Noise Figure Meter Sets Records for Accuracy, Repeatability, and Convenience

Noise figure measurements used to be mysterious, time consuming, difficult, and not very accurate. This instrument makes them quick, accurate, and easy.

by Howard L. Swain and Rick M. Cox

IT HAS BEEN TWENTY-FOUR YEARS since Hewlett-Packard introduced its first noise figure meter,[1,2] and noise figure is still a widely used figure of merit for the noise performance of devices, subassemblies, and complete systems.* As performance requirements have increased, it has become increasingly important to be able to measure noise figure accurately and conveniently. For example, a 0.34-dB difference in the noise figure of low-noise amplifiers for satellite earth stations can translate into a fivefold difference in price.[3]

The new HP Model 8970A Noise Figure Meter (Fig. 1) makes outstanding contributions to this field in accuracy, convenience, and flexibility. The use of a microprocessor enables it to correct several of the errors that have been accepted as part of the measurement uncertainty in the past. Its own broad tuning range, high sensitivity, and ability to control external local oscillators over the HP-IB (IEEE 488) provide a tremendous increase in convenience. The 8970A can also measure gain and display both swept noise figure and gain as functions of frequency on an oscilloscope. Thus it can gather data in a few minutes that would take hours by previous methods. It turns a mysterious, time consuming, difficult, often inaccurate measurement into one that is quick, accurate, easy, and pleasant.

*Some basic noise figure theory is presented in the Appendix at the end of this article.

8970A Contributions

Three effects that have caused errors in noise figure measurements in the past are eliminated by the 8970A. The first is the variation with frequency of the excess noise ratio (ENR) of the noise source used for the measurements. The ENR of the source must be known before the noise figure of the device under test can be measured. While some noise figure meters allow the entry of an arbitrary ENR, only one value can be entered. When a measurement is taken at a new frequency, a new value has to be entered. Thus measurements must be made at one frequency at a time.

The 8970A allows the entry of ENR for 27 frequencies at one time and stores this data in a table. When a measurement is made, the proper value is automatically used for that frequency. If necessary, the 8970A interpolates linearly between the entered data points.

The second error arises because the off or cold noise temperature of the noise source, T_C, is not equal to the standard reference temperature of 290K. T_C is equal to the physical temperature of the noise source termination, which is approximately room temperature or about 296K. This 6K error in T_C causes an error in noise figure of about 0.1 dB for noise figures below 1 dB. Assuming that T_C = 290K, as all previous noise figure meters do, simplifies the equation* enough that it can be solved by analog circuits. How-

*See Appendix.

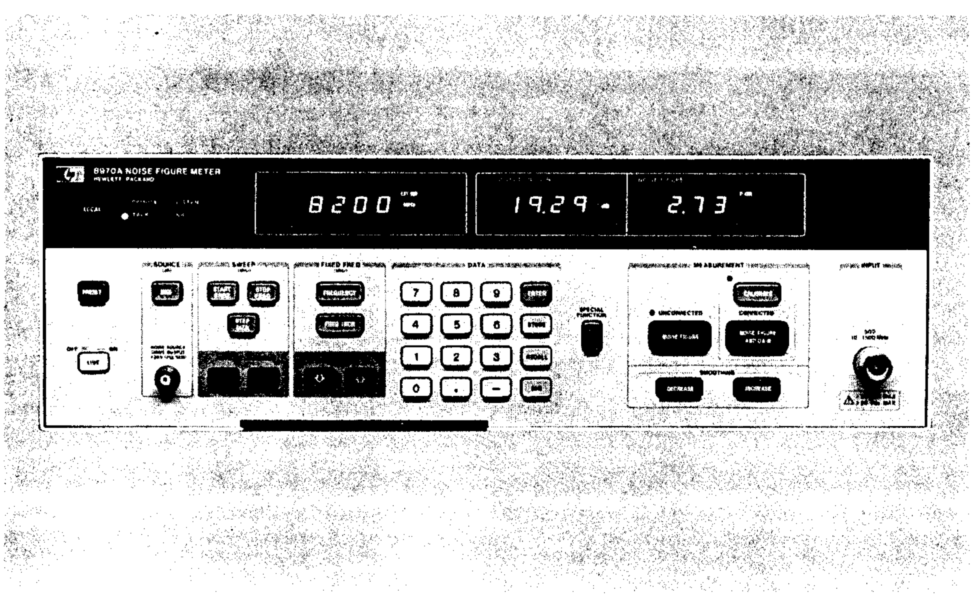

Fig. 1. *Model 8970A Noise Figure Meter measures noise figure, effective input noise temperature, and gain, contributing less than ±0.1 dB of uncertainty to noise figure measurements. A built-in microprocessor corrects automatically for ENR variations and second-stage effects.*

ever, the 8970A's microprocessor easily solves the exact equation, virtually eliminating this error.

The third error is caused by the second-stage noise figure. When the noise figure of two devices in cascade is measured, the result F_{12} includes contributions from both devices: $F_{12} = F_1 + (F_2 - 1)/G_{av1}$, where F_1 is the noise figure (in ratio form) of the device under test, F_2 is the noise figure of the second stage, and G_{av1} is the available power gain of the first stage. In a calibration step, the 8970A measures and stores its own noise figure F_2. Then when it measures the gain of the device and the total noise figure F_{12}, it can solve the above equation for the noise figure of the device alone, F_1. Since the gain and noise figure are measured at the same time, the 8970A can display the corrected noise figure in real time. This is extremely important, for example, when tuning a transistor to find the minimum noise figure. Because the tuning also affects the gain, it is essential to make this correction in real time to find the true minimum of F_1. If F_{12} is minimized and then that minimum is corrected for second-stage noise, the result is a different, higher, and wrong number because the tuning is a compromise between what minimizes F_1 and what maximizes G_{av1}. Before the 8970A it was almost impossible to make this measurement to find the true minimum of F_1.

Two additional factors can reduce the accuracy of the measured noise figure. The first is a lack of adequate resolution. The 8970A has 0.01-dB resolution to go with its 0.1-dB accuracy. To avoid any possibility of roundoff or other errors, the data sent out on the HP-IB has 0.001-dB resolution.

The second factor is lack of adequate smoothing or filtering to make use of the high resolution. The 8970A uses adjustable exponential averaging during fixed-frequency measurements.[4] Two front-panel keys, **INCREASE** and **DECREASE**, allow the user to adjust the smoothing for an optimum tradeoff between speed of response and jitter. Even when increased averaging is selected, the 8970A still updates the display three to five times per second, making adjustments easy. During sweeps, the 8970A averages the selected number of readings at each frequency before moving on.

Besides the errors described here, many inconveniences have plagued those who attempt to measure noise figure. First, all previous noise figure meters measure at only one or a few fixed frequencies. Measurements over a band of frequencies require an additional mixer and a tunable low-noise local oscillator. Some noise figure meters require the use of an additional amplifier. It is almost impossible to make swept, single-frequency measurements over a broad bandwidth. The 8970A remedies all of these shortcomings. It tunes from 10 to 1500 MHz in 1-MHz (or greater) steps, and it has enough internal gain so that no additional amplifiers are needed. Thus, the 8970A and a noise source can measure a great many devices without any additional equipment.

To measure noise figure at frequencies above 1500 MHz does require an external mixer and local oscillator. However, even this measurement is easy. All parameters (ENR, T_C, and frequencies) are still entered into the 8970A, and the 8970A acts as a limited HP-IB controller to set the frequency of the external local oscillator. Naturally, all the corrections mentioned above are still made. For example, using the HP 8672A as an external local oscillator allows the user to make a swept, fully corrected noise figure and gain measurement from 2 to 18 GHz and have a flicker-free display on a nonstorage oscilloscope.

Because the 8970A can correct for second-stage noise figure, it can measure the noise figure of lossy devices such as mixers. With its broad frequency range, the 8970A can even measure noise figure as a function of intermediate frequency (IF).

