Residual Phase Noise and AM Noise Measurement Techniques

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Acknowledgements

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Contents

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1

General Information

What is residual two-port noise?

Residual two-port noise is the noise added to a signal when the signal is processed by a two-port device. Such devices include: amplifiers, dividers, filters, mixers, multipliers, phase-locked loop synthesizers and any other two-port electronic network. Residual two-port noise contains both AM and ΦM components.

Residual two-port noise is the sum of two basic noise mechanisms:

1. Additive noise: This noise is generated by the two-port device, at or near the signal frequency, which adds in a linear fashion to the signal.

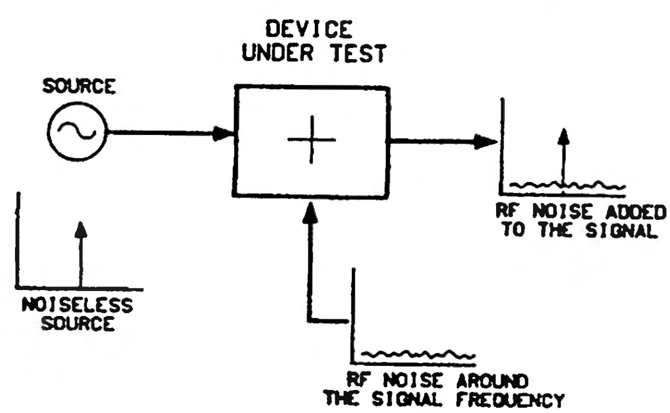


Figure 1-1. Additive Noise Components

2. Multiplicative noise: This noise has at least two mechanisms. The first is an intrinsic, direct, phase modulation with a $\frac{1}{f}$ spectral density, the origin of which is unknown. The second, in the case of amplifiers or multipliers, is noise which may modulate an RF signal by the multiplication of baseband noise with the signal. This mixing is due to non-linearities in the two-port network. The baseband noise may be produced by the active devices of the internal network, or may come from low-frequency noise on the signal or power supply.

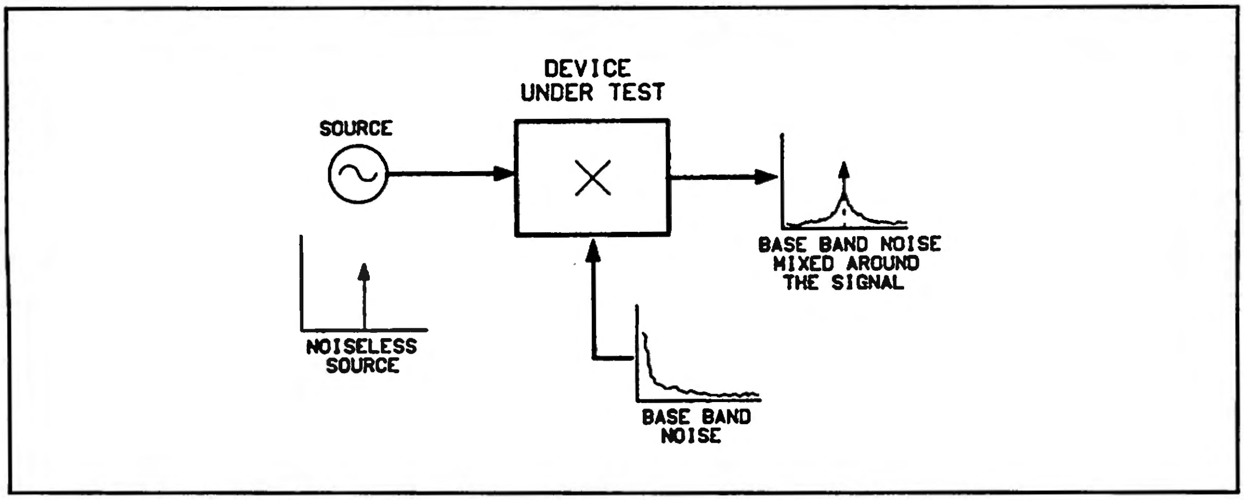


Figure 1-2. Multiplicative Noise Components

Why are residual and AM noise measurements important?

- In recent years it has become apparent to primary contractors that to ensure overall system noise performance, residual noise must be specified for all subsystems.
- The absolute noise of an oscillator is set by the residual noise of the active device, the residual noise of the resonator, and the bandwidth of the resonator.
- Oscillator noise is degraded by the residual noise of all the devices that follow it: amplifiers, dividers, filters, mixers, multipliers, phase-locked loops, synthesizers, and so forth.
- AM noise is important in generators for residual phase-noise testing or adjacent-channel receiver testing.
- Any active or non-linear device produces some level of AM to ΦM noise conversion. This AM can contribute to the residual phase noise. This includes AM noise in phase detectors.
- When troubleshooting unsatisfactory phase noise performance, it may be necessary to measure both the AM noise and the residual ΦM noise of the system components to locate the problem.

HP 3048A Description

The HP 3048A provides you standard process for measuring phase noise. It allows you to measure sources of many types with a flexible system configuration.

The HP 3048A Phase Noise Measurement System includes the following instruments and accessories.

- The HP 11848A Phase Noise Interface, an interface box specifically designed for high performance phase noise measurements. The HP 11848A supports several measurement techniques for phase noise and AM noise measurement. Built into the interface are phase detectors, amplifiers, filters, and switches necessary to measure phase noise over a frequency range of 5 MHz to 18 GHz. An input for an external phase detector outside the above mentioned frequency range is also provided. Internal sources are provided to allow the system to functionally check all of its signal handling circuits ensuring proper operation prior to making a measurement.
- The HP 3561A Dynamic Signal Analyzer, a Fast Fourier Transform analyzer of a wide frequency range (125 μHz to 100 kHz). The HP 3561A has built-in data averaging capabilities, large dynamic range, and fast measurement speed which make it ideal for quantifying demodulated phase noise (noise voltages).
- Measurement software, a program that includes all drivers necessary to run both standard and optional instruments of the HP 3048A system.
- Operator's Training, a training course that explains all of the operating modes and measurement techniques of the HP 3048A, when each technique is appropriate, and how to analyze the measured data.

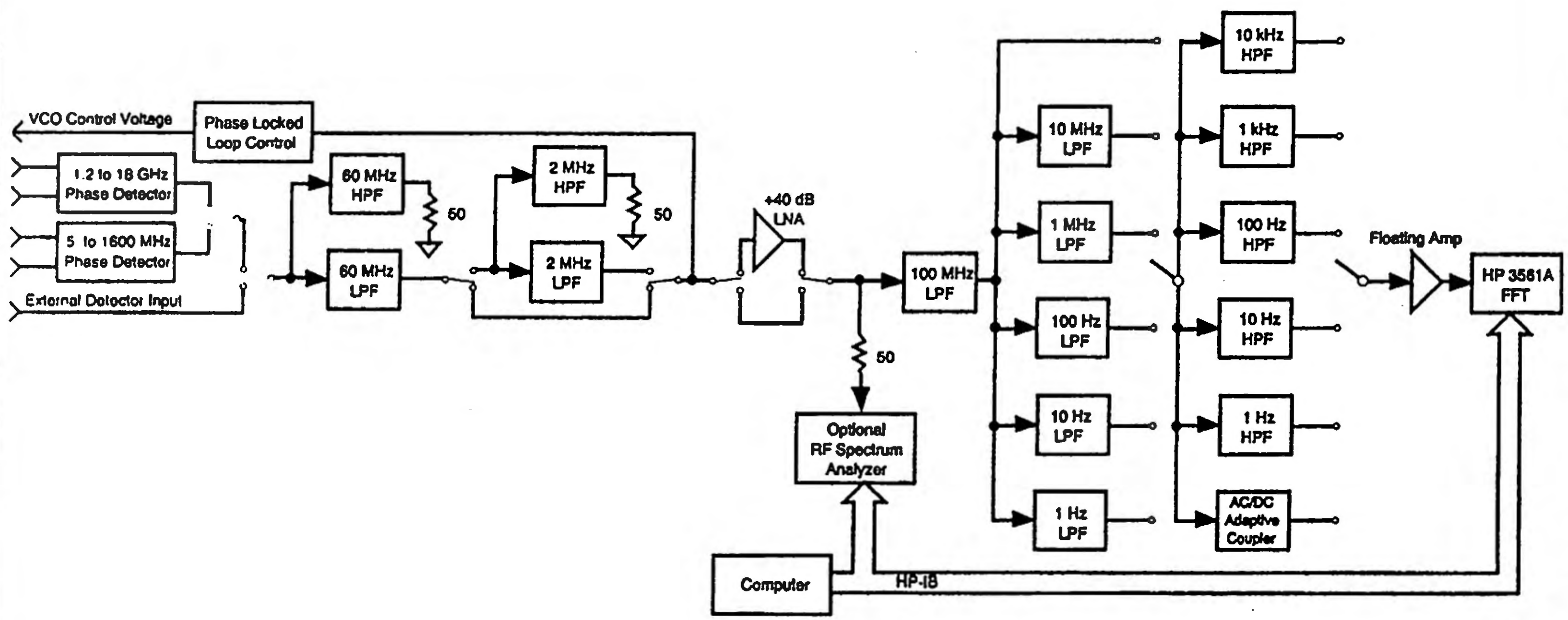


Figure 1-3. HP 3048A Block Diagram

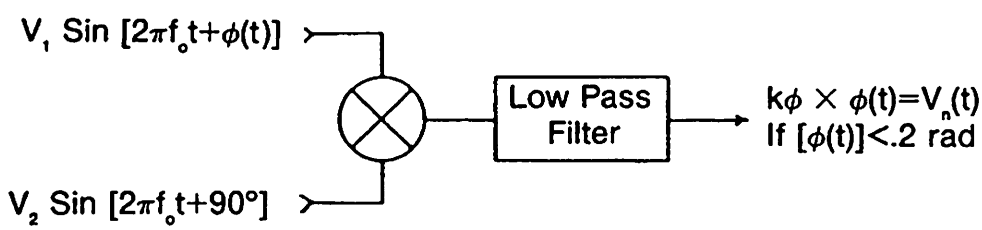
2

Residual Phase Noise Measurement

Basic Phase Noise Measurement Theory

Phase noise can be measured by demodulating the RF signal and analyzing it at baseband.

Doubly balanced mixer used as phase detector



$$S_\phi = \frac{S_n(f)}{K^2\phi}$$

$$\mathcal{L}(f) = \frac{S_\phi(f)}{2} = \frac{S_n(f)}{2K^2\phi} \quad \text{because} \quad |\phi(t)| < .2\text{rad}$$

1. Demodulate RF signal and analyze at baseband
2. Measure spectral density of noise fluctuation
3. Calculate phase noise

Figure 2-1. Double-Balanced Mixer Used as a Phase Detector

A double-balanced mixer is used as a phase detector to demodulate the RF signal for baseband analysis.

When operated as a phase detector, two signals are input to the double-balanced mixer at the same frequency. The phase difference between the signals is adjusted to 90° (quadrature) to minimize the detector's sensitivity to AM fluctuations, and to maximize its sensitivity to phase fluctuations. Any phase fluctuations not common to both signals (for example, $\phi(t)$) result in a voltage fluctuation proportional to the phase difference, provided the phase fluctuations are less than approximately 0.2 radians. This voltage output $\nu_n(t)$ is equal to the difference in phase fluctuations multiplied by the phase detector gain of the mixer, K_ϕ, in volts per radian. The spectral density of the phase fluctuations, $S_\phi(f)$, is calculated by measuring the spectral density of voltage fluctuations, $S_n(f)$ with a baseband spectrum analyzer. $S_n(f)$ is then divided by the square of the phase detector constant (squared because of the power relationship of spectral density) which results in $S_\phi(f)$.

The single-sided phase noise $\mathcal{L}(f)$, can then be calculated from the spectral density of phase fluctuation $S_\phi(f)$, (or frequency fluctuation, $S_\nu(f) = f^2 \times S_\phi(f)$) provided that the mean square phase fluctuations, $\phi^2(t)$, are small relative to 1 radian.

Region of Validity of $\mathcal{L}(f) = S_\phi(f)/2$

Because of the small-angle criterion, caution must be exercised when $\mathcal{L}(f)$ is calculated from the spectral density of the phase fluctuations. This plot (figure 2-2) of $\mathcal{L}(f)$ resulting from the phase noise of a free-running VCO illustrates the erroneous results that can occur if the instantaneous phase modulation exceeds a small angle. Approaching the carrier, $\mathcal{L}(f)$ is obviously increasingly in error as it reaches a relative level of +45 dBc/Hz at a 1 Hz offset (45 dB more power at a 1 Hz offset, in a 1 Hz bandwidth, than the total power in the signal). The −10 dB/decade line is drawn on the plot for an instantaneous phase deviation of 0.2 radians integrated over one decade of offset frequency. At approximately 0.2 radians the power in the higher-order sidebands of the phase modulation is still insignificant compared to the power in the first-order sideband, thus ensuring the validity of the calculation of $\mathcal{L}(f)$. Below the line, the plot of $\mathcal{L}(f)$ is correct; above the line, $\mathcal{L}(f)$ becomes increasingly invalid and $S_\phi(f)$ must be used to represent the phase noise of the signal.

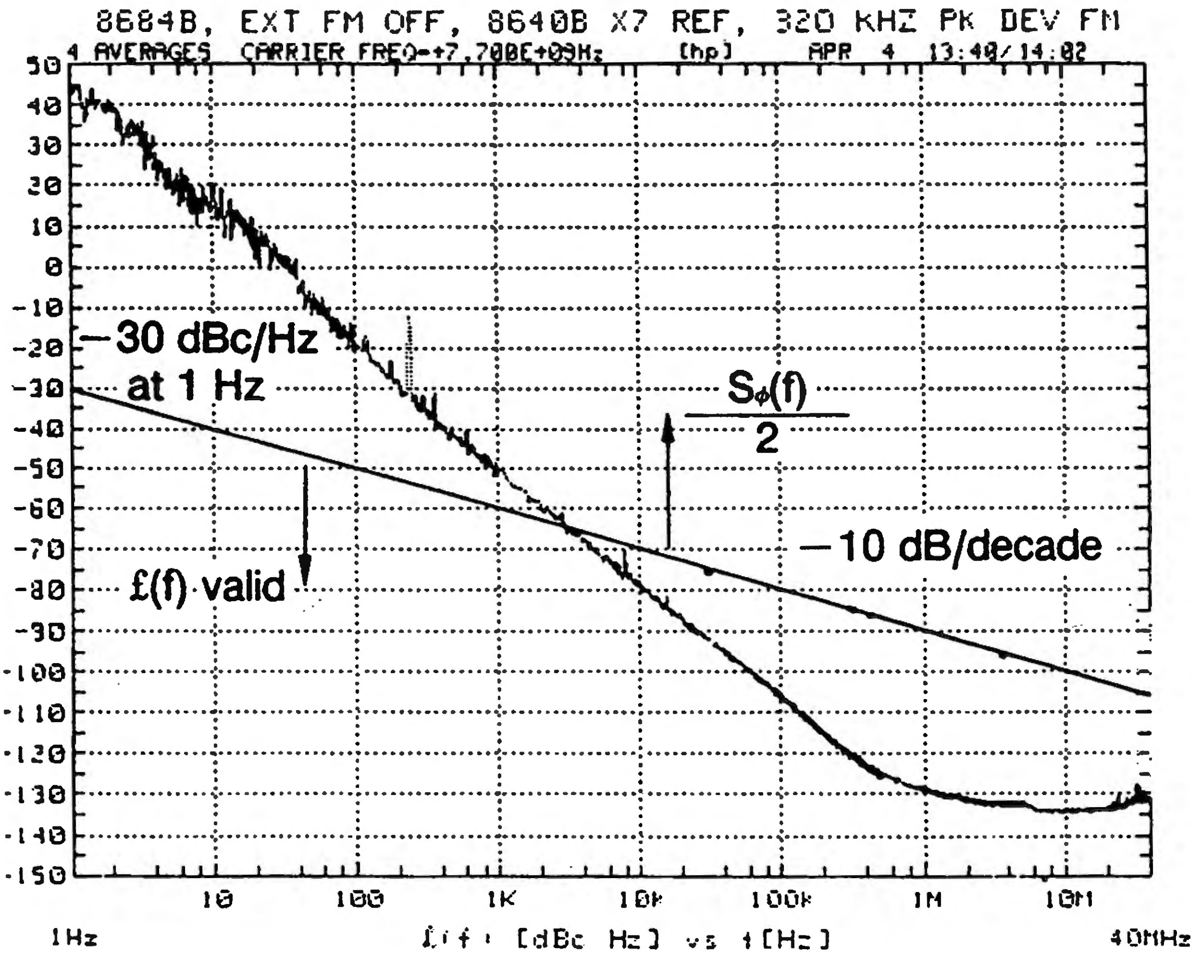


Figure 2-2. Region of Validity of $\mathcal{L}(f) = S_\phi(f)/2$

Conversion Between $S_\phi(f)$ and $S_\nu(f)$

Other than $S_\phi(f)$, the instability of a signal may also be represented with a plot of the spectral density of frequency fluctuations, $S_\nu(f)$. As illustrated below $S_\nu(f)$ is equal to $f^2 \times S_\phi(f)$ because $\nu(t)$ is the derivative of $\phi(t)$. These two graphs are from the same data with figure 2-3 a square root of $S_\nu(f)$. The graph of the square root of $S_\nu(f)$ indicates the power spectral density of the frequency modulation (FM) noise on the signal. Such a measurement of the spectral density of the FM noise versus the offset from the carrier is be very useful in the design of an FM system.

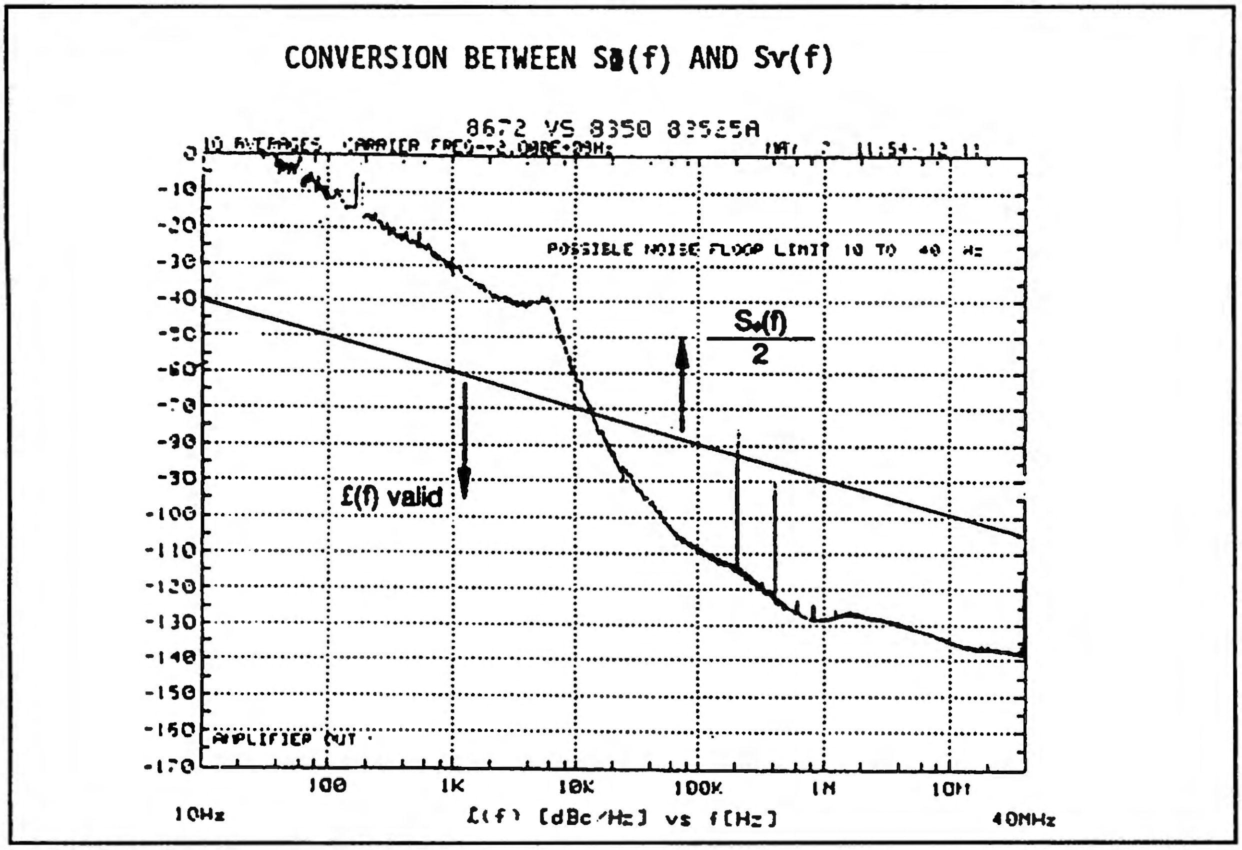


Figure 2-3. $S_\phi(f)/2$: 1/2 Spectral Density of Phase Fluctuation

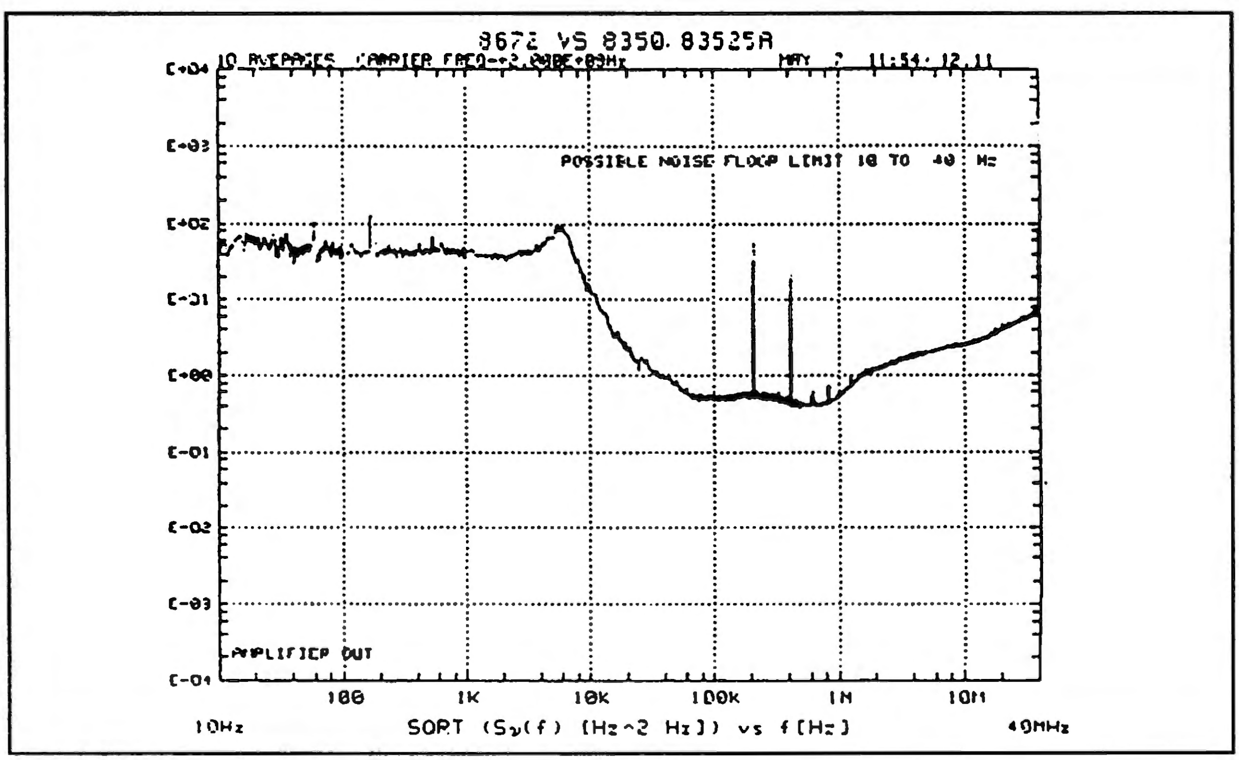


Figure 2-4. $S_\nu(f)$: Spectral Density of Phase Fluctuation

Residual Phase Noise Measurement: Basic Assumptions

- The source noise in each of the two phase-detector paths is correlated at the phase detector for the frequency-offset range of interest. (This assumption will be examined more closely.)
- Correlated phase noise at the phase detector will cancel.
- Source AM noise is small. A typical mixer-type phase detector only has about 20 to 30 dB of AM noise rejection.

Given these assumptions, if a device-under-test (DUT) is placed ahead of either of the two inputs of the phase detector, then all of the source noise will cancel and only the residual noise of the DUT will be measured.

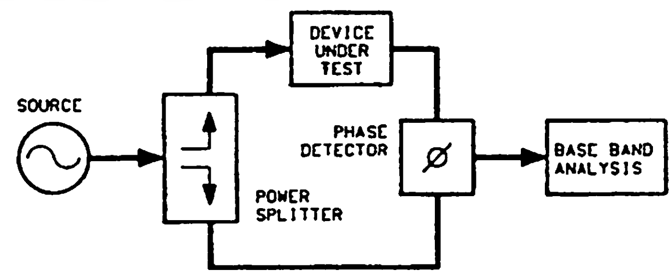


Figure 2-5.

If the DUT is a frequency translating device, then one DUT must be put in each path. The result will be the sum of the noise from each DUT. For most applications, if the DUTs are identical, it can be assumed that the noise of each is half the measured result or 3 dB less. All that can be concluded is that one of the DUTs is at least 3 dB better than the measured result. If a more precise determination is required, a third DUT must be measured against the other two DUTs. The data from each of the three experiments can then be processed by the system to give the noise of each of the individual DUTs.

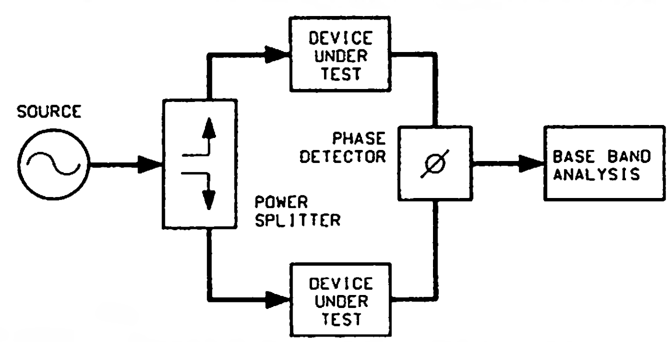


Figure 2-6.

Steps for Making Residual Phase Noise Measurements

- Connect the system hardware and load/run the software.
- Measure the system calibration data. The system calibration data is the correction data for all the signal paths in the interface box.
- Main Menu—Select the type of measurement to be made. All residual phase noise measurements will be "PHASE NOISE MEASUREMENT WITHOUT VOLTAGE CONTROL."
- Establish parameters.
  1. Source parameters
     a. Enter phase detector input frequency
     b. Enter carrier frequency
     c. Select internal or external phase detector (mixer)
     d. Select calibration option
  2. Measurement parameters
     a. Enter start and stop frequencies for the measurement data
        - HP 3048A without RF analyzer: 0.01 Hz to 100 kHz
        - HP 3048A with RF analyzer: 0.01 Hz to 40 MHz
     b. Enter number of sweeps averaged on FFT analyzer
  3. Plot parameters
     a. Select graph type (usually "SINGLE SIDEBAND PHASE NOISE")
     b. Select plotter type (if any)
     c. Enter minimum and maximum Y-axis values (dBc)
     d. Enter minimum and maximum X-axis values (Hz)
     e. Enter a title
- Make measurement.
  1. Connect the DUT and external hardware
  2. Measure calibration data by selected option
  3. Measure noise data
- Interpret the measurement results.

Choosing a Calibration Method

Method 1: User entry of phase detector constant

This calibration option requires that you know the phase detector constant for the specific measurement to be made. The phase detector constant can be estimated from the source power levels or it can be determined using one of the other calibration methods.

Once determined, the phase detector constant can be entered directly into the system software without going through a calibration sequence. Remember, however, that the phase detector constant is unique to a particular set of sources, the RF level into the phase detector, and the test configuration.

Advantages	Easy method of calibrating the measurement system.
	Requires little additional equipment, only an RF power meter to manually measure the drive levels into the phase detector.
	Fastest method of calibration. If the same power levels are always at the phase detector (as in the case of leveled outputs) the phase detector sensitivity will always be essentially the same (within 1 or 2 dB). If this accuracy is adequate, it is not necessary to recalibrate.
	Only one RF source is required.
	Quick method of estimating the phase detector constant and noise floor to verify other calibration methods and check available dynamic range.
Disadvantages	Least accurate of the calibration methods.
	Does not take into account the amount of power at harmonics of the signal.
	Does not take into account the power which may be generated by spurious oscillations, causing the power meter to measure more power than is at the distinct phase-detector frequency.

Method 2: Measured +/– dc peak voltage

This technique requires you to adjust off of quadrature to both the positive and the negative peak output of the Phase Detector. This is done by either adjusting the phase shifter or the frequency of the source. An oscilloscope or voltmeter is optional for setting the positive and negative peaks.

Advantages

Easy method of calibrating the measurement system.

This calibration technique can be performed using the HP 3561A.

Fastest method of calibration. If, for example, the same power levels are always at the phase detector, as in the case of leveled or limited outputs, the phase detector sensitivity will always be essentially equivalent (within 1 or 2 dB). Recalibration becomes unnecessary if this accuracy is adequate.

Only one RF source is required.

Measures the phase detector gain in the actual measurement configuration.

Disadvantages

Has only moderate accuracy compared to the other calibration methods.

Does not take into account the amount of phase detector harmonic distortion relative to the measured phase detector gain, therefore, the phase detector must operate in its linear region.

Requires manual adjustments to the source and/or phase shifter to find the phase detector's positive and negative output peaks. The system will read the value of the positive and negative peak and automatically calculate the mean of the peak voltages which is the the phase detector constant used by the system.

Method 3: Measured Beatnote

This calibration option requires that one of the input frequency sources be tunable such that a beatnote can be acquired for the two sources. For the system to calibrate, the beatnote frequency must be within the ranges shown in the table below. (You should also note that for beatnote frequencies below 20 Hz, it will take the system longer to determine the calibration constant.)

Carrier Frequency	Beatnote Frequency Range (f_B)
$<$ 95 MHz	1 Hz $< f_B <$ 1 MHz
$>$ 95 MHz	1 Hz $< f_B <$ 20 MHz

Advantages

Does not require an RF spectrum analyzer.

Simple method of calibration.

Disadvantages

It does not take into account the harmonics of the phase detector and all non-linearities thereof when using the HP 3048A.

It requires two RF sources separated by 1 Hz to 40 MHz at the phase detector. The calibration source output power must be manually adjusted to the same level as the power splitter output it replaces (requires a power meter).

It is less accurate than either the phase modulation method or the single sided spur method.

Method 4: Double-Sided Spur

This calibration option has the following requirements:

- One of the input frequency sources must be capable of being frequency or phase modulated.
- The resultant sideband spurs from the FM or ΦM modulation must have amplitudes that are >−100 dB and <−20 dB relative to the carrier amplitude.
- The offset frequency or modulation frequency must be between 20 Hz and 100 kHz if only the HP 3561A analyzer is configured in the system, or between 20 Hz and 20 MHz if an RF spectrum analyzer is also configured in the system.

Advantages — Requires only one RF source.

Calibration is done under actual measurement conditions so all non-linearities are calibrated out. Because the calibration is performed under actual measurement conditions, the Double-sided Spur Method and the Single-sided Spur Method are the two most accurate calibration methods.

Disadvantages — Requires an RF spectrum analyzer for manual measurement of ΦM sidebands.

Requires audio calibration source.

Requires a phase modulator which operates at the desired carrier frequency. (Most phase modulators are narrow-band devices, therefore a wide range of test frequencies will require multiple phase modulators.)

Method 5: Single-Sided Spur

This calibration option has the following requirements:

- A third source to generate a single-sided spur.
- An external power combiner (or adder) to add the calibration spur to the frequency carrier under test. The calibration spur must have an amplitude >−100 dB and <−20 dB relative to the carrier amplitude. The offset frequency of the spur must be >20 Hz and <20 MHz.
- A spectrum analyzer or other means to measure the single-sided spur relative to the carrier signal.

You will find that the equipment setup for this calibration option is similar to the others except that an additional source and a power splitter have been added so that the spur can be summed with the input carrier frequency.

