Practical Considerations for Modern RF & Microwave Phase Noise Measurements

Graham Payne
Applications Engineer
Agilent Technologies Ltd.(UK)

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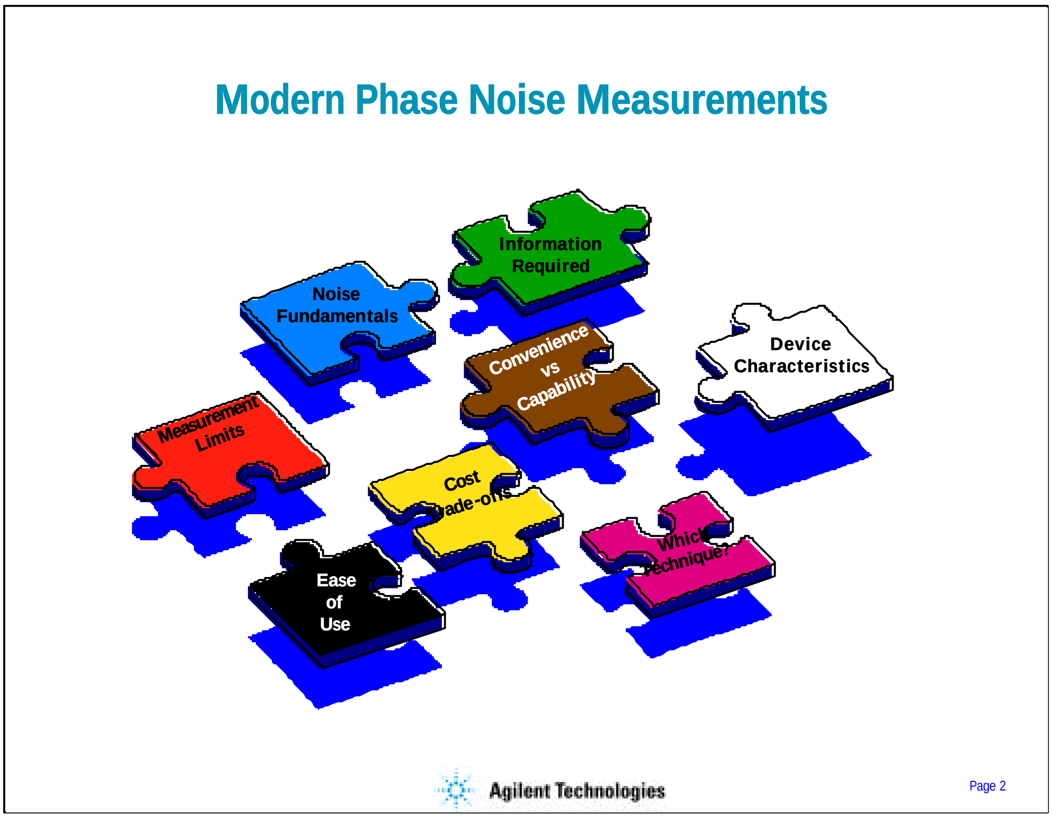
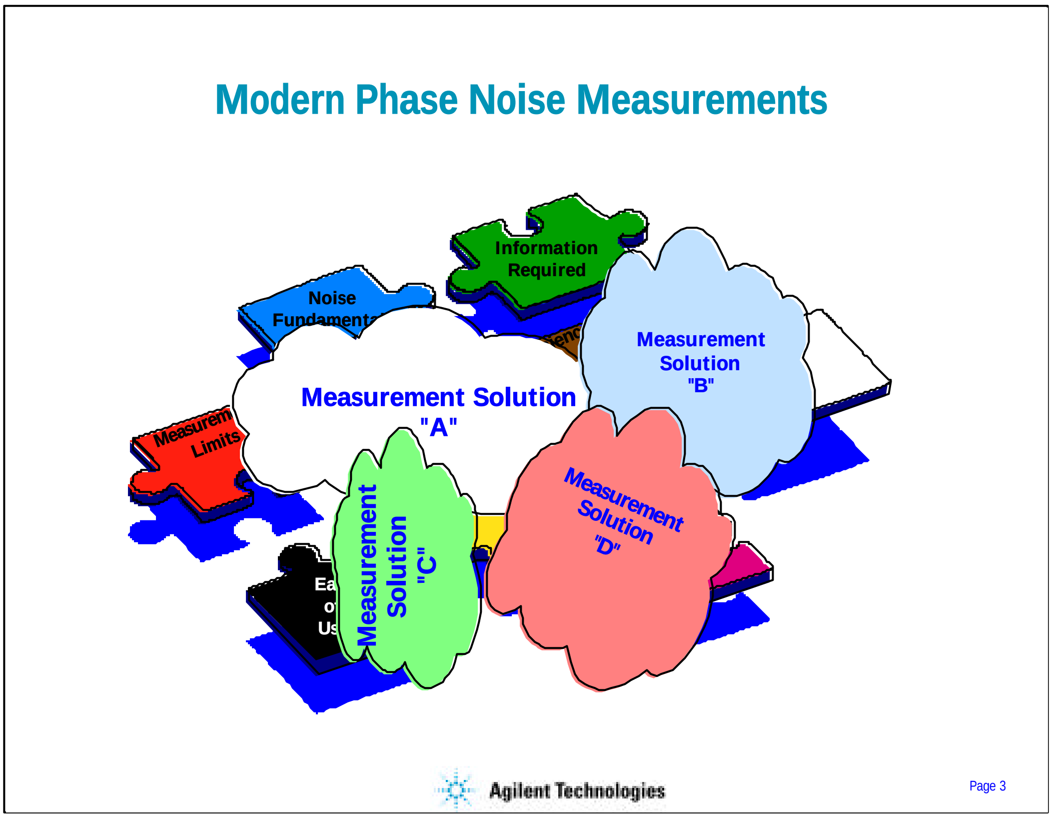


We want to encourage interactive discussion thoughout this seminar so we can all share your problems and experiences. To start, what issues and concerns have you brought which you would like to see resolved today?

As shown here, the most confusing aspect to making successful phase noise measurements is dealing with all of the different items of knowledge which seem to be important.



To complicate matters even further, different phase noise measurement solutions solve different portions of the puzzle. At the end of today's seminar, you should be able to identify how these pieces fit together, and which solutions help solve your measurement problems.