For low-noise devices and for use in systems calculations, many users want their measurement results in terms of effective input noise temperature T_e instead of noise figure. The 8970A can accommodate them. In addition to noise figure in dB and T_e, the 8970A can also display noise figure as a ratio and Y factor (see below) as a ratio and in dB.

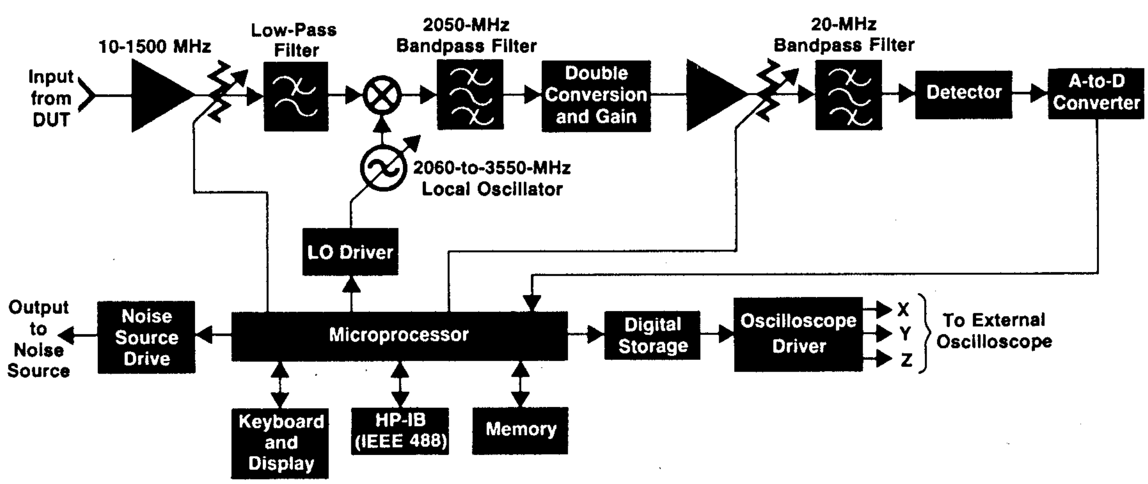

Fig. 2. *The 8970A Noise Figure Meter is basically a tunable power meter or receiver. The microprocessor controls the input and IF attenuators and the first LO, reads the analog-to-digital converter, provides output data to the digital storage circuits, and turns the noise source on and off.*

System Design

There are many ways to measure noise figure.[5] All techniques require at least two measurements of the output power from the device under test (DUT). For the first measurement, the signal connected to the input of the DUT is noise of a known power P_1. For the second, the signal is noise or a single-frequency continuous-wave (CW) signal of a different known power P_2.

The first thing to decide is whether to use noise or CW for P_2. The use of CW has several disadvantages: it requires that the bandwidth of the measurement be accurately known, it is difficult to know the absolute power of low-level CW signals, and a broadband (10 MHz to 18 GHz) CW source is rather expensive.

On the other hand, the advent of solid-state noise sources has provided a broadband, inexpensive source for P_2 that can be accurately calibrated in power level. Furthermore, the use of noise eliminates the need to know the measurement bandwidth accurately. Therefore, noise was chosen for P_2 in the 8970A.

The 8970A needs to measure the ratio of P_2 to P_1, known as the Y factor. There are many ways to do this, depending on whether P_2 is varied and whether Y is forced to be 2 (the 3-dB method) or not.[5] The most accurate way that is also convenient is to keep P_2 fixed, let Y vary as determined by the DUT, and measure the ratio of P_2 to P_1 with a power meter. This is the technique automated with the 8970A.

Although the quantity of interest is spot noise figure,[6] that is, noise figure at one particular frequency, in practice one can only measure noise averaged over a band of frequencies, the measurement bandwidth. Therefore, a noise figure meter must define the measurement bandwidth and select the center frequency of that bandwidth. Hence the 8970A is really a tunable power meter or receiver (see Fig. 2).

It is easiest and most accurate to do most of the signal processing at some intermediate frequency. Therefore, some frequency conversion is needed to translate the input frequency to that IF. To measure noise figure at only one frequency, the noise figure meter must reject the image and other responses of the input mixer.

There are two ways to do this. One is to use a preselector. The other is to have the first mixer up-convert the input signal to a first IF that is higher than any desired input frequency. This has two advantages over preselection. First, a simple fixed low-pass filter rather than a tunable low-pass or bandpass filter can be used. This is important when a range greater than two decades must be covered. Second, the first local oscillator (LO) has to tune less than an octave to provide greater than two decades of frequency coverage. Therefore, the 8970A uses the up-convert method.

As can be seen in Fig. 3, the 8970A down-converts from the first IF of 2050 MHz to the last IF of 20 MHz in two steps. This helps eliminate undesired responses. An isolator be-

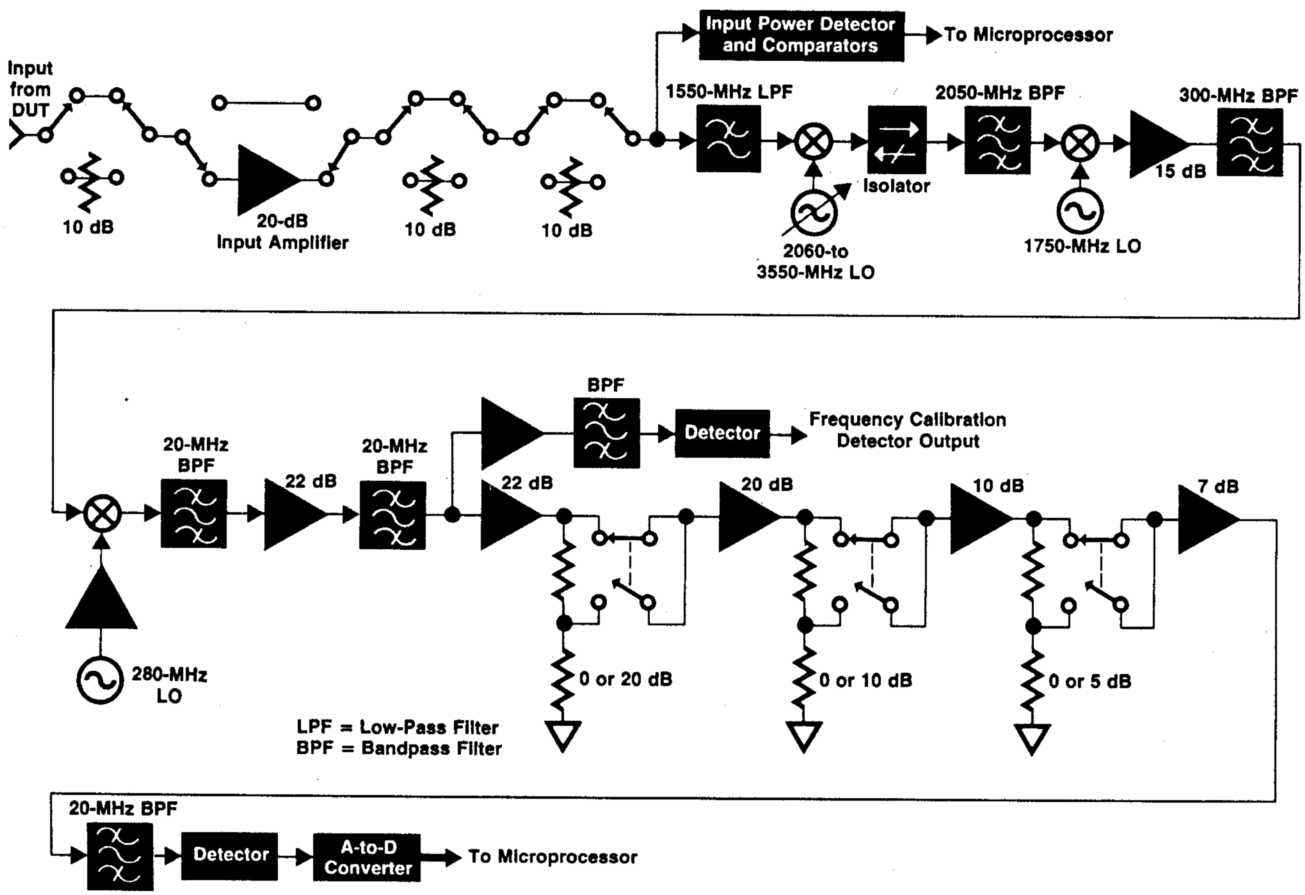

Fig. 3. *8970A RF block diagram. Input signals are converted to a 20-MHz IF for processing. Gain, filtering, and step attenuation are distributed along the IF chain.*

A Noise Source for Noise Figure Measurements

The noise source used for a noise figure measurement is the standard of the measurement, that is, the calibration and accuracy of the noise source are transferred directly to the measurement. For the noise characteristics of devices to be easily verifiable at vendors' and customers' plants, noise sources must be stable and accurately calibrated.

