APRIL 1982

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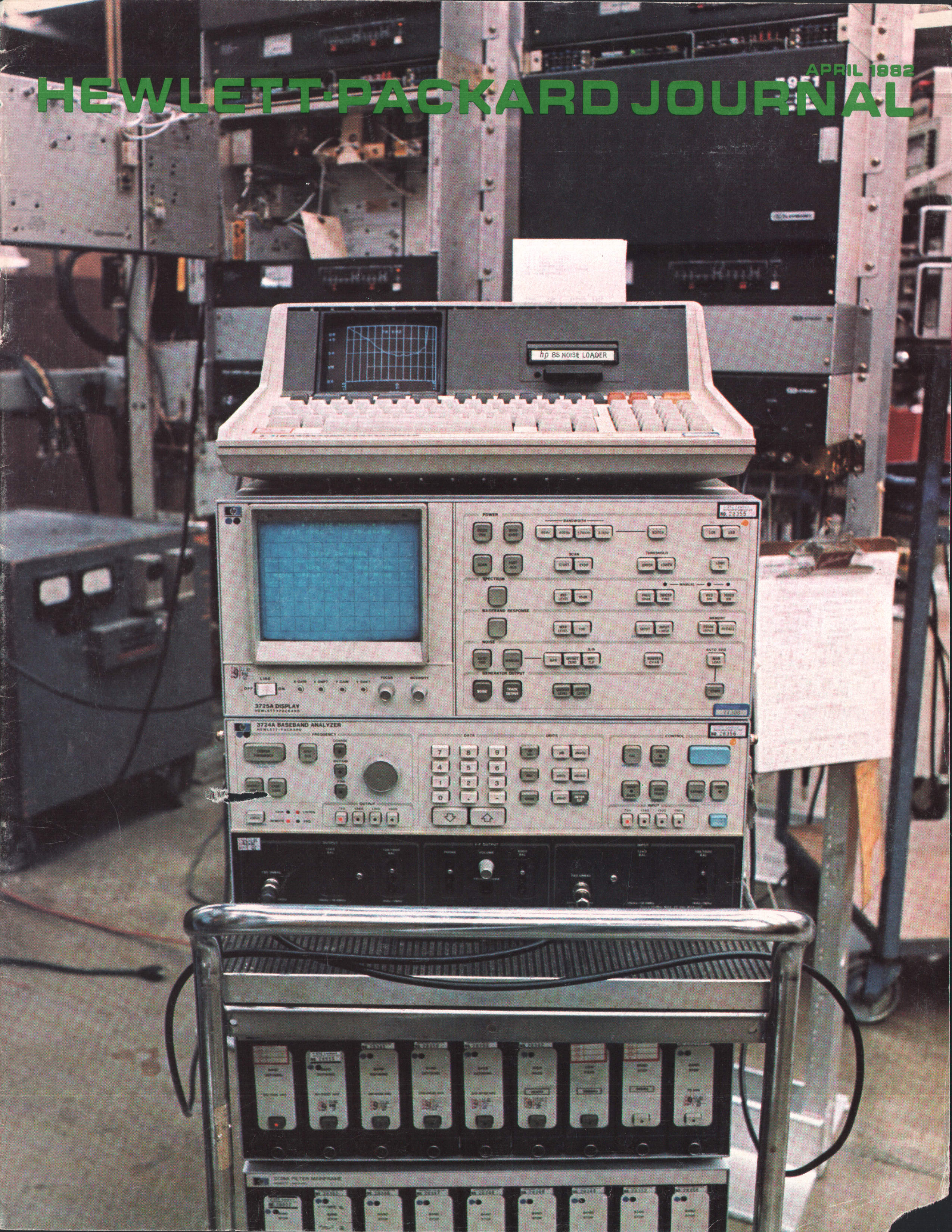

HEWLETT·PACKARD JOURNAL

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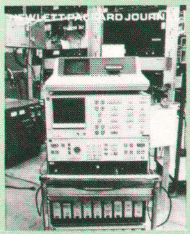


Between any two of the multitude of microwave antennas that dot our countryside there may be, to give a typical number, 2700 conversations going on simultaneously. They are kept from interfering with each other by assigning each conversation its own channel, a narrow band of frequencies within a band wide enough to hold all 2700 channels. The composite 2700-channel signal, called a baseband signal, is ultimately impressed on a high-frequency microwave carrier signal for transmission from antenna to antenna.

Frequency division multiplex microwave radio communications links of this type are complex systems that require frequent testing to maintain. The required tests, which are many and complex, are specified by various standards-setting organizations such as CCITT, CCIR, Bell, and Intelsat. Some tests are made on the baseband equipment, some on the microwave signal, and some at intermediate frequencies that exist within radio systems. Both in-service and out-of-service tests are done.

This month's cover subject, Model 3724A/25A/26A Baseband Analyzer, is designed to simplify and reduce the costs of radio system testing. It combines all of the instruments commonly used on the baseband of a microwave radio system in a single integrated test set with a common display and keyboard. Measurement modes can be changed at the press of a key without the recabling and retuning that have to be done if multiple test sets are used. The 3724A automates many routines that were formerly time-consuming and tedious, thereby lowering test and inspection times and improving accuracy, and it does the job of both in-service and out-of-service test sets. A major feature of the 3724A Baseband Analyzer is an advanced receiver that takes on many different personalities under control of the analyzer's built-in microprocessor. The design of this receiver is described in the article on page 8. Other 3724A articles are on pages 3, 18, and 22.

In today's electronic circuits, transitions often occur within nanoseconds (a nanosecond is a thousandth of a millionth of a second). To observe these fast events, the design engineer typically uses an oscilloscope. If the event of interest is a transient or glitch that occurs so seldom that an ordinary oscilloscope can't produce a usable trace, it may still be possible to capture and display the anomaly using a storage oscilloscope. The article on page 26 describes the design of a storage oscilloscope that's about as good as you can get for this kind of job. Combining an extremely fast writing rate of 2000 centimetres per microsecond with a high bandwidth of 275 megahertz, the HP 1727A Storage Oscilloscope is designed to display elusive events that are difficult to observe because they are fast and infrequent.

-R. P. Dolan

An Integrated Test Set for Microwave Radio Link Baseband Analysis

This instrument combines six traditional test instruments into one package for easy baseband measurements from 50 Hz to 18.6 MHz. An internal microprocessor simplifies test setup, improves accuracy, and enables the instrument to check itself.

by Richard J. Roberts

TODAY'S microwave radio systems are placing increasing demands on the qualified personnel available to maintain them. This in turn creates a strong need for more accurate, easier-to-use instrumentation that can be used either with local automatic control or as part of a centralized surveillance system. This can be difficult to achieve when one considers the number of instruments necessary to maintain a microwave radio station, each with its own interface and input/output structures. Similar problems confront manufacturers who desire high test throughput to minimize costs without sacrificing measurement accuracy. Reduced test time is even more important during the commissioning phase when high-caliber engineers have to carry out large numbers of measurements before a new radio link becomes operational.

The HP Model 3724A/25A/26A Baseband Analyzer (Fig. 1) contains all of the test instruments that are used on the baseband of a radio link (up to 18.6 MHz) and links them to a common CRT (cathode ray tube) display and keyboard. This convenient package can effectively replace the following instruments:

- Wideband power meter
- Selective voltmeter
- Synthesized signal generator
- Frequency counter
- Spectrum analyzer
- White-noise test set.

This new instrument provides a high degree of user flexibility, very high measurement accuracy, and new ways of doing measurements. Large time savings can be made when commissioning radio links, greater throughput can be achieved in manufacturers' final system tests, and there is less dependence on operator skill in achieving high accuracy measurements.

The instrument is contained in three boxes; the lower two contain the main instrument and the upper one is used to house the plug-in filters for white-noise testing. The instrument may be broken down into the following sections (Fig. 2).

- A versatile selective receiver with very low distortion, low noise and high accuracy. The appropriate signal routing and IF (intermediate-frequency) filtering are automatically selected by the internal microprocessor and power measurements are made by a true-rms detector or a logarithmic amplifier/detector.
- A low-noise, wide-range, true-rms wideband power measurement path.
- A synthesized local oscillator giving receiver tuning from 10 Hz to 18.6 MHz in 10-Hz steps. This oscillator is phase-locked to an internal HP 10811A Crystal Oscillator. The same local oscillator is also used in an open-loop swept mode for spectrum analysis.
- A generator providing either a tracking CW (continuous-wave) output signal or a white-noise load-

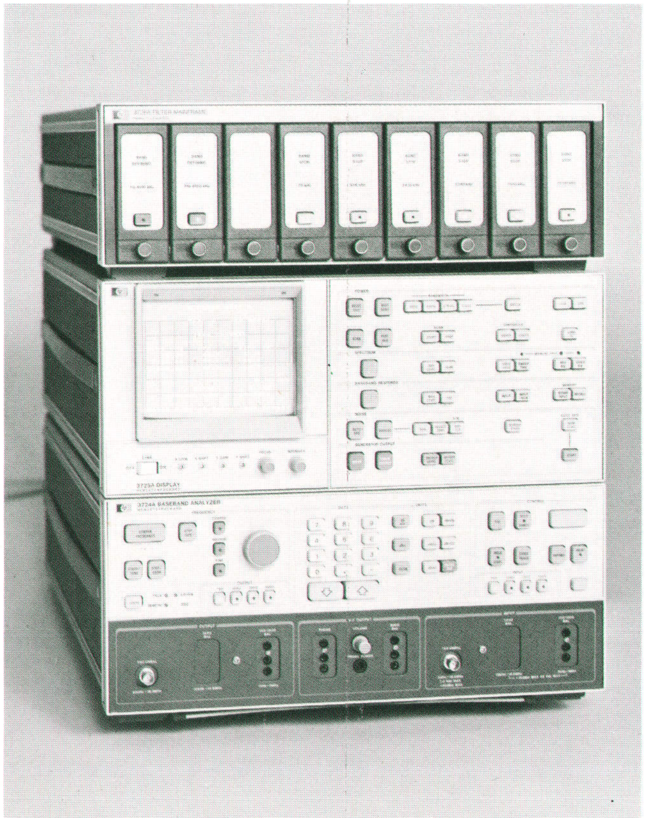


Fig. 1. *The HP Model 3724A/25A/26A Baseband Analyzer can measure all of the baseband parameters of a microwave FDM (frequency division multiplex) radio link that are required to characterize its performance.*

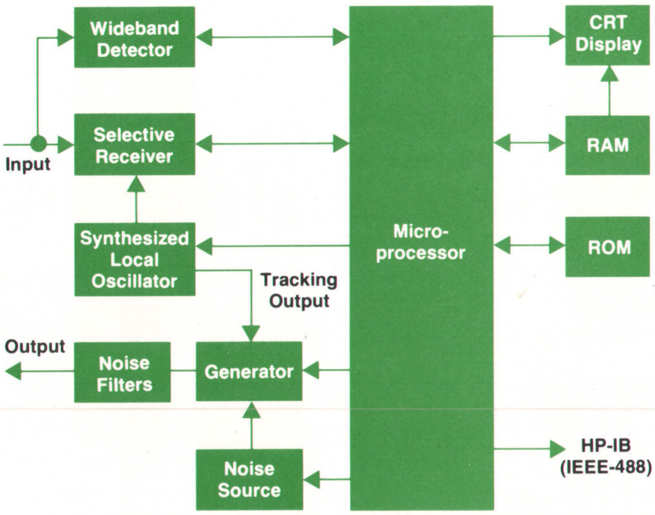


Fig. 2. *Simplified block diagram of the 3724A/25A/26A Baseband Analyzer.*

ing signal conforming to CCIR* recommendations.

- Plug-in filters for white-noise loading. These consist of band-defining and bandstop filters conforming to the CCIR and Intelsat recommendations.
- Overall control of the instrument with a 6800 microprocessor and 60K×8 bits of ROM (read-only memory) and 4K×8 bits of RAM (random-access memory). The RAM is also used by the display driver circuits via direct memory access. The CRT display is an HP 1340A Display Module.
- Remote control via the HP-IB.**

In addition to the capabilities of all the above instruments, the baseband analyzer allows rapid switching between these measurements without retuning or recabling. Thus a problem can always be diagnosed or analyzed using the optimum or alternate configurations of the instrument. Not only is the 3724A/25A/26A Baseband Analyzer a valuable instrument for setting up and troubleshooting microwave links, it is also an excellent tool for in-service measurement and analysis with dedicated algorithms for such use.

Wideband and Selective Power

True-rms wideband power measurments (Fig. 3a) can be made from +20 dBm to less than −76 dBm over a frequency range from 20 Hz to 18.6 MHz. Selective level measurements (Fig. 3b) may be made with either a 3.1-kHz channel filter, a 1.74-kHz noise filter, a 400-Hz narrowband filter, or a 40-Hz pilot filter. The noise floor is very low at less than −125 dBm in a 1.74-kHz bandwidth. The second and third-order intermodulation distortion products are less than −75 dB and −80 dB respectively. A variety of measurement units are used in the worldwide communications market and to avoid confusion or possible errors the appropriate units may be selected from the keyboard. Thus wideband power measurements may be made in dBm and pW. Also by entering a known test-level point, relative measurements in dBmO or pWO may be made. Selective level measurements in a 3.1-kHz bandwidth may also include true psophometric or C-message-weighting networks giving measurements in dBmp, dBmOp, pWOp, pWOC, dBrnC and dBrnCO.*

Noise-with-tone measurements can be made. In this case the tone in a channel is filtered out with a 1-kHz notch filter, leaving the noise in that channel to be measured.

The exact frequency of a tone within the measuring bandwidth may be counted to either 1-Hz or 0.1-Hz resolution. This resolution may be used for checking pilot frequencies or measuring low-level spurious tones as an aid to tracing their origin.

The tuned frequency can be selected in a variety of ways besides the normal keyboard entry or manual control using the rotary pulse generator. First, the frequency of a tone within the measurement bandwidth as measured by the internal frequency counter may simply be transferred to the tuned center frequency. This ensures maximum measurement accuracy and ease of use when using the narrow-bandwidth filters. Second, when making in-service measurements on a live system the telephone channels are described by tables giving each channel's virtual carrier or reference tone frequency and whether it is upper sideband or lower sideband. The tuned frequency of the baseband analyzer can be entered in terms of the carrier or tone frequency and upper or lower sideband, and the instrument will automatically tune the receiver to either the center of the channel with the 3.1-kHz and 1.74-kHz filters or the carrier frequency or tone frequency with the 400-Hz or 40-Hz filters. Third, a frequency step size can be entered and the center, carrier or tone frequency incremented by that amount using the ⇑ or ⇓ keys.

Tracking Generator

The tracking generator can be used in conjunction with most modes of the instrument. It has a range from +6 dBm to −60 dBm with 0.1-dB resolution and is flat within 0.2 dB from 10 kHz to 14 MHz. With the restriction that it always tracks the receiver setting this output can be used as a synthesized signal generator in most applications.

Scan and High-Level-User Algorithms

The scan mode is a general-purpose algorithm that consists of a sequence of selective level measurements between defined start and stop frequencies, each of which is compared with either or both upper and lower threshold limits. Any of the four measurement bandwidths may be selected and any frequency-step size may be used. Measurements outside of these limits are tabulated on the CRT (Fig. 3c), thus avoiding the need for a separate printer. The format gives the frequency of the violation, its maximum or minimum power level, and a count of the number of times that the frequency violation has occurred. A pointer may be moved to any frequency in the table (2164.00 kHz in Fig. 3c) whereupon that frequency will be automatically transferred to the selective-level or spectrum analysis modes for further investigation when either mode is selected.

The high-level-user algorithm addresses the need for

*International Radio Consultative Committee

**Hewlett-Packard's implementation of IEEE Standard 488 (1978).

*Editor's Note: Here O indicates a measurement value relative to a known test level, p indicates psophometric weighting, C indicates C-message weighting, and rn indicates reference noise level. Hence, dBrnCO is dB above reference noise level with C-message weighting corrected by the test-level value.

rapidly finding any signals in the actual baseband traffic that are above an upper threshold limit. This is an ever present problem, especially with the variety of equipment available that the consumer may connect to the telephone line. Location of these offending signal sources is usually time-consuming and ineffective, yet with this mode of operation, 1800 channels can be scanned in about 80 seconds, rapidly giving the frequency and level of any offending signal. If left running over a period of time the table displayed on the CRT will give an indication of the duration of each signal together with its frequency and maximum level.

Spectrum Analysis

The spectrum analysis mode (Fig. 3d) gives the operator a conventional spectrum analyzer with 2 or 10 dB/division resolution and automatic or fixed coupling between resolution bandwidth, video bandwidth and sweep time. This gives a choice between complete flexibility and ease of use with generally optimum settings. In addition there is a shifted key labeled **SPURIOUS** that presets the resolution and video bandwidths to the optimum parameters to find spurious tones on the unloaded baseband in the presence of noise. The tracking generator operates in a conventional manner in this mode.

Baseband Response

The selective-level, scan, and spectrum analysis modes allow amplitude response measurements of components or systems, with high accuracy on a point-by-point basis or high dynamic range on a swept basis. The baseband response mode combines high resolution and a fast swept response together with local or end-to-end measurements. Thus the baseband amplitude response of a radio link can be rapidly checked and adjusted with a clear display of the measurement in contrast with the conventional point-by-point method of measurement and adjustment.

The 1 dB/division or 0.1 dB/division display (Fig. 3e) consists of 33 equally spaced selective level measurements made between defined start and stop frequencies, the number of points being chosen to give the optimum balance between resolution and update rate. Although the residual response of the 3724A and the test cables connected in a back-to-back configuration is typically ±0.24 dB, there is also a memory which can be used to normalize any measurement against either a previous measurement or, since the data in the memory may be entered through the HP-IB, a predetermined standard or shape. Such normalization allows the residual responses to be removed from the measurement result.

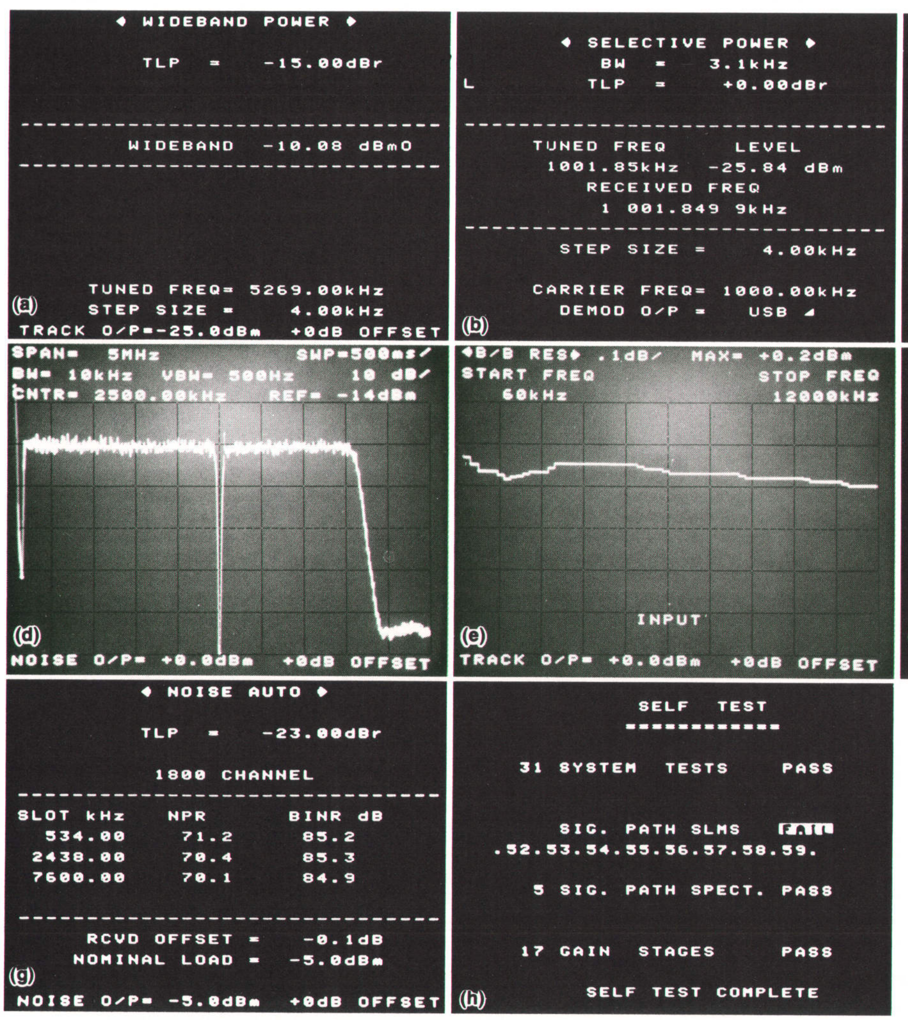


Fig. 3. *Typical displays for various modes of the 3724A/25A/26A Baseband Analyzer. (a) Wideband power. (b) Selective level. (c) Scan. (d) Spectrum analysis. (e) Baseband response. (f) Manual noise. (g) Automatic noise. (h) Self-test.*

White-Noise Testing of FDM Communication Links

Frequency-division-multiplex (FDM) has been the major method of combining individual telephone channels into one complex signal since 1918. The early communication systems were limited to about 24 channels by the open-wire lines and technology available at that time. Today, it is common for 2700 channels to be multiplexed together for transmission over a single microwave radio channel and over 10,000 channels can be multiplexed onto a coaxial cable system. A basic "group" consists of 12 telephone channels, each modulated onto a 4-kHz band between 60 and 108 kHz using single-sideband suppressed-carrier modulation. In the same manner "supergroups" and "mastergroups" are built up until a complete baseband signal of the required capacity is formed.