Agenda

- Review Basic Phase Noise Concepts

  Define Noise Measurement Techniques

  Select an Appropriate Measurement Technique

  Optimizing Modern Noise Measurements

  Survey of Measurement Solutions Available

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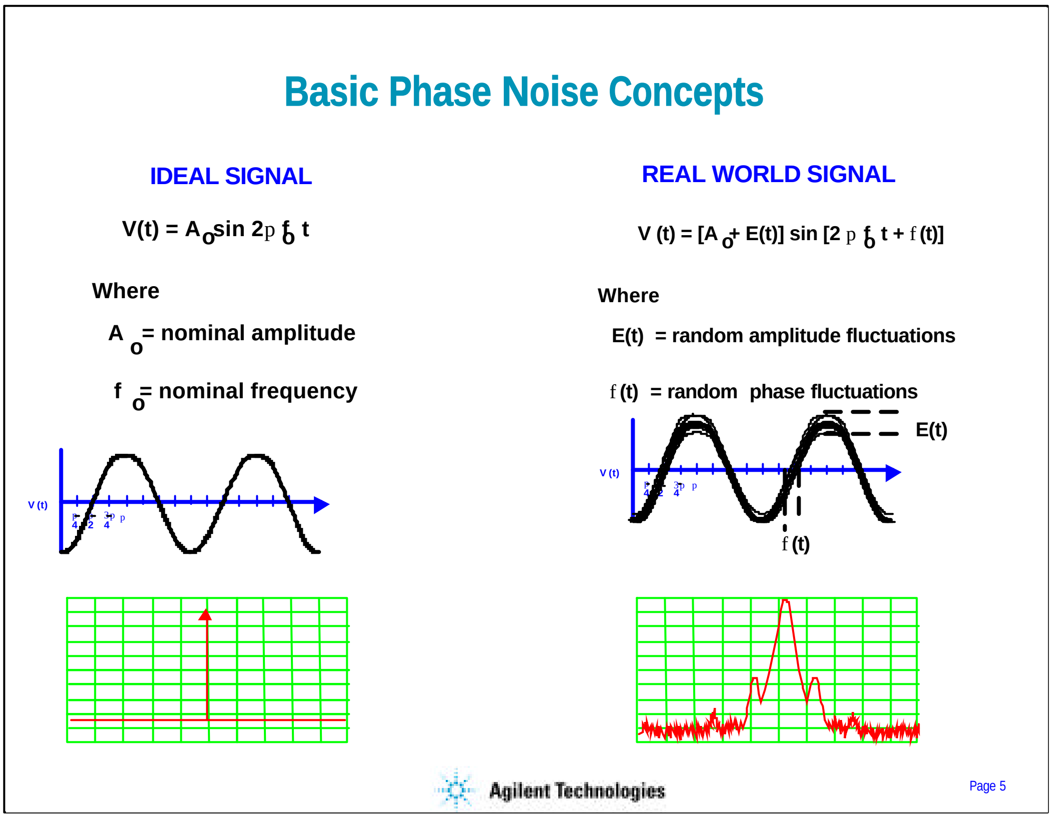
These are the topics that we will discuss today. Where practical, we will demonstrate the measurement concepts discussed. The seminar reviews basic noise concepts which are important to remember before we discuss different measurement techniques. Once the basic techniques are understood, we can determine which are most applicable to the devices we need to measure. Once we are making measurements, optimizing a measurement becomes important. The last section of the seminar will discuss the current phase noise measurement solutions available from Agilent.



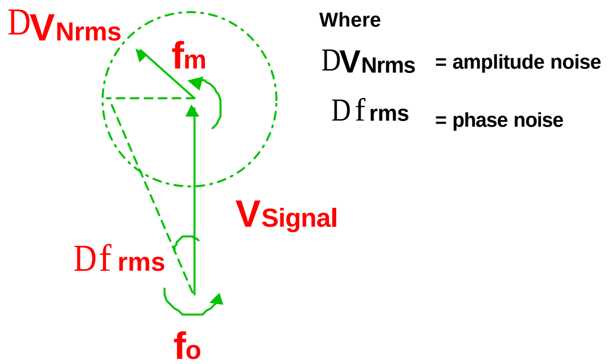
Before we get to the formal definitions of phase noise, let's look at the difference between an ideal signal (a perfect oscillator). In the frequency domain, this signal is represented by a single spectral line.

In the real world however, there is always small, unwanted amplitude and phase fluctuations present on the signal. Notice that frequency fluctuations are actually an added term to the phase angle portion of the time domain equation. Because phase and frequency are related, you can discuss equivalently about unwanted frequency or phase fluctuations.

In the frequency domain, the signal is no longer a discrete spectral line. It is now represented by spread of spectral lines - both above and below the nominal signal frequency in the form of modulation sidebands due to random amplitude and phase fluctuations.

Basic Phase Noise Concepts

USING PHASOR RELATIONSHIPS



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You can also use phasor relationships to describe how amplitude and phase fluctuations affect the nominal signal frequency.

Basic Phase Noise Concepts

UNIT OF MEASURE

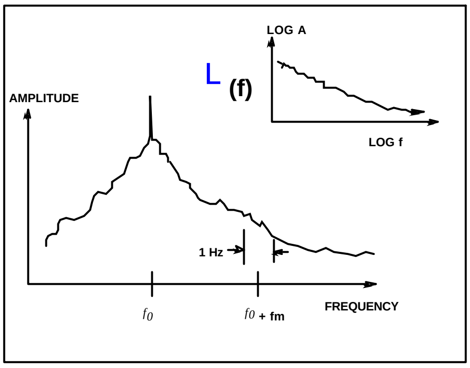
Usually phase noise is quantified as L *(f)*

L *(f) - defined as single sideband power due to phase fluctuations referenced to the carrier frequency power:*

- **In a 1 Hz bandwidth at a frequency f Hz from the carrier**
- **Divided by the signal's total power**

L *(f) has units of dBc/Hz*

$$\mathcal{L}(f) = \frac{\text{Area 1 Hz bandwidth}}{\text{Total area under the curve}}$$



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Historically, phase noise units of measure has been the single sideband power within a one hertz bandwidth at a frequency f away from the carrier referenced to the carrier frequency power. This unit of measure is described as script L(f) and the units are dBc/Hz.

Basic Phase Noise Concepts

Noise Power in a 1 Hz bandwidth

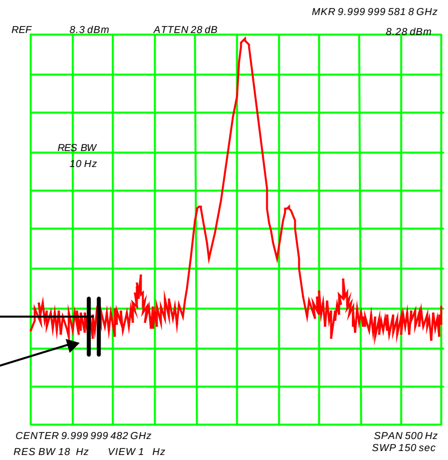
Total Signal Power

Ps (dBm)

$\mathcal{L}(f)$ = Pn (dBm/Hz) - Ps (dBm)

Pn (dBm/Hz)

1 Hz bw



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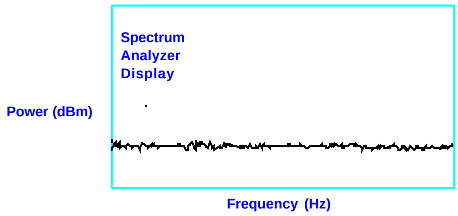
If you where measuring phase noise on a spectrum analyzer, this is basically what you would do.