The source impedance of an ideal noise source is exactly equal to the characteristic impedance of the system in which the device under test is intended to operate—50 ohms is the most common value. Because the measured noise figure of many devices is a sensitive function of the source impedance, any deviation of the noise source impedance from the nominal value introduces errors into the measurement. These errors require correction if accurate comparisons are to be made using different measurement systems. Therefore, for rapid, convenient, easily comparable measurements, the noise source should have an impedance as close to 50Ω as possible. The source should also be useful over a wide frequency range.

These requirements governed the design of the 346B Noise Source and had considerable impact on the specifics of its implementation. The noise generator in the 346B is a specially constructed silicon diode operated in avalanche breakdown to achieve the hot temperature or on condition.[1,2] To ensure that the diode breakdown characteristics and the resulting noise spectrum are consistent from diode to diode and each time each diode is turned on, the anode is surrounded with a guard ring, as depicted in Fig. 1. The guard ring region has a higher breakdown voltage than the anode and confines the breakdown region to the area directly under the anode, thereby eliminating edge effects that can cause fluctuations in the noise spectrum. The active area of the diode is small to minimize the effects of material inhomogeneities.

The diode chip is mounted directly on the metal cartridge for optimum heat sinking to maintain the lowest possible junction temperature and best long-term stability of the calibration.[3,4]

It was desired that the 346B Noise Source have wide bandwidth, both for measurement convenience and for the economy resulting from needing only one source for a broad range of applications. However, the diode impedance varies greatly over the desired frequency range of 10 MHz to 18 GHz, so it was essential to develop an imbedding circuit that matches the diode impedance to 50 ohms and simultaneously provides a flat ENR-versus-frequency characteristic. This is done by using a combination of loss and reactive matching as indicated in the equivalent circuit shown in Fig. 2.

As shown in Fig. 3, the noise diode and matching circuit are sealed in a hermetic package. A wideband hermetic feedthrough

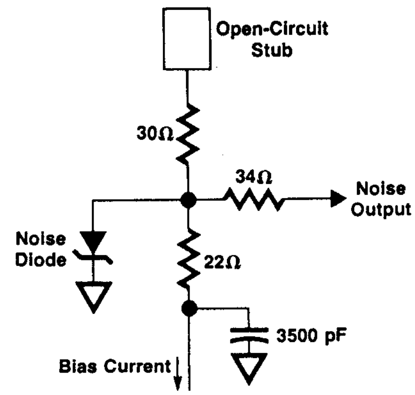

Fig. 2. *Schematic diagram of the broadband impedance-matching imbedding circuit for the 0.01-to-18-GHz noise diode.*

was specially developed to provide a low-SWR transmission structure and to form part of the mechanical transition from the 3.5-mm coaxial structure external to the noise cartridge to the microstrip matching circuit within the cartridge. An internal transition from 3.5 mm to 7 mm is used when type N or APC-7 connectors are desired. To improve the source match, a 6-dB pad is used between the external connector and the noise cartridge, providing an additional 12-dB improvement in the return loss. Fig. 4 shows the typical source match of the 346B Noise Source.

The available noise power produced by the avalanche process is inversely proportional to the diode current. Therefore, it is important that the current be well controlled to obtain repeatable results. The 346B contains a built-in current regulator to ensure the correct value of diode current under all conditions, such as might occur with different drive cable lengths or operation from a supply separate from the 8970A. The diode current for each 346B is factory adjusted to the optimum value during the calibration procedure.

ENR Calibration

Because solid-state noise sources are not fundamental standards, they must be calibrated using a fundamental standard, namely a physically heated or cooled resistor. The United States National Bureau of Standards (NBS) provides regular calibration service only at the frequencies shown in Fig. 5, using resistive terminations that are at the temperatures listed.

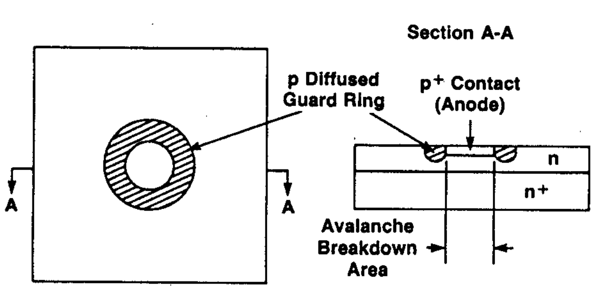

Fig. 1. *Noise diode guard ring detail. The guard ring ensures consistent avalanche breakdown characteristics.*

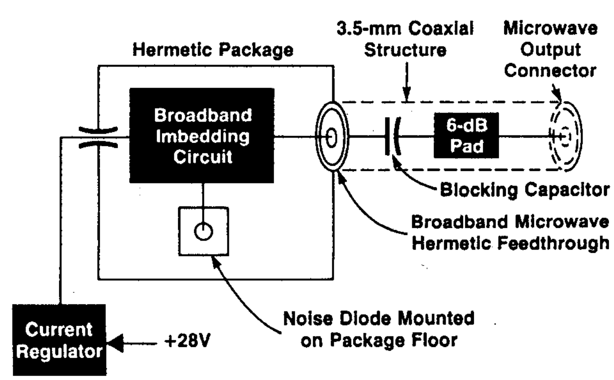

Fig. 3. *Block diagram of the HP 346B Noise Source.*

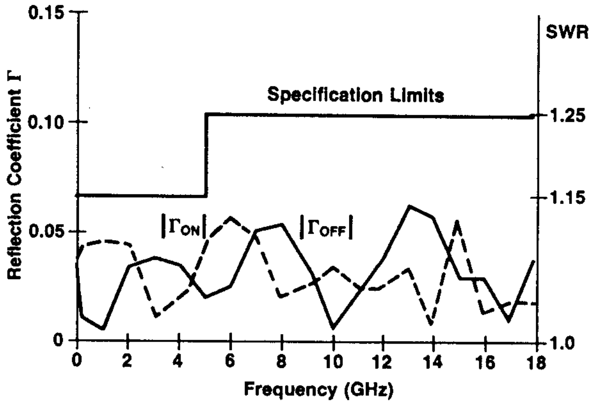

Fig. 4. *Typical source match of the 346B Noise Source.*

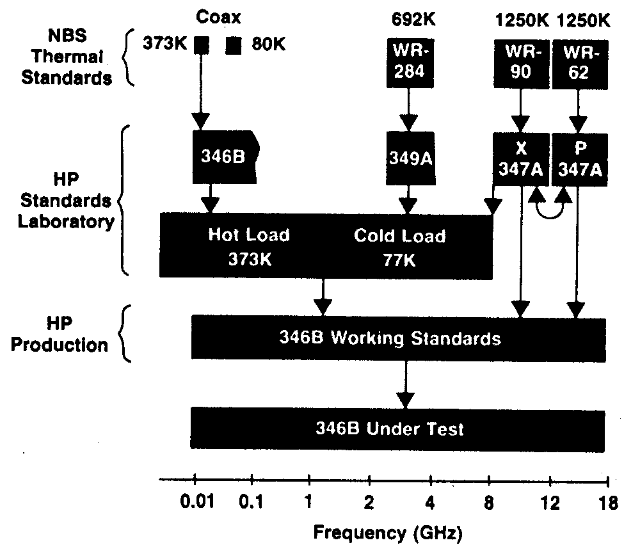

Fig. 5. *Traceability of 346B Noise Source calibration to the United States National Bureau of Standards.*

The standards laboratory of Hewlett-Packard's Stanford Park Division regularly sends its transfer standard noise sources to NBS for calibration. For the frequencies at which NBS does not provide service, the SPD standards lab maintains hot and cold loads for use as standards. Cross checks are done between the transfer standards and the hot and cold loads and between the two waveguide transfer standards to assure the highest possible confidence.

