a local oscillator in microwave spectrum analyzers can be set to a given frequency with an accuracy no better than a few megahertz because of nonlinearities in the magnetic tuning structure. However, for steady-state outputs, much better accuracy can be achieved by phase-locking the oscillator to a known harmonic of a stable lower-frequency reference. This principle is applied to the sweeping local oscillator of the 8566A by initially phase-locking the oscillator to a start frequency accurately synthesized from a stable reference,

*Hewlett-Packard's implementation of ANSI/IEEE 488-1978.

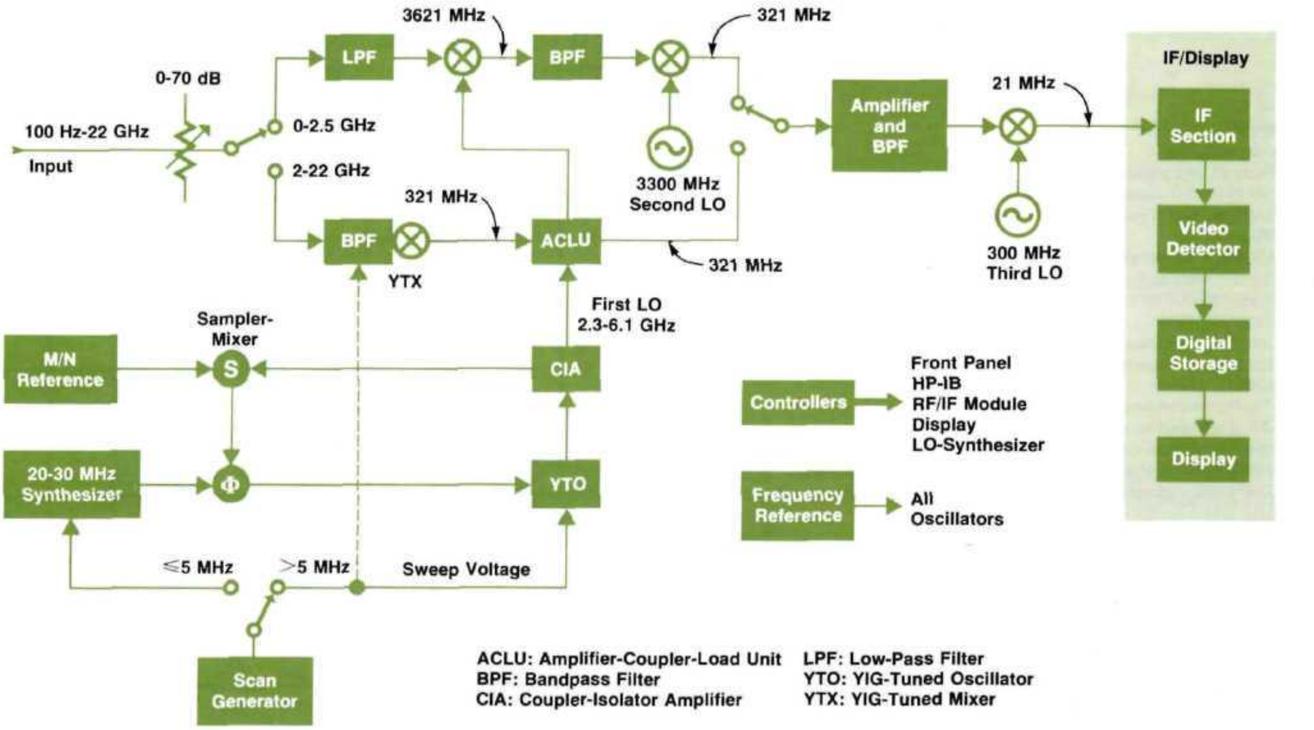


Fig. 3. *Simplified block diagram of the Model 8566A Spectrum Analyzer. Detailed descriptions of the front end and the synthesized first local oscillator will be found in the articles that follow.*

then opening the phase-lock loop while retaining the corrected tuning voltage on a storage capacitor, and adding a precision linear ramp to the tuning voltage to sweep the frequency.

If the selected frequency span exceeds the tuning range of the YIG oscillator, the sweep stops at the end of the tuning range, a new start frequency at the lower edge of the oscillator's range is synthesized, and the sweep continues using the oscillator's next higher harmonic. Known as the "lock-and-roll" technique, this occurs under microprocessor control without any intervention on the part of the operator.

The lock-and-roll approach is used in the 8566A for frequency spans as narrow as 100 Hz and as wide as 22 GHz—over eight decades of span width. An internal, oven-controlled, 10-MHz, crystal frequency standard with a stability of one part in 10^9 per day establishes the basic accuracy of the synthesizer. The start frequencies are synthesized with a resolution of one hertz, and the sweep contributes an error of less than 1% of span width to the resulting stop frequencies, e.g., ± 1 Hz for the 100-Hz span. This level of accuracy has not been available for microwave spectrum analysis in the past.

The exceptional stability of the local oscillator allows the use of a 10-Hz resolution bandwidth. Microwave signals, however, often do not have the frequency stability to permit measurements in a 10-Hz bandwidth. Therefore, a wide range of bandwidths is provided in the 8566A so an optimum bandwidth is available for any measuring situation. These range up to 3 MHz in a 1 - 3 - 10 sequence. With this large number of bandwidths to choose from, sweep time can be minimized since sweep time is generally constrained by the selected bandwidth (the sweep time must be slow enough to allow the resolution bandwidth filters to respond to changes in signal level, the allowed time being inversely proportional to the square of the bandwidth).

Sensitivity, Flatness, Distortion

In a broadband receiver such as a spectrum analyzer, a compromise must inevitably be made between sensitivity, frequency-response flatness, and distortion caused by overloading the input mixer. For the 2-to-22-GHz input frequency range, a YIG-tuned preselecting filter in the 8566A improves distortion performance by attenuating large signals lying outside the frequency range of interest. However, the presence of such a filter can degrade flatness because of standing waves that develop as a result of impedance mismatches between the filter and the input mixer. In the past, the solution to this problem was to insert attenuation between the two components. This, however, caused a loss of sensitivity.

Another way to avoid this problem is to keep the length of transmission line between the filter and the mixer much shorter than one-quarter wavelength at the highest frequency of interest, i.e., much less than 3 mm at 22 GHz. This is done in the 8566A by integrating the filter and mixer in a single unit using microcircuit technology. As a result, flatness is better than ± 2.2 dB up to 20 GHz and no lossy padding is required. Although the lower conversion efficiency of harmonic mixing degrades sensitivity, sensitivity at 22 GHz is still better than -114 dBm (the highest harmonic used is the fourth).

For the frequency range from 100 Hz to 2.5 GHz, the instrument uses a broadband single-balanced mixer preceded by a low-pass filter. This gives a sensitivity of -134

SUMMARY SPECIFICATIONS

HP Model 8566A Spectrum Analyzer

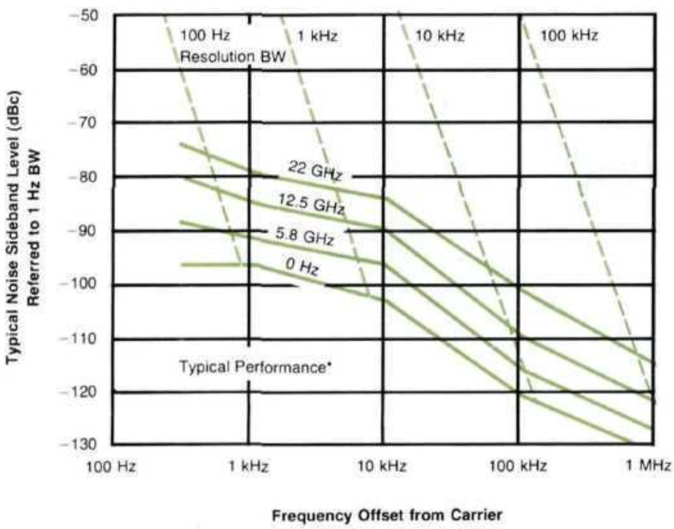
Frequency

RANGE:
100 Hz to 22 GHz.
2 GHz to 22 GHz preselected.
100 Hz to 22 GHz sweep in SINGLE SWEEP mode.

RESOLUTION: 3-dB bandwidths of 10 Hz to 3 MHz in a 1-3-10 sequence.

ACCURACY
FREQUENCY REFERENCE ERROR (AGING RATE):
$<1\times10^{-9}$/day and $<2\times10^{-7}$/year.
CENTER FREQUENCY:
Spans ≤5 MHz: ±(2% of frequency span + 10 Hz + frequency reference error × center frequency).
Zero Span: ± frequency reference error × center frequency.
FREQUENCY SPAN:
±1% of indicated separation for spans ≤5 MHz.
±3% of indicated separation for spans >5 MHz.
MARKERS: Same as Center Frequency.

SPECTRAL PURITY: Noise sidebands ≤80 dB below signal, 320 Hz offset, 100 Hz to 5.8 GHz tuned frequency (fundamental mixing) with 10-Hz resolution bandwidth.



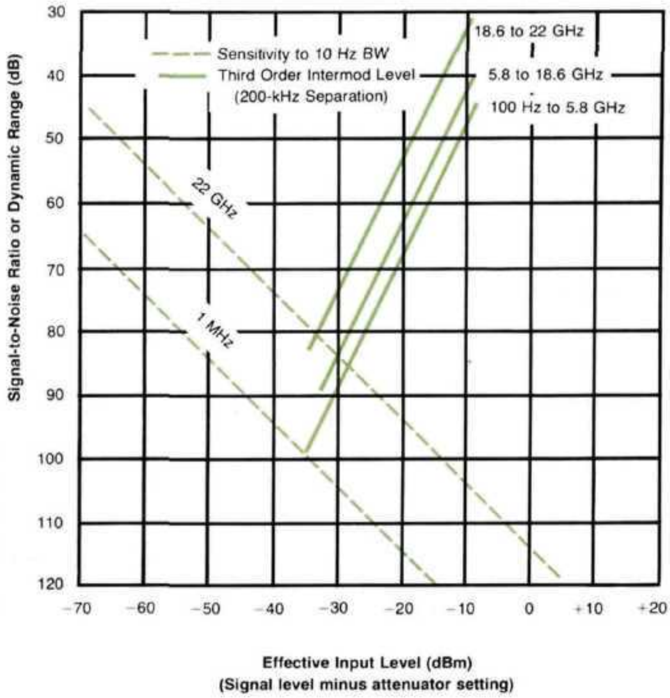
Typical Single Sideband Noise Normalized to 1 Hz BW.*

Amplitude

RANGE: −134 dBm to +30 dBm (32 nV to 7.07 volts, 50Ω). Displayed 10, 5, 2, or 1 dB/division or linear on a 10-division linear scale.