Advantages	Calibration is done under actual measurement conditions so all non-linearities and harmonics of the phase detector are calibrated out.
	The Double-sided Spur Method and the Single-Sided Spur Method are the two most accurate methods.
	Broadband couplers with good directivity are available, at reasonable cost, to couple-in the calibration spur.
Disadvantages	Requires two RF sources that must be between 20 Hz and 100 kHz if only an HP 3561A analyzer is configured in the system, or between 20 Hz and 20 MHz if an RF spectrum analyzer is configured in the system.
	Requires an RF spectrum analyzer for manual measurement of the signal-to-spur ratio and the spur offset frequency.

Calibration and Measurement General Guidelines

The following general guidelines should be considered when setting up and making a residual two-port phase noise measurement.

1. **For residual phase noise measurements, the source noise must be correlated.**
   a. The phase delay in the paths between the power splitter and the phase detector must be kept to a minimum when making residual noise measurements. In other words, by keeping the cables between the phase detector and power splitter short, τ will be small. The attenuation of the source noise is a function of the carrier offset frequency, and the delay time (τ) and is equal to:

$$\text{Attenuation (dB)} = 20\log|2\sin(\pi \times \text{f} \times \tau)|$$

$$\begin{aligned} \text{where: } \text{f} &= \text{carrier offset frequency} \\ \pi &= 3.14159 \\ \tau &= \text{time delay (sec.)} \end{aligned}$$

Note

For $f = 1/(2\pi\tau)$ the attenuation of the source is 0 dB.
For $f < 1/(2\pi\tau)$ the source noise is attenuated at the rate of 20 dB per decade.
For $\frac{1}{(2\tau)}$ there is 6 dB gain.
See appendix A.

b. The source should also have a good broadband phase noise floor because at sufficiently large carrier offsets it will tend to decorrelate when measuring components with large delays. A source with a sufficiently low noise floor may be able to hold an otherwise impossible measurement within the region of validity. Examples of sources which best meet these requirements are the HP 8640B and HP 8642A/B.

2. **The source used for making residual phase noise measurements must be low in AM noise because:**

   a. Source AM noise can cause AM to ΦM conversion in the DUT.

   b. Mixer-type phase detectors only provide about 20 to 30 dB of AM noise rejection in a ΦM noise measurement.

3. **It is very important that all components in the test setup be well shielded from RFI.** Unwanted RF coupling between components will make a measurement setup very vulnerable to external electric fields around it. The result may well be a setup going out of quadrature simply by people moving around in the test setup area and altering surrounding electric fields. A loss of quadrature stops the measurement.

4. **When making low-level measurements, the best results will be obtained from uncluttered setups.** Soft foam rubber is very useful for isolating the DUT and other phase-sensitive components from mechanically-induced phase noise. The mechanical shock of bumping the test set or kicking the table will often knock a sensitive residual phase noise measurement out of quadrature.

5. **When making an extremely sensitive measurement it is essential to use semi-rigid cable between the components.** The bending of a flexible cable from vibrations and temperature variations in the room can cause enough phase noise in flexible connecting cables to destroy the accuracy of a sensitive measurement. The connectors also must be tight; a wrench is the best tool.

6. **When measuring a low-noise device, it is important that the source and any amplification, required to achieve the proper power at the phase detector, be placed before the splitter so it will be correlated out of the measurement.** In cases where this is not possible; remember that any noise source, such as an amplifier, placed after the splitter in either phase detector path, will contribute to the measured noise.

7. **An amplifier must be used in cases where the signal level out of the DUT is too small to drive the phase detector, or the drive level is inadequate to provide a low enough system noise floor.** In this case the amplifier should have the following characteristics:

    a. It should have the lowest possible noise figure, and the greatest possible dynamic range.

    b. The signal level must be kept as high as possible at all points in the setup to minimize degradation from the thermal noise floor.

    c. It should have only enough gain to provide the required signal levels. Excess gain leads to amplifiers operating in gain compression, making them very vulnerable to multiplicative noise problems. The non-linearity of the active device produces mixing which multiplies the baseband noise of the active device and power supply noise around the carrier.

    d. The amplifier's sensitivity to power supply noise and the power supply noise itself must both be minimized.

Calibration and Measurement Procedures

The following procedures use the system noise floor measurement as an example.

Method 1: User entry of phase detector constant

1. Connect circuit as shown in figure 2-7 and tighten all connections.

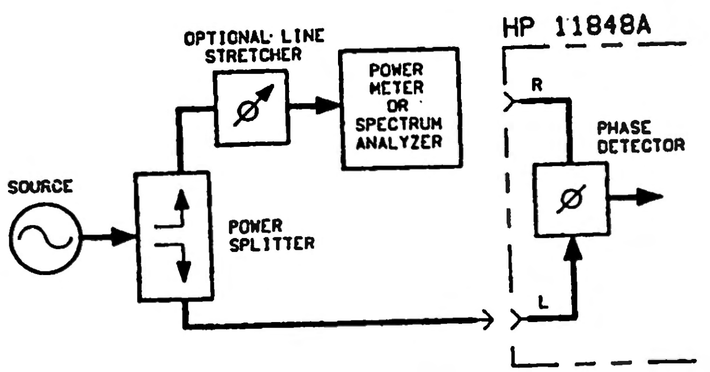


Figure 2-7. Measuring Power at Phase Detector R Port

2. Measure the power level that will be applied to the R port of the HP 11848A's Phase Detector. The following table shows the acceptable amplitude ranges for the HP 11848A Phase Detectors.

Phase Detector			
5 MHz to 1.6 GHz		1.2 GHz to 18 GHz	
L Port	R Port	L Port	R Port
+15 dBm to +23 dBm	0 dBm to +23 dBm	+7 dBm to +10 dBm	0 dBm to +10 dBm

3. Locate the power level you measured on the left side of the Phase Detector Sensitivity Graph (figure 2-8). Now move across the graph at the measured level and find the corresponding Phase Detector constant along the right edge of the graph. This is the value you will enter as the `Current Detector Constant` when you define your measurement. (Note that the approximate measurement noise floor provided by the R port level is shown across the bottom of the graph.)

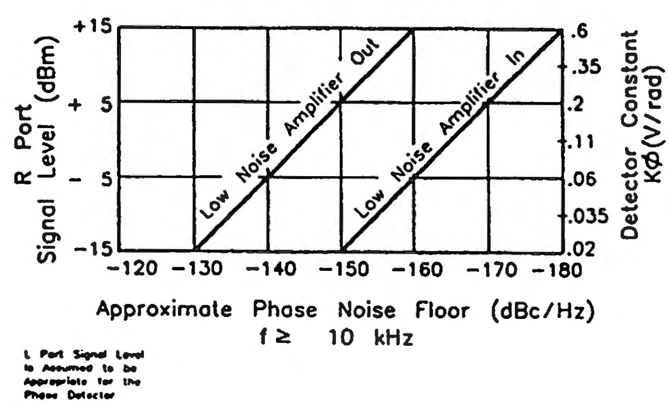


Figure 2-8. Phase Detector Sensitivity

4. If you are not certain that the power level at the L input port is within the range shown in the preceding graph, measure the level using the setup shown in figure 2-9.

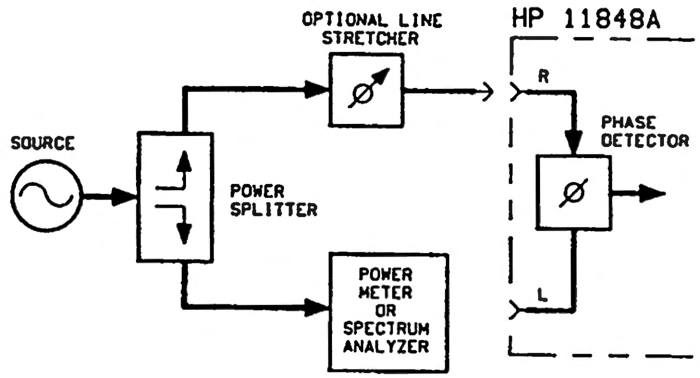


Figure 2-9. Measuring Power at Phase Detector L Port

5. After you complete the measurement set up procedures and begin running the measurement, the HP 3048A will prompt you to adjust for quadrature. Adjust the phase difference at the phase detector to 90 degrees (quadrature) by either adjusting the test frequency or by adjusting an optional variable phase shifter or line stretcher. Quadrature is attained when the meter on the front panel of the phase noise interface is set to center scale, zero.

Note

For the system to accept the adjustment to quadrature, the meter must be within the first small divisions around zero, and for the system to continue to take data it must stay within the second small divisions.

6. Once you have attained quadrature, you are ready to proceed with the measurement.

Method 2: Measured +/- DC Peak Voltage

1. Connect circuit as per figure 2-10, and tighten all connections.

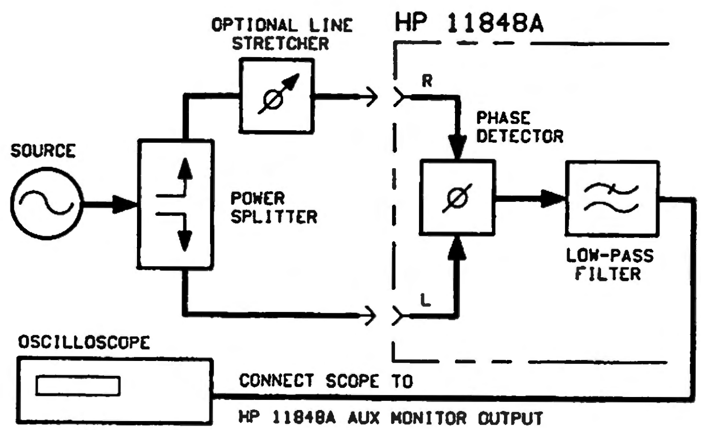


Figure 2-10. Connection to Optional Oscilloscope for Determining Voltage Peaks

2. Measure the power level that will be applied to the R port of the HP 11848A's Phase Detector. The following table shows the acceptable amplitude ranges for the HP 11848A Phase Detectors.

Phase Detector			
5 MHz to 1.6 GHz		1.2 GHz to 18 GHz	
L Port	R Port	L Port	R Port
+15 dBm to +23 dBm	0 dBm to +23 dBm	+7 dBm to +10 dBm	0 dBm to +10 dBm

3. Adjust the phase difference at the phase detector over a 360 degree range.
4. The system will measure the positive and negative peak voltage of the phase detector using the signal displayed on the HP 3561A. For more sensitivity, an oscilloscope or voltmeter can be connected to the AUX MONITOR port on the HP 11848A for determining the peaks. The phase may be adjusted either by varying the frequency of the source or by adjusting a variable phase shifter or line stretcher.

Note Connecting an oscilloscope to the AUX MONITOR port is recommended because the signal can then be viewed to give visual confidence in the signal being measured. As an example, noise could affect a voltmeter reading, whereas, on the oscilloscope any noise can be viewed and the signal corrected to minimize the noise before making the reading.

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5. The system software will then calculate the phase detector constant automatically using the following algorithm.

   $$Phase\ Detector\ Constant = \frac{((+V_{peak})-(-V_{peak}))}{2}$$

6. The system software will then adjust the phase detector to quadrature if the source can be controlled automatically, or will prompt you to set the HP 11848A meter to quadrature if the source is a manual instrument.
7. The system will now measure the noise data.

Method 3: Measured Beatnote Method

1. Connect circuit as per figure 2-11, and tighten all connections.

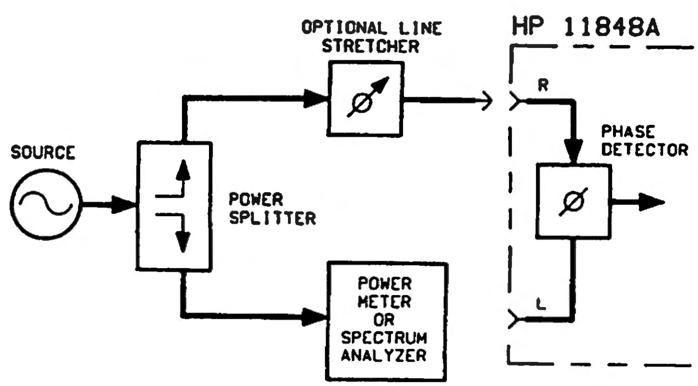


Figure 2-11. Measuring Power from Splitter

2. Measure the power level that will be applied to the R port of the HP 11848A's Phase Detector. The following table shows the acceptable amplitude ranges for the HP 11848A Phase Detectors.

Phase Detector			
5 MHz to 1.6 GHz		**1.2 GHz to 18 GHz**	
L Port	**R Port**	**L Port**	**R Port**
+15 dBm to +23 dBm	0 dBm to +23 dBm	+7 dBm to +10 dBm	0 dBm to +10 dBm

3. Measure the output power at one side of the power splitter, then terminate in 50 ohms.
4. Adjust the calibration source to the same output power as the measured output power of the power splitter.
5. Adjust the output frequency such that the beatnote frequency is between 1 Hz and 100 kHz, or to between 1 Hz and 20 MHz if an RF spectrum analyzer is included in the system. (Note that the beatnote frequency may be measured on the system spectrum analyzers.)
6. The system can now measure the calibration constant.
7. Disconnect the calibration source and reconnect the power splitter.
8. Adjust the phase difference at the phase detector to 90 degrees (quadrature) either by adjusting the test frequency or by adjusting an optional variable phase shifter or line stretcher. Quadrature is achieved when the meter on the front panel of the phase noise interface is set to zero.

Note

For the system to accept the adjustment, the meter needle must be between the first two small divisions around center scale (zero). For the system to continue to take data, the needle must stay within the second two small divisions around center scale.

9. Reset quadrature and measure phase noise data.

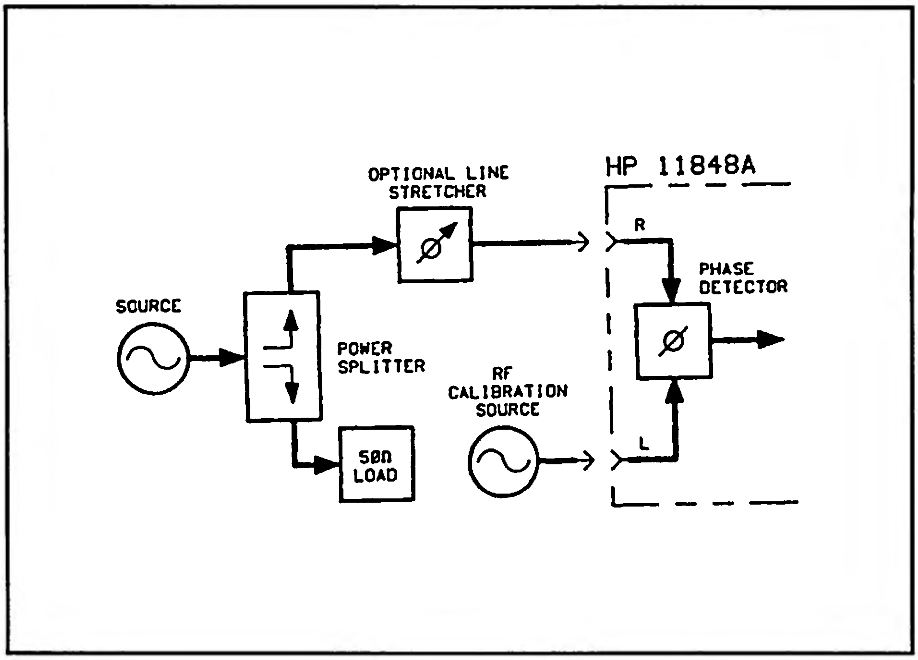


Figure 2-12. Calibration Source Beatnote Injection

Method 4: Double-sided Spur

1. Connect circuit as per figure 2-13, and tighten all connections.

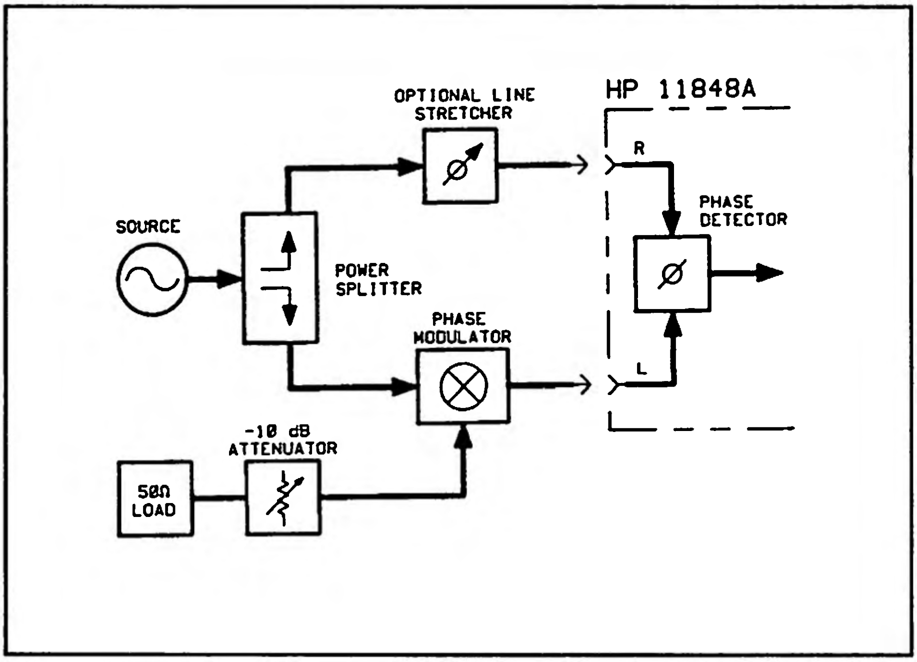


Figure 2-13. Double-Sided Spur Calibration Setup

2. Measure the power level that will be applied to the R port of the HP 11848A's Phase Detector. The following table shows the acceptable amplitude ranges for the HP 11848A Phase Detectors.

Phase Detector			
5 MHz to 1.6 GHz		**1.2 GHz to 18 GHz**	
L Port	**R Port**	**L Port**	**R Port**
+15 dBm to +23 dBm	0 dBm to +23 dBm	+7 dBm to +10 dBm	0 dBm to +10 dBm

3. Using the RF spectrum analyzer, measure the carrier-to-sideband ratio of the phase modulation at the phase detector's modulated port and the modulation frequency. The audio calibration source should be adjusted such that the sidebands are between −30 and −60 dB below the carrier and the audio frequency is between 20 Hz and 100 kHz (or between 20 Hz and 20 MHz if an RF spectrum analyzer is included in the system).

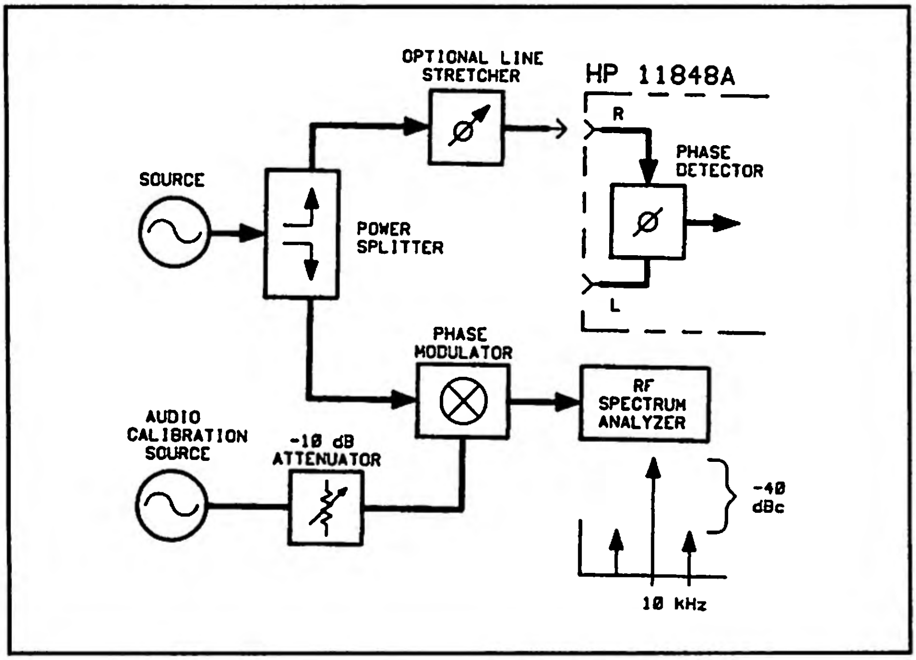


Figure 2-14. Measuring Carrier-to-sideband Ratio of the Modulated Port

4. Measure the carrier-to-sideband ratio of the non-modulated side of the phase detector. It must be at least 20 dB less than the modulation level of the modulated port. This level is necessary to prevent cancellation of the modulation in the phase detector. Cancellation would result in a smaller phase detector constant, or a measured noise level that is worse than the actual performance. The modulation level is set by the port-to-port isolation of the power splitter and the isolation of the phase modulator. This isolation can be improved at the expense of signal level by adding an attenuator between the phase modulator and the power splitter.
5. Connect the phase detector.

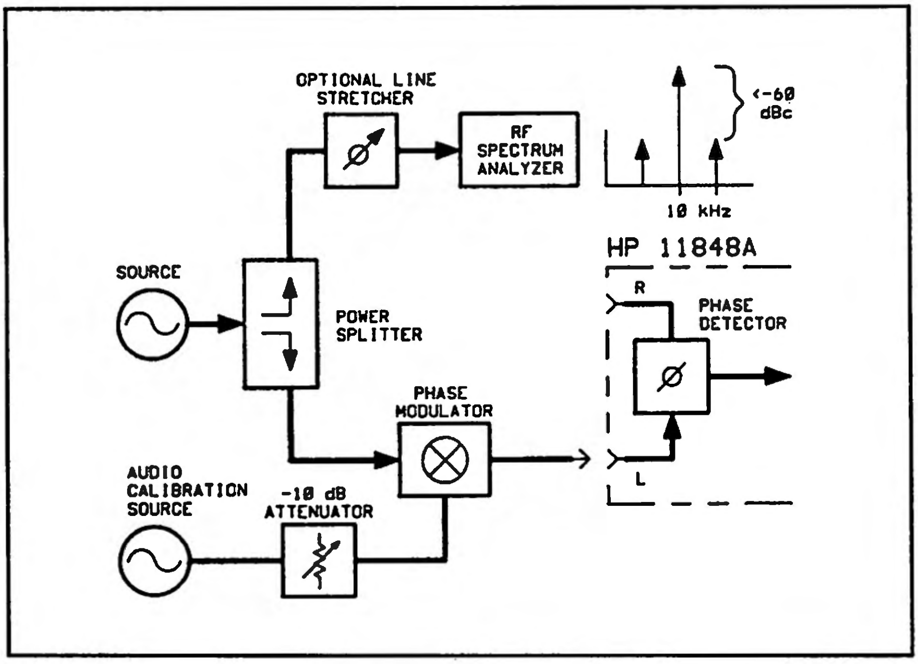


Figure 2-15.
Measuring Carrier-to-sideband Ratio of the Non-modulated Port

6. Adjust the phase difference at the phase detector to 90 degrees (quadrature) either by adjusting the test frequency or by adjusting an optional variable phase shifter or line stretcher. Quadrature is achieved when the meter on the front panel of the HP 11848A is set to center scale.

Note

For the system to accept the adjustment, the meter needle must be between the first two small divisions around center scale (zero). For the system to continue to take data, the needle must stay within the second two small divisions around center scale.

7. At the Connect Diagram access the Calibration Process display by pressing the `Calib Process` softkey.
8. Enter the sideband amplitude and offset frequency.
9. Press `Done` to return to the Connect Diagram.
10. Check quadrature and measure the phase detector constant by pressing `Proceed`.
11. Remove audio source.
12. Reset quadrature and measure phase noise data.

Method 5: Single-Sided Spur Method

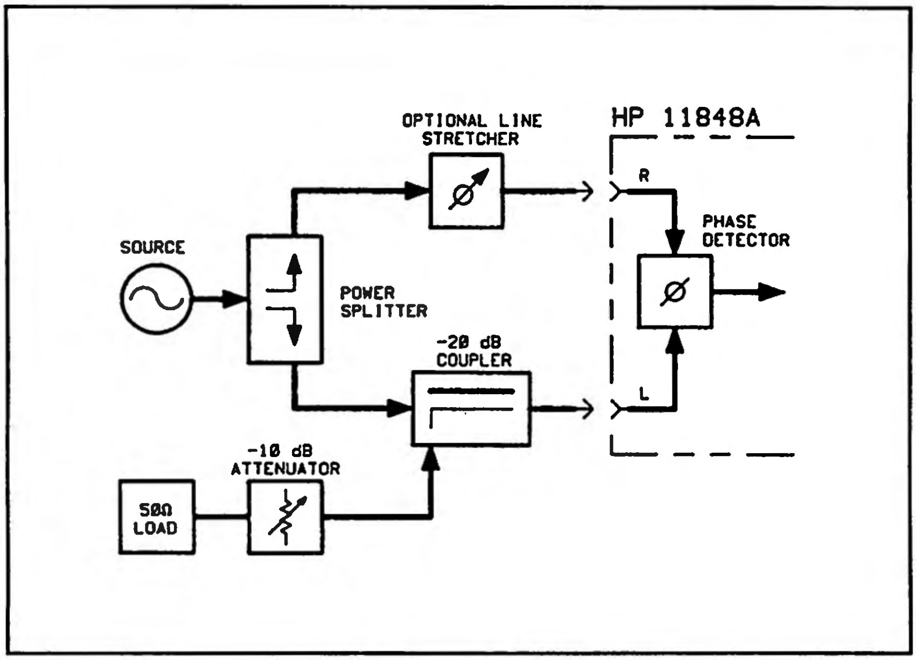


Figure 2-16. Single-Sided Spur Calibration Setup

1. Connect circuit as shown in figure 2-16 and tighten all connections.
2. Measure the power level that will be applied to the R port of the HP 11848A's Phase Detector. The following table shows the acceptable amplitude ranges for the HP 11848A Phase Detectors.

Phase Detector			
5 MHz to 1.6 GHz		1.2 GHz to 18 GHz	
L Port	R Port	L Port	R Port
+15 dBm to +23 dBm	0 dBm to +23 dBm	+7 dBm to +10 dBm	0 dBm to +10 dBm

3. Measure the carrier-to-single-sided-spur ratio out of the coupler at the phase detector's modulated port and the offset frequency with the RF spectrum analyzer. The RF calibration source should be adjusted such that the sidebands are between −30 and −60 dB below the carrier and the frequency offset of the spur between 20 Hz and 100 kHz (or between 20 Hz and 20 MHz if an RF spectrum analyzer is connected in the system).

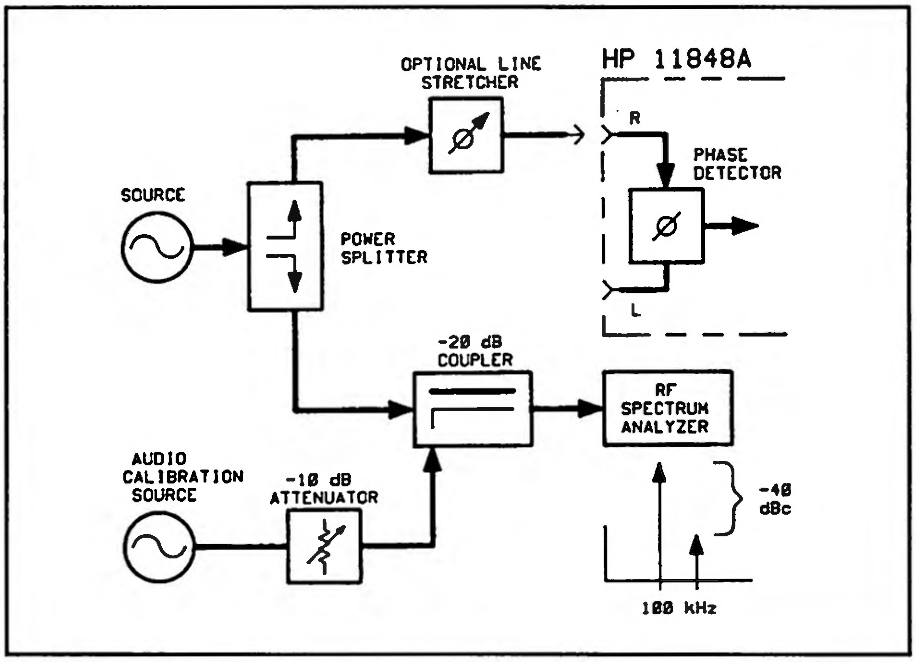


Figure 2-17. Carrier-to-sideband Ratio of the Modulated Signal

4. Measure the carrier-to-spur ratio of the non-modulated side of the phase detector. It must be at least 20 dB less than the spur ratio of the modulated port. This level is necessary to prevent cancellation of the modulation in the phase detector. Cancellation would result in a smaller phase detector constant, or a measured noise level that is worse than the actual performance. The isolation level is set by the port-to-port isolation of the power splitter and the isolation of the −20 dB coupler. This isolation can be improved at the expense of signal level by adding an attenuator between the coupler and the power splitter.

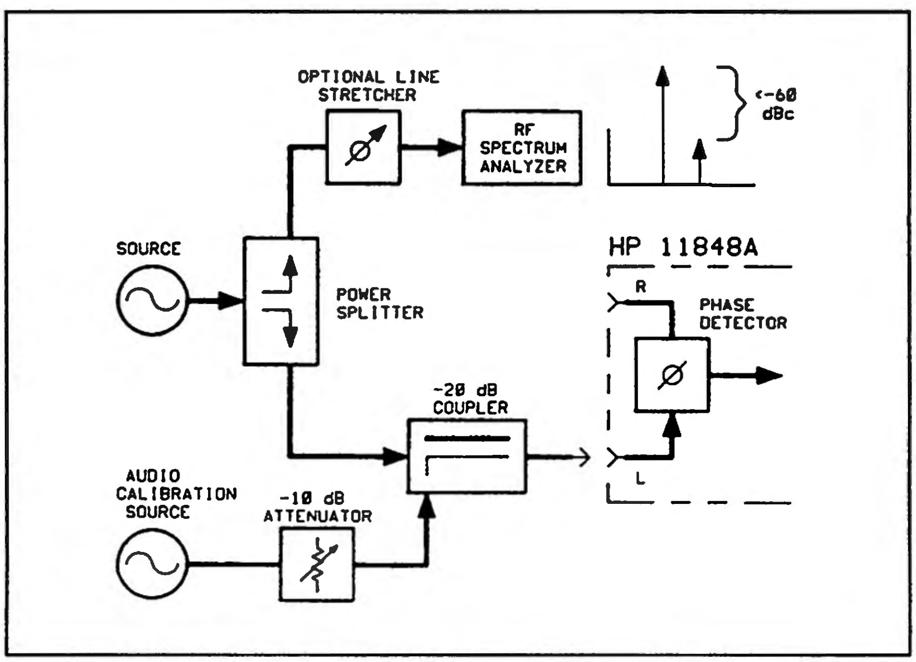


Figure 2-18. Carrier-to-spur Ratio of the Non-modulated Signal

5. Connect the phase detector.
6. Adjust the phase difference at the phase detector to 90 degrees (quadrature) either by adjusting the test frequency or by adjusting an optional variable phase shifter or line stretcher. Quadrature is achieved when the meter on the front panel of the HP 11848A is set to center scale.

Note

For the system to accept the adjustment, the meter needle must be between the first two small divisions around center scale (zero). For the system to continue to take data, the needle must stay within the second two small divisions around center scale.

---

7. Enter sideband level and offset.
8. Check quadrature and measure the phase detector constant.
9. Remove audio source.
10. Reset quadrature and measure phase noise data.

Residual Phase Noise Measurement Examples

System Noise Floor Measurement

The residual noise of the phase detector sets the noise floor performance of the HP 3048A. The system noise floor performance should be measured periodically to ensure measurement integrity.

Initial Setup

This measurement was performed using the User Entry of Phase Detector Constant calibration method.

In this example, the system noise floor is measured using the performance verification test fixture supplied with the system. The test fixture (a power splitter, and a short piece of coax) produces a 90° phase shift at approximately 400 MHz (0.625 ns). The test fixture will be driven by an HP 8640B signal generator, followed by a power amplifier.