Then the transfer standards and the hot and cold loads are used to calibrate four solid-state noise sources (346Bs) to be used as working standards in production. These working standards (one for each connector option) are used to calibrate the units for shipment.

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-Donald R. Chambers

tween the first mixer and first IF filter minimizes reflections and assures a smooth frequency response. The losses of the first two mixers, the isolator, and the 1550-MHz and 2050-MHz filters, combined with the noise figure of the 15-dB, 300-MHz amplifier result in a 15-dB noise figure looking into the first mixer.

Adding a low-noise 20-dB amplifier to the input reduces this noise figure to about 6 dB at low frequencies. High input powers from high-gain devices under test make it necessary to switch out the input amplifier and add some attenuation. The microprocessor reads a detector at the input to the first mixer to keep the level incident on the mixer below −20 dBm when the noise source is turned on.

Filtering is distributed in the 20-MHz IF. The first filters keep the 280-MHz LO feedthrough and the total broadband noise from overloading the amplifier stages. The last filter makes sure the detector responds to the noise in only the 4-MHz bandwidth. The step attenuators are also distributed so the noise figure can be kept low and the amplifiers are not overdriven. These IF step attenuators and the input RF step attenuators are kept fixed as the noise source is turned on and off for P_2 and P_1. Thus their attenuation errors do not affect the Y-factor accuracy.

There are several tradeoffs in selecting the IF bandwidth and the integration time of the analog-to-digital converter (ADC) that measures the detected power level. Whenever noise power is detected (converted to dc), the resulting dc is noisy. For the 8970A's detector the noise-to-signal ratio is $1/\sqrt{B\tau}$, where B is the predetection (IF) noise bandwidth and τ is the postdetection filter time constant. Thus, for minimum display jitter, both B and τ should be maximized. A large B also minimizes susceptibility to electromagnetic interference (EMI). Because the "signal" is noise, a larger B makes a larger "signal."

An ADC that integrates the input signal can provide the needed postdetection filtering. This is advantageous because the microprocessor can turn the integrator on and off in synchronism with the noise source drive. Thus, the 8970A does not have to wait many time constants for a separate filter to reach steady state before the A-to-D conversion can take place. Filtering and conversion happen at the same time with one circuit doing both. In this case, the filter time constant τ is just the integration time. Therefore, a large τ also makes it easier to build an ADC with the $\geq$14 bits required to make its quantization error less than 0.005 dB.

On the other hand, if B is too large, the measurement is no longer of spot noise figure, especially at the lowest RF measurement frequency of 10 MHz. Furthermore, it becomes harder to measure narrowband units. If τ is too large, the reading update rate will be too low, making it hard to adjust or tune a circuit for minimum noise figure.

Verifying the 8970A's Accuracy in Production

The signal path of the 8970A Noise Figure Meter can be described as a receiver that has up to 100 dB of very stable gain, followed by a noise power detector and an analog-to-digital converter. To verify that the 8970A meets its noise figure accuracy specifications, the linearity of these circuits must be measured during production testing.

To measure the 8970A's amplifier, detector, and analog-to-digital converter linearity over a 22-dB range with 0.001-dB resolution, the automatic test system shown in Fig. 1 is used. Typical repeatability of this system is about 0.005 dB.

The test system measures the 8970A's linearity by comparing the output of the 8970A's detector to that of an 8484A Power Sensor and a 436A Power Meter as their common input power level is stepped over the 22-dB range. Using a power meter to measure the input power to the 8970A at each step eliminates errors caused by test system input attenuator inaccuracies, nonrepeatability, and mismatch.

Errors caused by nonlinearity in the power sensor are greatly reduced by operating the sensor over only a 5-dB range, from −38 dBm to −43 dBm. Over this range, the sensor and power meter are linear within 0.005 dB and measurement noise and drift are negligible. Reading the power meter's recorder output with the 3456A Digital Voltmeter gives greater resolution than is available from the power meter's HP-IB output. The entire reference channel is periodically calibrated by the HP standards laboratory.

After the correct starting power levels are set up, readings are taken from each channel. Then the test system input attenuation is increased and the change in power in both channels is measured. This change in noise power should be equal for both the reference and test channels. Any difference between the two is the linearity error of the 8970A under test (see Fig. 9 on page 31).

This is repeated until the test system input attenuation has been increased by 5 dB. Then the reference channel attenuation is decreased by 5 dB, returning the sensor and power meter to the correct operating range for the next measurement. Whenever this attenuator is changed, errors can be introduced by its nonrepeatabilities and mismatch errors and by power splitter tracking errors. To reduce these errors, the test program makes a new reference measurement each time the reference channel attenuator is stepped.

Because noise is used as the "signal," there is noise or jitter in the reference and test channel measurements. This jitter is reduced by using the internal averaging available in the 8970A and the 3456A. Also, a curve fit is done on the final data.

In making a noise figure measurement, the 8970A must make two power measurements—one with the noise source on and one with it off. Their ratio is the Y factor, and its error can be read from the linearity error curve by taking the difference between the errors at the two power levels. For each possible Y factor, the test system program determines Y-factor error for all possible allowed combinations of the two powers from the linearity curve. Then, the worst error is found for each Y factor. Next, the ideal noise figure is calculated for each Y factor. Noise figure is also calculated using these Y factors and the previously calculated worst-case Y-factor errors. The difference between these two noise figures is the 8970A's noise figure measurement error, Fig. 2.

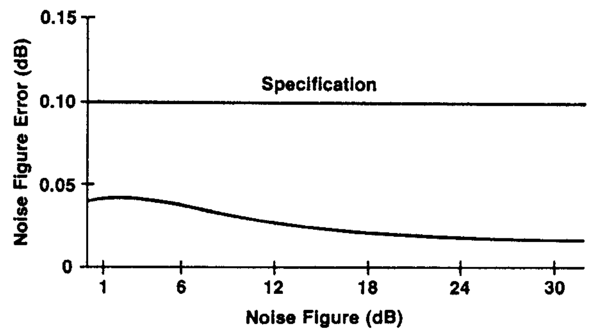

Fig. 2. *8970A noise figure error versus noise figure.*

The test system also measures the 20-MHz IF attenuator error, and by combining this with the detector linearity error (Fig. 9, page 31), determines the gain measurement accuracy of the 8970A.

A major benefit of this test system is its flexibility to change. The test noise source can be replaced by a signal generator or broadband noise source. By readjusting the input, reference, and test attenuator levels, power linearity measurements can be made over wide frequency ranges for many devices such as power sensors and microwave detectors.

-Harry Bunting

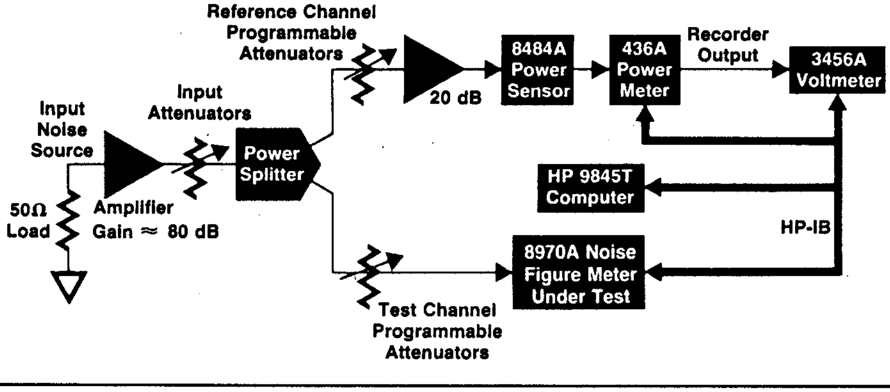

Fig. 1. *Test setup for verifying the linearity of the amplifiers, detector, and analog-to-digital converter of the 8970A Noise Figure Meter.*

Clearly, a compromise is required. The 8970A has an IF bandwidth B of 4 MHz and an ADC integration time τ of 16 ms. These combine to give a low jitter of 0.05 dB peak, a reasonable spot size, and a reading rate of about five per second, which includes time for two A-to-D conversions and time for the microprocessor to calculate, display, and perform other functions. This compromise made the EMI shielding not too difficult and eased the design of the 16-bit ADC. If less jitter is required, software smoothing can be selected.

