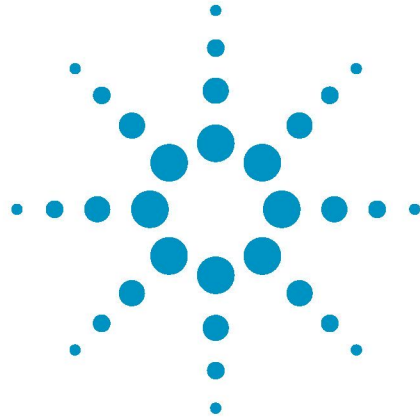


Phase Noise Characterization of Microwave Oscillators

Phase Detector Method

Product Note 11729B-1



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Product
Note
11729B-1

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Table Of Contents	Page
Chapter 1: Introduction	3
Chapter 2: Phase Noise and its Effect on Microwave Systems	4
What is Phase Noise?	4
Residual and Absolute Noise	5
Why is Phase Noise Important?	5
Chapter 3. SSB Phase Noise Measurements — Phase Detector Method	7
Common Measurement Techniques	7
Direct Measurement	7
Heterodyne Frequency Measurement	7
Carrier Removal/Demodulation	7
Measurement with a frequency discriminator	7
Measurement with a phase detector	8
The Phase Detector Method	8
Basic Theory	8
Importance of the Reference Source	9
Importance of Quadrature	10
Effect of Phase Lock Loop	10
General Measurement Procedure	11
Set-Up and Calibration	11
Lock/Quadrature	11
Measure	12
Corrections	12
Chapter 4: 11729B Theory of Operation and Measurement Considerations	13
Multiplier Chain	13
Demodulator and Baseband Signal Processing Section	14
First Downconversion	14
Phase Detector	14
Baseband Signal Processing	14
Phase Lock Loop/Quadrature Section	15
Loop Hold	15
Phase Locking via the 8662A Reference Oscillator	16
Loop Holding Range (LHR)	16
Loop Bandwidth (LBW)	16
Loop Bandwidth Limitations	16
Phase Locking via the 8662A dc FM	17
Loop Holding Range (LHR)	17
Loop Bandwidth (LBW)	17
Phase Locking using a 3rd Source	17
Effect of Phase Locking Method on System Noise Floor	18
Loop Capture	18
Loop Tune	19
Loop Characterization	19

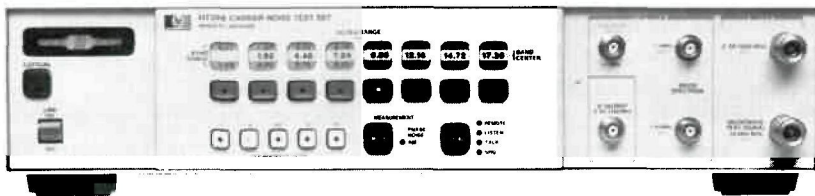
Chapter 5: Making Phase Noise Measurements with the 11729B	21
Properties of the Source Under Test	21
Considerations in Phase Locking	22
Phase Noise Measurement Procedures	22
System Setup	23
System Calibration	23
Lock	24
Measure	25
Apply Corrections	25
Measurement of Loop BW and Loop Noise Suppression	26
Chapter 6: Considerations in System Accuracy	27
1. Spectrum Analyzer Accuracy	27
2. Attenuator Accuracy of Calibration	27
3. Phase Detector Flatness	28
4. 11729B Baseband Signal Processing Section Flatness	28
5. Mixer dc Offset	29
6. Absolute Noise Accuracy due to System Noise	29
7. Accuracy of Measurement of Loop Noise Suppression	29
Overall Accuracy	30
Accuracy without Error Correction	31
Accuracy with Error Correction	31
Chapter 7: Using the 11729B with the HP 3047A for Automatic Measurements	32
System Set-Up	32
System Operation	32
Overall System Accuracy	33
Appendix A: Phase Detector Method Derivations and Advanced System Considerations	34
Appendix B: Phase Lock Loops: Advanced Considerations	37
Appendix C: Calculating System Phase Noise	41
Appendix D: Measurements with 11729B/8662A/3582A/8566A	43
Appendix E: 11729B HP-IB Codes	47
Appendix F: References	48

As the performance of microwave radar and communication systems advances, certain system parameters take on increased importance. One of these parameters that must be measured is the spectral purity of microwave signal sources.

In the past, many techniques for measuring spectral purity have used complex, dedicated instrumentation, often cumbersome in both size and operation. The broadening focus on spectral purity has created a need for measurement techniques that provide the high performance necessary for R&D requirements, and that can be automated for production environments. Also, service applications require a versatile system with a broad frequency and performance range.

The Hewlett-Packard 11729B Carrier Noise Test Set is a key element of a system that provides convenient manual or automatic phase noise and amplitude noise measurements. With appropriate companion instrumentation, noise measurements can be made on a broad range of sources, 5 MHz to 18 GHz.

This product note discusses phase noise and its effects on modern microwave systems in Chapter 2. Chapter 3 describes a sensitive technique for measuring the phase noise of sources. The implementation of this technique in the 11729B is shown in Chapter 4. Chapter 5 outlines the measurement steps needed to make a phase noise measurement, and the resultant measurement accuracy is derived in Chapter 6. Finally, Chapter 7 presents a way to automate the 11729B in an automatic phase noise measurement system.



2 Phase Noise and its Effect on Microwave Systems

What is Phase Noise?

Frequency stability can be defined as the degree to which an oscillating source produces the same frequency throughout a specified period of time. Every RF and microwave source exhibits some amount of frequency instability. This stability can be broken down into two components — long term and short term stability.

Long term stability describes the frequency variations that occur over long time periods, expressed in parts per million per hour, day, month, or year. Short term frequency stability contains all elements causing frequency changes about the nominal frequency of less than a few seconds duration. This product note deals with short term frequency stability.

Mathematically, an ideal sinewave can be described by

$$V(t) = V_0 \sin 2\pi f_0 t$$

where V_0 = nominal amplitude,

$2\pi f_0 t$ = linearly growing phase component,

and f_0 = nominal frequency.

But an actual signal is better modeled by

$$V(t) = |V_0 + \epsilon(t)| \sin |2\pi f_0 t + \Delta\phi(t)|$$

where $\epsilon(t)$ = amplitude fluctuations,

and $\Delta\phi(t)$ = randomly fluctuating phase term or phase noise.

This randomly fluctuating phase term could be observed on an ideal spectrum analyzer (one which had no sideband noise of its own) as in Figure 2.1a. There are two types of fluctuating phase terms. The first, deterministic, are discrete signals appearing as distinct components in the spectral density plot. These signals, commonly called spurious, can be related to known phenomena in the signal source such as power line frequency, vibration frequencies, or mixer products.

The second type of phase instability is random in nature, and is commonly called phase noise. The sources of random sideband noise in an oscillator include thermal noise, shot noise, and flicker noise.

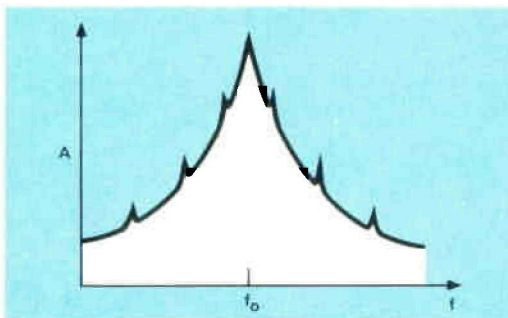
Many terms exist to quantify the characteristic randomness of phase noise. Essentially, all methods measure the frequency or phase deviations of the source under test in either the frequency or time domain. Since frequency and phase are related to each other, all of these terms are also related.

One fundamental description of phase instability or phase noise is the one-sided spectral density of phase fluctuations on a per-Hertz basis. The term spectral density describes the energy distribution as a continuous function, expressed in units of energy within a specified bandwidth. Thus, $S_{\Delta\phi}(f)$ (Figure 2.1b) is defined as

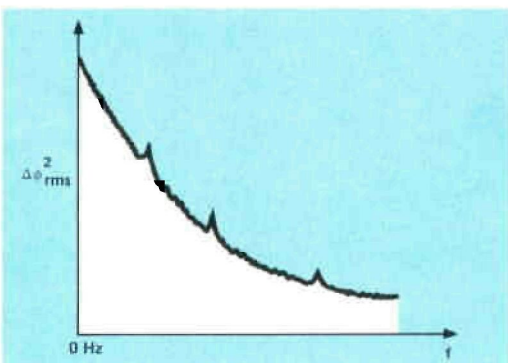
$$S_{\Delta\phi}(f) = \frac{\Delta\phi_{rms}^2}{\text{BW used to measure } \Delta\phi_{rms}} \quad \frac{\text{rad}^2}{\text{Hz}}$$

If the modulation sidebands are such that the **total phase deviations are $\ll 1$ radian** ($\Delta\phi_{pk} \ll 1$ radian), then another useful measure of the noise energy is $\mathcal{L}(f)$, which is then directly related to $S_{\Delta\phi}(f)$ by

$$\mathcal{L}(f) = \frac{1}{2} S_{\Delta\phi}(f).$$



2.1.a. RF sideband spectrum.



2.1.b. Phase noise sidebands.

Figure 2.1. CW signal sidebands viewed in the frequency domain.

$\mathcal{L}(f)$ is an indirect measure of noise energy easily related to the RF power spectrum observed on a spectrum analyzer. As shown in Figure 2.2, the U.S. National Bureau of Standards defines $\mathcal{L}(f)$ as the ratio of the power in one phase modulation sideband, on a per Hertz of bandwidth spectral density basis, to the total signal power, at an offset f_m Hertz away from the carrier, where f_m = the Fourier frequency or offset frequency.

$$\mathcal{L}(f) = \frac{\text{power density (in one phase modulation sideband)}}{\text{total signal power}} = \frac{P_{SSB}}{P_s}$$

= single sideband (SSB) phase noise to carrier ratio (per Hz)

$\mathcal{L}(f)$ is usually presented logarithmically as a spectral density plot of the phase modulation sidebands in the frequency domain, expressed in dB relative to the carrier per Hz (dBc/Hz), as shown in Figure 2.3. This product note, except where indicated otherwise, will use the logarithmic form of $\mathcal{L}(f)$ to describe and graph phase noise.

Another common term for quantifying short term frequency instability is $S_{\Delta f}(f)$, the spectral density of frequency fluctuations, related to $\mathcal{L}(f)$ as follows:

$$S_{\Delta f}(f) = 2f^2 \mathcal{L}(f)$$

Most common measurement systems measure the terms $S_{\Delta\phi}(f)$ or $S_{\Delta f}(f)$ directly. Throughout this note we shall use $\mathcal{L}(f)$, but if the other terms are more familiar, they can be related to $\mathcal{L}(f)$ as shown.

Residual and Absolute Noise

There are two different kinds of phase noise commonly specified. For devices (such as multipliers, mixers, amplifiers) and sources that operate from a common reference oscillator, the RESIDUAL phase noise is that noise **inherent** in the device, regardless of the noise of the reference oscillator used. The ABSOLUTE phase noise of the source or device is the total noise present at the output, and is a function of both the residual phase noise of the device AND the reference oscillator noise. You can see from this definition that fundamental sources only have absolute noise associated with them.

The procedures described in this note are for making absolute phase noise measurements on microwave sources. In general, it is the absolute phase noise of a source that is important in the final systems application. However, residual noise of devices or synthesized sources is often measured prior to system integration. For residual measurements on sources or devices, the HP 3047A Spectrum Analyzer System is a good solution.

Why is Phase Noise Important?

Phase noise on signal sources is a concern in frequency conversion applications where signal levels span a wide dynamic range. The frequency offset of concern and the tolerable level of noise at this offset vary greatly for different microwave systems. Sideband phase noise can convert into the information bandwidth and limit the overall system sensitivity.

This general case is illustrated in Figure 2.4. Suppose two desired signals f_1 and f_2 are input to the frequency conversion system, where

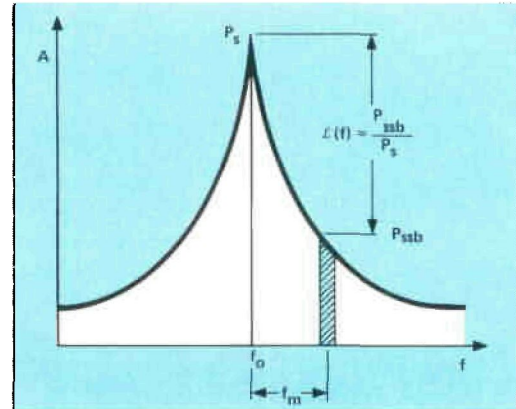


Figure 2.2. Deriving $\mathcal{L}(f)$ from a spectrum analyzer display.

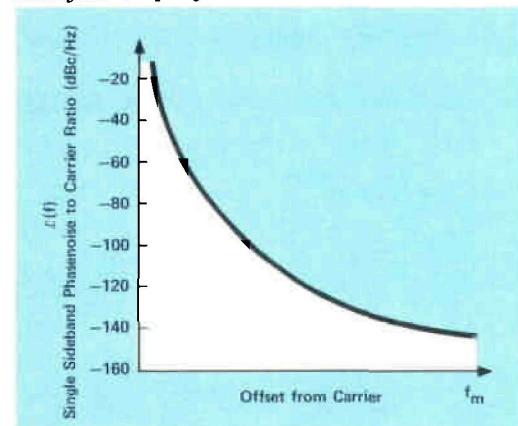


Figure 2.3. $\mathcal{L}(f)$ described logarithmically as a function of offset frequency.

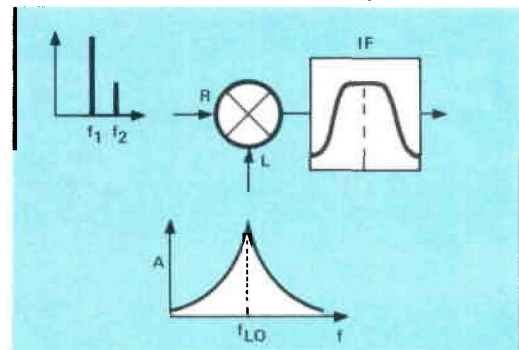


Figure 2.4a. Inputs to mixer.

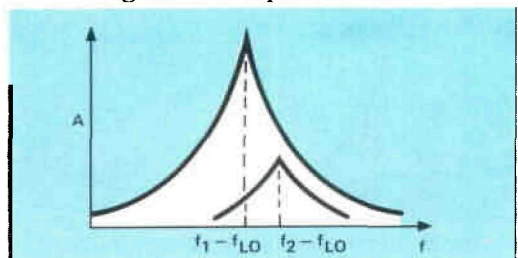


Figure 2.4b. IF output.

Figure 2.4. Effect of LO noise in frequency conversion application.

they are to be mixed with a local oscillator signal f_{LO} (Figure 2.4a) down to an intermediate frequency (IF) for processing. The phase noise of the local oscillator will be directly translated onto the mixer products (Figure 2.4b). Note that though the system's IF filtering may be sufficient to resolve the larger signal's mixing product ($f_1 - f_{LO}$), the smaller signal's mixing product ($f_2 - f_{LO}$) is no longer recoverable due to the translated local oscillator noise. The noise on the local oscillator thus degrades the system's sensitivity as well as its selectivity. Three specific examples of frequency conversion applications where phase noise is important follow.

Digital Communications System

In digital communications, phase noise very close to the carrier (less than 1 kHz) is important. Close-in phase noise (or phase jitter in the time domain) on the system local oscillator (LO) affects the system bit-error rate.

Analog Microwave Communications System

In many analog communications systems, modulation information is placed at least several hundred kHz away from the carrier. Initially, the signal to noise ratio is always sufficiently high. However, in each repeater station, the incoming signal is amplified, increasing the level of broadband noise on the carrier. If the signal passes through several repeater stations, the level of this broadband noise can increase and start to mask the information. Too high a level of broadband noise on each system local oscillator will affect the signal-to-noise (or system sensitivity) at the receiving end of a multiple hop system.

Doppler Radar System

Doppler radars determine the velocity of a target by measuring the small shifts in frequency that the return echoes have undergone. In actual systems, however, the return signal is much more than just the target echo. The return includes a large 'clutter' signal from the large, stationary earth (Figure 2.5). If this clutter return is decorrelated by the delay time difference, the phase noise from the local oscillator can partially or even totally mask the target signal. Thus, phase noise can set the minimum signal level that must be returned by a target in order to be detectable.

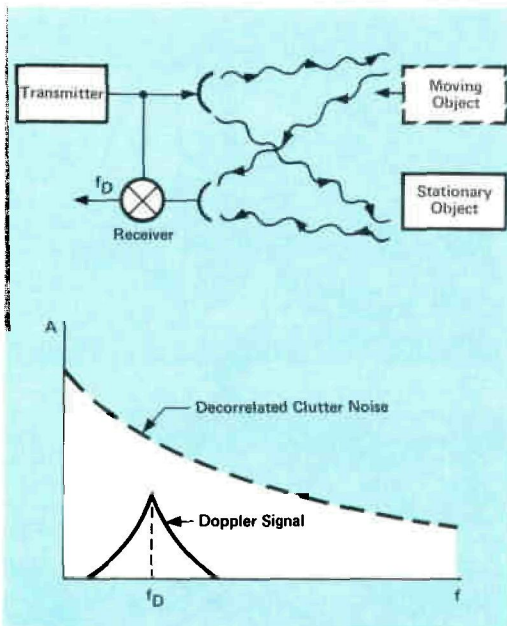


Figure 2.5. Effect of carrier phase noise in a doppler radar system.

Common Measurement Techniques

There are several methods of making phase noise measurements, each with its own set of advantages and disadvantages. Here is a brief summary of some of the most common methods, with a few comments about their applicability.

Direct Measurement

The most straightforward method of phase noise measurement is to input the test signal into a spectrum analyzer, directly measuring the power spectral density of the oscillator. However, this method may be significantly limited by the spectrum analyzer's dynamic range, resolution, and LO phase noise.

Though this direct measurement is not useful for measurements close-in to a drifting carrier, it is convenient for qualitative quick evaluation on sources with relatively high noise. The measurement is valid if the following conditions are met:

- A. The spectrum analyzer SSB phase noise at the offset of interest must be lower than the noise of the Device Under Test (DUT).
- B. Since the spectrum analyzer will measure total noise power, the amplitude noise of the DUT must be significantly below its phase noise. (Typically 10 dB will suffice.)

Heterodyne Frequency Measurement

This time domain method downconverts the signal under test to an intermediate frequency. Then a high resolution frequency counter repeatedly counts the IF signal frequency, with the time period between each measurement held constant. This allows several calculations of the fractional frequency difference, y , over the time period used. From these values for y , the Allan variance, $\sigma_y(\tau)$ can be computed. $\sigma_y(\tau)$ in the time domain corresponds to $\mathcal{L}(f)$ in the frequency domain.

This method is particularly useful for short-term frequency instabilities occurring over periods of time greater than 10 ms (less than 100 Hz offsets in the frequency domain), where the phase noise is falling rapidly. It is ideal for close-in measurements on frequency standards. However, it is not well suited for measurement of noise at offsets from the carrier greater than 10 kHz (Figure 3.3), or for measuring noise which is flat or decreasing slowly vs. offset frequency (as a function of $1/f$ or less).

Carrier Removal / Demodulation

Most of the techniques for phase noise measurements fall into this class. Increased sensitivity is obtained by nulling the carrier, or demodulating the carrier and then measuring the noise of the resulting baseband signal. Most common of this class are 1) measurements with a frequency discriminator, and 2) measurements with a phase detector. Figure 3.3 compares some typical sensitivities of these methods and the heterodyne frequency measurement.

Measurement with a frequency discriminator

This frequency domain method feeds the signal under test into a frequency discriminator and monitors the output on a low frequency spectrum analyzer (Figure 3.4). Common discriminators used include delay line/mixer, cavity, and bridge types.

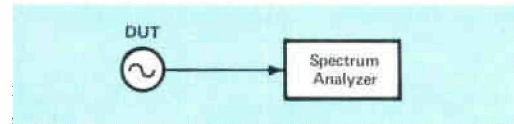


Figure 3.1. Direct measurement on a spectrum analyzer.