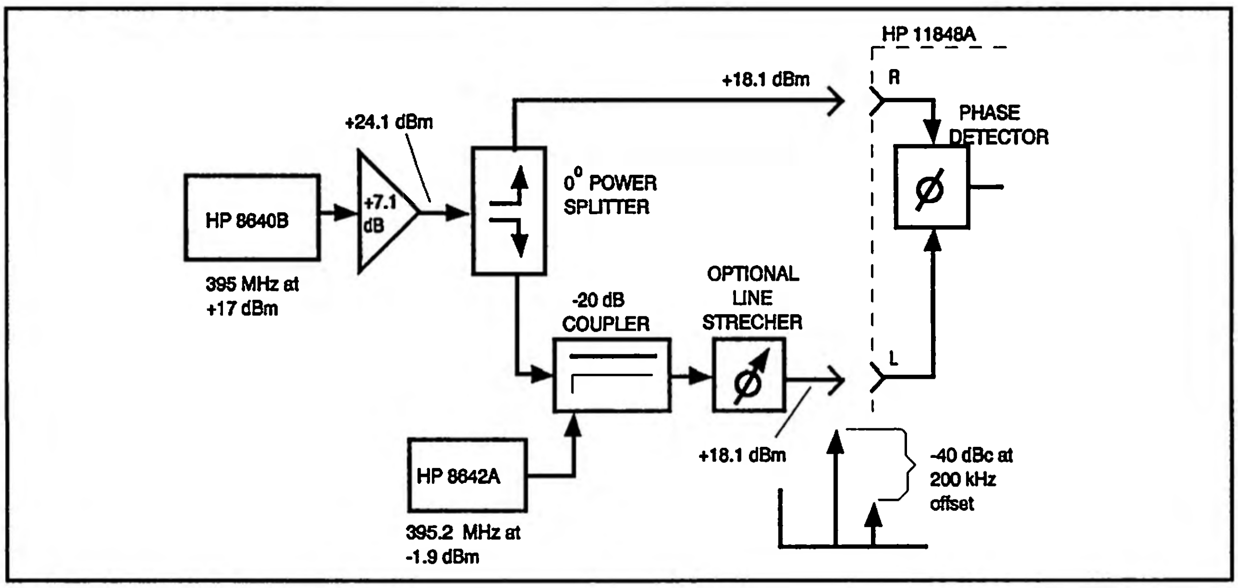


Figure 2-19. User Entry of Phase Detector Constant Calibration Setup

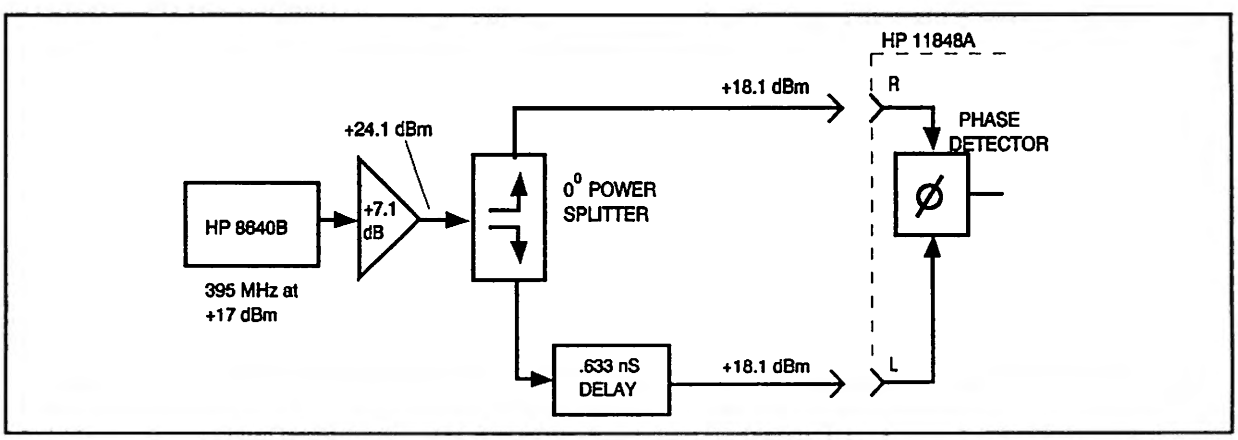


Figure 2-20. User Entry of Phase Detector Constant Measurement Setup

Conditions

This measurement was made under the following conditions.

- All the power required to drive the phase detector is supplied by the source.
- To minimize source noise decorrelation, the time delay is only long enough to produce quadrature at the phase detector.
- The source frequency is adjusted to set quadrature, eliminating the need for any phase-shifting device which might add noise or time delay.
- The source has a very low broadband phase-noise floor, <−160 dBc/Hz for offsets greater than 5 MHz in the frequency band used for this measurement.
- The phase detector constant was carefully measured using a single-sided-spur technique. The measurement was stopped after this measurement and the spur coupler and line stretcher (used for the measuring the phase detector constant) were removed to prevent excess time delay during the system noise floor measurement. The measurement was then restarted and the phase detector constant previously measured was entered into the system.

Results

The results of this measurement example are shown in figure 2-21. The following is an analysis of those results.

- The system noise floor is <−180 dBc/Hz at offsets greater than 20 kHz. The rise in the noise floor beyond 5 MHz is due to the decorrelation of the source noise. The noise specification in the region between 10 kHz and 40 MHz is −170 dBc/Hz.
- The noise between 1 Hz and 5 kHz has a −10 dB/decade slope with a 1 Hz intercept of −145 dBc/Hz. The noise specification in the region between 1 Hz and 10 kHz has a −10 dB/decade slope with a 1 Hz intercept of −130 dBc/Hz.
- The noise between 0.02 Hz and 1 Hz has a −30 dB/decade slope with a 0.02 Hz intercept of −90 dBc/Hz. The noise specification in this region has a slope of −30 dBc/Hz with a 0.02 Hz intercept of −79 dBc/Hz.
- The spurs between 60 Hz and 1 kHz are due to 60 Hz line spurs with all spurs well below the −112 dBc system specification.
- The discontinuity in the noise at 1 kHz is caused by the effective resolution bandwidth of the FFT analyzer being too wide to resolve the 60 Hz spurs in the region beyond 1 kHz.

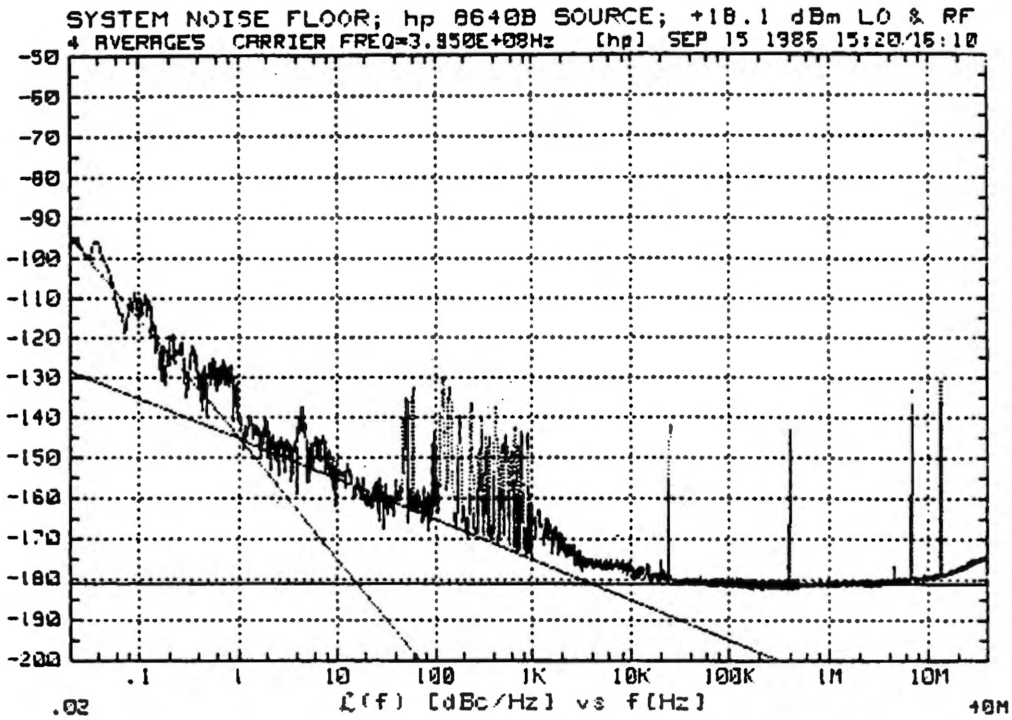


Figure 2-21. System Noise Floor Results Graph

PRESENT SOURCE CHARACTERISTICS
CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±10 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 3.95000E+08 Hz
CARRIER FREQ = 3.95000E+08 Hz
INTERNAL MIXER IS 0, (5 MHz - 1.6 GHz)

PRESENT MEASUREMENT CONSTANTS
VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.752 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz   LOOP BW3 = 0 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1   ATTEN2 = 1

Amplifier Noise and Dynamic Noise Figure Measurement

This measurement was performed using the Single-Sided-Spur calibration method.

The residual noise measurement of an amplifier can reveal two very important pieces of information:

1. The signal-to-noise ratio or dynamic range of the amplifier. The signal-to-noise ratio is a measure of the amount of noise floor during actual operating conditions.
2. The amplifier noise figure, can be calculated from the amplifier input power and $\mathcal{L}(f)$ data, at any measured offset where: $NF(dB) = \mathcal{L}(f) + P_i + 177$. (See appendix B.)

   The amplifier noise is measured under actual large signal conditions. It includes the multiplicative noise produced by the nonlinearity of the active device in the presence of a large signal. The small signal noise figure measured on a noise figure meter may vary greatly from the large signal measurement. As the input power increases, the active device starts to operate nonlinearly and the noise figure increases. This effect may appear with signal levels 10 dB below the 1 dB compression point.

Initial Setup

In this example, the DUT is an HP 8447D preamplifier.

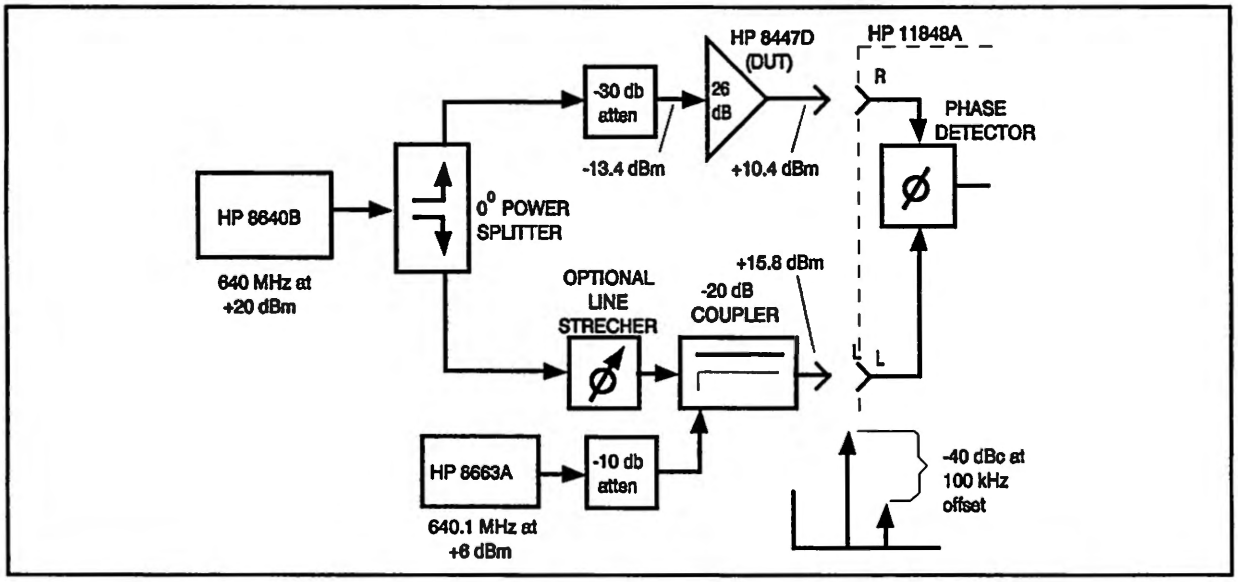


Figure 2-22. Dynamic Noise Figure Measurement Setup

Conditions

This measurment was made under the following conditions.

- All the power required to drive the phase detector comes from the source output. (The source is amplified before the power splitter if more power is needed to drive the phase detector.)
- The power level in the DUT is adjusted with an attenuator to set the desired test condition.
- If the DUT output is inadequate to drive the phase detector, an amplifier may be added to the DUT output. It is necessary to measure the amplifier's noise under this operating condition to ensure it does not limit the measurement.
- An HP phase shifter is used to obtain quadrature.
- The power supply is well filtered to prevent low-frequency noise from entering the DUT and degrading the performance from multiplicative noise.

Results

The results of this measurement are shown in figure 2-23. The following is an analysis of those results.

- The amplifier, measured at a carrier frequency of 640 MHz, appears to be well behaved. There are no major discontinuities in the graph and all the spurs are below −120 dBc. The noise floor is at about −157 dBc/Hz at a 10 kHz offset with a −139 dBc/Hz, 1 Hz intercept.
- At −13.4 dBm input power and 100 kHz offset, the calculated large signal-noise figure is 6.6 dB.
- The noise figure measured by an HP 8970A noise-figure meter was 6.5 dB at 640 MHz. In this case, at this input level, the amplifier is still operating very quietly.

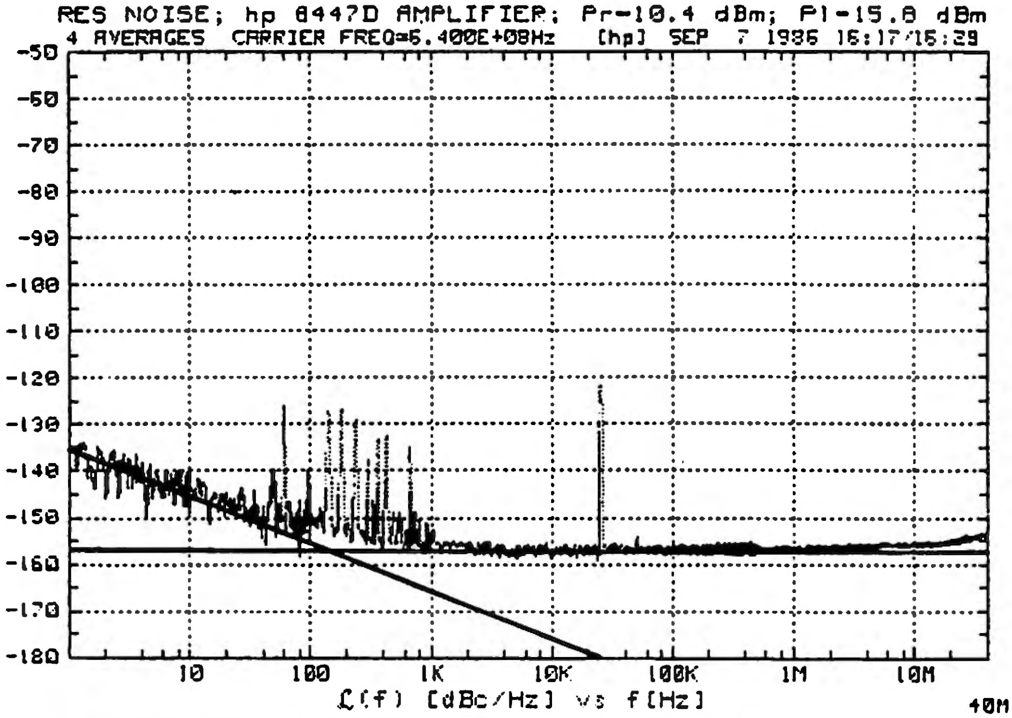


Figure 2-23. Dynamic Noise Figure Results Graph

```
            PRESENT SOURCE CHARACTERISTICS
CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±10 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 6.40000E+08 Hz
CARRIER FREQ = 6.40000E+08 Hz
INTERNAL MIXER IS 0, (5 MHz - 1.6 GHz)

            PRESENT MEASUREMENT CONSTANTS
VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.399 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz    LOOP BW3 = 0 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1      ATTEN2 = 1
```

Measurement of a Device with Large Time Delay (τ)

This measurement was made using the Single-Sided-Spur calibration method.

The measurement of a device with long delay usually has two special considerations:

1. Delays exceeding 1 μs tend to have a large amount of signal path loss. This loss makes it necessary to follow the DUT with an amplifier having the properties discussed in "Calibration and Measurement General Guidelines."
2. The long delay will decorrelate the source noise. The attenuation of the source noise is equal to:

$$\text{Attenuation(dB)} = 20\log|2\sin(\pi f \tau)|$$

$$\text{where } \pi = 3.14159$$

$$f = \text{frequency offset (Hz)}$$

$$\tau = \text{time delay (seconds)}$$

Note

At $\frac{1}{2\pi\tau}$ the source noise will be completely decorrelated and at $\frac{1}{2\tau}$ there is an actual enhancement of 6 dB to the source noise.

The source noise will be periodic in the region beyond $\frac{1}{2\pi\tau}$. The noise peaks are the sum of the source and DUT noise. The bottom of the nulls is residual noise.

Initial Setup

In this example, the DUT is a 236 ns SAW delay line, followed by the amplifier used in the Amplifier Noise and Dynamic Noise Figure measurement.

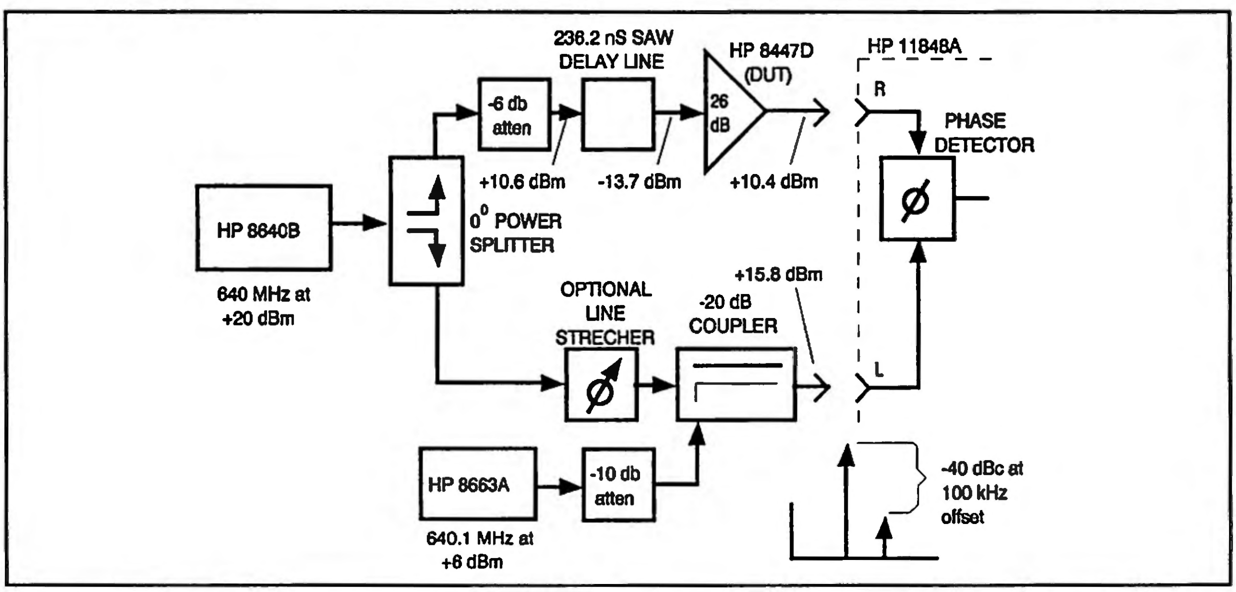


Figure 2-24. Time Delay Measurement Setup

Conditions

This measurement was made under the following conditions.

- All the power required to drive the phase detector comes from the source output.
- The power level into the DUT is adjusted with an attenuator to set the desired test condition.
- The DUT output is inadequate to drive the phase detector, thus an amplifier has been added to the DUT output. It is necessary to measure the amplifier's noise under this operating condition to ensure it does not limit the measurement.
- An HP phase shifter is used to obtain and maintain quadrature. The phase shift through this SAW delay line is very sensitive to temperature change and therefore, it drifts with time. It was necessary to adjust the phase shifter very slowly during the measurement to maintain quadrature. A sudden movement in the phase correction will look like phase noise close to the carrier, and invalidate the close-in data.
- The source decorrelation is plotted for this example to provide an idea of what the noise to be measured should look like.

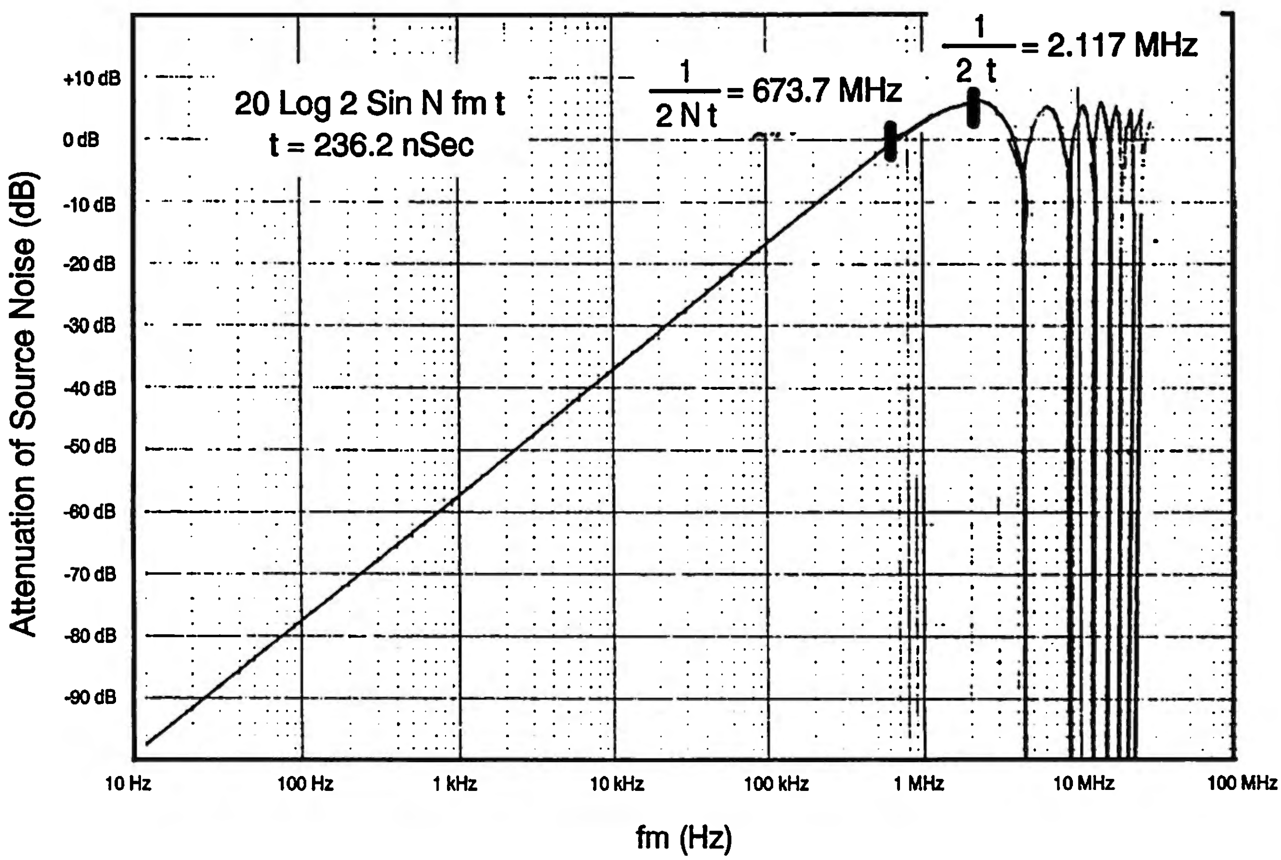


Figure 2-25. Graph of Source Decorrelation

- The attenuation of the DUT must either be known or it must be measured.
- The test-setup noise floor is measured, substituting an attenuator for the DUT. This is necessary to sort the DUT noise from the test-setup noise.

Results

The results of this measurement are found in figure 2-26. The following is an analysis of those results.

- The SAW delay line, measured at the frequency of 640 MHz, appears to be very well behaved. There are no discontinuities in the graph and all the spurs are <10 dB out of the noise. The noise floor is at about −155 dBc/Hz at a 10 kHz offset with a −118 dBc/Hz, 1 Hz intercept.
- The $\frac{1}{f}$ noise region out to about 4 kHz offset is very typical of this DUT.
- The floor region between 4 and 200 kHz approaches the test system noise floor. Data in this region is being degraded by insufficient dynamic range of the test setup. This problem may be remedied either by operating the DUT at a higher output level to increase its output signal-to-noise ratio, or by using an amplifier with a lower noise figure. The test-setup noise floor must be 10 dB below the measured noise to ensure less than a 1 dB measurement error.
- The region beyond 200 kHz is a very good example of the periodic nature of source decorrelation. At $\frac{1}{2\tau}$, noise is almost exactly 6 dB higher than the phase noise of a typical HP 8642A at that offset and carrier frequency. The noise nulls are at the measured test-system noise floor.

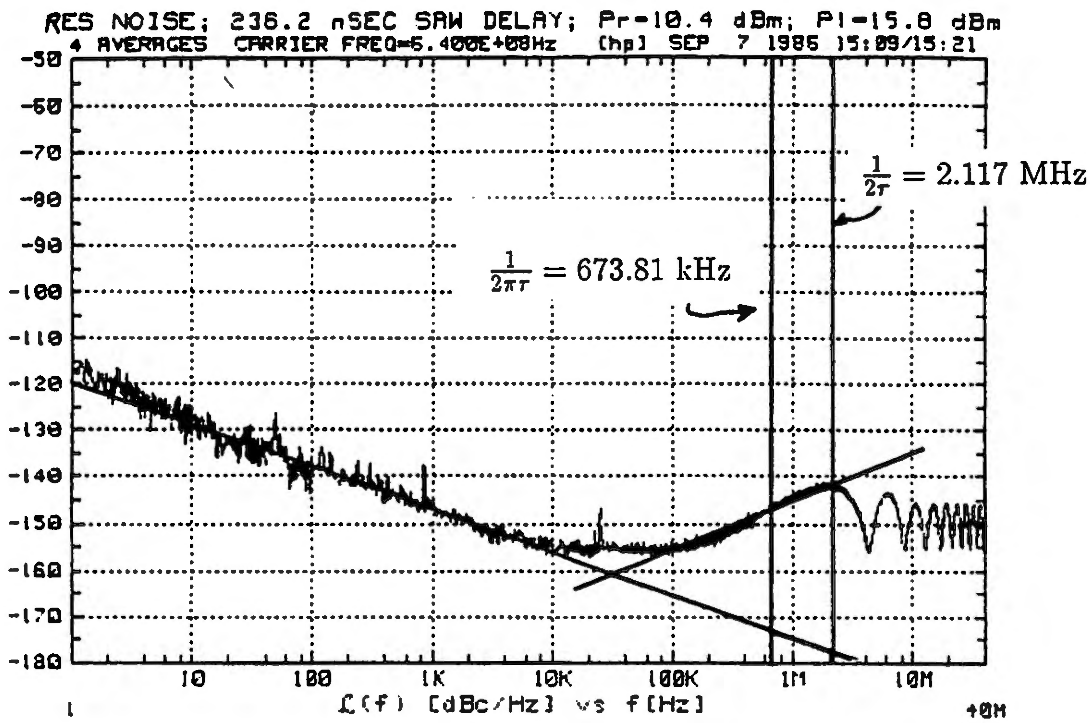


Figure 2-26. Time Delay Results Graph

```
            PRESENT SOURCE CHARACTERISTICS
CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±10 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 6.40000E+08 Hz
CARRIER FREQ = 6.40000E+08 Hz
INTERNAL MIXER IS 0, (5 MHz - 1.6 GHz)

            PRESENT MEASUREMENT CONSTANTS
VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.413 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz     LOOP BW3 = 0 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1       ATTEN2 = 1
```

Measurement of a Crystal Resonator

This measurement is made using the Single-Sided-Spur calibration method.

The measurement of a crystal resonator has four special considerations.

1. Its loss usually makes it necessary to follow the DUT with an amplifier having the properties discussed in "Calibration and Measurement General Guidelines."
2. Crystals are usually high-Q devices and Q is proportional to delay or:

$$\tau = \frac{Q}{\pi f_o}$$

$$\text{Where } \tau = \text{time delay (seconds)}$$

$$Q = \frac{\text{3 dB bandwidth}}{\text{resonant frequency}}$$

$$\pi = 3.14159$$

$$f_o = \text{resonant frequency}$$

   The source must have good close-in phase noise performance, otherwise the source noise, discriminated by the delay of the high-Q resonator, will dominate the measurement.

3. Crystal resonators are very narrow bandwidth devices. The noise measurement must be performed within a few parts-per-million of the resonant frequency. The source must therefore have fine frequency resolution and high stability.
4. Noise data taken at offsets greater than $\frac{\textit{resonator bandwidth}}{2}$ will be attenuated by the resonator itself.

Initial Setup

In this example, the DUT is a 93 MHz SC-cut crystal, followed by a low-noise amplifier.

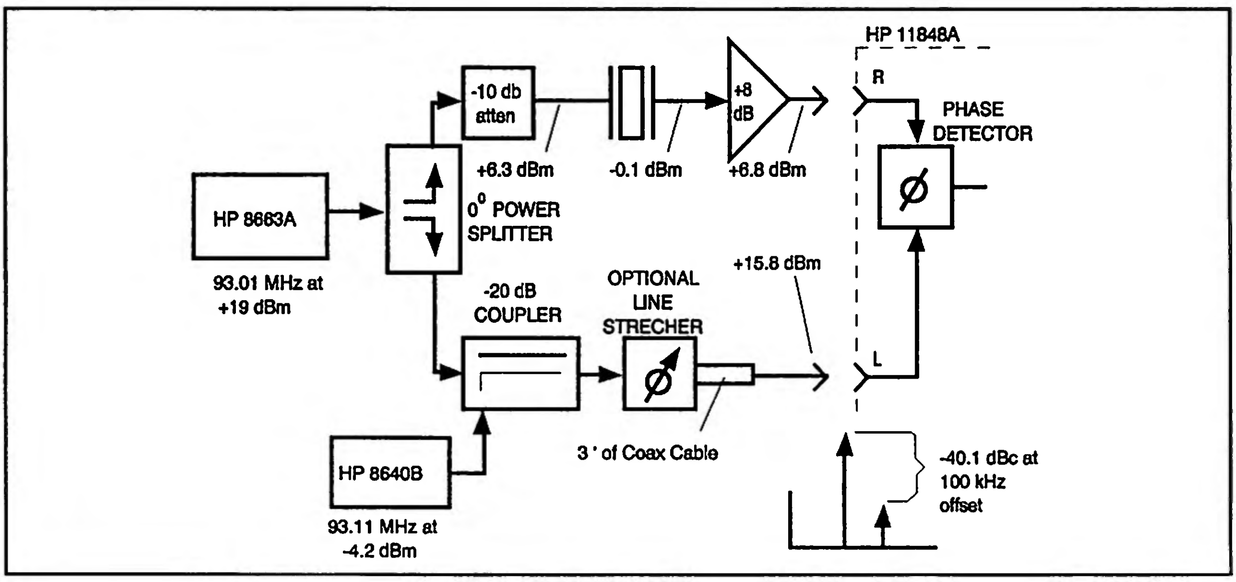


Figure 2-27. Crystal Measurement Setup

Conditions

In this example the DUT output is inadequate to drive the phase detector, thus a low-noise amplifier has been added to the DUT output. To ensure the validity of the resonator measurement, the setup noise floor must first be measured.

Measuring the Noise Floor of the Test Setup..

1. The insertion loss of the DUT must be measured so it can be accounted for in the noise floor measurement.
2. The HP 8663A was selected as the measurement source because of its low close-in phase noise and its ability to supply sufficient power so that an amplifier was not needed in the phase detector LO path.
3. The power level into the DUT (simulated by a 6 dB attenuator) is adjusted with an attenuator to set the desired test condition. A maximum input power of 7 dBm was specified by the manufacturer.