It is essential that the parameters of the whole transmission system are maintained so that the other simultaneous conversations do not degrade any one conversation below an acceptable level. White-noise testing is a means of establishing the overall subjective quality of a system by measuring one parameter instead of all the parameters of the individual parts.[1] This is done by replacing the traffic being carried by the baseband by white noise that is band-limited by filters to the bandwidth of the baseband under test. This gives a very good simulation of the normal traffic, and by varying the overall applied power level one can simulate conditions varying from light loading to severe overloading.

Very narrow-bandwidth bandstop filters (ideally 4 kHz wide) can be inserted at three or four frequencies in this white-noise signal to simulate quiet channels in the normal baseband traffic (Fig. 1). If this signal is now passed through the system under test there will be a certain increase in the noise in these quiet channels, or slots, because of thermal-noise limitations and intermodulation distortion within the system. This can be measured as the noise power ratio (NPR), which is the ratio in dB of the noise density P_1 with the bandstop filter switched out to the noise density P_2 in the slot with the bandstop filter switched in; it is normally measured in a 1.74-kHz bandwidth. Other units are often used to describe this signal-to-noise measurement but the technique is essentially the same in each case. A further measurement is usually made to distinguish between the thermal or intrinsic noise contribution of the system under test and the intermodulation noise. This intrinsic noise is either measured as an absolute power or as an NPR with the noise switched out, P_1 being the power level at nominal loading had the noise signal been applied. This measurement is called the basic intrinsic noise ratio (BINR).

Thus an overall figure of merit for the system can be obtained for low, middle, and high-frequency parts of the baseband at the nominal loading level. In addition, these measurements can help indicate the direction for troubleshooting when the performance of the system is below an acceptable level.

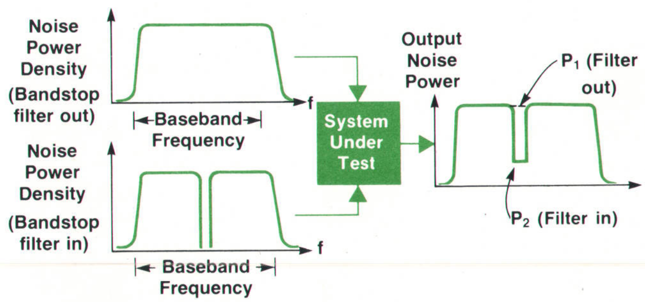


Fig. 1. *By measuring the noise power output from a system under test with a white noise input signal with (P_2) and without (P_1) a narrow bandstop filter switched in, the noise power ratio (NPR) can be determined from the equation NPR = 10 log (P_1/P_2) dB.*

Reference:
1. M.J. Tant, "The White Noise Book," White Crescent Press Ltd, Luton, 1974.

Measurements may be made either on a local basis using the tracking generator output or on a remote basis using the receiver of one instrument and the tracking generator of another. In the latter case the receiver locks onto the incoming signal without any additional interconnection.

White-Noise Testing

For white-noise testing the generator output is a uniform spectrum of white noise used to simulate the traffic on a radio link (see box above). The filters for this mode, together with the noise source, are housed in the 3726A mainframe. The low-distortion baseband analyzer meets the recommended CCIR test instrument specification of a noise power ratio (NPR) greater than 67dB without the need for matching bandpass prefilters as normally used in white-noise test sets. If higher NPR measurements are needed, such as for component testing, then bandpass filters can be added easily to the baseband analyzer.

The instrument may be operated in either a manual (Fig. 3f) or an automatic sequence (Fig. 3g) mode. The former is compatible with existing white-noise instrumentation and gives the operator complete measurement flexibility. Since bandpass filters normally are not used in the 3724A the receiver may be tuned to any slot frequency in the baseband, or when traffic is on the system, to any inband or out-of-band monitoring slot.* In the **AUTO SEQ**uence mode a series of measurements is made to display (Fig. 3g) the noise power ratio or signal-to-noise ratio and the thermal noise (noise generator switched off) at up to four slot frequencies across the baseband without any intervention by the operator. This measurement also can be made on a remote basis without any additional interconnection.

HP-IB Interface

The simplest use of the HP-IB interface is with either an HP 7225A Graphics Plotter or 7245B Plotter/Printer without an additional controller. Any of the CRT displays may be obtained in hard copy by pressing the **PRINT** key on the 3724A Baseband Analyzer. In this way three or four plots can produce all the information needed to document the performance of a microwave radio link on a routine maintenance basis.

The addition of a controller, such as an HP-85A Personal Computer, makes all the front-panel controls accessible

*Slot frequencies are specified by CCIR, Intersat, etc. for sampling the performance of the baseband.

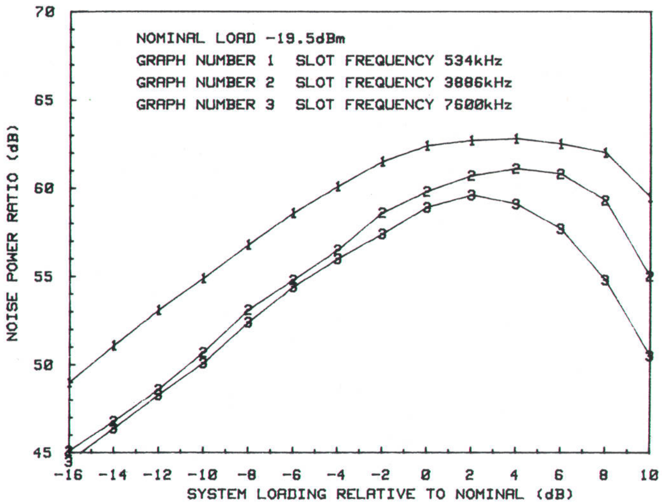


Fig. 4. *White-noise V-curves. Using the HP-IB interface and a plotter, one can directly record many measurements in a convenient format.*

through the HP-IB, plus some others that are dedicated to specific functions available only over the HP-IB. Thus the setup of measurements and large specific sequences of measurements may be rapidly carried out, analyzed and documented. An example is shown in Fig. 4 where a complete set of white-noise V-curves has been measured and plotted using a simple program. This can be extended to form part of a large centralized surveillance system for monitoring and analyzing a complex microwave radio link system, or to form part of a manufacturer's final system test. An HP-85 software package (37018A) is available to help a user develop dedicated system software by selecting an appropriate set of software modules from the group that covers each mode of operation of the instrument.

Reliability and Self-Test

It is essential that the user of the instrument have confidence that it is fully functional when operated either manually or remotely up to hundreds of kilometres away. Specific emphasis was put on reliability throughout the design phase. Examples of this are the low internal ambient and component temperatures and the scrutiny given to every component, based on its performance history within Hewlett-Packard, before being included in the 3724A/25A/26A Baseband Analyzer.

Software is used in three ways to monitor and help service the instrument. First, the instrument is monitored continuously during normal operation to flag errors in operation such as incorrect key sequences. The errors are cleared by the next valid key sequence. The instrument also detects overall hardware fault conditions such as calibration faults, synthesizer unlocked or generator unleveled conditions. The resulting fault message indicates each hardware fault affecting one or more modes of operation. This leads to the second area of fault diagnosis. At instrument switch-on or reset the microprocessor carries out 31 checks on itself and its interface with the rest of the instrument. When the **SELF TEST** key is pressed this is repeated and extended with an additional 33 checks that include tests on virtually all the analog functional blocks of the instrument (Fig. 3h). This is done by routing the calibration signal through each possible signal path in turn and checking the absolute power reading on the detector against predetermined limits. There are 15 such signal paths, each giving an independent fault code. All of the programmable attenuators and gain stages are also checked against each other for both functionality and accuracy. In this way faults can often be diagnosed to board level before the covers of the instrument are removed. The third area where software is used is in providing a set of specific test programs, accessed with an internal switch, that can be used to simplify the internal detailed fault-finding, readjustment and calibration of the instrument.

Acknowledgments

Clearly, many people are involved during the development cycle of a product of this size and complexity. Apart from the major contributions of the authors of the subsequent articles, credit must also be given to David Heath for his elegant yet simple product design of the 3724A, John Struthers for his work on front-panel design and the 3725A, and Arthur Thornton for the product design of the 3726A and its many plug-in filters. Bob Hoffman, John Fisher, Mike Crabtree and Robin Sharp contributed to the early definition phase with many marketing inputs, and Reid Urquhart helped with later phases and the 37018A. Wullie Keir and David Webster contributed throughout the development, adding to the instrument's serviceability and ease of production. Ian Howett's help was invaluable during the transfer to production.

Richard J. Roberts

Richard Roberts received his BSc and PhD degrees in 1967 and 1971 from the University College of North Wales. He joined HP in 1972 and for a while continued to work on microwave integrated circuits. He initiated the 3724A/25A Baseband Analyzer and contributed to various parts of its design as well as serving as project leader. In his spare time, Richard and his wife enjoy skiing and boardsailing. His other interests include woodworking, photography and learning to play a recently acquired electronic organ.

Design of a Precision Receiver for an Integrated Test Set

by J. Guy Douglas and David Stockton

THE HP Model 3724A/25A/26A Baseband Analyzer combines the functions of several normally separate instruments under control of an internal microprocessor for use in characterizing the baseband performance of microwave radio equipment. To do this requires a versatile receiver that can be reconfigured by the microprocessor. According to the measurement mode selected by the user, this receiver must function as a

- Wideband power meter that covers a frequency range from 20 Hz to greater than 18.6 MHz and a level range from +20 dBm to a typical noise floor of −76 dBm.
- Selective level meter that can measure true-rms power within a selected bandwidth of 40 Hz, 400 Hz, 1.74 kHz, or 3.1 kHz over a frequency range from 50 Hz to 18.6 MHz and a level range from +20 dBm to −130 dBm.
- Spectrum analyzer that covers a frequency range from 100 Hz to 20 MHz, handles an input level range from +20 dBm to −130 dBm, and has a dynamic range of 80 dB.
- White-noise loading receiver that can handle systems of up to 2700 telephone channels and measure noise power ratios (NPRs) from 0 to 67 dB and signal-to-noise ratios from −18.8 dBmOp to −85 dBmOp.
- High-level-user (HLU) detector that can scan up to 1800 channels within 80 seconds to locate signals above a selected threshold level.

A simplified block diagram of the precision receiver design meeting these requirements is shown in Fig. 1.

The wideband power meter path fulfills a function similar to that of the HP 3400A RMS Voltmeter and consists of fixed gain stages and switchable attenuators. The rms detector is the monolithic thermal converter first used in the leveling loop of the HP 3336A/B/C Synthesizer/Level Generator.[1]

The wideband power meter also serves a useful purpose in the other measurement modes since the microprocessor can monitor the broadband input power to the selective signal path to determine the optimum input attenuation setting. For reasons that will be discussed later the main input attenuator consists of 37.5 dB of attenuation in 2.5-dB steps. To avoid excessive autoranging time and relay switching, the processor measures the most recent approximation of the input power and calculates the new required value of input attenuation. This reduces the maximum number of input attenuator autoranging steps from 16 to 3.

The other functions are variations on the theme of wave analysis, so the heart of the structure is an autoranging superheterodyne receiver capable of making frequency-selective measurements over a wide frequency range.

However, there are several conflicting design requirements.

A selective level meter measures the power of input signal frequency components within the bandwidth of an IF (intermediate frequency) filter whose frequency response has a flat top and steep sides for high selectivity. The detector must be true-rms and have a resolution of 0.01 dB.

A spectrum analyzer presents continuously updated data representing a wide range of frequency components whose levels may differ considerably. The accuracy and resolution are less important than those of a selective meter. The IF filter frequency responses must be Gaussian in shape for speed of settling and the detector must be fast-responding and logarithmic, providing a wide dynamic range.

A white-noise loading receiver computes the ratio of two measurements of the noise power in a 1.74-kHz channel at a predetermined frequency.* The first measurement is made with the noise generator's bandstop or slot filter out of the circuit. The second is made with the slot filter in the circuit. The ratio of the first measurement to the second is known as the noise power ratio (NPR). This receiver requires a flat-topped, steep-sided 1.74-kHz IF filter. The detector should have a measurement resolution of 0.1 dB together with good averaging to provide a result stable enough for display purposes. Of paramount importance, however, is the need to maintain the integrity of the signal as it passes through the receiver. The receiver circuitry is essentially a continuation of the system being measured and as such must have very low thermal noise and low intermodulation distortion (i.e., good back-to-back NPR) so that its performance is not significant in relation to the system being tested.

The back-to-back NPR performance of a white-noise loading generator and receiver should be greater than 67 dB. This assures that the test equipment will therefore contribute less than 0.4 dB error to the NPR measurement of a typical single-hop radio link, which will achieve a 55 to 57 dB NPR.[2]

With the above design requirements and constraints in mind, let us discuss in more detail the important sections of the versatile precision receiver used in the 3724A/25A/26A Baseband Analyzer. These sections (see Fig. 1) are the front end, first mixer, first LO (local oscillator), IF stage, programmable-gain amplifier, selective detector, and tracking calibration amplifier.

An Accurate, Low-Distortion Front End

The major objective in the design of the receiver's front end was to create a selective-level-meter type of structure with flat, broadband input circuits to facilitate accurate

*Bell, CCITT, CCIR and Intelsat have defined a standard range of slot frequencies covering all radio basebands from 12 kHz to 12360 kHz. For example, CCIR slot frequencies are 70, 270, 534, 1248, 2438, 3886, 5340, 7600, and 11700 kHz.

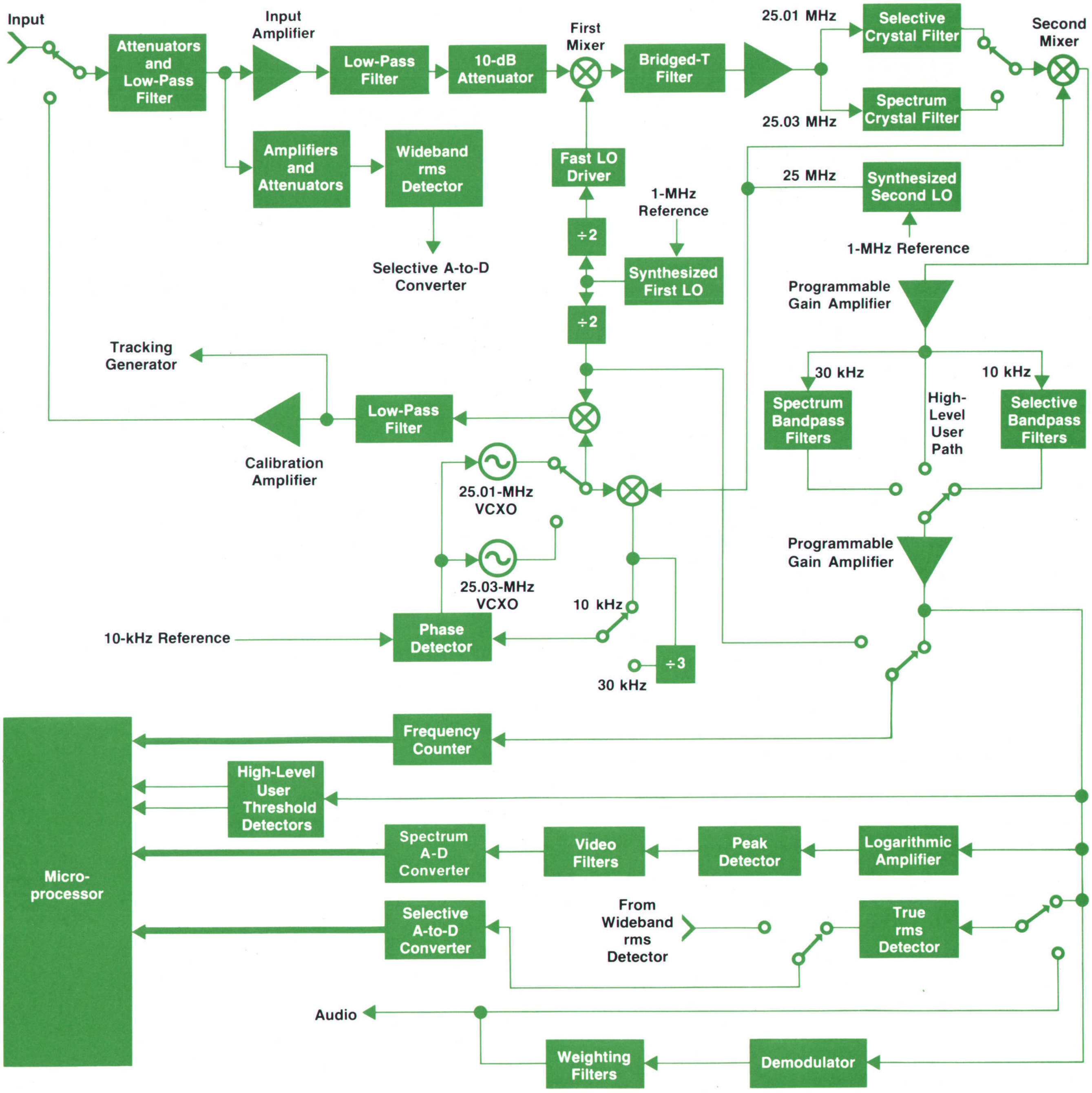


Fig. 1. *Block diagram of receiver used in the 3724A Baseband Analyzer.*

measurements over a wide frequency range. However, it also had to be sufficiently low in distortion and have a sufficiently low noise floor to permit white-noise loading measurements without the need for a range of bandpass filters at the receiver input. Such bandpass filters would not only be expensive and bulky, but would introduce difficult switching problems in an effort to achieve good frequency flatness and input return loss along with accurate level measurement.

Existing selective level meters achieve NPRs from 55 to 60 dB on a 2700-channel noise band. On some selective level meters it is possible to select either low-noise or low-distortion performance. Obtaining the best noise floor may allow the presence of large signals to introduce some distortion, and conversely, a poorer noise floor results when low distortion is selected for signal integrity.