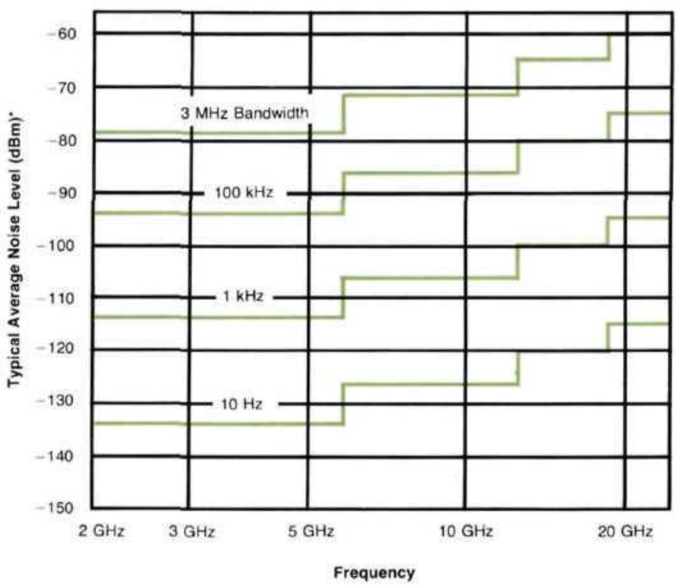
DYNAMIC RANGE
SECOND-HARMONIC DISTORTION:
<−80 dBc, 100 Hz to 700 MHz } −40 dBm mixer level
<−70 dBc, 700 MHz to 2.5 GHz } −40 dBm mixer level
<−100 dBc, 2.0 GHz to 22 GHz −10 dBm mixer level
THIRD-ORDER INTERMODULATION DISTORTION:
THIRD-ORDER IM INTERCEPT:
Specified { +7 dBm, 100 Hz to 5.8 GHz. +5 dBm, 5.8 GHz to 18.6 GHz.
Typical* { +12 dBm, 100 Hz to 5.8 GHz. +10 dBm, 5.8 GHz to 18.6 GHz. +5 dBm, 18.6 to 22 GHz.
IMAGE AND MULTIPLE RESPONSES: <−70 dBc, 100 Hz to 18.6 GHz.
AVERAGE NOISE LEVEL (SENSITIVITY): For 10-Hz resolution BW:
−134 dBm, 1 MHz to 2.5 GHz.
−132 dBm, 2 GHz to 5.8 GHz.
−125 dBm, 5.8 GHz to 12.5 GHz.
−119 dBm, 12.5 GHz to 18.6 GHz.
−114 dBm, 18.6 GHz to 22 GHz.

ACCURACY: Measurement accuracy is a function of technique. The following sources for uncertainty can be summed to determine achievable accuracy (at constant ambient temperature, assuming the error correction function and preselector peak have been used, and avoiding unnecessary control changes between calibration and measurement).
CALIBRATOR UNCERTAINTY: ±0.3 dB.



Typical Optimum Dynamic Range.*

FREQUENCY RESPONSE UNCERTAINTY (FLATNESS):
±0.6 dB, 100 Hz to 2.5 GHz.
±1.7 dB, 2.5 GHz to 12.5 GHz.
±2.2 dB, 12.5 GHz to 20 GHz.
±3 dB, 20 GHz to 22 GHz.
COMPARISON UNCERTAINTY (resulting from one of the following techniques for comparing the unknown signal with the calibrator):
REPOSITIONING SIGNAL TO CALIBRATION LEVEL: ±1.2 dB*.
USING MARKER: ±3.2 dB*.



Average Displayed Noise Level

*Typical, non-warranted performance parameters useful in applying the instrument.

(continued)

Sweep

SWEEP TIME: 20 ms full span to 1500 s full span. Zero Frequency Span, 1 μs full sweep (of display) to 1500 s full sweep.

Input

RF INPUT: 100 Hz to 22 GHz, 50Ω dc coupled. Precision Type N female. Diode limiter 100 Hz to 2.5 GHz. Preselected 2.0 to 22 GHz.

MAXIMUM INPUT LEVEL

AC: +30 dBm (1 watt) continuous power.

DC: <100 mA current damage level.

ATTENUATOR: 70-dB range in 10-dB steps.

Outputs

FIRST LOCAL OSCILLATOR: 2.3 GHz to 6.1 GHz; >+5 dBm into 50Ω.

DISPLAY: X, Y, and Z outputs for auxiliary CRT display.

RECORDER: Horizontal sweep output (X), video output (Y), and penlift/blanking output (Z) to drive an X-Y recorder.

Instrument State Storage

Up to 6 sets of user defined control settings may be saved and recalled.

Remote Operation

All analyzer control settings (with the exception of video trigger level, focus, align, intensity, frequency zero, amplitude cal, and line power) may be programmed via the Hewlett-Packard Interface Bus (HP-IB).

General

ENVIRONMENTAL: Operation 0°C to 55°C. <95% relative humidity, 0°C to 40°C.

EMI: Conducted and radiated interference is within the requirements of CE 03 and RE 02 of MIL STD 461A, and within the requirements of VDE 0871 and CISPR publication 11.

WARM-UP TIME

OPERATION: Requires 30 minute warm-up from cold start, 0° to 55°C. Internal temperature equilibrium is reached after 2 hr. warm-up at stable outside temperature.

FREQUENCY REFERENCE: Aging rate attained after 24 hr. warm-up from cold start at 25°C. Frequency is within 1×10^{-8} of final stabilized frequency within 30 minutes.

POWER REQUIREMENTS: 50 to 60 Hz; 100, 120, 220 or 240 volts (+5%, −10%); approximately 650 VA (40 VA in standby). 400 Hz operation available.

WEIGHT: 50 kg (112 lb).

DIMENSIONS: 280 mm H × 426 mm W × 600 mm D (11 × 16¾ × 23½ in).

MANUFACTURING DIVISION: SANTA ROSA DIVISION
1400 Fountain Grove Parkway
Santa Rosa, California 95404 U.S.A.

dBm, a flatness of ±0.6 dB, and a third-order intermodulation intercept point of +7 dBm for closely-spaced signals. The input signal is automatically switched to the appropriate mixer by the microprocessor according to the frequency span selected.

The combination of high-performance analog circuit components and internal data handling by microprocessors has resulted in a spectrum analyzer that sets new standards for measurement capability and user convenience in the microwave frequency range.

Acknowledgments

The development of the Model 8566A Spectrum Analyzer extended over several years and involved a great number of people. Many different approaches to the microwave front end, the LO synthesizer, and the measurement display were investigated. Some proved to be impractical, others became key elements for other instruments, such as the 86290A Sweeper and its YIG-tuned multiplier, and the signal synthesis scheme used in the 8672A Synthesized Signal Generator.

Much help with the system definition, plus encouragement and support despite lengthening schedules, came from Rit Keiter, Santa Rosa engineering lab manager (now general manager for the Santa Rosa spectrum analyzer operations). His ideas, together with those of Dave Eng, industrial design manager, and of many others, formed the basis for an easy-to-interface front panel that combined the feel of a knob-controlled analog instrument with the precision of a keyboard-controlled digital machine.

In addition to those mentioned in the following articles, recognition should go to the following for their contributions to the 8566A: Irv Hawley, spectrum analyzer section manager (now R&D manager for the network measurement operation at Santa Rosa), who always maintained the flow of resources and, with critical questions, kept the project on track; Ron Trelle, for the overall product design with particular attention to cooling, shielding and vibration control; Art Upham, for the design of the third converter, YTX driver, and miscellaneous circuitry whenever the need arose; Rich Pope, for the second LO phaselock loop and electronic test tools for the YTX production; Rex Bullinger, for the HP-IB interface; Lynn Wheelwright, for his support in developing the digital control system, and Dee Humphreys and his design team for the development of the IF/display section.

Numerous other people in many areas of Hewlett-Packard contributed to the project. Many of their contributions, like GaAs FETs and YIG spheres, are buried deep inside the instrument and are barely visible even on a detailed block diagram. Yet it was this joint effort—true teamwork—that made the 8566A project succeed.

References

1. S.N. Holdaway and M.D. Humpherys, "The Next Generation RF Spectrum Analyzer," Hewlett-Packard Journal, June 1978.

2. S.N. Holdaway, D.H. Molinari, S.H. Linkwitz, and M.J. Neering, "Signal Processing in the Model 8568A Spectrum Analyzer," Hewlett-Packard Journal, June 1978.

Siegfried H. Linkwitz

Born in Bad Oeynhausen, Germany, Siegfried Linkwitz is a 1961 graduate (Diplom Ingenieur) of Darmstadt University, Germany. Siegfried joined HP shortly after graduation and has since worked on vector voltmeters, signal generators, sweepers, and spectrum analyzers (including the 8566A, for which he was program manager). A resident of Santa Rosa, California, Siegfried is married and has one son, 14, and one daughter, 16. In his spare time, he enjoys windsurfing, skiing, and designing hi-fi equipment. A member of the Creative Initiative Foundation, he also leads groups on fulfilled living.

Broadband Input Mixers for a Microwave Spectrum Analyzer

by John C. Lamy and Frank K. David

TO PROVIDE THE USER with state-of-the-art spectrum analyzer performance over a broad range of frequencies, a dual front-end approach was chosen for the Model 8566A Spectrum Analyzer. The input section has two independent heterodyne conversion channels: a 2-to-22 GHz YIG-preselected harmonic mixer chain, and a 100-Hz-to-2.5-GHz up-down converter chain. As indicated in Fig. 1, these alternative paths are selected automatically by a mechanical relay that is under control of a microprocessor.

The dual approach was chosen because present-day YIG technology, which provides the analyzer's spurious-free microwave performance, is not applicable at lower frequencies. The use of two techniques enables top performance over the full 100-Hz-to-22-GHz operating range of the instrument.

2-to-22-GHz Band

The 2-to-22-GHz band on the 8566A is a modern implementation of a classical concept: a multiconversion heterodyne receiver with a broadband input mixer. Much of the system's performance depends on the input mixer, which converts the input signal frequency to a fixed IF. When the frequency of the sweeping local oscillator (YTO) equals the input signal plus or minus the first IF, a response is generated in the IF detector. The response appears as a pulse on the display, which sweeps in synchronism with the LO as the LO sweeps the signal past the IF.