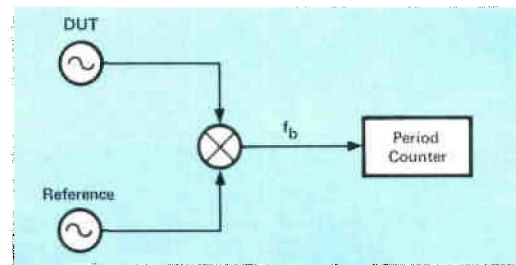


Figure 3.2. Heterodyne frequency measurement.

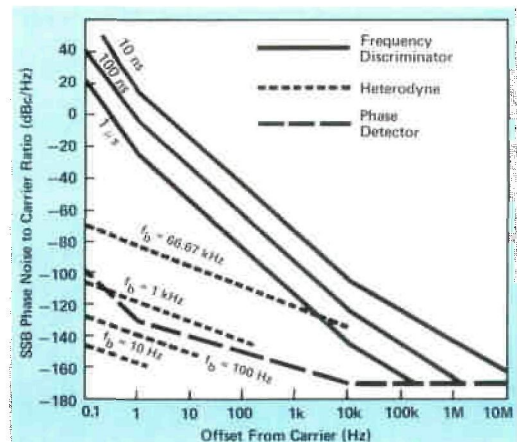


Figure 3.3. Comparison of typical system sensitivities at 10 GHz.

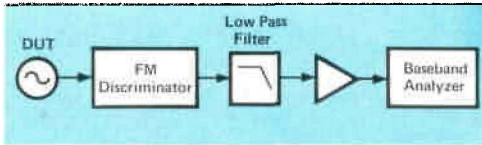


Figure 3.4. Noise measurement with a frequency discriminator.

This method is useful for measuring sources with large, low rate phase instabilities such as free-running sources. However, due to the inherent relationship between frequency modulation and $S_{\Delta\phi}(f)$, frequency discriminator system sensitivity degrades as $1/f^2$ as the carrier under test is approached (Figure 3.3). Frequency discriminators thus are limited for measurements on stabilized sources. Also, the physical discriminators commonly implemented tend to be narrow band. The advantage of the discriminator technique is that it does not require a second source for down-conversion; it is sometimes referred to as the one-oscillator technique.

Measurement with a phase detector

This method yields the overall best sensitivity, as shown in Figure 3.3. This is one method used in the 11729B and will be explored in detail in this and subsequent chapters. Since this method requires two sources input to a phase detector, it is sometimes called the two-oscillator technique.

The Phase Detector Method

Basic Theory

The fundamental block diagram of this technique is shown in Figure 3.5. (NBS has utilized this technique for phase noise measurements since 1967.) The basis of this method is the double-balanced mixer used as a phase detector. Two signals at identical frequencies and nominally in phase quadrature (90° out of phase) are input to the phase detector. At quadrature, the output of the phase detector is a difference frequency of 0 Hz and an average voltage output of 0V.

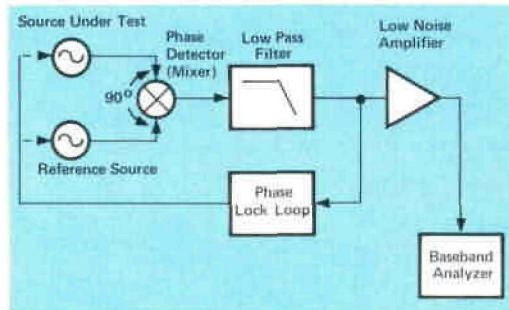


Figure 3.5. Phase Detector method — basic block diagram.

There is also a small fluctuating voltage, ΔV . For small phase deviations ($\Delta\phi_{pk} \ll 1$ radian), this fluctuating voltage is proportional to the fluctuating phase difference between the two signals. (The assumption that $\Delta\phi_{pk} \ll 1$ radian is equivalent to saying that the higher order Bessel or modulation components are insignificant compared to the carrier and first sideband. For nearly all high quality sources, this is a good assumption.) This phase difference represents the COMBINED phase modulation sidebands of the two input signals.

Stated simply, when the two input signals are identical in frequency and in phase quadrature, the output of the phase detector is a voltage directly proportional to the combined phase modulation sidebands of the two input signals.

To define how the signals measured using the phase detector method result in an expression of phase noise, first return to the definition of $\mathcal{L}(f)$.

$$\mathcal{L}(f) = \frac{P_{ssb}}{P_s} = \frac{\text{power density (in one phase modulation sideband)}}{\text{total signal power}}$$

Expressing $\mathcal{L}(f)$ logarithmically yields

$$\mathcal{L}(f) \text{ dBc} = 10 \log \frac{P_{ssb}}{P_s} .$$

Thus, to measure the SSB phase noise of a source under test, the system is first calibrated by establishing a carrier reference level, P_s . This can be done by offsetting one of the two sources to the mixer and observing the resultant beat note V_b (Figure 3.6). The level of this beat note V_b corresponds to the total energy of the carrier (P_s in dBm).

There is typically a low noise amplifier in the phase detector system. To prevent this amplifier from saturating during the calibration procedure, the input level to the R port of the phase detector/mixer is usually reduced — for example, by 40 dB. Then the resultant beat note V_b can be consi-

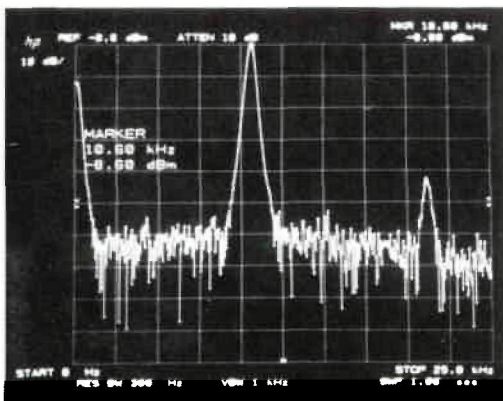


Figure 3.6. $V_b = P_s(\text{dBm}) - 40 \text{ dB}$.

dered as representing the level of the carrier under test reduced by 40 dB (P_S (dBm) = V_b (dBm) + 40 dB).

The frequency and amplitude offsets are then removed such that the two input signals are again at identical frequencies, and are set in phase quadrature. To properly use a double balanced mixer as a phase detector, it is important to use the mixer in its linear range (where the voltage output is directly proportional to the phase difference of the input signals by a constant K). Operation at quadrature assures linear operation.

At quadrature, the output of the phase detector V_S (sideband level in dBm) contains the desired information about the level of the SSB noise modulation. It also contains the power in the second sideband, due to the mixer operation that must be subtracted out.

To understand this correction factor, first let's look at the operation of a mixer in standard frequency modulation theory. Let the two inputs to the mixer be as shown in the frequency domain in Figure 3.7a. The LO signal is assumed ideal, and the signal to the R port is at the same frequency but has a single tone of FM modulation with a small modulation index. The output of the mixer is shown in Figure 3.7b. As before, the mixer sum component is filtered off, leaving the difference component. This difference signal is dc, but the lower FM sideband has been folded into the upper FM sideband around 0 Hz. The amplitudes of the two modulation sideband voltages add linearly in the output of the mixer, resulting in four times as much power in the one resultant sideband as in one original sideband. If we looked at this output signal on a spectrum analyzer, and wanted to know the level of one of the sidebands on the original RF signal, we could take the measured power level and subtract 6 dB.

This is analogous to the effect of the mixer/phase detector on the noise sidebands. The continuous noise sidebands can be thought of as a series of discrete modulation sidebands. Each of these modulation sidebands is folded around dc, and their amplitudes add linearly in the output of the mixer. Since $\mathcal{L}(f)$ involves the noise power in just ONE sideband, the measured noise sideband level V_S is 6 dB higher than the original single sideband. Therefore, in quadrature, the power density in one phase modulation sideband is $V_S - 6$ dB. Thus, for

V_b = calibration beat note in dBm, and P_S in milliwatts, then

$$10 \log P_S = V_b + 40 \text{ dB},$$

and for

V_S = measured sideband level in dBm, and P_{SSb} in milliwatts, then

$$10 \log P_{SSb} = V_S - 6 \text{ dB}.$$

Therefore,

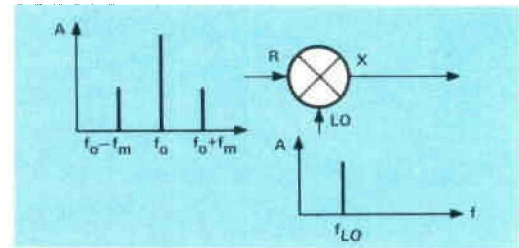
$$\begin{aligned} \mathcal{L}(f) \text{ dBc} &= 10 \log (P_{SSb}) - 10 \log (P_S) \\ &= V_S - 6 \text{ dB} - V_b - 40 \text{ dB}. \end{aligned}$$

(The above discussion of measuring the level of noise sidebands is applicable for $\Delta\phi \ll 1$ radian. However, even if this assumption is not true, the mixer will still act as a phase detector. However, the phase detector output will be $S\Delta\phi(f)$, but this will not be equal to $\mathcal{L}(f)$.)

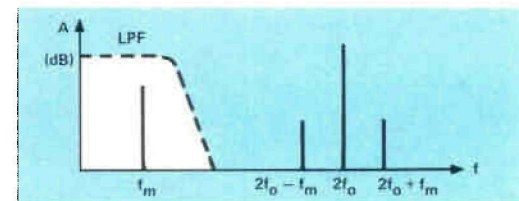
For a much more structured derivation of the mixer outputs and how they relate to $\mathcal{L}(f)$, see Appendix A.

Importance of the Reference Source

The resultant phase noise measured at the output of the mixer is equal to the COMBINED noise of the two sources. Therefore, it is desirable to use as low a noise reference as possible. The noise floor of the phase detector method was shown in Figure 3.3. In actual measurements, how-



a. Mixer Inputs



b. Mixer Outputs

Figure 3.7. Effect of mixer on sidebands when downconverting with $f_{LO} = f_o$.

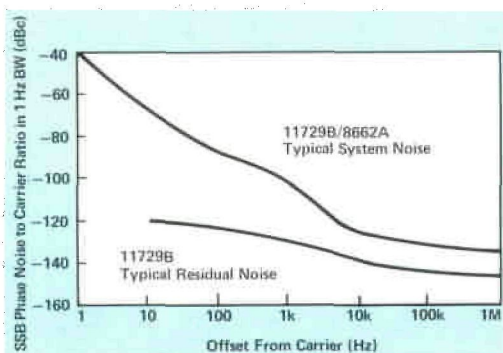


Figure 3.8. Typical 11729B residual noise and system absolute noise at 10 GHz.

ever, the real limitation on system noise will be the noise floor of the reference source.

The 11729B, when driven with the appropriate signals, provides a low noise microwave reference source for use in a phase noise measurement system. Figure 3.8 shows the typical system noise floor at 10 GHz when driving the 11729B with an HP 8662A Synthesized Signal Generator.

Since the voltage out of the phase detector represents the combined noise of the two input sources, there is an error introduced by the finite noise contribution of the reference source. If the reference has significantly lower noise than the DUT, the error is negligible. However, the error increases for reference noise approaching the noise of the test source. The following equation can be used to estimate this error. P_{ref} and P_{dut} are the noise power of the reference and the DUT, in watts, at a particular offset.

$$\text{error (dB)} = 10 \log \left(1 + \frac{P_{ref}}{P_{dut}} \right)$$

$$\text{or error (dB)} = 10 \log \left[1 + \text{antilog} \left(\frac{\mathcal{L}_{ref} - \mathcal{L}_{dut}}{10} \right) \right]$$

The error is defined as $\mathcal{L}(f)$ measured minus $\mathcal{L}(f)$ dut, \mathcal{L}_{dut} = SSB noise power ratio of the source under test, \mathcal{L}_{ref} = SSB noise power ratio of the reference, with all \mathcal{L} in dBc.

Included in Chapter 6 is a tabulation of the measurement error for several values of the difference in noise powers of the reference and DUT.

Importance of Quadrature

For the mixer to act as an accurate phase detector, it must remain within several degrees of quadrature. In a user-designed noise measurement system, the deviation from quadrature should be monitored and its error contribution accounted for. But as will be seen in Chapter 4, the 11729B Carrier Noise Test Set ensures good phase quadrature by controlled phase locking. This frees the measurement system from any error that might be associated with quadrature maintenance.

At quadrature, a double balanced mixer also has minimum sensitivity to AM noise; typical AM noise suppression is on the order of >20 dB. Again, in a user-designed system, if the two phase detector inputs drift out of quadrature, the AM noise sensitivity increases, and the phase noise sensitivity decreases. Such increased sensitivity to AM noise would cause an additional measurement error if the source under test has high AM noise.

To ensure that the final phase noise measurement has not been influenced by AM noise, the 11729B provides good quadrature maintenance. The 11729B also features optional AM noise detection, with a noise floor typically -165 dBc for offsets greater than 1 MHz.

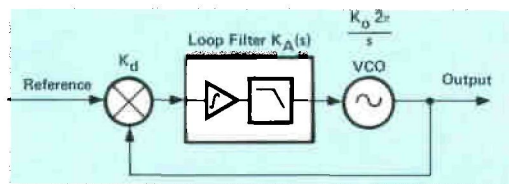


Figure 3.9. Basic phase lock loop model.

Effect of Phase Lock Loop

The phase detector method requires the two signals to remain in quadrature for the duration of the measurement. This can be accomplished by phase locking the sources together using the dc mixer output connected to a frequency control line on one of the sources (acting as the VCO). The loop feedback causes the VCO to tune until it is at the same frequency as the second source, and the error voltage at the phase detector output returns to nominally zero volts. Therefore, the source under test can drift during the measurement and the reference source will follow it.

Figure 3.9 shows the model of a basic phase lock loop. This model shows an EFFECTIVE loop filter after the mixer determining the bandwidth of the phase lock loop. (This effective loop filter consists of an integrator and low pass filter, with actual loop bandwidth being a function of several loop components. See Appendix B for a more complete discussion of phase lock loops in the 11729B.)

Inside this loop bandwidth, the sources track in phase. For example, a 100 Hz loop bandwidth would allow the VCO to track any phase change of the reference occurring at rates less than 100 Hz. Note that the use of a second order loop in the 11729B allows the sources to track slow frequency changes many times larger than the loop bandwidth. The sources can track in frequency only over the tuning range of the VCO; this range is defined as the loop holding range (LHR).

The tracking of phase variations inside the loop bandwidth attenuates phase noise (measured at the phase detector output) within this bandwidth. In normal measurement operation, a phase noise measurement at offset frequencies inside the loop bandwidth will appear to have phase noise lower than it actually is. This is shown by the actual noise plots of Figure 3.10. A noise measurement was made on a 10 GHz source for offsets from 0 to 100 Hz, first with a 1 Hz loop bandwidth, and then with a 100 Hz loop bandwidth. The MEASURED noise with the 100 Hz measurement loop BW is lower than the correct noise measured with the 1 Hz loop BW. This figure shows the need for correction factors for accurate measurements at offsets inside the loop bandwidth.

General Measurement Procedure

The phase detector method of phase noise measurements can be broken down into five main steps: 1) set-up, 2) calibration, 3) lock, 4) measure, and 5) corrections.

Set-Up and Calibration

System set-up follows the basic phase detector block diagram of Figure 3.11.

The calibration procedure involves offsetting one of the sources to produce a beat note, used to set the carrier power reference level. The actual phase noise measurement readings will then be made relative to this reference carrier level.

To prevent overloading the low noise amplifier and spectrum analyzer by the beat note, the amplitude of the signal applied to the R port of the phase detector is reduced in level by 40 dB for calibration. In practice, best measurement accuracy is obtained if sufficient attenuation is included in the calibration stage so that the spectrum analyzer input sensitivity need not be changed between calibration and measurement.

If using more than one spectrum analyzer for the baseband analysis, each analyzer should have a reference calibration performed. The calibration should be made with the low noise amplifier (if used) included in the system.

Lock/Quadrature

After calibration, the DUT and the reference oscillator are returned to the same frequency and the original power levels.

The reference oscillator must now be locked to the DUT in quadrature. The phase lock loop error voltage is connected to the frequency control of the VCO. The bandwidth of the phase lock loop is then selected wide enough to follow the large amplitude, low rate phase instabilities of the

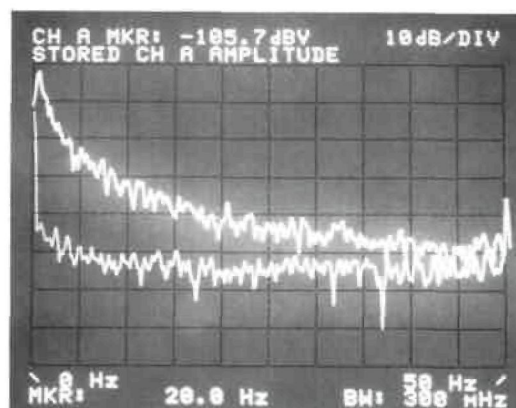


Figure 3.10. Noise plots showing effect of loop bandwidth setting. Upper trace: LBW = 1 Hz. Lower trace: LBW = 100 Hz.

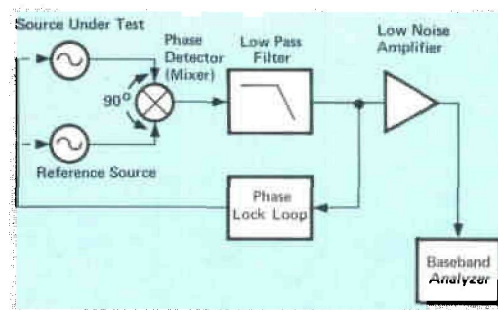


Figure 3.11. Phase detector method.

source under test which would otherwise drive the phase detector out of its linear range.

With careful phase lock loop design, the phase lock loop holding range will be equal to the total VCO tuning range. Loop bandwidth is a function of several loop parameters, including the VCO slope, the loop gain, and the phase detector phase slope. See Appendix B for a more complete phase lock loop model.

The phase lock loop suppresses noise within its loop bandwidth. In normal measurement operation, the low frequency cut-off of the phase lock loop must be below the lowest offset frequency to be analyzed. However, techniques exist to correct for the noise suppression within the bandwidth of the phase lock loop. Again, see Appendix B for more information.

Measure

After the sources are locked in quadrature, the combined phase noise of both sources can be measured on the spectrum analyzers. The analyzers are set up to span the offset frequencies of interest with sufficient resolution bandwidth to resolve spurious. Because phase noise exhibits a random density spectrum, it is desirable to use some form of averaging if available.

Care should be exercised while making phase noise measurements; a list of potential problems to be considered to avoid erroneous data is given in Appendix A. Chapter 5 discusses in detail proper system operation with the 11729B to eliminate most potential system errors.

Corrections

After the phase noise level has been measured, a number of correction factors must be applied to translate the measured noise value to $\mathcal{L}(f)$.

The subtraction of the level of the calibration beat note, the calibration attenuation and the 6 dB for mixer operation have already been discussed. Additional correction factors include the normalization of the equivalent noise bandwidth of the spectrum analyzer to a 1 Hz bandwidth. This bandwidth normalization factor is simply $10 \log$ (equivalent noise BW in Hz). Note that a spectrum analyzer's noise bandwidth is generally wider than the nominal front panel resolution bandwidth setting. Most Hewlett-Packard spectrum analyzers have equivalent noise bandwidths approximately 1.2 times the 3 dB resolution bandwidth. However, for best accuracy, the analyzer equivalent noise bandwidth should be measured. See Hewlett-Packard Application Note 150-4 "Spectrum Analysis . . . Noise Measurements".

In addition to bandwidth normalization, other correction factors may be required, depending on the type of spectrum analyzer used. Most analog spectrum analyzers use logarithmic amplifiers and peak detectors. When measuring random noise, these circuits indicate a value on the spectrum analyzer display about 2.5 dB less than the actual noise level. Thus an additional 2.5 dB correction factor must be added. Again, see AN 150-4 for further explanations of spectrum analyzer corrections.