To make white-noise loading measurements (see article on page 6), we require low noise and low distortion simultaneously. The slot-out reference level is measured in a 1.74-kHz bandwidth, so for a 2700-channel noise loading signal:

$$P_{in} = 10\log\left(\frac{12360-316}{1.74}\right) + \left(\begin{matrix}\text{slot-out}\\ \text{reference level}\end{matrix}\right)$$

$$\therefore P_{in} = 38.5 \text{ dB} + \text{slot-out reference level} \qquad (1)$$

where P_{in} is the total input power to the front end when measuring a 2700-channel noise band (316 to 12,360 kHz).

This means that to measure an NPR of 70 dB the slot-in power level measured by the receiver must be 108.5 dB below the broadband input power; that is, the dynamic range required of the receiver is almost 110 dB.

The baseband analyzer has a noise figure of 14 dB and the noise floor is typically −127.5 dBm (1.74 kHz). From Fig. 2 it can be seen that P_{in} (with zero input attenuation) must be greater than −14 dBm for the measured signal to exceed the thermal noise of the receiver by 5 dB when measuring an NPR of 70 dB. The input attenuator uses 2.5 dB resolution as mentioned earlier because the input level must be tightly controlled to maintain the correct compromise between headroom above the noise floor and tendency to overload the input amplifier and mixer.

Ideally the attenuators should be followed by a high-gain (say 15 dB), low-noise amplifier; this would improve the receiver's noise figure dramatically. However, with an expected maximum input power of −13 dBm this would mean driving the first mixer with a very high signal level. A compromise gain of 6 dB was implemented in the input amplifier which is optimized for low distortion with particular attention paid to even-order distortion. Two identical amplifiers are connected such that the bottom amplifier derives its input from the output of the top amplifier. The differential output of the top and bottom amplifiers across a 75-ohm resistor forms the final output. A characteristic of this design is that a measure of even-order distortion cancellation is achieved. At maximum drive level the two-tone intermodulation products are less than −110 dB (third order) and −90 dB (second order) relative to the wanted signals at the output of the amplifier. The amplifier, if measured in isolation, typically achieves an NPR of 70 dB at a level 10 dB above the maximum encountered in normal receiver operation.

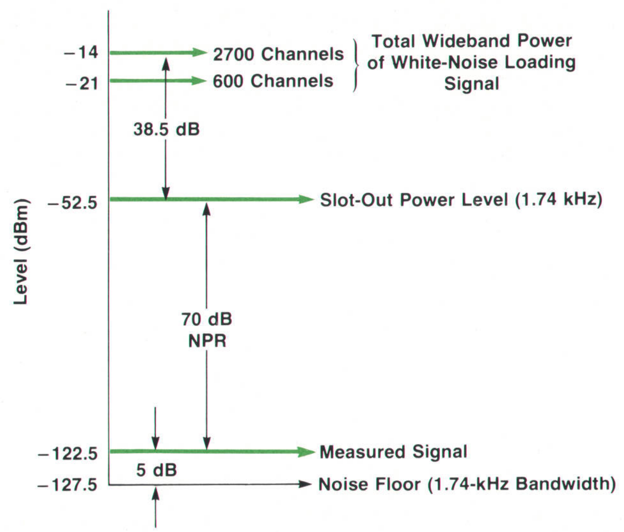


Fig. 2. *Given the receiver noise floor and the required NPR of greater than 70 dB, the total input power to the receiver's front end can be defined as indicated above.*

Two inductors, whose polarities are connected in series opposition from the virtual ground of the top amplifier to chassis ground, provide a dc return for the input (the attenuators are also dc coupled). This is necessary for white-noise testing of certain intermediate test points on some radio designs that require dc bias from the load device. A threshold detector protects the baseband analyzer input by allowing up to ±100 mA of bias current to flow before energizing a protection relay to open-circuit the input.

The input amplifier is preceded by a fifth-order elliptic low-pass filter and followed by a ninth-order elliptic low-pass filter. The filters provide first-IF rejection at 25.01 and 25.03 MHz, and first-IF image rejection in the 50-to-70-MHz range.

First Mixer

Much development effort was concentrated on the design of the first mixer and its associated circuitry. The IF output is derived in the usual way from the difference between the local oscillator frequency and the baseband input signal spectrum so that further signal processing may occur at a fixed frequency. The character of the baseband signal converted to IF must be preserved. However, since the derivation of the IF output is a nonlinear differential process, it is not unlikely that other undesirable, nonlinear products will appear at the IF.

The first mixer in the baseband analyzer is basically a conventional class II (type 2) Schottky-diode ring mixer (Fig. 4).[3] The major sources of signal impairment in a mixer of this type are nonlinearity of the diodes, phase modulation of the LO switching signal by the baseband signal at the instant of switching, and phase noise on the LO, both close-in and far-out. The first two sources result in intermodulation distortion even with a perfectly clean LO while the third introduces sidebands onto the IF signal by a reciprocal mixing process.

Ideally at any instant in time the LO waveform biases two diodes (e.g., D1 and D2) to conduct with a low linear resistance while the other two diodes (D3 and D4) are reverse-biased and exhibit zero conductance. In practice, when reverse-biased, the diodes behave as voltage-dependent capacitors. Forward-biased diodes behave according to the following equation:

$$V \approx (kT/q) \ln (I/I_s) \qquad (2)$$

where I_s is the reverse saturation current of the diode.

By considering the signal flow during one-half cycle of the first LO it can be shown that the two-tone third-order intermodulation intercept (3OI)* for this mixer is[4]

$$3OI = 10\log\left[2000\,(R_s+R_L+R/2)^3\,(8q/R_skT)\,I^3\right] \text{ dBm} \qquad (3)$$

where R_s is the IF terminating impedance, R_L is the

*3OI is the theoretical signal power level at which the $2f_2 - f_1$ and $2f_1 - f_2$ (third-order) intermodulation products have the same power as the combined input power of f_1 and f_2.

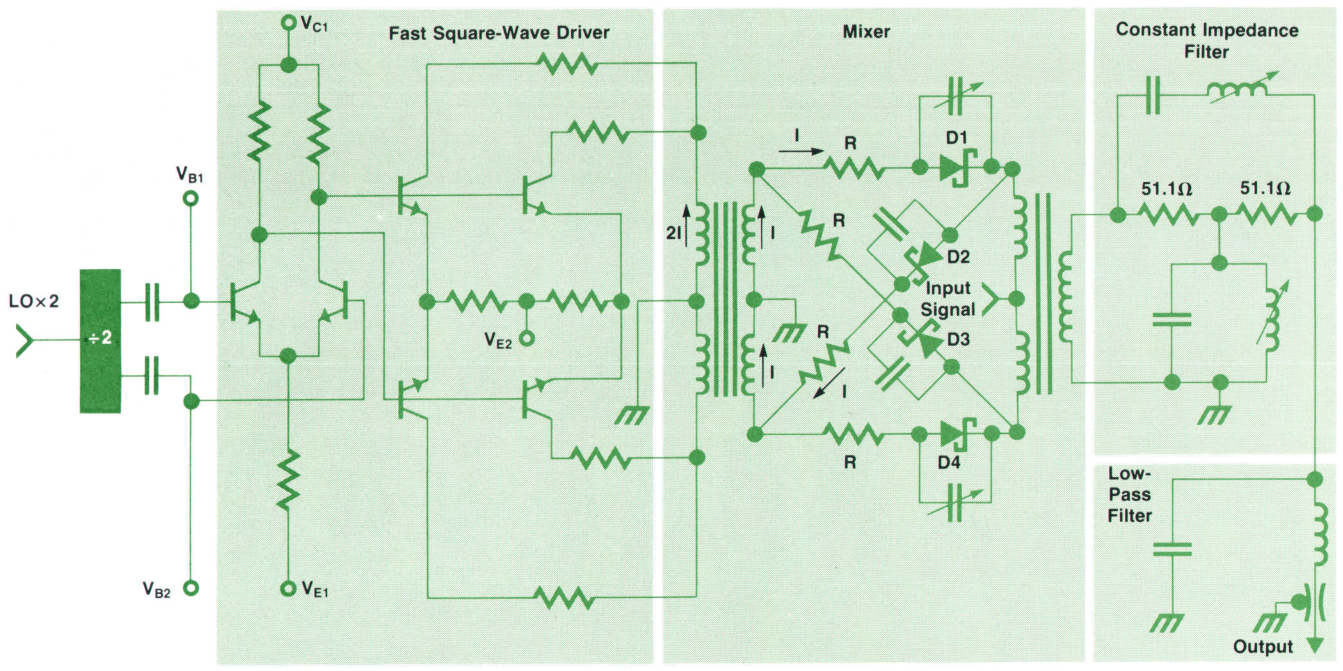


Fig. 3. *The first mixer in the baseband analyzer is a class II Schottky-diode ring mixer.*

baseband terminating impedance, R is the linearizing resistor in series with the diodes, and I is the diode current due to the first LO. From this it can be seen that the first LO switching current I should be as high as possible to force the on diodes as far into their linear region as possible. A diode current of 45 mA is used in the baseband analyzer's first mixer.

Two overload conditions can occur.[5] The first is a result of the tendency of the peak signal voltage to forward-bias a reverse-biased diode and the second is a result of the tendency for the peak signal current to reverse-bias a forward-biased diode. R appears in the above equation because it reduces the signal current in a forward-biased diode, alleviating the second overload condition. Increasing R has the additional benefit of counteracting the first overload by increasing the LO voltage across reverse-biased diodes. However, increasing R above a certain value eventually becomes counterproductive because it also increases the conversion loss of the mixer as follows:

$$\text{Conversion loss} = 3.92 + 20\log\left[1 + \frac{r_f + R}{2(R_s + R_L)}\right] \quad (4)$$

$$+ \text{ transformer loss (dB)}$$

where r_f is the on resistance of a forward-biased diode.

A compromise of R = 50 ohms increases the conversion loss from 6 to 8 dB, but also increases the voltage across a reverse-biased diode from 1 to 3V, introducing a very significant improvement in 3OI.

Even with ideal diodes, unless the switching caused by the LO waveform is instantaneous, the signal voltage will modulate the switching function at the instant of switching (see Fig. 4). It can be shown that the two-tone 3OI caused by the finite rise time t_r of the LO waveform is[4]

$$3OI = 10\log\left[2000V_c^2/(t_r^2\ \omega_{LO}^2\ R_s)\right]\ \text{dBm} \quad (5)$$

where V_c is the peak-to-peak LO voltage, and ω_{LO} is the angular frequency of the LO.

As expected this equation shows that the highest ratio of the LO voltage to its rise time should be sought. The first mixer of the 3724A uses two high-frequency emitter-coupled transistor pairs connected in parallel to switch a primary current of 90 mA in the transformer. This results in a secondary square-wave voltage of 6V peak-to-peak with approximately 1-ns rise time. This yields a theoretical 3OI from equation (5) of +44 dBm at an LO frequency of 37 MHz (approximately 12-MHz baseband frequency). This is a good reason for keeping the first IF, and consequently the first LO, as low in frequency as possible to reduce the significance of the rise time of the LO.

All of the measures mentioned so far for improving

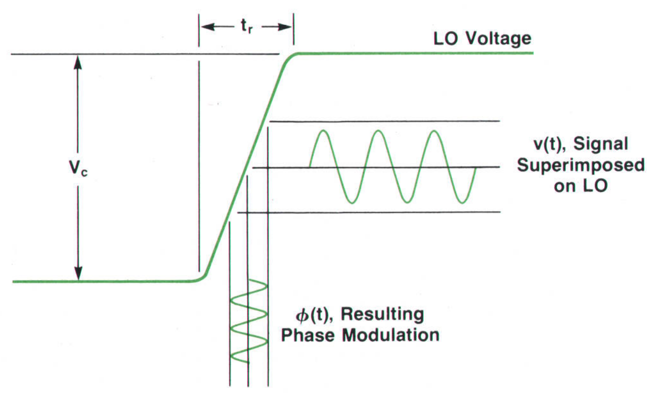


Fig. 4. *Phase modulation of the LO switching function by the input signal is caused by the finite rise time of the function.*

third-order distortion will also improve second-order distortion, provided the balance of the mixer is preserved. Any unbalance, however, will degrade the second-order distortion behavior which is also crucial to good NPR performance. Odd-order distortion levels are more important for a wave analyzer since these products lie close to the wanted signals and could fall within the IF bandwidth.[4] For good NPR performance, however, even-order distortion levels must also be low since the noise signal is broadband and sum and difference products will appear at all slot frequencies.

A large contribution to second-order distortion comes from capacitive imbalance in the reverse-biased diode pair of the ring mixer. The signal voltage modulates the reverse bias voltage of these diodes which in turn modulates the diode capacitances. It can be shown that the second-order intercept 2OI for this configuration is given by:†

$$2OI = 10\log\left[4000V_{DC}^2/(R_s\,(\Delta C)^2\,(r_f+R)^2\,\omega^2)\right]\ \text{dBm}$$

where V_{DC} is the diode contact potential plus half the LO peak-to-peak voltage and ΔC is the capacitive imbalance between the two reverse-biased diodes (at zero reverse bias). The diode capacitance is assumed to be inversely proportional to the square root of the reverse bias voltage.

For example, a capacitive imbalance of 0.5 pF between any diode pair will result in second-order intermodulation products at a −80-dB level for this mixer at normal drive levels. The baseband analyzer uses balancing capacitors in parallel with the diodes (see Fig. 3) to cancel distortion introduced by differences in the diode capacitances, transformer imbalance, and circuit stray capacitance.

To prevent this balance from being upset by variations in the even-order harmonic content of the LO, an emitter-coupled-logic (ECL) divide-by-2 circuit is incorporated immediately before the fast LO driver.

Ideally a narrowband crystal filter is connected directly to the IF port of the mixer to reduce the effect of distortion in following stages. However, for good conversion-loss flatness and high baseband-port return loss, it is important to terminate the IF port in a broadband, nonreactive load.‡ Instead, a compromise solution of a constant-impedance bandpass filter is adopted in the 3724A. This bridged-T circuit (see Fig. 3) terminates the IF port correctly while offering some bandwidth reduction (2.5-MHz 3-dB bandwidth). This has important advantages. It simplifies the first-IF amplifier design because return loss must be high only within ±2 MHz of the IF and the signal bandwidth reduction achieved by the bridged-T prevents the possibility of second-order distortion in the first-IF amplifier. The noise figure of this amplifier is 2.5 dB and its gain is 22 dB, both essential to achieving the necessary overall 14-dB noise figure of the receiver. Its two-tone third-order distortion products are −100 dB at normal maximum operating power, again necessary only over the bandwidth of the bridged-T filter.

The 10-dB switchable attenuator immediately preceding the first mixer is switched out in white-noise loading measurement modes for the best noise figure, and switched in when best flatness is required because the input port of the mixer then sees a good termination for frequencies well beyond the baseband upper limit. Thus the attenuator is permanently selected in the baseband-response and selective-level modes when the measured signal level is greater than −85 dBm. The processor automatically switches the attenuator out if the measured level is less than

†2OI is the theoretical input signal power level at which the $f_1 \pm f_2$ (second order) intermodulation products have the same power as the combined input power of f_1 and f_2.

‡Other mixer products are then terminated properly at the IF, in particular the upper sideband and the LO frequency.

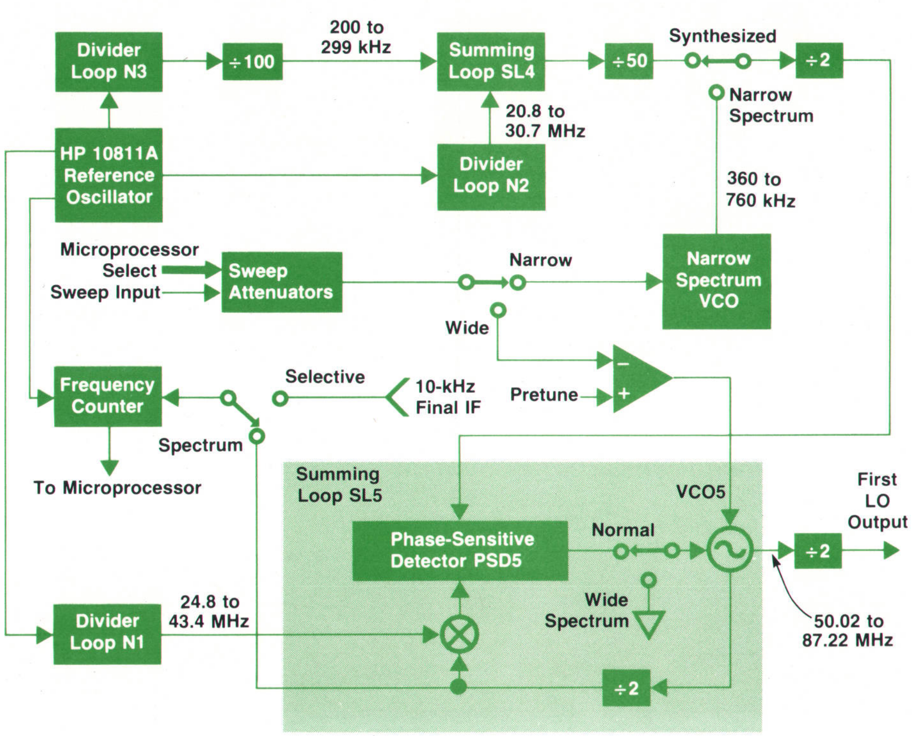


Fig. 5. *Block diagram of first-LO section.*

−85 dBm (approximately), provided the broadband input power threshold has not been exceeded. In this way a good compromise between flatness and signal-to-noise ratio is achieved.

In the noise measurement modes, the drive level into the first mixer, as controlled by the input attenuator autorange firmware, is dependent on the channel capacity of the incoming signal. Since, for example, a 600-channel noise signal constitutes a lower broadband input power for a given power level with the slot filter out than a 2700-channel signal (see Fig. 2), the drive level to the first mixer is reduced by the autorange firmware as the user-entered channel capacity value decreases. This has the effect of reducing the distortion produced in the input circuits without compromising the headroom above the noise floor (in fact it is sometimes improved). In this way the back-to-back NPR of the baseband analyzer is optimized for each channel capacity value.

In the selective-level and spectrum analysis modes the character of the input signal is undefined so the highest-power/Hz signal must be assumed by the autorange firmware and the drive level to the mixer is fixed at −22 dBm.

The overall effect is a receiver front end that has a typical back-to-back NPR of 71 dB with 2700-channel loading using an external local oscillator (e.g., an HP 8640B AM/FM Signal Generator) in the noise measurement modes and typically 0.05-dB flatness from 10 kHz to 17 MHz in the selective-level and baseband response modes.