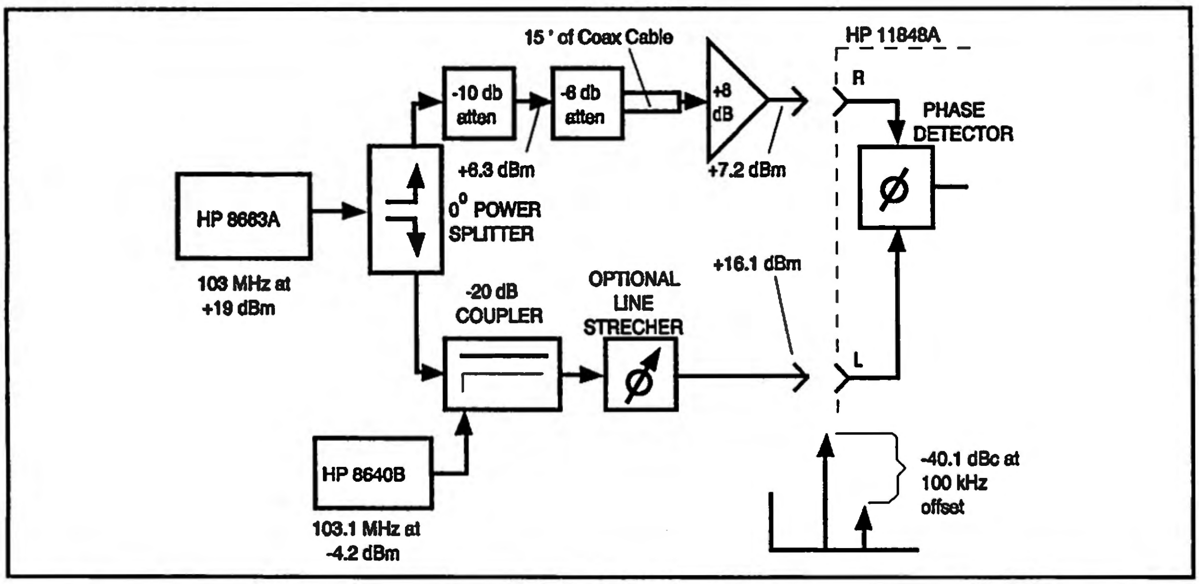


Figure 2-28. Noise-Floor Measurement Test Setup

4. Quadrature was obtained by adding about 3 meters of coax (about 25 ns) to the DUT path as a coarse adjustment and then using the HP phase shifter for the fine adjustment.

Note Coaxial delay lines for low-frequency measurements may become excessively long because:

$$\text{length for } 90° = \frac{c(v_r)}{4(f_o)}$$

$$\text{Where } c = 3 \times 10^8 \text{meters/second}$$

$$v_r = \text{relative velocity of the coax } (\approx 0.65)$$

$$f_o = \text{test frequency}$$

At 10 MHz, about 4.88 meters of coax are required. At this point, a lumped-element delay may be more desirable.

Measuring the Crystal Resonator.

1. The 6 dB attenuator is replaced by the crystal of equal loss.
2. The HP 8663A is adjusted to the crystal resonant frequency.
3. Note where the calibration spur is injected into the measurement. The DUT has a Q of about 100 in a 50 ohm system, which reduces the modulation bandwidth to about 465 kHz for modulation which must pass through it. A calibration signal after the DUT or in the other path does not have this restriction.
4. Quadrature is obtained by:
   a. Removing the three meters of coax used to obtain quadrature in the test-setup noise-floor measurement.
   b. Adding coax as needed as a coarse phase adjustment and using the phase shifter as the fine adjustment.

Results of the Setup Noise Floor Measurement

The results of the Setup Noise Floor Measurement are shown in figure 2-29. The following is an analysis of those results.

- The test setup, measured at a carrier frequency of 93.01 MHz, appears to be a valid measurement. The discontinuity at 1 kHz was caused by unresolved 60 Hz spurs which are resolved below 1 kHz.
- The noise floor was approximately −170 dBc/Hz at 50 kHz offset, with a −139 dBc/Hz, 1 Hz intercept. The large signal noise figure, calculated at 50 kHz offset, with 0 dBm input level, was 7 dB.
- The rise in the noise after 50 kHz offset was produced by the decorrelation caused by the 25 ns of delay in the three meters of coax used to set quadrature.

Results of the Crystal Resonator Measurement

The results of the Crystal Resonator Measurement are shown in figure 2-30. The following is an analysis of those results.

- The DUT noise data appears to be valid, with no major discontinuities.
- The Noise floor was approximately −142 dBc/Hz offset, with a −110 dBc/Hz, 1 Hz intercept. A broadband noise hump of about −140 dBc/Hz was observed from 1 kHz to the resonator half-bandwidth near 500 kHz.
- Data measured beyond 500 kHz was attenuated by the DUT.

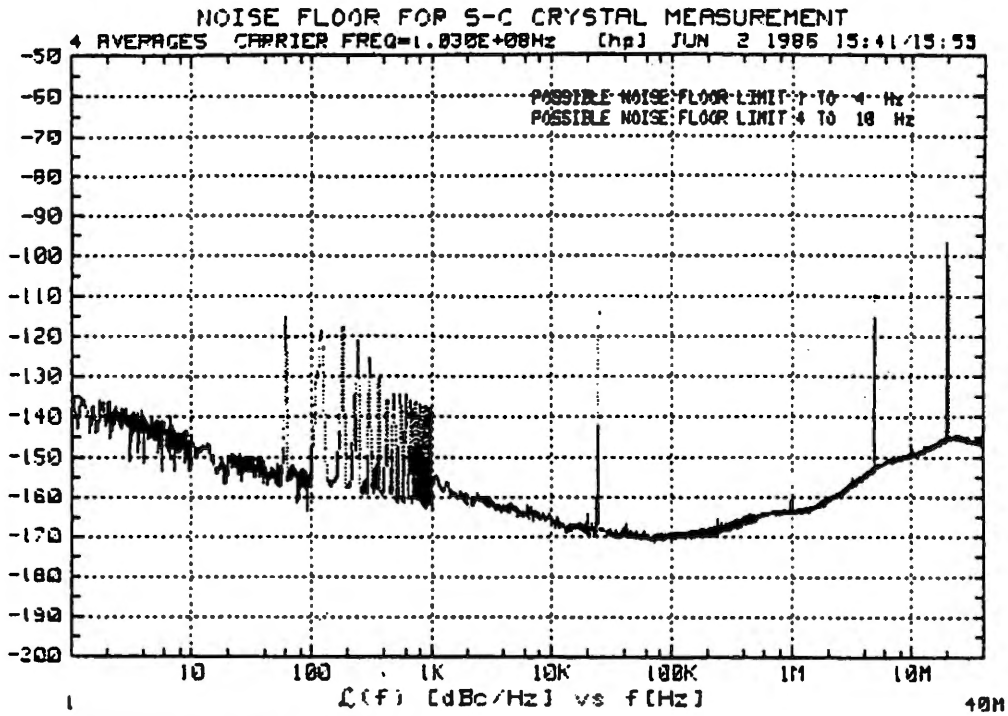


Figure 2-29. Noise Floor Measurement Results

```
            PRESENT SOURCE CHARACTERISTICS
CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±4.995 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 1.03000E+08 Hz
CARRIER FREQ = 1.03000E+08 Hz
INTERNAL MIXER IS 0, (5 MHz - 1.6 GHz)

            PRESENT MEASUREMENT CONSTANTS
VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.262 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz     LOOP BW3 = 33.51 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1       ATTEN2 = 1
```

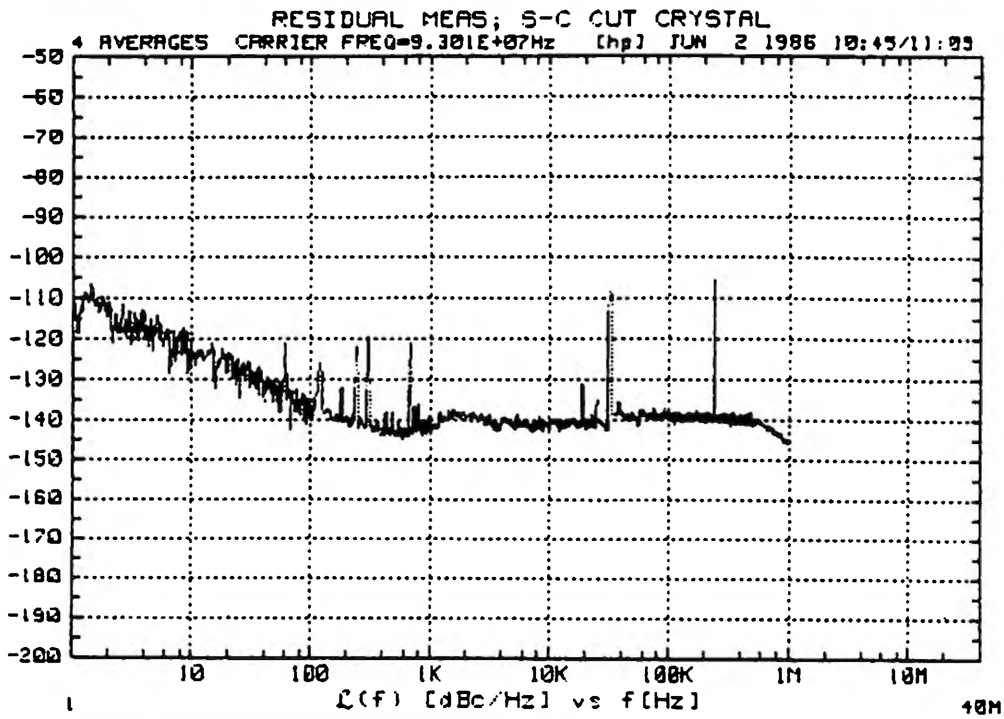


Figure 2-30. Noise Floor Measurement Results

```
            PRESENT SOURCE CHARACTERISTICS
CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±9.99 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 9.30140E+07 Hz
CARRIER FREQ = 9.30140E+07 Hz
INTERNAL MIXER IS 0, (5 MHz - 1.6 GHz)

            PRESENT MEASUREMENT CONSTANTS
VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.246 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz     LOOP BW3 = 33.51 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1       ATTEN2 = 1
```

Measurement of a Low-Noise Phase Modulator

This is a measurement of an extremely low-noise device. Several conditions must be met when measuring very quiet passive devices.

- All the power needed to drive the phase detector is supplied by the source. If more power is needed to provide adequate phase detector sensitivity, an amplifier may be placed between the source and the power splitter. Any noise source, such as an amplifier, placed in either phase detector path will contribute to the measured noise, and in this case, dominate the measurement.
- It is essential to keep the path lengths as short as possible between the phase detector and the power splitter to prevent decorrelation of the source noise.
- It is important to use a source with a good, low, noise floor so that the measurement will not be degraded in the event of a small amount of decorrelation.
- The test setup must be free of mechanically induced noise. Use semi-rigid cables and tighten all connections with a wrench.
- The test setup must be free of RFI-induced noise. All components in the setup must have adequate RFI shielding.

Initial Setup

In this example, the DUT is a varactor-tuned phase modulator.

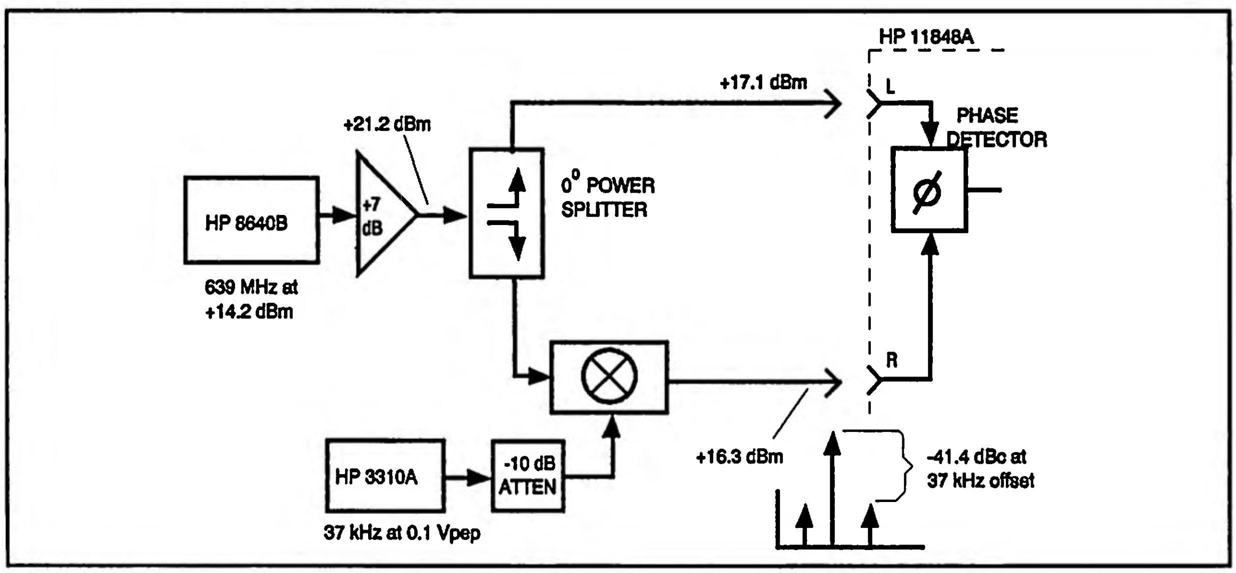


Figure 2-31. Low Noise Phase Modulator Measurement Setup

Conditions

This measurement was made under the following conditions.

- Because this is a very quiet device with approximately one octave of bandwidth from 500 MHz to 1000 MHz, quadrature will be established by allowing a very small difference in the path lengths between the power splitter and the phase detector and adjusting frequency.
- The modulation sideband levels in the reference path (without the DUT) must be measured to ensure that they are at least 20 dB below the sidebands of the phase modulator and

cause no cancellation of the calibration signal (that is, the power splitter must provide sufficient isolation).

- The RF and LO drive levels should be measured and checked against the phase detector sensitivity graph to ensure an adequate system noise floor.

Results

The results of this measurement are shown in figure 2-32. The following is an analysis of those results.

- This phase modulator has very low noise. It measured a noise floor of −178 dBc/Hz from 20 kHz to where the path length delay caused decorrelation above 500 kHz. It probably had about a 1 Hz intercept of −143 dBc/Hz. The fast rise in noise between 1 and 5 Hz was probably caused by the operator bumping the measurement table. The data should be retaken to verify this hypothesis.
- Between 60 Hz and 1 kHz, 60-Hz spurs dominate the measurement. Unresolved 60-Hz spurs probably cause the sharp rise in the noise between 1 and 10 kHz. This should be verified in Noise Monitor Mode using the FFT analyzer to look at the noise between the spurs.
- The phase modulator noise is so low that the system noise floor must be questioned. The phase detector constant for this measurement was 0.575 V/rad. Locating the phase detector constant on the phase detector sensitivity graph, figure 2-7, the corresponding measurement noise floor was −178 dBc/Hz. This implies that the DUT was at least 3 dB better than the measured data because the DUT and the system floor are equal.
- The system floor can be further investigated. For this particular HP 3047A system, the noise floor was measured at −181 dBc/Hz at 100 kHz offset with a phase detector constant of 0.752 V/rad. The phase noise floor can be calculated from this data for a test setup with a different phase detector constant. The difference in the system noise floor due to a change in detector constant is:

$$\begin{aligned}\Delta \text{ floor} &= \frac{20\log(\text{system floor detector constant})}{\text{new detector constant}} \\ &= 20\log\left(\frac{0.752}{0.575}\right) \\ &= 2.33\text{dB}\end{aligned}$$

The system noise floor, with 0.575 V/rad phase detector constant, is 2.33 dB higher, or −178.7 dBc/Hz at 100 kHz offset. It can be concluded from this that the actual noise of the phase modulator is at least 2 dB better than the measured data and was degraded by the noise floor contribution.

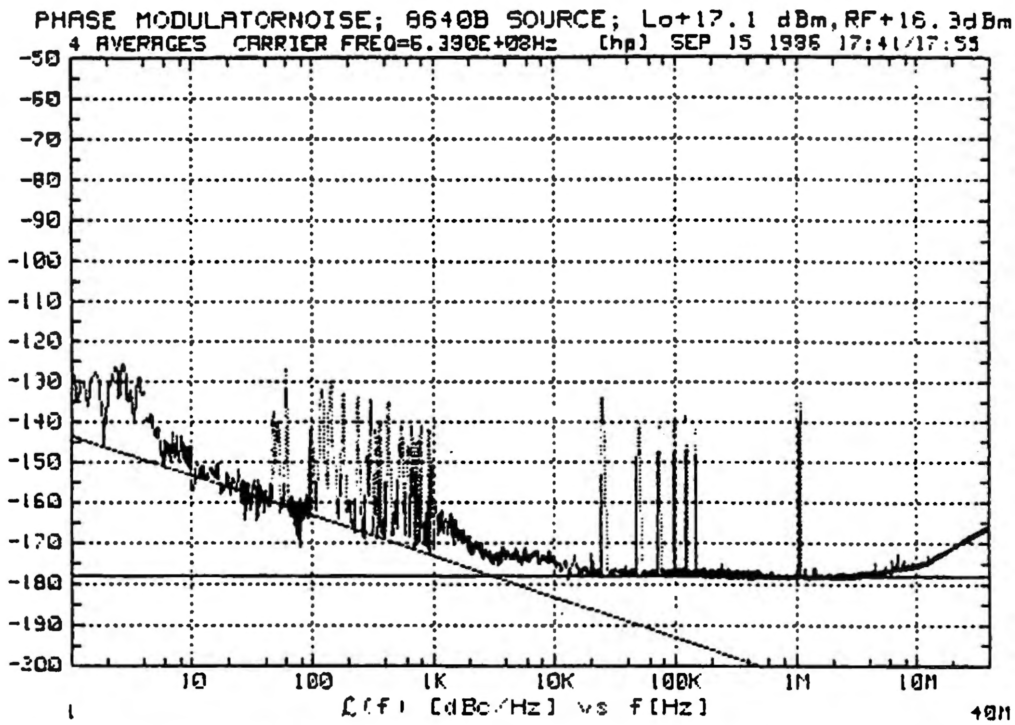


Figure 2-32. Phase Modulator Noise HP 8640B Source

PRESENT SOURCE CHARACTERISTICS

CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±10 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 6.39000E+08 Hz
CARRIER FREQ = 6.39000E+08 Hz
INTERNAL MIXER IS 0, (5 MHz - 1.6 GHz)

PRESENT MEASUREMENT CONSTANTS

VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.575 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz   LOOP BW3 = 0 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1   ATTEN2 = 1

Residual Noise of a Frequency Synthesizer

This measurement is made using the Beatnote Calibration Method.

The residual noise measurement of a frequency translating device, such as a divider, mixer, multiplier, phase-locked loop, or synthesizer, must meet several important requirements.

- Two units must be measured at the same time so that the signals at the phase detector are at the same frequency and in quadrature.
- The measured noise data is the sum of the two DUTs. If it can be assumed that the noise contribution of each DUT is equal, the noise of each individual DUT is 3 dB lower. If it cannot be assumed that the noise contributions are equal, then a third DUT must be measured against each of the first two DUTs. The data is then processed through a three source comparison program. The output will be the noise of each individual DUT.

Initial Setup

In this example, the DUTs were two HP 8663A synthesizers at 639 MHz and +19 dBm output level.

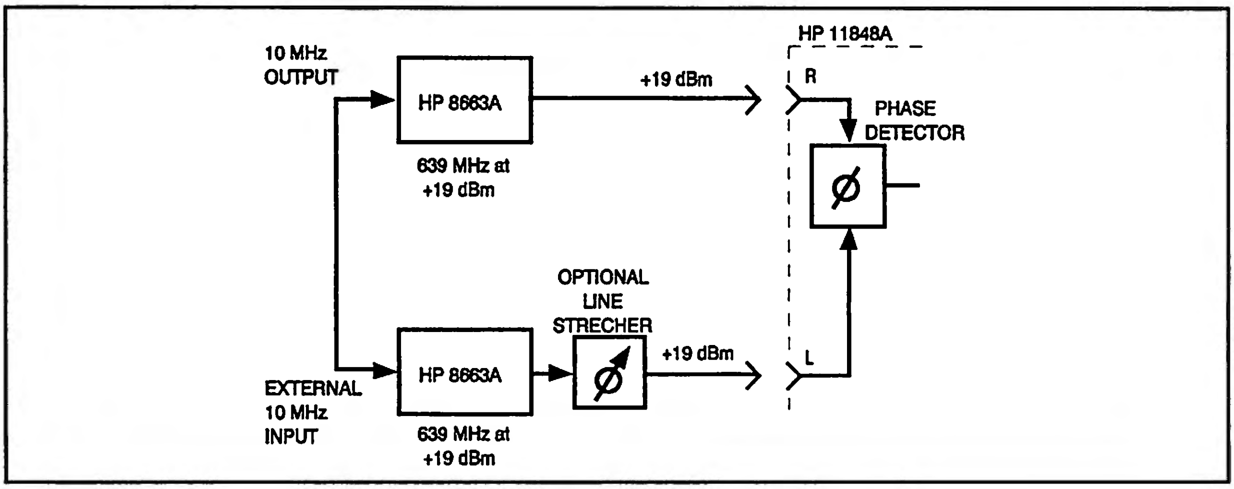


Figure 2-33. Frequency Synthesizer Measurement Setup

Conditions

This measurement was made under the following conditions.

- The two synthesizers are connected to the same high-stability time base, in this case, located inside one of the DUTs. The second DUT is connected through the time base output of the first DUT. It is critical that the phase shift from the semi-rigid cable and connectors between the two sources be minimized, as it will be multiplied by the DUT multiplication factor at the phase detector in this case, 35.11 dB.
- The beatnote was generated by setting one of the DUTs to 639.1 MHz during the calibration step, instead of adding a second calibration generator as in the beatnote calibration procedure.
- The DUTs must be allowed several hours to warm up. This helps to remove phase drift produced by the thermal effects of warm-up.
- An HP phase shifter was used to obtain and maintain quadrature.

  After a 30-minute warm-up period, there was still a significant amount of phase drift between the two instruments. (One DUT had been on all day and was stable.) It was necessary to adjust the phase shifter very slowly during the measurement to maintain quadrature. A sudden movement in the phase correction will look like phase noise close to the carrier and invalidate that data.

Results

The results of this measurement are shown in figure 2-34. The following is an analysis of those results.

- It is important to notice the "amplifier out" indicator in the lower left corner of the phase noise plot. This indicates that the low-noise amplifier (LNA) in the interface box, after the phase detector, was not used in this measurement. Without the LNA, with phase slope of 0.742 V/rad, the system noise floor is about −160 dBc/Hz.
- The measured data is for two DUTs. The correction for a single DUT is −3 dB at all frequency offsets. With that in mind, the broadband noise floor (offsets >2 MHz) is about −154 dBc/Hz. The noise pedestal at 10 kHz offset is −138 dBc/Hz, with 1 Hz intercept of −90 dBc/Hz.
- There are also some 60 Hz spurs mixed with some unknown noise or spurs. This region should be investigated using the Real-Time Measurement mode. This mode allows the use to manually adjust the spectrum analyzer bandwidth and frequency in the region of interest. The system can then be instructed to make calibrated, single-point measurements of both noise and spurs.
- The spurs beyond 10 kHz are probably generated inside the HP 8663s. They are all within the – dBc customer specification.

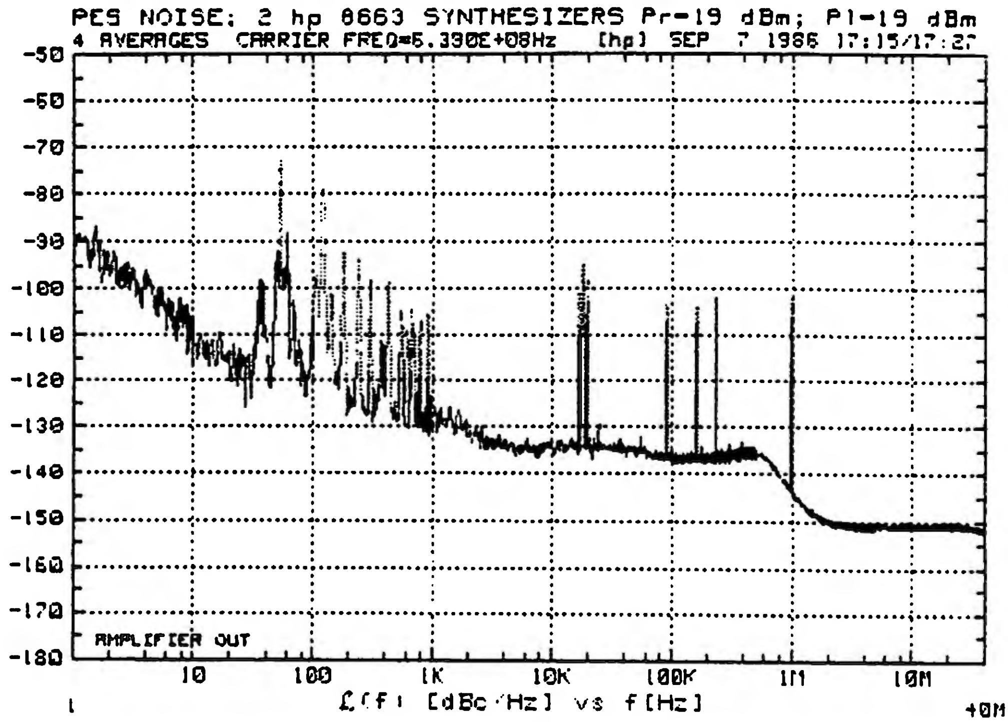


Figure 2-34. Residual Noise: Two HP 8663A Synthesizers

PRESENT SOURCE CHARACTERISTICS

CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±10 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 6.39000E+08 Hz
CARRIER FREQ = 6.39000E+08 Hz
INTERNAL MIXER IS 0, (5 MHz - 1.6 GHz)

PRESENT MEASUREMENT CONSTANTS

VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS OUT
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.742 VOLTS
DC OFFSET OF MIXER = 0.0026 VOLTS
LOOP BW1 = 0 Hz LOOP BW3 = 0 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1 ATTEN2 = 1

Residual Noise of a Comb Generator Multiplier

This measurement was performed using the Measured +/− DC Peak calibration method.

The following is a list of special considerations when measuring a comb generator.

- A comb generator is a frequency translating device. Two units must therefore be measured at the same time so that the signals at the phase detector are at the same frequency and in quadrature.
- Comb lines must be measured one at a time. If more than one comb frequency is allowed to enter the phase detector, the phase sensitivity may cancel.
- Extreme care must be taken between the power splitter and multiplier to avoid noise and phase glitches induced by loose connections and flexible cables, especially if the input frequency is above a few hundred MHz.
- Microwave residual noise measurements, in general, are much more difficult than RF measurements. This is largely because of the shorter wavelengths involved and the greater vulnerability to mechanically-induced noise at the higher frequency.
- Step-recovery-diode comb generators are very vulnerable to AM and ΦM conversion, especially if they are biased to increase efficiency. The source must be low in AM noise to prevent degradation of the measurement.

Initial Setup

In this example, the DUTs are a pair of HP 330C step-recovery diodes (SRDs) with an input frequency of 640 MHz.

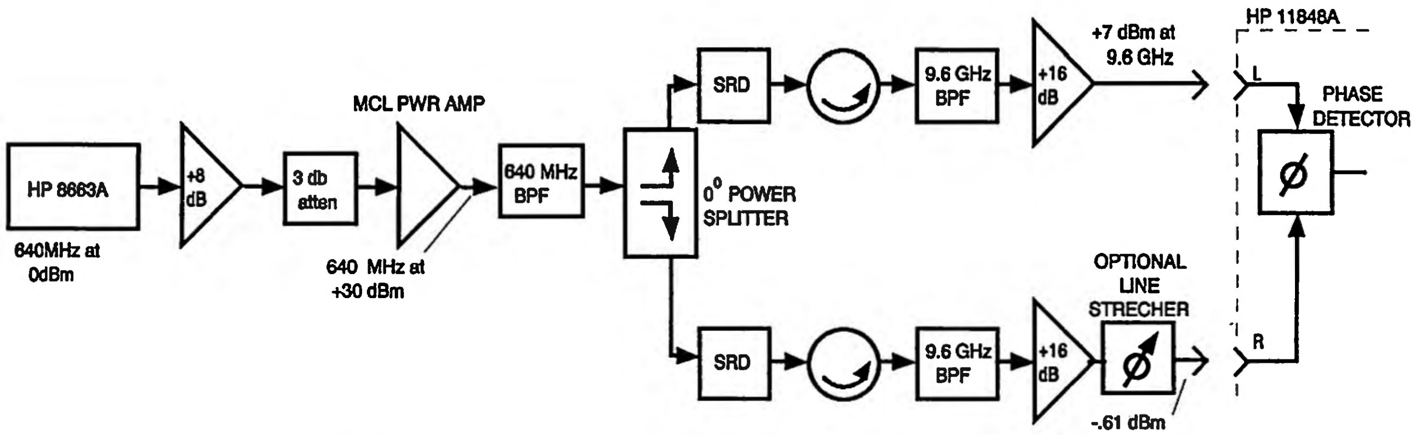


Figure 2-35. Residual Noise of Step-Recovery Diode Setup

Conditions

This measurement was made under the following conditions.

- The source used in this measurement is the 640 MHz auxiliary output of an HP 8663A which is then amplified to over +30 dBm. It is then filtered with a 640 MHz 5-pole bandpass filter before it reaches the power splitter. The auxiliary output of the HP 8663A avoids the noise pedestal of the HP 8663A output phase-locked loops and the noise generated by the GaAs FET output section.
- The output of the step-recovery diode (SRD) must be terminated in 50 ohms to prevent parametric oscillations. These oscillations will make it impossible to make a meaningful phase-noise measurement. In this case, a microwave circulator was used to provide the impedance match.
- The 15th harmonic at 9.6 GHz is selected by a bandpass filter to avoid harmonic cancellation of the phase slope.
- The power necessary to drive the phase detector is provided by a pair of 16 dB low-noise GaAs FET amplifiers. The test system noise floor should be measured to ensure that the system noise floor should be measured to ensure that the amplifiers do not dominate the SRD measurement. The 9.6 GHz microwave source needed for this measurement was not available at the time.
- An HP phase shifter was used to vary the phase at the phase detector through 360° while monitoring its output voltage on an oscilloscope. The positive and negative peaks were then measured and the phase slope was calculated.
- The phase shifter was then used to obtain and maintain quadrature. Phase drift through the SRDs (resulting from changes in temperature caused by the 0.5 Watt dissipation in varying air currents) made it necessary to adjust the phase shifter very slowly during the measurement in order to maintain quadrature. A sudden movement in the phase correction will look like phase noise close to the carrier and invalidate that data.

Results

The results of this measurement are shown in figure 2-36. The following is an analysis of those results.

- The data is for two DUTs measured at their outputs. The correction for a single DUT is −3 dB at all frequency offsets. With that in mind, the broadband noise floor with offsets >400 kHz is about −153 dBc/Hz. The noise at the 1 Hz intercept is about −103 dBc/Hz.
- The Equivalent input noise may be derived form measured output noise by the following.

  $$input\ noise = output\ noise + 20log\left(\frac{F_{in}}{F_{out}}\right)$$

  Where output noise of 1 DUT is 3 dB less than the plotted data.

  This corresponds to a broadband noise floor (>400 kHz) of −176.5 dBc/Hz noise at 1 kHz of −156 dBc/Hz and a 1 Hz intercept of −126.5 dBc/Hz.
- The noise hump at 15 MHz off the carrier results from parametric amplification of the noise floor in the SRDs.
- The discontinuity at 1 kHz is the result of unresolved 60 Hz spurs.

- The forest of spurs between 60 Hz and 1 kHz are 60 Hz spurs. It is important to note that the actual phase noise is measured at the bottom of the spurs, and that the spurs can be so bad that they obscure the phase-noise data entirely. The $\frac{1}{f}$ noise slope line accurately depicts the phase noise in this case. In general, slope lines may be added to the drawing to help determine the actual phase noise (which is a well-behaved response) and differentiate it from the irregularities of spurious interference.