The broadband front-end mixer traditionally reduces to an extremely simple piece of hardware: a single diode in a

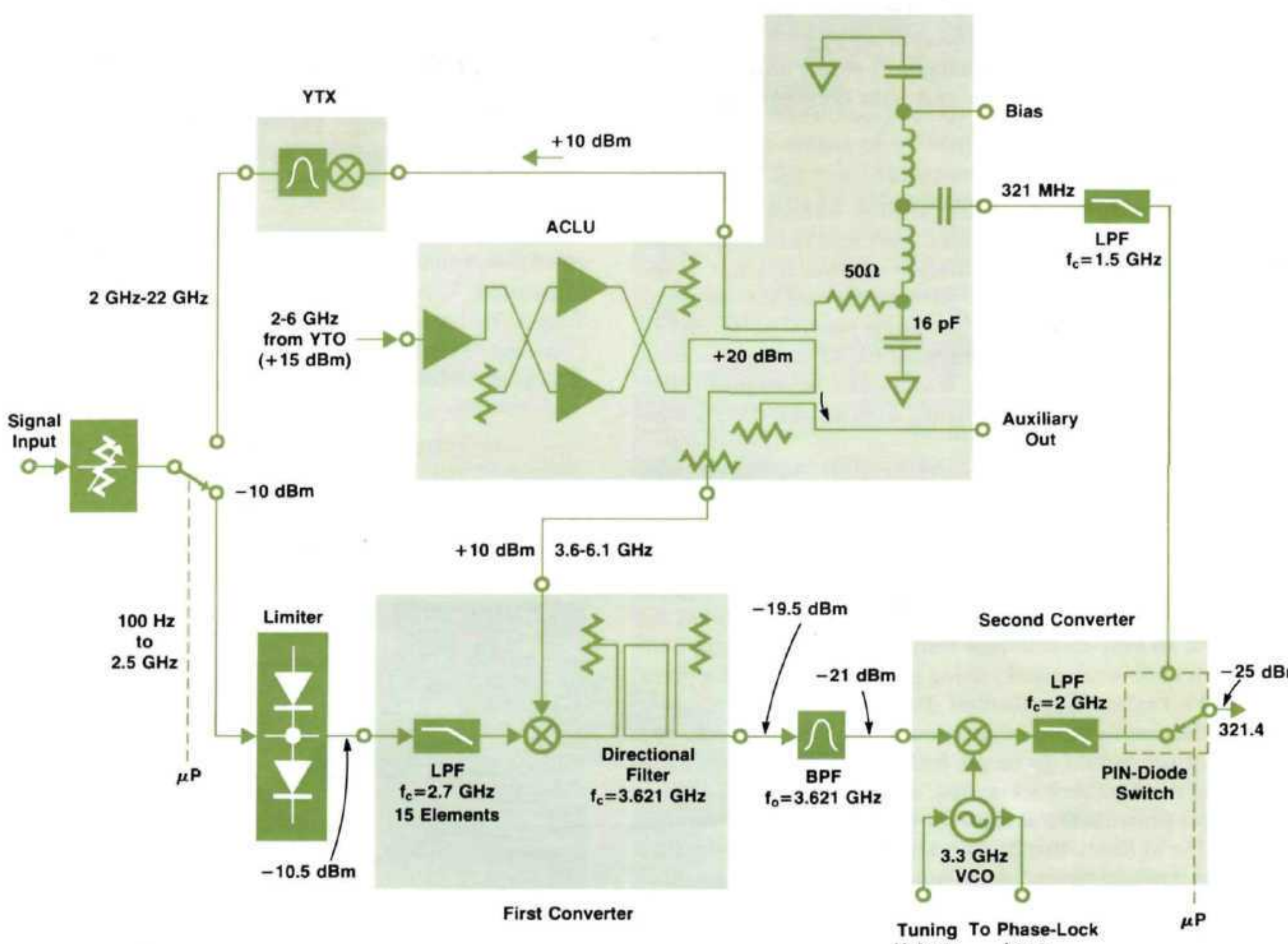


Fig. 1. *Block diagram of the input section of the Model 8566A Spectrum Analyzer. The control microprocessor selects the appropriate signal path according to the frequency span selected.*

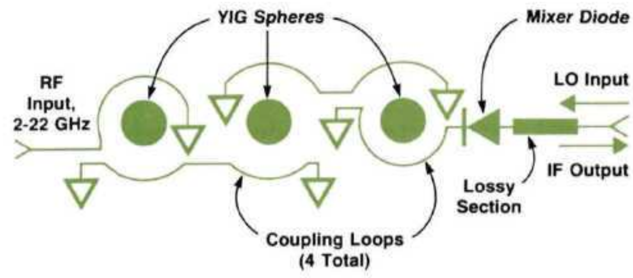


Fig. 2. *Schematic representation of the YIG-tuned filter and mixer (YTX) used in the 8566A. Although this shows all the coupling loops in the same plane, the two loops for each YIG sphere cross at an angle so coupling occurs only at the resonant frequency of the YIG sphere, which is a function of the dc magnetic field strength.*

structure made small compared to the wavelength of the signal, LO, and IF frequencies.

Simple? Yes, except that there is a great deal more involved than appears on the surface. Four kinds of unwanted responses are generated along with the wanted response: those resulting from image frequencies, multiples (frequencies that beat with harmonics of the LO), out-of-band signals, and signal harmonics. The traditional solution to the problem is to precede the input mixer with a tunable narrow-band filter that tracks the frequency tuning of the analyzer. This suppresses the unwanted responses but trades off performance because of the impedance mismatch between the preselector filter and the mixer. The mismatch causes standing waves to develop on the transmission line between them, introducing substantial variations in the analyzer's frequency response. The solution to that problem has been to insert attenuation between the preselector and the mixer, which results in some loss of sensitivity.

In short, the preselector eliminates unwanted responses but forces a trade-off between flatness and sensitivity, two key performance characteristics of the analyzer.

The YIG-Tuned Mixer

This trade-off was eliminated in the 8566A by integrating the mixer diode with the YIG-tuned preselector such that there is essentially zero line length between them. The complete structure is called the YIG-tuned mixer, or YTX.

Operation of the high-band front-end system can be explained with reference to Fig. 1. The 2-to-6-GHz swept LO signal is applied to the ACLU (amplifier-coupler-load unit) where it is amplified and leveled by a saturating FET amplifier. It is then coupled to the main line where it travels to the YTX diode, switching the diode on and off. The on-off ratio (conduction angle) is controlled by the dc bias applied through the bias port of the ACLU. Different conduction angles are chosen for the various harmonics of the LO, enabling operation to 22 GHz.

Three YIG spheres with their coupling loops form the YTX bandpass filter, as shown in Fig. 2. The filter is tuned to the analyzer's instantaneous frequency by varying the applied magnetic field. An input signal passing through the filter is alternately transmitted and reflected back by the mixer diode as the diode switches on and off. Frequencies above 2 GHz in the transmitted part terminate in the 50Ω resistor of the ACLU. The 321-MHz IF generated at the mixer diode passes through the ACLU tuned circuit and into the IF section. Because of the preselector, it is not necessary to use a relatively high first IF to space the image frequency far enough from the desired signal to make it easy to deal with, another advantage of the preselector approach.

The YTX is designed to mix on the first through fourth harmonics of the LO. The control microprocessor selects the appropriate harmonic so the harmonic bandwitching is transparent to the user.

Design Considerations

Although the use of microcircuitry solved the major problems, there were others that had to be dealt with. The first was how to guarantee 70-dB rejection of unwanted signals. A rule of thumb in YIG filter design is that each resonator sphere can contribute about 25 dB of stopband isolation, so three spheres seemed to be about right. But then, how do you guarantee that all three spheres are subject to the same magnetic field intensity as the intensity is varied over a 10-to-1 range? The error can be no more than 1 part in 2000, equivalent to a tuning error of 10 MHz, or one-half the preselector 3-dB bandwidth at a center frequency of 20 GHz. Since the H field is controlled by the 1.35-mm (0.054-inch) gap in the electromagnet, a 1/2000 error is less than 0.75 μm (30 millionths of an inch) of pole-face nonparallelism, virtually impossible to hold in production since it would be the sum of machining tolerances in three relatively large pieces of steel.

The usual way to maintain acceptable parallelism in a multi-sphere structure has been to shim the pieces with tape at strategic points. The way used for the YTX is to intentionally grind the pole faces with a slight tilt and then,

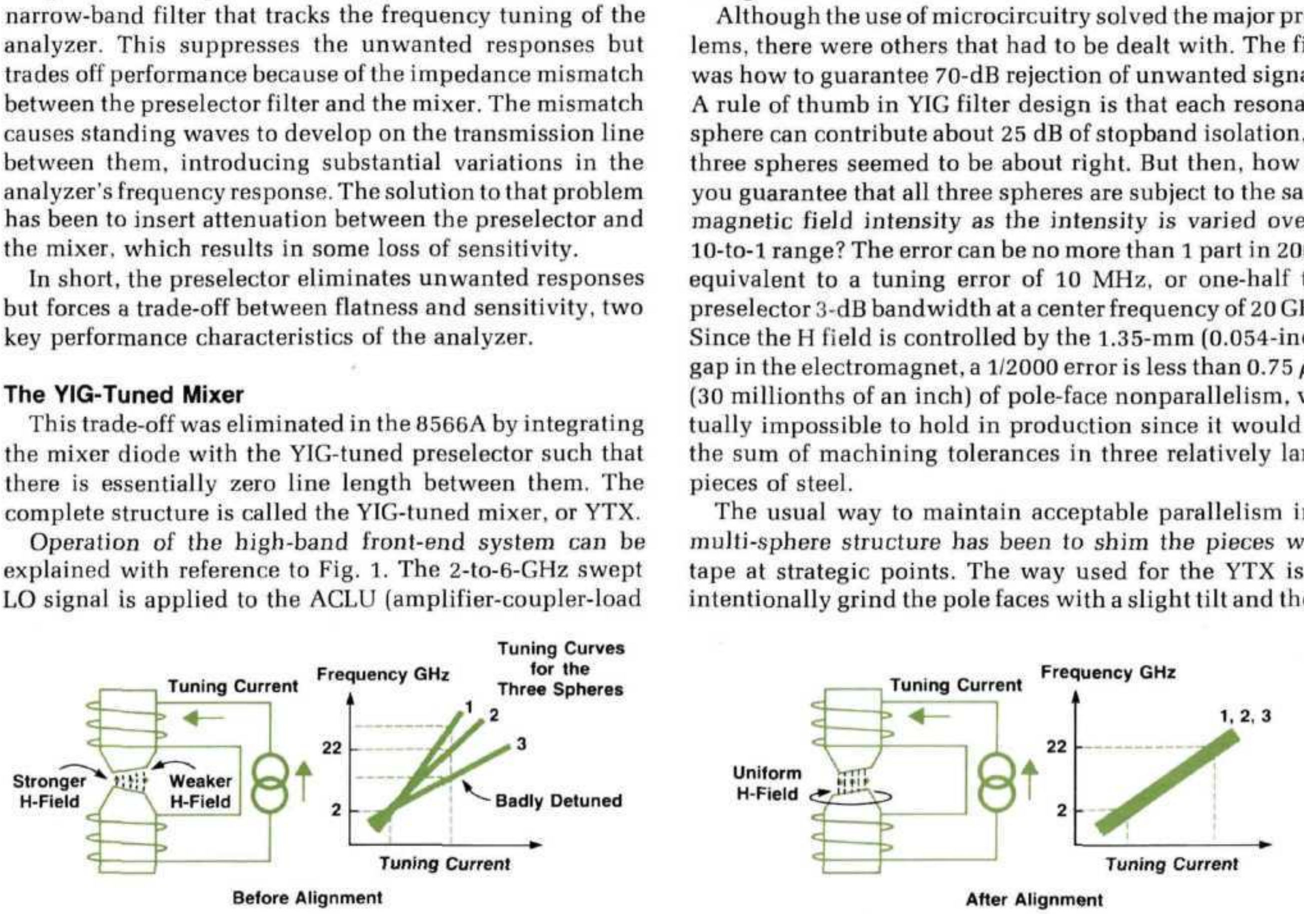


Fig. 3. *Misalignment of the tuning magnet pole faces, shown exaggerated here, causes the three YIG spheres to have different tuning characteristics. Rotating the pole pieces to bring the tilted faces into parallel alignment achieves identical tuning characteristics for the three spheres.*

Precision Assembly of a YIG-Tuned Mixer

Examination of the YIG-tuned mixer for the Model 8566A Spectrum Analyzer indicated that it could not be manufactured by traditional manufacturing methods. To maintain the designer's intent while achieving efficient fixturing it was concluded that the engineer who did the mechanical design of the YTX should also develop the tooling for manufacturing. In this way, it would be possible to avoid the problems that occur when a manufacturing engineer develops concepts that differ from the design engineer's. This placed a heavy burden on the design engineer as the assembly fixturing was constantly being evaluated and changed, but it paid off in that the final design was nicely manufacturable.