Input Assembly

The input assembly (see Fig. 4) serves as a low-noise, broadband, variable-gain front end for the rest of the instrument. It consists of three switchable 10-dB attenuators, a switchable 20-dB low-noise, broadband amplifier, and a power detector. Input noise power incident on the first converter is detected and converted to a dc voltage by the input power detector circuit. This voltage is then sent to a pair of comparators that signal the controller assembly that the input power is either too high, at an acceptable level, or too low. The controller uses this information to adjust the overall gain of the input assembly appropriately. Gain of the assembly is variable in 10-dB steps from −30 dB to +20 dB. Optimum power level to the first converter is between −30 dBm and −20 dBm.

A few circuit details of the input amplifier are noteworthy. The amplifier consists of two independent high-frequency-transistor series-shunt feedback stages. This configuration gives an acceptable compromise for overall amplifier noise figure, gain flatness, input match, and bandwidth for the number of stages used. The input amplifier is realized using printed circuit board technology. This requires parallel combinations of capacitors and resistors in several places to reduce the effects of parasitic lead inductance. The variable inductor in the collector of transistor Q1 is realized by a variable-length high-impedance transmission line. It is used to adjust amplifier gain at high frequencies. Performance specifications for the amplifier are:

Frequency Range	10 to 1500 MHz
Input SWR	<1.5
Gain	20 dB ±1 dB
Noise Figure	3.5 dB at 10 MHz 5.0 dB at 1500 MHz
P_{out} at 1 dB compression	>10 dBm

The input power detector consists of a zero-bias Schottky detector diode operating in the square-law region followed by a high-dc-gain operational amplifier stage. Feedback around the op amp includes a thermistor to compensate for changes in the detection sensitivity of the Schottky diode with temperature.

20-MHz IF Assembly

The 20-MHz IF circuit (see Fig. 3) consists of a series of filters, amplifiers and attenuators that determine the bandwidth and power level of the 20-MHz noise signal sent to the noise power detector. Nominal bandwidth of the

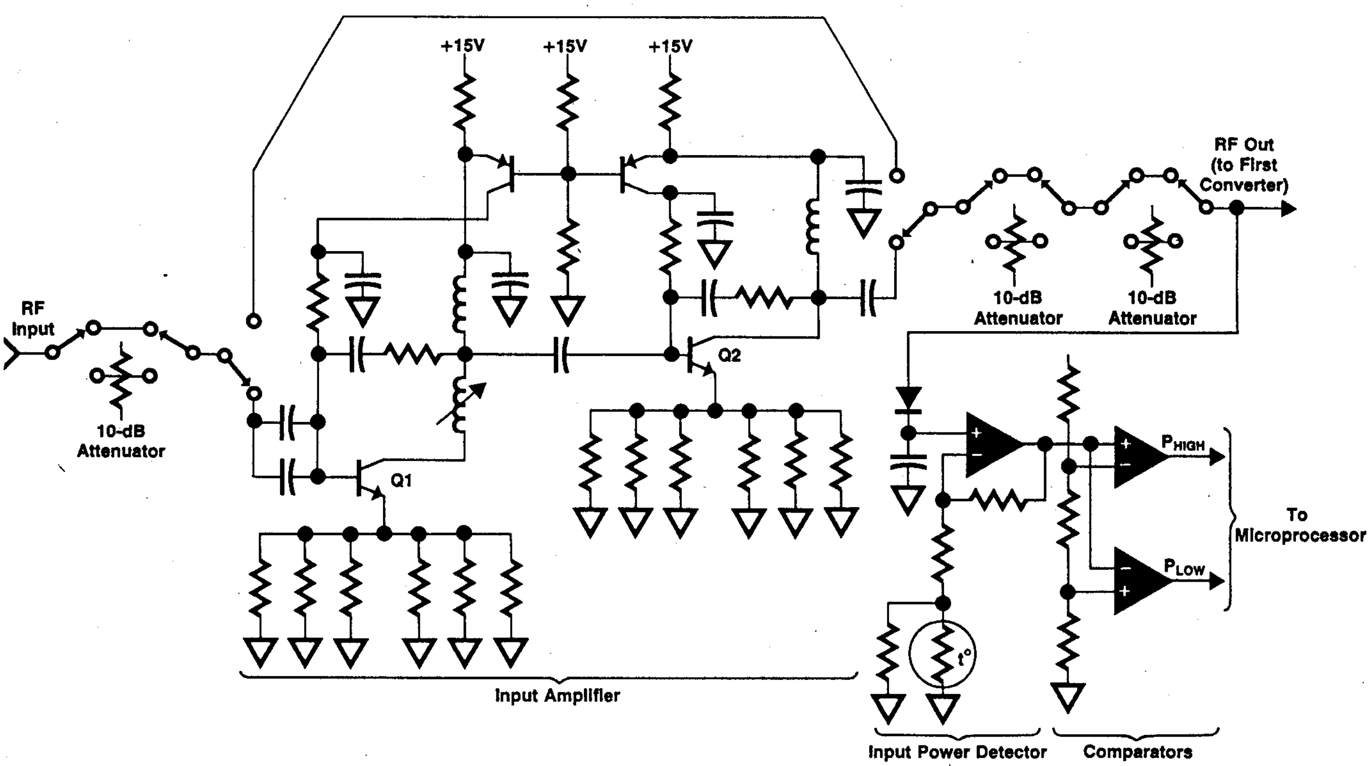

Fig. 4. *The 8970A's input assembly serves as a low-noise, broadband, variable-gain front end. It consists of three 10-dB attenuators, a 20-dB amplifier, and a power detector.*

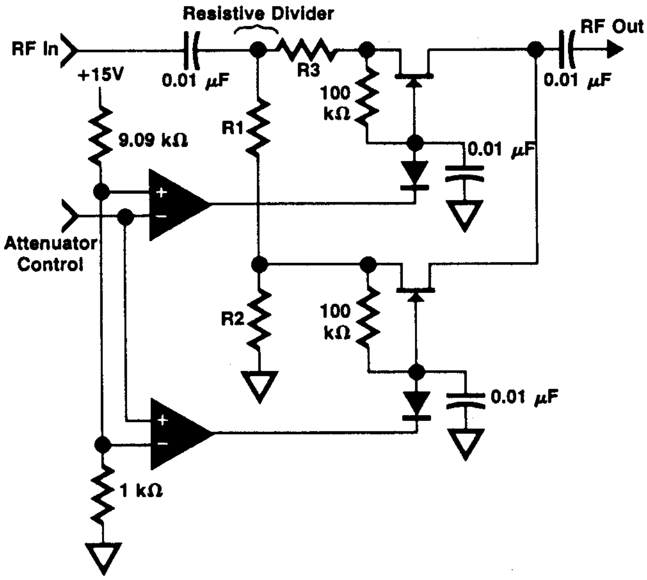

Fig. 5. *An attenuator section representative of the 20, 10, and 5-dB step attenuators in the 20-MHz IF circuit.*

complete circuit is about 4 MHz. Gain of the 20-MHz IF circuit is variable under control of the microprocessor in 5-dB steps from +40 dB to +75 dB. A detector is included in the IF circuit to detect feedthrough from the first LO when the LO is tuned near 2050 MHz during a frequency calibration.