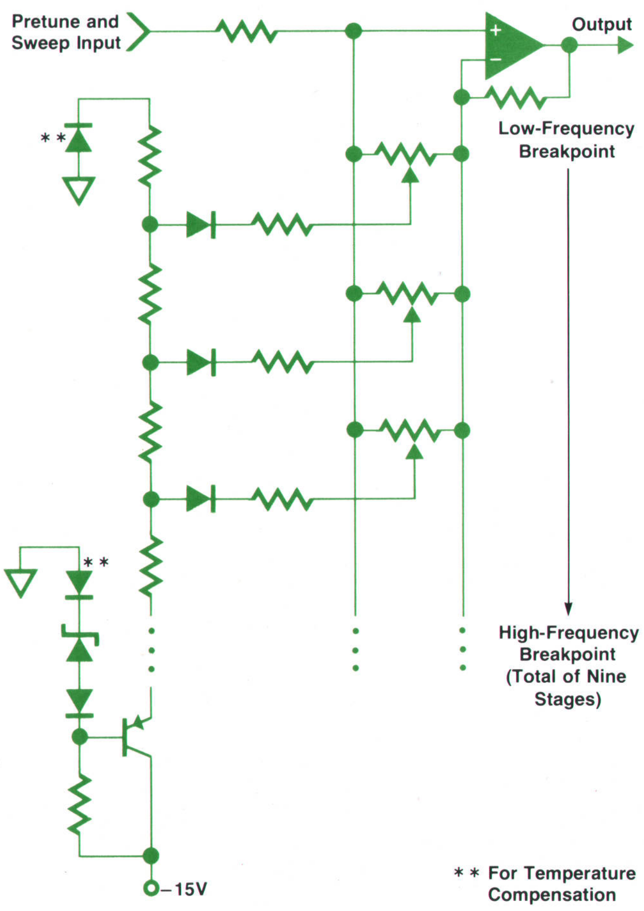


Fig. 6. *The shaping network shown above is used to compensate for voltage-to-frequency nonlinearity of VCO5.*

First Local Oscillator

The first LO for the receiver has to operate as an accurate, low-noise oscillator for all selective measurements of the instrument and as a continuously swept oscillator for the spectrum analyzer measurements. This leads to three modes of operation for the first LO (Fig. 5):

- As a conventional synthesizer with three divider loops and two summing loops to give an output frequency between 25.01 MHz and 43.61 MHz corresponding to the 25.01-MHz first IF.
- For spectrum frequency spans greater than 100 kHz, VCO5 is swept open-loop by the appropriate sweep voltage together with the pretuned center-frequency voltage.
- For spectrum frequency spans less than or equal to 100 kHz, the N1 and SL5 loops are phase-locked, giving accurate 100-kHz steps at the output. The narrow-spectrum VCO (voltage-controlled oscillator) is swept to provide interpolation between these points and replaces the normal signal fed to the phase-sensitive detector PSD5.

The reference frequency for the synthesizer is an HP 10811A Crystal Oscillator which has an aging rate of $<1 \times 10^{-7}$ parts/year. In the spectrum analysis modes the oscillators are running open-loop and therefore need some means of locking the center frequency to the requested frequency. This is achieved by momentarily halting the sweep at its center and counting the local oscillator frequency. The microprocessor compares this counted frequency (minus 25.03 MHz) with the requested frequency and, if the difference is significant, adjusts the pretune data to minimize the error. This process is carried out on every sweep so that the effects of temperature, aging, and digital-to-analog conversion inaccuracies are removed.

The narrow-spectrum VCO is a current-controlled multivibrator whose current-to-frequency output is linear over the required range of 360 to 760 kHz. However, VCO5 does not have sufficient linearity for wide spectrum spans. To compensate the nonlinear voltage-to-frequency characteristic of VCO5, a shaping network was designed with a minimum of required adjustments and interaction. This network consists of nine fixed-position breakpoints. One potentiometer for each breakpoint can increase or decrease the gain without the need for jumpers or switches (Fig. 6). When set midway, a breakpoint introduces identical attenuation into both the input and feedback networks and hence has no effect. Moving the respective potentiometer in either direction unbalances this situation and either increases or decreases the overall gain in that breakpoint's active region. These gain adjustments do not interact provided that they are adjusted in sequence from low-frequency to high-frequency breakpoints.

The purity of the local oscillator is the final critical factor for good NPR measurement. Referring to Fig. 7 we can discriminate between close-in and far-out phase noise. Far-out noise from both sidebands of the LO will mix with the white-noise loading signal, effectively increasing the noise floor in the IF. To achieve a back-to-back NPR of 70

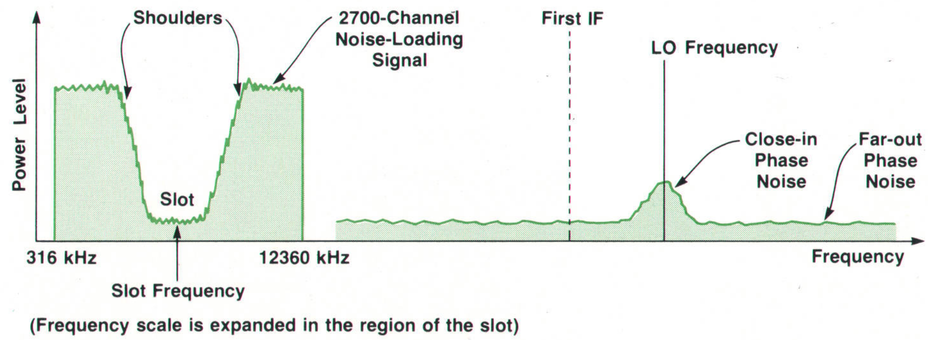


Fig. 7. *The LO mixes the slot frequency into the first IF. The shoulders of the slot mix with the close-in phase noise of the LO to appear in the IF and the rest of the noise-loading signal mixes with the far-out noise sidebands of the LO and also appears in the IF.*

dB, the far-out noise integrated over a 2700-channel bandwidth on either side of the LO must be less than −70 dBc. That is, the LO far-out noise is

$$\leqslant -70 - 10\log\left[(12360-316)\times 2\times 10^3\right]$$

$$= -143.8 \text{ dBc/Hz single sideband (SSB)}$$

Using a low-noise oscillator that generates a large output signal and a low-noise buffer amplifier, the baseband analyzer's first LO achieves typically −157 dBc/Hz after the divide-by-2 stage.

The major NPR degradation comes from mixing between close-in phase noise (less than ±15 kHz from the carrier) and the noise shoulders of the bandstop slot filter. The 3-dB bandwidths of the generator bandstop slot filters vary considerably. The narrowest is ±4 kHz, so phase noise in this area is not important because there is no signal that can mix with this noise and appear in the IF. The responses of the most significant loops have been optimized for NPR measurement by shaping the noise within the loop to be as close-in to the carrier as possible.

In the ±4-to-±15-kHz range the LO noise is typically −115 dBc/Hz (SSB) which would result in an NPR of 72 dB for the narrowest bandstop filter, assuming a perfect receiver. The combination yields a receiver with a synthesized LO that achieves an NPR of typically 69.5 dB in the 3886-kHz slot* and 71.5 dB in the 2438-kHz slot** with 2700-channel loading. When 900-channel loading is used, the 2438-kHz slot measures typically 73 to 74 dB NPR.

IF Stage

In the IF structure, the major design conflict centers around the type and range of IF filters required. For spectrum measurements over a 20-MHz span, 10 kHz is the minimum resolution bandwidth that will result in an acceptable sweep rate, given a display with digital storage. The filters available in the 3724A range from 100 Hz to 10 kHz in a 1, 3, 10 sequence. It was desirable to realize these using a conventional five-stage inductor-capacitor (LC), synchronously tuned, variable-Q filter. A final spectrum IF of 30 kHz was chosen because a 100-Hz bandwidth is not practical for higher intermediate frequencies using LC filters.

The IF filters required for selective-power and white-noise loading measurements, 3.1 kHz, 1.74 kHz, 400 Hz and 40 Hz, are high-order filters to achieve a rapid cutoff rate. Consideration of available coil Q rules out the possibility of a 30-kHz IF. Instead, the 40-Hz filter† is centered on a 485-Hz IF, which requires an extra mixing stage from the 10-kHz IF at which the other three filters are centered. These three filters are LC elliptic or Chebyschev designs.

Therefore, referring to Fig. 1, the baseband analyzer has two second IFs, one at 10 kHz for selective-power-measurement-path filters and one at 30 kHz for spectrum-analysis-path filters. Crystal filters used in the first-IF structure are similar to those used in the HP 3586A/B/C Selective Level Meter; they provide image rejection close to the passband and so allow mixing down to 10 or 30 kHz directly.[6] The necessity for two second IFs implies the need for two first IFs. An alternative would have been to switch the second LO between two frequencies locked to the reference frequency, but the image rejection and passband requirements for the first-IF filter would then have been extremely difficult to achieve. The selective-power-measurement-path crystal filter (Fig. 8a) has a 3-dB bandwidth of approximately 6 kHz centered on 25.01 MHz and has greater than 85 dB rejection at 24.99 MHz (Fig. 8b). The spectrum-analysis-path filter has a 3-dB bandwidth of about 17 kHz centered on 25.03 MHz with its image rejection of 85 dB at 60 kHz away (24.97 MHz). The additional advantage of using a crystal filter is that later stages of signal processing have only a narrowband signal to deal with and distortion is less of a problem following the filter.

To detect high-level users in the channel traffic there is an initial fast measurement loop that uses the selective-power-measurement-path crystal filter in the first IF and bypasses all the second-IF filters. Because this filter has a 6-kHz bandwidth, the synthesizer can be tuned in 6-kHz‡ steps. Together with the fast-responding high-level-user threshold detector, this allows the instrument to scan live traffic on an 1800-channel radio link to identify high-level users in about 80 seconds.

Programmable-Gain Amplifier

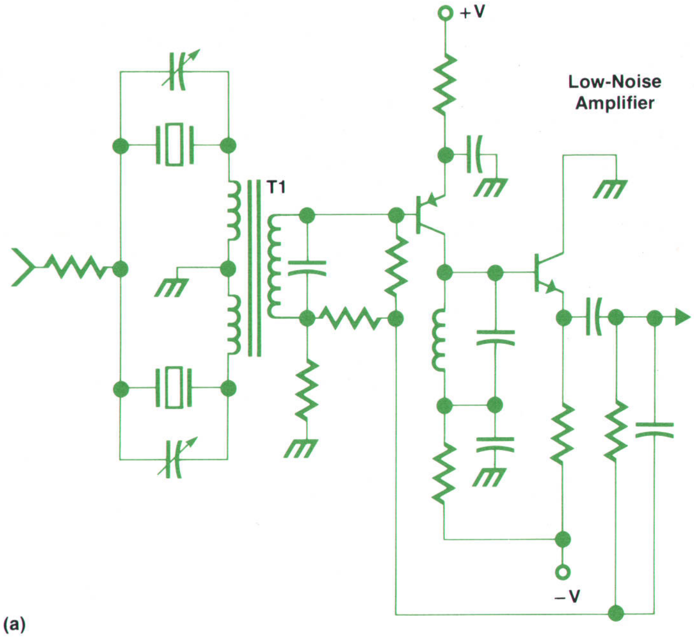
For measurement accuracy over a wide range of levels, an accurate programmable-gain amplifier in the IF stage is essential (see Fig. 1). In the 3724A this amplifier operates at

*The 3886-kHz slot is the lowest-frequency filter to be realized as a crystal filter and hence has a very narrow 3-dB bandwidth.

**The 2438-kHz slot is the highest-frequency filter using inductors and capacitors and hence has a wide 3-dB bandwidth.

†This is an active filter similar to that used in the HP Model 3745A Selective Level Measuring set, see H.P. Walker, "Designing Precision into a Selective Level Measuring Set," Hewlett-Packard Journal, Vol. 27, no. 5, January 1976, p. 11.

‡This IF bandwidth allows the threshold to be set within 6 dB of the normal maximum channel loading level. A wider bandwidth can result in continuous false triggering due to the combined power of, say, three normally loaded channels marginally exceeding the threshold.



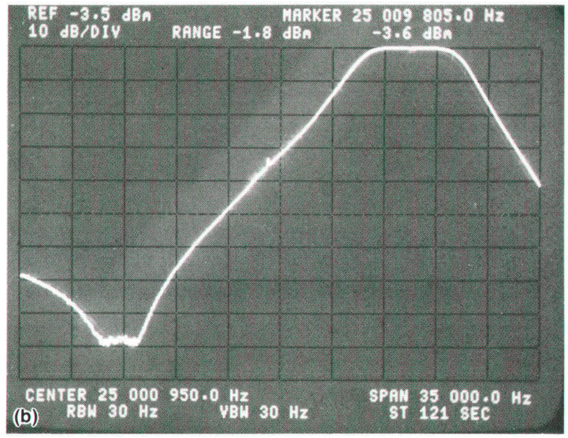


Fig. 8. *(a) One section of the two-stage first-IF crystal filter in the selective circuit path. T1 is both a component of the lattice structure of the filter and a matching transformer for the low-noise amplifier that follows. (b) Filter response.*

the 30-kHz and 10-kHz second IFs and the 485-Hz (for the pilot filter) third IF, providing a range of gain from 0 to 100 dB in 10-dB steps. Up to 20 dB of gain precedes the second-IF filters, and 80 dB is available following the filters.

The structure consists of fixed-gain amplifiers buffering four ratio transformer stages (Fig. 9). This approach has several benefits: the need for precision resistors is eliminated, and the problems associated with switched-gain amplifiers, such as changes in loop gain and on-resistance compensation of the electronic switches, are nonexistent.

The only constraint on the fixed-gain amplifiers is that their ratio of input to output impedance must be at least 10,000:1 (for 0.001-dB error in a 40-dB step). This is relatively easy to achieve even at 30 kHz. The transformers use a conventional winding technique with taps at exactly the −20-dB and −40-dB points (−10 dB and −20 dB on two of the transformers). It is not necessary to use two-stage transformers or other techniques to obtain errors of less than 1 in 10^4 at 10 kHz.[7] Transformer errors are proportional to the square of frequency and come from leakage inductance and unbalanced self-capacitance of the windings. Hence, the errors are significantly increased at 30 kHz, the spectrum analysis IF. This is not critical, however, since the display errors in the spectrum analysis mode are at least an order of magnitude higher.

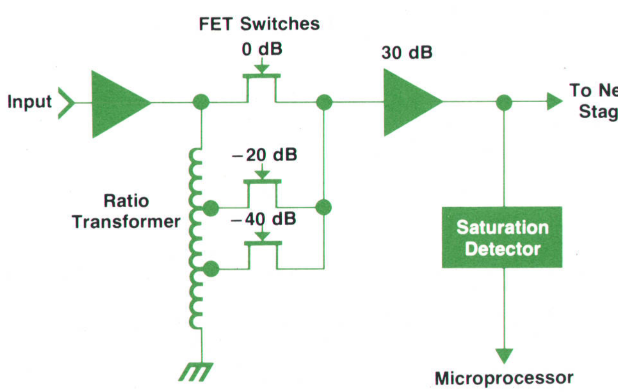


Fig. 9. *One section of the programmable-gain amplifier.*

When the −40-dB transformer tap is selected the accuracy is affected by the off-to-on resistance ratio of the switches. The FET (field-effect transistor) switches used give an isolation of 130 dB at 10 kHz.

Three saturation detectors provide the microprocessor with information about how much gain to remove when the IF signal undergoes a large change in level. Consequently the IF gain can autorange very quickly. Over an 80-dB range the programmable-gain amplifier error is no more than 0.01 dB at 10 kHz.

Selective Detector

The selective detector is a commercially available integrated circuit that uses a squaring technique to produce a dc output proportional to the true-rms value of the input. A logarithmic amplifier and peak detector with video filtering selectable up to 5 kHz provides a wide dynamic range, fast-responding detector for spectrum analysis measurements at the 30 kHz IF.

The selective detector is operated over a 14-dB range, so its linearity is of importance. The transfer characteristic of an ideal detector is a straight line with a slope of 45 degrees passing through the origin. In practice there is normally an offset which can be nulled by adjustment at zero input. This, together with a single-point calibration (e.g., CAL1 in Fig. 10) somewhere in the detector range, achieved by applying a signal of known value to the input, completely characterizes the detector transfer function.

However, the bandwidth of this device is dependent on the signal level applied. For example, at 3V rms the input response peaks at 500 kHz, but at 10 mV rms the response peaks at 6 kHz. The frequency response is also temperature dependent. Fig. 10 shows typical transfer characteristics for this device at temperature extremes (the nonlinearity is exaggerated). Severe non-monotonic behavior occurs at

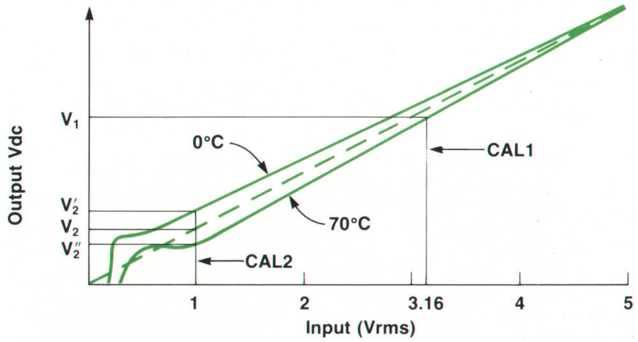


Fig. 10. *The selective detector has a nonlinear characteristic (exaggerated here for illustrative purposes) that is corrected by a calibration procedure discussed in the text.*

low input levels. The detector is used between 1 and 5V and clearly, nulling the offset at zero input is unlikely to make the desired correction at 1V.

Instead a two-point calibration is used. CAL1 is generated by connecting the calibration signal into the receiver in the normal way. A second point CAL2, still within the monotonic range of the detector characteristic, is applied by reducing the second-IF programmable gain by 10 dB. Because of the ratio transformers this is accurate within ±0.005 dB. The system's microprocessor then iteratively adjusts the offset via a digital-to-analog (D-to-A) converter until the detector reads a 10-dB±0.01-dB difference between CAL1 and CAL2. A firmware logarithm routine is implemented in the processor to convert the linear detector result to appropriate units for measurement and display.

The two-point calibration occurs at turn-on and when the **CAL** key is pressed. In this way linearity is maintained within 0.02 dB over the temperature range.

Tracking Calibration Amplifier

The calibration amplifier produces a signal of fixed level (−30 dBm) at the frequency to which the receiver is tuned. This signal is sufficiently accurate and stable to allow calibration of the receiver signal path. The calibration process is initiated whenever the **CAL** key is pressed or automatically by the microprocessor when the receiver signal path is changed in any way (e.g., change of measurement filter, change of input impedance).

Because the calibration signal tracks the tuned frequency it can be used to correct for nonflatness in attenuators, particularly close to the band edges. In the baseband analyzer the input attenuator autorange state is held (up to a maximum of 30 dB) during calibration, and the ratio transformer IF gain is introduced as required to return the signal to the correct level at the detector. In this way the precise attenuator configuration used for measurement is compensated for any attenuation and flatness errors that may be present.

The calibration signal is a square wave whose fundamental frequency component is used to calibrate the receiver. The fundamental must therefore be flat with frequency. From Fourier analysis it can be shown that for a square wave of frequency f and finite rise time t_r,

$$A(t_r)/A(0) = 20\log\left[\sin(\pi f t_r)/\pi f t_r\right] \text{ dB}$$

where $A(t_r)$ is the amplitude of the fundamental with rise time t_r, and $A(0)$ is the amplitude with zero rise time. Thus a rise time of 1 ns is required at 18.6 MHz to achieve a theoretical flatness of better than 0.01 dB.

In Fig. 11, CR1 and CR2 are Schottky barrier diodes whose minority carrier lifetime is much less than 1 ns. The reference current I is switched alternately through CR1 and

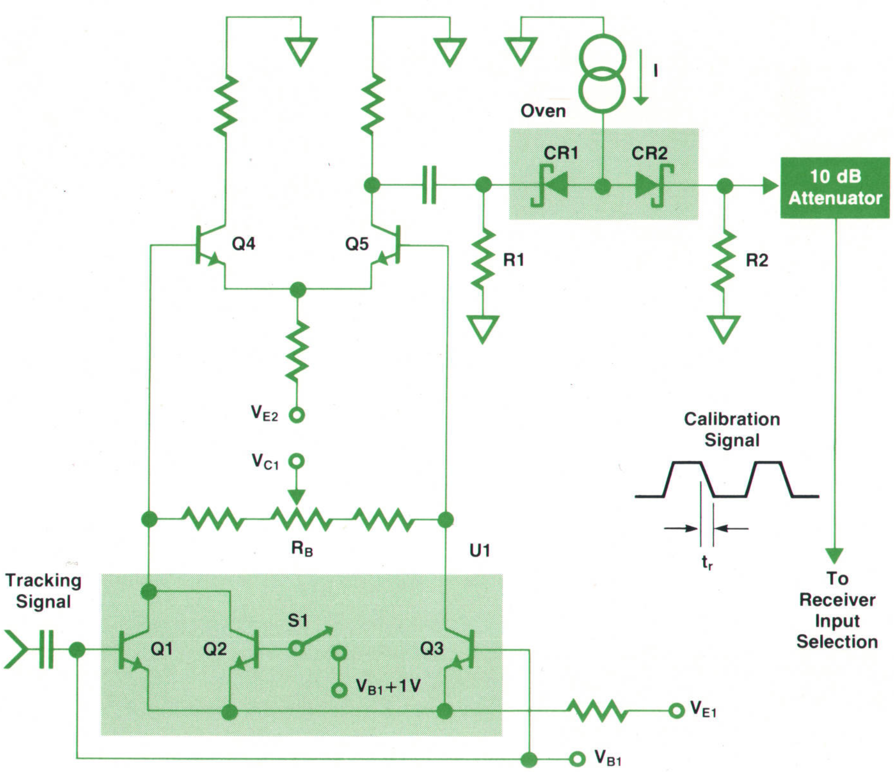


Fig. 11. *Simplified schematic of calibration amplifier.*

System Software Package for the Baseband Analyzer

The 3724A/25A/26A Baseband Analyzer was designed not only as an integrated measurement instrument but also as a powerful system component that can be controlled via the HP-IB.