Basic Phase Noise Concepts

THERMAL NOISE



$$N_p = kTB$$

k=Boltzman's Constant   T=Temperature K   B=Bandwidth

For T=290K $N_p = -204\ \dfrac{\text{dB(Watts)}}{\text{Hz}} = \dfrac{-174\text{dBm}}{\text{Hz}}$

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Thermal noise is the mean available noise power per Hz from a resistor at a temperature T(kTB). As the temperature of the resistor increases, the kinetic energy of its electrons increase and more power becomes available. Thermal noise is broadband and virtually flat with frequency.

Basic Phase Noise Concepts

Thermal Noise Effects on Phase Noise Measurements

$\mathcal{L}(f) = P_n\,(\text{dBm/Hz}) - P_s\,(\text{dBm})$

Total Power (kTB) = P_n (kTB) = -174 dBm/Hz
Phase Noise and AM noise equally contribute
Phase Noise Power (kTB) = -177 dBm/Hz

Theoretical kTB Limits to Phase Noise Measurements For Low Signal Levels

P_s(dBm)	$\mathcal{L}(f)$ [dBc/Hz]
+10	-187
0	-177
-10	-167
-20	-157

Note: There are other measurement factors besides kTB limitations which can reduce the theoretical measurement limit significantly.

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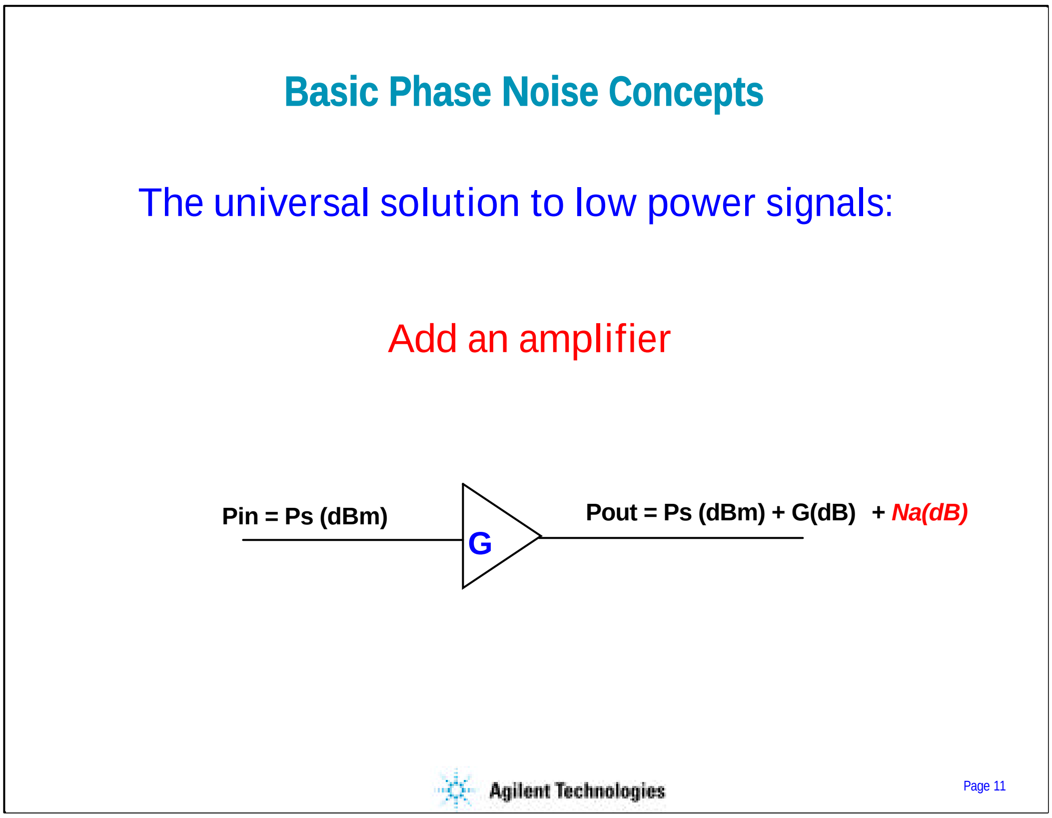
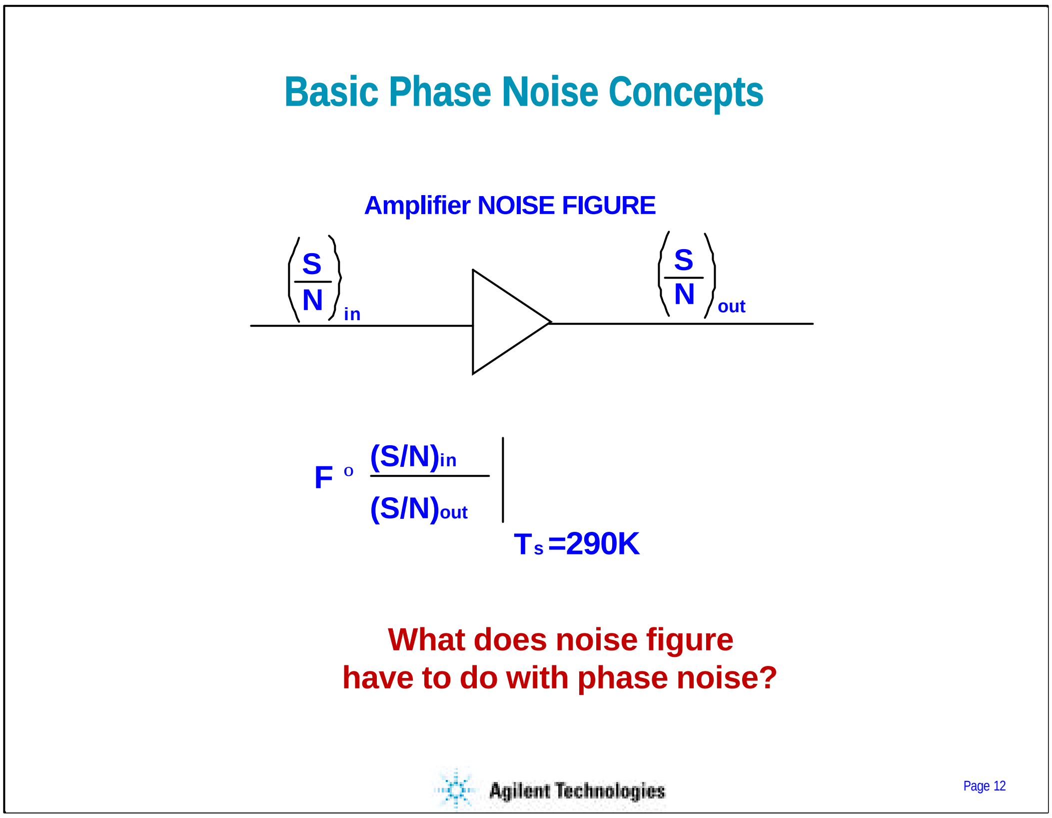
Thermal noise can limit the extent to which you can measure phase noise. Thermal noise as described by kTB at room temperature is -174 dBm/Hz. Since phase noise and AM noise contribute equally to kTB, the phase noise power portion of kTB is equal to -177 dBm/Hz (3 dB less than the total kTB power).