These correction factors, summarized, are

$$\begin{aligned} \mathcal{L}(f), \text{ in dBc in a 1 Hz bandwidth } f_m \text{ Hz from the carrier,} \\ = V_s, \text{ sideband level (dBm)} \\ \text{minus } V_b, \text{ carrier level (dBm)} \\ \text{minus 40 dB for the attenuation added during calibration (or} \\ \text{whatever attenuation value was used)} \\ \text{minus 6 dB to convert measured value to } \mathcal{L}(f) \\ \text{minus } 10 \log \text{ equivalent noise BW (to convert reading from} \\ \text{the measurement bandwidth used to a 1 Hz BW)} \\ \text{plus 2.5 dB if analog spectrum analyzers are used.} \end{aligned}$$

11729B Theory of Operation and Measurement Considerations 4

The 11729B Carrier Noise Test Set uses the phase detector method of phase noise measurement, allowing test signals from 5 MHz to 18 GHz to be demodulated to baseband for phase noise analysis.

When used with the necessary driving signals, the 11729B first provides the critical low noise microwave reference signal. This reference signal is electronically tunable to track the drift in the source under test. In addition, the 11729B includes all the necessary demodulation circuitry plus variable phase-lock-loop/quadrature circuitry. A loop characterization feature allows a broad range of offset frequencies to be measured. And because the 11729B is HP-IB programmable, these measurements can also be made automatically.

First, let's understand how the 11729B works, so that we can better determine the proper measurement considerations. This chapter discusses the three major functional blocks of the Carrier Noise Test Set: the multiplier chain, the demodulator and baseband signal processing section, and the phase-lock-loop/quadrature section. A simplified 11729B block diagram is shown in Figure 4.1. See Figure 4.15 for a more complete block diagram.

Multiplier Chain

To obtain a low noise microwave signal, the 11729B capitalizes on available low noise performance at RF frequencies. The 11729B requires a fixed frequency 640 MHz drive signal for multiplication to microwave. The basic multiplying scheme utilizes a step-recovery diode multiplier to generate a comb of frequencies.

An RF source optimized for this application is the HP 8662A Synthesized Signal Generator. The 8662A front panel output covers 10 kHz to 1280 MHz with good phase noise characteristics (Figure 4.2). In addition, the 8662A has an optional auxiliary 640 MHz fixed frequency output with state-of-the-art spectral purity. This 640 MHz signal has SSB noise -149 dB below the carrier at a 10 kHz offset, reaching a noise floor of -159 dBc at a 100 kHz offset (again see Figure 4.2). It is this signal that will be used as the input to the 11729B multiplier chain. (For more information on the 8662A, see HP 8662A Technical Data Sheet and HP Application Note 283-1, "Applications and Measurements of Low Phase Noise Signals Using the 8662A Synthesized Signal Generator.") The model HP 8663A, 10 kHz to 2560 MHz, could also be used.

The 11729B multiplier chain is shown in detail in Figure 4.3. The 640 MHz signal first passes through a 640 MHz bandpass filter, to reject 10 MHz reference or 20 MHz reference harmonic spurious sidebands that may be present due to the synthesis process in the 8662A. A power amplifier then provides sufficient drive level to the step recovery diode (SRD) multiplier, which generates a comb of frequencies spaced 640 MHz apart extending to 18 GHz.

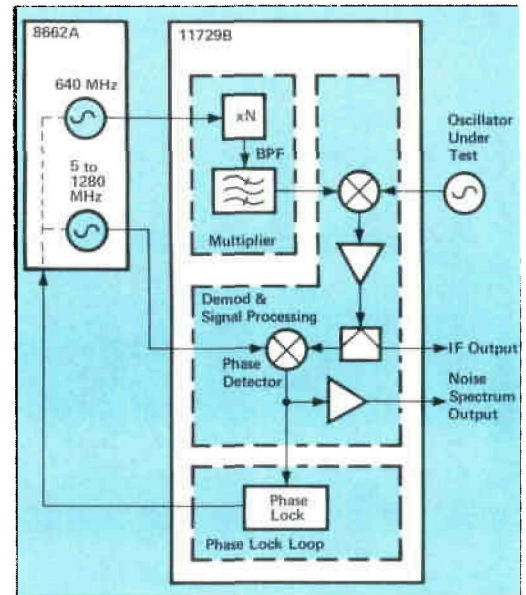


Figure 4.1. Simplified 11729B block diagram.

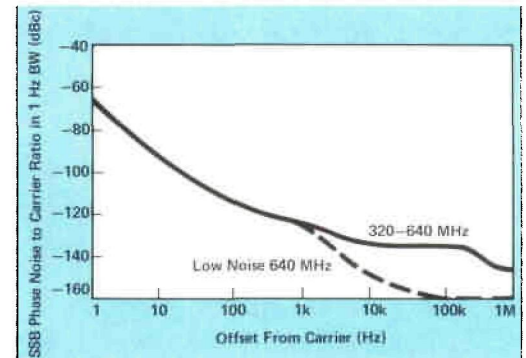


Figure 4.2. Typical 8662A low phase noise signals.

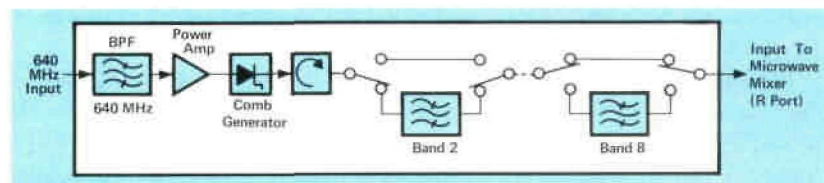


Figure 4.3. 11729B multiplier chain.

A circulator (or isolator) absorbs unwanted reflections from feeding back into the SRD. Then a microwave bandpass filter selects a comb line close to the frequency of the device under test. The result is an ultra-clean multiple of the 640 MHz signal, within 1280 MHz of the source under test.

The standard 11729B has 8 microwave switches with 7 bandpass filters installed, allowing a single 11729B to downconvert test signals from 5 MHz to 18 GHz. For frequencies less than 1.28 GHz, one switch bypasses the microwave mixer and applies the test signal directly to the RF mixer — see Demodulator section following. A single filter version of the 11729B, for narrowband or single test frequency applications, retains the bypass switch for low frequencies and one user-defined bandpass filter.

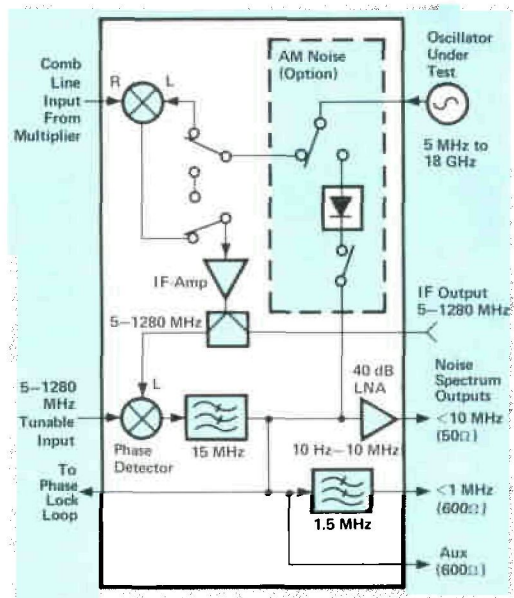


Figure 4.4. 11729B demodulator and baseband signal processing section.

Demodulator and Baseband Signal Processing Section

First Downconversion (microwave mixer)

The selected harmonic of the 640 MHz drive signal mixes with the microwave source under test in the microwave mixer to yield an IF of 5 to 1280 MHz (Figure 4.4). Because of the low level of the higher frequency comb lines, the source under test must provide the local oscillator (LO) drive power to the mixer. The resultant IF signal amplified to $\geq +7$ dBm is available at the front panel of the 11729B for monitoring, or for input to the HP 3047A System (see Chapter 7).

Because of the need to avoid carrier signal feedthrough in the baseband signal processing section (see below), the IF stage is limited to 5 MHz on the low end. This means that the frequency of the microwave source under test must be at least 5 MHz away from the selected comb line.

Phase Detector

The resultant IF signal is then phase-detected against a 5 MHz to 1280 MHz signal (available from the 8662A front panel), resulting in a baseband signal. An RF mixer operates in quadrature as the system phase detector. The IF from the first downconversion drives the L port. The 8662A front panel signal is applied to the R port, allowing amplitude changes on the 8662A to be directly translated to the baseband signal during the measurement calibration procedure.

Baseband Signal Processing

The resultant baseband signal is filtered through a 15 MHz (3 dB BW) low pass filter (LPF) to remove unwanted mixing products. The filtered signal is available at the 11729B front panel Auxiliary Noise Spectrum Output, and can be used as an additional quadrature monitor, or as a tracking monitor for noise measurements on sources with high drift (see Chapter 5, Loop Tracking).

The baseband signal is further processed through a low noise amplifier (LNA), and then brought to the <10 MHz Noise Spectrum output. The amplifier has about 40 dB of gain (coupled into 50Ω), a bandwidth of about 10 Hz to 30 MHz, typical flatness of less than 1 dB, and a noise figure of about 1.6 dB. This LNA permits the 11729B phase detected noise output to be viewed on standard lab spectrum analyzers.

For offsets close to the carrier, the phase noise level is usually sufficiently high that additional amplification is not required. The 10 MHz low pass filtered signal is filtered again through a 1.5 MHz LPF to remove any close-in mixing products or other high level spurious, and is available at the <1 MHz Noise Spectrum Output for viewing on a suitable low frequency spectrum analyzer.

Care should be taken when operating with IF frequencies less than 20 MHz. One purpose of the low pass filters is to avoid overloading the spectrum analyzers with high level mixer outputs (for example, LO feed-through from low 11729B IF frequencies) while attempting to view lower level noise. Because the final low pass filter is placed at 1.5 MHz, IF frequencies are limited to 5 MHz when the amplifier is not used, and to about 20 MHz if the LNA is included. If a lower IF frequency is critical, additional low pass filtering can be user-added.

The bandwidth of the 11729B IF amplifier limits the bandwidth of the IF stage to 1500 MHz. At the phase detector, the 8662A provides a tunable signal to 1280 MHz. This allows for test signals 1280 MHz either side of the selected comb line, or 2560 MHz of maximum phase detection bandwidth for each filter. For example, if the 9.6 GHz filter is selected (15th harmonic of 640 MHz), test frequencies from 8.32 GHz to 10.88 GHz (excluding ± 5 MHz centered on 9.6 GHz) can be phase-detected. The frequency of the selected comb line, as well as the range of input signals that can be downconverted with each comb line, are indicated on the front panel of the 11729B.

8662A phase noise is a function of output frequency selected. Therefore, when using the 8662A at the RF port of the phase detector, system noise can be affected by the value of the frequency necessary to phase detect against the 11729B IF. See Appendix C for more information on computing system noise.

Phase-lock Loop/Quadrature Section

The input signals to the mixer must be maintained in quadrature for the duration of the measurement. This is done through the phase lock loop section as shown in the 11729B phase lock loop block diagram (Figure 4.5). An understanding of several characteristics of this phase lock loop that can affect the phase noise measurement is necessary; this section will summarize these important characteristics. For a more complete derivation of the phase lock loop performance and how to change the loop characteristics for advanced applications, see Appendix B.

Loop Hold

During a normal locked measurement, a second order phase lock loop maintains the phase detector input signals in quadrature; this condition can be called loop hold. In this locked condition, one of the phase detector input signals MUST function as the VCO.

There are four basic ways to lock the microwave source under test when using an 11729B/8662A phase noise measurement system: locking via the 8662A internal reference oscillator, locking via the 8662A dc FM, locking via the frequency control of a third source, and locking via the frequency control of the DUT. Each of these methods will be described in this section.

For each of these locking methods, we need to know two important characteristics of the loop: the loop holding range and the loop band

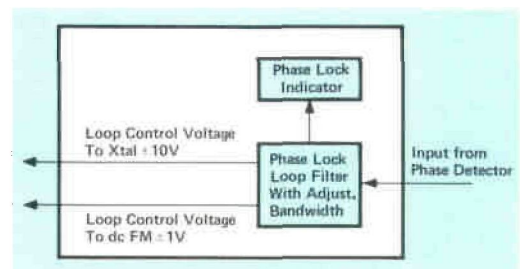


Figure 4.5. 11729B phase lock loop.

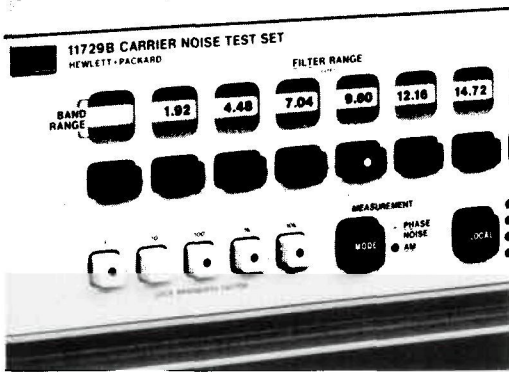


Figure 4.6. 11729B lock bandwidth factor.

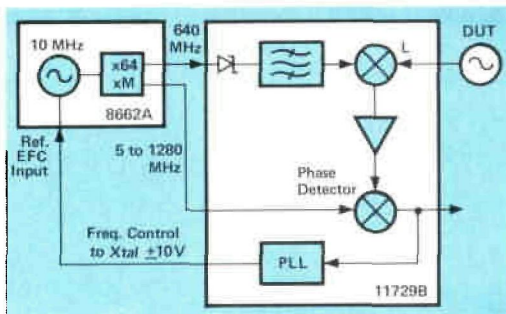


Figure 4.7. Phase locking through the 8662A reference oscillator.

width. As discussed in Chapter 3, the loop holding range (LHR) is how far the VCO can track the DUT in frequency. In practice, LHR is the amount of frequency drift on the source under test that can be tolerated and the loop remain locked. The loop bandwidth (LBW) is that range of offset frequencies where the loop is tracking in phase. LBW determines the offsets where the measured phase noise will be reduced by the lock loop. In general, the LBW is a function of the 11729B Lock Bandwidth Factor (Figure 4.6), the tuning slope of the VCO, and the amplitude of the VCO. In normal operation as described in this chapter, we will assume that the amplitude of the VCO is fixed at 0 dBm during measurement, and thus does not contribute to changing loop bandwidth. Appendix B discusses effect of changing VCO amplitude.

Phase Locking via the 8662A Reference Oscillator

The first way to lock the microwave DUT is through the crystal reference oscillator of the 8662A acting as the VCO in the phase lock loop (Figure 4.7). This oscillator has an Electronic Frequency Control (EFC) input, which allows the 11729B Frequency Control to XTAL $\pm 10V$ signal to tune the 10 MHz oscillator nominally ± 1 Hz. The advantage of locking via EFC is lower noise, but with narrow tuning range.

Loop Holding Range (LHR)

Nominal loop holding range (LHR) when locking thru the EFC of the 8662A can be expressed simply by

$$\text{nominal EFC LHR (Hz)} = \pm \frac{f_{\text{dut}} \text{ (Hz)}}{10^7}$$

Locking by tuning the 8662A Reference Oscillator yields a somewhat limited loop holding range (1 part in 10^7). For DUT's such as stabilized sources, this loop holding range will probably be sufficient. However, for free-running sources and some stabilized sources, the drift characteristics or the rate of large phase instabilities in the DUT may be such that larger loop holding range and loop bandwidth are required. For these sources, the 11729B provides a second method using the 8662A as a VCO (see Phase Locking via 8662A dc FM).

Loop Bandwidth (LBW)

When locking via the 8662A Reference Oscillator, the nominal loop bandwidth is determined only by the 11729B Lock Bandwidth Factor (LBF) and the frequency of the DUT (which determines the effective tuning range of the crystal, acting as the VCO). Nominal loop bandwidth is given by

$$\text{nominal EFC LBW (Hz)} = \frac{11729B \text{ LBF} \times f_{\text{dut}} \text{ (Hz)}}{10^{10}}$$

Loop Bandwidth Limitations

The maximum LBW has certain limitations when locking via the 8662A Reference Oscillator (as explained in detail in Appendix B). When locking via the EFC of the 8662A Reference Oscillator, the phase lock loop has an effective low pass filter cutoff < 3 kHz. If a bandwidth approaching this cut-off frequency is selected, the loop will be unstable. For example, for a test frequency of 10 GHz, the 10k Lock Bandwidth Factor would always yield an unstable loop. And, depending on the exact frequency of this LPF, the 1k Lock Bandwidth Factor might also be unstable.

Phase Locking via the 8662A dc FM

Loop Holding Range (LHR)

The second method for maintaining quadrature at the phase detector permits greater loop holding range and loop bandwidth. By applying the 11729B $\pm 1V$ Frequency Control dc FM signal to the 8662A dc coupled FM input, the 8662A front panel signal acts as the loop VCO (Figure 4.8). The 8662A has ± 200 kHz maximum dc FM deviation. The loop holding range is set only by the FM deviation selected on the 8662A, and is not a function of DUT frequency. (See Appendix B for loop models.)

$$\text{nominal dc FM LHR (Hz)} = \pm \text{FM deviation set on 8662A}$$

Maximum FM deviation of the 8662A is ± 200 kHz, but allowable deviation is a function of frequency. For example, for frequencies from 320 to 640 MHz, the 8662A maximum FM deviation is halved to ± 100 kHz. Thus, the maximum LHR is a function of the 8662A frequency required. (8662A frequency required = 11729B IF frequency = $f_{\text{dut}} - f_{\text{filter}}$.)

Appendix C shows how better system noise performance is achieved with lower 8662A front panel frequencies. Since lower 8662A front panel frequencies also yield lower maximum FM deviations, and thus lower LHR, there is a trade-off between loop holding range and system noise floor that needs to be considered in certain cases.

Loop Bandwidth (LBW)

Using the $\pm 1V$ control signal yields

$$\text{nominal dc FM LBW} = \frac{(\text{8662A front panel FM deviation}) \times 11729B \text{ LBF}}{10^3}$$

(assuming 8662A amplitude is 0 dBm)

Thus, the loop bandwidth can be changed by varying the Lock Bandwidth Factor plus varying the peak FM deviation. Note that changing the Lock Bandwidth Factor changes the loop bandwidth in decade steps, while changing the FM deviation can change the loop bandwidth in smaller increments.

Phase Locking Using a 3rd Source or the DUT as the Loop VCO

The 8662A provides 200 kHz maximum LHR (maximum peak dc FM deviation), and limits the loop bandwidth to less than 100 kHz. For some free-running sources, the magnitude of phase instabilities are such that wider bandwidth phase lock loops are required to keep the phase detector in its linear range, or the drift is such that more LHR is needed. This requires wider dc FM bandwidth and LHR than is available on the 8662A. For these types of sources, two alternate locking techniques are available.

First, the 8662A signal to the phase detector can be replaced with another RF signal with wider dc FM bandwidth. A good source to use is the HP 8640B Option 002 Signal Generator, as shown in Figure 4.9. The 8640B provides a 0.5 to 1040 MHz tunable signal with up to 5 MHz peak dc coupled FM deviation (when driven with a 1V signal). Using the 8640B or other appropriate RF source would thus enable wider LHR and

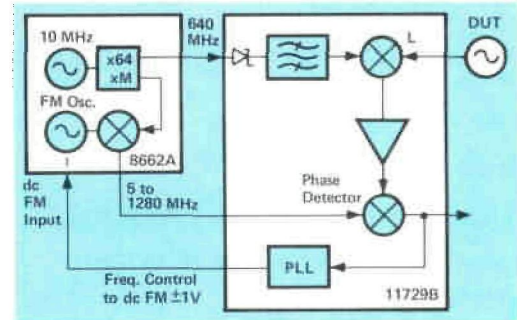


Figure 4.8. Phase locking through the 8662A dc FM.

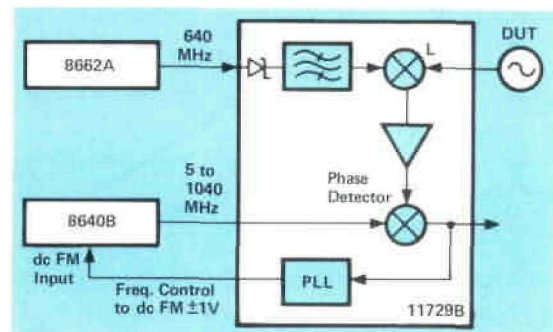


Figure 4.9. Phase locking using a third source as the loop VCO.