Writing and debugging a dedicated software package for such an instrument can present the user with a significant development task. However, the 37018A System Software allows the user to make automatically controlled measurements immediately. The software is modular so that dedicated packages can be rapidly assembled, or different measurement or output formats added with minimal effort.

The system controller is an HP-85 (or HP-83) Personal Computer that gives a low-cost and flexible approach to data storage and hard-copy output. The software guides the user, via simple prompts, through the creation of a measurement file and the assembly of a sequence of up to 40 measurements. The data for each measurement is requested and checked for validity as the file is built up together with predefined limits for that measurement. Halting, continuing or branching may be performed as a result of each test which enables diagnostic measurements to be built into the file when an item under test fails one of the main measurements.

CR2 by the switching signal at the collector of Q5, which is an amplified version of the tracking signal. The result is a fast-rise-time square-wave voltage developed across R2 whose amplitude is controlled by I.

When S1 is closed, the calibration amplifier is turned off by Q2 which prevents Q1 and Q3 from conducting. Q1 through Q3 are all on the same integrated circuit (IC) in U1 so that, whether the calibration amplifier is on or off, virtually the same power is dissipated in the IC. In this way the mark-to-space ratio at the output, as set by R, stabilizes very quickly.

Using this calibrator, the 3724A can achieve typically ±0.035-dB absolute error over much of its measurement range in the selective-level mode.

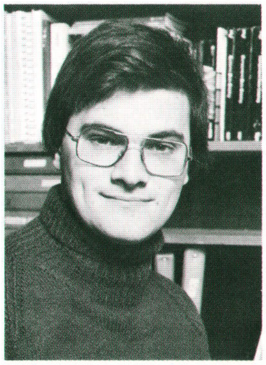
Acknowledgments

The crystal filters in the first IF and the selective filters in the second IF were designed by Stephen Biddle. Daya Rasaratnam designed the power supplies for the 3724A and the 3725A. Andy Batham made early contributions to the spectrum and selective detector circuitry. Thanks go to Eng Chong Meng for designing the wideband power section and contributing much hard work in various areas during prototyping. Keith Willox designed the demodulation section of the receiver. We wish to thank Hugh Walker for continuous help and encouragement throughout the project.

References

1. P.M. O'Neill, "A Monolithic Thermal Converter," Hewlett-Packard Journal, Vol. 31, no. 5, May 1980, p. 12.
2. CCIR Recommendation 399-1.
3. D. Cheadle, "Selecting Mixers for Best Intermod Performance," Microwaves, November 1973, pp. 48-52.
4. H.P. Walker, "Sources of Intermodulation in Diode-Ring Mixers," The Radio and Electronic Engineer, Vol. 46, no. 5, May 1976, pp. 247-255.
5. J.G. Gardiner, "Local-Oscillator-Circuit Optimization for Minimum Distortion in Double-Balanced Modulators," Proceedings of the IEE, Vol. 119, no. 9, September 1972.
6. P.L. Thomas, "A Programmable Selective Level Meter (Wave Analyzer) with Synthesized Tuning, Autoranging, and Automatic Calibration," Hewlett-Packard Journal, Vol. 31, no. 5, May 1980, p. 6.
7. W.C. Sze and F.R. Kotter, "The Design of Near-Perfect Instrument Transformers of Simple and Inexpensive Construction," Journal of Applied Measurements, no. 2, 1974, pp. 22-27.

David Stockton



David Stockton spent one year with HP while gaining his BS degree from Leeds Polytechnic University in 1976. He returned to HP and developed the synthesized local oscillators and spectrum filters for the 3724A Baseband Analyzer. He is now working part time to complete his MS degree. In his spare time he designs high-fidelity audio equipment, has built his Mini Cooper from a shell, goes sailing and spends many hours riding horses. When relaxing he enjoys reading and listening to music.

J. Guy Douglas

From Ayrshire in Scotland, Guy Douglas joined HP in 1973, working first in the systems department. After joining R&D he contributed to the early development of the baseband analyzer, concentrating on the front end of the receiver. This led to his supervising the entire receiver hardware development. Later he became project leader and guided the 3724A/25A/26A into production. Guy is married and has two young daughters. He is a graduate of Glasgow University (1973) and Heriot-Watt University, Edinburgh (an MS degree in 1977). A Church of Scotland elder at Dalmeny Kirk, which celebrated its 850th anniversary in 1980, Guy organizes a youth group there. He and his wife play guitar and sing together. Besides activities for various Christian groups, Guy enjoys skiing and can be persuaded to play squash occasionally.

Control and Display System for a Baseband Analyzer

by Lawrence Lowe and Brian W. Woodroffe

THE HP Model 3724A/25A/26A Baseband Analyzer is a complex instrument that can be reconfigured to perform a variety of widely differing measurements, each of which requires extensive control, computation and display facilities (Fig. 1). The control and computational system is based on a 68B00 microprocessor linked to 56K bytes of program code in read-only memories (ROMs), 5K bytes of random-access memory (RAM), and memory-mapped I/O (input/output) facilities. The 3725A Display produces alphanumeric characters and flicker-free spectrum-analyzer-type outputs on a cathode-ray tube (CRT). Both the characters and the spectrum analysis trace are generated from data stored in RAM. The character data is placed in the RAM by the microprocessor while the trace data is loaded directly from the spectrum-analysis-mode analog-to-digital (A-to-D) converter although the computational system may override the A-to-D converter and load a computed trace. Because both the processor and display circuits require access to the display data RAM, the memory system is set up such that any part of memory may be accessed by either system.

Multiplexed Memory System

Memory sharing is achieved by the use of a multiplexed direct-memory-access (DMA) system that switches the memory data and address buses between the microprocessor and other memory accessing circuits once every processor clock cycle.[1] The operation of multiplexed DMA, as applied to the 68B00 microprocessor (Fig. 1) relies on the memory not being accessed by the processor during the period that system clock $\phi1$ is high. Thus the display and the spectrum analysis A-to-D converter may access memory when $\phi1$ is high and the microprocessor has access when $\phi2$ is high ($\phi1$ low). The advantage of multiplexed DMA over cycle-stealing or halting DMA is that the processor continues to operate at all times, the DMA operation being transparent. The display, A-to-D converter and microprocessor memory accesses are synchronized through a common reference clock. The high-speed 68B00 version of the 6800 allows the use of standard-speed ROMs even though the time available for access is reduced by the DMA system.

Memory Map

The memory system is designed to fit within the 64K bytes of memory that the microprocessor can address directly. Fig. 2 illustrates how the 64K memory is used and shows the memory sharing described above. The 8K-to-10K area has two devices mapped into it: a program ROM which is accessed by the processor during $\phi2$ high and a character ROM which is only accessed by the display during $\phi1$ high.

Display System

The 3725A Display may be considered as two systems in one. The first is the alphanumeric generator used to provide data output and assist user control, and the second is the analog display used to provide graphical output for the spectrum analysis and baseband response modes. By using a raster-scan system, sixteen rows of thirty-two characters or eight rows of characters and a digitally generated analog trace may be displayed over the full CRT face.

The sixteen rows of thirty-two characters on the CRT are represented by 512 memory locations in the main memory RAM. A character code written to one of these locations by

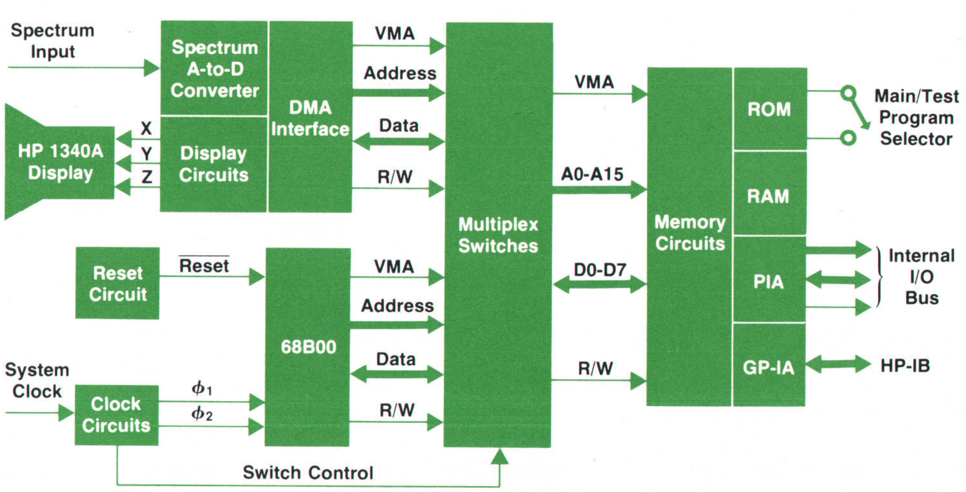


Fig. 1. *Block diagram of control and display system for the HP Model 3724A/25A/26A Baseband Analyzer.*

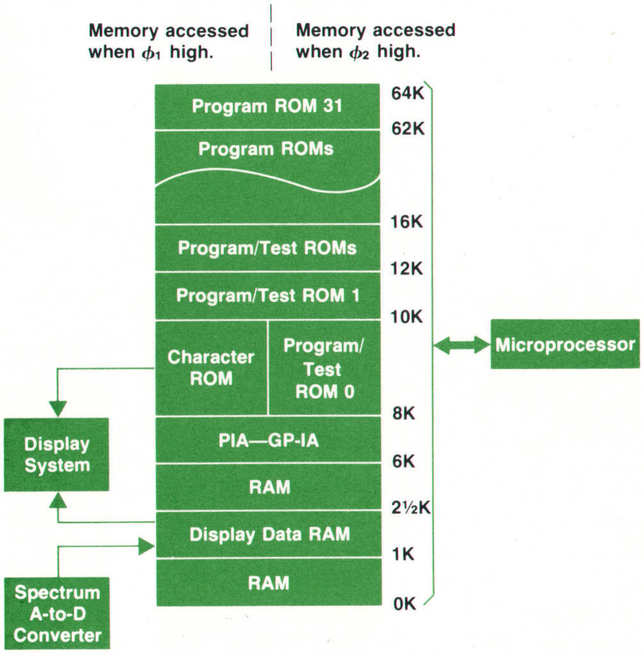


Fig. 2. *Memory map. An internal switch selects the test mode by deselecting the program ROMs from 8K to 16K and replacing them by 8K of test code.*

the processor will appear on the CRT within 15 milliseconds at the next screen refresh. Writing the character code into RAM is the last action the processor has to perform. The display system then accesses the RAM independently via the multiplexed DMA. The alphanumeric display system reads a row-scan character ROM to generate the special set of characters used in the 3724A/25A/26A Baseband Analyzer. The two most-significant bits of the 8-bit character word are used to produce any of the 64 characters in inverse video, flashing, flashing inverse video or the normal mode. The CRT provides an 8×8 matrix per character into which 5×7 dot matrices are written with the spare columns used for automatic spacing.

When using measurement modes such as selective level or wideband power it is often necessary to be some distance away from the test equipment when attempting to read the output on the screen. To help overcome this problem the 3725A has an expanded output format that displays the critical parameters of the particular measurement in characters 16 times the standard size.

A novel approach was used to generate a lowercase p, which is widely used in displays, by using a spare row-bit in the character ROM to shift the raster as shown in Fig. 3 by applying a voltage to the vertical position amplifier.

The spectrum analysis display consists of 1024 points across the screen horizontally and 256 points vertically. The data is written directly from the A-to-D converter to the RAM at the chosen sweep rate. The display accesses this data at the refresh rate and uses an interpolation circuit to join the points into a continuous line. The **STORE TRACE** key causes the processor to freeze alternate RAM locations of the analog data store while still allowing the remaining locations to be updated. The resultant display shows two independent traces whose horizontal resolution is halved.

In the baseband response mode the processor generates level measurements at 33 equally spaced frequencies. These results are then passed through a smoothing algorithm and written to the display area of the RAM to appear as a continuous trace.

I/O and Interrupts

In addition to the I/O function performed by the DMA system there are also 52 eight-bit I/O ports (data latches and buffers for transfer of data to and from the processor) distributed around the instrument and its optional filter mainframes. All I/O ports are on the same 8-bit data bus and 6-bit address bus and have a common strobe line for timing. The I/O buses are driven from a single peripheral interface adapter (PIA) via buffer-drivers.

I/O ports are normally written to or read from when the operating code requires the ports to output or receive data. However, the front-panel keyboards, 3726A Filter Mainframe keys and HP-IB* inputs require immediate attention from the processor whenever a key is pressed or information is passed on the bus. The maskable interrupt (IRQ) facility of the 68B00 is used to achieve the required speed of response. The IRQ line is taken around the instrument with all interrupting circuits pulling directly upon it. Each circuit that can raise an interrupt has a data port which may be polled by the processor to determine where the interrupt came from. The various interrupt sources are allotted priorities within the instrument and polled accordingly.

The keyboard drivers are designed to provide an interface betweeen an 8×8 matrix keyboard and the 68B00 microprocessor.[2] The design offers a number of advantages over the more conventional scanned-keyboard designs at minimal hardware cost. It is static, so there is no need to generate a separate low-frequency clock that might generate noise or interference. This is a significant advantage if the keyboard is used close to sensitive analog circuitry. Besides providing key rollover, the keyboard is arranged to provide multiple interrupts on the ⇑ (up) and ⇓ (down) keys (if held depressed) to allow the operator to view the effect of continuously varying a control parameter.

Self-Test and Troubleshooting

The self-test features for testing the analog functions have already been described on page 7. A combination of signa-

*Hewlett-Packard's implementation of IEEE Standard 488 (1978).

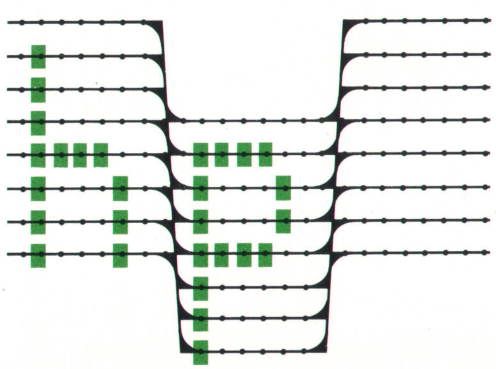


Fig. 3. *A spare row bit in the character ROM is used to shift the vertical raster to generate a lowercase p as shown above.*

ture analysis and self-test was also developed for the display and control system. The test procedure allows boards to be installed and their operation verified in a serial manner such that the features of each may be used to assist in the testing of the next board. In this way the tests that may be performed can grow in complexity as the proven capability of the system increases.

Once the processor/DMA board has been verified as operational by free-running the microprocessor and taking address signatures, this same address cycling is used to stimulate each of the ROM boards in turn, using signature analysis to verify their correct operation (the memory boards all have output-disable switches to prevent their corrupting the data bus when under test). With the ROM and processor boards operational the processor is able to run test programs that verify the RAM and I/O boards. The program that verifies the operation of these two functions can signal an overall go or no-go on light-emitting diodes (because the CRT may not be functional). When a signature analyzer is used in conjunction with the program any fault in the RAM board can be indicated down to the particular device that returns faulty data or, in the case of the I/O system, whether or not data written to a test latch can be read back via an associated test buffer correctly. This test shows whether there are any faults on the data bus lines throughout the instrument. The modules installed next in the system are the keyboards and display circuits which can be tested by means of a combination of signature analysis and internal test programs. An internal switch enables the

Checking 458,752 Bits of Program Memory

It is essential for the proper operation of the 3724A/25A/26A Baseband Analyzer that the central processing unit function correctly at all times because just one misread bit of program data (56K of code is over 450K bits) can cause the control program to be lost. In larger computers the expense of error-correcting circuits is easily justified. In some microprocessor systems, checksums over memory blocks are calculated during self-test. The 3724A uses a more secure method than checksums (a bit dropped here can be cancelled by one set elsewhere), that of cyclic redundancy checking (CRC).[1] This technique is widely used in fields where long serial bit streams are found (e.g., disc memory systems and signature analysis).

In this technique the program data, considered as a serial bit stream, is fed into a shift register (Fig. 1) with feed-forward and EXCLUSIVE-OR gates. This is mathematically equivalent to dividing the incoming data word by a number whose value is equal to that of the feed-forward taps. At any time the value in the shift register is the remainder of that division, so if the check character is included in the division, the result must be zero. Finally the divisor should be chosen so that the check word has maximum "uniqueness."

CRC-16 is designed for serial streams of data using simple gates and a shift register. In the baseband analyzer it is the ability of the processor to read memory that determines the reliability of the control program and hence the microprocessor is used to calculate the CRC word. The processor, modeling the CRC-16 algorithm (i.e., $x^{16}+x^{15}+x^2+1$), can calculate the new CRC word an input byte at a time using a recurrence relationship that is a matter of a few shifts and logical operations between the old CRC word and the incoming data. Each ROM (read-only memory) is checked separately to allow servicing to the chip level. Further, the address of each chip is included in the CRC generation to ensure correct location of the memory devices. The CRC character is generated during software development on an HP 1000 Computer before programming the ROMs, and is displayable by the baseband analyzer so that the revision of each ROM may be readily checked in service.

Reference

1. S. Vasa, "Calculating an Error Checking Character in Software," Computer Design, Vol. 15, no. 5, 1976.

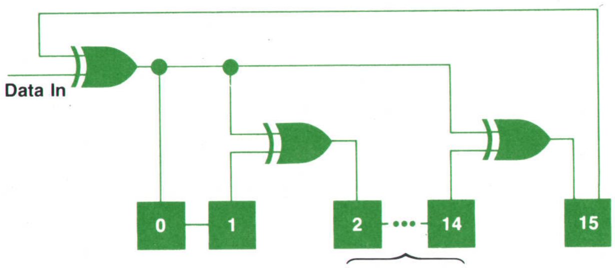


Fig. 1. *Shift register generation of CRC-16 (cyclic redundancy checking).*

Brian W. Woodroffe

Brian Woodroffe obtained his BA degree in engineering and economics from Downing College, Cambridge, England in 1970 followed by an MA degree in 1975. Before joining the R&D group at HP he worked in production and systems engineering. In the lab he has worked on the 3790A Microwave Link Analyzer and 3780A Pattern Generator/Error Detector projects. He is married and has recently spent much time extending his home. Although English by birth he has represented Scotland internationally in full-bore rifle shooting and was adjutant for the Tartan Target Team's tour of southern Africa.

Lawrence Lowe

Lawrence Lowe graduated from the Lanchester Polytechnic, Coventry, England in 1970 after serving a four-year student apprenticeship in a telecommunications company. In 1971 and 1973 he earned his MSc and PhD degrees from the University of Warwick. He joined HP in 1973 and worked briefly on the 3745A Selective Level Measuring Set before he began his contributions to the 3724A. Lawrence has been an active author and lecturer on the subject of microprocessors and digital troubleshooting. He is married, has two children, and spends his spare time as commodore of both the HP Queensferry Sailing Club and the local Port Edgar Yacht Club. He is also a steam railway enthusiast and enjoys acting as consultant to his son's model railway layout.