If the power in the carrier signal becomes a small value, for example -20 dBm, the limit to which you can measure phase noise power is the difference between the carrier signal power and the phase noise portion of kTB (-177 dBm/Hz - (-20 dBm) = -157 dBc/Hz). Higher signal powers allow phase noise to be measured to a lower level.

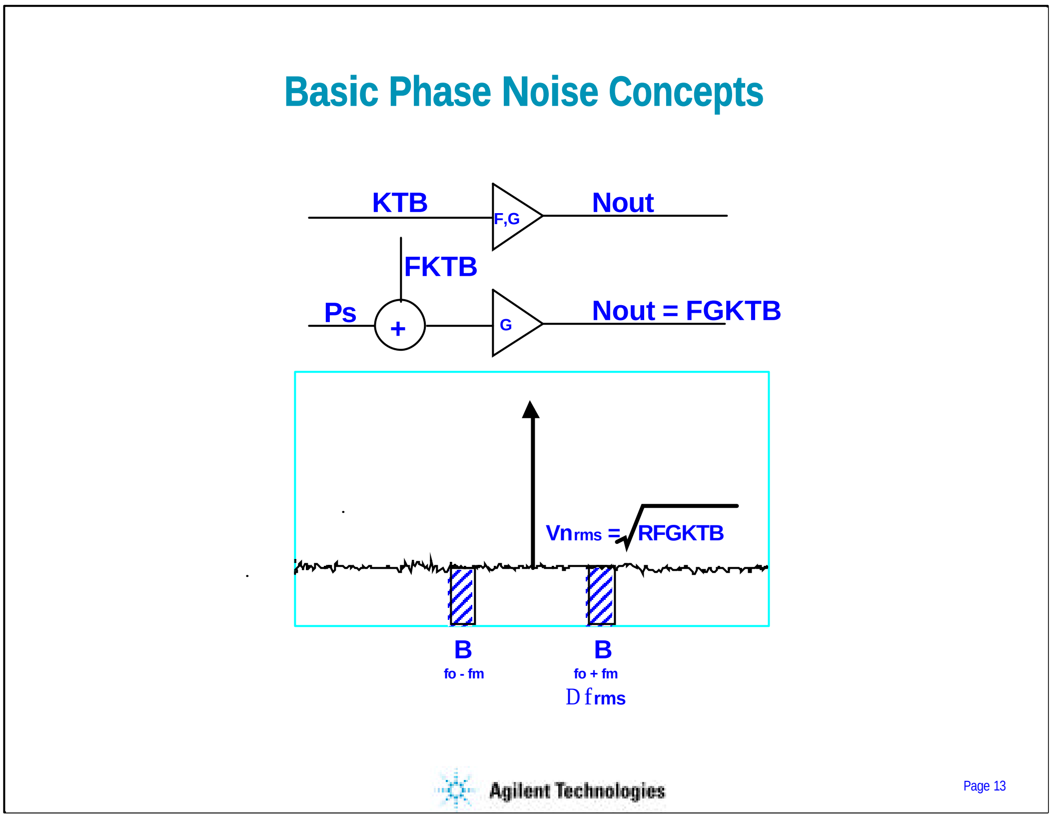


But the amplifier itself adds noise. Adding amplification also adds noise, which we need to account for within our measurement.



Amplifiers boost not only the input signal but also the input noise. The input signal-to-noise ratio is only preserved when the amplifier itself does not add noise.

Noise Figure is simply the ratio of the signal-to-noise at the input of a two-port device to the signal-to-noise ratio at the output, in dB, at a source temperature of 290K. In other words, noise figure is a measure of the signal degradation as it passes through the device - due to the addition of noise by the device. What does this have to due with phase noise measurements?



The noise power at the output of an amplifier can be calculated if its gain and noise figure are known. The noise at the amplifier output is given by N(out) = FGkTB.

The display shows the rms voltages of a signal and noise at the output of the amplifier. We want to see how this noise affects the phase noise of the amplifier.

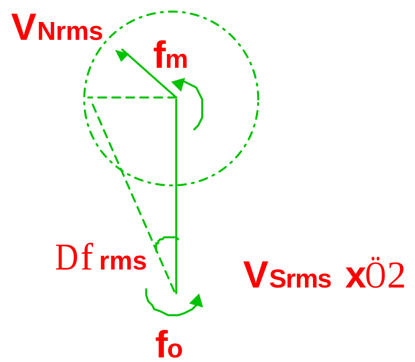
Basic Phase Noise Concepts

USING PHASOR RELATIONSHIPS

$$D\,f_{peak} = \frac{V_{Npeak}}{V_{speak}}$$

$$D\,f_{rms} = \frac{V_{Nrms}}{V_{speak}}$$

$$D\,f_{rms} = \frac{V_{Nrms}}{V_{srms} \cdot \sqrt{2}}$$



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Basic Phase Noise Concepts

USING PHASOR RELATIONSHIPS

For small $\Delta\phi_{rms}$

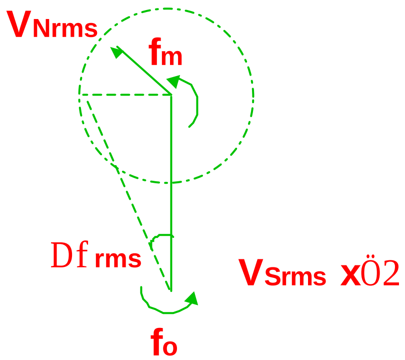
$$\Delta\phi_{rms} = \frac{V_{Nrms}}{V_{Srms} \times \sqrt{2}} = \sqrt{\frac{FGkTB}{2P_sG}}$$

$$\Delta\phi_{rms\,total} = \sqrt{\Delta\phi_{rms_1}^2 + \Delta\phi_{rms_2}^2} = \sqrt{\frac{FkTB}{P_s}}$$

$$S_\phi(f) = \frac{\Delta\phi^2_{rms}(f)}{B} = \frac{FkT}{P_s}\left[\frac{rad^2}{Hz}\right]$$

$$L(f) = \frac{S_\phi(f)}{2}$$

$$\therefore\ L(f) = \frac{FkT}{2P_s}\left[\frac{rad^2}{Hz}\right]$$



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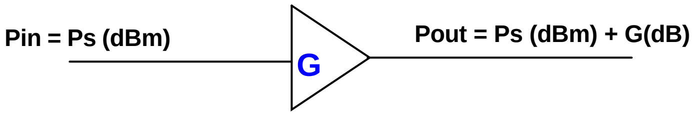
Using phasor methods, we can calculate the effect of the superimposed noise voltages on the carrier signal. We can see from the phasor diagram that V(norms) produces a delta phi (rms) term. For small angles, delta phi (rms) = V(nrms)/Vs(peak). The total delta phi(rms) can be found by adding two individual phase components powerwise. Squaring this result and dividing by the bandwidth gives the spectral density of phase fluctuations or phase noise. The phase noise is directly proportional to the thermal noise at the input and the noise figure of the amplifier.