The two bandpass filters at the front end of the 20-MHz IF circuit combine with the bandpass filter at the output of the 7-dB amplifier to form a composite eight-pole filter. These filters determine the nominal 4-MHz bandwidth of the

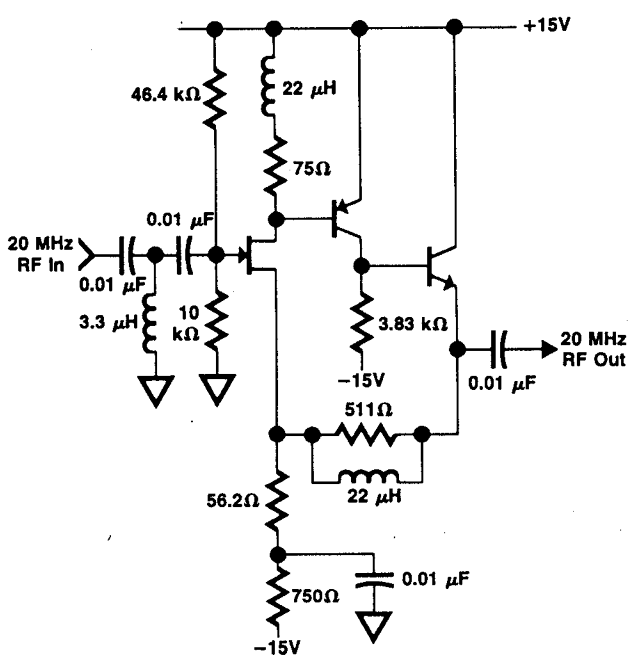

Fig. 6. *Representative 20-MHz IF amplifier. Feedback creates a high input impedance and a low output impedance and determines the gain.*

complete IF circuit. They also serve to reject unwanted first-LO feedthrough.

The 20, 10 and 5-dB attenuators are realized by simple resistive dividers that are switched by complementarily driven FET stages. A representative attenuator is shown in the schematic of Fig. 5. The state of the control line causes either the upper or the lower FET to be switched into the on state, thereby coupling either the unattenuated or the attenuated version of the input signal to the output. Resistor R3 causes the attenuator to present a nearly constant impedance to the following stage regardless of the attenuator's state. This nearly constant impedance minimizes the error resulting from loading by the following stage. This attenuation scheme was selected over a mechanically switched attenuator for reasons of cost, repeatability, and reliability.

The method chosen for attenuation imposes some requirements on the design of the amplifiers between the attenuator sections. Each of these amplifiers must have a low output impedance to appear as an ideal voltage-source driver for the resistive divider of the attenuator. Each amplifier must also have a high input impedance to minimize loading effects on the attenuator stage preceding it. The schematic for these amplifiers is shown in Fig. 6. The FET input transistor, feedback topology, and shunt resonating input inductor combine to realize the required high input impedance. The emitter follower output stage and feedback serve to lower the output impedance of the amplifier.

Frequency calibration of the 8970A is accomplished by tuning the first LO near 2050 MHz and then observing the resultant LO feedthrough by means of the frequency calibration detector located in the 20-MHz IF assembly (see Fig. 3). Feedthrough from the first LO is passed through a narrow bandpass filter centered in the 20-MHz IF passband. Output from this filter is detected, resulting in a dc level that is related to LO frequency. The maximum dc level corresponds to the center of the 20-MHz IF passband or zero instrument input frequency. If the LO is tuned below 2050 MHz and then stepped up in frequency, the output of the frequency calibration detector always increases in level relative to the preceding output until a maximum is reached at 2050 MHz. After this, the output begins to decrease. Only information about whether the output of the frequency calibration detector is increasing or decreasing is really necessary to locate the zero frequency point. This allows the use of an economical slope detector circuit (see Fig. 7). A

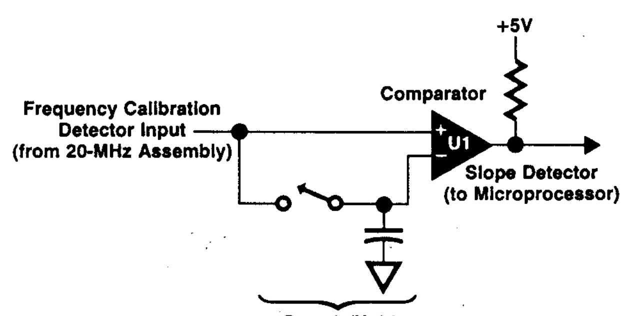

Fig. 7. *A simple, economical slope detector is used in automatic frequency calibration.*

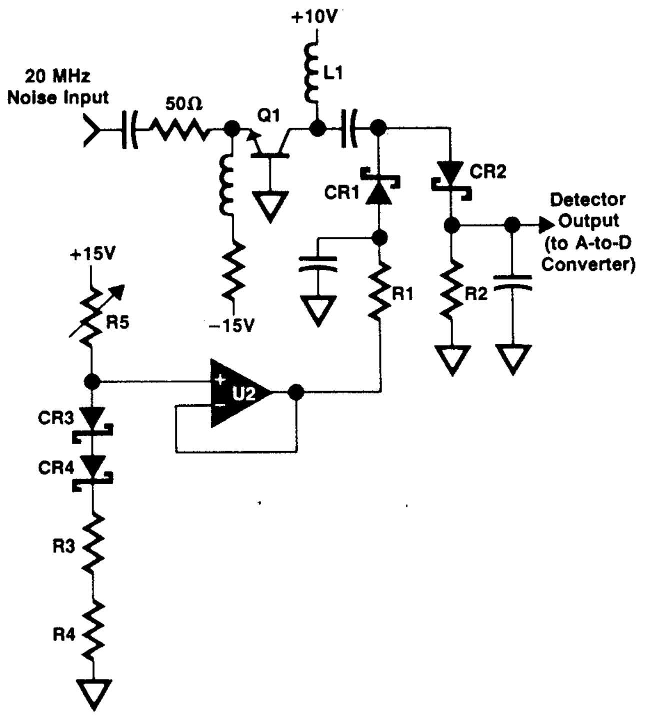

Fig. 8. *The noise power detector converts RF noise signals to dc levels that the microprocessor can measure by means of the analog-to-digital converter.*

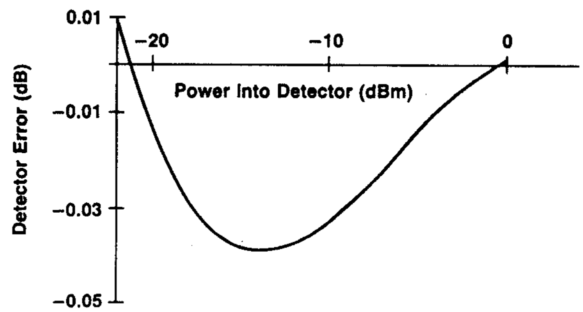

Fig. 9. *Typical performance curve for the noise power detector.*

sample-and-hold circuit saves the previous output of the detector circuit and U1 compares it to the current output. As long as successive levels increase, U1's output to the controller remains in a high state. At the maximum detector output level the comparator switches state, signaling the controller that zero instrument frequency has been located.

Noise Power Detector Assembly

The noise power detector circuit, shown in Fig. 8, converts RF noise signals to dc levels that the controller can interpret by means of the analog-to-digital converter (ADC). The basic detector consists of a common-base transistor stage Q1 driving a pair of rectifying Schottky diodes CR1 and CR2. The common-base stage provides a good 50Ω termination for the last 20-MHz IF bandpass filter while at the same time functioning as a high-impedance current source driver for the detector diodes. Diode CR2 rectifies positive-going half-cycles, producing an average dc current in resistor R2. The resultant dc voltage across this resistor is the detected signal that is sent to the ADC and is proportional to the square root of the RF noise power. Diode CR1 and resistor R1 are present to provide a symmetrical load for the negative-going half-cycles. Inductor L1 resonates with the capacitance present at the collector of Q1 to realize more closely the high-impedance current source needed for the diodes.