Microprocessor Contributions to Baseband Analyzer Accuracy and Speed of Measurement

In the 3724A/25A/26A Baseband Analyzer it is the microprocessor's function to optimize the configuration of the receiver hardware in real time appropriate to each measurement mode. The aim is to maximize measurement accuracy, minimize distortion and make the result available to the user in the shortest time possible. However, the emphasis placed on each of these characteristics differs from mode to mode. The front-end attenuators are shared by all modes and are used to control the signal level into the first mixer. In the selective-level and scan modes, where good amplitude flatness with frequency is important (see below), this signal level is lower than it is in the noise modes where good headroom above the noise floor is more important and must be closely controlled.

In the baseband response mode and the initial fast search in the high-level-user scan the attenuator and the IF (intermediate-frequency) gain settings are determined purely by the reference/threshold level entered by the operator, since the required dynamic range of the analog circuits is small.

In all modes the processor makes a logarithmic conversion of the relevant detector's output and the resolution of conversion is varied to maximize throughput. For autoranging, 1-dB resolution is sufficient, but for measurements either 0.1-dB resolution for noise or 0.01-dB resolution for the selective-level and scan modes is used.

Wideband Power Mode

Wideband power is the simplest autorange mode. It is also fundamental to the other modes because it is used by them to maintain the correct input level to the first mixer. The purpose of the wideband power algorithm is to maintain within bounds the signal level at the wideband detector. Because the detector has a long time constant to accommodate low frequencies, the processor increases the path gain in bigger steps than it decreases it (i.e., the processor responds dramatically to a falling signal level, whereas the hardware responds more quickly to a rising signal level). By this method the response to both rising and falling levels is optimized. Once the processor has decided that autoranging is required and has taken the appropriate action, it makes a logarithmic conversion of the instantaneous analog-to-digital converter output word while waiting for the wideband detector analog circuits to settle. This result, corrected by the calibration data, is displayed as the best estimate of the wideband input power. Once autoranging is complete a real-time clock is started that interrupts the processor after a time related to the time constant of the analog circuits. This next result is made available to the HP-IB interface. In the meantime the displayed result has been slewing continuously to this final result. In this manner the front-panel user always gets a best estimate, much like an analog meter, whereas the HP-IB user will get only the accurate settled value. This settling time software is shared by all modes.

Spectrum Analysis Mode

In this mode the RF (radio-frequency) attenuation is autoranged by the processor if the wideband input power is greater than the reference level. A typical application of the baseband analyzer that requires this function is the measurement of a white-noise signal. When the in-band noise is displayed at the reference level, the broadband power considerably exceeds the reference level and the RF attenuation must be increased accordingly to prevent distortion. The IF gain is increased by the processor to compensate for this greater attenuation, thus maintaining correct calibration of the display.

Selective-Level, Scan and Noise Modes

The power of the microprocessor is fully used in these modes. The timer mechanism for generating both front-panel results and HP-IB results is the same as for the wideband power mode and the method of using the wideband path for monitoring input power is the same as for the spectrum analysis mode. However, the IF gain is autoranged in three sections: 20 dB before the second-IF measurement filters, 80 dB following the filters, and 50 dB in the demodulation signal path for noise-with-tone measurements. Autorange speed is of the essence and so the processor monitors strategically placed threshold and overload detectors to permit rapid recovery from massive overload. The processor also switches the narrowband detector into a short-time-constant state during autorange to minimize the time taken to reach the new required IF-gain value. One further refinement is possible. A 10-dB RF attenuator immediately precedes the first mixer (see page 12). When this attenuator is selected, the receiver flatness is optimum and when deselected the noise floor is optimum. In the selective-level and scan modes the processor's decision to select this attenuator or not is therefore based not only on the wideband input power as for the other RF attenuators, but also on the narrowband signal level in the IF stage. At a low signal level in the IF stage the attenuator is deselected to improve signal-to-noise ratio (provided the input power does not demand that it be selected because of a large out-of-band signal) while at other powers the optimum receiver flatness is preferable.

In the noise modes the RF attenuation must be computed to within 2.5 dB because it is essential to optimize the dynamic range of the receiver continuously to this resolution rather than to the 10-dB resolution that is adequate in other modes. Information about the energy distribution of the white-noise input signal is contained in the channel capacity value entered by the operator. The processor then calculates the required attenuation from the equation:

$$\text{RF attenuation} = \text{wideband input power} - \text{autorange level}$$

The value of the autorange level is the desired input power to the first mixer and is related to the channel capacity by a table. Thus, as the channel capacity increases, the RF attenuation decreases. The processor makes the best updated estimate of the input power and then calculates the RF attenuator setting. The RF attenuators can be switched from all-in to all-out in three steps, whereas if a state-machine approach had been used, 16 steps would have been required. Hence a considerable improvement in response time and relay reliability is made. Also the wideband power algorithm is used in this mode to determine whether a noise-on or noise-off condition exists so that the processor can decide whether the current narrowband result is an NPR (noise power ratio) or a BINR (thermal noise) measurement (see box on page 6). This is particularly important in the autosequence noise mode when this result is used to synchronize the receiver to a remote generator.

internal set of test programs to be accessed (Fig. 2) once the microprocessor and keyboards are operational.

Acknowledgments

Eric Paterson helped with the development of the microprocessor hardware and firmware and wrote the set of test menu programs. Our thanks also go Andrew Batham who made many contributions to the design of the display system. The 37018A software package was developed by Reid Urquhart and Ian Johnston.

References

1. L. Lowe, and B. Woodroffe, "Multiplexed DMA with the 6800," Microprocessors and Microsystems, Vol. 3, no. 4, 1979.
2. B. Woodroffe, "Non-Clocked Encoded Keyboard Design," Microprocessors and Microsystems, Vol. 3, no. 3, 1979.

A Combined Tracking and White-Noise Generator

by John R. Pottinger and Stephen A. Biddle

TWO TYPES OF STIMULUS are necessary for testing microwave radio equipment at the baseband. For baseband amplitude response measurements a swept sine wave of accurately programmable frequency and level is required; the sine wave also is used as a fixed-frequency test tone. For white-noise testing a defined band of white noise of known characteristics is required (see box on page 6). The HP Model 3724A/25A/26A Baseband Analyzer provides both stimuli through a combined tracking and white-noise generator design (Fig. 1).

Generator System Configuration

The leveling loop control circuits, power amplifier and attenuator are mounted on a single circuit board below the synthesizer in the 3725A Display. The output from the attenuator is routed via an impedance switching board in the 3724A Baseband Analyzer to its front panel.

This configuration enables a 3724A/25A combination to produce a tracking sine-wave source that covers a frequency range from 300 Hz to 18.6 MHz and is flat within 0.2 dB peak-to-peak from 10 kHz to 14 MHz at levels between

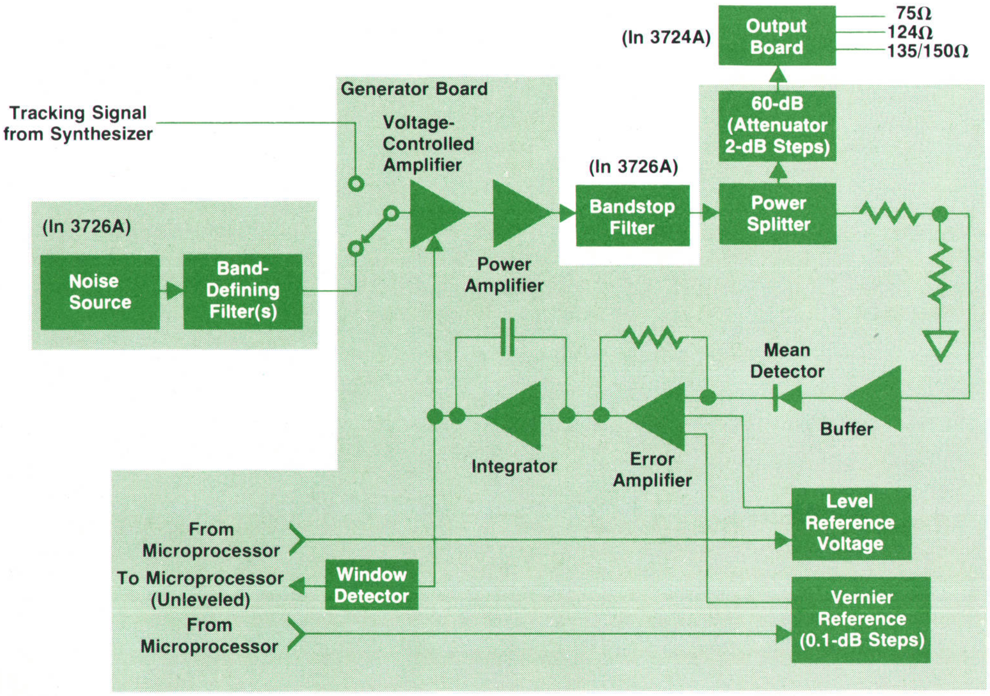


Fig. 1. *Block diagram of the white-noise and tracking generator in the 3725A Display.*

+6 dBm and −50 dBm. A 3724A/25A with a 3726A Filter Mainframe contains a noise source and all the necessary band-defining and bandstop filters for white-noise loading measurements. The white-noise generator and band-defining filters produce defined bands of noise in the range from 10 kHz to 13 MHz (the required bands being defined by CCIR,* Intelsat, etc.). Typical flatness is 0.6 dB peak-to-peak with a power range of +12 dBm to −60 dBm, and the crest factor exceeds 12 dB (crest factor = ratio of peak power to rms power). When more than nine filters are required, up to three 3726As may be used on any one system. The filters are designed to plug directly into the front of the mainframe for easy interchangeability.

The generator is designed to level either the sine-wave tracking signal or the output from the noise source. A fundamental difference between the two modes is that, for the sine-wave tracking signal, the flatness is defined only by the detector and the radio-frequency (RF) circuits between the detector and the front-panel output. In the noise mode, the signal flatness is defined by all the RF circuit components from the noise source to the front-panel output; the detector only measures the total power level of the defined band of noise.

One of the most important parameters is the inherent noise power ratio (NPR) of the white-noise test signal, which dictates the structure of the generator. The notch (slot) produced by the bandstop filters cannot be followed by any active circuits because even very small amounts of nonlinearity would reduce the notch depth. Hence, the bandstop filters must follow the power amplifier, which means the bandstop filters must be able to achieve the required notch at relatively high power levels.

Noise Source

The noise source is the Johnson noise generated by a resistor and amplified by a five-stage, low-noise amplifier. Three transistors are used per stage. A low-noise amplifier is necessary to avoid the effects of the additional noise generated by the transistors, which does not have the desired white-noise spectrum. Adjustment of the amplifiers makes the noise level flat typically within 0.2 dB peak-to-peak from 10 kHz to 14 MHz.

*International Radio Consultative Committee

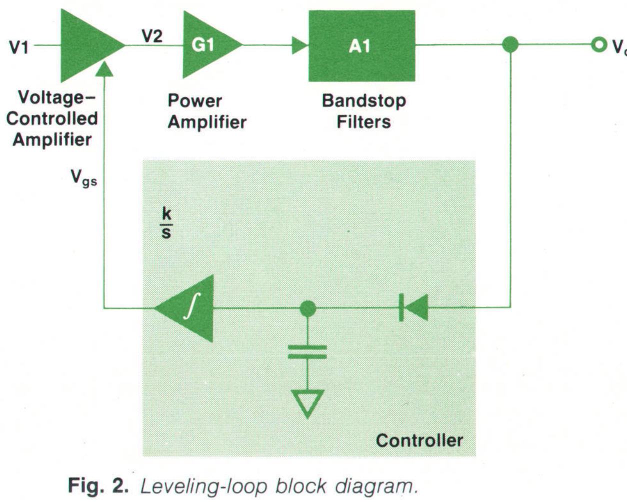


Fig. 2. *Leveling-loop block diagram.*

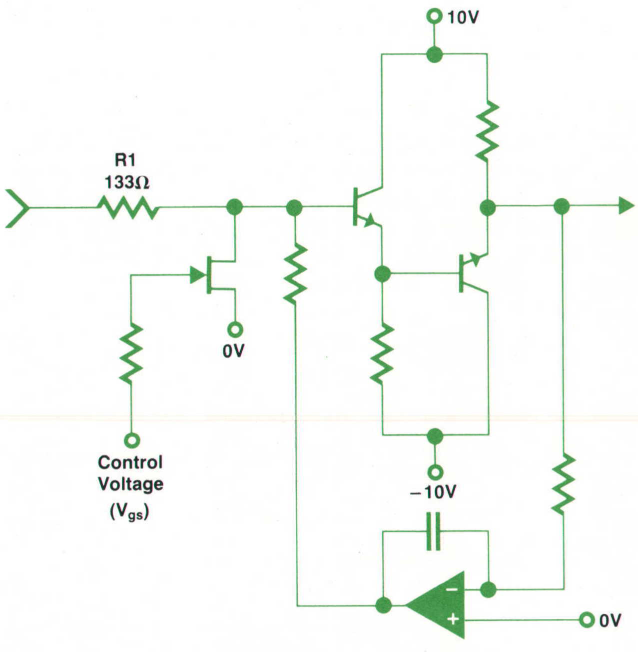


Fig. 3. *Voltage-controlled amplifier stage.*

Filters

The higher-frequency bandstop filters use quartz crystals, and all the other filters use passive LC (inductance-capacitance) elements. The band-defining filters use series combinations of high-pass and low-pass sections. Many of the inductors use stranded Litz** wire to achieve the high Q necessary. The filters are hermetically sealed to prevent the Litz wire's absorbing moisture, since this would reduce their Q. The hermetic seal is effected by printing a broad rectangular copper border on the component side of the printed circuit board and soldering a steel cover to it. A small quantity of silica gel is enclosed to absorb any moisture trapped at the time of sealing. This avoids any necessity to fill the enclosure with specially dried gas. The filter printed circuit board is mounted within its case, an edge connector at one end interfaces to the motherboard in the 3726A, and a plastic molding and keyswitch at the other end becomes the front panel. The filter is switched in or out by a pair of relays on the filter board to minimize stray coupling.

The 3726A also may be used to house bandpass filters that correspond to the bandstop filters. The bandpass filters may be used to extend the receiver NPR measurement sensitivity for applications such as radio system component testing.

Leveling-Loop Control

Fig. 2 shows a simplified diagram of the leveling loop. For both modes the RF input V1 is known within ±2 dB. In the noise mode each band-defining filter has an individual attenuator that defines a constant output power. Hence V1 is effectively eliminated as a primary factor influencing the loop gain. The voltage-controlled amplifier consists of three stages, one of which is shown in Fig. 3. A JFET (junc-

**A stranded conductor whose strands are individually insulated and wound so that each strand varies its position in the conductor's cross-section to minimize skin effect.

tion field-effect transistor) is used as the control device, the circuit attenuation being controlled by the applied gate voltage. This approach yields consistent flatness at all values of attenuation and good high-level signal handling capabilities. As the JFET's drain-source resistance is varied, the bias conditions for the emitter-follower stages are changed. This would allow a transient to propagate around the loop to the level detector, which could cause the entire loop to become unstable. The local integrator loop prevents this effect by controlling the bias conditions to hold the mean output of the emitter follower at zero volts.

Two different leveling-loop and local-loop time constants are used. A 100-ms leveling-loop settling time is used for the noise mode and the sine-wave mode above 10 kHz. A 500-ms settling time is used for the sine-wave mode from 200 Hz to 10 kHz. A fast settling time is needed to ensure a rapid update rate for the baseband response mode.

Leveling-Loop Gain

In a leveling loop the loop gain and hence stability and settling time are usually dependent on the input signal amplitude, the multiplier function, and in this case the attenuation of the bandstop filter. The input V1 has already been defined to within ±2 dB so this parameter is not significant. Referring to Fig. 2,

$$\text{Loop Gain} \quad G = \frac{\partial V2}{\partial V_{gs}} \times G1 \times A1 \times \frac{k}{s} \qquad (1)$$

For the three-stage voltage-controlled amplifier (Fig. 3),

$$B = \frac{V2}{V1} \approx \left[\frac{R_{DS}/R1}{(R_{DS}/R1) + 1 - (V_{gs}/V_p)} \right]^3 \qquad (2)$$

where B is the gain of the RF path of the voltage-controlled amplifier, R_{DS} is the JFET drain-to-source resistance at $V_{gs}=0$, V_p is the JFET pinchoff voltage and V_{gs} is the gate-to-source voltage. Differentiating (2) with respect to V_{gs} gives:

$$\frac{\partial V2}{\partial V_{gs}} = \frac{3V2\sqrt[3]{B}}{V_p(R_{DS}/R1)}$$

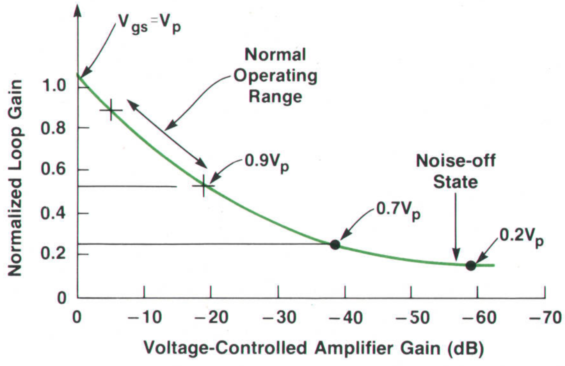


Fig. 4. *Loop-gain variation with voltage-controlled amplifier gain.*

From (1) and (3) and using $V_o=V2\times G1\times A1$ it follows:

$$G=KV_o\sqrt[3]{B}$$

where K is a constant.

The important characteristic of this equation is that a large gain change in the voltage-controlled amplifier results in a relatively small change in the loop gain. The result is shown in Fig. 4. Despite the high degree of nonlinearity in the JFET characteristic, an acceptable variation of settling time can be achieved for the main control loop.

Power Amplifier and Power Splitter

The output stage of the power amplifier is shown in Fig. 5. The peak noise power of 650 mW requires approximately ±7V into 75 ohms. A class-A push-pull current-source output stage is used since it can use the standard ±15V supply on the circuit board. The output impedance is defined by a 75-ohm shunt resistor.

The power-splitter arrangement is switched to give the best compromise for both noise and sine-wave modes. The sine-wave configuration is a conventional power splitter, which offers the best accuracy and least sensitivity to load impedance variations. The noise configuration is optimized for minimum loss at the expense of higher sensitivity to load impedance variations. This maximizes the power output available at the front panel. For all low-power measurements there is some attenuation between the power splitter and the load. This reduces errors caused by load variations and is also the dominating influence in defining the output return loss. High-power outputs are primarily required for overload testing where some loss of accuracy is acceptable.

Loop Detector

The baseband amplitude response mode requires a loop settling time of less than 100 ms to within 0.01 dB. This rules out the possibility of using a thermal true-rms detector, which would have been ideal for detecting both noise and sine-wave signals. Instead, a biased Schottky-diode bridge is used to recover the full-wave-rectified mean of the input and provide a good balance between accuracy and speed. The difference in detected voltage between the noise and sine-wave signals is compensated by the microprocessor-controlled dc reference voltage which also provides the fine output adjustment over a 2-dB range.

Noise-off State

The automatic sequence noise mode of the baseband analyzer performs a sequence of NPR and thermal noise (BINR) measurements on up to four slot frequencies and tabulates the results on the CRT display. In this mode the generator continuously cycles its noise output on (at the level entered by the operator) and off (to allow the BINR measurements to be made). For end-to-end measurements the remote receiver locks onto this generator sequence.