Basic Phase Noise Concepts

$$\mathcal{L}(f) = P_{in\ KTB}\,(\mathrm{dBm/Hz}) + NF(\mathrm{dB}) + G(\mathrm{dB}) - P(out)$$



$$\mathcal{L}(f) = P_{in\ KTB}\,(\mathrm{dBm/Hz}) + NF(\mathrm{dB}) - Ps(\mathrm{dBm})$$

Conclusion:

Amplifiers help boost the device carrier power signal to levels necessary for successful measurements, but the theoretical phase noise measurement limit is reduced by the noise figure of the amplifier and the low signal power from the signal-under-test.

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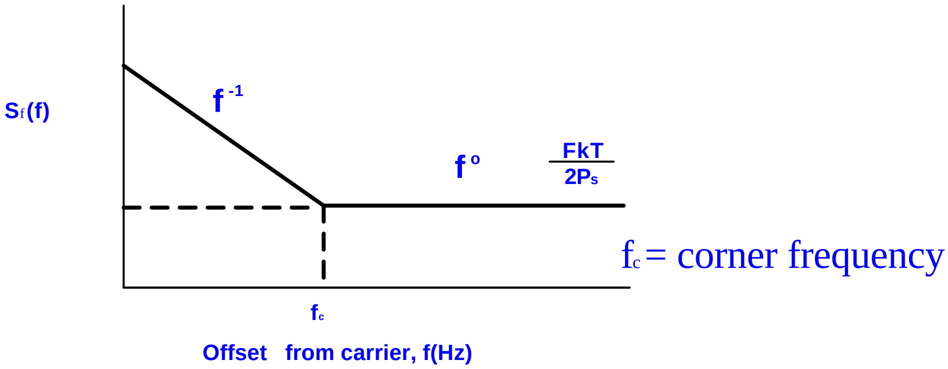
To summarize, amplifiers help boost the device carrier power signal to levels necessary for successful measurements, but the theoretical phase noise measurement limit is reduced by the noise figure of the amplifier and the low signal power from the signal-under-test.

Basic Phase Noise Concepts



- Phase noise "flicker" appears < f_c
- "flicker" noise is approximately -120 dBc/Hz @ 1 Hz offset

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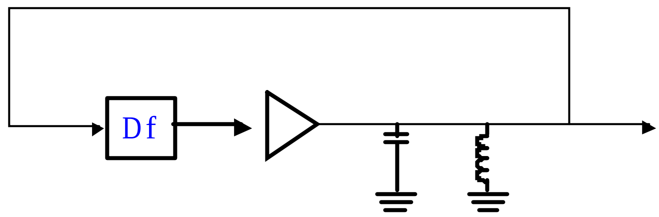
In addition to a thermal noise floor of approximately constant level with frequency, active devices exhibit a noise flicker characteristic which intercepts the thermal noise at an empirically determined corner frequency (fc). For offsets below fc, So(f) increases with f-1. This noise is caused by defects within semiconductor lattice structures resulting in combination-recombination of charge carriers.

Noise processes in an oscillator



$$Df = Df \frac{f_o}{2Q}$$

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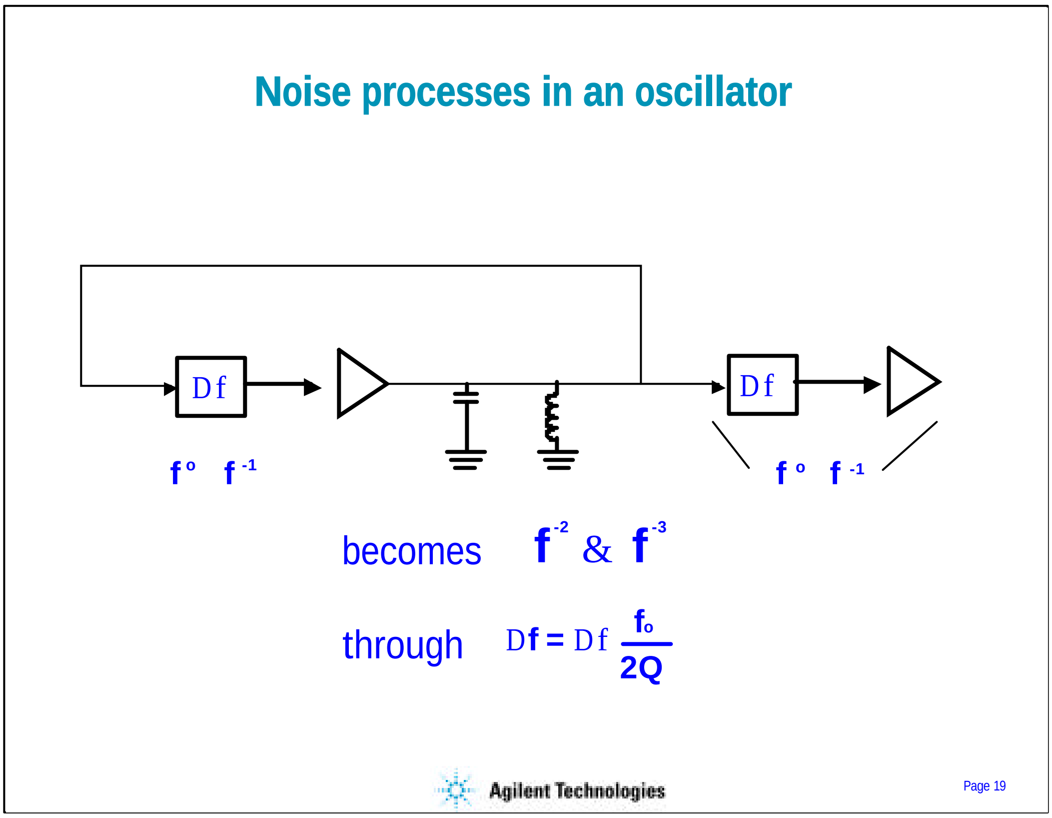
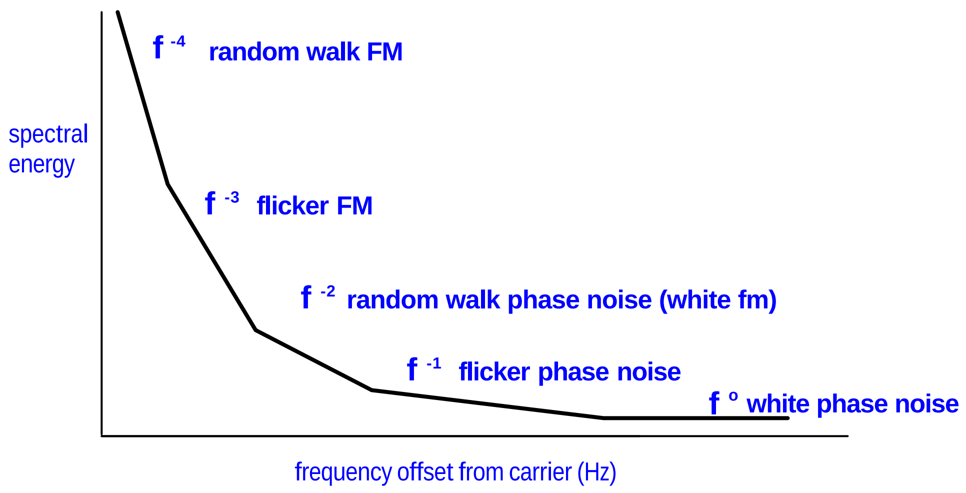
In an oscillator, the white (fo) and flicker (f-1) phase modulations cause even greater slopes of noise spectra. To demonstrate this, add a resonator of some quality factor Q to the output of an amplifier. Then connect the resonator output back to the amplifier input in the proper polarity for positive feedback. Consider the fo and f-1 of the amplifier to be represented by the phase modulator (delta phi) with a perfect amplifier. Any oscillator will shift frequency in response to a phase change anywhere in its loop. Since fo and Q are constants, phase modulation is converted directly to frequency modulation. This makes their spectral slopes two units more negative.