Detector linearity is improved by biasing the diodes slightly. This is accomplished by applying a voltage to R1, CR1, CR2, and R2 with op amp U2. This voltage must be temperature compensated to ensure that detector sensitivity remains constant. R5 drives R3, R4, CR3, and CR4 with a constant current. The resultant voltage at the noninverting terminal of U2 is sensed by this unity-gain op amp stage and applied to the detector circuit.

The performance of the noise power detector is quite good. Its dynamic range extends from 0 dBm to about −22 dBm. Deviation from linearity over this range is less than 0.04 dB. A typical performance curve for the detector is shown in Fig. 9.

Applications

Fig. 10 shows an example of a VHF measurement using the 8970A Noise Figure Meter. The meter is making a swept measurement of gain and noise figure from 10 to 500 MHz. No additional amplifiers, local oscillators, filters, or mixers are needed. The 8970A is automatically correcting for ENR variations with frequency, using the correct T_C, and removing the second-stage noise. The results are digitally stored in memory and are refreshed at a rapid rate for display on a nonstorage oscilloscope.

To measure at frequencies above 1500 MHz, an external mixer and LO are required (Fig. 11). However, the 8970A controls the LO's frequency and level via the HP-IB. Thus all corrections can be made and the results digitally stored for the oscilloscope display. In this measurement the 8970A

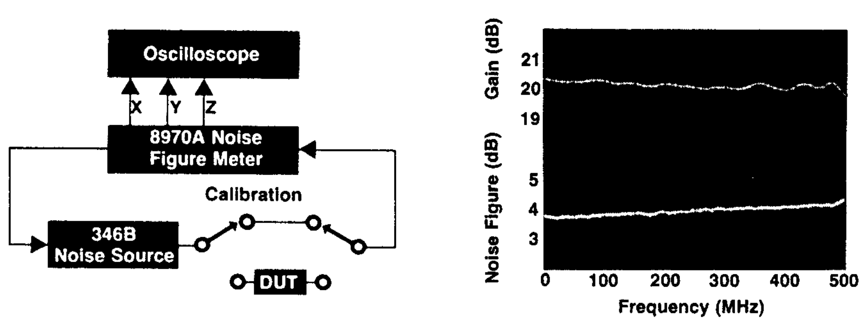

Fig. 10. *A swept measurement of gain and noise figure from 10 to 500 MHz.*

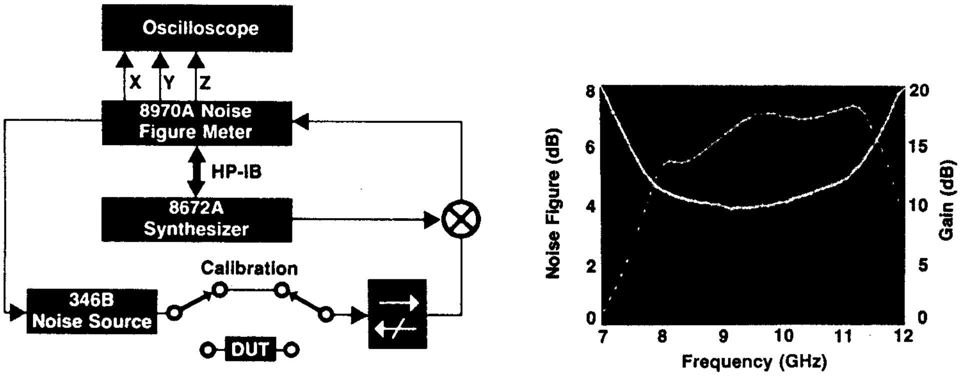

Fig. 11. *Using an external local oscillator (an HP 8672A) and mixer to measure above 1500 MHz.*

is tuned to 30 MHz and tunes the 8672A local oscillator from 7 to 12 GHz. An isolator in front of the RF port of the mixer reduces the mismatch uncertainty of the gain measurement and reduces the sensitivity of the mixer's noise figure to the output impedance of the device under test.

Electromagnetic interference (EMI) can cause errors of several dB. A spurious signal will add proportionately more power to the noise-source-off power measurement than to the noise-source-on measurement. This decreases the Y factor and thus increases the measured noise figure.

EMI can be detected by making swept noise figure measurements. For wide frequency sweeps, EMI appears as peaks at the frequencies of the spurious signals. Narrow sweeps over frequencies where a broad band of interference exists may show broad humps.

Fig. 12 shows the effect of EMI on a DUT that is not in a tightly shielded box. EMI can also enter through connectors (such as BNC), through single-shielded coax, or over the power supply leads, and can be picked up by tuners during transistor measurements.

Acknowledgments

Product manager Mike Cuevas' research paper helped launch the 8970A project. John Page, R&D manager during the project, encouraged us to go for a broad frequency range and external LO control via the HP-IB. Steve Adam was section manager during the development phase. Nick Kuhn, product marketing engineer, provided many valuable inputs that helped define the product. Mike Pozzi designed the digital and low-frequency circuits including the ADC and the CRT driver. Takashi Yoshida developed most of the 24K bytes of firmware. Tim Kelly and Patty Decker Morton did the product design. Production section manager Tom Neal assembled a fine crew including Sandi Molina, assembly supervisor, and Ray Lew, test supervisor, that achieved a smooth buildup to volume production. Production engineer Mark Haaland showed fine attention to detail in helping introduce the 8970A to production. Harry Bunting, production engineer, built the test system, provided advice on producibility, and helped in environmental testing.

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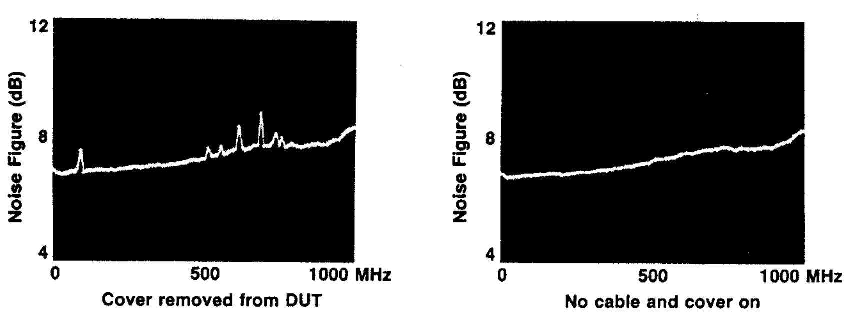

Fig. 12. *Electromagnetic interference appears as peaks during swept measurements of noise figure.*

Appendix
Noise Figure Basics

Noise limits the detection of weak signals. As a signal passes through each stage of a system, noise is added. Thus the signal-to-noise ratio is reduced. Noise figure is a measure of this degradation.

Two main mechanisms cause the noise that is responsible for the signal degradation measured by noise figure. The first is the random motion of charge carriers caused by thermal agitation. This noise is called thermal noise, Nyquist noise, or Johnson noise.[1,2] It exists in all conductors whose temperature is above absolute zero.

This thermal noise generates an available power of kTB where k is Boltzmann's constant, T is absolute temperature, and B is bandwidth. This is the power that can be delivered to a conjugately matched load. Note that the available thermal noise power is independent of the resistance R. If the resistor is at 290K, the available power is 4.00×10^{-21} watts (or −174 dBm) in a 1-Hz bandwidth.

The second mechanism is the random, corpuscular flow of carriers in transistors and diodes. Each charge carrier (electron or hole) is emitted individually and flows across the device on its own. Thus the current arriving at the collecting electrode is a series of little pulses randomly distributed in time, each having the charge of one electron. This is a noisy process, and the noise generated is called shot noise.[3]

Noise figure is defined as the total output noise of the device under test (DUT) caused by both the DUT and the source, divided by the output noise caused by the source alone.[4] Because this ratio depends on the amount of noise produced by the source, it is essential to specify that noise. Therefore, noise figure is defined referred to a source that has the available output noise power of a resistor at 290K.*

$$F = \left. \frac{N_{out} \text{ caused by DUT} + N_{out} \text{ caused by source}}{N_{out} \text{ caused by source}} \right|_{T_{source} = 290K}$$

Device noise can be modeled using the fact that the available noise power produced by a resistor is kTB. In Fig. 1a, the total output noise power is the sum of the noise caused by the source (kT_sB) and the noise added by the noisy amplifier with gain G. To model this added noise, the amplifier is made noiseless and an additional fictitious resistor is added at the input, as shown in Fig. 1b. The temperature of this resistor, T_e, is chosen so that the total output noise power remains the same as from the noisy amplifier in Fig. 1a. In this model, the power from both resistors is added, and neither loads the output of the other.