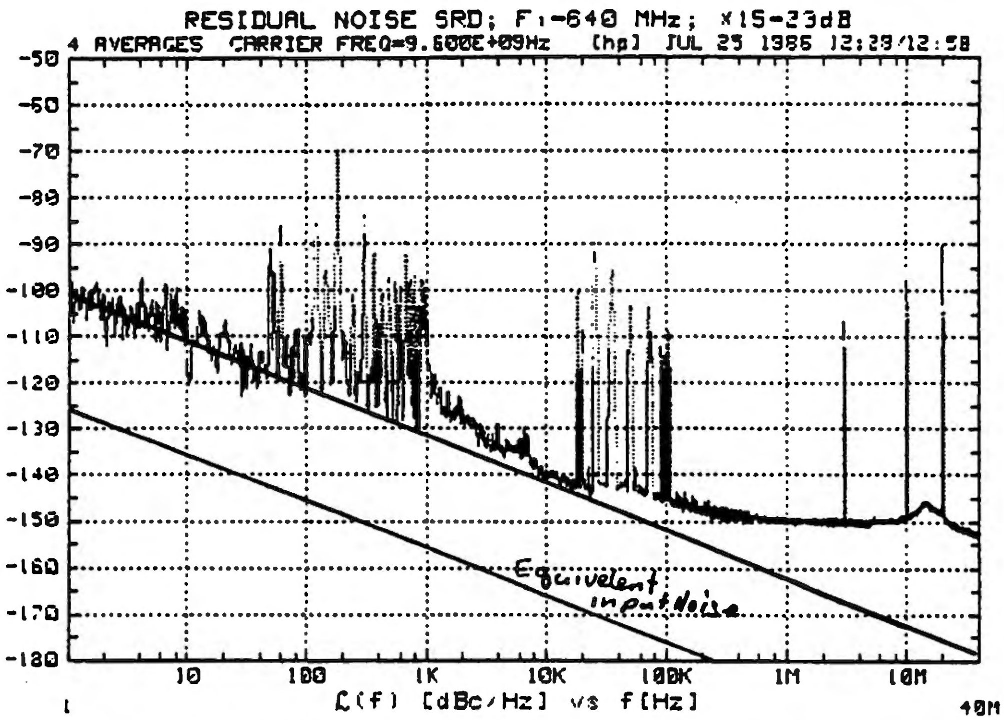


Figure 2-36. Residual Noise: Step Recovery Diode

```
              PRESENT SOURCE CHARACTERISTICS
CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±4.995 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 9.60000E+09 Hz
CARRIER FREQ = 9.60000E+09 Hz
INTERNAL MIXER IS EXTERNAL

              PRESENT MEASUREMENT CONSTANTS
VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.19 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz     LOOP BW3 = 0 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1      ATTEN2 = 1
```

Measurement of a Device Using an External Phase Detector

This measurement was performed using the Beatnote Calibration Method.

The external phase detector input extends the carrier frequency range to whatever frequencies are acceptable for the external detector, provided that:

- For carrier frequencies lower than 5 MHz, a low-pass filter is provided after the phase detector to attenuate the phase detector carrier feedthrough and all non-baseband mixer products to <50 dB below a beatnote of 0 dBc. The system must also be told that the phase detector frequency is 5 MHz or greater.
- For carrier frequencies less than 5 MHz, the largest offset frequency measured is inside the bandwidth of the phase detector filter.
- For carrier frequencies less than 5 MHz, the system must be told that the phase detector frequency is 5 MHz or greater.
- For carrier frequencies between 5 MHz and 95 MHz, the largest carrier offset available is 1 MHz or less. An alternative is an external low-pass filter added between the phase detector and the external input which meets the filter requirements of item 1. In this case, the system must be told that the phase detector frequency is greater than 95 MHz. The proper filter can make possible the measurements of carriers less than 95 MHz out to 40 MHz offset.

Note Data is only valid inside the bandwidth of the filter.

Initial Setup

The devices to be tested are two frequency dividers which are used in the HP 3325A frequency synthesizer.

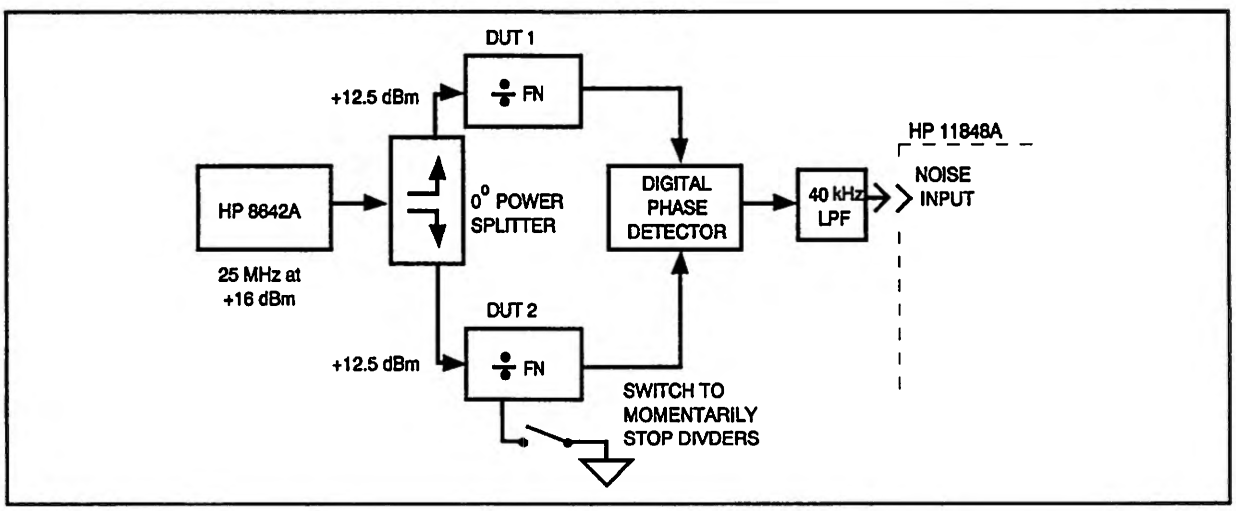


Figure 2-37. External Phase Detector Setup

Conditions

This measurement was made under the following conditions.

- This measurement is made using an external digital phase detector operating at 50 kHz. An external, 40-kHz, 7-pole, 5-zero, elliptical, low-pass filter following the phase detector keeps unwanted mixer products from saturating the low-noise amplifier that follows the phase detector.
- Care must be taken to minimize ground loops and sources of 66 Hz noise when using the external input. This prevents 60 Hz spurs from masking the noise data.
- The calibration beatnote was generated by changing the divide number of one of the two dividers which produced a frequency difference at the phase detector.
- Quadrature is achieved by starting dividers in the correct phase. This is accomplished by letting one divider operate continuously, while momentarily pausing the other. It may take several tries to get the second divider to start in the right phase.

Results

The results of this measurement are shown in figure 2-38. The following is an analysis of those results.

- The data is for two DUTs measured at their outputs. The correction for a single DUT is −3 dB at all frequency offsets. With that in mind, the broadband noise floor with offsets less than 100 Hz is about −172.5 dBc/Hz. The noise at the 1 Hz intercept is about −153 dBc/Hz. The ability of the system to draw slope lines at any offset and slope often simplifies interpretation of the data.
- The graph looks very well-behaved. The spurs in the region between 60 Hz and 1 kHz are 60 Hz spurs which are all below −120 dBc.
- The spurs at 50 kHz and its multipliers are phase-detector feedthrough and undesired mixer products which are all below −100 dBc.
- The data beyond 40 kHz is attenuated by the feedthrough filter.

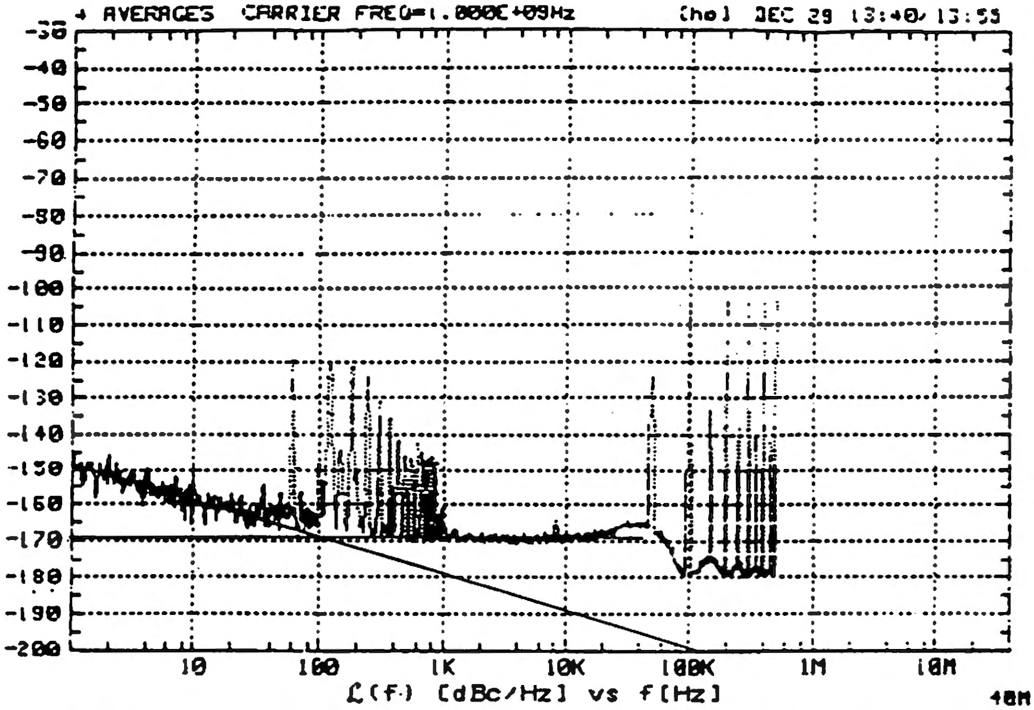


Figure 2-38. HP 3325A Fractional Divider/Phase Detector Measurement Results

```
          PRESENT SOURCE CHARACTERISTICS
CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±10 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 5.00000E+06 Hz
CARRIER FREQ = 1.00000E+06 Hz
INTERNAL MIXER IS EXTERNAL

          PRESENT MEASUREMENT CONSTANTS
VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.84 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz     LOOP BW3 = 0 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1      ATTEN2 = 1
```

3

AM-Source Noise Measurement

AM-Noise Measurement Theory of Operation

Basic Noise Measurement

The HP 3048A phase noise measurement system measures noise by:

- Calibrating of the noise detector sensitivity.
- Measuring the recovered baseband noise out to the detector.
- Calculating the noise around the signal by multiplying the measured data by the detector sensitivity.
- Displaying the measured noise data in the required format.

Once the detector is calibrated, the system looks at the signal out of the detector as just a noise voltage which must be measured over a band of frequencies regardless of the signal's origin.

The detector calibration is accomplished by applying a known signal to the detector. The known signal is then measured at baseband. Finally, the transfer function between the known signal and the measurement baseband signal is calculated.

Phase Noise Measurement

In the case of small angle phase modulation (<0.1 rad), the modulation sidebands' amplitude is constant with increasing modulation frequency. The phase detector gain can thus be measured at a single offset frequency, and the same constant will apply at all offset frequencies.

- In the case of calibrating with phase modulation sidebands, the system requires the carrier-to-sideband ratio and the frequency offset of the sidebands. The offset frequency is equal to the baseband frequency where the signal can be found. The ratio of the baseband signal voltage to the carrier-to-sideband ratio is the sensitivity of the detector.
- In the case of calibrating with a single-sided spur, it can be shown (see appendix C) that a single-sided spur is equal to a ΦM signal plus an AM signal. The modulation sidebands for both are 6 dB below the original single-sided spur. Since the phase detector attenuates the AM by more than 30 dB, the calibration constant can be measured as in the previous case, but with an additional 6 dB correction factor.

Amplitude Noise Measurement

The level of amplitude modulation sidebands is also constant with increasing modulation frequency. The AM detector gain can thus be measured at a single offset frequency and the same constant will apply at all offset frequencies. Replacing the phase detector with an AM detector, the AM noise measurement can be calibrated in the same way as ΦM noise measurement, except the phase modulation must be replaced with amplitude modulation.

The AM noise measurement is a source-type measurement. The residual AM noise of a DUT can only be calculated by measuring the source's AM noise, then subtracting that from the measured output noise of the DUT. The noise floor of this technique is the noise floor of the source.

AM Noise Measurement System Block Diagram

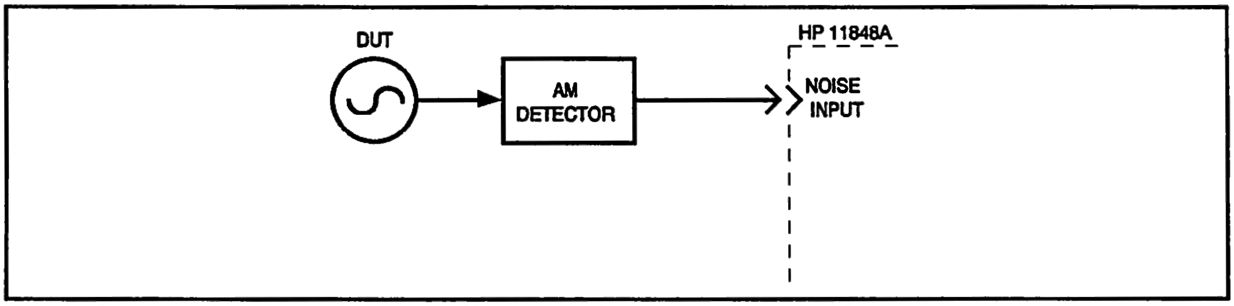


Figure 3-1. System Block Diagram

The noise measurement block diagram consists of adding an AM detector and an AM Detector filter to the external noise input of the HP 3048A.

AM Detector

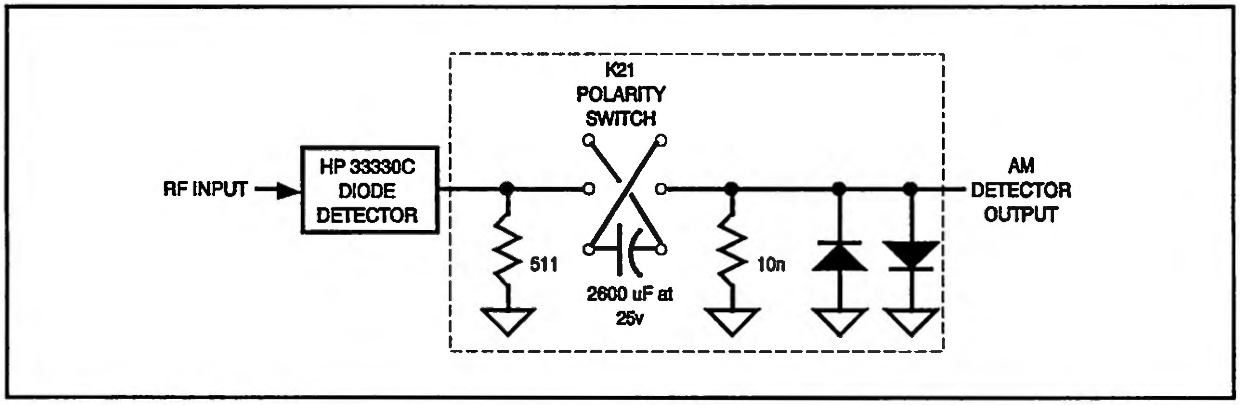


Figure 3-2. AM Detector Schematic

AM Detector Specifications

Detector type	low barrier Schottky diode
Carrier frequency range	10 MHz to 26.5 GHz
Maximum input power	+23 dBm
Minimum input power	0 dBm
Output bandwidth	1 Hz to 40 MHz

AM Detector Considerations

- The AM detector consists of an HP 33330C Low-Barrier Schottky Diode Detector and an AM detector filter (HP 3048A K21, see appendix G).
- The detector, for example, is an HP 33330C Low-Barrier Schottky-Diode Detector. The Schottky detectors will handle more power than the point contact detectors, and are equally as sensitive and quiet.
- The AM detector output capacitor prevents the dc voltage component of the demodulated signal from saturating the system's low noise amplifier (LNA). The value of this capacitor sets the lower frequency limit of the demodulated output. The cutoff frequency can be decreased by increasing the value of the dc blocking capacitor.

- Carrier feedthrough in the detector may be excessive for frequencies below a few hundred megahertz. The LNA is protected from saturation by the internal filters used to absorb phase detector feedthrough and unwanted mixer products. This limits the maximum carrier offset frequency to 1 MHz for input frequencies of less than 95 MHz and 40 MHz for carriers above 95 MHz.
- The 511 ohm resistor in the AM detector sets the dc bias-current in the diode detector. The ac load is 50 ohms, set by the input impedance of the test system. The 50 ohm load increases the detector bandwidth to greater than 40 MHz.
- A high impedance monitor port is provided on the AM detector for measuring calibration constants. This port must be bypassed with a feedthrough capacitor to prevent noise from entering the main signal path. It must not be connected during the actual noise measurement.
- The HP 11848A Phase Noise Interface must be dc blocked when using its NOISE INPUT. The interface will not tolerate more than ± 2 mV DC Input without overloading the LNA. A DC block must be connected in series after the AM Detector to remove the dc component. The HP 3048A Option K21 is designed specifically for this purpose.

Steps for Making AM Noise Measurements

1. Connect the system hardware and load/run the software.
2. Measure the system calibration data. (System calibration data is the correction data for all signal paths in the interface box.)
3. Main menu - Select the type of measurement to be made.
4. Establish parameters
   a. Source parameters
      - Phase detector input frequency
      - Carrier frequency
      - Select external detector
      - Select calibration option
   b. Measurement parameters
      - Start and stop frequency of measurement data
        - ☐ Without RF analyzer 0.01 Hz to 100 kHz
        - ☐ With RF analyzer 0.01 Hz to 40 MHz
      - Number of sweeps averaged on FFT analyzer

c. Plot parameters

- Graph type (usually Single-Sideband Phase Noise)
- Plotter type (if any)
- Minimum and maximum Y-axis (dBc)
- Minimum and maximum X-axis (Hz)
- Title

d. Measure

- Connect-up the device-under-test (DUT) and external hardware
- Measure calibration data by selected option
- Measure noise data

e. Interpret the measurement result

Choosing a Calibration Method

Method 1: User entry of phase detector constant

Method 1, Example 1

Advantages	Easy method of calibrating the measurement system
	Will measure DUT without modulation capability.
	Requires only an RF power meter to measure drive levels into the AM detector.
	Fastest method of calibration. If the same power levels are always at the AM detector, as in the case of leveled outputs, the AM detector sensitivity will always be essentially the same.
	Super-quick method of estimating the equivalent phase detector constant.
Disadvantages	It is the least accurate of the calibration methods.
	It does not take into account the amount of power at harmonics of the signal.
	It does not take into account the power which may be generated by spurious oscillations, causing the power meter to measure much more power than is at the AM detector.

Method 1, Example 2

Advantages	Easy method of calibrating the measurement system.
	Will measure DUT without modulation capability.
	Requires little additional equipment: only a voltmeter or an oscilloscope.
	Fastest method of calibration. If the same power levels are always at the AM detector, as in the case of leveled outputs, the AM detector sensitivity will always be essentially the same.
	Measures the AM detector gain in the actual measurement configuration. Super-quick method of estimating the equivalent phase detector constant.
Disadvantages	Has only moderate accuracy compared to the other calibration methods.

Method 2: Double Sided Spur

Method 2, Example 1

Advantages	Requires only one RF source (DUT) Calibration is done under actual measurement conditions so all non-linearities and harmonics of the AM detector are calibrated out. The double-sided spur method and the single-sided-spur method are the two most accurate methods for this reason.
Disadvantages	Required that the DUT have adjustable AM which may also be turned off. Requires the AM of the DUT to be extremely accurate; otherwise an RF spectrum analyzer, or modulation analyzer, for manual measurement of AM sidebands is required.

Method 2, Example 2

Advantages	Will measure source without modulation capability Calibration is done under actual measurement conditions so all non-linearities and harmonics of the AM detector are calibrated out. The double-sided spur method and the single-sided-spur method are the two most accurate methods for this reason.
Disadvantages	Requires a second RF source with very accurate AM modulation and output power sufficient to match the DUT. If the AM modulation is not very accurate, a spectrum analyzer or modulation analyzer must be used to make manual measurement of the AM sidebands.

Method 3: Single-Sided-Spur Method

Advantages	Will measure source without modulation capability. Calibration is done under actual measurement conditions so all non-linearities and harmonics of the AM detector are calibrated out. The double-sided spur method and the single-sided-spur method are the two most accurate methods for this reason.
Disadvantages	Requires 2 RF sources, which must be between 1 Hz and 40 MHz apart in frequency. Requires an RF spectrum analyzer for manual measurement of the signal-to-spur ratio and spur offset.

Calibration and Measurement General Guidelines

- **The AM detector must be well shielded from external noise especially 60 Hz noise.** The components between the diode detector and the test system should be packaged in a metal box to prevent RFI interference. Also, the AM detector should be connected directly to the test system if possible, to minimize ground loops. If the AM detector and test system must be separated, semi-rigid cable should be used to keep the shield resistance to a minimum.
- **Although AM noise measurements are less vulnerable than residual phase-noise measurements to noise induced by vibration and temperature fluctuation,** care should be taken to ensure that all connections are tight and that all cables are electrically sound.
- **The output voltage monitor on the AM detector must be disconnected** from digital voltmeters or other noisy monitoring equipment before noise measurement data is taken.
- **The $\frac{1}{f}$ noise floor of the detector may degrade as power increases above +15 dBm.** Noise in the $\frac{1}{f}$ region of the detector is best measured with about +10 dBm of drive level. The noise floor is best measured with about +20 dBm of drive level.
- **An amplifier must be used in cases where the signal level out of the DUT is too small to drive the AM detector or is inadequate to produce a low enough measurement noise floor.** In this case the amplifier should have the following characteristics.
  - It should have the lowest possible noise figure, and the greatest possible dynamic range.
  - The signal level must be kept as high as possible at all points in the test setup to avoid noise floor degradation.
  - It should have only enough gain to get the required signal levels. Excess gain leads to amplifiers operating in gain compression, increasing their likelihood of suppressing the AM noise to be measured.
  - The amplifier's sensitivity to power supply noise and the supply noise itself must both be minimized.

Calibration and Measurement Procedures

Method 1: User Entry of Phase Detector Constant

Method 1, example 1

1. Connect circuit as shown in figure 3-3, and tighten all connections.

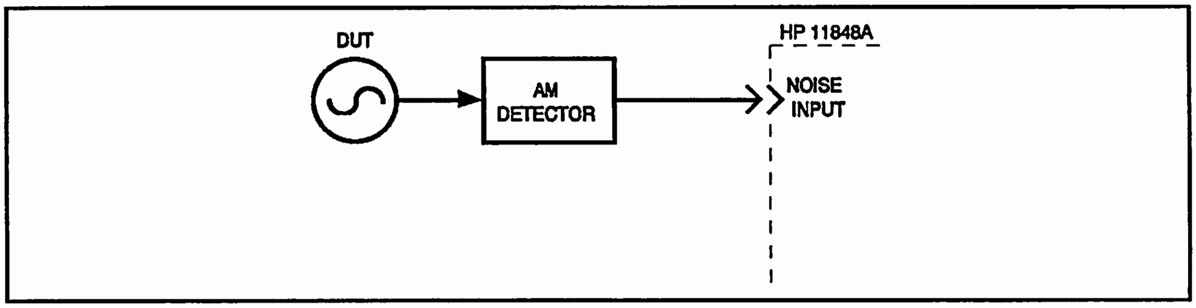


Figure 3-3. AM Noise Measurement Setup

2. Measure the power which will be applied to the AM detector. It must be between 0 and +23 dBm.

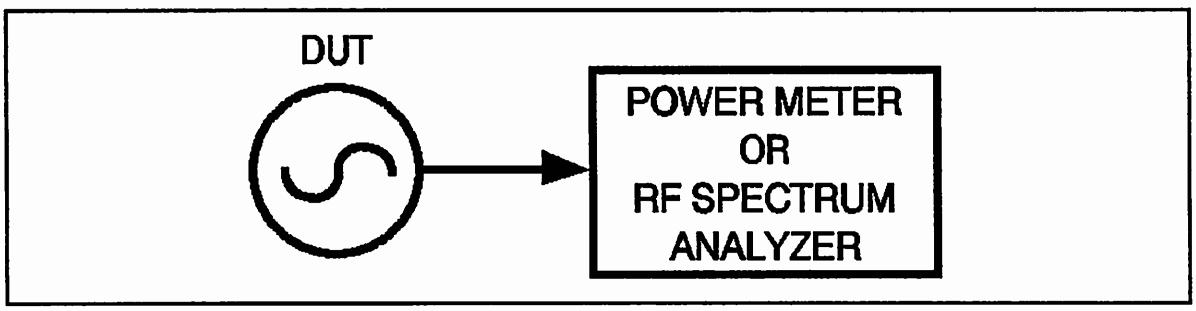


Figure 3-4. AM Noise Calibration Setup

3. Locate the drive level on the AM sensitivity graph (figure 3-5), and enter the data.

4. Measure the noise data and interpret the results. The measured data will be plotted as single-sideband AM noise in dBc/Hz.

Note The quadrature meter should be at zero volts due to the blocking capacitor at the AM detector's output.

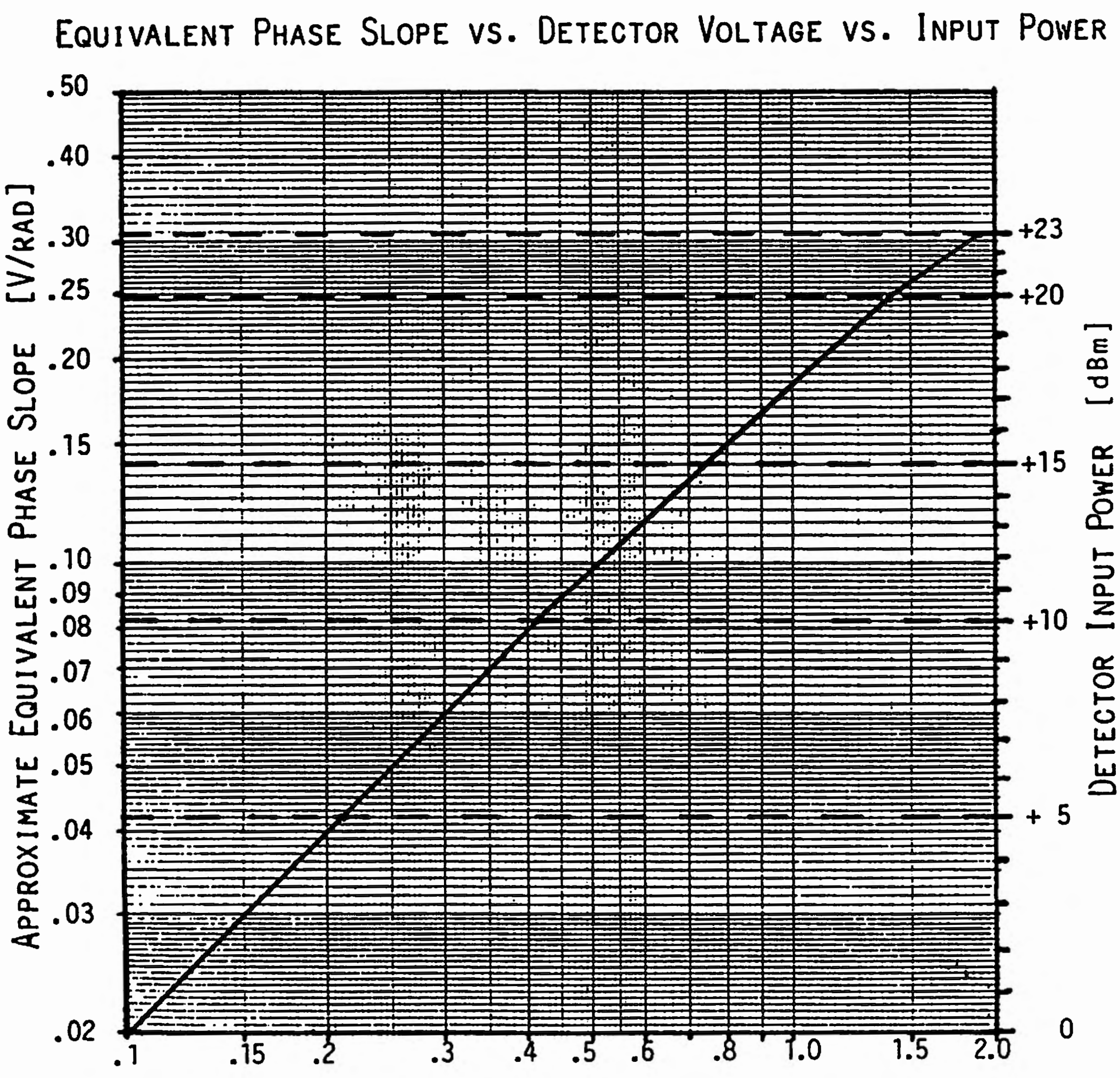


Figure 3-5. AM Detector Sensitivity Graph

Method 1, Example 2

1. Connect circuit as per figure 3-6, and tighten all connections.
2. Measure the power which will be applied to the AM detector. It must be between 0 and +23 dBm.

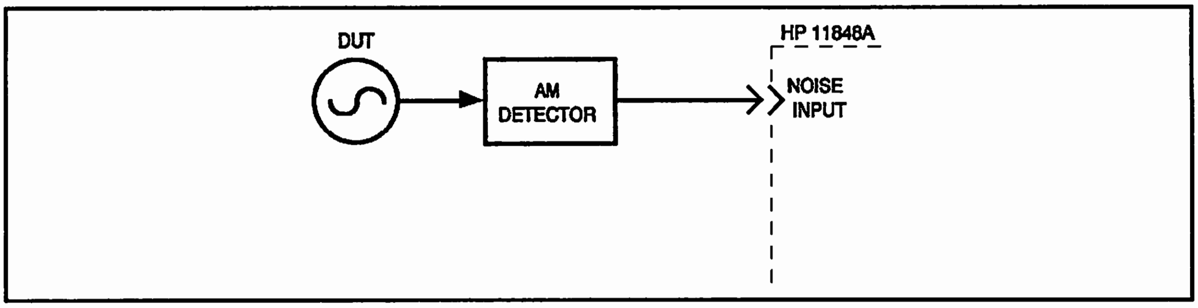


Figure 3-6. AM Noise Measurement Setup

3. Measure the monitor output voltage on the AM detector with an oscilloscope or voltmeter. Locate the diode detector's dc voltage along the bottom of the AM sensitivity graph (figure 3-7). Moving up to the diagonal calibration line and over, the equivalent phase detector constant can then be read from the left side of the graph. The measured data will be plotted as single-sideband AM noise in dBc/Hz.

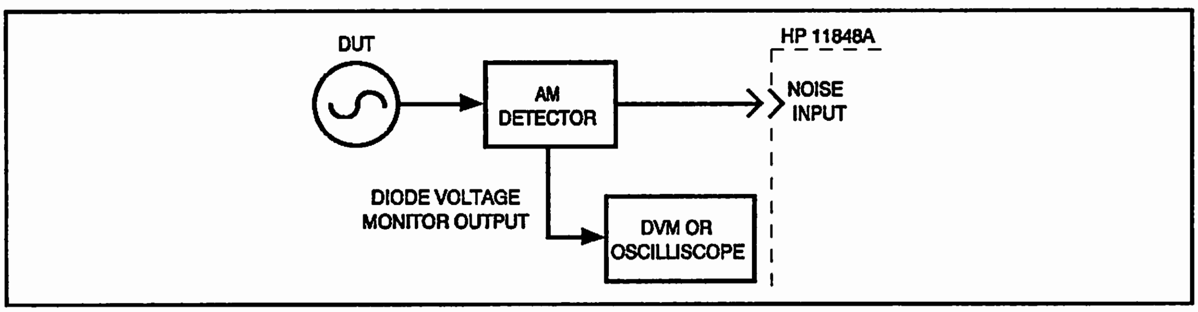


Figure 3-7. Modulation Sideband Calibration Setup

4. Measure noise data and interpret the results.

Note The quadrature meter should be at zero volts due to the blocking capacitor at the AM detector's output.