The oscillating loop itself will have noise slopes of f-2 and f-3, but the buffer amplifier found in most oscillators adds its own fo and f-1 noise slopes to the output signal

Typical Phase Noise Distribution



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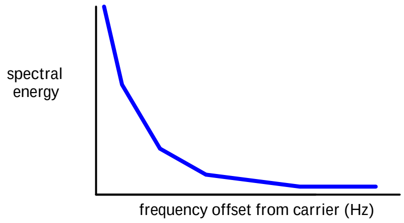
The distribution of phase noise energy can be described in the terms given above. Each of these characteristic noise distributions is due to a distinct process in the source circuitry.

Basic Phase Noise Concepts

Quantifying Phase Noise in Terms of Power Spectral Density



$S_{\phi}(f)$ Spectral Density of Phase Fluctuations

$\mathcal{L}(f)$ Single Sideband phase noise relative to total signal power

$S_{\nu}(f)$ Spectral Density of Frequency Fluctuations

$S_y(f)$ Spectral Density of Fractional Frequency Fluctuations

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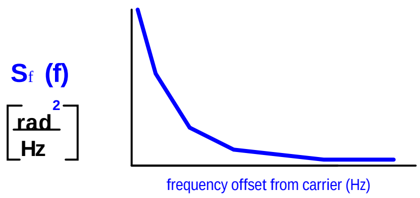
Due to the random nature of the instabilities, the phase deviation is represented by a spectral density distribution plot. The term spectral density describes the energy distribution as a continuous function, expressed in units of energy within a given bandwidth. The short term instability is measured as low-level phase modulation of the carrier and is equivalent to phase modulation by a noise source. There are four different units used to quantify spectral density:

Sphi(f), L(f), Snu(f), and Sy(f).

Spectral Density of Phase Fluctuations



Demodulate phase modulated signal with a phase detector (convert phase fluctuations to voltage fluctuations)

$\Delta V_{out} = K_\phi \Delta\phi_{in}$ $\quad K_\phi$ = V/radian

Measure the voltage fluctuations on a spectrum analyzer:

$\Delta V_{rms}(f) = K_\phi \Delta\phi_{rms}(f)$ [V]

$$S_\phi(f) = \frac{\Delta\phi^2_{rms}(f)}{B} = \frac{\Delta V^2_{rms}(f)}{K_\phi^2 B} = \frac{S_{V_{rms}}(f)}{K_\phi^2} \left[\frac{rad^2}{Hz}\right]$$

$S_{Vrms}(f)$ = the power spectral density of the voltage fluctuations out of the phase detector

$S_\phi(f)$ is expressed in dB relative to 1 radian

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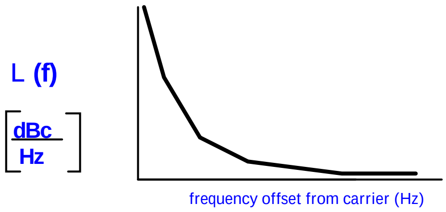
A measure of phase instability often used is Sϕ(f), the spectral density of phase fluctuations on a per-Hertz basis. The term spectral density describes the energy distribution as a continuous function, expressed in units of phase variance (radians) per unit bandwidth. If we use 1 radian(rms)/Hz as the phase variance comparison, we can express Sϕ(f) in terms of dB.

For large phase variations (>> 1 radian rms/rt Hz), Sϕ(f) will be greater than 0 dB. For small phase variations (< 1 radian rms/rt Hz), Sϕ(f) will be less than 0 dB.

Single Sideband Phase Noise

L (f)

L *(f)* = *single sideband power due to phase fluctuations referenced to the carrier frequency power:*



$$\mathcal{L}(f) = \frac{\text{Power density (one phase modulation sideband)}}{\text{Carrier Power}} \left[\frac{\text{dBc}}{\text{Hz}}\right]$$

$$L(f) = \frac{S_{\phi}(f)}{2}$$

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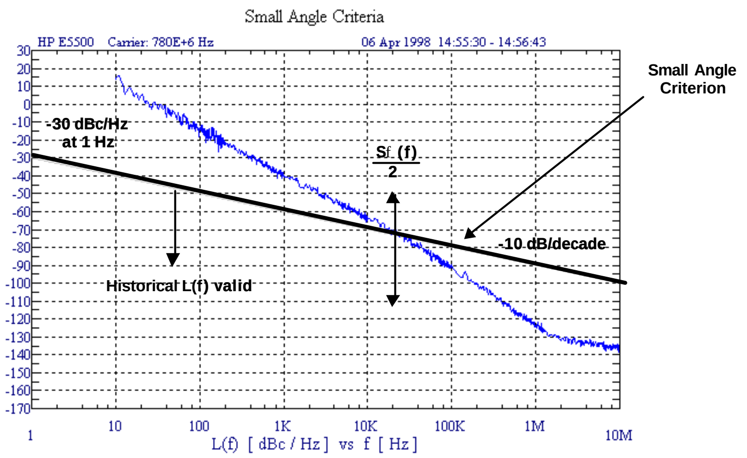
L(f) is an indirect measure of noise energy easily related to the RF power spectrum observed on a spectrum analyzer. The historical definition is defined as the ratio of the power in one phase modulation sideband on a per-hertz basis, to the total signal power. L(f) is usually presented logarithmically as a plot of phase modulation sidebands in the frequency domain, expressed in dB relative to the carrier power per Hertz of bandwidth [dBc/Hz].

This historical definition is confusing when the phase variations exceed small values because it is possible to have phase noise values than are greater than 0 dB even though the power in the modulation sideband is not greater than the carrier power.