A photograph of the interior of the YTX is shown in Fig. 1. The main part is the circular piece with the three holes where the YIG spheres are located. This part is molded of Fiberite and then gold-plated. The D-shaped part surrounding it is made of copper for good heat conduction. The small microcircuit at the upper right corner is a heater and the device at the lower right corner where the flying lead connects is a thermistor. The heater maintains the YIG spheres at a slightly elevated temperature so their temperature can be held constant during fluctuations in the ambient.

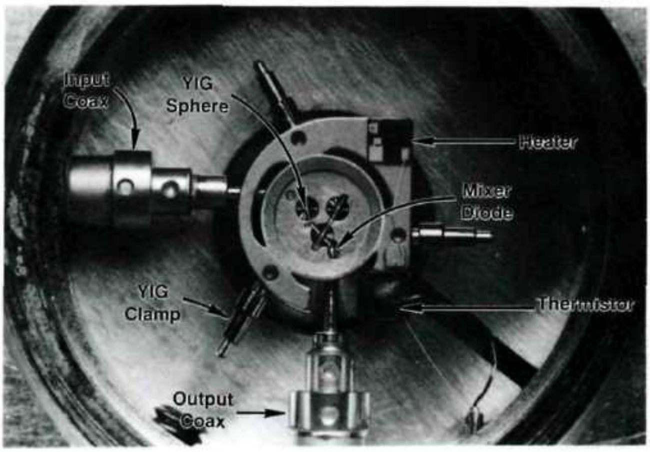


Fig. 1. *Internal view of the YIG-tuned mixer (YTX). The circular part that has the three holes for the YIG spheres is only 9 mm in diameter.*

The thermal circuit to the three YIG spheres is completed by metal rods that are held in cylindrical spring-loaded clamp circuits, shown protruding from the D ring. The rod ends are accessible through plug-holes in the external magnet structure so, with the magnets in place, the rods can be rotated and moved axially to find the position for best temperature compensation. This adjustability is obtained while maintaining good thermal contact. The clamps are firm enough to obviate the need for any additional clamping device or cement to prevent movement during shock and vibration.

The YIG sphere coupling loops are just visible in the photo. Only 50 μm (0.002 inch) thick, these have to be dimensioned and positioned very precisely. For example, the distance between a loop and a YIG sphere must not vary by more than 25 μm (0.001 inch). It would be very difficult for even a highly skilled person to form and position these loops with the precision required. This problem was solved by using state-of-the-art chemical milling techniques to mill the loops from beryllium copper, obtaining sturdy parts with a tolerance of ± 5 μm (0.0002 inch) and excellent repeatability. These are checked in the assembly area with an optical comparator just before assembly.

Molding on the Production Line

Molding of the dielectric for the coaxial sections presented some special problems. Injection molding was considered but rejected because of the fragility of the coax parts (there is a 665-to-172-μm center conductor transition) and the fear that the air-filled microballoons required to obtain the correct dielectric constant could separate from the resin and fillers when exposed to restricted flow under high pressure. Tooling was developed that would allow the center conductor to be inserted and positioned accurately after the outer section is filled with the dielectric in a plastic state. The assembled unit is then cured in a pressure vessel to minimize any expansion of trapped air due to the elevated curing temperature. After the curing, the lead lengths and excess epoxy are trimmed, and then the coax assembly is tested with a time-domain reflectometer to assure its acceptability. All of these operations are performed in the final assembly area to assure maintenance of the tolerances desired.

Special Soldering Techniques

A key element in the mounting of the very small parts was the use of solder cream (solder granules suspended in flux). It is used as a liquid preform that enabled us to solder parts that could not be soldered by conventional means.

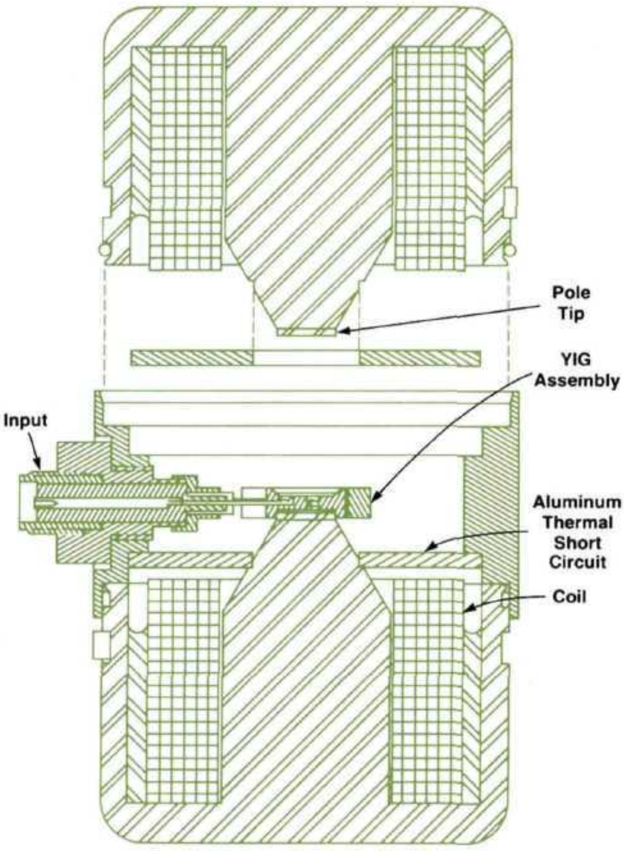


Fig. 2. *Cross-section of the YTX.*

It became apparent that soldering the entire structure at one time was impractical, so some means had to be found that would allow soldering without reflowing previously-made solder joints. The solution was to use solders with higher melting temperatures for the early operations. As a result, it is possible to solder the coax sections to the assembled central unit during final operations, obtaining a well-bonded grounding structure.

Thermal Design of the Magnet

A major objective was to provide a constant, uniform magnetic flux under all environmental conditions. A cross-section of the magnetic structure with the YIG-sphere assembly in place is shown in Fig. 2. The pole tips of the magnet are made of a material that has a different thermal expansion ratio from the main body so the magnet gap remains constant despite changes in the ambient temperature.

However, some short-term differential heating of the magnet occurs whenever the tuning coil current is changed with the greatest change occurring at the high end of the YIG tuning range where the current is proportionately high. Because of the relatively poor thermal *conductivity of the nickel-iron alloys used, a temperature gradient* would exist between the poles and the main body since the outside surface dissipates heat more readily than the pole faces. This results in changes in the gap spacing, hence in the YIG tuning. This problem was minimized by installing aluminum discs between the pole faces and the outer ring as shown in Fig. 2. These serve as thermal short circuits between the pole faces and the magnet body reducing the differential heating by a factor of 10 (a single slot cut radially in each disc prevents the formation of eddy currents). Residual short-term temperature differentials are compensated for by the microprocessor control system (see box, page 15).

Acknowledgments

Special thanks are due Earl Heldt of the Stanford Park Division for the original mechanical design. Dick Lyon made major contributions to the design. Pete Planting developed the microballoon material and provided technical support for the soldering processes.

-Lee Olmstead

with the structure assembled, use a desktop computer test program that indicates to the test technician how much the pole pieces should be rotated with respect to each other around their common axis so the tilt in one pole face will match the tilt in the other (Fig. 3). The computer controls test instruments that make frequency response curves as the filter is tuned throughout the 2-to-22-GHz range. Misalignment of the pole faces causes the frequency response curves to have three humps. The technician measures the frequency separation between adjacent humps and enters this value into the computer, which then calculates the amount of rotation required to bring the pole faces into alignment. Usually, one pass is sufficient to achieve virtually perfect parallelism. This procedure takes less than 15 minutes.

A second problem concerned the impedance match looking from the diode toward the ACLU load where the input signals would normally terminate. To meet our flatness goals, the return loss had to exceed 20 dB over the entire 2-to-22-GHz range. Although this can be done up to 6 GHz, it is unrealistic to expect this performance of the SMA connector pairs used for the connections, not to mention the transitions from coax to microstrip and so on. To surmount this problem, the YTX uses polyiron-loaded dielectric in a specially designed molded coax assembly (Fig. 2). The polyiron is lossless at the IF and LO frequencies, but above 6 GHz it becomes lossy and acts as a good termination to signals above 12 GHz, giving outstanding flatness.

Amplifier/Coupler/Load Unit

Of the several functions that the amplifier/coupler/load unit (ACLU) performs, its contribution to the high-band flatness is most noteworthy.

First, it provides a relatively constant LO power to the YTX's single-diode harmonic mixer. For some harmonics, the mixer's conversion efficiency is a function of the diode's conduction angle, and the conduction angle changes with changing LO power level. The FET amplifier in the ACLU operates in a saturated mode, and therefore provides gain compression that greatly reduces the YTO power variations that exist unit-to-unit and as a function of frequency and temperature. The amount of gain compression is shown in Fig. 4.

The second contribution to flatness is the isolation and match that the coupler load portion of the ACLU provides for the signal input to the YTX. Input signals in the 2-to-22-GHz range go into the YTX, mix in the diode, and then go into the ACLU. If any significant portion of this signal reflects off the ACLU's input and returns to the YTX, ripple will be produced in the YTX's conversion efficiency. This doesn't happen, due in part to the isolation from the amplifier's output provided by the coupler, and also to the termination provided by the 50Ω thin-film resistor. The 50Ω resistor is ac-coupled to ground through the 16-pF capacitor (see Fig. 1) such that it terminates frequencies of 2 GHz on up, but allows 321 MHz to exit the IF port with only small loss (about 3 dB). The 16-pF capacitor and some thin-film inductors provide a 321-MHz tuned matching network to match the nominal 50Ω IF impedance to the series combination of the 50Ω resistor and the mixing diode in the YTX (a combination of approximately 150Ω).