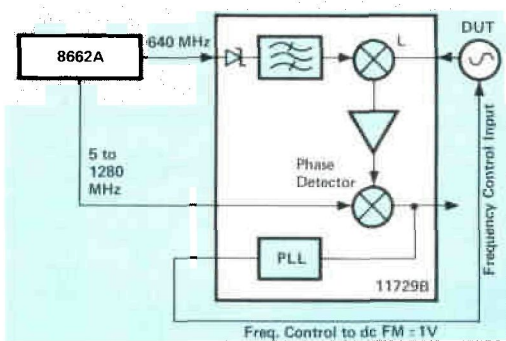


Figure 4.10. Phase locking using the frequency control of the DUT.

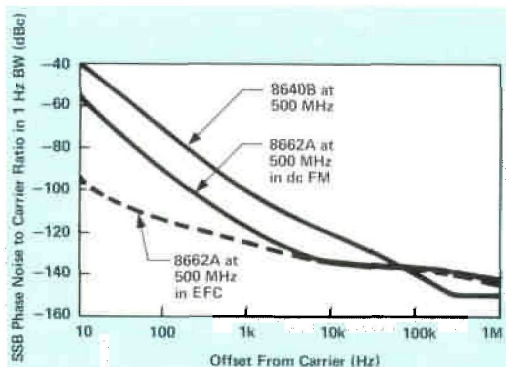


Figure 4.11. Comparison of the SSB noise of 3 sources for phase locking.

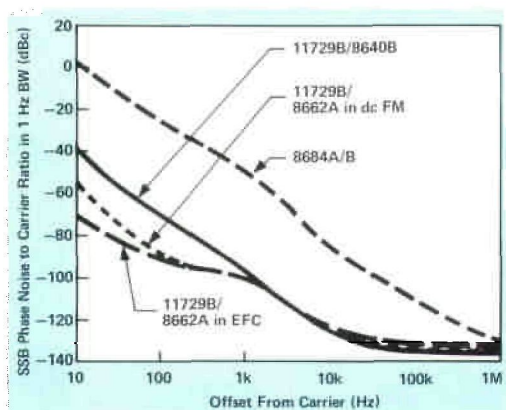


Figure 4.12. Effect of method of phase locking on system noise floor at 10 GHz.

LBW. (Note: a source other than the 8640B may be required if tuning range to 1280 MHz is necessary.)

An alternate scheme is to use the tuning capability (if available) of the source under test, as shown in Figure 4.10. Typical microwave VCO's have tuning sensitivities on the order of 10 MHz/volt. With the 11729B $\pm 1V$ Frequency Control signal, this would result in a typical loop holding range (VCO tuning range) of 10 MHz.

Effect of Phase Locking Method on System Noise

Absolute system noise as a function of offset frequency, as discussed in Appendix C, is always a function of the noise contributions from 3 sources: the noise on the 640 MHz reference signal (multiplied to microwave), the noise on the RF source added at the phase detector, and the residual noise of the 11729B. The method of phase locking can affect the noise of the RF source used at the phase detector, and therefore the noise of the measurement system.

When phase locking by tuning via the EFC of the 8662A reference oscillator, neither the noise performance of the 640 MHz signal nor the 8662A front panel signal is affected. The noise of the front panel signal of the 8662A at 500 MHz is shown in Figure 4.11.

When phase locking by using the 8662A dc FM, the noise of an internal free-running 140 MHz oscillator is added to the synthesized front panel signal, slightly degrading the 11729B/8662A system noise floor. The resultant noise of the 8662A at 500 MHz in dc FM mode is also shown in Figure 4.11.

The noise of the 8662A FM oscillator does not affect the noise on the 640 MHz reference signal. However, since the 8662A multiplies and divides its fundamental 320 to 640 MHz band to achieve its other bands, the noise of the FM oscillator is also multiplied or divided. Therefore, the added noise of the FM oscillator is a function of selected 8662A frequency, and should be accounted for in calculation of system noise (Appendix C).

If the 8640B or other RF source is chosen as the locking source, its total noise will affect system noise. Also shown in Figure 4.11 is the noise of the 8640B at 500 MHz.

Figure 4.12 shows the 11729B/8662A system noise at 10 GHz for these three choices of phase locking. The graph shows the trade-off between system noise and maximum loop holding range and loop bandwidth. However, in general, DUT's requiring wider loop bandwidth are free-running microwave sources, with noise considerably higher than the noise on the free-running RF oscillators used for phase locking. For example, also shown in Figure 4.12 is the noise of the HP 8684B, a cavity-tuned microwave source. System noise when locking via 8662A dc FM or via the 8640B is well below the measured noise on the 8684B.

Loop Capture

In normal measurement operation it is desired to keep the 11729B phase lock loop (PLL) bandwidth as narrow as possible to have minimum loop noise suppression. However, if a narrow bandwidth PLL is chosen, it may be very difficult for the PLL to acquire lock after the calibration procedure. To facilitate locking the DUT, the 11729B has a Loop CAPTURE feature.

When Loop CAPTURE is pressed (or enabled over HP-IB), a wider first-order PLL enables the two signals to acquire lock. Then when the CAP-

TURE button is released, the normal second order loop is again engaged, the signals will be in quadrature, and a measurement can be made.

Loop Tune

Loop CAPTURE can also be used for frequency tuning. When the CAPTURE first order loop is enabled, the voltage out of the phase detector (phase error) is proportional to the **frequency** difference between the two input signals. The 11729B front panel LED PHASE LOCK INDICATOR display monitors that voltage. So when CAPTURE is held down (either manually or over the HP-IB), the PHASE LOCK INDICATOR can be used to tune the RF source to a frequency such that the inputs to the phase detector are within the loop holding range. There is also an HP-IB signal available indicating when the phase error is within the loop holding range.

For many DUT's, tuning the RF source at the phase detector into the center of the loop holding range is required to allow the source to drift the maximum amount in either direction and still remain locked. For DUT's exhibiting a drift characteristic in one direction, it may be desirable to tune the RF source to one end of the loop holding range. This allows the source to drift twice the maximum tuning range and remain locked.

Loop Characterization

As discussed in Chapter 3, a PLL (phase lock loop) forces the VCO to track the reference in phase for offsets less than the bandwidth of the PLL. This tracking inside the loop BW results in suppression of phase noise at the output of the phase detector. This property normally limits a phase noise measurement to offsets from the carrier greater than the loop BW. However, the 11729B enables the phase lock loop to be characterized, allowing corrected phase noise measurements to be made inside the LBW.

The 11729B Loop Test Input (as shown in Figure 4.13) allows a test disturbance — for example, a random noise source or a tracking generator — to be applied to the loop. Then, by measuring the response of the loop to the Test Input at the Loop Test Output, the transfer characteristic of the phase lock loop can be determined. (See Appendix B for a more thorough description of Loop Characterization.) During Loop Characterization, the loop remains locked and in quadrature; that is, the loop is characterized in the same state that it was in during the phase noise measurement.

The loop filter transfer function as displayed on the spectrum analyzer yields two important pieces of information. First, the loop BW can be determined, which designates the offset frequencies for which an uncorrected phase noise measurement can be made.

Secondly, the degree of noise suppression vs. offset frequency can then be used to correct the value of DUT noise measured when the system was in the normal loop hold mode. For example, Figure 4.14 shows a typical phase lock loop transfer characteristic, with a BW of about 100 Hz. At a 10 Hz offset, the loop suppresses the noise 20 dB. Prior to adding the test disturbance, the normal loop hold measurement on the DUT yielded a noise measurement of -90 dBc/Hz at 10 Hz. The loop noise suppression correction is added to this number, yielding the actual phase noise of the DUT at 10 Hz offset of:

measured noise level:	-90 dBc/Hz
loop suppressed noise:	+20 dB
actual noise level:	-70 dBc/Hz

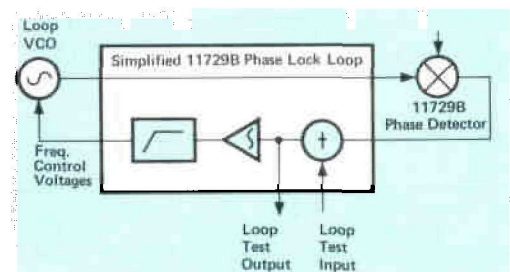


Figure 4.13. 11729B phase lock loop characterization.

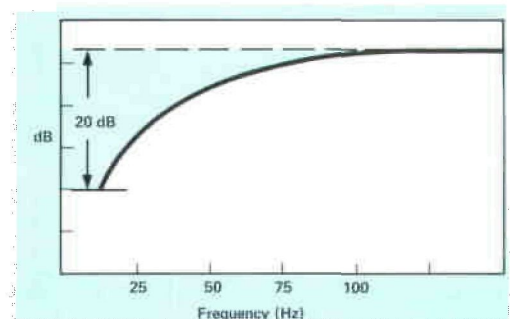


Figure 4.14. Typical phase lock loop filter transfer characteristic.

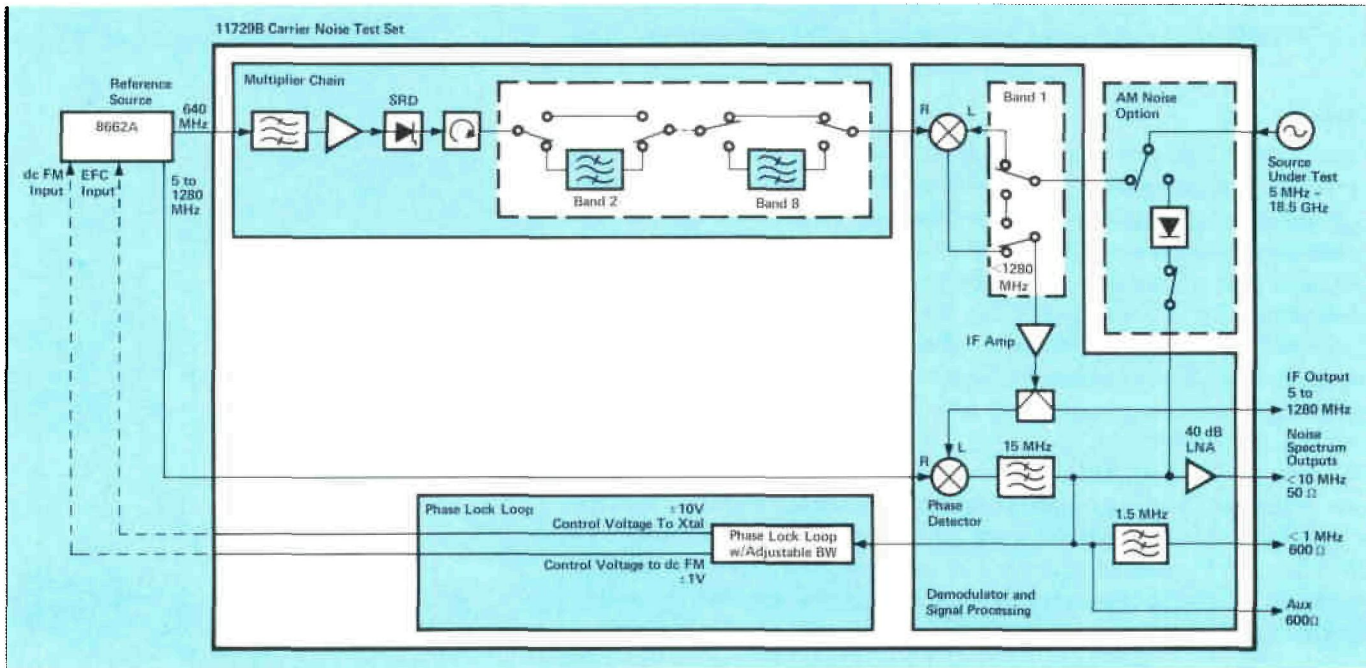


Figure 4.15. 11729B block diagram.

This chapter integrates the theory of the phase detector method (Chapter 3) and its implementation in the 11729B (Chapter 4) into procedures for making phase noise measurements on microwave sources. Specific instrument operation for the 11729B and 8662A as well as generalized instructions for spectrum analyzers are given. (Appendix D gives specific keystroke sequences for two Hewlett-Packard spectrum analyzers.) As in Chapter 3, the measurement steps will be broken down into five main procedures: 1) system setup; 2) calibration; 3) lock; 4) measure; 5) corrections.

Properties of the Source Under Test

To help determine system parameters (such as the necessary 11729B filter band and the method of phase locking to be used), certain information about the source under test is required before making a phase noise measurement.

If a microwave spectrum analyzer is available, first measure the output signal from the device under test (DUT) directly. If the phase noise sidebands of the DUT are sufficiently high relative to its AM sidebands and to the phase noise of the spectrum analyzer, direct measurement might be possible. Compare the level of the displayed noise sidebands with the phase noise specification of the spectrum analyzer. If the measured level is several dB greater than the noise of the spectrum analyzer, then the display is the phase noise of the DUT (provided the amplitude noise of the DUT is significantly below its phase noise). With sufficient spectrum analyzer dynamic range and resolution, direct measurement would be valid.

If a microwave spectrum analyzer is not available, the IF output of the 11729B (after first downconversion stage) can be used to examine the characteristics of the test signal on a lower frequency spectrum analyzer. Before connecting the DUT into the 11729B Microwave Test Signal input, verify approximate signal frequency and level. (CW microwave input level to the 11729B must be between +7 and +18 dBm.) This enables the proper microwave bandpass filter to be selected as well as protects the input mixer from damage. Input levels less than +7 dBm can be used with a degradation in system noise floor.

Connect the 640 MHz Auxiliary Output signal from the 8662A to the 11729B 640 MHz Input, and select the appropriate filter as shown on the 11729B front panel display. No lock connections are necessary at this time. If a signal $\geq +7$ dBm at a frequency between 5 MHz and 1280 MHz is obtained at the 11729B IF Output, the signal input is at a sufficient level, and the microwave chain and first downconversion stage are operating.

Examine the IF signal on a spectrum analyzer; measure the level of the noise sidebands. As in the case of the microwave spectrum analyzer, determine whether direct noise measurement is possible at the IF frequency. The 11729B first down conversion allows a microwave signal to be examined at a much lower frequency, where the LO of a spectrum analyzer might have substantially lower noise characteristics.

If direct measurement is not possible, examine the microwave DUT signal or the 11729B IF signal to determine the DUT's characteristics. Phase noise measurements can be very sensitive to a noisy environment, and proper system set-up is important. Is the source behaving as expected — i.e., not taking large random frequency jumps or exhibiting excessive drift.

If the source is not behaving as expected, a number of system factors should be examined. The DUT should be properly warmed up and heat sunk



Figure 5.1. Selecting phase noise measurement.

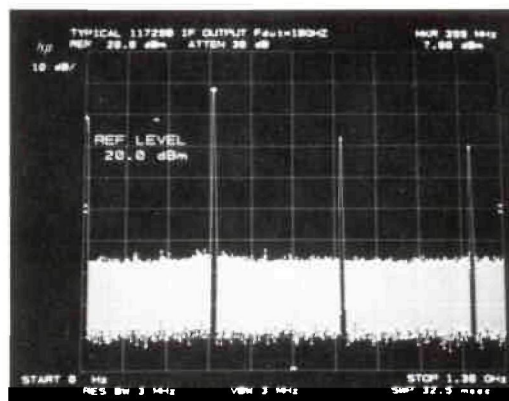


Figure 5.2. Typical 11729B IF output.

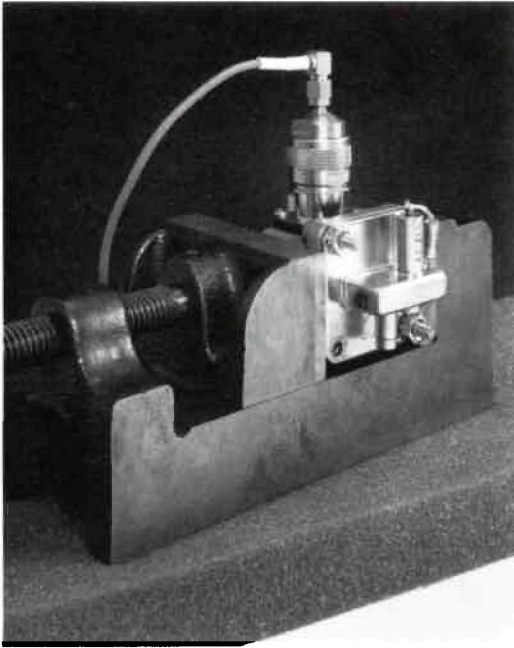


Figure 5.3. Typical oscillator setup for phase noise measurement.

to reduce drift. Induced microphonics should be kept to a minimum by tight connections; an isolator might be needed at the output of the DUT to reduce frequency pulling from movement of the connecting cable. Often, to reduce induced signals, it is necessary to have short lead lengths from the power supply to the device under test. The power supply might need to be regulated to avoid large line frequency sidebands from being induced into the measurement circuitry. Another alternative is to run the DUT off a battery.

Sometimes, microphonic isolation might be necessary; try placing the DUT on a vibration absorbing pad. Flex cable, though lossy at microwave, might minimize vibration transmission to the DUT. What is the RFI environment of the room; is there any RFI shielding? High power radio stations have been known to appear as spurious signals in the noise spectrum!

Considerations in Phase Locking

As mentioned in Chapter 4, a locked or stabilized DUT can usually be held in phase quadrature at the 11729B phase detector by phase locking through the 8662A Reference Oscillator. For free-running sources or locked sources with large level phase instabilities, increased loop holding range or wider phase lock loop bandwidth is needed. These sources can be locked via the dc FM of the 8662A or another source. If in doubt as to which lock technique to use, first try to lock via the 8662A Reference, as this results in lower system noise performance. If lock cannot be maintained, relock using dc FM.

What are the observed drift characteristics of the source? The drift must be less than the loop holding range (LHR) for the duration of the measurement. The total measurement can be broken into a series of successive spectrum analyzer sweeps; for example, 100 Hz to 1 kHz, 1 kHz to 10 kHz, 10 kHz to 100 kHz, etc. Then the DUT drift rate must be less than the time for one of these sweeps. From Chapter 4, we found $LHR (EFC) = 1 \text{ part in } 10^7$; $LHR (\text{dc FM } 8662A \text{ or third source}) = \text{maximum deviation per volt}$.

Also, any large phase instabilities must be at rates inside the loop BW. That is, the bandwidth must be selected large enough to track the large, low rate phase instabilities which would otherwise drive the phase detector out of its linear range. Again in Chapter 4, we found

$$\text{loop BW (EFC)} = \frac{f_{\text{dut}} \times 11729B \text{ LBF}}{10^{10}}$$

$$\text{and loop BW (dc FM)} = \frac{11729B \text{ LBF} \times \text{FM deviation}}{10^3}$$

If the source has phase instabilities such that it will not stay locked under either of these two locking procedures, try locking via a third source or via the DUT. If lock cannot be established, another method of phase noise measurement is recommended. Using a frequency discriminator might be appropriate. This can be done with the 11729B, or with the HP 3047A Spectrum Analyzer System, or with other instrumentation; see references for more information.

Phase Noise Measurement Procedures

This section discusses the measurement procedures to use with the 11729B in a phase noise measurement system, including key-stroke

sequences for the 11729B and 8662A and generic spectrum analyzer instructions.

For additional system operation, Appendix D gives specific instructions for using the HP 3582A and HP 8566A spectrum analyzers as the 11729B Noise Spectrum baseband analyzers. The 3582A covers 0.02 Hz to 25 kHz, and also features a random noise source output useful for characterizing the 11729B phase lock loop. The 8566A covers 100 Hz to 22 GHz, and is useful for measuring other parameters of the DUT as well as for use as a baseband analyzer. Other spectrum analyzers with the necessary sensitivity covering the offset frequencies of interest, such as the HP 3585A and 8568A, are also good alternatives.

System Set-Up

Before connecting to the noise measurement system, it is good practice to measure the frequency of the DUT or the resultant 11729B IF frequency with 100 Hz or 1 kHz resolution. This will determine the needed frequency on the tunable RF source so that it can be easily tuned to within the LHR.