Method 2: Double-Sided Spur Method

Method 2, Example 1

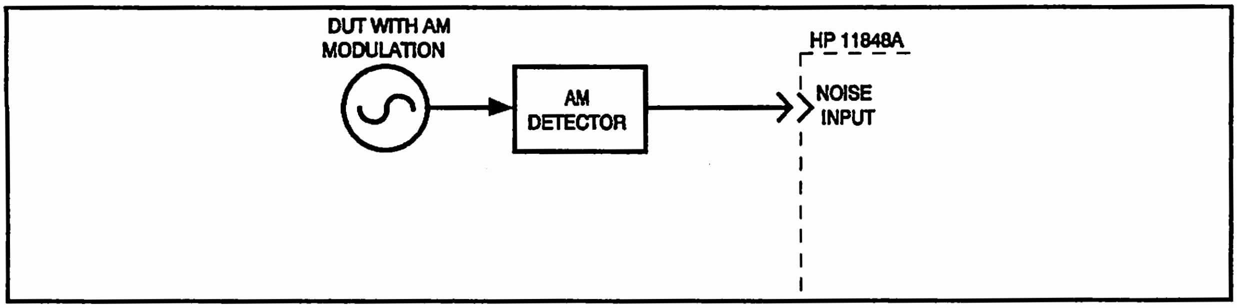


Figure 3-8. AM Noise Measurement Setup

1. Connect circuit as shown in figure 3-8, and tighten all connections.
2. Measure the power which will be applied to the AM detector. It must be between 0 and +23 dBm.
3. Measure the carrier-to-sideband ratio of the AM at the AM detector's input with an RF spectrum analyzer or modulation analyzer. The source should be adjusted such that the sidebands are between −30 and −60 dB below the carrier with a modulation rate between 1 Hz and 20 MHz.

Note

The carrier-to-sideband ratio ($\frac{C}{sb}$) for AM is:

$$\frac{C}{sb} = 20log(\frac{\%AM}{100}) - 6\ dB$$

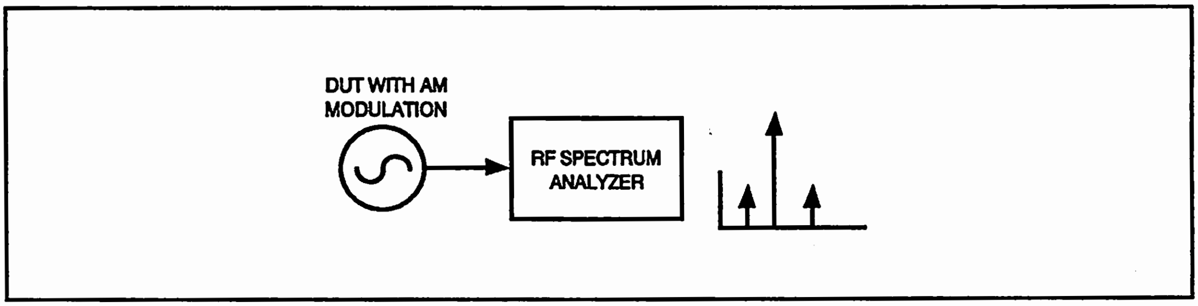


Figure 3-9. Measuring the Carrier-to-Sideband Ratio

4. Reconnect the AM detector and enter the carrier-to-sideband ratio and modulation frequency.

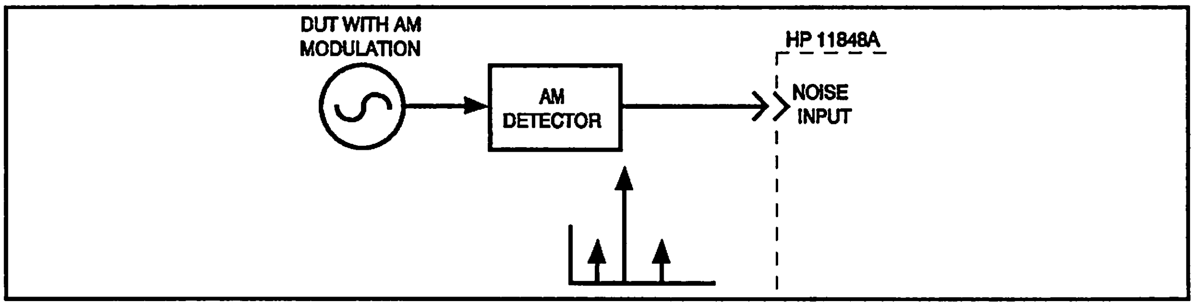


Figure 3-10. Measuring the Calibration Constant

5. Measure the AM detector calibration constant.
6. Turn off AM.
7. Measure noise data and interpret the results.

Note

The quadrature meter should be at zero volts due to the blocking capacitor at the AM detector's output.

Method 2, Example 2

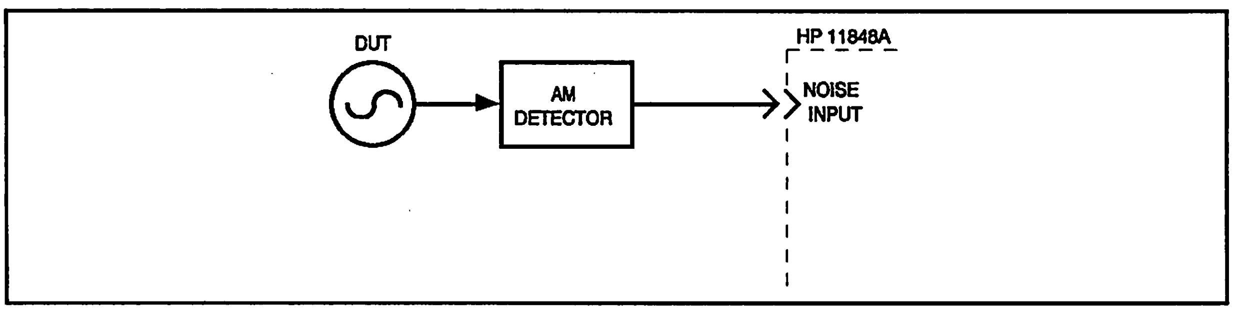


Figure 3-11. AM Noise Measurement Setup

1. Connect circuit as shown in figure 3-11, and tighten all connections.
2. Measure the power which will be applied to the AM detector. It must be between 0 and +23 dBm.

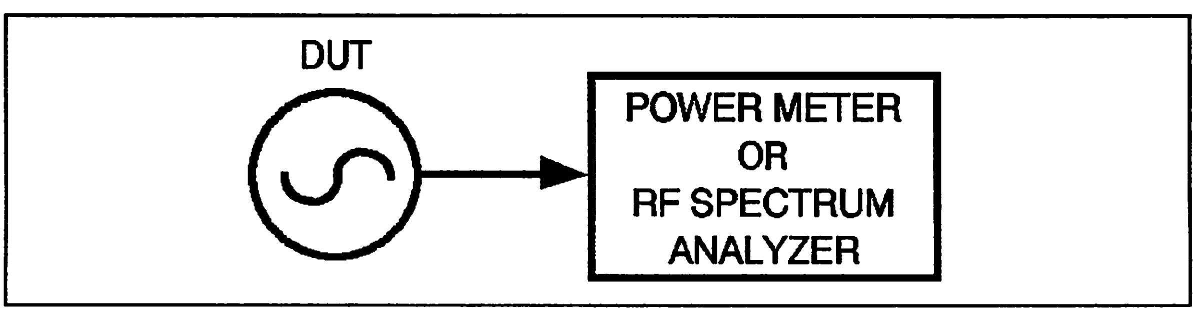


Figure 3-12. Measuring Power at the AM Detector

3. Using a source with AM, set its output power equal to the power measured in step 2. The source should be adjusted such that the sidebands are between −30 and −60 dB below the carrier with a modulation rate between 1 Hz and 20 MHz.

Note

The carrier-to-sideband ratio $\frac{C}{sb}$ for AM is:

$$\frac{C}{sb} = 20log(\frac{\%AM}{100}) - 6\ dB$$

To check the AM performance of the source, measure the carrier-to-sideband ratio of the AM at the source output with an RF spectrum analyzer or modulation analyzer.

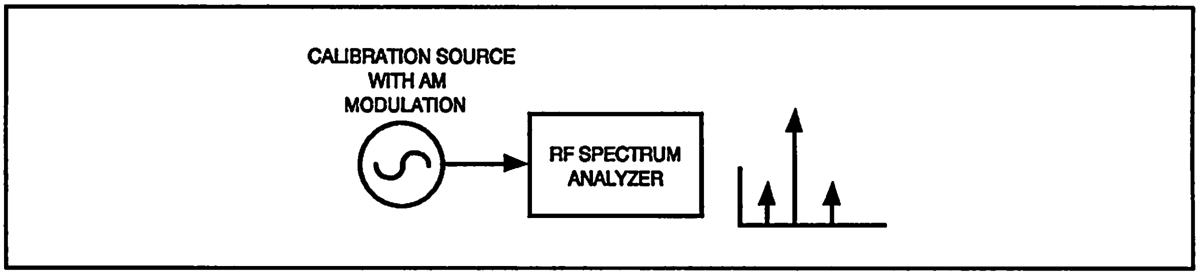


Figure 3-13. Measuring Carrier-to-Sideband Ratio

4. Enter the carrier-to-sideband ratio and offset frequency, then measure the calibration constant.

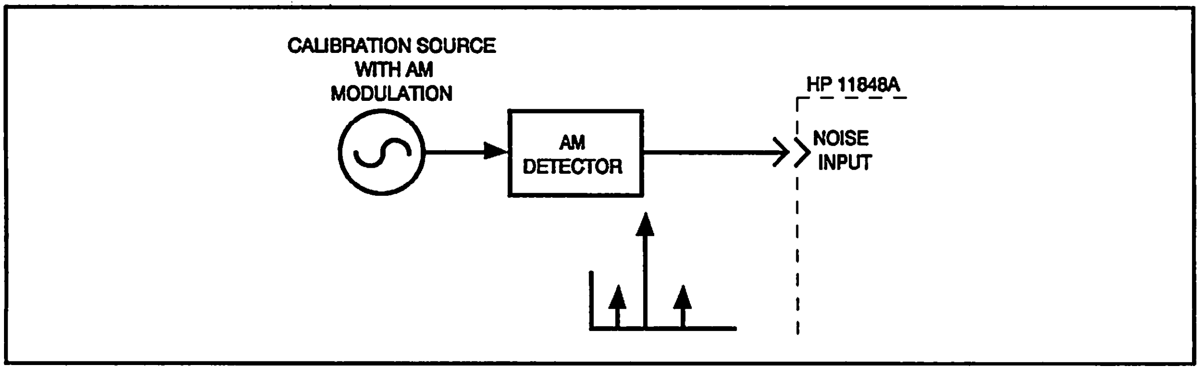


Figure 3-14. Measuring the Calibration Constant

5. Remove the AM source and reconnect the DUT.
6. Measure noise data and interpret the results.

Note

The quadrature meter should be at zero volts due to the blocking capacitor at the AM detector's output.

Method 3: Single-Sided-Spur Method

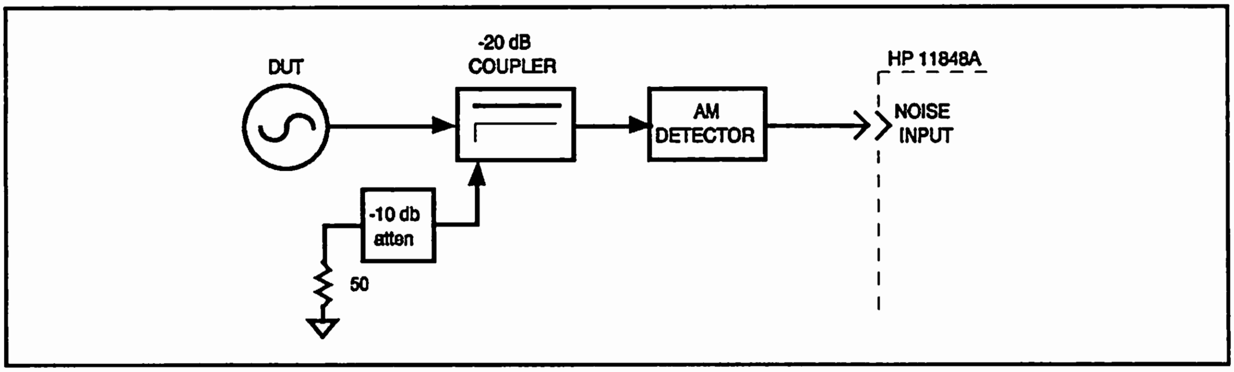


Figure 3-15. AM Noise Measurement Setup Using Single-Sided-Spur

1. Connect circuit as shown in figure 3-15, and tighten all connections.
2. Measure the power which will be applied to the AM detector. It must be between 0 and +23 dBm.
3. Measure the carrier-to-single-sided-spur ratio and the spur offset at the input to the AM detector with an RF spectrum analyzer. The spur should be adjusted such that it is between −30 and −60 dBc, with a carrier offset of 1 Hz to 20 MHz.

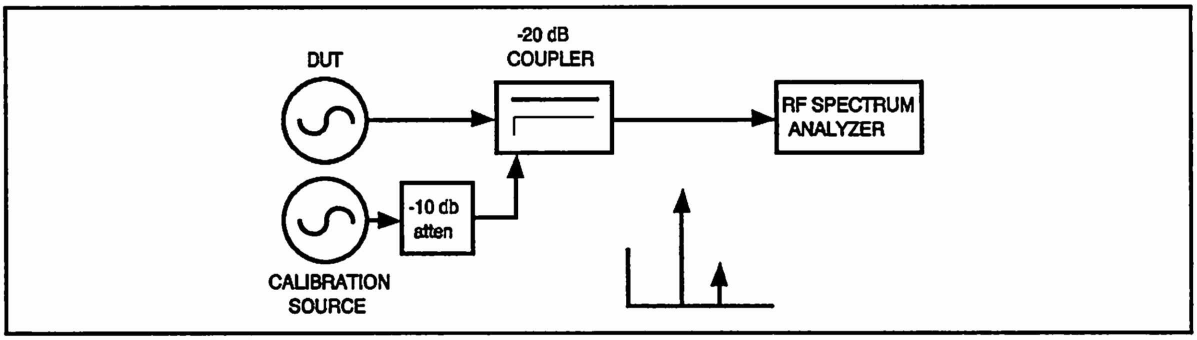


Figure 3-16. Measuring Relative Spur Level

4. Reconnect the AM detector and measure the detector sensitivity.

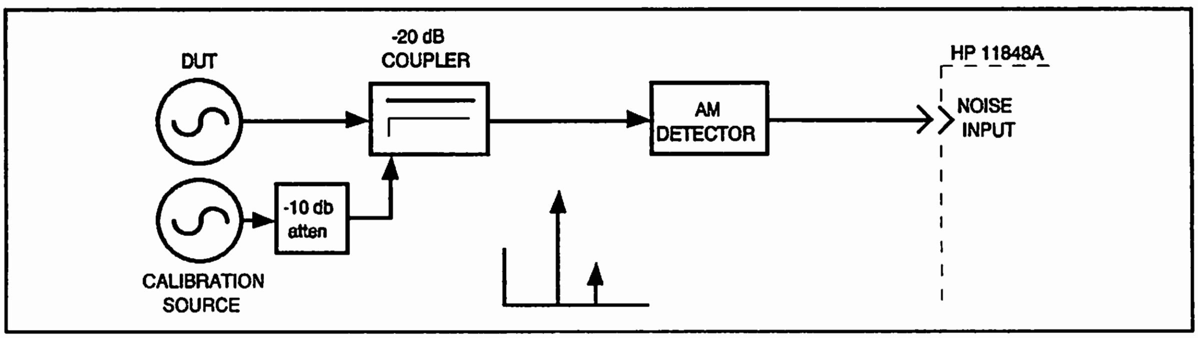


Figure 3-17. Measuring Detector Sensitivity

5. Turn off the spur source output.
6. Measure noise data and interpret the results.

Note

The quadrature meter should be at zero volts due to the blocking capacitor at the AM detector's output.

Examples of AM Noise Measurements

Measurement of a Source with AM

This measurement uses the Double Sided Spur Calibration Method.

The measurement of a source with amplitude modulation capability is among the simplest of the AM noise measurements. The modulation sidebands used to calibrate the AM detector are generated by the DUT. In cases where the percent modulation and modulation rate are known very accurately, the carrier-to-sideband ratio may be calculated, where:

$$\frac{c}{sb} = 20log(\frac{\%AM}{100}) - 6\ dB$$

The percent modulation may also be measured with a modulation analyzer, such as an HP 8901A/B or HP 8902A, or the modulation sidebands may be measured directly with an RF spectrum analyzer.

Initial Setup

In this example, the DUT is an HP 8642A signal generator, with an output frequency of 640 MHz and an output power of +15 dBm.

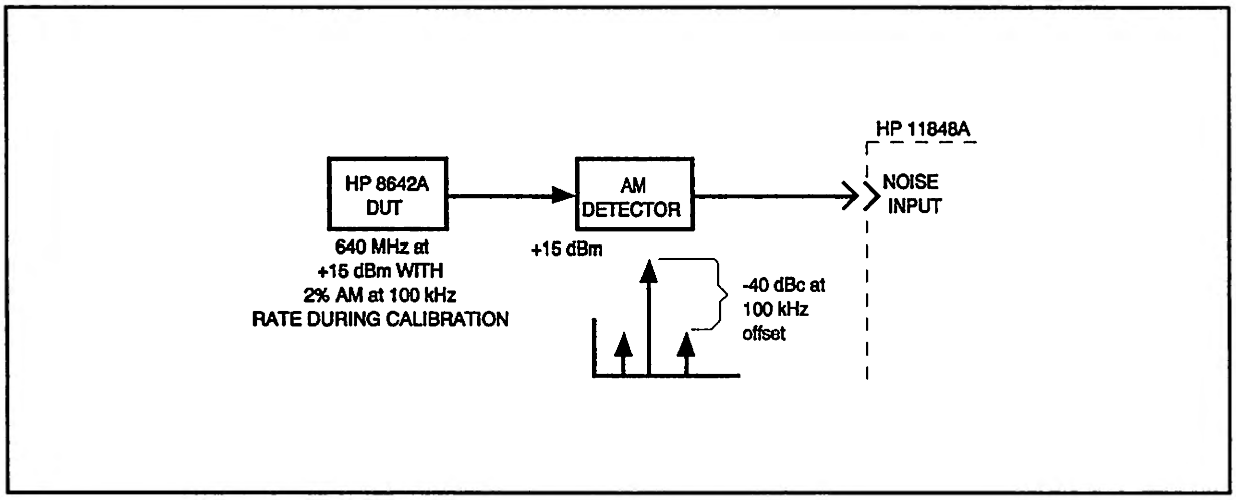


Figure 3-18. AM Noise Measurement Setup for a Source with AM

Conditions

This measurement was performed under the following conditions.

- All the power required to drive the AM detector is provided by the DUT.
- The AM detector is calibrated by setting the DUT to 2% AM at a 100 kHz modulation rate. This produces −40 dBc amplitude modulation sidebands with an offset of 100 kHz.
- The modulation sideband amplitude was verified with an HP 8568A RF spectrum analyzer.
- The signal power must not significantly change when the calibration modulation is removed.
- The quadrature meter should be at zero volts due to the blocking capacitor in the AM detector.

Results

The results of this measurement are shown in figure 3-19. The following is an analysis of those results.

- The signal generator, measured at 640 MHz, has an AM noise floor of −160 dBc/Hz at carrier offsets greater than 50 kHz, with a 1 Hz intercept of −123 dBc/Hz. It is important to note that the HP 8642A signal generator has one of the lowest AM noise floors available, which makes it an excellent signal source for residual 2-port phase noise measurements.
- The spurs between 60 Hz and 1 kHz are due to 60 Hz line spurs, possibly induced by a ground loop between the DUT and the test system.
- The discontinuity at 1 kHz is caused by unresolved 60 Hz spurs.
- All the remaining spurs above 10 kHz are less than the DUT's −100 dBc spur specification.

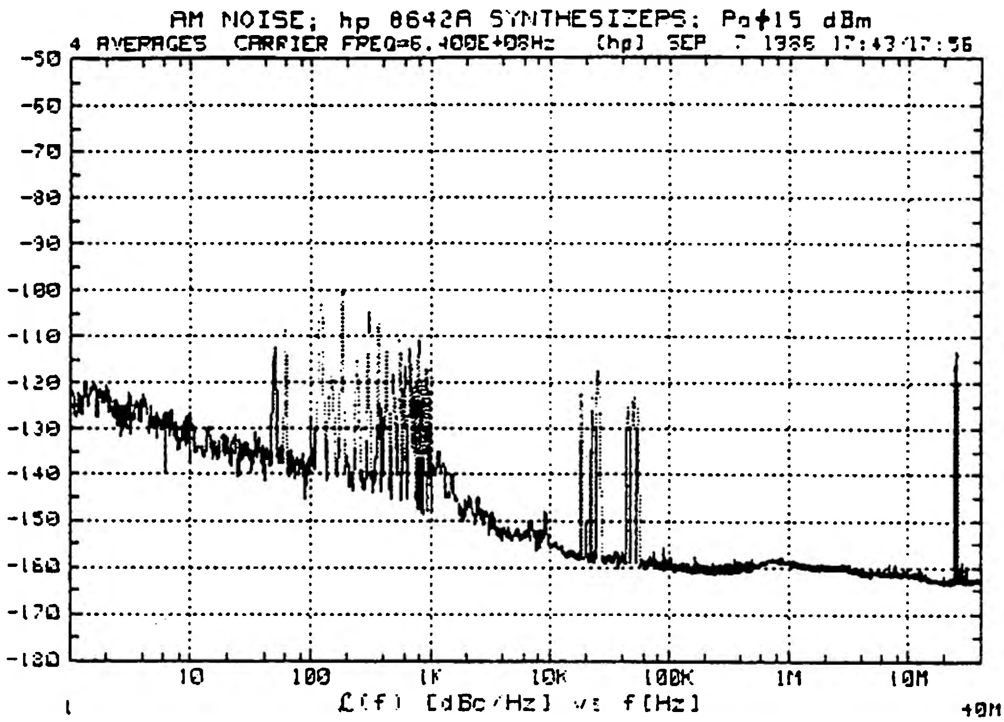


Figure 3-19. AM Modulation Measurement Results

```
            PRESENT SOURCE CHARACTERISTICS
CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±10 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 6.40000E+08 Hz
CARRIER FREQ = 6.40000E+08 Hz
INTERNAL MIXER IS EXTERNAL

            PRESENT MEASUREMENT CONSTANTS
VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.143 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz     LOOP BW3 = 0 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1      ATTEN2 = 1
```

Measurement Of A Source Without AM

This measurement uses the Single-Sided-Spur Calibration Method.

The single-sided-spur method is the most accurate calibration technique for sources without amplitude modulation capability. It requires that a single-sided-spur be added to the signal. It can be shown (see appendix C) that the single-sided-spur is equal to amplitude modulation plus phase modulation, both with sidebands 6 dB below the single-sideband spur. Since the AM detector is not sensitive to phase modulation, the ΦM sidebands are stripped away, and the AM sidebands are demodulated. The sensitivity of the AM detector is equal to the ratio of the recovered baseband signal to the signal-to-spur ratio minus 6 dB.

Initial Setup

In this example, the DUT is a 100 MHz voltage-controlled crystal oscillator followed by a Mini Circuits power amplifier with an output power of +33.4 dBm at 100 MHz.

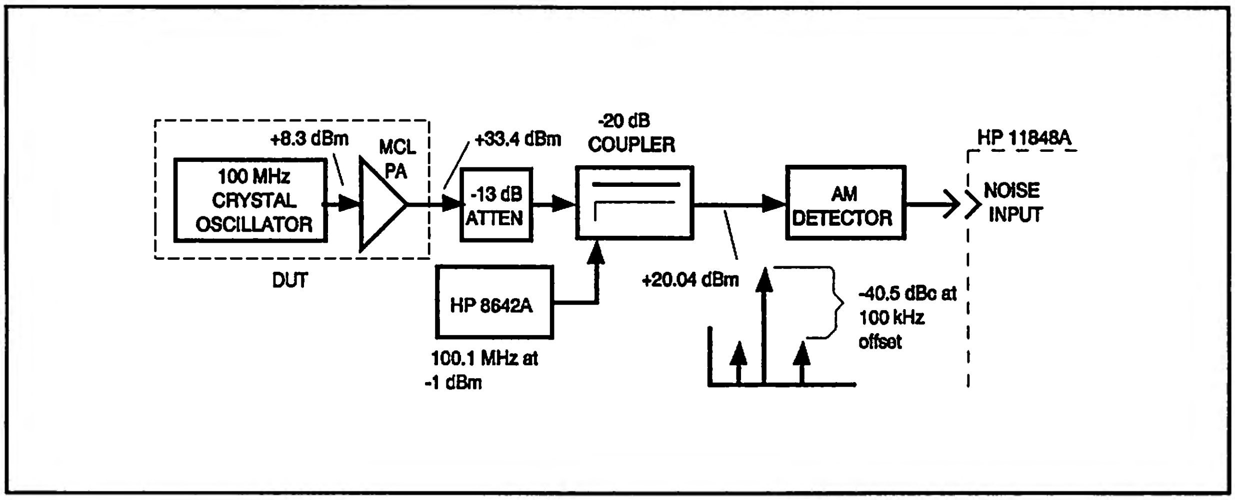


Figure 3-20. Source Without AM Measurement Setup

Conditions

This measurement was performed under the following conditions.

- The power out of the DUT is greater than the maximum power rating of the diode detector. Thus, the output must be attenuated to less than +23 dBm, but still remain large enough to provide an adequate AM-detector sensitivity, as detector sensitivity is directly proportional to the detector input power.
- The AM detector is calibrated by adding a −40.5 dBc spur to the main signal through a −20 dB coupler. The spur has an offset frequency of 100 kHz. After the detector is calibrated, the spur is removed by setting the calibration generator output power to the noise floor, while maintaining the impedance match of the coupler's coupled port.
- The carrier-to-spur ratio is measured with an HP 8568A RF spectrum analyzer.
- The signal power must not significantly change when the calibration modulation is removed.
- A signal level of +20 dBm provides the lowest detector noise floor.
- The quadrature meter should be at zero volts due to the blocking capacitor in the AM detector.

Results

The results of this measurement are shown in figure 3-21. The following is an analysis of the results.

- The crystal-oscillator/power-amplifier combination measured at 100 MHz has a noise floor of at least −170 dBc/Hz at offsets greater than 1 MHz. The system noise floor can be estimated by comparing the equivalent phase slope to the Phase Detector Sensitivity Graph (figure 2-8).
- The 1 Hz intercept noise is at least −116 dBc/Hz. The large $\frac{1}{f}$ noise region is probably due to one of two mechanisms:
  - The noise of the power amplifier. This should be investigated by removing the power amplifier, and remeasuring the oscillator's AM noise.
  - The diode detector in the AM detector is operating at a very high power level to measure the noise floor performance. The high power may be degrading the $\frac{1}{f}$ performance of the detector. The $\frac{1}{f}$ region of the noise data should be remeasured with an additional 10 dB of attenuation placed before the AM detector, which will lower the input level to +10 dBm.
- The spurs between 60 Hz and 1 kHz are due to 60 Hz line spurs, possibly induced by a ground loop between the DUT and the test system.
- The discontinuity at 1 kHz is caused by unresolved 60 Hz spurs.
- Spurs in the 1 to 2 MHz region are produced by the Shared Resource Management System multiplexer with the new HP 50961A SRM coax adapter.

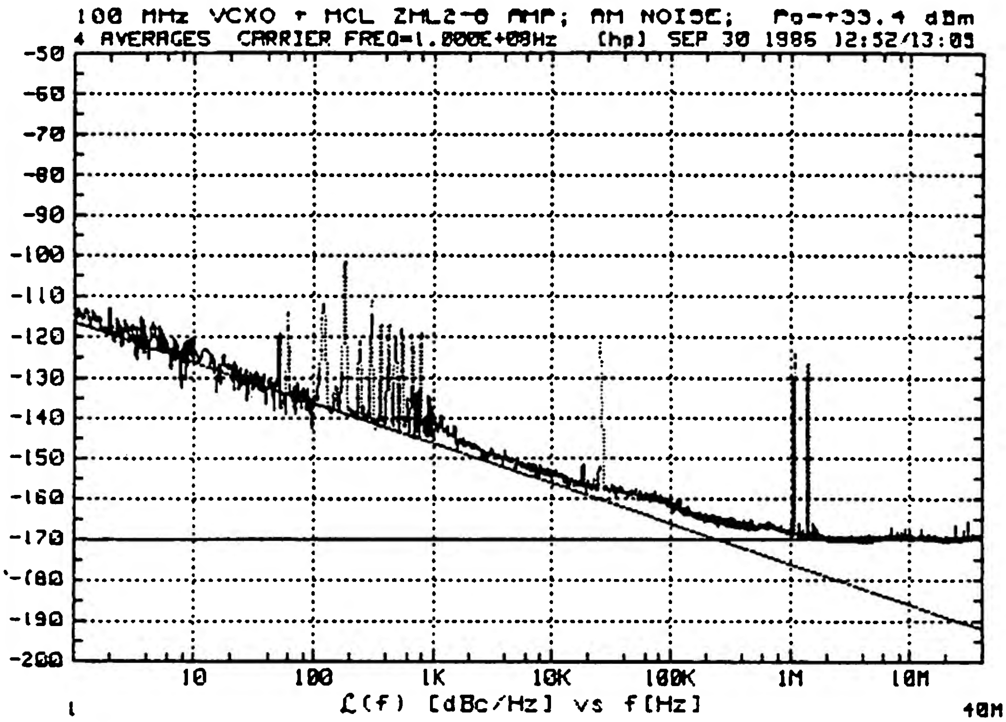


Figure 3-21. Source Without AM Modulation Measurement Results

```
            PRESENT SOURCE CHARACTERISTICS
CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±10 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 1.00000E+08 Hz
CARRIER FREQ = 1.00000E+08 Hz
INTERNAL MIXER IS EXTERNAL

            PRESENT MEASUREMENT CONSTANTS
VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.264 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz     LOOP BW3 = 0 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1       ATTEN2 = 1
```

Measurement of a Microwave Source without AM Modulation

This measurement uses the User Entry of Phase Detector Constant Calibration Method.

This is an example of a microwave device with a large frequency drift (during warmup) and no AM capability. The User-Entry Method was selected because:

- Its accuracy is not affected by device frequency drift. The customer required that the data be taken during the first 5 minutes of operation before the DUT's temperature stabilized. The single-sided-spur method requires that the calibration spur and the DUT frequency be separated by 1 Hz to 100 kHz during the calibration period for an HP 3048A without an RF spectrum analyzer. A source with poor frequency stability may drift outside the 100 kHz range of the FFT analyzer before the calibration data can be measured.
- No microwave test equipment was available at the time of the measurement except for the DUT and AM detector used for the noise measurement.

Note

It is very useful to observe the signal before noise data is measured. This helps to eliminate erroneous results caused by spurious oscillation or by DUTs operating improperly.

Initial Setup

In this example, the DUT is a Gunn-Diode oscillator operating at 10.525 GHz, with an output power of +11.5 dBm.

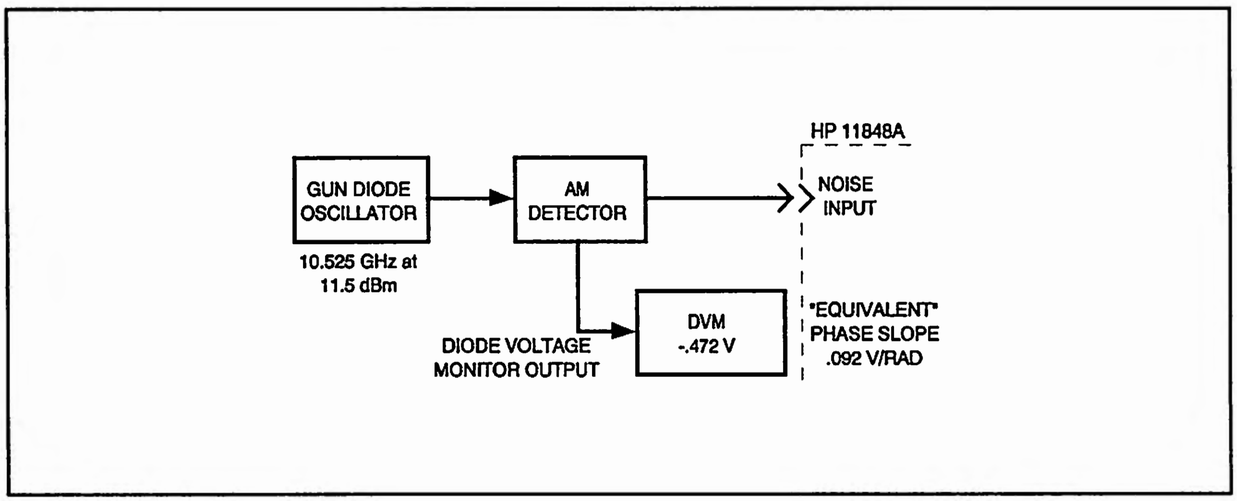


Figure 3-22. Microwave Source Without AM Setup

Conditions

This measurement was made under the following conditions.