Instead of switching the attenuator to cycle the generator on and off, the leveling-loop output is reduced to a low level (Fig. 4) by setting the JFETs in the voltage-controlled amplifier to their minimum resistance. This achieves a level change greater than 40 dB, which is more than adequate and avoids premature wear of the attenuator relays.

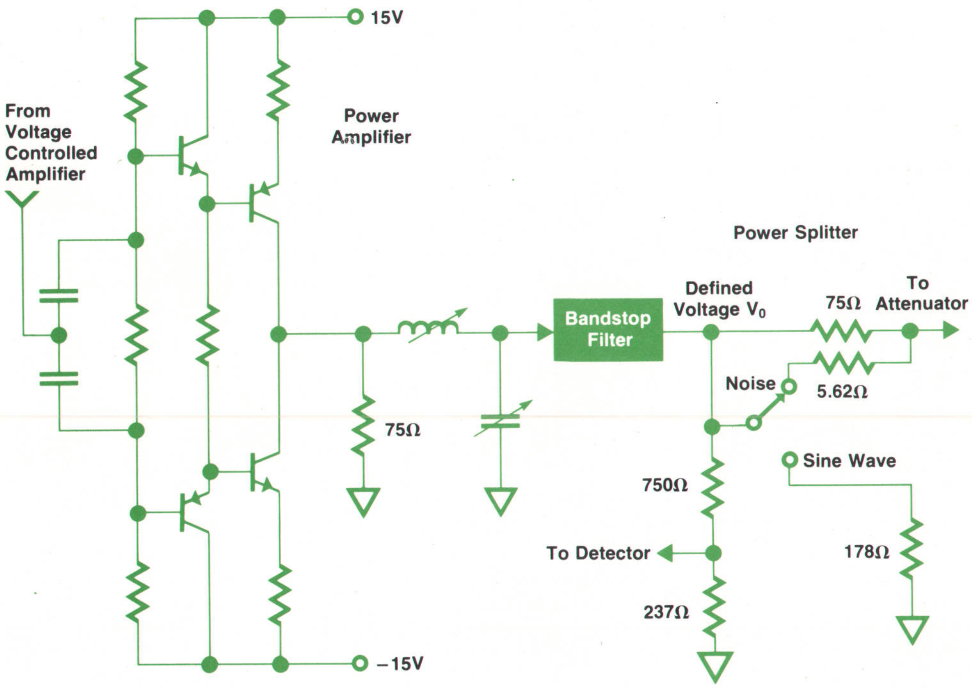


Fig. 5. *Power amplifier and power splitter circuit.*

System Flexibility

An important goal during the development of the generator was to give the customer maximum flexibility in configuring the system. The design allows interchangeability of filter type and position, and even variation in the number of filter mainframes. These variations in system topology have required that each circuit element be designed to be flat with frequency and exhibit high return loss. In this way variations caused by interactions between elements are minimized.

Considerable attention to circuit layout and configuration was necessary to eliminate ground loops which, because of finite ground path impedance, would otherwise result in degradation of bandstop filter notch depth.

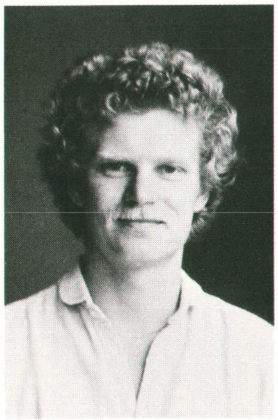
Up to 100 dB of gain is necessary to amplify the Johnson noise of the noise source to the level necessary to produce +12 dBm at the generator output. Careful positioning of RF screening was essential to prevent crosstalk between the bandstop filter path on the 3726A motherboard (carrying high-level signals) and the early stages of the noise source contained within the same mainframe.

Acknowledgments

Much credit must go to John Cooke who developed the programs that were used to design all the filters, and who worked on the early prototype filter designs. Daya Rasaratnam helped with the design of the bandstop and band-defining filters.

John R. Pottinger

John Pottinger joined HP in 1978 after gaining several years experience with various U.K. electronics companies. He designed the generator circuits for the 3724A/25A/26A Baseband Analyzer. John received the BS degree in 1970 from North Staffordshire Polytechnic University and is currently working on his MS dissertation. He is a member of the IEE. Originally from the south of England, John now lives in Dumfermline, Scotland. He is married, and enjoys mountain climbing, skiing, playing snooker and working on his radio-controlled model yacht.



Stephen A. Biddle

Stephen Biddle joined HP in 1976 after obtaining the BS degree in electronic engineering from the University College of North Wales, England. He first worked on a selective-power meter investigation before joining the baseband analyzer project team in 1977. Steve developed the 3726A Filter Mainframe and a large number of the band-defining and bandstop filter plug-ins associated with it. He also designed the first-IF and selective second-IF filters for the receiver section. Steve recently left HP. He is married and enjoys horse riding, photographic modeling, and having dinner parties.

Wideband, Fast-Writing Oscilloscope Solves Difficult Measurement Problems

A new expansion storage cathode ray tube and a wideband amplifier design extend the writing rate frontier to 2000 cm/μs.

by Danny J. Oldfield and James F. Haley

COMBINING EXTREMELY FAST WRITING RATE (2000 cm/μs) with high-bandwidth signal fidelity (275 MHz), the HP Model 1727A Oscilloscope (Fig. 1) captures single-shot or low-repetition-rate signals with rise times as fast as 1.27 ns, four divisions high. The key to this high performance is Hewlett-Packard's state-of-the-art expansion storage technology. The 2000 cm/μs writing speed is available in the variable persistence mode, which provides the maximum light integrating capability needed to produce bright, crisp displays of low-repetition-rate signals.

Consider the need to measure a very fast transition that occurs only once every few seconds. With a conventional oscilloscope, this measurement is virtually impossible. With the 1727A, however, the transition need occur only once for it to be captured for complete characterization (Fig. 2).

Another difficult problem, one that arises in logic design, is identifying a setup or hold violation on a D flip-flop: when the clock arrives before the data has been present long enough, or the data is removed too soon after the clock has occurred, the output is indeterminate. If the violation occurs only a small percentage of the time, a conventional oscilloscope probably will not produce a display bright enough to be noticed. The 1727A, on the other hand, if set for long persistence, will need just one or two occurrences to display the violation (Fig. 3).

The high bandwidth needed for such applications is obtained using an amplifier system originally developed for the 1720A Oscilloscope.[1] The major 1727A design challenge was to develop a cathode ray tube (CRT) that could accommodate this amplifier system and take advantage of Hewlett-Packard's expansion storage technology. During the development of the original expansion storage CRT, currently used in the 1744A Oscilloscope, extensive computer modeling was used to optimize the expansion lens design. Leveraging that effort by adding high-speed vertical plates to the 1744A CRT design resulted in a CRT with

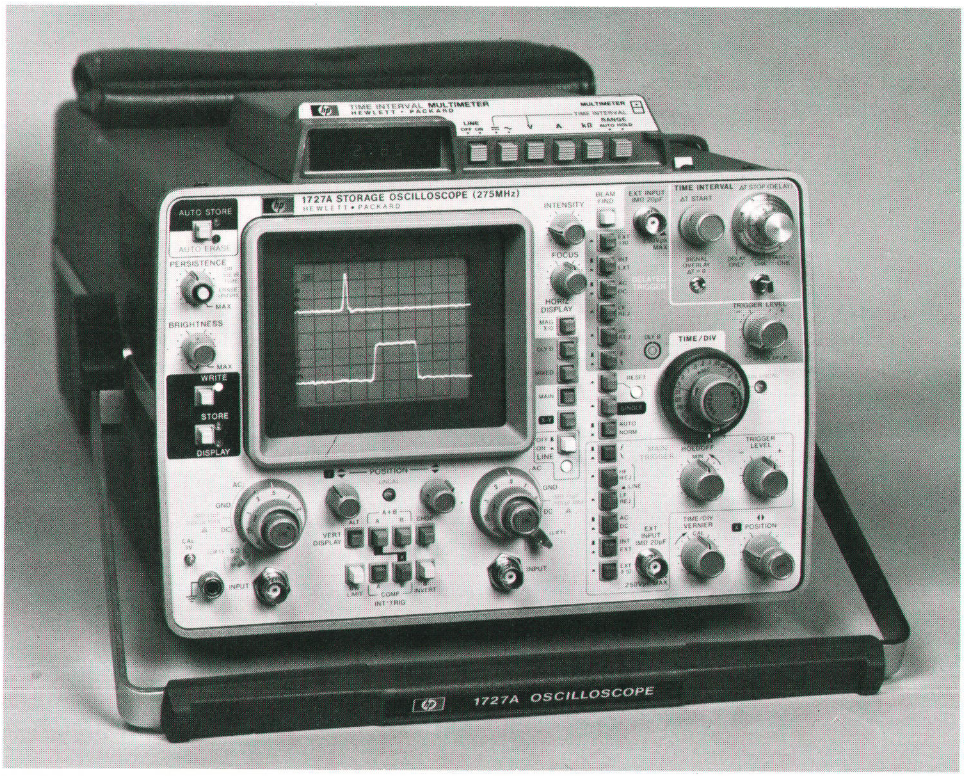


Fig. 1. *With its 275-MHz bandwidth, variable persistence, and high 2000-cm/μs writing capability, Model 1727A Oscilloscope can capture single-shot or low-repetition-rate signals that would be difficult to observe otherwise.*

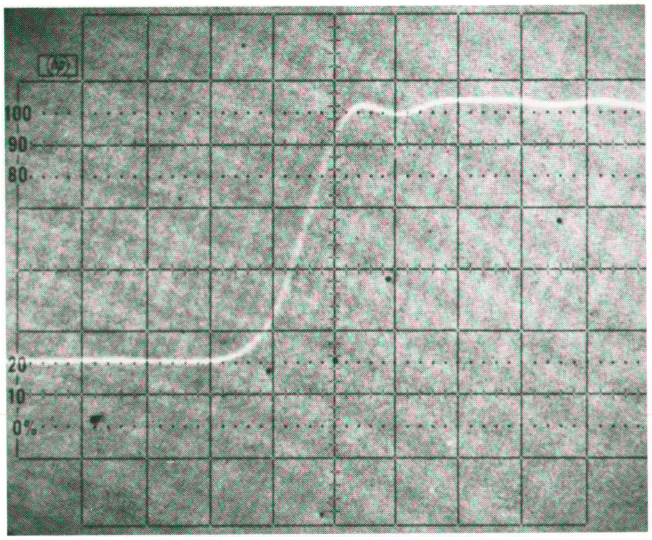


Fig. 2. *A single-shot signal with a 1.27-ns displayed transition time is easily captured by the 1727A Oscilloscope. (Sweep speed = 1 ns/div.)*

higher performance but without the long development normally associated with a completely new CRT. Complementing the 1727A CRT's high performance is an auto-intensity circuit designed to minimize blooming and virtually eliminate operator concern about burning the storage surface.

Expansion Storage

Expansion storage combines a precision storage mesh with an electronic lens system that magnifies and projects the stored image. The storage mesh used in an expansion storage CRT is much smaller than the storage mesh used in earlier variable persistence/storage tubes, just as the 35-mm transparencies used for home entertainment purposes are much smaller than the screen upon which their images are projected. The electrostatic expansion lens provides a linear magnification of approximately six. This results in a displayed image approximately 5.5 cm × 7.5 cm. Since the lens action is fully static, no additional switching is required and all of the advantages of flexibility noted for variable persistence CRTs are retained.

The expansion storage mesh is sandwiched between the collector and accelerator meshes, directly in front of the collimation lens (see Fig. 4). Although it is physically dense (1000 lines/inch) and has a target area of only 1 cm^2, the mesh operates much like a standard storage mesh.

Several design advantages were gained by reducing the storage target size. With a smaller target, better dielectric deposition uniformity can be achieved than with larger meshes. The write gun can be shorter and more efficient than one needed to scan a larger area. The writing rate can be improved further by raising the write gun cathode potential, although this results in a tradeoff with deflection sensitivity.

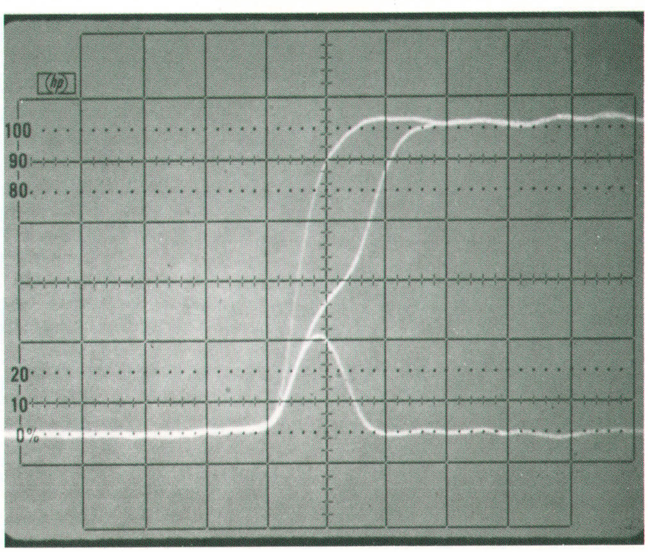


Fig. 3. *D flip-flop setup violation showing a metastable state. (Sweep speed = 5 ns/div.)*

Distributed Deflection Plates

The 1727A uses a transmission-line deflection plate structure that appears to the vertical output amplifier as a resistive load rather than a capacitive load (Fig. 5). With this structure, response is limited not by available drive current but by the speed of the overall amplifier design. This results in a much higher bandwidth than is possible with conventional deflection plates.

The structure is manufactured by winding helixes of metallic ribbon. Each helix is analogous to a lumped-parameter transmission line. After solidly mounting a pair of helixes on glass beading rods in the electron gun assembly, no further mechanical adjustment for electrical characteristics is needed. By matching the signal delay between adjacent segments to the velocity of the write gun electrons, a given electron is deflected by a certain part of the input signal for the entire time it is between the plates. Thus, the electron beam is exposed to a deflection field that travels from segment to segment at the speed of the electrons.

Optimizing the Expansion Lens

The usual problems encountered in the design of good lens systems that project visual images are inherent in the design of an electrostatic lens. Limited material choices, constantly varying "refractive index," and modeling difficulties characterize the major obstacles. Furthermore, holding linear distortions to less than 4% and focused spot size variations to a ratio of 2:1 or less at 6× magnifications presented challenging problems.

The lens used to establish feasibility of the projection storage method was a simple cylindrical design from Liebmann.[2] The results of the initial experiment were mixed. The feasibility of obtaining a real inverted image of the storage mesh on the phosphor screen was demonstrated, but the image quality was terribly poor. Several more lenses were tried with similar results, and a computer-aided design model was developed to assist in the task. An extremely useful program was written as a result of these efforts.

The CAD program is capable of accurately modeling rotationally symmetric lens systems of the general type. It allows plotting equipotential contours within the boundaries

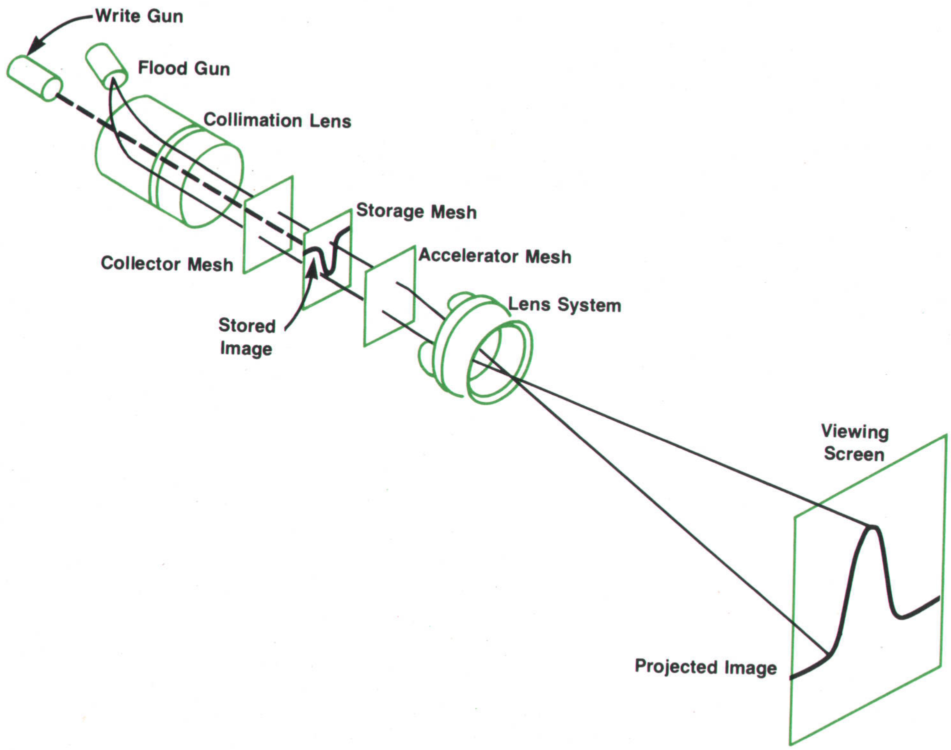


Fig. 4. *Model 1727A's expansion storage CRT combines a small precision storage mesh with an electrostatic lens system.*

of the defined system and includes field computation subroutines for use in computing ray-trace trajectories through the lens. Fig. 6 shows a sample set of equipotentials and ray traces obtained by using the program to model one of the lens configurations studied. When the predicted results were compared with experimental results on actual tubes, it was found that agreement within 5% could be expected. This accuracy proved to be more than sufficient for the program's desired uses.

Although computer modeling was extremely useful in approaching the final lens design, the program's limitation to rotationally symmetric cases prevented its use in the final

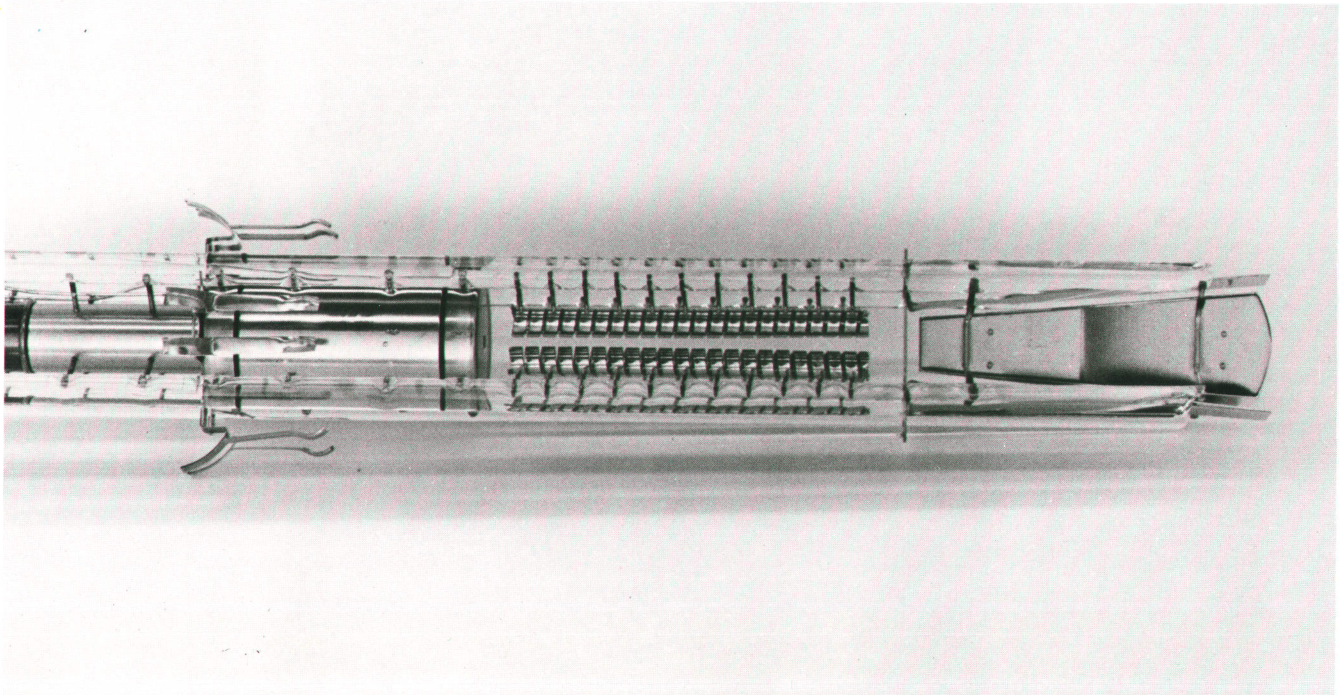


Fig. 5. *The 1727A CRT uses a transmission line vertical deflection plate structure to help achieve high bandwidth.*

Variable Persistence

The analysis of transient electrical signals, whether repetitive or single-shot, has long been a problem for users of cathode ray tube oscilloscopes. Conventional CRTs may be able to produce a screen trace during the occurrence of the transient, but the few milliseconds of phosphor persistence usually does not allow enough time for anything except a fleeting glimpse of the waveform.