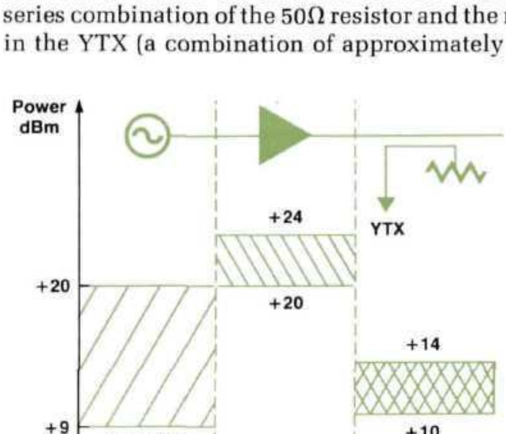


Fig. 4. *Variations in the local oscillator power supplied to the mixer diode are compressed by operating the FET amplifier in a saturated mode.*

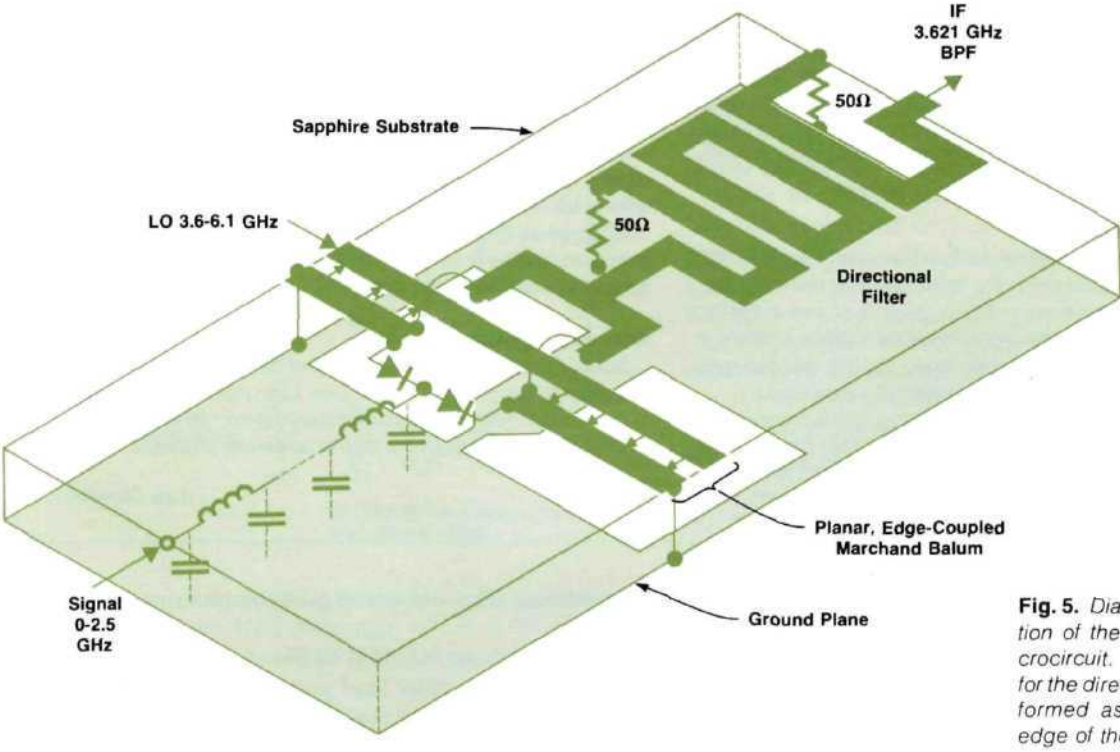


Fig. 5. *Diagrammatic representation of the second converter microcircuit. The 50Ω terminations for the directional filter are actually formed as thin films along the edge of the substrate.*

0-to-2.5-GHz Band

As shown in Fig. 1, the ACLU is the supplier of local oscillator power for both frequency bands. To remove the problem of image response in the 0-to-2.5-GHz band, a double frequency conversion is performed on the input signal. First there's an up-conversion to a fixed 3.6-GHz IF and then, after some filtering, a down-conversion to a 321-MHz IF.

Although the image response is effectively removed by this technique, the added components compound the problem of trying to fit maximum performance into minimum space. This problem was kept in check by using thin-film technology and a high degree of integration of components. Microcircuit technology is used in the assembly of the limiter, first converter, second converter, and ACLU.

Limiter

The limiter's function is to reduce the burn-out susceptibility of the first-converter mixing diodes from an overload at the input. It does this very well for CW input powers between 1 milliwatt (onset of limiting action) to 10 watts. When not limiting, the device is virtually transparent to the incoming signal (loss <1 dB, VSWR <1.25). A back-to-back diode arrangement eliminates the need for dc blocks or returns allowing the limiter to operate to very low frequencies, even dc.

First Converter

The first converter is designed to achieve a good balance of flatness, low distortion, and conversion efficiency. Because the frequency response and distortion characteristics in the 0-to-2.5-GHz range are determined primarily by the first converter's conversion flatness and distortion, these parameters were optimized in a trade-off with conversion efficiency, giving a flatness of ±0.6 dB for the input signal range of 100 Hz to 2.5 GHz. For a -30-dBm input to the first converter, the harmonic distortion and third-order distortion products are 70 dBc and 90 dBc respectively over the range of 100 Hz to 700 MHz, which covers the important communication bands.

Several specific parts of the first converter, shown in Fig. 5, contribute significantly to its good performance. The planar, edge-coupled, realization of a Marchand balun,[1] which carries the LO signal to the diodes, contributes to good flatness by isolating the signal and IF paths from the LO port. Distortion is reduced by bringing the LO in on the broadband, balanced structure. Distortion is further reduced by use of an integrated dual beam-lead diode that has closely matched parameters for each of the diodes in the pair. Finally the directional filter is crucial in achieving good flatness without excessive reduction in conversion efficiency. This filter passes the 3.6-GHz IF with minimal insertion loss while providing a good resistive termination for all of the other frequencies produced in the mixer.

Second Converter

The second converter achieves its function in a small volume with good efficiency and low phase noise.

The 3.3-GHz VCO in the second converter is a push-pull two-transistor oscillator using a microstrip horseshoe resonator.[2] Typically, microstrip resonators have comparatively low unloaded Q (around 250), and hence are not noted for particularly low phase noise when used in oscillators. By optimizing the resonator's dimensions (for maximum unloaded Q) and by running low bias current in the transistors (7 mA per transistor) the loaded Q of the oscillator was improved to the point that at 1 MHz away from the carrier, phase noise in this oscillator is as low as

John C. Lamy

With HP since 1968, John Lamy was project leader for the 85660A RF Module, the 8557A Spectrum Analyzer, and the 435A/8481A Power Meter, and was design engineer for the 8555A spectrum analyzer tuning section. John was born in Kansas City, Missouri, and received his BSEE degree in 1968 from Massachusetts Institute of Technology. Now a resident of Santa Rosa, California, he is married and has a seven-year-old son. In his spare time, John enjoys backpacking and sailing and is actively involved in the Creative Initiative Foundation.

Frank K. David

An HP employee since 1969, Frank David was project engineer for the 8566A microcircuits, including the low-range front end mixer, filters, oscillator, and limiter. Frank received his BSEE degree in 1965 from the University of California at Berkeley and his MSEE degree in 1975 from Oregon State University. A native of Sacramento, California, he is married and has two "active" children (ages six and three). Frank spends much of his leisure time landscaping his yard, doing masonry work, and camping with his family.

that in a typical YIG-tuned oscillator (−135 dBc/Hz). The close-in phase noise performance is improved by phase-locking the oscillator to a harmonic of the instrument's 100-MHz reference frequency. A microcircuit sampler serves as the phase detector. It samples the 3.3-GHz signal at the 100-MHz rate and the resulting dc is used to control the VCO.

The conversion efficiency of this converter was enhanced by taking advantage of the fact that the frequencies at all three mixer ports are fixed. These ports are tuned so as to reflect the energy of higher mixing products back to the mixing diodes where they are reconverted to the IF frequency. This raises the conversion efficiency, which initially was 25%, up to 40%.

Acknowledgments

Far more people participated in the design and introduction of these microwave components than we can adequately acknowledge here—it has been truly a large scale team effort. Two contributors stand out on the YTX: Rit Keiter for initial concept and continuing guidance, and Robert Joly for establishing the basic realization.

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2. R.E. Pratt, R.W. Austin, and A Dethlefsen, "Microcircuits for the Microwave Sweeper." Hewlett-Packard Journal, November 1970.

A Synthesized Microwave Local Oscillator with Continuous-Sweep Capability

by Larry R. Martin, Kenneth L. Lange, and Stephen T. Sparks

THE FREQUENCY ACCURACY and stability of the Model 8566A Spectrum Analyzer allows the use of a resolution bandwidth of 10 Hz anywhere within the 100-Hz-to-22-GHz range of the instrument. The instrument's accuracy, stability, and sensitivity also give it the ability to measure microwave frequencies at very low signal levels with an accuracy approaching that of the best microwave frequency counters. The low phase noise enables the analyzer to make measurements in the audio frequency range as well as the RF and microwave ranges, and in many cases allows it to measure the phase noise of microwave oscillators directly.

Performance of this caliber requires synthesizer accuracy in the local oscillators. To achieve this accuracy along with a sweeping capability, the frequency-conversion chain in the Model 8566A uses seven phase-locked loops, two of which have direct sweeping capability. A block diagram is shown in Fig. 1.

When the analyzer's input frequency is in the 0-to-2.5 GHz range where the broadband front end is used, the instrument operates as a conventional high-IF machine with a first IF of 3.6214 GHz. A 3.3-GHz source reduces this IF to a second IF of 321.4 MHz, which is further down-converted by a 300-MHz signal to a final IF of 21.4 MHz. When the input frequency is in the 2-to-22-GHz range where the preselector is used, the 321.4-MHz IF becomes the first IF and the 3.3-GHz source is disabled.

The YTO (YIG-tuned oscillator) loop shown in Fig. 1 is the final summing loop of the synthesizer system that functions as the first local oscillator in both frequency ranges. The inputs to this loop are a fixed reference frequency from the M/N loop, a fixed or swept frequency from the low-

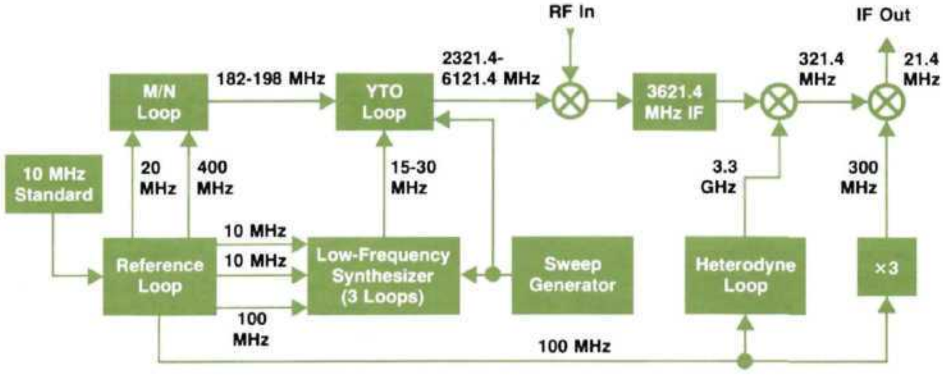

Fig. 1. *Frequency conversion chain uses seven phase-locked loops, two of which can be swept. All are referenced to a 10-MHz oven-controlled crystal frequency standard that has an aging rate of less than one part in 10^9 per day.*

frequency synthesizer, depending on the sweep range, and a sweep control from the sweep generator, also depending on the sweep range. The M/N loop provides YTO tuning steps in 10-MHz increments and the low-frequency synthesizer interpolates between the 10-MHz steps.