Connect the system as shown in Figure 5.4, with the appropriate phase lock loop control voltage line.

1. Select carrier frequency.

Set the device under test to the desired carrier frequency, f_{dut} , at a level of +7 to +18 dBm. On the 11729B, select the appropriate bandpass filter (Figure 5.5), and a Lock BW Factor of 100.

2. Set the IF frequency on the tunable RF source.

Set the tunable source to 0 dBm. The frequency of this source (usually the 8662A) should be set to f_{IF} , which has either been directly measured or can be calculated from $f_{\text{IF}} = f_{\text{dut}} - f_{\text{filter}}$. Verify on the spectrum analyzers that any remaining beat note at the Noise Spectrum Outputs is less than the loop holding range of the method of phase lock that will be selected.

System Calibration

The calibration procedure involves measurement of a carrier level, so that noise can be measured relative to that level.

3. Set up beat note.

a) For synthesized sources, set up a beat note by offsetting the tunable RF source by a small Δf . For a locked DUT, set a 10 kHz frequency increment on the 8662A by pressing FREQUENCY INCREMENT SET 10 KHz. Set an amplitude increment by pressing AMPLITUDE INCREMENT SET 40 DB. Press INCREMENT DOWN (output should now be at -40 dBm). Press FREQUENCY INCREMENT UP.

Verify the presence of the beat note on the spectrum analyzers (Figure 5.6). The 8662A front panel signal provides the R port drive to the phase detector, and any change in level will be translated to the Noise Spectrum Outputs of the 11729B. The beat note will be the reference level for subsequent phase noise measurements, and represents the frequency translated carrier reduced by 40 dB.

b) For free-running sources, follow the same procedure but choose a larger beat frequency (50 or 100 kHz), that will not drift past 0 Hz during calibration.

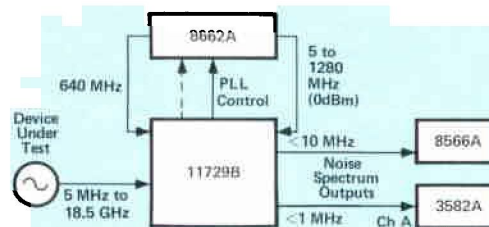


Figure 5.4. Typical phase noise measurement setup.



Figure 5.5. Selecting 11729B band.

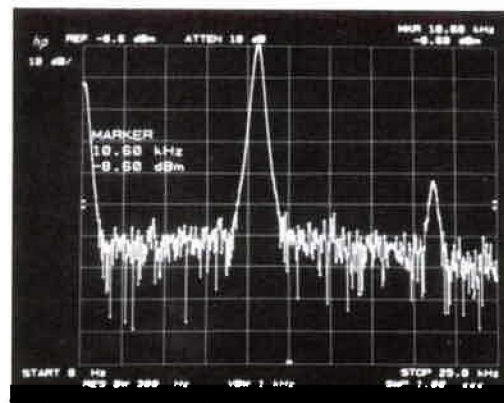


Figure 5.6. Calibration beat note.

4. Measure level of beat note.

Set up the spectrum analyzer to display the beat note. Best spectrum analyzer accuracy is obtained if the spectrum analyzer input sensitivity is not changed between Calibration and Measure. Therefore, it is recommended to adjust the spectrum analyzer input sensitivity so that the beat note is near the top of the screen (Figure 5.6). Establish a reference level by placing the beat note at the top of the screen, or setting a reference marker, or simply by recording the beat note amplitude V_b in dBm.

Lock

5. Lock up the sources in quadrature.

Remove the frequency and amplitude increments on the 8662A by pressing FREQUENCY INCREMENT DOWN, AMPLITUDE INCREMENT UP.

a) For sources locked via EFC

Set the Lock BW Factor (LBF) on the 11729B to 100. Press CAPTURE and verify that the signals lock. If they do not, use the CAPTURE—LOOP TUNE feature described in Chapter 4 to tune the 8662A to the center of the loop holding range. Release CAPTURE. If the source under test has the proper characteristics, the signals should now lock. If they do not, try the 1k LBF. If they still do not lock, reconnect for locking via dc FM.

If the sources remain locked in the 100 Lock BW Factor and close-in measurements are desired, try reducing the LBF, making sure the 11729B PHASE LOCK INDICATOR remains green or within the wide section (Figure 5.7). If lock is broken, increase the LBF and press CAPTURE to re-enable lock. Of course, for accurate measurements without correcting for loop suppression, the BW only needs to be reduced to below the lowest offset frequency of interest.

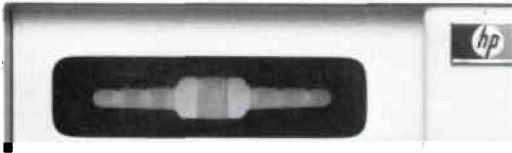


Figure 5.7. 11729B phase lock indicator.

b) For sources locked via 8662A dc FM

Set the 8662A to dc coupled FM, 50 kHz deviation. Select the 100 Lock BW Factor on the 11729B. (Selecting FM on the 8662A introduces a free-running FM oscillator into the system, which might cause a finite frequency jump on the output frequency of the 8662A.) The largest frequency deviation is not initially chosen because enabling LOOP CAPTURE will momentarily set up a wider band phase lock loop. If too large an FM deviation is initially chosen, the effective loop BW might be unstable. After phase lock has been established, a wider loop holding range (FM deviation) and BW (FM deviation and Lock BW Factor) can be selected.

Use the CAPTURE—LOOP TUNE feature to tune the 8662A so that the sources are in the center of the loop holding range. Release CAPTURE and observe if the signals remain locked. If they drift quickly out of lock, center them again in the loop holding range. Release CAPTURE, then immediately increase the FM deviation to 100 kHz. Again monitor lock.

Continue trying to establish a stable lock condition (i.e., setting up a wide enough LBW) by increasing FM deviation after loop centering. If there is not sufficient FM deviation to maintain lock, try increasing Lock BW Factor after loop centering. Remember that increasing lock BW factor by one position increases Loop BW by a factor of 10, while increasing FM deviation from 50 kHz to 100 kHz increases LBW only by a factor of two. Also, increasing FM deviation increases the Loop Holding Range, while the Lock Bandwidth Factor has no effect on LHR.



Figure 5.8. Using CAPTURE—LOOP TUNE to center the sources in the loop holding range.

c) Lock tracking

Lock up in the narrowest LBW possible to maintain quadrature, yet with the widest LHR to keep the DUT from drifting out of lock. Note that lock needs to be maintained only for one series of spectrum analyzer measurement sweeps; after one series, the 8662A can be retuned, and lock reestablished. This can be done in a manual measurement system by using the CAPTURE—LOOP TUNE feature to re-tune the 8662A using the PHASE LOCK INDICATOR display. Alternatively, a counter can be placed at the IF Output of the 11729B to monitor the frequency changes of the DUT and then use this information to increment/decrement the 8662A appropriately to center the LHR.

In automatic applications, there is an Out-of-Lock indicator provided over HP-IB that can be used for lock tracking.

Measure

6. Measure the phase noise on the spectrum analyzer.

After lock has been established, the level of the phase noise can be measured. Set up the spectrum analyzer span to cover the offset frequencies of interest. Increase the spectrum analyzer input sensitivity only if the displayed noise is too low to be measured. Select a Resolution BW that is in keeping with the chosen frequency span; for example, if a span from 0 Hz to 25 kHz is used, Resolution BW should be chosen ≤ 300 Hz.

Because phase noise is a random quantity, some sort of averaging is desired. If available, digital averaging can be selected; an alternative method is to use some analog averaging by reducing the video BW.

After averaging, take a reading V_S in dBm at the offset frequency of interest, noting the Resolution BW setting (Figure 5.9). Set other frequency spans and make measurements as desired.

In general, it is not advisable to take measurements on a portion of the spectrum analyzer display where the noise level is falling very rapidly (>20 dB per major division). For example, if a span from 0 Hz to 100 kHz was displayed, and the noise was falling rapidly around the desired 5 kHz measurement offset frequency, it would be better to set up a narrower span around 5 kHz and then measure.

It is not recommended to measure noise levels that are in the bottom 10 dB of the display. If possible, increase the spectrum analyzer input sensitivity and repeat the measurement.

Apply Corrections

7. Apply corrections to translate measured value to $\mathcal{L}(f)$

Subtract the level of the calibration signal V_b from the level of the noise measured. This is equal to the phase noise of the source under test with the corrections described in Chapter 3 and summarized again below.

$\mathcal{L}(f) = V_S$ minus V_b
minus 40 dB for the attenuation added during calibration
minus 6 dB to convert measured value to $\mathcal{L}(f)$,
minus $10 \log$ (equivalent noise measurement BW) dB,
plus 2.5 dB if an analog spectrum analyzer is used,
plus loop noise suppression correction at appropriate offset frequency.

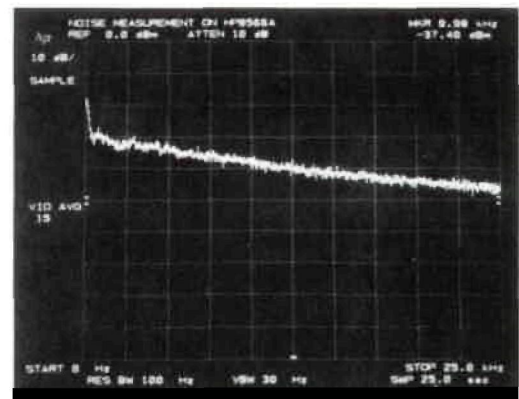


Figure 5.9. Typical noise spectrum output displayed on a spectrum analyzer.

Using the measurement examples as shown in Figures 5.6 and 5.9, the level of the beat note V_b was -8.6 dBm. The noise level at a 10 kHz offset V_s was measured at -35.8 dBm, in a 100 Hz nominal resolution bandwidth. Thus, since there was no loop suppression at this offset frequency, the SSB phase noise level on the 10 GHz carrier is

$$\begin{aligned}
 & -37.4 \text{ dBm } (V_s) \\
 & -8.6 \text{ dBm } (V_b) \\
 & -40 \text{ dB (calibration attenuation)} \\
 & -6 \text{ dB (effect of mixer)} \\
 & -20.8 \text{ dB (conversion to nominal 1 Hz noise bandwidth)} \\
 & +2.5 \text{ dB (since 8566A is an analog spectrum analyzer)} \\
 \hline
 & -93.1 \text{ dBc/Hz}
 \end{aligned}$$

8. Measure loop BW and loop noise suppression if necessary

To use the loop characterization features of the 11729B, connect the system as shown in Figure 5.10. The Loop Test Input can be a random noise source, a tracking generator, or even the variable frequency sine-wave output of a function generator. Loop Test Output is applied to the spectrum analyzer.

Apply the test disturbance with the loop in a locked condition (with the same Lock BW Factor and FM deviation as set during the noise measurement). Too high a level at the loop test input will cause large oscillations on the 11729B front panel PHASE LOCK INDICATOR. Adjust the input level such that very small oscillations around quadrature can be detected, and then decrease the level slightly.

Since we are adding a high level of noise into the phase lock loop, the amplitude levels of the Loop Test Output will generally be higher than those of the measured phase noise. Therefore, increase spectrum analyzer input attenuation, and span the desired offset frequencies. From the formulas given in Chapter 4, approximate LBW can be calculated; thus, a good starting point is to span from 0 Hz to 1.5 times the calculated LBW. The effective loop filter can then be traced out on the spectrum analyzer display, and the appropriate correction values can be measured and stored.

Measure these correction factors by determining that frequency where the loop test output level becomes constant. This will be outside the loop BW, where the loop no longer suppresses the phase noise. For offset frequencies greater than this, no correction is needed. Set a marker at this frequency, and note its level. Enter a delta marker or a relative mode if possible, and move the marker to lower offset frequencies. Measure and record the amount of loop noise suppression (Figure 5.11). The amount of loop noise suppression will be added to the measured phase noise value. If more exact information about the loop suppression is desired, the frequency span can be narrowed to the range of interest. However, for a random noise source input it is not recommended to take Loop Suppression measurements greater than a decade inside the PLL bandwidth. For characterization further inside the loop BW, a tracking generator or function generator input is recommended.

Note 1: Some phase lock loops will display a small amount of peaking at offsets greater than the corner frequency of the PLL. This amount of loop peaking should be SUBTRACTED from the measured value of phase noise at the appropriate offset frequencies.

Note 2: Be sure to disconnect the Loop Test Input signal before proceeding with another phase noise measurement.

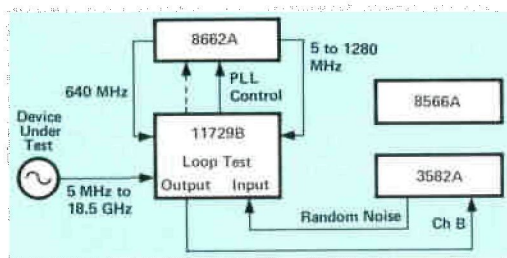


Figure 5.10. Sample system setup for loop characterization.

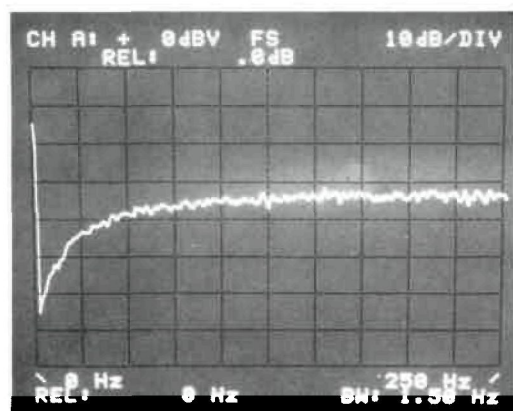


Figure 5.11. Measurement of loop suppression.

After configuring a phase noise measurement system, it is often desirable to determine the accuracy of the measurement. This chapter will discuss some of the elements that can affect overall system accuracy, as well as typical system accuracy for a phase noise measurement system consisting of an 11729B/8662A/3582A/8566A. With careful system design, phase noise measurements can be made to typical accuracies of less than ± 2.5 dB. (Even without extensive correction routines, typical accuracies of ± 3 to ± 5 dB can be expected.) The overall accuracy is a function of 1) the accuracy of the instrumentation used to measure the noise and 2) certain system parameters of the 11729B/ 8662A. Looking at the individual contributions to system accuracy isolates the areas where accuracy can be improved.



Figure 6.1. Typical phase noise measurement system.

1. Spectrum Analyzer Accuracy

The MAJOR contributor to system accuracy is the relative amplitude accuracy of the spectrum analyzer used to measure the SSB phase noise. Spectrum analyzer accuracy is usually a function of several components, and can be minimized with certain measurement techniques. Also, since this system makes a noise measurement RELATIVE to the carrier level, overall accuracy is better than if absolute noise levels were measured.

3582A:

For the 3582A, amplitude accuracy = sum of the accuracy at the passband center and the filter accuracy as given below.

AMPLITUDE ACCURACY (full scale):

- Absolute Accuracy at Passband Center: ± 0.5 dB
- Flat Top Filter: +0, -0.1 dB
- Hanning Filter: +0, -1.5 dB
- Uniform Filter: +0, -4.0 dB

8566A:

For the 8566A, amplitude accuracy varies significantly with measurement procedure. For example, if absolute noise values were required, overall spectrum analyzer inaccuracy could be worse than 6 dB. However, proper measurement technique as discussed in this chapter can significantly improve that accuracy. The 8566A Amplitude Fidelity specification affects the measurement accuracy as follows:

Fidelity Log:

Incremental	Cumulative
± 0.1 dB/dB over 0 to 80 dB display	≤ 1.0 dB max over 0 to 80 dB display, 20-30 C
	≤ 1.5 dB max over 0 to 90 dB display

2. Accuracy of Calibration (Attenuator Contribution)

The phase noise measurement system measures noise relative to a carrier level. Since two numbers are subtracted for a relative measurement, careful system operation during the phase noise calibration step can improve overall system accuracy.

During calibration, the R port to the phase detector is attenuated 40 dB. This 40 dB step in attenuation during Calibration has an absolute accuracy not associated with the spectrum analyzer. The accuracy to which the reference level is set will be dependent only on the accuracy of the attenuation step.

If the 8662A is used for this signal, the absolute amplitude accuracy of the 8662A (including flatness, attenuator accuracy, ALC detector error and measurement uncertainty) is ± 1 dB. Thus, assuming mixer and spectrum analyzer flatness to be $\ll 1$ dB (or calibrated out), the beat note represents the frequency translated carrier reduced in level by 40 dB, with ± 1 dB accuracy.

The relative accuracy between two amplitude levels on the 8662A is even better than 1 dB. Much of this relative inaccuracy can be further calibrated out by using a power meter monitoring 8662A output power to accurately set the attenuation in the Calibration step. Alternatively, a “standards lab” attenuator could be used at the output of the 8662A.

Note: it is generally advisable to adjust spectrum analyzer controls such that a reference level is in the top 10 dB of the analyzer display. Then, the noise measurement has effectively 80 dB of dynamic range in the display (minus top and bottom 10 dB), plus the 40 dB in the calibration stage, allowing noise measurements to -126 dBc/Hz without changing the spectrum analyzer controls. (-120 dB minus the 6 dB correction factor from the use of a mixer as a phase detector.)

If a wider spectrum analyzer measurement bandwidth is used, for example 1 kHz, then an effective dynamic range of 156 dB is obtained. (80 dB of display, 40 dB from the Calibration stage, 6 dB for the operation of the mixer as a phase detector, and 30 dB for conversion of the 1 kHz measurement bandwidth to a 1 Hz bandwidth.) Since this is less than the 11729B/8662A system noise floor, this dynamic range should be sufficient for most measurements.

Changing spectrum analyzer settings between calibration and measurement generally results in additional errors. If necessary, redo the Calibration step with increased attenuation added such that the noise measurement can be made without changing spectrum analyzer settings.

3. Phase Detector Flatness

A Calibration step should be made for each new test frequency (that yields a different IF frequency into the 11729B phase detector) to eliminate the small error associated with the flatness of the phase detector.

4. 11729B Baseband Signal Processing Section Flatness

The 11729B Signal Processing Section is typically flat to within ± 1.0 dB (1 Hz to 10 MHz). If a very flat spectrum analyzer or other measurement instrument is available, this inaccuracy can be reduced by one of two methods. If a noise measurement at only a few offset frequencies is desired, a Calibration step (i.e., setting up a reference level) could be done at each offset frequency of interest. For a more complete error correction, the 11729B signal processing section can be swept characterized. This swept characterization as a function of frequency may be done by applying a varying frequency into the Signal Processing

section, and measuring the resultant output signals at the <1 MHz and <10 MHz Noise Spectrum Output ports.

5. Mixer dc Offset

Some very small error is introduced because the mixer will actually operate slightly out of quadrature (error = $20 \log \cos \theta$). For operation within 5 degrees of quadrature, the slight mixer dc offset will affect accuracy by less than 0.1 dB.

6. Absolute Noise Accuracy due to System Noise

As discussed in Chapter 3, the noise measured at the output of the phase detector is equal to the combined noise of the two sources. Thus, the finite noise contribution of the 11729B/8662A does contribute to the accuracy to which the DUT noise can be measured. This error is given by

$$\text{error (dB)} = 10 \log \left(1 + \text{antilog} \frac{\mathcal{L}_{\text{ref}} - \mathcal{L}_{\text{dut}}}{10} \right)$$

This error is tabulated in the following table for several values of noise power differences.

$\mathcal{L}_{\text{dut}} - \mathcal{L}_{\text{ref}}$ (dB)	0	1	2	3	4	5	10	15
correction (dB)	3.0	2.5	2.1	1.8	1.5	1.2	0.4	0.2

This error can be corrected out by accurately characterizing the noise of the 11729B/8662A system, and then using this known value of noise to correct for the measured value as shown in the table. The noise of the 11729B/8662A can be characterized using what is commonly called the ‘three source comparison’. (Note that the noise of the 11729B/8662A can only be characterized if a cleaner microwave source is available.) If the noise of three sources is unknown (but within about 10 dB of each other), three sets of pair-wise measurements can be made, and the absolute value of the noise of each source calculated. See HP Application Note 283-1 for the equations to calculate the phase noise of three unknown sources.