T_e is called the effective input noise temperature.[5] It is very useful in systems analysis and in specifying very low-noise components. The relationship between T_e and noise figure will be derived later.

To measure T_e, one might try connecting a power meter to the output of the device, measuring $P=k(T_s+T_e)BG$, and solving for T_e. However, BG is not usually known. B is really the noise bandwidth, which is the integral of the gain over frequency divided by the gain at the center of the band. In addition, it is difficult to measure absolute power accurately.

Because there are really two unknowns (T_e and BG), two independent equations are needed. Making another power measurement with the source resistor at a different temperature will yield a second equation, independent of the first. Thus P_2 and P_1 are measured with T_s equal to a hot (T_H) and a cold (T_C) temperature, respectively.

$$P_2 = k(T_H+T_e)BG$$

$$P_1 = k(T_C+T_e)BG$$

Dividing these two equations causes the unknown BG to cancel out. An added benefit is that only the ratio of two powers is required and not absolute power. This ratio is called the Y factor.

$$\frac{P_2}{P_1} = Y = \frac{T_H+T_e}{T_C+T_e}$$

Solving for T_e yields

$$T_e = \frac{T_H-YT_C}{Y-1}$$

This involves only the two known source temperatures and the measured Y factor.

The definition of noise figure and the model of Fig. 1b are now used to express noise figure in terms of T_e. The model shows that the output noise caused by the device is kT_eBG, and the output noise caused by the source is kT_sBG. Because noise figure is referred to a source at 290K, the output noise caused by the source becomes k290BG.

Putting these into the definition and solving yields F in terms of T_e.

$$F = \frac{kT_eBG + k290BG}{k290BG}$$

*This definition applies only to single-input, single-output devices. For devices that have multiple inputs, outputs, and/or responses, see reference 4.

$$F = 1 + T_e/290$$

Inserting the equation for T_e gives F in terms of Y, T_H, and T_C.

$$F = 1 + \frac{T_H-YT_C}{290(Y-1)}$$

$$F = \frac{(T_H/290-1) - Y(T_C/290-1)}{Y-1}$$

10 log (T_H/290−1) is defined as the excess noise ratio or ENR of the source. The T_H or on temperature of an electronic noise source is usually specified by its ENR.

$$ENR = 10 \log (T_H/290-1)$$

If the cold source temperature T_C is approximately 290K, then the second term in the numerator of the equation for F is approximately zero, and noise figure in dB simplifies to:

$$F_{dB} \approx ENR - 10 \log (Y-1)$$

In general T_C is not 290K, and making this approximation can result in an error of 0.1 dB or more.

The method just described for measuring noise figure requires changing the effective temperature of the source from T_H to T_C. It is difficult and slow to heat a resistor physically. A solid-state noise source provides a convenient solution. The kT_HB noise is generated by avalanche breakdown in a diode driven by a constant current. The noise can be switched on and off by turning the current on and off. An attenuator at the output provides a good match and serves as the source of noise (kT_CB) when the diode is off.

Noise Figure of Cascaded Networks or Second-Stage Contribution

Noise contributed by stages following the device under test can cause an error in the measurement of noise figure. Again, this effect can be modeled using the concept of effective input noise temperature T_e. Because k and B will be the same everywhere, the noise powers can be simply modeled using temperatures only. Consider two cascaded amplifiers with gains G_1 and G_2 connected to a source at temperature T_s. Let T_{e1} and T_{e2} represent the noise added by the two amplifiers referred to their inputs, and let T_s represent the source noise. The total noise output of the cascaded system is $k(T_s + T_{e1})G_1G_2B + kT_{e2}G_2B$.

Next refer the noise of the second amplifier, T_{e2}, all the way back to the input of the first amplifier. To do this, divide T_{e2} by G_1, yielding T_{e2}/G_1.

Now the total effective input noise temperature, T_{eT}, is just the sum of all the noise temperatures except the source, T_s, at the input of the first amplifier.

$$T_{eT} = T_{e1} + T_{e2}/G_1$$

Solving $F = 1 + T_e/290$ for T_e and putting this in for each T_e yields the expression for the total noise figure for the cascaded network.

$$T_e = 290\,(F-1)$$

$$F_T - 1 = F_1 - 1 + (F_2-1)/G_1$$

$$F_T = F_1 + (F_2-1)/G_1$$

Thus, the second stage causes an error equal to $(F_2-1)/G_1$. This error can be removed if F_2 and G_1 can be measured. F_2 is easy to measure—just connect the noise source and read the result.

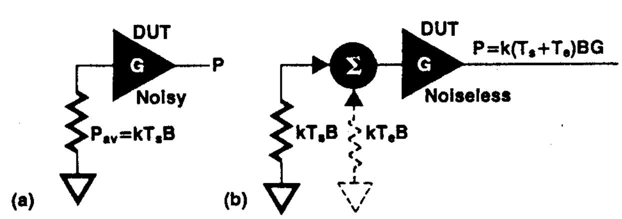

Fig. 1. *How effective noise temperature T_e is defined.*

How the 8970A Measures Gain

Gain measurement requires a little more work. All gain measurements require two steps: connect a signal source to a receiver and set a reference. Then insert the DUT and read the change in output power. The same thing can be done using a noise source as the signal source. However, because the "signal" power levels are so near the inherent noise levels, the noise figures of the DUT and meter must be taken into account. This yields four unknowns: T_{eD}, T_{eM}, G_D, G_M. Four power measurements are needed to set up four equations.

1. Connect noise source to 8970A
   Measure: $P_1 = k(T_C + T_{eM})BG_M$
   $P_2 = k(T_H + T_{eM})BG_M$

2. Insert DUT
   Measure: $P_3 = k(T_C + T_{eT})BG_DG_M$
   $P_4 = k(T_H + T_{eT})BG_DG_M$

T_{eM} and T_{eT} are eliminated by subtracting P_1 from P_2 and P_3 from P_4. Then kBG_M is eliminated by dividing the two differences. The result is the gain of the DUT, G_D, which can now be used in subtracting out the second-stage noise figure.

$$\frac{P_4 - P_3}{P_2 - P_1} = \frac{k(T_H - T_C)BG_DG_M}{k(T_H - T_C)BG_M}$$

$$G_D = \frac{P_4 - P_3}{P_2 - P_1}$$

This technique also allows the measurement of the noise figure and gain (loss) of devices whose gain is less than unity, such as mixers and attenuators.

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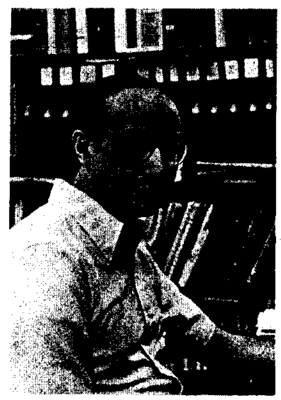

Howard L. Swain

Howard Swain was project manager for the 8970A Noise Figure Meter. He received his BSEE degree from Iowa State University in 1969 and his MSEE from Stanford University in 1972. With HP since 1969, he has designed several low-noise oscillators and contributed to the design of the 8640B Signal Generator. He's a member of the IEEE and the author of a paper on accurate noise figure measurement techniques, and has been a guest lecturer at Stanford on RF amplifier design. Born in Des Moines, Iowa, he now lives in Palo Alto, California and enjoys square dancing, table tennis and playing a harpsichord and a pipe organ that he built from kits.