IEEE STD 1139 has been changed to define L(f) as Sphi(f)/2 to eliminate the confusion.

Single Sideband Phase Noise
L (f)

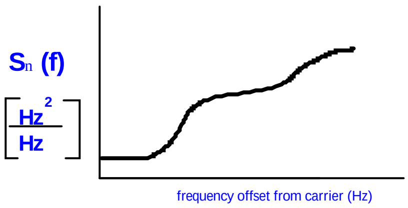
$$L(f) = \frac{S\phi\,(f)}{2}$$



Historical measurements of L(f) with a spectrum analyzer typically measured phase noise when the phase variation was much less than 1 radian. Phase noise measurement systems however measure Sphi(f) which allow the phase variation to exceed this small angle restriction. On this graph, the typical limit for the small angle criterion is a line drawn with a slope of -10 dB/decade that passes through a 1 Hz offset at -30 dBc/Hz. This represents a peak phase deviation of approximately 0.2 radians integrated over any one decade of offset frequency.

This plot of L(f) resulting from the phase noise of a free-running VCO illustrates the confusing display of measured results that can occur if the instantaneous phase modulation exceeds a small angle. Measured data, $S\phi(f)/2$ (dB), is correct, but historical L(f) is obviously not an appropriate data representation as it reaches +15 dBc/Hz at a 10 Hz offset (15dB more power at a 10 Hz offset than the total power in the signal). The new definition of L(f) = Sphi(f)/2 allows this condition since Sphi(f) in dB's is relative to 1 radian. Exceeding 0 dB simply means than the phase variations being measured are greater than 1 radian.

Sy(f) Spectral Density of Fractional Frequency Fluctuations



Sv(f) can be derived from Sφ(f)

$$\Delta f(t) = \frac{1}{2\pi}\,\frac{d\Delta\phi(t)}{dt}$$

Transformed into the frequency domain......

$$\Delta f(f) = \frac{2\pi f}{2\pi}\,\Delta\phi(f)\ \ [\mathrm{Hz}]$$

$$\Delta f^2_{rms}(f) = f^2\,\Delta\phi^2_{rms}(f)\ \ [\mathrm{Hz}]^2$$

$$S_\nu(f) = \frac{\Delta f^2_{rms}(f)}{B} = \frac{f^2\,\Delta\phi^2_{rms}(f)}{B} = f^2 S_\phi(f)\ \ \left[\frac{\mathrm{Hz}^2}{\mathrm{Hz}}\right]$$

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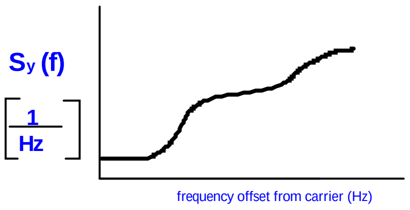
Another common term for quantifying short term frequency instability is Snu(f), the spectral density of frequency fluctuations. In this expression, the energy distribution is in frequency variance per unit bandwidth. When expressed in terms of dB, the dB value is relative to the frequency fluctuation of 1 Hz.

Sy(f) Spectral Density of Fractional Frequency Fluctuations



Sy(f) is related to Sφ(f)

$$y(t) = \frac{\Delta f(t)}{f_o} = \frac{\frac{1}{2\pi}\frac{d\Delta\phi(t)}{dt}}{f_o}$$

Transformed into the frequency domain......

$$y(f) = \frac{2\pi f}{2\pi f_o}\Delta\phi(f)$$

$$y^2_{rms}(f) = \frac{f^2}{f_o^2}\Delta\phi^2_{rms}(f)$$

$$S_y(f) = \frac{f^2}{f_o^2}\frac{\Delta\phi^2_{rms}(f)}{B} = \frac{f^2}{f_o^2}S_\phi(f)$$

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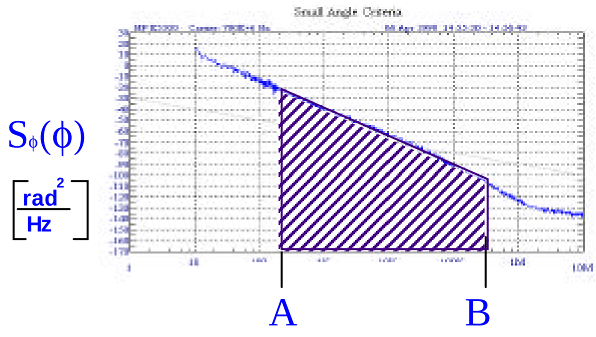
Sy(f), the spectral density of fractional frequency fluctuations allows direct comparison between sources of different carrier frequencies. Sy(f) is also related to Sφ (f) and Sv(f). Using the same Laplace transform approach we see that the spectral density of fractional frequency fluctuations is equal to the spectral density of frequency fluctuations divided by fo^2.

Sf(f) to Residual FM Conversion



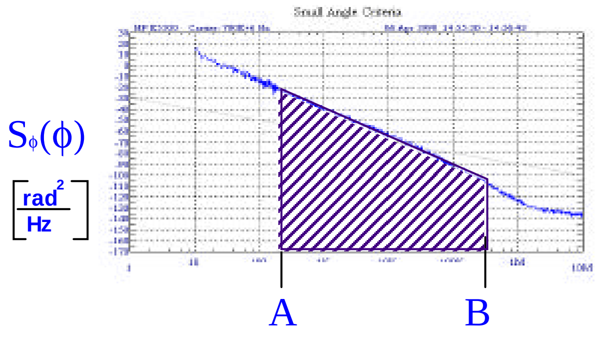
Residual FM = Total rms frequency deviation within a specified bandwidth

$$\text{Residual FM} = \sqrt{\int_A^B f^2 S_f(f)\,df} = \sqrt{\int_A^B S_n(f)\,df} \quad [\text{Hz}]$$

Residual FM is a familiar measure of frequency instability commonly used to specify noise inside a data communications bandwidth. Related to Sφ (f), residual FM is the total rms frequency deviation in a specified bandwidth. Commonly used bandwidths have been 50 Hz to 3 kHz, 300 Hz to 3 kHz, and 20 Hz to 15 kHz. Only the short-term frequency instability occurring at rates within the bandwidth are included. No information regarding the relative weighting of instability is conveyed. Therefore the energy distribution within the bandwidth is lost.

Since spurious signals are detected as FM sidebands, the presence of large spurious signals near the signal under test can greatly increase the measured level of residual FM.