7. Accuracy of Measurement of Loop Noise Suppression

For a ‘corrected’ phase noise measurement — a measurement made inside the bandwidth of the phase lock loop — the accuracy of measuring the loop noise suppression also contributes to total measurement accuracy. This accuracy is a function of the flatness of the test signal input, the relative amplitude accuracy of the spectrum analyzer, and how far within the loop bandwidth it is desired to measure. Assuming that the loop test signal is flat (or that its flatness has been calibrated out), and that correction is only made to a decade within the loop bandwidth, the accuracy to which the suppression can be measured is equal to the relative spectrum analyzer accuracy.

If the Loop Test Input is a random noise source, measurements greater than a decade inside the loop bandwidth are subject to increasingly higher inaccuracy. In general, the loop characterization can be made further inside the loop bandwidth if a tracking generator or variable

function generator is used as the Loop Test Input. Alternatively, a model for the phase lock loop can be developed (see Appendix B), and the model used to correct for loop suppression decades inside the loop bandwidth.

For best accuracy, be sure to measure the loop response outside of the loop bandwidth as well as inside, to examine the effects of loop peaking. A critically damped loop will have about 2.2 dB of peaking. If this peaking is not calibrated out, measurements in the area of loop peaking will have this additional 2 dB of error.

Overall Accuracy

With the seven components of system inaccuracy listed above, the overall accuracy for a phase noise measurement can be calculated. This, of course, assumes that all necessary correction factors to the noise measurement (such as normalization to a 1 Hz Noise bandwidth, 6 dB correction for noise foldover, etc.) have been properly made.

The Calibration signal (V_b pk) can be measured to xx dB, where xx = accuracy of calibration attenuation step plus repeatability of the spectrum analyzer.

In general, since the Calibration step begins with the displaying of V_b , the only errors which will result are those of subsequent CHANGES in the attenuator settings and relative errors to the V_b measurement. Errors which remain the same as when the V_b measurement was made will not appear in the result.

First, let's examine the typical accuracy that can be obtained if no extra effort is made to calibrate out system errors. Assume that the system noise of the 11729B/8662A is >10 dB below the noise of the DUT, and that sufficient attenuation is added during Calibration so that the spectrum analyzer attenuation setting is not changed between Calibration and Measure. The 11729B Phase Detector Flatness is not an error source because the Calibration is performed at the IF of interest. Using the 8662A attenuator, the calibration is made to ± 1 dB accuracy, plus the repeatability of the spectrum analyzer. V_s can be measured to the accuracy of the spectrum analyzer. The total resultant errors are tabulated as follows:

Table 6.1. Overall System Accuracy

1) Accuracy of calibration step	± 1 dB + spectrum analyzer repeatability (typically a few tenths dB)
2) Relative accuracy of measuring V_s	relative accuracy of spectrum analyzer
3) Phase Detector Flatness	—
4) 11729B Baseband Signal Processing Section Flatness	± 1 dB
5) Mixer dc Offset	± 0.1 dB
6) Accuracy due to System Noise	± 0.2 dB
7) Accuracy of Loop Noise Suppression (with loop peaking measured)	relative accuracy of spectrum analyzer

Accuracy without Error Correction

Therefore, typical system accuracy without extra error correction using the two HP analyzers discussed would be:

For the 3582A:

Outside Loop Bandwidth: +3, -4dB
Inside Loop Bandwidth: +3.5,-6.5 dB

For the 8566A:

Outside Loop Bandwidth: ± 3.5 dB
Inside Loop Bandwidth: ± 5 dB

Potentially, some of these errors would occur in the same direction and cancel. Of course, the error or correction factor due to system noise varies with the relative noise level of the DUT and the system.

These numbers are WORST CASE assuming that all errors add in the worst case way. A more realistic approximation can be obtained by examining each inaccuracy and its cause to determine if it is random or systematic. Often, this results in relative measurements having errors that partially cancel out. For a probabilistic error estimate, some errors could be combined by a root sum of the squares method, instead of by simple addition.

Also remember that phase noise is a random quantity of which any measurement is only an estimate. Averaging, whether video or digital, significantly improves the accuracy and repeatability of a random measurement. Though a single sweep can be measured with the summation of the accuracies given above, this single sweep does not characterize the statistical randomness of the signal.

Accuracy with Error Correction

Careful measurement procedure can reduce the error in the Calibration step, and characterize the flatness of the 11729B Baseband Signal Processing Section. Therefore, overall system accuracy could probably be improved by about 1 to 1.5 dB, for those applications where high accuracy is needed.

For example, in a typical system the Calibration step can be made to ± 0.4 dB accuracy (typical accuracy of attenuators = $\pm 0.1/10$ dB), and the 11729B Signal Processing Section flatness can be measured to ± 0.5 dB.

Thus, typical system accuracies (still assuming errors add in the WORST CASE way!) using the two HP analyzers would be:

For the 3582A:

(Outside Loop Bandwidth): +1.8, -1.9 dB

For the 8566A:

(Outside Loop Bandwidth): ± 2.3 dB.

7 Using the 11729B with the HP 3047A for Automatic Measurements

The HP 3047A Spectrum Analyzer System provides calibrated, automatic measurement and analysis of phase noise. It contains an HP 3582A Spectrum Analyzer (20 mHz to 25 kHz), an HP 3585A Spectrum Analyzer (20 Hz to 40 MHz), an HP 35601A Spectrum Analyzer Interface (contains phase detector and phase lock circuitry), and an HP 9845B or 9836A Desktop Computer. Also included is system software for automatic system calibration, testing and graphic output.

The 3047A provides three modes of phase noise analysis: 1) direct spectrum mode for test frequencies to 40 MHz, 2) phase noise analysis using an external frequency discriminator, and 3) phase noise analysis using the phase detector method. Overall 3047A measurement system accuracy when used in the phase detector mode is ± 2 dB to 1 MHz offsets, ± 4 dB for 1 MHz to 40 MHz offsets.

Figure 7.1 repeats the basic set-up for the phase detector method. There are three fundamental components of this method: 1) the reference source, 2) the phase detector/quadrature maintenance circuitry, and 3) the baseband analysis section. Note that the 3047A provides two of these components: the phase detector/quadrature section and the baseband analysis section. However, a phase noise measurement system using the 3047A in the phase detector mode still requires an external reference. The 11729B/8662A combination can be used to provide the required low noise microwave reference.

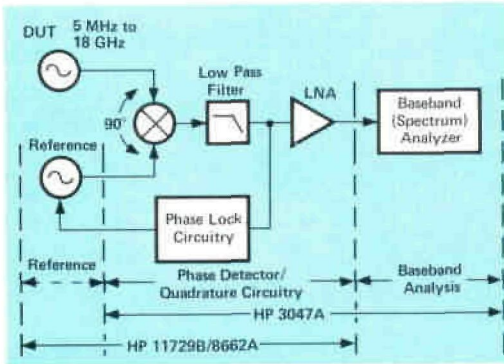


Figure 7.1. Basic phase detector method.

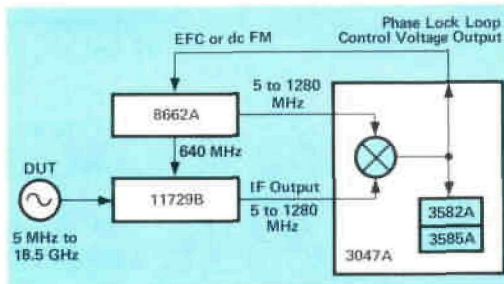


Figure 7.2. 11729B/8662A/3047A system set-up.

System Set-Up

The 11729B/8662A can be configured with the 3047A as shown in Figure 7.2. Though the 11729B has its own phase detector/quadrature section, the 3047A's phase detector/phase lock section is used when these instruments are used together. The phase detector/quadrature section of the 3047A is used because the 3047A software includes procedures for calibration and error analysis on this resident circuitry. The 11729B is used primarily as a low noise downconverter, to translate the microwave signal into an IF range that can be input to the 3047A. This IF signal is then mixed in quadrature with a suitable low noise RF signal, exactly as is done in the 11729B systems previously described.

System Operation

Because the 11729B/8662A are used primarily to downconvert the microwave DUT under test, the additional software necessary to the existing 3047A software is minimal. The only software command necessary to the 11729B is to select the phase noise measurement mode, and the correct microwave bandpass filter, F2 thru F8.

Since the 8662A is used as the source into the phase detector and also sometimes as the VCO for phase lock, a few more HP-IB commands might be needed. First, the 8662A frequency must be selected to provide the IF signal. The microwave signal under test must be known to within 40 MHz. Knowing the selected comb frequency, and the selected IF signal, the resultant beat note will be within the 40 MHz spectrum analyzer range. (As before, it is useful to first measure the microwave DUT frequency and amplitude before input to the 11729B. This will prevent damage to the instruments as well as enable easy selection of the IF frequency to within 40 MHz.)

When using the internal 11729B phase detector, the 8662A provides the R port drive to the phase detector. For low system noise floor, the 3047A uses a high level mixer as a phase detector. Thus, when using the 11729B/8662A with the 3047A, the IF signal out of the 11729B is the R port, and the 8662A provides the L port drive. The 8662A amplitude should be set to the +16 dBm overrange level.

The 3047A software will guide the user through the measurement, subject to certain user-input data. This includes frequency of device under test, phase detector input frequency (=IF Frequency out of 11729B), VCO tuning range, VCO voltage tuning requirements, measurement limits, graphical limits, etc.

When using the 3047A and locking through the dc FM of any source, the maximum FM deviation available should be selected. This allows the most stable bandwidths to be selected, and the 3047A provides mathematical algorithms to correct for the loop far inside the loop BW.

Once the initial parameters have been set, the 3047A will instruct the user to connect the Loop Control Voltage Output to the VCO input. This input will either be the 8662A EFC port ($\pm 10V$), the 8662A dc FM port ($\pm 1V$), or the dc FM port of whatever source is being used as the VCO. The 8662A FM deviation can either be set automatically or manually. Then if a beat note <20 MHz is measured on the 3585A Spectrum Analyzer, the 3047A will continue with the measurement.

The 3047A might prompt the user to manually tune one of the sources so that the beat note is within the tuning range of the VCO. The second signal input to the phase detector is the usual choice for tuning, and this is often the 8662A. This tuning can be done either manually, or an additional software driver can be added to measure the beat note and increment or decrement the 8662A as needed.

Overall System Accuracy

The 3047A measurement system accuracy is ± 2 dB to 1 MHz offsets, ± 4 dB to 40 MHz offsets. This does not include the potential noise contribution of the noise of the reference source, as discussed in Chapter 3. If the noise of the reference is 10 dB below the noise of the DUT, the actual phase noise measured is about 0.4 dB in error. If the noise of the reference is significantly below the noise of the DUT, the additional measurement error is small. If the margin between reference noise and DUT noise is small or unknown, and three sources are available, the 3047A has automatic software to compute the absolute noise level of each of the three sources. Once the absolute noise of the reference is known, its noise contribution can be subtracted from all future noise measurements.

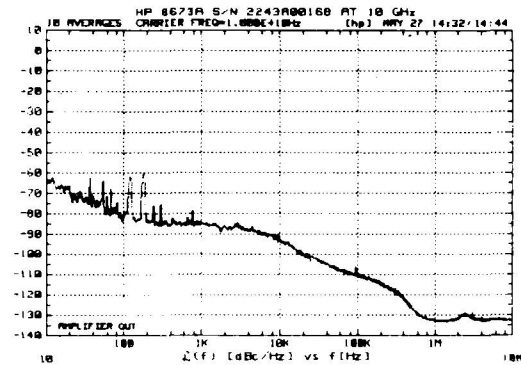


Figure 7.3. Typical 11729B/8662A/3047A output.

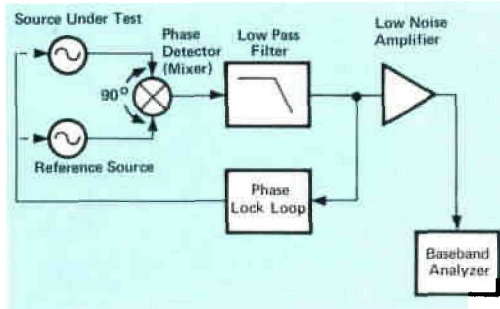


Figure A.1. Basic phase detector method.

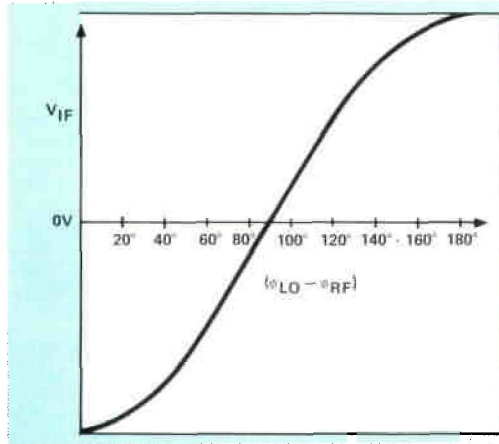


Figure A.2. Typical double-balanced mixer phase detector characteristic.

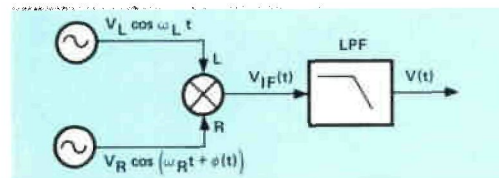


Figure A.3. Mixer operation.

The basic block diagram of the phase detector method of phase noise measurements is shown in Figure A.1. The basis of this measurement is the double balanced mixer used as a phase detector. Therefore, it is useful to first examine the operation of a mixer used in this way.

Figure A.2 shows a typical mixer-phase detector characteristic. When operated as a phase detector, the mixer outputs a voltage $V(t)$ proportional to the fluctuating phase difference between the two input signals $\phi_{LO} - \phi_{RF}$. The point of maximum phase sensitivity (the greatest voltage change per degree of phase change) and the center of the region of most linear operation occur where the phase difference between the two inputs is equal to 90 degrees, or phase quadrature.

To understand how a mixer operates as a phase detector, let's first examine a normal mixer output (Figure A.3).

Let the L port signal be described by $V_L \cos \omega_L t$, and the R port signal by $V_R [\cos \omega_R t + \phi(t)]$. Then the output of the mixer $V_{IF}(t)$ is the product of the two signals:

$$1) \quad V_{IF}(t) = K_L V_R \cos |(\omega_R - \omega_L) t + \phi(t)| + K_L V_R \cos |(\omega_R + \omega_L) t + \phi(t)| + \dots$$

The low pass filter in the block diagram of Figure A.3 removes the higher frequency components, leaving $V(t)$, as shown in Figure A.4.

$$2) \quad V(t) = K_L V_R \cos |(\omega_R - \omega_L) t + \phi(t)|$$

Let the peak amplitude of $V(t)$ be defined as $V_{b \text{ peak}}$ (peak voltage of the beat signal), equal to $K_L V_R$, where K_L = mixer efficiency.

$$3) \quad V_{b \text{ peak}} = K_L V_R$$

Then,

$$4) \quad V(t) = \pm V_{b \text{ peak}} \cos |(\omega_R - \omega_L) t + \phi(t)|$$

When operating the mixer as a phase detector, the input signals must be at the same frequency and 90° out of phase. That is, at quadrature,

$$5) \quad \omega_L = \omega_R, \text{ and } \phi(t) = (k + 1) 90^\circ + \Delta\phi(t)$$

Therefore, substituting in equation 4, the output of the mixer at quadrature is described by

$$6) \quad \Delta V(t) = \pm V_{b \text{ peak}} \sin \Delta\phi(t),$$

where $\Delta V(t)$ = instantaneous voltage fluctuations around 0 V, and $\Delta\phi(t)$ = instantaneous phase fluctuations.

For $\Delta\phi_{\text{peak}} \ll 1$ radian, $\sin \Delta\phi(t) \cong \Delta\phi(t)$, and equation 6 becomes

$$7) \quad \Delta V(t) = \pm V_{b \text{ peak}} \Delta\phi(t)$$

Note that this yields a direct linear relationship between the voltage fluctuations at the mixer output and the phase fluctuations of the input signals, or

$$\Delta V = K_d \Delta\phi$$

where $K_d = V_{b \text{ peak}} =$ phase detector constant (volts/radian), which is equal to the slope of the mixer sine wave output at the zero crossings.

To determine this phase detector constant, K_d , the mixer is operated not in quadrature, but with the inputs at two different frequencies, resulting in $V(t)$ as described in equation 4. The IF output signal measured on a spectrum analyzer provides the rms value of the signal ($V_b \text{ rms}$). The phase detector constant K_d , equal to $V_p \text{ peak}$ is the measured value $V_b \text{ rms} \times \sqrt{2}$.

When the mixer is again operated as a phase detector (input signals in quadrature), the voltage output of the mixer as a function of frequency will be directly proportional to the input phase deviations from equation 7.

$$8) \quad \Delta V(f) = K_d \Delta \phi(f)$$

$$9) \quad \Delta V(f) = \sqrt{2} V_b \text{ rms} \Delta \phi(f)$$

Then $\Delta \phi_{\text{rms}}(f)$ as measured on the spectrum analyzer is

$$10) \quad \Delta \phi_{\text{rms}}(f) = \frac{1}{K_d} \Delta V_{\text{rms}}(f) = \frac{1}{\sqrt{2} V_b \text{ rms}} \Delta V_{\text{rms}}(f)$$

In Chapter 2, the spectral density of phase fluctuations $S_{\Delta \phi}(f)$ was defined as

$$S_{\Delta \phi}(f) = \frac{\Delta \phi_{\text{rms}}^2}{\text{bandwidth used to measure } \Delta \phi_{\text{rms}}} \frac{\text{rad}^2}{\text{Hz}}$$

Therefore

$$S_{\Delta \phi}(f) = \Delta \phi_{\text{rms}}^2(f) = \frac{1}{2} \frac{\Delta V_{\text{rms}}^2}{V_b \text{ rms}^2} \quad (\text{in 1 Hz measurement BW})$$

For $\Delta \phi \ll 1$ radian,

$$\mathcal{L}(f) = 1/2 S_{\Delta \phi}(f) = \frac{1}{4} \frac{\Delta V_{\text{rms}}^2}{V_b \text{ rms}^2}$$

Or in logarithmic terms:

$$\mathcal{L}(f) \text{ dBc} = 10 \log \left(\frac{1}{4} \frac{\Delta V_{\text{rms}}^2}{V_b \text{ rms}^2} \right)$$

$$= 20 \log \Delta V_{\text{rms}} - 20 \log V_b \text{ rms} + 10 \log 1/4$$

$$= V_s \text{ (dBm)} \quad \text{level of the phase noise spectrum in a 1 Hz bandwidth}$$

$$- V_b \text{ rms (dBm)} \quad \text{level of the calibration beat signal}$$

$$-6 \text{ (dB)} \quad \text{accounts for rms value of beat signal and conversion of } S_{\Delta \phi}(f) \text{ to } \mathcal{L}(f)$$

In practice, other correction factors may be necessary as explained in Chapter 3. In particular, there is often a correction for attenuation added during calibration, and for effects of the spectrum analyzer itself on measured noise.

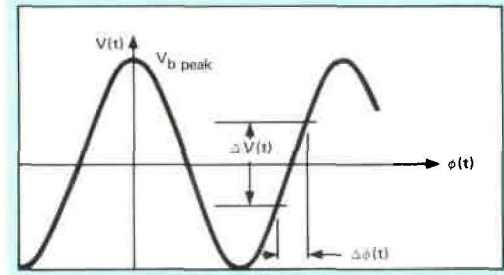


Figure A.4. Filtered mixer output.

The Importance of Quadrature

As shown in Figure A.2, phase quadrature is the point of maximum phase sensitivity and the region of most linear operation. Any deviation, ($\Delta\phi$), from quadrature results in a measurement error given by

$$\mathcal{L}_e = \text{error (dB)} = 20 \log |\cos(\text{magnitude of the phase deviation from quadrature})|$$

where error is defined as $\mathcal{L}(f)$ measured - $\mathcal{L}(f)$ at $\Delta\phi = 0$ in dB. Notice that the error in dB is always negative, since $\mathcal{L}(f)$ measured is always $\leq \mathcal{L}(f)$ at $\Delta\phi = 0$.