Before the start of a frequency scan, the YTO is tuned by a 12-bit digital-to-analog converter (DAC) such that its unlocked frequency lies 20 to 30 MHz below the Nth harmonic of the M/N loop output. The YTO output is mixed with the Nth harmonic and the resulting difference frequency is applied to a phase-frequency detector, as shown in Fig. 2. The other input to the phase-frequency detector comes from the low-frequency synthesizer. When the loop is closed the output of the phase detector is applied to the YTO tuning coils, forcing the YTO to track a combination of the M/N loop and low-frequency synthesizer frequencies.

The output frequency of the M/N loop, which ranges between 182 and 198 MHz, is defined by the equation:

$$f_{M/N} = 200 - 10M/N \text{ MHz}$$

so the Nth harmonic is 200N−10M. Thus, the frequency of the YTO loop in the locked condition is:

$$f_{YTO} = 200N - 10M - f_{lfs} \text{ MHz}$$

where f_{lfs}, the output of the low-frequency synthesizer, can be varied between 20 and 30 MHz in 1-Hz steps when setting a start frequency.

For frequency sweeps greater than 5 MHz, the M/N loop and the low-frequency synthesizer establish a precise start frequency, as determined by the main control microprocessor in response to the front-panel or HP-IB inputs. The YTO control voltage for this frequency is then retained on a capacitor, the YTO phase-lock loop is opened, and the sweep voltage is added to the YTO tuning voltage (see Fig. 2). When the frequency sweep is less than 5 MHz, the frequency sweep is generated within the low-frequency synthesizer and the YTO remains phase-locked to it and the M/N loop throughout the sweep.

10-MHz Increments

The high-purity M/N output originates in a voltage-controlled oscillator (VCO) that uses the same type of fore-

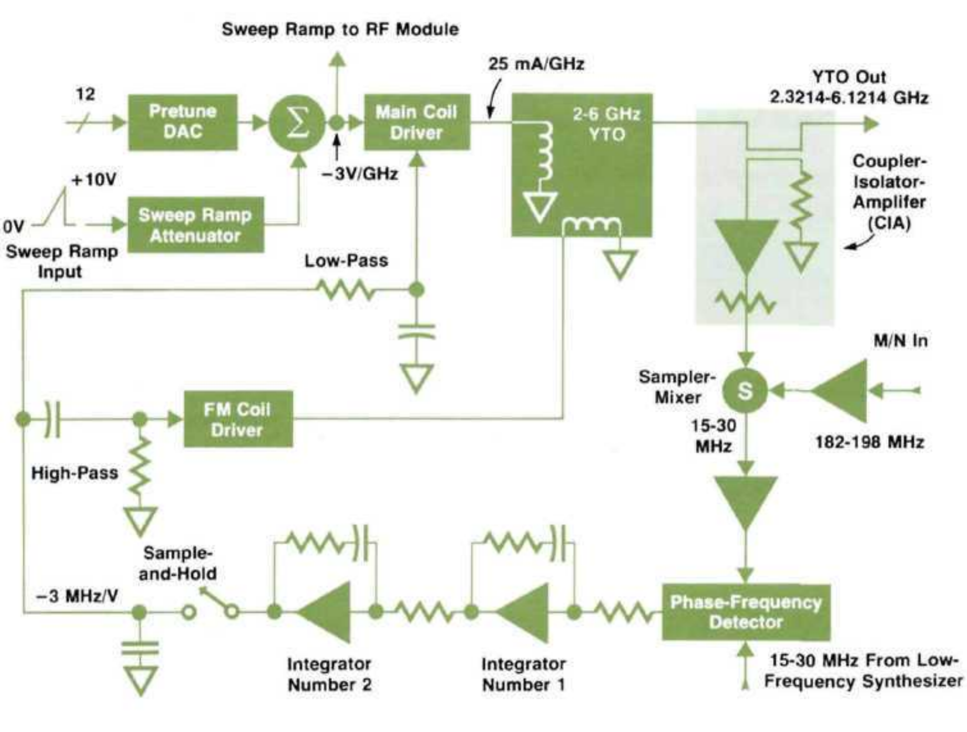

Fig. 2. *YTO (YIG-tuned oscillator) control loop locks the YTO to a combination of the M/N and low-frequency-synthesizer loop frequencies.*

Some Microprocessor Contributions to Spectrum Analyzer Performance

by Michael S. Marzalek

The new Model 8566A Spectrum Analyzer uses virtually the same digital hardware as the Model 8568A.[1] In fact, both instruments share the same IF/Display unit, which means they are identical from the 21-MHz third IF on through the video detector to the display processing circuits. The differences, from the digital point of view, are in the control of the front end and the first and second IF stages.

Because of the fundamental differences between the analog circuits in the new 8566A and those in the 8568A, there were new opportunities to use the power of microprocessor control to enhance performance. One area in which digital control contributes to the 8566A's performance is in helping the YTX (YIG-tuned mixer) track the YTO (YIG-tuned local oscillator). Through phase-locking techniques, the YTO is always tuned precisely at the start of a sweep but, by necessity, the YTX operates open loop. Delays between the actual frequency scan and the tuning ramp, hysteresis in the tuning magnets, and self-heating of the tuning coils all contribute to mistracking of the YTO and the YTX. Since it is necessary for the YTX to track the YTO within ±10 MHz (3-dB bandwidth) in spite of step changes as wide as 22 GHz, methods of compensating for mistuning of the YTX had to be developed.

Sweep delays are equalized by resistor-capacitor time constants inserted in the faster circuits. Hysteresis effects are compensated for by tuning the YTX well below the start frequency before the start of a sweep so it approaches the start frequency from below and continues upwards during the scan. The decision made by the microprocessor on how far to go below the start frequency and how long to hold it there before starting the sweep is based on the relationship between the previous frequency of the YTX and the new frequency.

Heating Effects

The effects of differential thermal expansion in the YTX tuning magnet, already reduced a factor of 10 by aluminum thermal short circuits (see page 10), are further reduced by firmware routines to a point where they are negligible for most applications. When a frequency change is called for, the microprocessor calculates the final temperature, T_x, that the YTX magnet would arrive at, proportional to the power dissipated in the YTX ($T_x = K_1 \times f_c^2$, where the center frequency f_c, is proportional to the coil current that causes the heating). The microprocessor also stores a variable that is indicative of the magnet's present temperature, T_i. At the end of each sweep, the microprocessor then calculates a new value of T_i such that:

$$T_i \leftarrow T_i + K_2 (T_x - T_i)(t_s + t_p)$$

where K_2 is the self-heating thermal conductivity of the magnet, t_s is the sweep time, and t_p is the nominal end-of-scan processing time.

The microprocessor then uses T_i to calculate a tuning coil current offset, applied through a DAC, to compensate for the short-term self-heating effect. The only restriction in the application of this technique is that the analyzer should be kept sweeping since it is the only way the processor has of keeping track of time.

Another new function performed by the microprocessor is determination of the time required for phase-error transients to die down before a sweep starts. An example will illustrate the need. If the YTO were sweeping a 5-MHz scan, the narrowest open-loop scan, a change in center frequency could cause the frequency to step over a band-switching point, causing the YTO to step from one end of its tuning range to the other. This frequency step could be as large as 3 GHz, yet the sweep start must be within 2% of 5 MHz or 100 kHz when the sweep starts. The microprocessor, by knowing the frequency the YTO is coming from, the frequency to which it is going, and the span width, calculates the time the YTO must be held in the phase-lock state to allow transients to die down to the point where the frequency specification can be met.

Preselector Peaking

In cases where a critical level measurement requires the YTX to be tuned exactly to a signal's frequency, the microprocessor can perform a preselector peaking routine. This routine is called when the display marker is positioned on the signal of interest and the PRESEL PEAK pushbutton is pressed. Using the offset DAC, the microprocessor tunes the YTX to maximize the amplitude response of the displayed signal and then retains the offset. The offset is also stored along with the control settings whenever the SAVE pushbutton is pressed.

Reference

1. M.S. Marzalek and L.W. Wheelwright, "Developing the Digital Control System for the Model 8568A Spectrum Analyzer," Hewlett-Packard Journal, June 1978.

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shortened coaxial cavity resonator as the Model 8672A Synthesized Signal Generator.[1,2] To obtain the desired tuning range with a high-Q resonator, the VCO runs at twice the M/N output frequency, or 364 to 396 MHz. The oscillator output is mixed with a 400-MHz reference to derive a difference frequency in a range of 4 to 36 MHz (Fig. 3). The mixer output is amplified and then divided down in frequency by the factor M, which ranges from 8 to 27. This is applied to one input of a phase-frequency detector. The 20-MHz reference is divided by N and applied to the other input of the phase-frequency detector, the output of which controls the VCO. When locked, the VCO frequency must then satisfy the relationship:

$$20/N = (400 - f_{VCO})/M \text{ MHz}$$

or,

$$f_{VCO} = 400 - 20\,M/N \text{ MHz}$$

This is divided by 2 to derive the M/N output frequency:

$$f_{M/N} = 200 - 10\,M/N \text{ MHz}$$

Since the YTO is locked to the Nth harmonic of this frequency such that:

$$f_{YTO} = 200N - 10M - f_{lfs} \text{ MHz},$$

it can be seen that the Nth harmonic moves in 10-MHz steps when M is incremented, and 200-MHz steps when N is incremented. M and N were selected such that N changes after M steps through its range. Stepping M through its

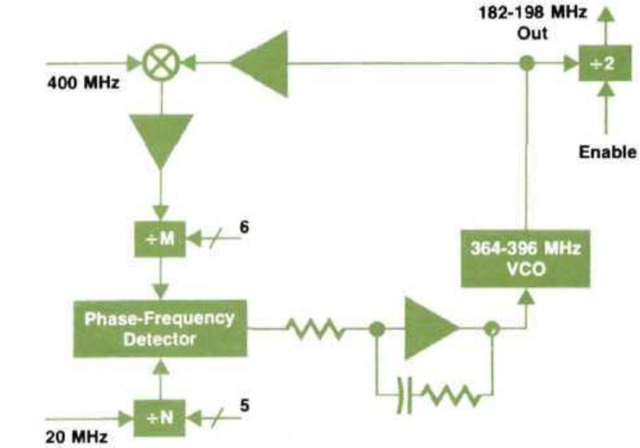

Fig. 3. *Block diagram of M/N loop.*

range causes the Nth harmonic to step across 200 MHz in 10-MHz increments.