- No additionally microwave test equipment other than the DUT and the AM detector is required.
- The DUT output power is estimated by measuring the diode detector output voltage with a digital voltmeter, ant then using the AM Sensitivity Graph (figure 3-5) to estimate the output power and equivalent phase slope.
- The quadrature meter should be at zero volts due to the blocking capacitor in the AM detector.

Results

The results of this test are shown in figure 3-23. The following is an analysis of those results.

- The Gunn diode oscillator, measured at 10.525 GHz, has a noise floor of at least −164 dBc/Hz at offsets greater than 100 kHz. The system noise floor may be limiting this measurement. The system floor can be estimated by comparing the equivalent phase slope to the Phase Detector Sensitivity Graph (figure 2-8). The 1 Hz intercept is at −125 dBc/Hz.
- This Gunn diode has very low AM noise, and makes an excellent signal source for residual two-port phase noise measurement.
- The spurs between 60 Hz and 1 kHz are due to 60 Hz line spurs, possibly induced by a ground loop between the DUT and the test system.
- Spurs in the 1 to 2 MHz region are produced by the Shared Resource Management System multiplexer connected to the test system controller. This RFI can be greatly reduced by replacing the multiplexer with the new HP 50961A SRM coax adapter.

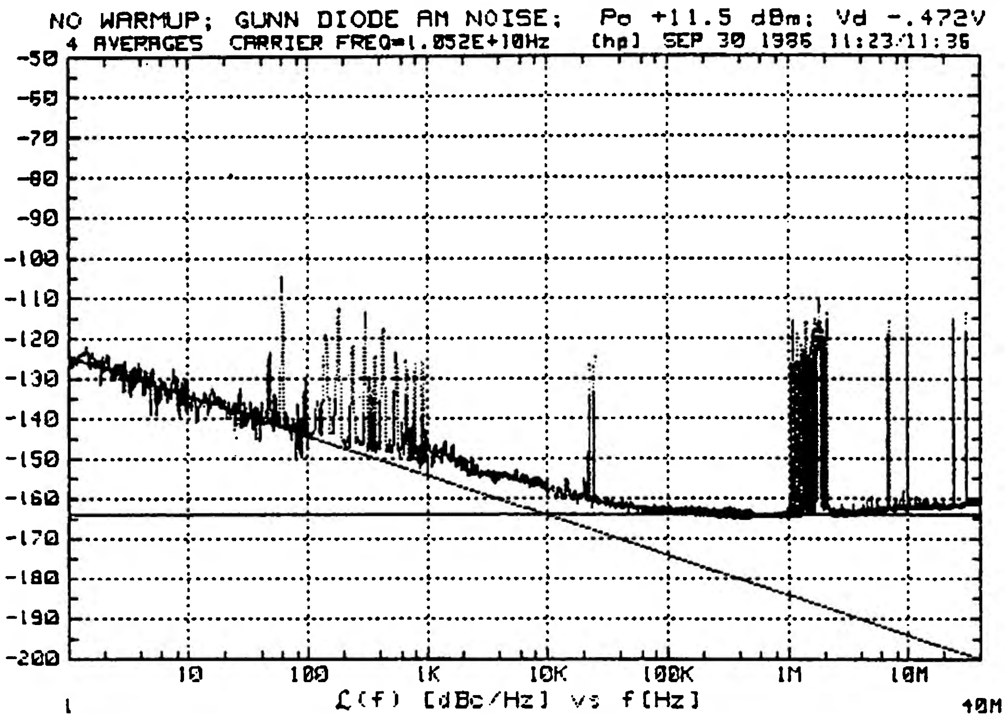


Figure 3-23. Microwave Source Without AM Measurement Results

```
            PRESENT SOURCE CHARACTERISTICS
CENTER VOLTAGE OF TUNING CURVE = 0 Volts
VOLTAGE TUNING RANGE = ±10 Volts
TOTAL FREQUENCY TUNING RANGE IS ≤1 MHz
PHASE DETECTOR INPUT FREQ = 1.05250E+10 Hz
CARRIER FREQ = 1.05250E+10 Hz
MIXER IS EXTERNAL

            PRESENT MEASUREMENT CONSTANTS
VCO SLOPE = 0 Hz/V
LOW NOISE AMPLIFIER IS IN
ACCURACY SPEC DEGRADATION = 0 dB
PHASE DETECTOR CONSTANT 0.092 VOLTS
DC OFFSET OF MIXER = 0 VOLTS
LOOP BW1 = 0 Hz     LOOP BW3 = 0 Hz
ZERO FREQUENCY IN LAG-LEAD = 1.59154943092 E+9 Hz
ATTEN1 = 1      ATTEN2 = 1
```

4

Baseband Noise Measurements

A baseband noise measurement measures the noise voltage of a device.

This measurement type uses the FFT Analyzer to directly measure the noise voltage out to 100 kHz. To extend the measurement range to 40 MHz, the HP 35601A or HP 11848A Phase Noise Interface is used to direct the noise voltage to the FFT analyzer or to the RF analyzer.

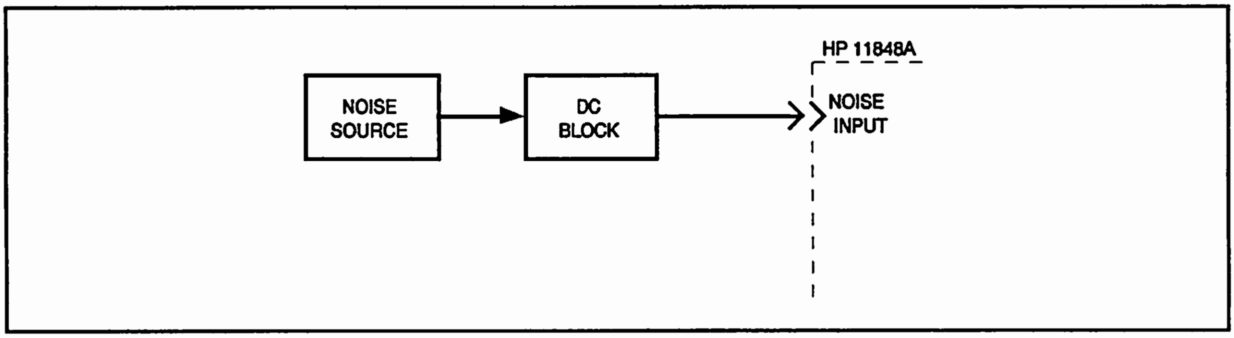


Figure 4-1. Baseband Noise Measurement Block Diagram

Measurement Considerations

- It may be necessary to use a dc blocking filter (HP 3048A K23, see appendix G) to enable the HP 3561A or HP 3585 to have a maximum dynamic range.
- The measurement will only measure out to 2 MHz offset if a Phase detector input frequency was set to less than 95 MHz.

Steps for Making Baseband Noise Measurements

1. At the Main Menu, select type of measurement.
2. Establish parameters.
   a. Source parameters
      - Enter phase detector input frequency
      - Enter carrier frequency
      - Select external detector
      - Select calibration option.

Input the gain of device-under-test, taking into account the effects of the dc blocking filter (6 dB loss for HP 3048 K23 DC Blocking Filter).

b. Measurement parameters

- Enter start and stop frequency of measurement data.

  For HP 3048: without RF analyzer, 0.1 to 100 kHz; with RF analyzer, 0.01 to 40 MHz for $f_c > 95$ MHz or 0.01 to 2 MHz for $f_c < 95$ MHz.

c. Plot parameters

- Select Graph type
- Enter Min and Max Y-Axis (dBv/Hz)
- Enter Min and Max X-Axis (Hz)
- Enter title

3. Measure

   a. Connect DUT to dc blocking filter and to phase noise system.

   b. Measure noise data

4. Interpret Results

   a. HP 3048 system plots data in dBv/Hz.

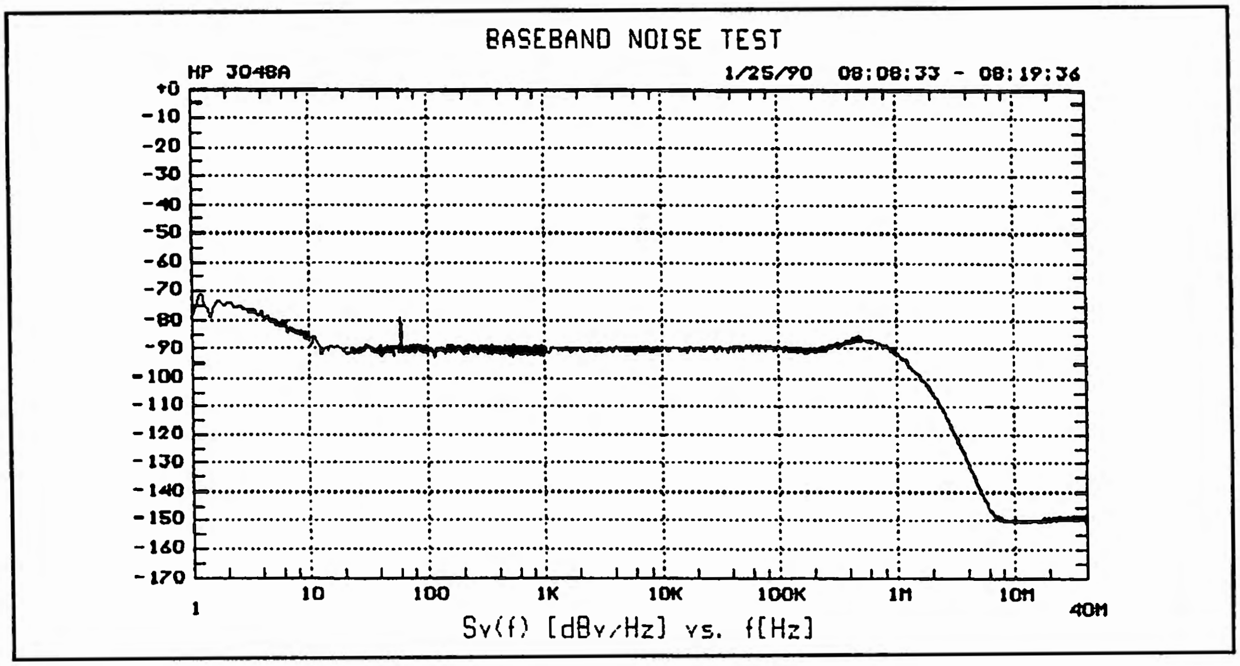
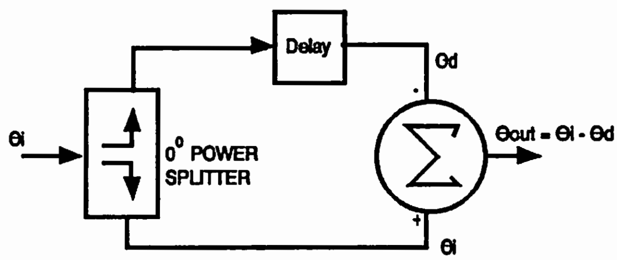


Figure 4-2. S_ν(dBv/Hz) versus f (Hz)

A

Calculation of Source Noise Attenuation

Source Noise Attenuation versus Carrier Offset and Time Delay



The input signal is expressed as a function of phase:

$$\theta_i = \underbrace{\omega_c t}_{carrier} + \underbrace{\Delta\theta \sin \omega_m t}_{phase\ modulation}$$

The delay path phase is:

$$\theta_d = \underbrace{\omega_c (t-\tau)}_{carrier\ delayed\ in\ time} + \underbrace{\Delta\theta \sin \omega_m (t-\tau)}_{phase\ modulation\ delayed\ in\ time}$$

Therefore, the ouput phase expression is:

$$\theta_{out} = \theta_i - \theta_d = \omega_c \tau + \Delta\theta[\sin \omega_m t - \sin \omega_m (t-\tau)]$$

From trigometric identities:

$$\sin \alpha - \sin \beta \equiv 2 \sin \left(\frac{\alpha-\beta}{2}\right) \cos \left(\frac{\alpha+\beta}{2}\right)$$

Which allows us to break the output signal into its component parts where:

$$\theta_{out} = \underbrace{\omega_c \tau}_{static\ phase\ error} + \underbrace{\Delta\ \theta}_{modulation\ amplitude} \underbrace{[2 \sin \frac{\omega_m \tau}{2} \cos \omega_m t]}_{sensitivity\ to\ modulation\ rate}$$

Because the measurement is made with the phase detector in quadrature, the static phase error is zero.

Also, for small values of $\omega_m \tau$, for example, $\omega_m \ll \pi$, the output phase expression can be reduced to:

$$\theta_{out} = \underbrace{\Delta\,\theta}_{modulation\ amplitude} \quad \underbrace{[2\sin\frac{\omega_m \tau}{2} \cos \omega_m t]}_{sensitivity\ to\ modulation\ rate}$$

Therefore the sensitivity to the input phase modulation, or the phase noise of a source is:

$$Sensitivity = 2\sin\frac{\omega_m \tau}{2}$$

or expressed in terms of carrier offset frequency:

$$\text{Sensitivity} = 2\sin \pi f_m \tau$$
$$\text{Where}: \ f_m = \text{offset frequency}$$
$$\tau = \text{time delay}$$

Finally, expressing the detector's sensitivity to source phase modulation in dB, the source's phase noise is attenuated by:

$$Att(dB) = 20\log|2\sin \pi f_m \tau|$$

It is important to note that the attenuation of the source's phase moise is a periodic function:

For offset frequencies of $\frac{n}{2\tau}$, when $n = 1, 3, 5, \ldots$, the input signal modulation and the delayed signal modulation add in phase. This results in the source's phase contribution having 6 dB of gain at the detector output.

For offset frequencies of $\frac{n}{2\tau}$, when $n = 0, 2, 4, 6, \ldots$, the input signal modulation and the delayed signal modulation are out-of-phase, which results in the total cancellation of the source's phase noise contribution at the detector output.

Finally, at approximately $\frac{1}{2\pi\tau}$ offset frequency, there is 0 dB of attenuation to the source noise contribution. From frequency discriminator theory, this corresponds to the offset frequency where the phase detector gain and the discriminator gain are equal.

A practical example of the determination of source noise attenuation can be found in figure 2-25.

B

Noise Figure Versus Dynamic Noise Figure

Noise figure is the ratio of the output noise of an amplifier referred back to the input divided by the thermal noise floor.

The noise figure of a linearly operating amplifier is defined by the expression:

$$NF = \frac{P_{out}}{KTBG}$$

$$\begin{aligned} \text{Where}: \ P_{out} &= \text{Noise power at amplifier output with the input terminated} \\ K &= \text{Boltzman's constant} \left(\frac{1.374 \times 10^{-23} \text{ joules}}{°K}\right) \\ T &= \text{Absolute temperature of amplifier } (°K) \\ B &= \text{Measurement bandwidth} \\ G &= \text{Amplifier gain} \end{aligned}$$

Expressed in dB:

$$NF\ (dB) = 10 \log \frac{P_{out}}{KTBG}$$

A noiseless amplifier would have a noise figure of 0 dB, that is, all the noise appearing at the output would be due to the noise generated by the input termination.

The noise power (P_n) of the termination is equal to:

$$P_n = KTB$$

$$\begin{aligned} \text{Where}: K &= \text{Boltzman's constant} \left(\frac{1.374 \times 10^{-23} \text{ joules}}{°K}\right) \\ T &= \text{Room temperature of } 290°K \\ B &= \text{Bandwidth of 1 Hz} \end{aligned}$$

Substituting in:

$$\begin{aligned} P_n &= \frac{1.374 \times 10^{-23} \text{ joules}}{°K} \times 290°K \times 1\text{Hz} \\ &= 3.985 \times 10^{-21} \text{watts} \\ P_n &= -174 \text{ dBm} \end{aligned}$$

The termination noise power consists of two equal contributors: AM noise and ΦM noise. This results in a termination noise floor of −177 dBc/Hz for both. or a dynamic rang of −177 dBc/Hz referred to 0 dBm.

We shall now calculate:

$$\begin{aligned}\mathcal{L}(\mathrm{f}) = {} & \Phi\mathrm{M\ noise\ floor\ (dBc/Hz)} \\ & + \mathrm{noise\ figure\ (dB)} \\ & + \mathrm{amplifier\ gain\ (dB)} \\ & - \mathrm{amplifier\ output\ power(dBm)}\end{aligned}$$

But:

$$amplifier\ output\ power\ (dBm) = input\ power\ (dBm) + gain\ (dB)$$

Therefore:

$$\mathcal{L}(\mathrm{f}) = \Phi\mathrm{M\ noise\ floor\ (dBc/Hz)} + \mathrm{noise\ figure\ (dB)} - \mathrm{amplifier\ input\ power\ (dBm)}$$

$$\text{That is}: \ \mathcal{L}(\mathrm{f}) = -177\ \mathrm{(dBc/Hz)} + \mathrm{NF\ (dB)} - \mathrm{P_i\ (dBm)}$$

Or the equation can be used to calculate noise figure,

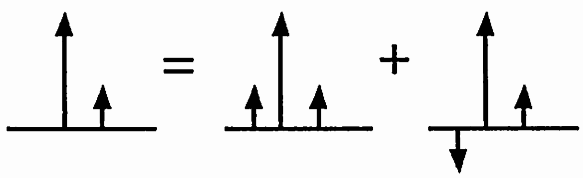
$$NF\ (dB) = \mathcal{L}(f)\ (dBc/Hz) + P_i\ (dBm) + 177\ (dBc/Hz)$$

This dynamic noise figure is measured under actual large signal conditions and may differ from the small signal noise figure. It includes the multiplicative noise produced by the non-linearities of the active device, in the presence of a large signal. This noise is negligible for very low input levels.

C

Single-Sided Spur

In this section we will show that a single-sided spur is equal to amplitude modulation plus phase modulation.



The instantaneous AM signal can be expressed as:

$$\Phi_{AM} = [A + f(t)]\cos\omega_c t$$

or

$$\Phi_{AM} = \underbrace{\mathrm{A}\cos\omega_c\mathrm{t}}_{\textit{carrier}} + \underbrace{\mathrm{F(t)}\cos\omega_c\mathrm{t}}_{\textit{modulation sidebands}}$$

Where f(t) is the modulation information.

Let $f(t) = a\cos\omega_m t$ so the modulation is a signal of amplitude and at a modulation frequency of

Substituting in f(t):

$$\Phi_{AM} = A\cos\omega_c t + a\cos\omega_m t \bullet \cos\omega_c t$$

From trigonmetric identities:

$$\cos\alpha\cos\beta = \frac{1}{2}\cos(\alpha - \beta) + \frac{1}{2}\cos(\alpha + \beta)$$

or

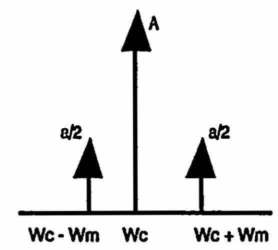
$$\Phi_{AM} = A\cos\omega_c t + \frac{a}{2}\cos(\omega_m - \omega_c)t + \frac{a}{2}\cos(\omega_m + \omega_c)t$$

also from trigonometric identities:

$$\cos\alpha \equiv \cos(-\alpha)$$

so:

$$\Phi_{AM} = \underbrace{\mathrm{A}\cos\omega_c \mathrm{t}}_{carrier} + \underbrace{\frac{\mathrm{a}}{2}\cos(\omega_\mathrm{c} - \omega_\mathrm{m})\mathrm{t}}_{lower\ sideband} + \underbrace{\frac{\mathrm{a}}{2}\cos(\omega_\mathrm{c} + \omega_\mathrm{m})\mathrm{t}}_{upper\ sideband}$$



The instantaneous ΦM signal can be expressed as:

$$\Phi_{\Phi M} = A\cos[\omega_c t + k_p \Phi(t)]$$

Where k_p is a constant and is small, such that the total phase deviation is less than 0.1 radian and the small angle criterion applies. The small angle criterion simply stated is that the majority of the modulation power is contained in the first pair of sidebands.

Where Φ(t) is the modulation information, let $\Phi(t) = \beta \sin\omega_m t$.

Substituting in Φ(t):

$$\Phi_{\Phi M} = A\cos[\omega_c t + k_p\beta\sin\omega_m t]$$

From trigonometric identities:

$$\cos(\alpha+\beta) \equiv \cos\alpha\cos\beta - \sin\alpha\sin\beta$$

or

$$\Phi_{\Phi M} = A\cos\omega_c t \bullet \cos[k_p\beta\sin\omega_m t] - A\sin\omega_c t \bullet \sin[k_p\beta\sin\omega_m t]$$

Because the small angle criterion applies, and the total amount of deviation is small, $k_p\beta$ must also be small.

Also, the cosine of a small number is approximately 1. The sine of a small number is approximately the small number.

With this in mind:

$$\cos[k_p\beta \sin\omega_m t] \rightarrow 1$$

and

$$\sin[k_p\beta \sin\omega_m t] \rightarrow k_p\beta \sin\omega_m t$$

or

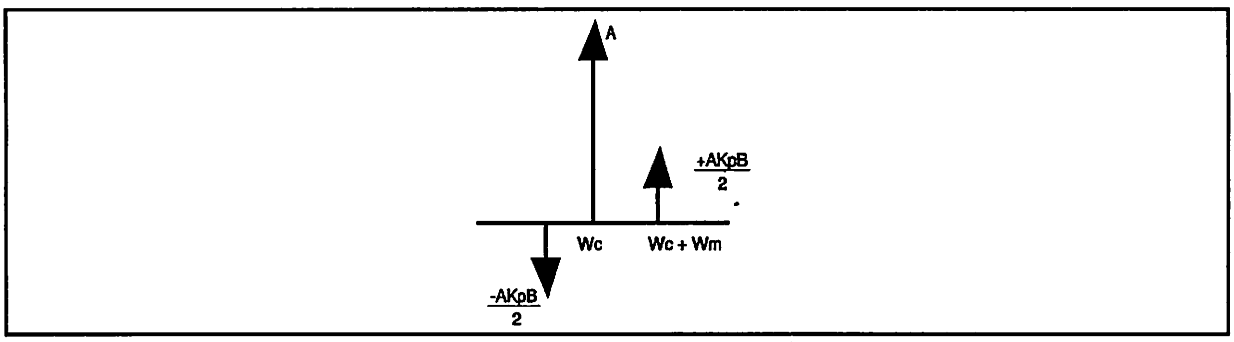
$$\Phi_{\Phi M} = A\cos\omega_c t - Ak_p\beta \sin\omega_c t \bullet \sin\omega_m t$$

From trigonometric identities:

$$\sin\alpha \sin\beta \equiv \frac{1}{2}\cos(\alpha - \beta) - \frac{1}{2}\cos(\alpha + \beta)$$

Substituting in:

$$\Phi_{pm} = \underbrace{\mathrm{A}\cos\omega_c \mathrm{t}}_{carrier} - \underbrace{\frac{\mathrm{Ak_p}\beta}{2}\cos(\omega_c - \omega_\mathrm{m})\mathrm{t}}_{lower\ sideband} + \underbrace{\frac{\mathrm{Ak_p}\beta}{2}\cos(\omega_c + \omega_\mathrm{m})\mathrm{t}}_{upper\ sideband}$$



If we let:

$$\frac{Ak_p\beta}{2}\cos(\omega_c - \omega_m)t = \frac{A}{2}\cos(\omega_c - \omega_m)t$$

then

$$\frac{Ak_p\beta}{2} = \frac{A}{2}$$

or

$$\beta = \frac{a}{Ak_p}$$

Substituting $\frac{a}{2}$ into $\Phi_{\Phi M}$

$$\Phi_{\Phi M} = A\cos\omega_c t - \frac{a}{2}\cos(\omega_c - \omega_m)t + \frac{a}{2}\cos(\omega_c - \omega_m)t$$

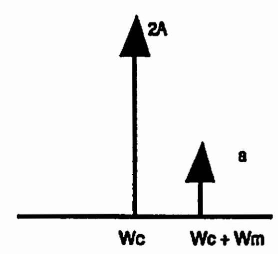
adding the AM signal and ΦM signal

$$\Phi_{AM} + \Phi_{\Phi M} = 2A\cos\omega_c t + \frac{a}{2}\cos(\omega_c + \omega_m)t + \frac{a}{2}\cos(\omega_c + \omega_m)t$$

or

$$\Phi_{AM} + \Phi_{\Phi M} = 2A\cos\omega_c t + a\cos(\omega_c + \omega_m)t$$

which can be written graphically:



This graph shows that half the power is in the amplitude modulation component and the other half is in the phase modulation component. It also shows that the height of the modulation sidebands are 6 dB lower than the spur height.

D

Carrier-to-Sideband Modulation Equations

AM Carrier-to-Modulation Sideband Ratio

$$\frac{C}{sb} = 20\log\left(\frac{\%AM}{100}\right) - 6\ dB$$

where %AM is the percentage depth of modulation.

Example

What is the $\frac{C}{sb}$ for 2% AM?

$$\begin{aligned}\frac{\mathrm{C}}{\mathrm{sb}} &= 20\log\left(\frac{2}{100}\right) - 6\ \mathrm{dB} \\ &= 20\log(0.02) - 6\ \mathrm{dB} \\ &= -33.98\ \mathrm{dBc} - 6\ \mathrm{dB} \\ \frac{\mathrm{C}}{\mathrm{sb}} &= -39.98\ \mathrm{dBc}\end{aligned}$$

ΦM Carrier-to-Modulation Sideband Ratio

For small-angle phase-modulation, where $\beta < 0.1$ radian:

$$\frac{C}{sb} = 20\log\left(\frac{\beta}{2}\right)$$

where β is the peak phase deviation in radians.

Example

What is the $\frac{C}{sb}$ for 0.02 radians of phase deviation?

$$\begin{aligned}\frac{\mathrm{C}}{\mathrm{sb}} &= 20\log\left(\frac{0.02}{2}\right) \\ &= 20\log(0.01) \\ \frac{\mathrm{C}}{\mathrm{sb}} &= -40\ \mathrm{dBc}\end{aligned}$$

FM Carrier-to-Modulation Sideband Ratio

For small angle frequency modulation, where:

$$\beta < 0.1\text{radian}$$

$$\text{and } \beta = \frac{\text{peak frequency deviation}}{\text{modulation rate}}$$

$$\text{or } \beta = \frac{\Delta \text{f}}{2\text{f}_\text{m}}$$

$$\frac{\text{C}}{\text{sb}} = 20\log\left(\frac{\Delta\text{f}}{2\text{f}_\text{m}}\right)$$

Example

What is the $\frac{C}{sb}$ for 2 kHz of deviation at a 100 kHz rate?

$$\frac{\text{C}}{\text{sb}} = 20\log\left(\frac{2\text{kHz}}{2(100\text{kHz})}\right)$$

$$= 20\log(0.01)$$

$$\frac{\text{C}}{\text{sb}} = -40\text{dBc}$$

E

Common Equipment Used in RF Noise Measurements

Note Not all equipment listed is fully supported by HP 3048A software.

Sources

HP 8642A	100 kHz to 1057.5 MHz synthesizer
HP 8642B	100 kHz to 2115 MHz synthesizer
HP 8662A	10 kHz to 1280 MHz synthesizer
HP 8663A	10 kHz to 2560 MHz synthesizer
HP 8656B	100 kHz to 990 MHz synthesizer
HP 8657A	100 kHz to 1040 MHz synthesizer
HP 8657B	100 kHz to 2060 MHz synthesizer
HP 8640B	5 MHz to 1100 MHz signal generator

Spectrum Analyzers

HP 3585	20 Hz to 40 MHz
HP 71000	Series modular spectrum analyzers
HP 8568	20 Hz to 1500 MHz
HP 8566	100 Hz to 22 GHz
HP 8558	100 kHz to 1500 MHz
HP 8559	10 MHz to 21 GHz

Modulation Analyzers

HP 8901A/B	150 kHz to 1300 MHz
HP 8902A	150 kHz to 1300 MHz

Power Meters

HP 436A with HP 8482A sensor	100 kHz to 4.2 GHz, input range −30 to +20 dBm
HP 437A with HP 8482A sensor	100 kHz to 4.2 GHz, input range −30 to +20 dBm
HP 438 with HP 8482A sensor	100 kHz to 4.2 GHz, input range −30 to +20 dBm

Diode Detectors

HP 33330C	10 MHz to 26 GHz P_{max} = +23 dBm
HP 33330D	0.01 to 33 GHz, P_{max} = +23 dBm
HP 8474C	0.01 to 33 GHz, P_{max} = +23 dBm

AM Detector

For HP 3048A K21 specifications, see appendix G.

HP 3048A K21	1 Hz to 40 MHz, ±5 V dc maximum.

DC Block

For HP 3048A K23 specifications, see appendix G.

HP 3048A K23	5 Hz to 40 MHz, ±30 V dc.

Attenuators

HP 8491	Series of fixed attenuators
HP 8493	Series of fixed attenuators
HP 8494	1 dB step attenator
HP 8495	10 dB step attenuator

Amplifiers

Specifications for the HP 3048A K22 can be found in appendix G.