This error contribution is very small for small deviations around quadrature, as shown in the table below.

<u>Offset from quadrature</u>	<u>Error</u>
1°	- 0.001 dB
3°	- 0.01 dB
10°	- 0.13 dB

Even though the error for small deviation around quadrature is small, in a user-designed noise measurement system, this deviation from quadrature would have to be monitored. However, the 11729B's quadrature maintenance section monitors and ensures good phase quadrature, freeing the measurement from the errors associated with maintaining quadrature.

General System Precautions

As mentioned in Chapter 3, care should be exercised when making phase noise measurements. The following potential problems should be considered to avoid erroneous data.

The first group of potential errors can be minimized by careful procedure. The second group are minimized by the 11729B, making phase noise measurements easier.

- * Non-linear operation of the mixer will result in a calibration error.
- * The amplifier or spectrum analyzer can become saturated during calibration or by high spurious signals such as line frequency multiples.
- * Closely spaced spurious may give the appearance of continuous phase noise when insufficient spectrum analyzer resolution and averaging are selected.
- * Vibration can excite significant noise in device under test.
- * Suppression or peaking of phase noise close to the phase lock loop bandwidth.
- * A non-sinusoidal beat note causes an error in the calibration constant.
- * Impedance interfaces should remain unchanged between calibration and measurement.
- * Noise from power supplies can be a dominant contributor to phase noise.
- * AM noise of the DUT signal can be measured if the quadrature condition is not maintained.
- * Injection locking of the DUT can occur.
- * Deviation from phase quadrature

Phase Lock Loops: Advanced Considerations

B

The input signals to the 11729B phase detector are maintained in quadrature by the 11729B internal phase lock loop. This loop, during normal measurement operation, is a second order phase lock loop and can be further modeled as shown in Figure B.1.

In this model, under normal locked conditions, the phase detector output V_d follows the phase fluctuations of the device under test, ϕ_{DUT} , according to the following equation:

$$V_d(s) = \phi_{DUT} \frac{K_d}{1 + \frac{2\pi}{s} K_o F K_a(s) K_d} = \phi_{DUT} \frac{K_d}{1 + G_{OL}}$$

where V_d = phase error voltage = output of the mixer, proportional to the fluctuating phase difference between the inputs,

K_d = phase slope or phase detector gain factor (volts/rad) (= calibration constant),

K_o = VCO slope (Hz/volt),

$\phi_o(s)$ = phase error,

F = 11729B Lock Bandwidth Factor,

$K_a(s)$ = loop amplifier gain,

$V_c(s)$ = VCO control voltage (or phase lock loop control voltage),

$s = 2\pi j f$ (Hz) (where f = offset frequency)

and $V_{ti}(s)$ = the loop test perturbation input which is = zero in normal locked operation

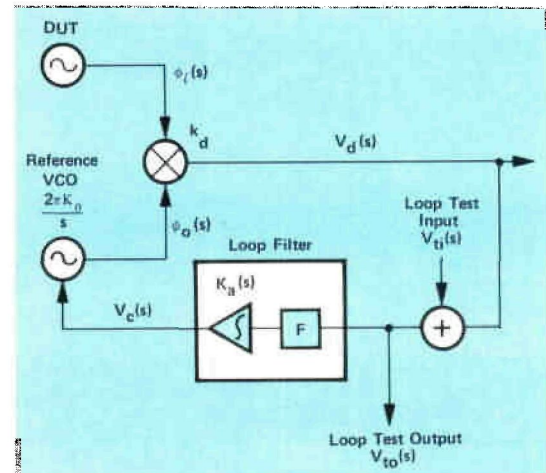


Figure B.1. Second order phase lock loop model.

Loop Parameters

There are four ways to hold the phase detector input signals in quadrature, with one of the phase detector inputs acting as a VCO. The phase lock loop parameters vary with the method of phase lock chosen, and we will look at each model individually.

Analyzing the loop models will yield the two important lock parameters, LHR (loop holding range) and LBW (loop bandwidth). The LHR is set by the tuning range of the VCO.

The loop bandwidth (or loop gain) is variable, and is a product of the loop gains: K_d = the phase detector gain, F the 11729B front panel Lock Bandwidth Factor, $K_a(s)$ the loop amplifier gain (or loop transfer characteristic), and K_o the VCO slope.

Note that the phase detector gain K_d is affected by the amplitude of the VCO. Since the VCO signal is the RF input to the phase detector (mixer), changing the VCO amplitude changes K_d and thus the total loop gain. The bandwidths computed in this section ASSUME THAT THE AMPLITUDE LEVEL OF THE VCO IS AT 0 dBm. In actual measurements, if for example, the VCO amplitude is changed to -6 dBm, the loop bandwidth will be halved from that computed here.

In general, the 11729B open loop gain can be described by

$$G_{OL} = K_d F K_a(s) K_o 2\pi/s.$$

The loop bandwidth is that frequency where the open loop gain is equal to one.

$$G_{OL} = 1 = K_d F K_a(s) K_o/f.$$

Then $f = \text{loop BW} = K_d F K_a(s) K_o$.

The 11729B provides two frequency control voltages $|V_c(s)|$ for these locking mechanisms.

Phase Locking via the 8662A Reference Oscillator

LHR

Locking through the EFC of the 8662A crystal oscillator can be more completely modeled by Figure B.2, which includes a xN block in the loop feedback. Note that the amount of frequency tuning range on the crystal determines the loop holding range at any test frequency. The nominal ± 1 Hz tuning at the 10 MHz Reference is multiplied up to ± 64 Hz on the 640 MHz signal, and then multiplied again in the 11729B multiplier chain. This is modeled by the xN block in the loop model, where $N = f_{dut}/10$ MHz.

For example, the nominal effective loop holding range at 10 GHz is ± 1 kHz, while at 5 GHz the nominal loop holding range is ± 500 Hz (100 Hz per GHz of carrier). Nominal loop holding range (LHR) when locking through the EFC of the 8662A can be expressed simply by

$$\text{nominal LHR (EFC) (Hz)} = f_{dut}/10 \text{ MHz.}$$

LBW

As in the general case, the 11729B loop bandwidth when locking via EFC is a function of the product of the loop gains. In this loop model, the xN also figures into the loop gain.

$$\text{The open loop gain is } G_{OL} = \frac{K_d F K_a(s) K_o 2\pi N}{s}$$

When locking via the 8662A EFC, $K_d K_a K_o$ is equal to 10^{-3} . The loop bandwidth ($G_{OL} = 1$) is therefore $FN/10^3$, or

$$\text{loop BW (EFC)} = \frac{\text{Lock Bandwidth Factor} \times N}{10^3} = \frac{\text{LBF} \times f_{dut}}{10^{10}}$$

$$\text{where } N = f_{dut}/10 \text{ MHz.}$$

Loop Bandwidth Limitations

The loop BW has certain limitations when locking via the 8662A reference. As shown in Figure B.2, there is a crystal filter (physically in the 8662A reference section) in the feedback chain. This bandpass filter is about 6 kHz wide, centered at 40 MHz. For the purposes of the phase feedback loop, the bandpass filter is transformed into an effective low pass filter of half the bandwidth (3 kHz). In fact, if the crystal filter is not exactly centered at 40 MHz, the effective low pass filter might be less than 3 kHz.

If a bandwidth close to the effective LPF cut-off frequency is selected, the loop will be unstable. For example, for a test frequency of 10 GHz, the 10k and 100k loop bandwidth factors would always yield an unstable loop. And if the crystal is not precisely centered, the 1k loop bandwidth factor might also be unstable, and will always give a loop amplitude response which peaks up several dB before rolling off. This peaking will apply uncalibrated gain to the phase noise measurement in the vicinity of the loop bandwidth.

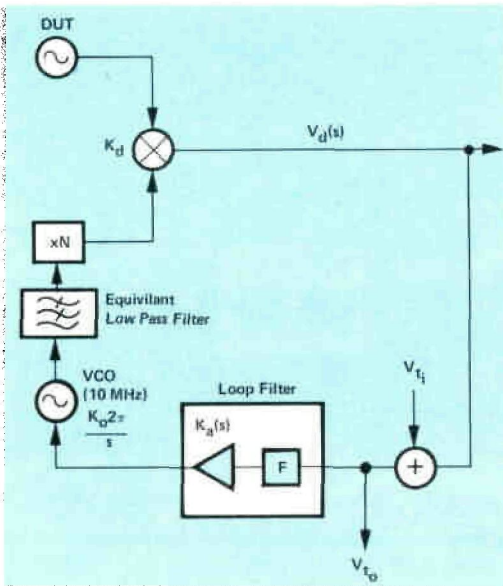


Figure B.2. PLL model when using the 8662A 10 MHz reference as loop VCO.

Phase Locking via the 8662A dc FM

LHR

This method applies the $\pm 1V V_c(s)$ signal to the 8662A dc coupled FM input (modulation input). The 8662A has selectable FM sensitivity (K_o), to a maximum of ± 200 kHz/V.

When locking via the dc FM of the 8662A, Figure B.1 can again be used as the loop model. The VCO is now the 8662A synthesized signal with a 140 MHz dc coupled FM oscillator added in. The loop locks the 8662A signal at the phase detector input. Thus, the loop holding range is set only by the FM deviation selected on the 8662A, and is not a function of the selected comb frequency (the FM deviation is not multiplied).

LHR (via 8662 dc FM) = FM deviation set

FM deviation on the 8662A is ± 200 kHz maximum, but is less for certain frequency ranges of the 8662A. Therefore, maximum obtainable LHR is dependent on the 11729B IF frequency that results after the first microwave down-conversion. Table B.1 shows the allowable 8662A FM deviations as a function of output frequency.

LBW

Again, 11729B loop bandwidth is a product of the loop gains. $BW = K_d F K_a(s) K_o$. In dc FM mode, the product of the 11729B loop gains $K_a K_d$ is equal to 10^{-3} volts/rad. Thus, loop BW is equal to

$$LBW \text{ (dc FM)} = \frac{F \times K_o}{10^3} = \frac{11729B \text{ LBF} \times 8662A \text{ deviation setting (Hz)}}{10^3}$$

Loop bandwidth can therefore be changed by varying the Lock Bandwidth Factor on the 11729B, or varying the peak FM deviation on the 8662A.

Loop Bandwidth Limitations

When locking via the 8662A dc FM, LBW is limited primarily by the 8662A maximum FM modulation rate of 100 kHz. Significant phase noise peaking can be expected for loop bandwidths approaching 100 kHz. Best results will be obtained with loop bandwidths <50 kHz.

Maximum LHR vs. Minimum LBW

In general, it is desirable to keep the loop BW as narrow as possible to avoid correction of the measured noise for loop suppression effects. It is also desired to keep the loop holding range (LHR) at maximum to keep the loop from unlocking as the DUT drifts. However, since in the dc FM mode of phase locking, the loop bandwidth is a function of the FM deviation set, and the LHR is equal to the FM deviation set, some trade-off is involved. If phase noise measurements at offsets close to the carrier are desired, LHR should be selected wide enough to keep the sources locked for a series of measurement sweeps on the spectrum analyzer, but not necessarily at maximum. After the series of sweeps, the 8662A can be retuned for the next series of sweeps as described in Loop Tune. This will minimize loop BW and the amount of loop noise correction that will be needed. Also remember that the loop bandwidth can be reduced by the LBF without sacrificing LHR.

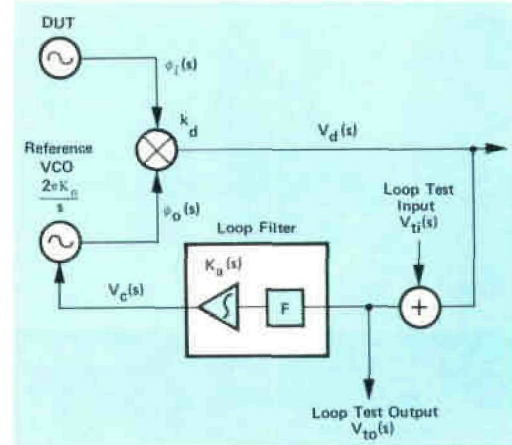


Figure B.3. Second order phase lock loop model.

Table B.1. Allowable 8662A dc FM deviation

8662A Front Panel Frequency (MHz)	Allowable Peak dc FM Deviation (kHz)
0.01 to 120	100
120 to 160	25
160 to 320	50
320 to 640	100
640 to 1280	200

Loop Capture

When Loop CAPTURE is pressed (or enabled over the HP-IB), a first-order phase lock loop replaces the second order loop. (CAPTURE removes the loop integrator, or equivalently removes the lock bandwidth factor F and replaces the loop amplifier K_A with a new wider K_A capture.) This yields a 3 dB bandwidth of $K_d K_o K_A$ capture (rad/s), a wider PLL that enables the two signals to acquire lock. Then when CAPTURE is released, the second order loop is again engaged, the signals will be in quadrature (forced by the higher dc gain in the 2nd order loop), and a measurement can be made.

Loop Characterization

Returning to Figure B.1, the perturbing input $V_{ti}(s)$ is normally zero. However, if a perturbation is input to the loop at this point — for example, a random noise source — the transfer characteristic of the phase lock loop can be determined. Note that the loop remains locked and in quadrature during the loop characterization.

The ratio of the voltage at the Loop Test Output V_{to} to the signal at the Loop Test Input V_{ti} can be described by:

$$\frac{V_{to}}{V_{ti}} = \frac{\text{Forward Gain}}{1 - \text{Open Loop Gain}} = \frac{1}{1 - G_{OL}}$$

Thus, the filter transfer characteristic can be traced out on the spectrum analyzer. It can be modeled as a simple one-pole hipass which will be 3 dB down from flat at $f_c = K_d K_o K_A(s)F$ and rolls off at 6 dB/octave.

The loop filter transfer function as displayed on the spectrum analyzer yields two important pieces of information. First, the loop BW can be determined, which designates the offset frequencies for which an uncorrected phase noise measurement can be made.

Secondly, the degree of noise suppression inside the loop BW can be measured. The amount of noise suppression vs. offset frequency can then be used to correct the value of DUT noise measured when the system was in the normal loop hold mode.

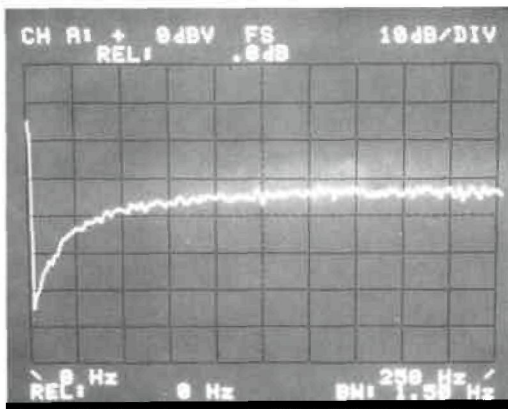


Figure B.4. Typical loop filter transfer function.

Calculating System Phase Noise

Since the output of a phase detector represents the combined noise of the two input signals, it is necessary for the noise of the reference source to be \leq the noise of the DUT. Therefore, before starting any phase noise measurement with the 11729B, determine whether the noise of the 11729B measurement system is less than the expected noise of the DUT. This appendix describes how to compute the 11729B measurement system noise dependent on the system parameters of the given measurement.

From Figure C.1 (a very simplified block diagram of the 11729B/8662A in a noise measurement system), the contributors to system noise are easily identified. System noise is effected by the multiplied noise on the 640 MHz reference source, the noise on the 5 to 1280 MHz tunable source, and the residual noise contribution of the 11729B (contributed by components such as the multiplier, amplifier, phase detector, etc.)

If N = harmonic of 640 MHz filtered by the selected bandpass filter, then the total system noise can be completely defined by:

$$\mathcal{L}_{\text{system}} \text{ (dBc)} = 10 \log \left(N^2 \times 10^{\frac{\mathcal{L}_1}{10}} + 10^{\frac{\mathcal{L}_2}{10}} + 10^{\frac{\mathcal{L}_3}{10}} \right)$$

\mathcal{L}_1 = Absolute SSB Phase Noise of the 640 MHz reference signal (dBc/Hz).

\mathcal{L}_2 = Absolute SSB Phase Noise of the 5 to 1280 MHz tunable signal (dBc/Hz).

\mathcal{L}_3 = Residual noise of 11729B (dBc/Hz).

Noise from the 640 MHz Source

Whenever a signal is multiplied, its noise is also multiplied. Therefore, the noise contribution from the 640 MHz reference is dependent on which harmonic is selected by the bandpass filter. This in turn is a function of the frequency of the device under test. Because the noise of the 640 MHz signal is multiplied, it is usually the major contributor to system noise. The typical and specified noise on the 640 MHz reference signal is shown in Table C.2.

N for the eight 11729B frequency bands is given in Table C.1.

Remember that for test frequencies less than 1.28 GHz, the multiplied 640 MHz signal is not used, and thus system noise is dependent only on the second two contributions in the formula.

Noise from the 5 to 1280 MHz Tunable Source

The second contributor to system noise is the noise on the tunable 5 to 1280 MHz signal. Several system parameters can affect the noise on this tunable signal. For example, if the 8662A is being used as the tunable source, the noise on the output signal is a function of frequency selected. This is because the 8662A obtains wide frequency coverage by frequency translating — multiplying, dividing, and heterodyning — its main band, processes which affect phase noise. The noise of the 8662A in its main 320 to 640 MHz band is given in Table C.2. The phase noise in its other bands can be computed or is given in the 8662A Technical Data Sheet.

Since 8662A noise is a function of frequency, the effect on 11729B system noise is dependent on the frequency to which the 8662A must be tuned ($f_{IF} = f_{\text{dut}} - f_{\text{filter}}$).

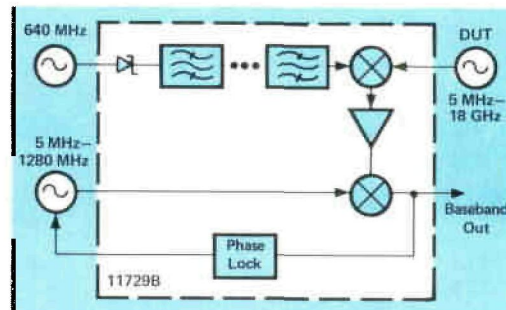


Figure C.1. Simplified 11729B block diagram.

Table C.1. 11729B harmonic numbers.

Frequency Range of Signal under Test (GHz)	Center Frequency of Microwave BPF	N
0.005 to 1.28	—	—
1.28 to 3.20	1.92	3
3.20 to 5.76	4.40	7
5.76 to 8.32	7.04	11
8.32 to 10.88	9.60	15
10.88 to 13.44	12.16	19
13.44 to 16.00	14.72	23
16.00 to 18.56	17.28	27

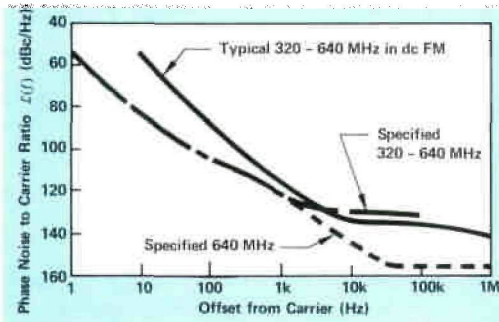
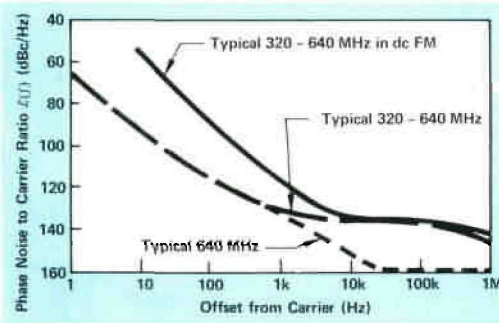


Figure C.2. Comparison of noise of 8662A signals.

Table C.2. Phase noise of 8662A signals.