Interpolation

The low-frequency synthesizer has three VCO loops that are interconnected in various ways to enable the analyzer to scan frequency spans as narrow as 100 Hz with the same relative accuracy and low phase noise as the broadband sweeps. This is done with a divide-and-upconvert scheme.

The basic source of the low-frequency synthesizer sweep frequencies is loop 1 in Fig. 4. It has a 75-150 MHz VCO whose output frequency is divided by 5 to generate an output in the 15-30 MHz range. This output is used directly by the YTO phase-lock loop for frequency sweeps ranging from 100 kHz to 5 MHz. In this range, the start frequency is settable with 1-kHz resolution.*

For narrower frequency spans, the other two loops come into play. The loop 1 output is then divided down by a factor of 5 or 100, depending on the sweep range. Loop 2 serves to up-convert the divided-down sweep frequency to a range between 160.15 and 166 MHz (see Fig. 4).

The final frequency translation takes place in loop 3. This loop has a VCO operating in a range of 199 to 300 MHz. This VCO output is applied to a mixer whose other input is the output of loop 2. The difference frequency output of the mixer is divided by 2 and the result is divided down to 5 MHz by a factor N_1, using fractional-N techniques,[2] and

*These numbers for span width and start frequency settability apply when the input frequency is below 5.8 GHz where LO fundamental mixing is employed. For higher input frequencies where harmonic mixing is used, the numbers are multiplied by the harmonic number. For example, at 20 GHz (4th harmonic mixing), the span width can be as wide as 20 MHz with the YTO locked to the low-frequency synthesizer, and the start frequency is settable with 4-kHz resolution.

Switch Position	Output Range	Synthesis Resolution	Scan Width
1	15-30 MHz	1 kHz	0.1-5 MHz
2	19.9-30	20 Hz	5-100 kHz
3	19.995-30	1 Hz	0.1-5 kHz

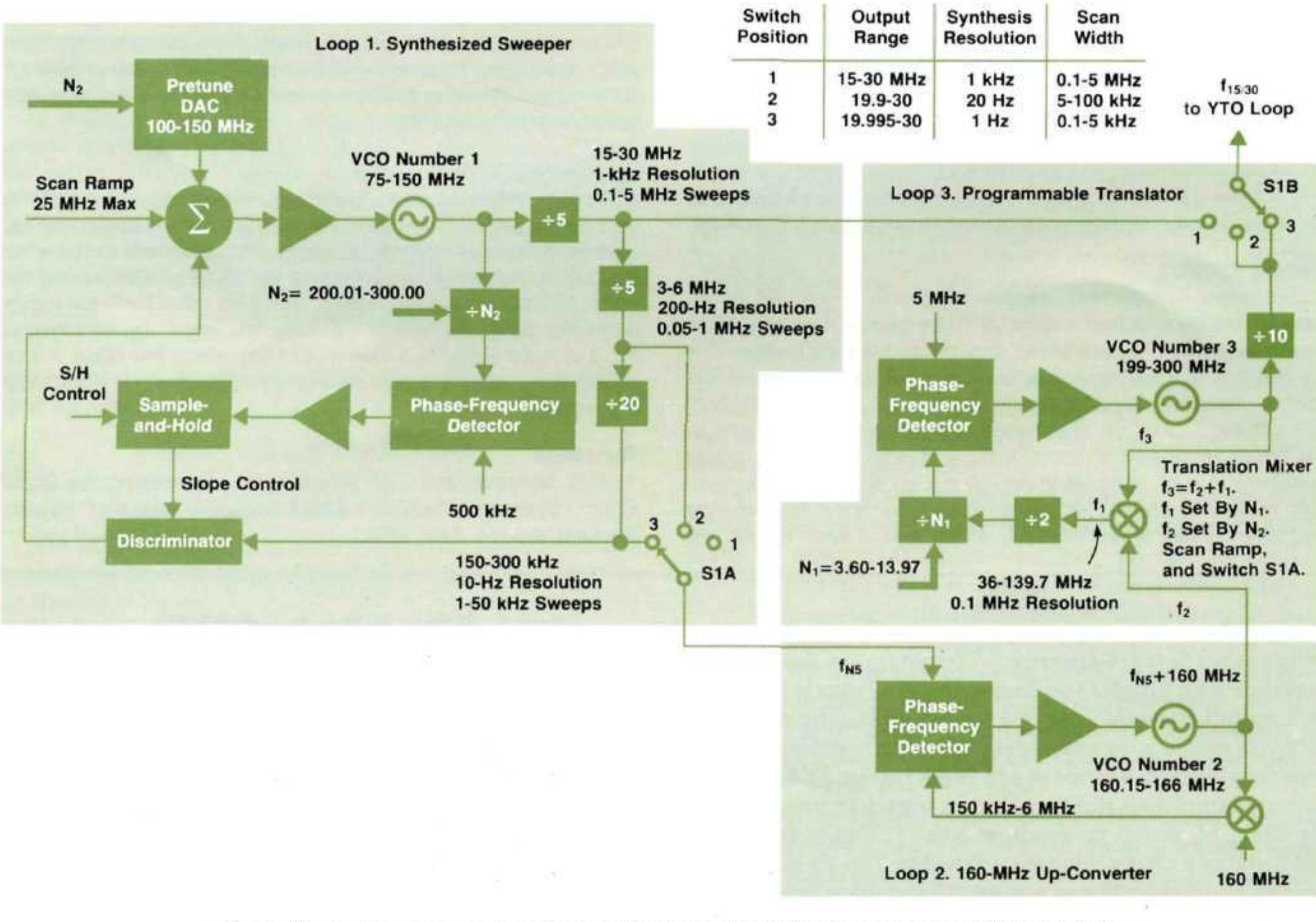

Fig. 4. *The low-frequency synthesizer has three loops. Loop 1 is the basic source of a frequency that is divided down then up-converted by loop 2 and translated by loop 3 to derive an appropriate frequency for control of the YTO.*

A Precision Discriminator with a Controllable Slope

by Stephen T. Sparks

The VCO that generates precision sweeps in the low-frequency synthesizer of the Model 8566A Spectrum Analyzer is controlled through a feedback loop that includes a discriminator (see Fig.4, page 16). A key characteristic of this discriminator is that the slope of its voltage/frequency response curve is adjusted by a phase-lock loop, thus enabling highly accurate narrowband sweeps.

A block diagram of the sweep-frequency control system is shown in Fig. 1. The discriminator is the pulse-count type in which the input signal triggers a pulse generator, which generates a single pulse of constant width and height for each cycle of the input. These pulses are integrated to obtain a dc current that is proportional to the input frequency.

The transfer characteristic of the discriminator is characterized by the equation:

$$I_2 = Kf + B$$

where K is the slope of the response curve, f is the input frequency, and B is an offset (see Fig. 1). A frequency change Δf results in an output change ΔI that is a function of K only. In the usual discriminator, the term K is subject to warm-up and long-term drift that exceeds the 0.1% accuracy desired for the 8566A. Therefore, some precise means of modifying K is needed to maintain accuracy.

Given that two points determine the slope of a line, fixing two points will fix K. In the discriminator described here, care was taken to reduce the offset B to negligible proportions. Thus, the origin is one of the fixed points on the response curve.

The second point is fixed during the pre-sweep interval while phase lock is being established. During this interval, the scan input where the sweep ramp is applied is zero (I_3 in Fig. 1). A precise pre-tune current (I_4) corresponding to the start frequency is fed into the VCO control loop. The N_2 phase-lock loop (Fig. 4, page 16), also programmed to the start frequency, then functions to adjust current source I_1 until lock is achieved. The discriminator response curve is thereby rotated to the point where the pre-tune current drives the loop to the desired start frequency. This, in effect, sets the second point, fixing K. The discriminator response is thus tied to the analyzer's frequency standard by way of the 500-kHz reference.

When the scan starts, the voltage that sets I_1 is stored in a sample-and-hold, opening this leg of the phase-lock loop. The discriminator is in the feedback loop that controls the VCO and it causes the VCO frequency to closely follow the tuning ramp, eliminating the effects of any tuning nonlinearities or temperature coefficient that may be in the VCO.

Precision Pulses

If the discriminator is to be truly linear, the width and height of the

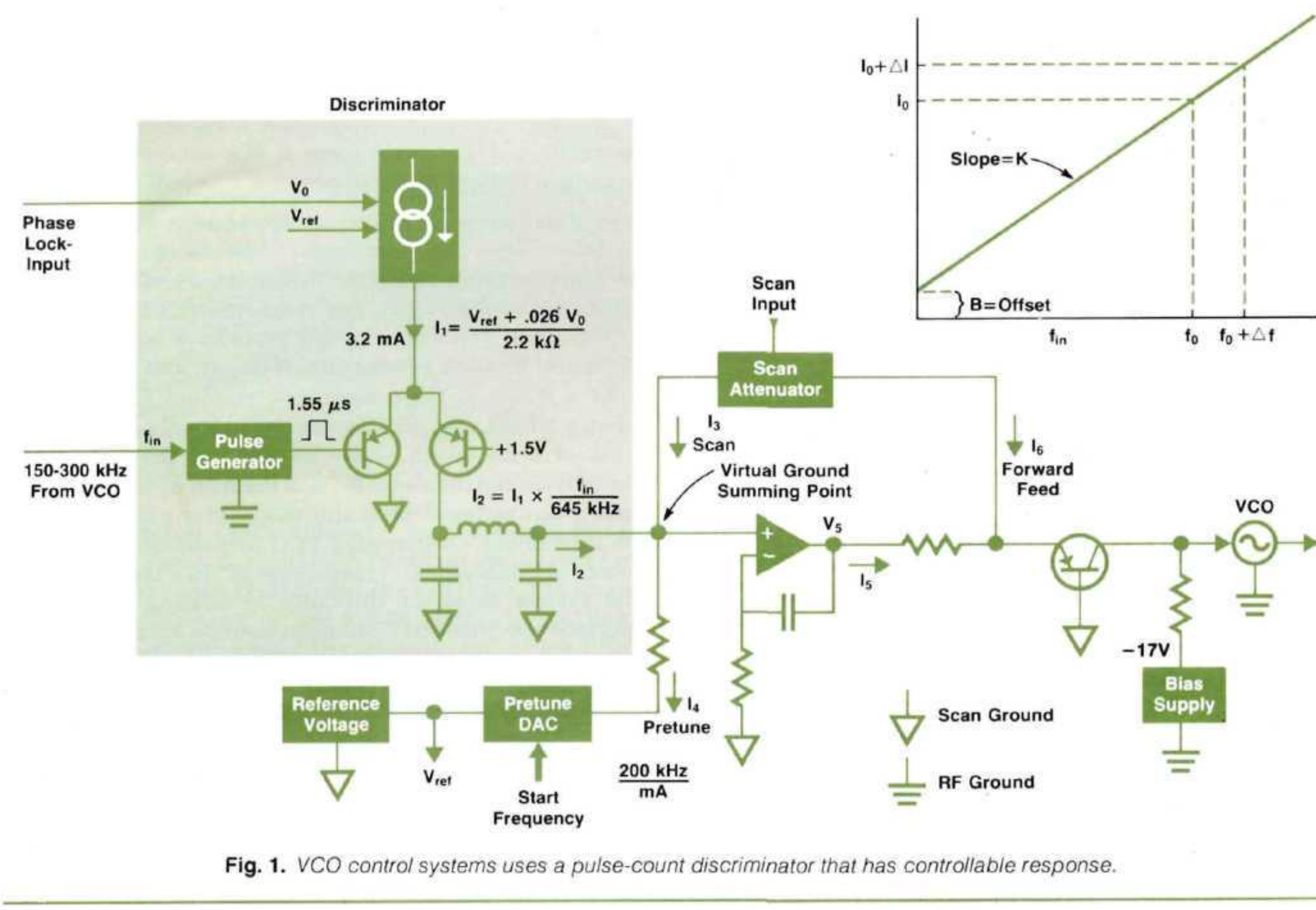

Fig. 1. *VCO control systems uses a pulse-count discriminator that has controllable response.*