Phosphors generate photon energy from electron excitation. This photon energy consists of a primary emission, which occurs at the time of excitation, and a secondary emission that continues after the stimulus has been removed. The duration of the secondary emission, called persistence, is frequently used to characterize phosphors. P31, for example, decays to 10% of the peak light output level in 38 μs.

Optimum trace quality could be maintained under varying conditions by switching phosphors, thereby compensating for sweep speeds, repetition rates, and other signal characteristics, but that is not practical. Variable persistence oscilloscopes, on the other hand, afford control over the perceived persistence of a CRT phosphor and therefore can be tailored to the characteristics of the signal being measured.

Hewlett-Packard introduced variable persistence/storage CRTs in 1966. These tubes used a dielectrically coated fine wire mesh located near the phosphor surface (Fig. 1). When the write beam impinges on the storage surface, it writes by creating areas of localized positive charge that can be differentiated from the unwritten background areas of the target. The localized charging is caused by secondary emission, a process that can be illustrated best by the following example. Suppose that three primary write beam electrons strike the storage surface with just enough energy to dislodge three secondary electrons. Since the charge is zero, the net effect of this interaction is zero. However, if the primary electrons strike the surface with substantially more energy, each one may dislodge several secondary electrons. If the secondary electrons can be captured, or collected, by a nearby electrode so that they do not return to the storage surface, the required positive charging of written areas will be accomplished.

The secondary emission ratio, defined as the number of secondary electrons produced by each primary electron, is a function of many variables. Write beam energy, dielectric material characteristics, and secondary collection efficiency all determine the overall efficiency of the process. A well-designed storage CRT optimizes all these factors to produce controlled charge patterns on the storage surface, so that written and unwritten areas can be easily differentiated.

After completion of the sweep, flood gun electrons produce a visual image of the stored trace. The low-velocity electrons from the flood gun are repelled by negatively charged areas of the storage surface and are gathered by the positive potential on the collector mesh nearby. Where the surface has a net positive charge—the written areas—the flood gun electrons approaching the mesh are pulled through by the high accelerating potential on the phosphor surface. Once through the storage mesh, the electrons strike the screen with sufficient energy to excite the phosphor, produce photon emission, and display the stored trace (Fig. 2).

To erase the stored image, the storage mesh is raised to a potential equal to that of the collector mesh (120V) for approximately 50 ms. This accelerates the flood gun electrons with enough energy to create a secondary emission ratio greater than one and thereby produces a net positive charge over the entire storage surface. The surface then is returned to approximately 10V for the next write cycle.

Variable persistence is obtained by shortening the erase cycle so only partial erasure occurs. Imagine a stored image being projected onto the phosphor. While the image remains stored, it continues to be displayed on the screen of the CRT. If that stored image is erased slowly, the image on the phosphor fades away. This process results in a persistence dependent on the erasure of the storage mesh. Applying small erase pulses (4 to 8 volts) slowly erases a stored image. By varying the frequency of the pulses, different persistence levels can be achieved. At the minimum-persistence control setting, the persistence is approximately 300 ms. When no pulses are applied, the persistence is maximum.

The front-panel brightness control determines the background brightness, which is a function of the number of flood gun electrons striking the phosphor. Increasing the brightness control increases the potential on the entire storage mesh. This causes the written and unwritten areas to pass more flood gun electrons, thereby increasing both background and trace brightness. The relative contrast between the trace and the background remains constant until the trace saturates. At this point, a reduction in contrast results. Contrast adjustment is achieved through the interactive adjustment of both the intensity and the persistence controls. Normally the brightness control is positioned at minimum and increased as required by the application.

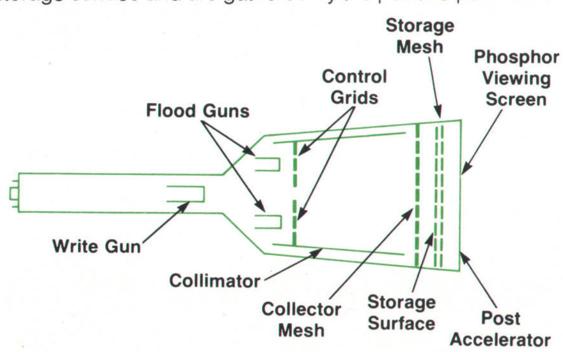


Fig. 1. *Simplified variable persistence CRT construction.*

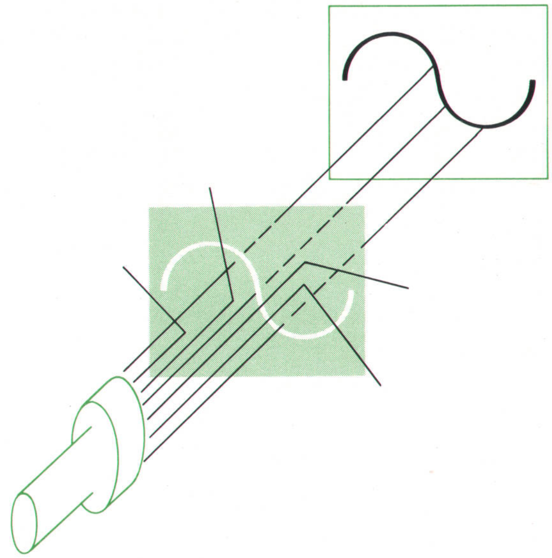


Fig. 2. *Once a trace is written on the storage mesh of a variable persistence CRT, flood gun electrons can penetrate the storage mesh in the written areas and continue on to the phosphor. Flood gun electrons are repelled by unwritten areas of the storage mesh.*

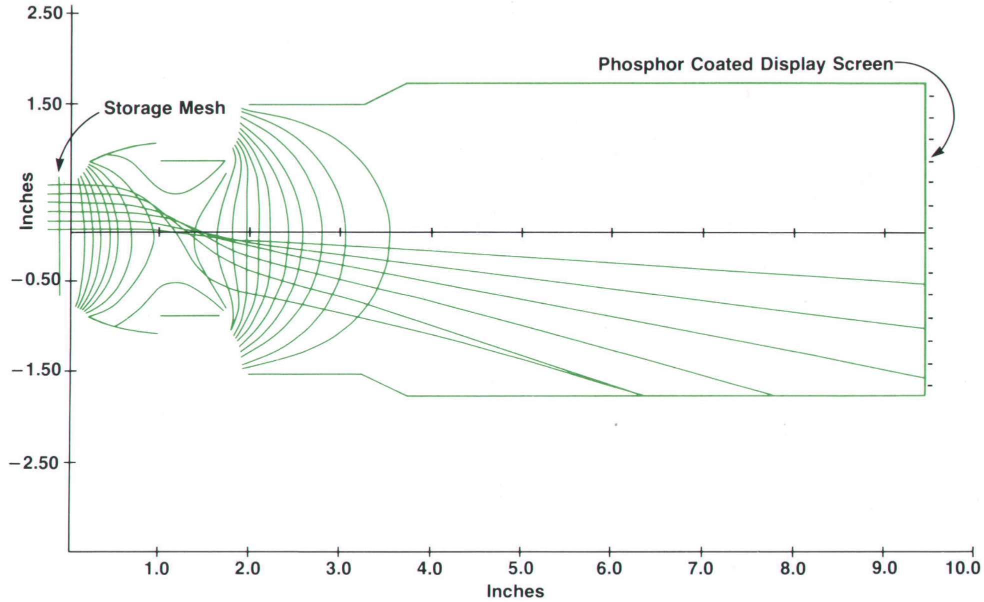


Fig. 6. *A sample set of equipotential lines and ray traces obtained from a computer CRT modeling program that was developed to assist in the design of the 1744A and 1727A CRTs.*

design phases. Because of the rectangular display format, it was found that optimum lens performance could be most easily achieved by strongly correcting the lens in the four corner regions.

The lens is operated at a lower potential than the accelerator. This decelerates the electrons leaving the accelerator mesh and deflects them toward the axis of the lens system. The electrostatic field is strong enough to cause the electrons to cross completely through the lens axis. This causes divergent beams of electrons from the accelerator mesh to converge and produce an inverted real image of the storage mesh at the phosphor screen. On the viewing side of the lens system, the electrons are accelerated again to regain their original velocity, and the image is magnified when projected on the phosphor. A displayed spot on the phosphor moves six times farther than on the storage surface, thereby increasing the writing speed by a factor of six.

A new write gun, optimizing both current density and spot size, was developed to enhance the trace characteristics of the expansion storage display. High current density is necessary to achieve fast writing speeds, while small spot size compensates for the effects of magnification.

Auto-Intensity Circuit

The 1727A's auto-intensity circuit is a significant contribution to HP's variable persistence/storage oscilloscope technology (Fig. 7). The circuit, which includes the CRT in a feedback loop, senses a small percentage of the beam current, and when the intensity exceeds a preset level, clamps the unblanking gate voltage applied to the write gun. This minimizes blooming and significantly reduces the chances that the storage surface will be burned.

Within the CRT is an accelerator cup that shapes the electron beam and intercepts any off-axis electrons emitted from the cathode. When the stray electrons strike the cup wall, a small current is generated that is directly proportional to the true beam current. This small current, typically nanoamperes, is applied to the integrator in the auto-intensity circuit. When the integrator input current exceeds a preset level, the output starts ramping to the threshold level, which is set on an individual basis during calibration to provide maximum viewing capability over a wide range of sweep speeds. The integrator charges during a sweep and slightly discharges during the retrace when the beam is blanked. When the integrator reaches its threshold value, it clamps the gate voltage, which then limits the CRT beam current.

Integrator output varies depending on operating conditions. For example, all repetitive waveforms, regardless of sweep speed, reach threshold after five or six sweeps and force the gate voltage to clamp. For single-shot events at fast sweep speeds, the integrator charge time is insufficient for it to reach the threshold level, while single-shot measurements at mid-range sweep speeds usually do allow the integrator to charge to the threshold and clamp the gate voltage. With slow sweep speeds, the trace intensity starts high and then decreases to an acceptable level as the integrator reaches the threshold.

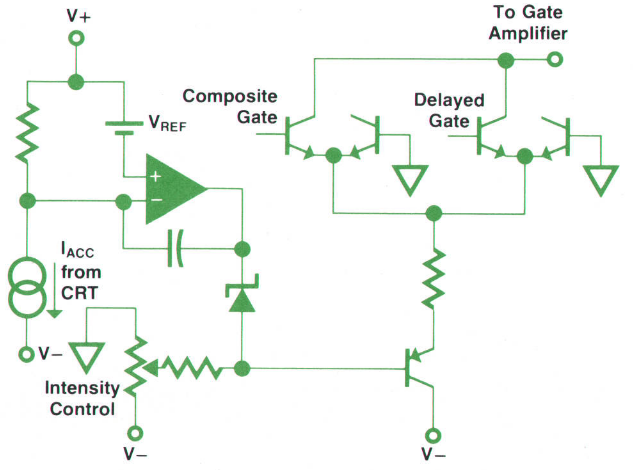


Fig. 7. *Auto-intensity circuit minimizes blooming and possible storage surface damage.*

Fast Vertical Amplifier System

The 1727A's high bandwidth results from using custom 2-GHz integrated circuits developed originally for the 1720A, a high-performance conventional oscilloscope.

The vertical amplifier system consists of two vertical channels having separate attenuators with selectable input impedance (1 MΩ/50Ω). The attenuators give full-bandwidth performance and have very low input capacitance (11 pF). A custom monolithic preamplifier in a differential cascode configuration conditions each channel's attenuated signal. The circuit contains 35 2-GHz transistors that control channel and trigger selection as well as vertical positioning. The two chips for the two vertical channels are mounted on one alumina substrate that includes 16 laser-trimmed thick-film resistors. The combined output of the preamplifiers goes through a 50-ns delay line to the output amplifier. The output amplifier has two ICs. One is in a 16-lead dual in-line package and the other is on an alumina substrate with thick-film resistors. The thick-film amplifier is a "sliding cascode" configuration with high current-drive capability and large dynamic range for driving the CRT's deflection plates.

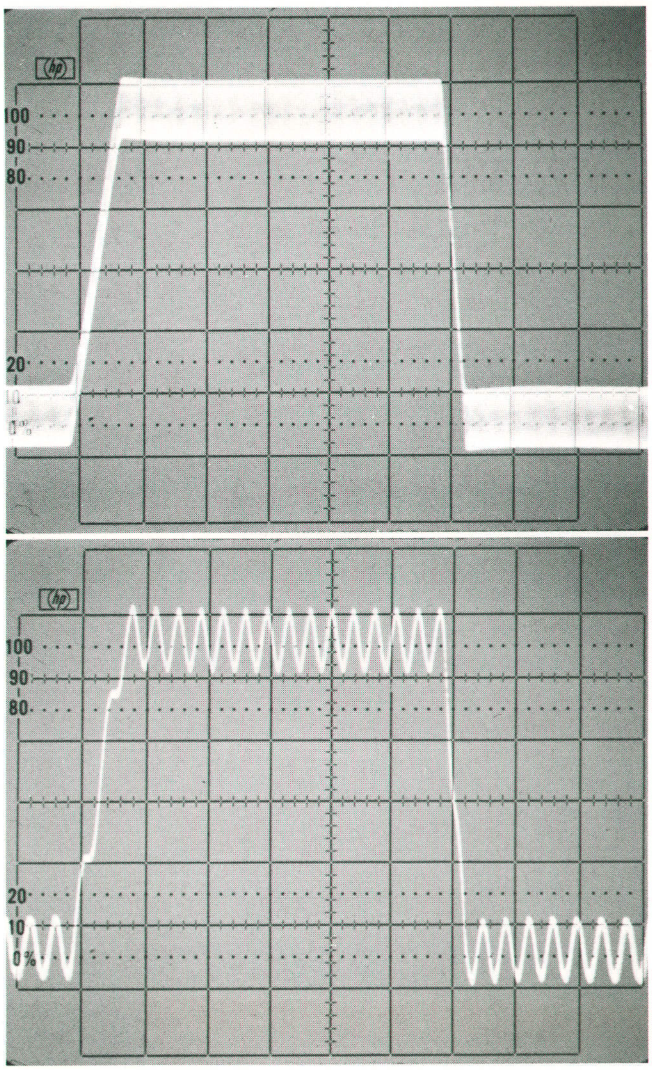


Fig. 8. *Auto-erase mode is a repetitive single-shot mode that provides a continuous sequence of updated displays. Here it helps in the observation of noise on a pulse train. Top: conventional display. Bottom: auto-erase display.*

Storage Modes

Like other HP 1700-Series variable persistence/storage oscilloscopes, the 1727A has two modes of waveform storage: auto-store and store. The auto-store mode, useful for single-shot events, is selected by pressing both the **AUTO STORE** and **SINGLE**-sweep pushbuttons. Auto-store automatically sets the persistence control to maximum for the fastest writing speed. After the event triggers the sweep and the signal is captured, the 1727A automatically switches to the store mode for maximum storage time. Front-panel indicators show which state the instrument is in and clearly indicate when the scope is triggered and when it switches to store. Pressing the **STORE/DISPLAY** button produces a viewable trace.

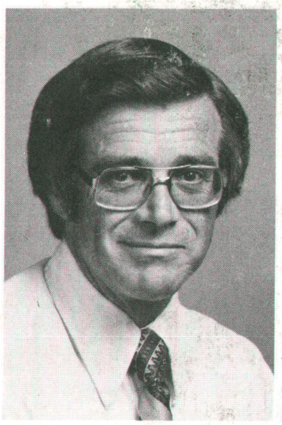
The 1727A also provides an auto-erase mode, which is a repetitive single-shot mode. Once set up, auto-erase provides a continuous sequence of updated displays. The view time is selected by the user to allow hands-off operation. Signals can be measured and circuits adjusted without the need to reset the oscilloscope. If a dual-channel alternate-sweep display mode is selected, the auto-erase circuits wait until both sweeps have completed before starting the view time. An example of the use of the auto-erase mode is shown in Fig. 8.

Acknowledgments

We wish to acknowledge the many people who contributed to the development of the 1727A. Paul Carnahan and Frank Balint did the initial design of the CRT, and Vic Petrosky finalized that design. Felipe Borrego made many contributions from CRT production engineering. Jim Felps contributed to the power supply and auto-intensity circuit. George Blinn did the mechanical design, and Lee Mack provided manufacturing support. Jim Egbert, Roy Wheeler, and Ron Westlund contributed significantly to the implementation of the 1727A, and Stan Lang gave unending support and guidance throughout.

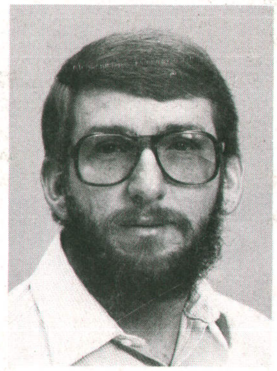
References

1. P.K. Hardage, S.R. Kushnir, and T.J. Zamborelli, "Optimizing the Design of a High-Performance Oscilloscope," Hewlett-Packard Journal, September 1974.
2. G. Liebmann, "Measured Properties of Strong 'Unipotential' Electrostatic Lenses," Proceedings of the Physical Society, Vol. 62, Part 4, Section B, April 1949.



James F. Haley

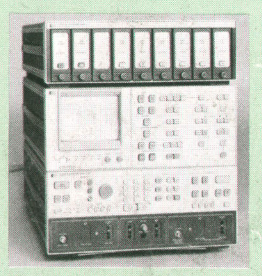
Jim Haley received his BSEE degree from the University of Colorado in 1964. He began his career as a control systems and circuit designer for a large semiconductor company, then joined Hewlett-Packard in October 1969 as a designer of electronic tools for the CRT R&D lab. He later supervised the lab's tube design group and assisted in the development of CAD tools for CRT design. He's now manager of manufacturing engineering for the oscilloscope product line. Jim is married and has two teenage children, and he enjoys the great fishing and backpacking available in southern Colorado.



Danny J. Oldfield

Dan Oldfield joined Hewlett-Packard's Colorado Springs division in 1978 after receiving his BSEE degree from the University of Tennessee. After a year in oscilloscope customer service working with 1700-Series products, Dan moved into the oscilloscope R&D lab to finalize the design of the 1727A and transfer it to the manufacturing phase. Born in Huntsville, Alabama, Dan is married and has a newborn daughter. He works with the local Big Brother organization, plays piano, and is currently renovating his home.

PRODUCT INFORMATION



HP Model 3724A/25A/26A Baseband Analyzer

MANUFACTURING DIVISION:
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South Queensferry, West Lothian
Scotland EH30 9TG

TECHNICAL DATA: HP Publications 5952-3282, 5952-3285, 5953-6686.

PRICES IN U.S.A.:
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3725A Display, $14,715.
3726A Filter Mainframe, $2620.

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