Offset from carrier	Specified 640 MHz low noise output	Typical 640 MHz low noise output	Specified 320-640 MHz tunable 8662A output	Typical 320-640 MHz tunable 8662A output	Typical 320-640 MHz 8662A dc FM mode
1 Hz	-54 dBc	-64 dBc	-54 dBc	-64 dBc	
10 Hz	-84 dBc	-94 dBc	-84 dBc	-94 dBc	-55 dBc
100 Hz	-104 dBc	-114 dBc	-104 dBc	-114 dBc	-90 dBc
1 kHz	-121 dBc	-126 dBc	-121 dBc	-125 dBc	-116 dBc
10 kHz	-145 dBc	-149 dBc	-131 dBc	-136 dBc	-135 dBc
100 kHz	-157 dBc	-159 dBc	-132 dBc	-136 dBc	-135 dBc
1 MHz	-157 dBc	-159 dBc		-146 dBc	-142 dBc

The method of phase locking also affects the noise of the 8662A. If lock is established via the EFC of the 8662A internal reference, no additional phase noise is contributed. But when locking via the dc FM of the 8662A, a free-running 140 MHz oscillator is added to the 8662A synthesized signal. The noise on this 140 MHz oscillator degrades the total 8662A output signal phase noise, and thus total system noise. Since the FM oscillator is added before the 8662A frequency translation, its noise is subject to the same effects as the normal synthesized output signal. The total phase noise on the 8662A output signal when in dc FM mode in the main band is also given in Table C.2.

Figure C.2 graphs the absolute phase noise of the 640 MHz signal, the 8662A main band signal, and the 8662A main band signal in dc FM mode. It is clear that the 640 MHz signal is chosen for multiplication because of its much lower broadband noise. In general, for test frequencies >10 GHz, the noise on the multiplied 640 MHz signal is the dominant noise contributor. But for $f_{dut} < 10$ GHz, the noise contribution of the 5 to 1280 MHz signal must be considered, as it can increase overall broadband system noise floor by several dB.

If using a third source such as the 8640B for the 5 to 1280 MHz source, its noise must be used as the second noise contribution (L_2).

Table C.3. Typical 11729B residual noise.

Offset from carrier (Hz)	Typical residual noise	
	at 5 GHz (dBc)	at 10 GHz (dBc)
1 Hz	-99	-93
10 Hz	-108	-106
100 Hz	-120	-116
1 kHz	-130	-125
10 kHz	-137	-132
100 kHz	-146	-141
1 MHz	-148	-142

11729B Residual Noise

The typical residual noise of the 11729B is given in Table C.3. In general, the residual noise of the 11729B is well below the absolute noise contribution of the multiplied 640 MHz reference signal, and so makes very little contribution to overall system noise floor and need not be computed. However, for measurements where the noise of the DUT approaches the noise of the overall measurement system, overall measurement error increases. Thus, the 11729B residual noise should be included in the system noise floor calculation.

Measurements with 11729B/8662A/3582A/8566A **D**

The HP 3582A and 8566A are good choices for use as the baseband analyzers in an 11729B/8662A Phase Noise Measurement System. The 3582A is a digital Fast Fourier Transform analyzer, covering 0.02 Hz to 25 kHz. Optimized for low frequency analysis, the 3582A provides quicker measurement of phase noise close to the carrier than could be obtained from wider frequency range spectrum analyzers. The 8566A is a high performance microwave spectrum analyzer covering 100 Hz to 22 GHz. It enables high sensitivity phase noise measurements to be made down to the upper end of the HP 3582A frequency range (and lower if desired) with narrow resolution bandwidths down to 10 Hz. Also, it is very useful for characterizing the DUT at the microwave frequency before applying the signal to the 11729B input.

This appendix gives the necessary settings for the HP 3582A and HP 8566A to make a phase noise measurement. The 11729B and 8662A settings are the same as given in Chapter 5. The numbers on the following steps correspond to the step numbers in Chapter 3 and Chapter 5.

Set-Up

Connect the system as shown in Figure D.1, with the 11729B <1 MHz Noise Spectrum Output connected to CH.A of the 3582A, and the <10 MHz Noise Spectrum Output input to the 8566A.

1. Select carrier frequency.
2. Set 8662A (or third source if being used) to the appropriate IF frequency.

Calibrate

3. Set up a beat note by offsetting the RF source in frequency and 40 dB in output level.

Note: The 11729B phase lock loop amplifier might drift slightly when not in a phase-locked condition. When measuring a synthesized source, this will result in some drift of the Calibration beat note. This small amount of drift can be eliminated by holding in CAPTURE during the calibration procedure.

4. Measure the level of the beat note.

On the 3582A:

Select A Coupling ac, Input MODE A, REPETITIVE TRIGGER, DISPLAY AMPLITUDE A, 0 to 25 kHz frequency span, SCALE 10 dB/DIV, Flattop passband shape, averaging off. Turn up CH. A input sensitivity until overload indicator just remains unlit. Enable the marker and place it on the beat note. Set a reference at this carrier level by pressing SET REF. Enter the relative mode by pressing REL. Now all further measurements will be in dB relative to this carrier level.

On the 8566A:

Start with INSTRUMENT PRESET. Then select the desired frequency span by pressing START FREQUENCY 0 Hz, STOP FREQUENCY 200 kHz. Change the input attenuation until the beat note is at the top of the screen. Record the level of the beat note, V_b in dBm.

Lock

5. Lock the signals in quadrature.

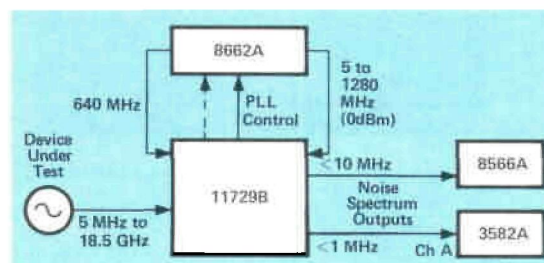


Figure D.1. System connection.

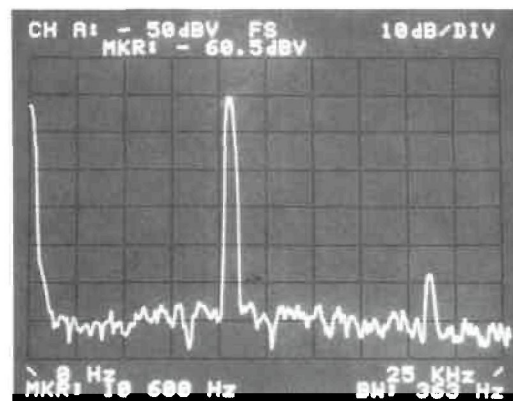


Figure D.2. Calibration beat note on 3582A.

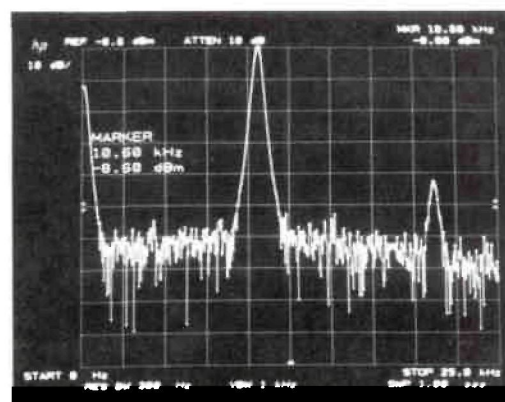


Figure D.3. Calibration beat note on 8566A.

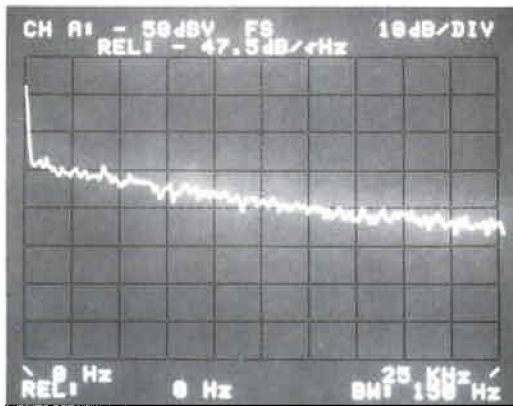


Figure D.4. Noise plot on 3582A.

Measure

6. Measure the phase noise on the spectrum analyzers.

On the 3582A:

Set the 3582A to span the desired offset frequency. For best accuracy, the noise measurement should be above the bottom 10 dB division on the 3582A.

Select the Hanning Passband shape, place the 3582A in the RMS AVERAGE mode, select the desired number of averages and press RE-START. As the 3582A takes readings, monitor the 11729 PHASE LOCK INDICATOR to ensure that the inputs to the mixer remain within the desired limits about quadrature.

To obtain readings directly in dBc/Hz on the 3582A, enable the automatic bandwidth normalization by pressing $\div \sqrt{BW}$. When the 3582A is finished sweeping, move the marker to the desired offset frequency and record the reading on the screen ($V_{\text{noise } 3582}$).

Repeat for different offset frequencies as desired.

On the 8566A:

Select the desired frequency span of interest, either by setting the START FREQUENCY and STOP FREQUENCY, or by selecting CENTER FREQUENCY and FREQUENCY SPAN. Choose a resolution BW significantly less than the frequency span selected by pressing RES BW xx Hz. A good rule of thumb is to use a resolution BW $\leq 10\%$ of the selected frequency span.

The 8566A provides for either digital or analog averaging. Both forms of averaging (or some combination) should provide the same measured value, and the choice of averaging mode is based on user-preference.

The analog averaging (video filtering) in general provides more averaging in the same period of time. Analog averaging effectively takes all the averages of one data point, and then displays that point on the screen. When the entire trace is visible on the screen, all the data has been averaged.

On the other hand, the digital averaging mode first displays the entire measurement sweep, and then updates the display by averaging successive sweeps. Therefore, some users prefer digital averaging as it quickly gives the overall shape of the noise spectrum.

Enable digital averaging by pressing SHIFT G bb HZ, where bb is the desired number of averages. For analog averaging, press VIDEO BW cc HZ, where cc is the desired video bandwidth.

The 8566A provides a convenient way to display the measured noise spectrum normalized to a NOMINAL 1 Hz noise measurement bandwidth, plus account for the 2.5 dB correction factor necessary for the effect of the 8566A's log amplifiers and peak detectors on noise. Enable the Noise Marker mode by pressing SHIFT M. Re-enable the marker by pressing NORMAL, and move the marker to the desired offset frequency. Record the displayed marker level, $V_{\text{noise } 8566}$ in dBm/Hz.

However, for best accuracy, the actual noise measurement bandwidth should be measured and then used to correct the noise reading taken. Enable the marker and record the displayed noise level at the desired offset frequency, $V_{\text{noise } 8566A}$ (in the selected nominal resolution bandwidth) in dBm.

Repeat for different offset frequencies as desired.

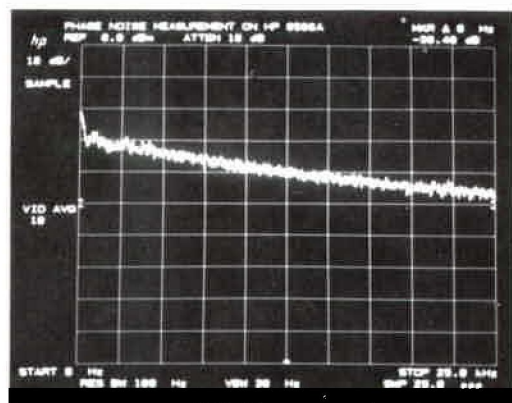


Figure D.5. Noise plot on 8566A.

Apply Corrections

7. Apply the necessary correction factors.

On the 3582A:

The 3582A noise measurement is already relative to the carrier level and normalized to a 1 Hz noise bandwidth. Therefore, the only corrections necessary are as follows:

$$\begin{aligned} \mathcal{L}(f) \text{ (dbc/Hz)} &= V_{\text{noise}3582A} \\ &-40 \text{ dB for calibration} \\ &-6 \text{ dB for mixer action} \\ &+ \text{loop noise suppression if applicable.} \end{aligned}$$

On the 8566A:

If the convenient Noise Marker mode was selected, the 8566A reading is also already normalized to a nominal 1 Hz noise measurement bandwidth, plus corrected by 2.5 dB. The necessary corrections to convert the measured values to $\mathcal{L}(f)$ are then:

$$\begin{aligned} \mathcal{L}(f) \text{ (dBc)} &= V_{\text{noise}8566A} \text{ (dBm/Hz)} \\ &- V_b \text{ (dBm)} \\ &-40 \text{ dB for calibration} \\ &-6 \text{ dB for effect of mixer used as phase detector} \\ &+ \text{loop noise suppression if applicable.} \end{aligned}$$

If the Noise Marker mode was not used, $\mathcal{L}(f)$ is obtained by:

$$\begin{aligned} \mathcal{L}(f) \text{ (dBc)} &= V_{\text{noise}8566A} \text{ (dBm)} \\ &-10 \log(\text{equivalent noise bandwidth}) \\ &- V_b \text{ (dBm)} \\ &-40 \text{ dB for calibration} \\ &-6 \text{ dB for effect of mixer used as a phase detector} \\ &+2.5 \text{ dB for effect of 8566A circuitry} \\ &+ \text{loop noise suppression if applicable.} \end{aligned}$$

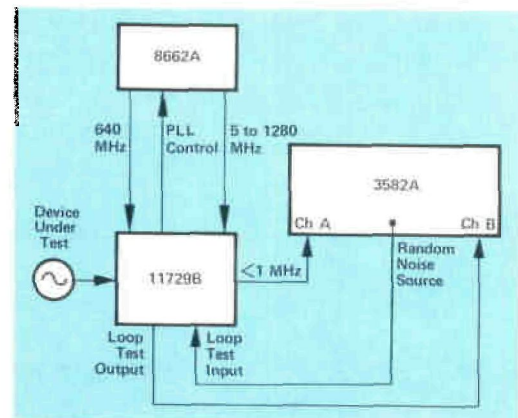


Figure D.6. System connection for measuring loop noise suppression with 3582A.

Measurement of Loop Noise Suppression

If the loop bandwidth selected is less than 25 kHz (see Chapter 4), all loop noise suppression measurements can be made on the 3582A with its noise source. If the BW selected is greater than 25 kHz, an alternative loop test source must be used and the measurements made on the 8566A.

On the 3582A:

Connect the system as shown in Figure D.6. The 3582A random noise source is used as the input to the 11729B Loop Test Input. To avoid re-cabling, it is convenient to connect the Loop Test Output to Channel B of the 3582A. Select B Coupling ac, Input MODE B, REPETITIVE TRIGGER, DISPLAY AMPLITUDE B, frequency span 0 to approximately the 1.5 x expected loop BW, SCALE 10 dB/dIV, Hanning Passband shape. Turn up CH B sensitivity until the overload indicator remains just unlit. Place the 3582A in the RMS AVERAGE mode, select the desired number of averages and press RESTART. As the 3582A takes readings, monitor the 11729B Lock Indicator to ensure that the mixer inputs remain in nominal quadrature.

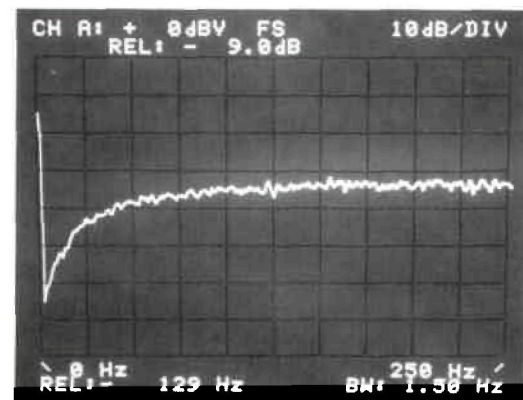


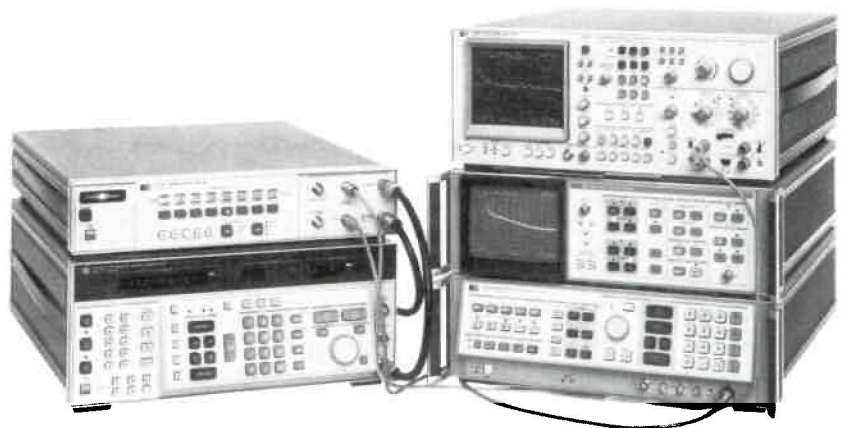
Figure D.7. Loop noise suppression measurement on 3582A.

The displayed loop transfer function should level out in the sweep. If it does not, select a larger frequency span, and RESTART the 3582A. When the display shows all relevant data about the effective loop HPF, enable the marker and place it where the level becomes constant. Set a reference at this level by pressing SET REF. Enter the relative mode by pressing REL. Move the marker to the desired offset frequency of interest. The delta marker reading will give the amount of loop noise suppression. (Remember that the displayed marker frequency reading will also be relative to the reference point rather than reading directly in offset from the carrier frequency). If more exact information about the loop suppression is desired, the frequency span can be narrowed to the range of interest. However, it is not recommended to take Loop Suppression measurements greater than a decade inside the PLL bandwidth.

On the 8566A:

For bandwidths greater than 25 kHz, an alternative Loop Test Input is simply the variable sine wave output of a function generator. Connect the function generator to the Loop Test Input, and the Loop Test Output to the 8566A. Select a 1 kHz sine wave on the function generator, and adjust its level until a reasonable beat note can be seen at 1 kHz on the 8566A.

Set the 8566A to span 500 Hz to a frequency slightly greater than the expected loop bandwidth. Increase the frequency of the function generator until the resultant beat note levels out. Enable the marker on this beat note, and then enter into a relative mode by pressing DELTA. Decrease the frequency of the function generator to the offset frequency of interest; the delta marker reading will indicate the amount of loop noise suppression.



11729B/8662A/3582A/8566A phase noise measurement.

11729B HP-IB Programming Codes

Code	Function
AM	Selects AM Noise Measurement Mode
PH	Selects Phase Noise Measurement Mode
FT 1	Selects first filter band (5 to 1280 MHz)
FT 2	Selects second filter band (1.28 to 3.2 GHz)
FT 3	Selects third filter band (3.2 to 5.76 GHz)
FT 4	Selects fourth filter band (5.76 to 8.32 GHz)
FT 5	Selects fifth filter band (8.32 to 10.88 GHz)
FT 6	Selects sixth filter band (10.88 to 13.44 GHz)
FT 7	Selects seventh filter band (13.44 to 16 GHz)
FT 8	Selects eighth filter band (16 to 18.56 GHz)
LK 1	Selects Lock Bandwidth Factor of 1
LK 2	Selects Lock Bandwidth Factor of 10
LK 3	Selects Lock Bandwidth Factor of 100
LK 4	Selects Lock Bandwidth Factor of 1k
LK 5	Selects Lock Bandwidth Factor of 10k
CA 1	Enables Capture
CA 0	Disables Capture
LP	Learn Front Panel
RO	Read Option List
RM	Read RSQ
@	Accept RSQ
CS	Clear Status Byte

Hewlett-Packard Application Notes

- AN 246-2 Measuring Phase Noise with the HP 3585A Spectrum Analyzer.
- AN 207 Understanding and Measuring Phase Noise in the Frequency Domain.
- AN 150-4 Spectrum Analysis . . . Noise Measurements.
- AN 283-1 Applications and Measurements of Low Phase Noise Signals Using the 8662A Synthesized Signal Generator.

Other References

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Howe, D., "Frequency Domain Stability Measurements: A Tutorial Introduction", N.B.S. Technical Note 679, National Bureau of Standards, Boulder, CO, 1976.

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Shoaf, J.H., Halford, D., Risley, A.S., "Frequency Stability Specification and Measurement: High Frequency and Microwave Signals", N.B.S. Technical Note 632, January, 1973.

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