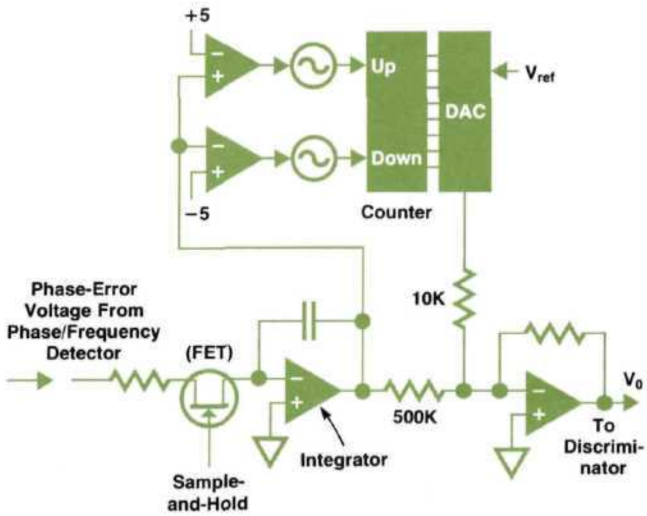

Fig. 2. *Digital sample-and-hold.*

discriminator output pulses must be uniform at all input frequencies. Uniform pulses are obtained in this discriminator by using the input pulses to drive a common-emitter transistor that has a resonant LC circuit in its collector circuit. Each pulse saturates the transistor, causing the LC circuit to ring at 5.2 MHz. The ringing waveform is applied to a ÷16 counter. Prior to each input pulse, the counter is preset to 6. Its MSB output then goes high on the second zero crossing of the ringing waveform (count = 8) and goes low again eight cycles later (count = 16). The width of the pulse generated by the MSB output is thus determined precisely by the frequency of the ringing circuit. These pulses are integrated to obtain I_2 (Fig. 1).

An important feature of this arrangement is that presetting the counter and using the MSB output causes it to ignore the first cycle of the ringing waveform. The first cycle is of low amplitude and of variable period because of the finite saturation time of the driving transistor. Also, at high input frequencies the LC circuit may not have settled down from the previous pulse, causing further variations in the first cycle. Skipping the first cycle results in a significant improvement in discriminator linearity.

The discriminator response deviates from a straight line less than ±0.002% for an integrated pulse current (I_2 in Fig. 1) ranging from 0 to 2/3 of I_2 max, and the incremental linearity, which determines the accuracy of narrow scans, is within ±0.03%. The discriminator actually operates between 1/3 and 1/2 of I_2 max, corresponding to an input range of 200 to 300 kHz.

Fast Lag and Slow Droop

Current I_6 in Fig. 1 is a portion of the sweep ramp that is applied directly to the VCO, reducing the excursions of voltage V_5. This reduces sweep lag, the amount by which the VCO frequency lags the scan ramp. The lag, which is proportional to scan rate, occurs because of the finite bandwidth of the discriminator, which was made low to reduce noise contributed by the discriminator. The I_6 feedforward current reduces sweep lag by a factor of 10, so the maximum sweep lag is only 0.15% of scan width.

Very slow sweeps, on the other hand, are a potential source of another problem: droop in the sample-and-hold circuit. Since scans can take as long as 1500 seconds each, leakage in the sample-and-hold capacitor would have to be less than 50 pA to maintain scan accuracy.

The sample-and-hold voltage is held in an integrator, shown in Fig. 2. To reduce the effects of sample-and-hold drift, the oscillator system was designed so the integrator has only a limited influence on the frequency; it can adjust the frequency by only ±0.04%. This range, however, is insufficient to correct for long-term drift in the discriminator. Thus, a digital-to-analog converter (DAC) and related circuitry was added to the integrator, as shown in Fig. 2. If the output of the integrator attempts to exceed ±5V during the phase-lock interval, one of the two oscillators shown in Fig. 2 is turned on, either clocking the counter up or down. The counter drives the DAC whose output is summed with the analog integrator output, thus contributing to the output V_0. Counting continues until the total output steps past the value required to achieve phase lock. The phase error voltage then causes the analog circuit to retreat, cancelling the overshoot. This shuts off the oscillator, and leaves the analog circuit in the middle of its range. The contribution of droop in the analog circuit after the sample-and-hold switch opens is thus reduced to negligible proportions because the major portion of V_0 is held in the DAC.

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compared to a 5-MHz reference. The following equation thus applies:

$$(f_{VCO3} - f_{loop2})/2N_1 = 5 \text{ MHz}$$

or,

$$f_{VCO3} = 10N_1 + f_{loop2}.$$

The VCO3 frequency is thus offset from the loop 2 frequency by $N_1 \times 10$ MHz, where N_1 ranges from 3.60 to 13.97.

The VCO3 frequency is divided by 10 to place it in the 19.9-to-30-MHz range and then supplied to the YTO loop. The output of loop 1 is thus divided down by factors of 50 or 1000 and then offset in frequency to place it in the proper range for the YTO.

What this scheme accomplishes is the reduction of the phase noise, residual FM, and synthesis-related spurious outputs of loop 1 by as much as 60 dB (20 log 1000) for frequency spans of less than 5 kHz, thus enabling the analyzer to meet the more stringent requirements of narrow-band scans. In addition, for spans less than 25 kHz, the gain of the YTO loop is increased about 10 dB. Although this degrades the far-out phase noise somewhat, it improves the close-in phase noise and line-related spurious performance. Of course, loops 2 and 3 make their own phase-noise and spurious contributions but these have been held to a minimum through careful circuit design and shielding.

As a result of these measures, residual FM, start frequency offsets, and drift during a scan are always less than 0.1% of scan width (one "bucket" of the 1000-point digitally-stored display is 0.1% of the scan width). Note that there is no overlap of the tuning ranges of any of the VCOs in the instrument. Overlapping VCO frequencies are a prime source of troublesome "crossing spurs" in a synthesizer, so this system eases the difficulty of meeting the desired spurious specifications. Spurious outputs are at least 90 dB below the carrier while broadband phase noise is at least 10 dB below that of the YTO loop.

For frequency spans greater than 5 MHz, where the sweep ramp is applied directly to the YTO, the synthesizer sweeping function is disabled and it then provides start frequencies with 1-Hz resolution.

Discriminator-Stabilized VCO

Loop 1 uses a stabilization system that includes a frequency discriminator in the phase-lock loop, as shown in Fig. 4. During a sweep, the loop 1 VCO is controlled by the discriminator output and is therefore sensitive to the discriminator's frequency-to-current response. During the pre-sweep interval, the phase-lock error voltage tunes the VCO by adjusting the slope of the discriminator's response while a known input current is applied to the discriminator, as explained on page 17. Then, when the VCO output divided down to 500 kHz becomes phase-locked to a 500-kHz reference, the discriminator slope is precisely known.

When a sweep is initiated, the error voltage from the phase detector is stored in a sample-and-hold circuit, effectively opening this feedback loop. A voltage ramp for sweeping the VCO is applied to a summing network along with the discriminator output (and the dc pretune voltage). The discriminator output thus functions as a negative feedback signal to cause the VCO frequency to closely track the ramp, thus correcting for any nonlinearities in the VCO tuning curve. Since the discriminator frequency-to-current response has been set accurately by the phase-lock loop, outstanding sweep accuracy results.

Controlling Residual Responses

Because the M/N output remains enabled during these sweeps (it is disabled during sweeps greater than 5 MHz), there is a possibility that residual responses could be generated by M/N harmonics that mix with the YTO in the input mixer to produce sum or difference frequencies near the first IF. An analysis of this situation determined that such mixing products could occur. Therefore, the YTO output is coupled to the phase-lock loop through a microcircuit directional coupler that has a 2-to-6-GHz GaAs FET amplifier in the coupled arm (see Fig. 2). Pads in the amplifier input and output attenuate the YTO signal such that the net gain through the pads and amplifier is 0 dB. The reverse isolation, however, is increased by the pads which, with the reverse isolation of the amplifier, gives a total reverse isolation more than 40 dB greater than the reverse isolation of the coupler itself. Residuals due to the M/N loop have thus been reduced to the point that most lie below the noise floor of the analyzer, even with the 10-Hz resolution bandwidth.

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Stephen T. Sparks

Steve Sparks received his BSEE degree in 1968 and MSEE degree in 1969 from the University of California at Berkeley. AT HP, he contributed to the circuit design of the 8601A and 8605A Sweepers, the 8407A and 8754A Network Analyzers, and the 8672A Signal Generator (he joined HP shortly after graduation from high school). Steve keeps busy with a variety of activities: backpacking, jeep trips, photography, restoring antique radios, brewing beer at home, cooking, winetasting, and flying small planes. He also grazes cattle and plans to develop a vineyard on his 50-acre ranch.

Kenneth L. Lange

A native Oregonian, Ken Lange received his BSEE and BAEE degrees in 1967 from Oregon State University and his MSEE degree in 1968 from Stanford University. Ken joined HP in 1973 and was responsible for circuit design in the 8557A Spectrum Analyzer and the 85660A RF Module. A resident of Santa Rosa, California, Ken is married and has two young sons, ages three and one. Handball, gardening, and flying small planes keep him busy in his spare time.

Larry R. Martin

Larry Martin received his BSEE degree in 1967 from Kansas State University. An HP employee since that same year, Larry was a project manager for the 8672A Synthesized Signal Generator, contributed to the design of the 8555A spectrum analyzer tuning section, and earned an MSEE from Stanford University along the way. Larry lives in Santa Rosa, California, is single, and keeps busy with photography, wine tasting, softball, tennis, and basketball. He also enjoys flying a sailplane occasionally.

Michael S. Marzalek

Mike Marzalek received his BSEE degree in 1969 from the University of California at Berkeley and his MSEE degree in 1972 from Stanford University. With HP since 1969, Mike designed the microprocessor software for the 8566A, designed hardware for the 8568A Spectrum Analyzer, and contributed to several synthesized signal generators. Born in Springfield, Missouri, and raised in southern California, Mike is married and spends many of his leisure hours crosscountry skiing, down-hill skiing, playing folk guitar, and working in stained glass.