HP 3048A K22 Dual Amplifier	9 dB gain each, 5 MHz to 1500 MHz, NF < 7.5 dB, output power > +15 dBm
HP 8447A preamp	26 dB gain, 0.1 MHz to 400 MHz, NF < 5 dB, output power > 6 dBm
HP 8447D preamp	26 dB gain, 0.1 MHz to 1300 MHz, NF < 8.5 dB, output power > 7 dBm
HP 8447E preamp	22 dB gain, 0.1 MHz to 1300 MHz, NF < 11 aB, output power > 15 dBm
MCL ZHL-2-8 power amplifier	27 dB gain, 10 MHz to 10 MHz, NF < 10 dB, output power > +29 dBm
ANZAC AMC-123	10 dB gain, 5 MHz to 500 MHz, NF < 3.5 dB, output power > +16 dBm

Phase Shifters

HP 11609 Option K08 mechanical line stretcher	2.5 to 3.4 nanosecond variable delay.
ARRA	Line stretchers and phase shifters

0° Power Splitters

MCL ZFSC-2-2500	10 MHz to 2.5 GHz, P_{max} 1 watt
MCL ZFSC-2-5	10 MHz to 1500 MHz, P_{max} 1 watt
MCL ZAPD-21	500 MHz to 2 GHz, P_{max} 10 watts
NARDA 4456-2	2 GHz to 18 GHz

Couplers

MCL ZFDC-20-5	0.1 to 2000 MHz 20 dB coupler, P_{max} 2 watts
MCL ZFDC-10-2	10 MHz to 1000 MHz coupler, P_{max} 1.5 watts
NARDA 4227-16	1.7 to 26.5 GHz coupler

Instrument Suppliers

Anzac Division of Adams/Russel
80 Cambridge Street
Burlington, MA 01803-0964
(617) 273-3333

ARRA
15 Harold Court
Bayshore, Long Island, NY 11706
(516) 231-8400

MCL (Mini-Circuits)
P.O. Box 166
Brooklyn, NY 11235
(212) 934-4500

F

Components of the Phase Noise Accessory Kits

Components for Residual Measurements

Description	Part Number	Qty.	A	B	C
Dual RF Amplifiers	HP 3048A Option K22	2	*		*
Refurbished Phase Shifter	HP 11609A Option K08	1	*	*	*
Power Splitter (10 MHz to 2.5 GHz)	0955-0504	1	*		*
Power Splitter (2 GHz to 18 GHz)	0955-0517	1		*	*
20 dB Directional Coupler (0.1 GHz to 2 GHz)	0955- 0516	2	*		*
Directional Coupler (1.7 GHz to 18 GHz)	0955- 0125	2		*	*

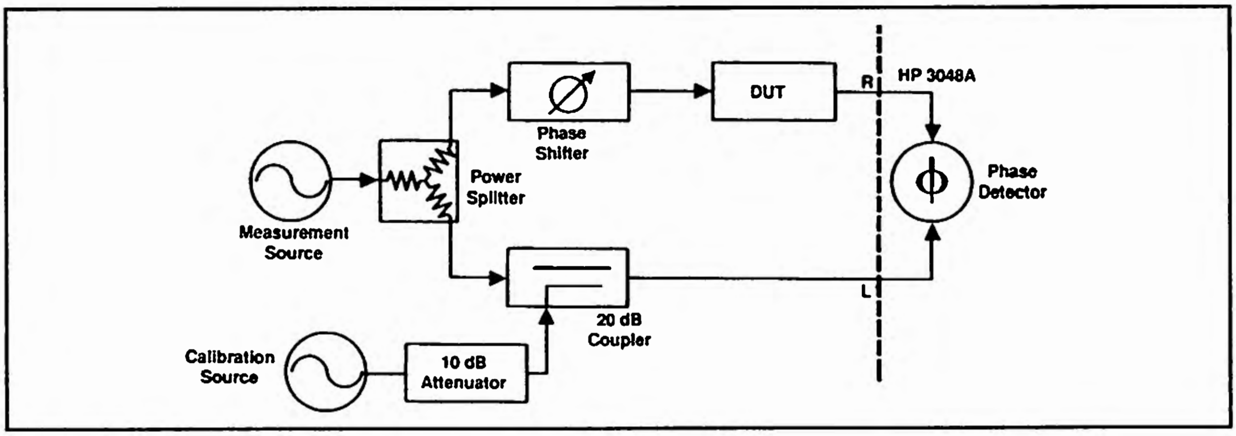


Figure F-1. Single-Sided Spur Calibration Setup for Residual Measurement

Description	Part Number	Qty.	A	B	C
AM Detector Filter	HP 3048A Option K21	1	*	*	*
AM Detector	HP 33334C	1	*	*	*

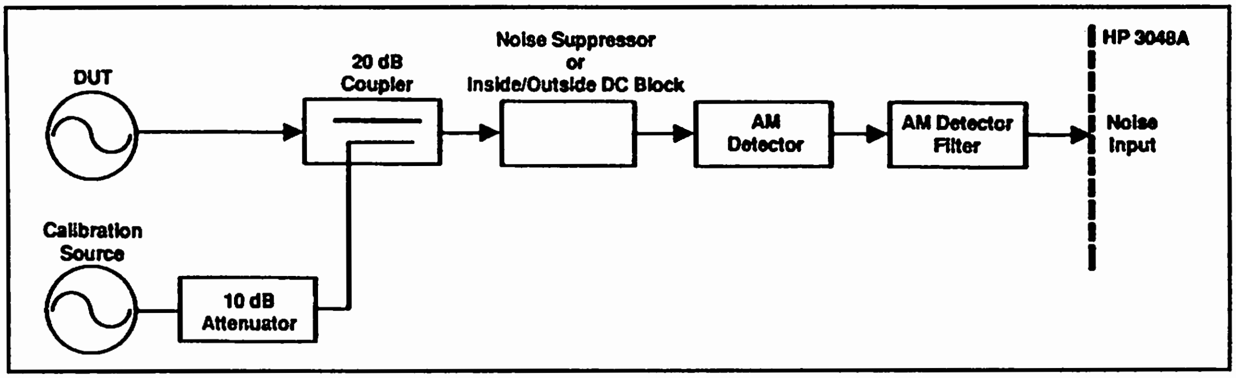


Figure F-2. Single-sided Spur Calibration Setup for AM Noise Measurement

This setup is for measuring of a device without amplitude modulation.

Miscellaneous Components

Description	Part Number	Qty.	A	B	C
DC Blocking Filter[1]	HP 3048A Option K23	1	*		*
Power Supply 20 volts, 0.5 amperes)[2]	HP 6236B	1	*	*	*
Delay Line (50 nanoseconds)[3]	5021-9670	2	*	*	*
Noise Suppressor[4]	0960-0798	2	*		*
Inside/Outside DC Block[5]	0960-0797	2		*	*
3 dB Attenuators	HP 8493B Option 003	3	*	*	*
6 dB Attenuators	HP 8493B Option 006	3	*	*	*
10 dB Attenuators	HP 8493B Option 010	3	*	*	*
20 dB Attenuators	HP 8493B Option 020	1	*	*	*
50 Ω Termination	HP 909D	2	*	*	*
Test Leads	HP 11002A	1	*	*	*
Wrench	8710-1765	1	*	*	*

[1] DC blocking filter is used to measure the noise of power supplies.
[2] Power supply is required for the HP 3048A Option K22 amplifier.
[3] Delay line is used in discriminator measurements.
[4] Noise suppressor is used to break ground loops.
[5] Inside/outside DC block is used to break ground loops.

Cables

Description	Part Number	Qty.	A	B	C
12-inch SMA Cable	8120-5386	3	*	*	*
12-inch Right Angle SMA Cable	8120-5387	2	*	*	*
24-inch SMA Cable	8120-5389	3	*	*	*
24-inch Right Angle SMA Cable	8120-5388	2	*	*	*
72-inch SMA Cable	8120-5390	2		*	*
48-inch BNC Cable	8120-1840	2	*		*

Adapters

Description	Part Number	Qty.	A	B	C
SMA(f) to SMA(f)	1250-1158	3	*	*	*
SMA(f) to BNC(m)	1250-2015	2	*		*
APC7 to SMA(f)	HP 11534A	2	*	*	*
SMA(f) to N(m)	1250-1250	5	*	*	*
SMA(m) to N(f)	1250-1562	2	*		*
SMA(m) to BNC(f)	1250-1200	2	*		*
BNC(f) to N(m)	1250-0780	2	*		*
SMA(m) to SMA(m)	1250-1159	2	*	*	*
N(f) to APC3.5(f)	1250-1745	2	*		*

Case Layout

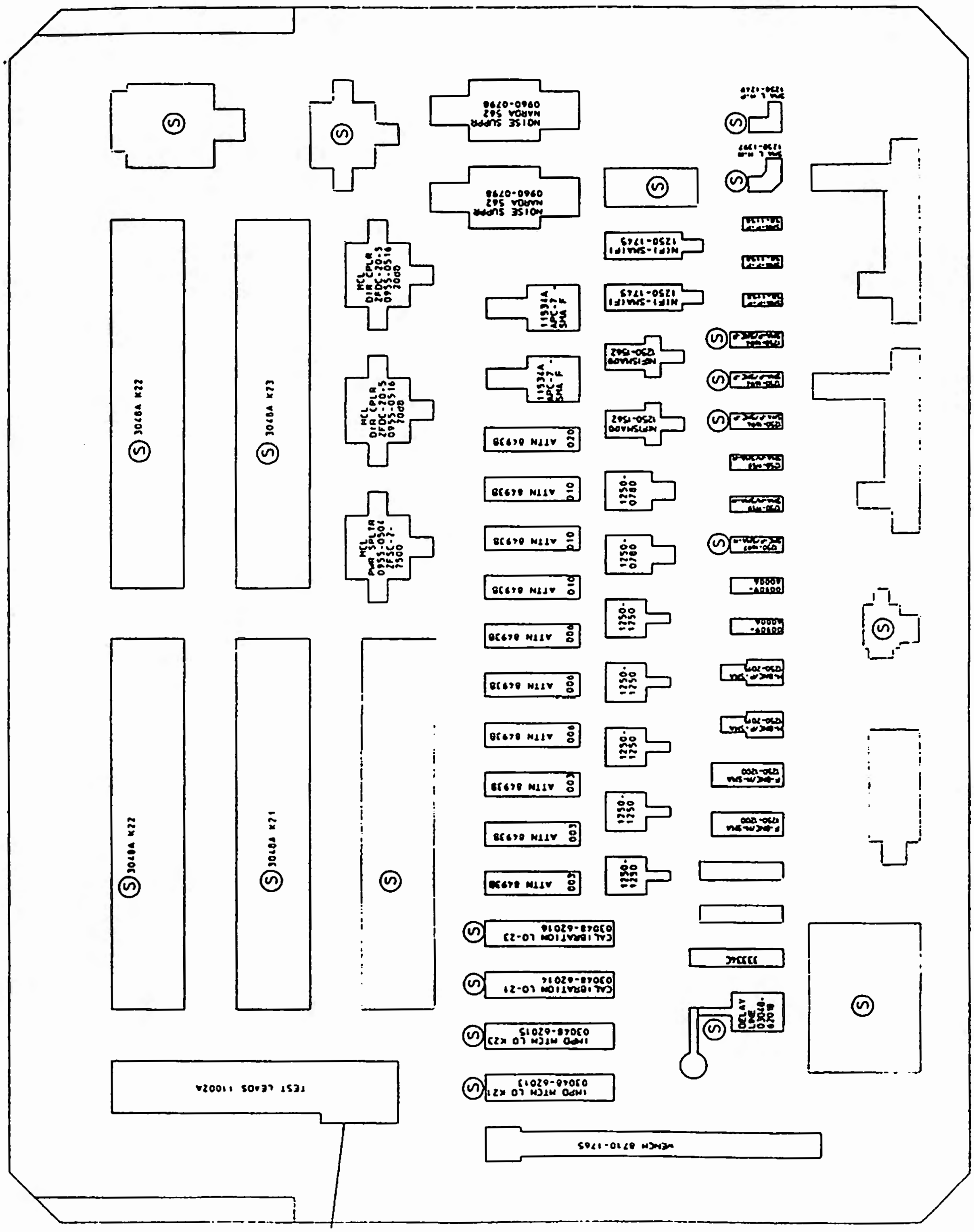


Figure F-3. HP 11826A Small Case Layout

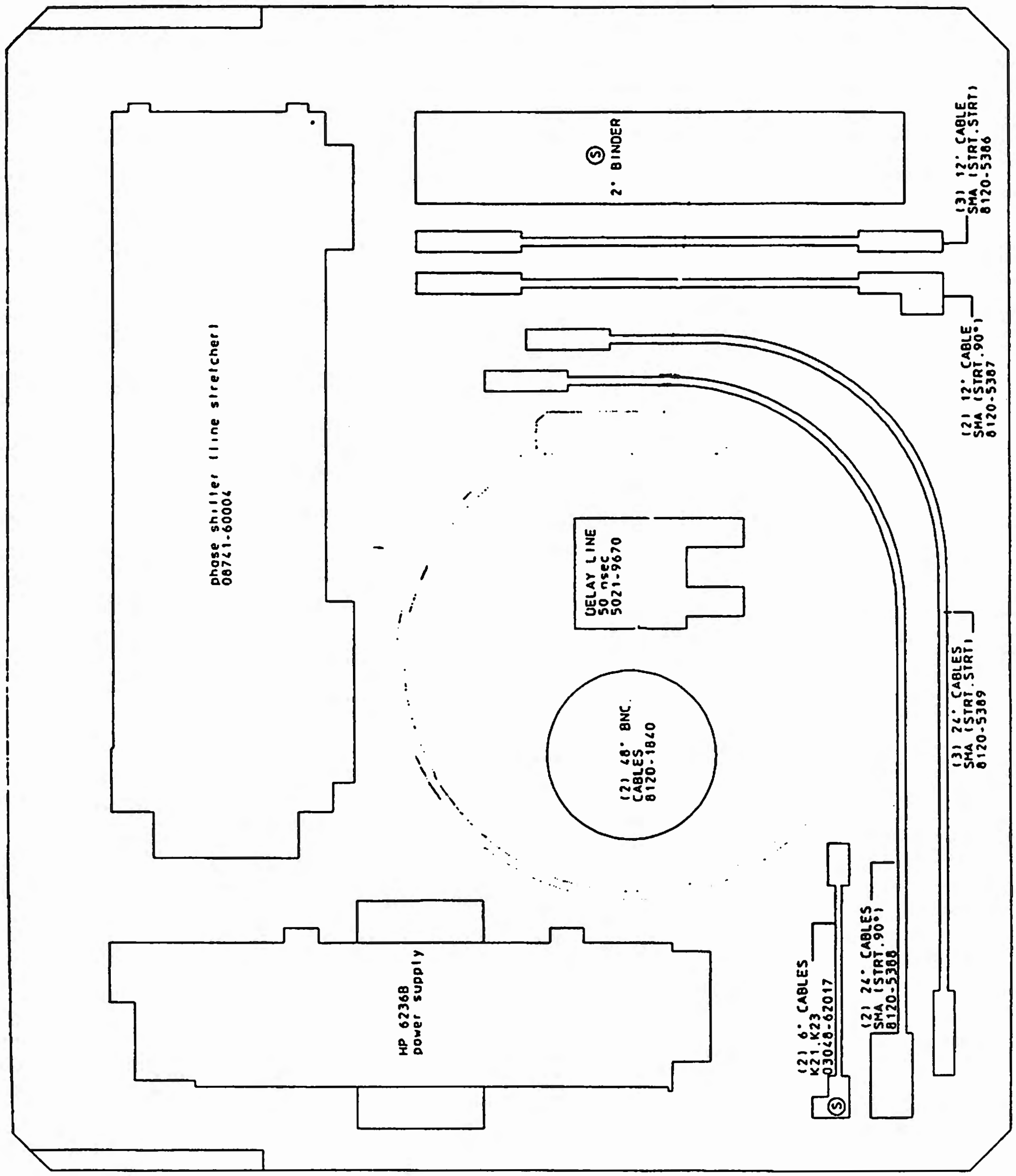


Figure F-4. HP 11826A Large Case Layout

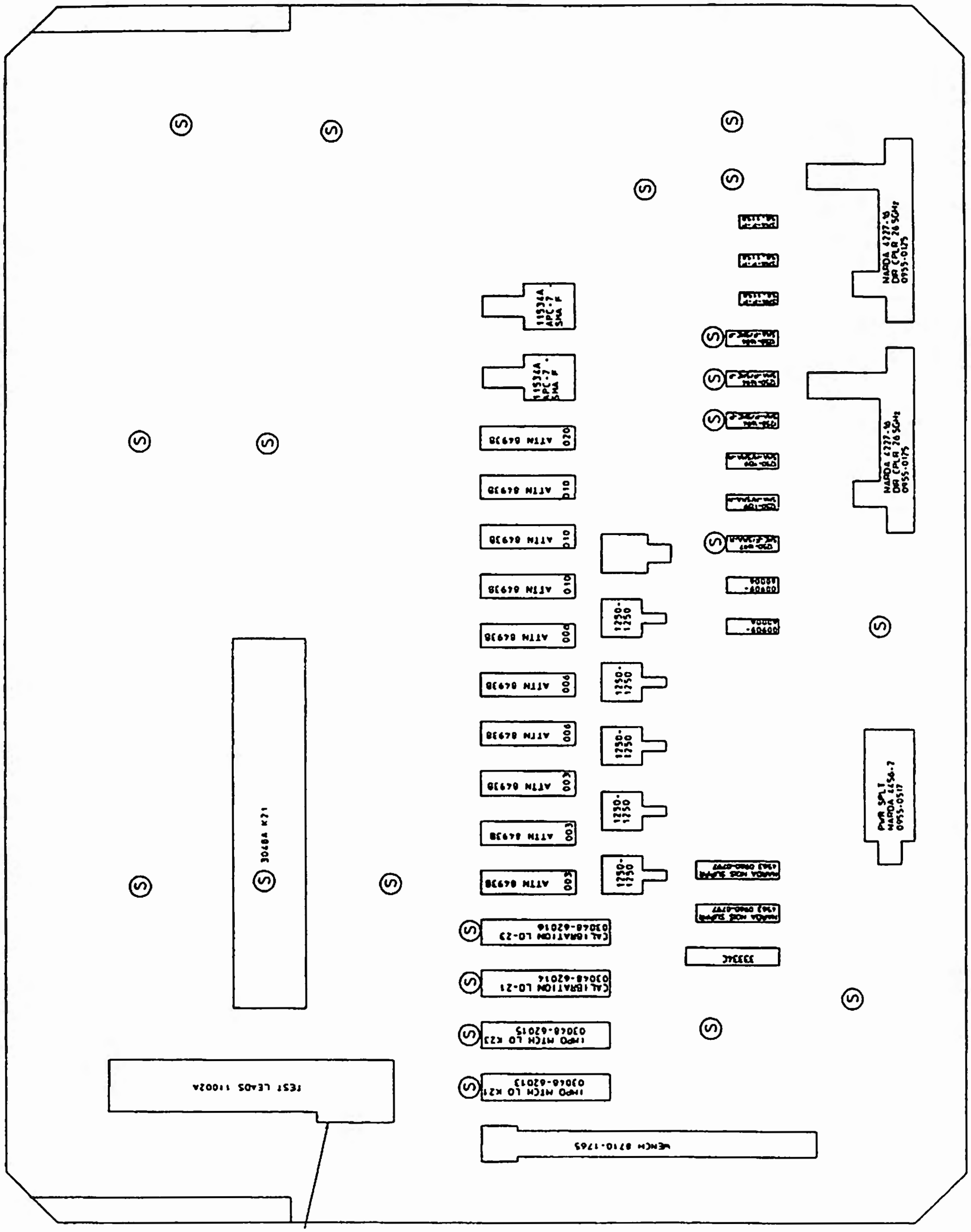


Figure F-5. HP 11826B Small Case Layout

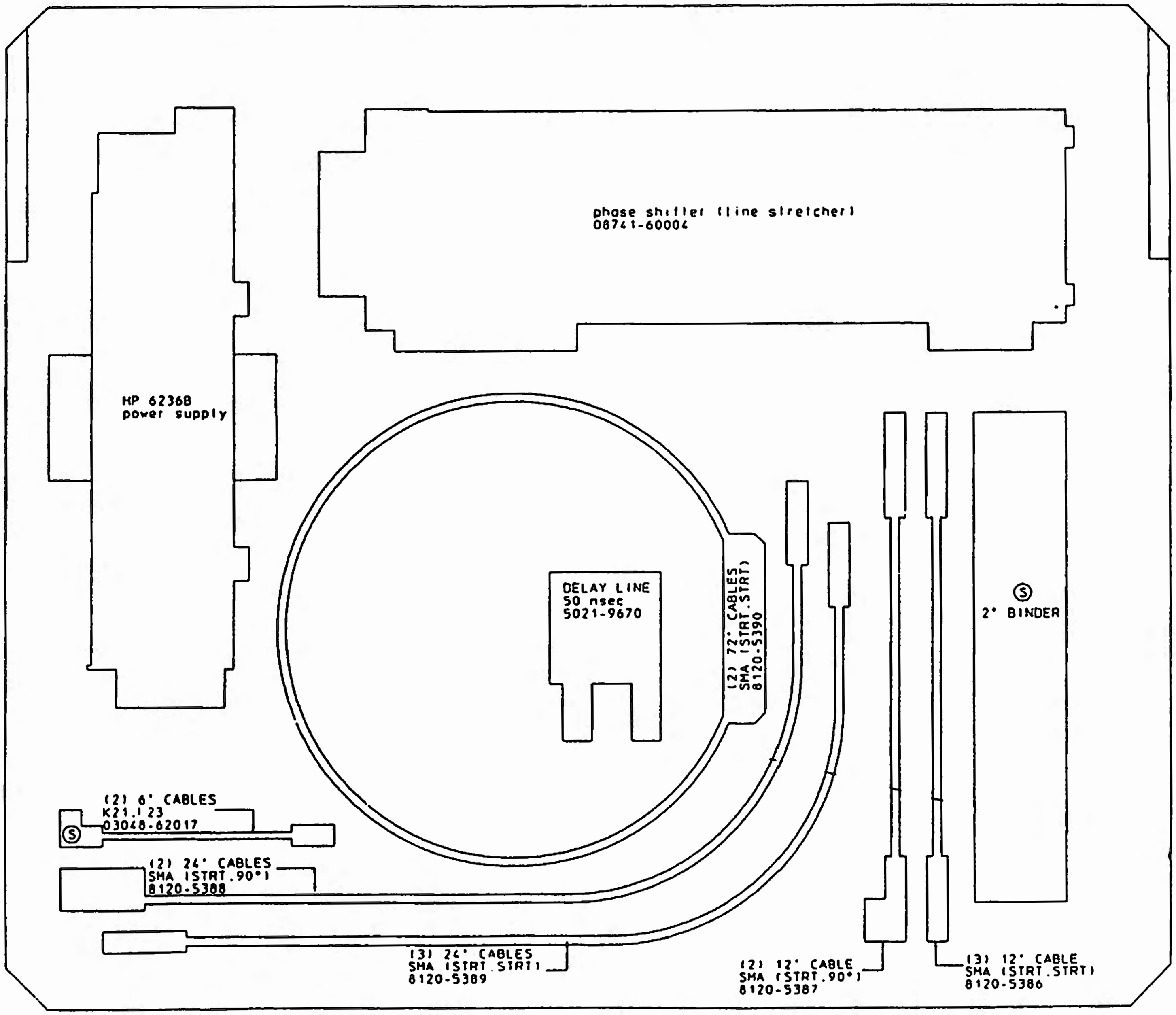


Figure F-6. HP 11826B Large Case Layout

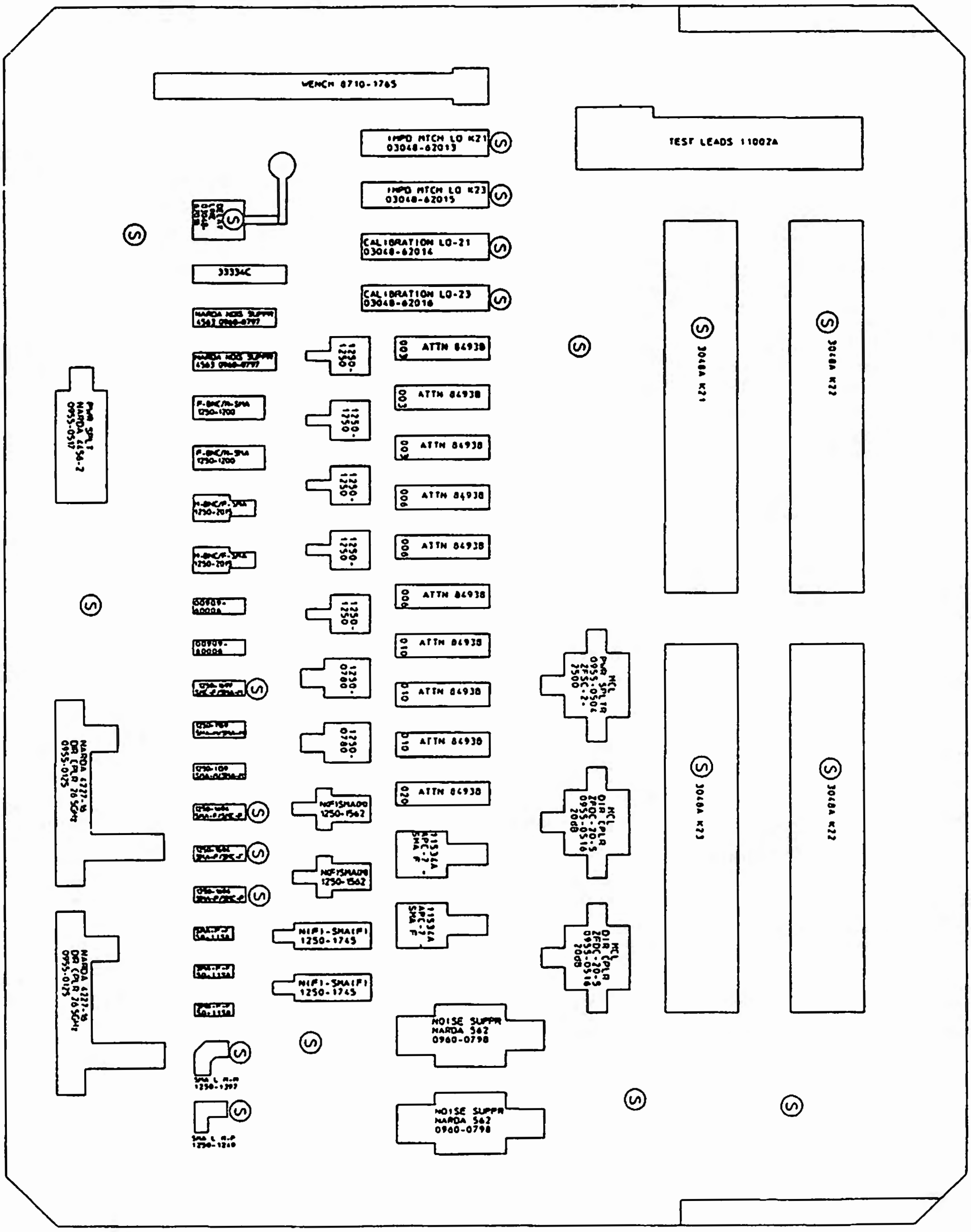


Figure F-7. HP 11826C Small Case Layout

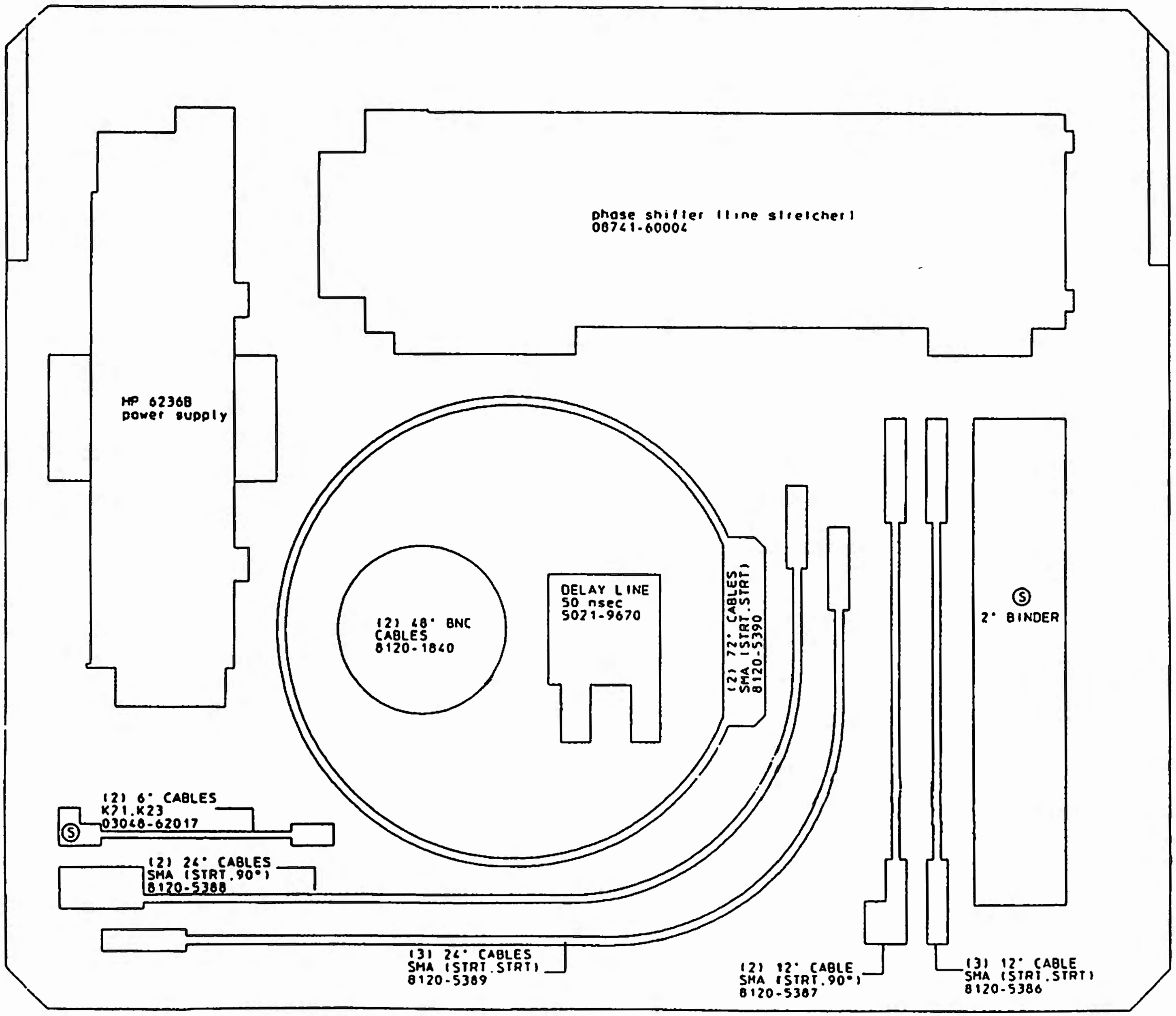
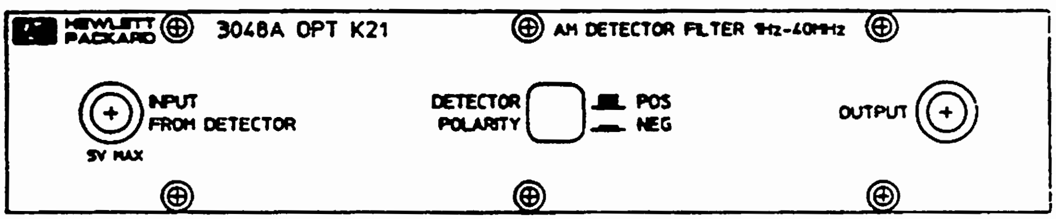


Figure F-8. HP 11826C Large Case Layout

G

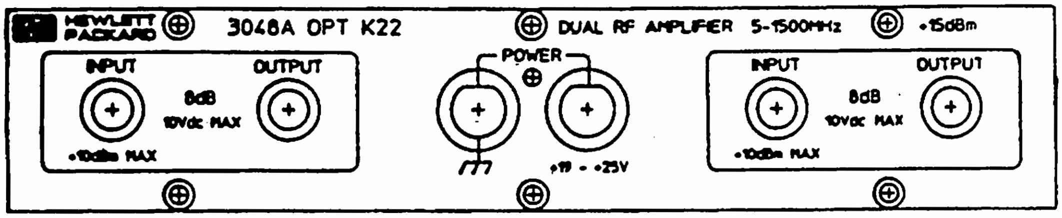
HP 3048A Options K21, K22, and K23 Specifications

HP 3048A Option K21 AM Detector Filter



Frequency Range:	1 Hz to 40 MHz
Insertion Loss:	< 1 dB
Flatness	< 1 dB
Input Level	±5 V dc maximum
(10.0 Ω fusable resistor)	

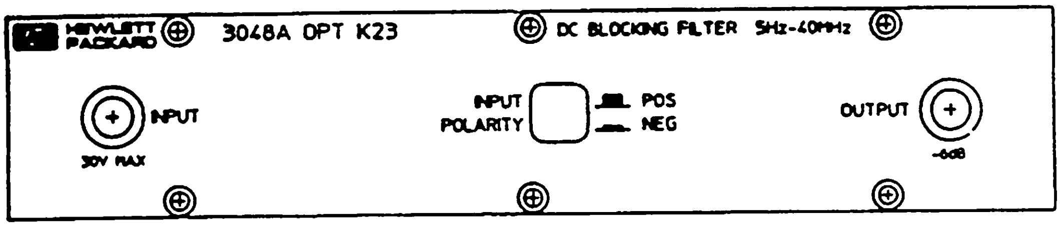
HP 3048A Option K22 AM Dual RF Amplifier



Typical Characteristics

Frequency Range:	*5 Hz to 1500 MHz*
Insertion Loss:	*< 9 dB ±1.5 dB*
Noise Figure:	*< 7.5 dB (typ. 6 dB), 50 MHz to 1500 MHz*
Dynamic Range:	*For f_c of 50 MHz to 1500 MHz: Meets HP 3048A system phase noise specifications.*
	For f_c of 5 MHz to 50 MHz @ >10 kHz offset: 10 dB degradation (typ. −170 dBc/Hz).

HP 3048A Option K23 AM DC Blocking Filter



Frequency Range:	5 Hz to 40 MHz
Input Impedance:	100 ohms
Insertion Loss:	6 dB
Flatness	< 1 dB
Input Level	±30 V dc maximum
(51.1 Ω fusable resistor)	