$S_\phi(f)$ to Phase Jitter Conversion



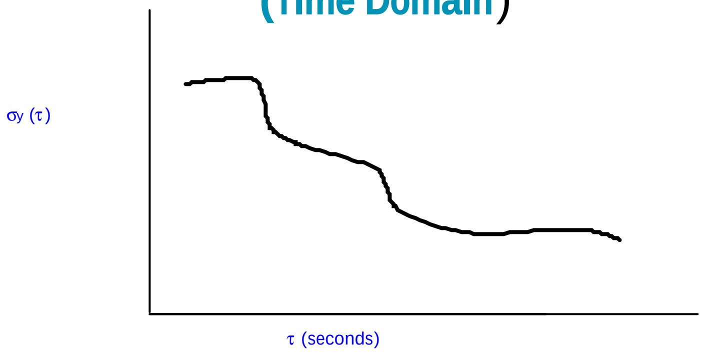
Phase Jitter = Total rms phase deviation within a specified bandwidth

Phase Jitter = $\sqrt{\int_A^B S_\phi(f)}$ [rad] Relative to 1 rad_{rms} = [dB]

Phase Jitter = $\frac{360}{2\pi}\sqrt{\int_A^B S_\phi(f)}$ [degrees]

Phase jitter can also be obtained by integrating $S\phi$ (f) over a specified bandwidth. The results can be expressed in degrees, radians or dB relative to one radian(rms).

σy(τ) Standard Deviation of Fractional Frequency Fluctuations (Time Domain)



- Statistical measure of how much the frequency varies from one time interval to the next
- Low τ represents short term frequency deviations of the carrier

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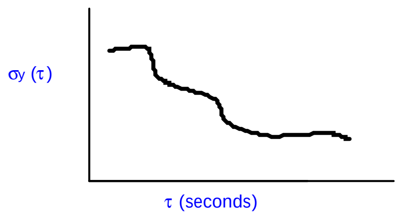
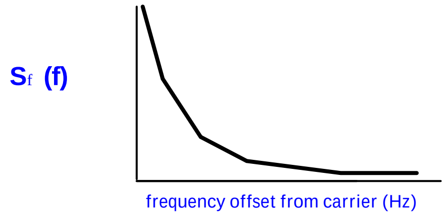
Sigma (N,T,tau,fh) for N = 2, T = tau, is the Allan Variance, is a time domain representation of frequency stability. Sigma (N,T,tau,fh) is the standard deviation of fractional frequency fluctuations, delta f/fo.

A short tau produces short-term information, while for a long tau, the short term instabilities will tend to average out producing long term information.

Relating $\sigma_y(\tau)$ to $S_\phi(f)$





$$<\sigma_y(N,T,\tau)> = \frac{N}{(N-1)} \int_0^{f_h} \frac{f^2 S_\phi(f)}{f_0^2} \frac{[\sin^2(\pi f \tau)]}{(\pi f \tau)^2} \left\{1 - \frac{\sin^2(\pi r f N \tau)}{N^2 \sin^2(\pi r f \tau)}\right\} df$$

where:

N = number of samples

T = total time (dwell plus samples)

r = T/τ

τ = sample time

f = offset frequency (variable of integration)

f_h = upper cutoff frequency of measured data

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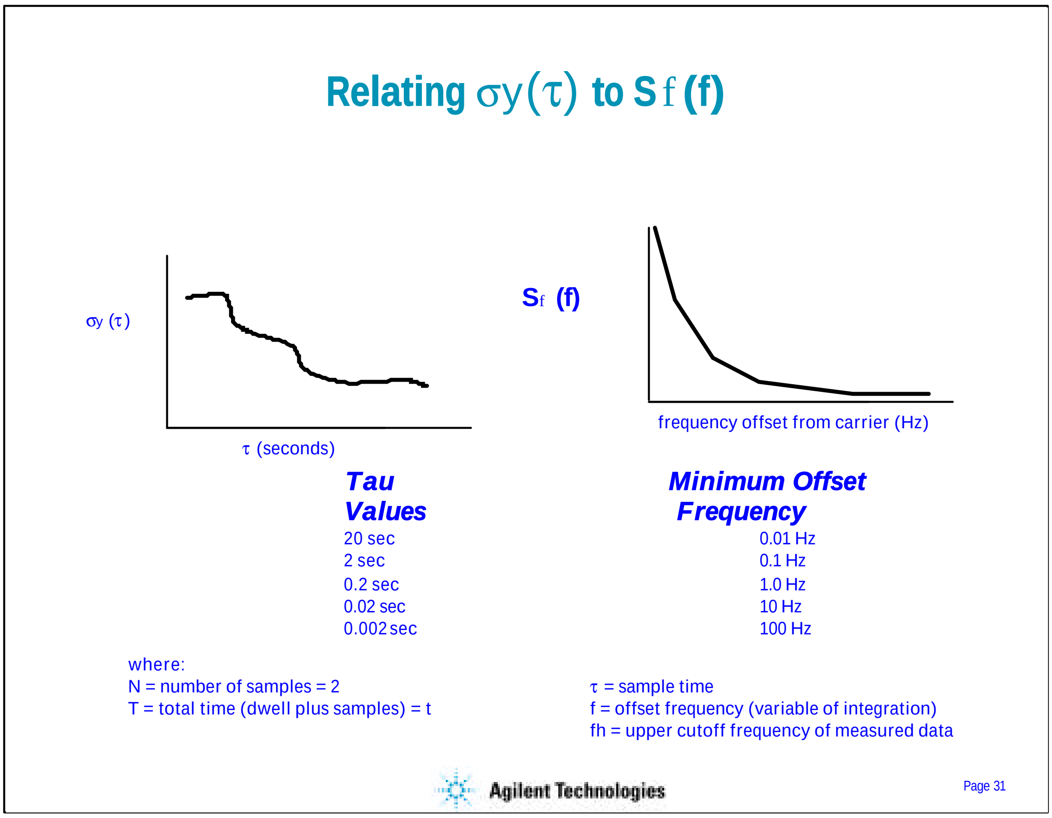
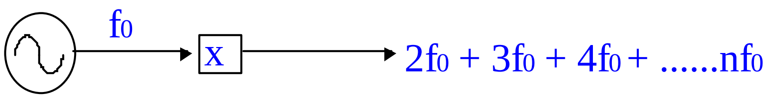
The very complex equation shown will convert phase noise from the frequency domain to the time domain. An example of a sigma tau number converted from Sϕ (f) is Sϕ(1 kHz) = -157 dBc/Hz which translates to sigma(1 msec) = 3.9 x 10(-11) when the carrier frequency is equal to 10 MHz. Note that the lower limit of the integral (0) has been replaced by the minimum offset frequency f(min).



Allan variance is a specific implementation of the sigma vs tau time domain measure of frequency instability (measurement of time domain zero-crossings). For this variance, N has been set equal to the value 2, and T (total time including dwell and samples) has been set equal to tau (sample time).

To obtain a plot of variance vs tau in the time domain, measurements are made with different time samples (tau varies). To obtain the equivalent plot of variance vs tau from measured phase noise data, make sure you have measured to the lowest offset frequency corresponding to the largest tau you wish to plot.

Noise of multiplied signals



$$\frac{V_{ssb}}{V_s} \approx J1 \approx \frac{\Delta f_{peak}}{2f_m} \approx \frac{\Delta f_{peak}}{2}$$

$$L(f_1) = \left|\frac{V_{ssb}}{V_s}\right|^2 = -6\text{ dB} + 20\log\frac{\Delta f_{peak}}{f_m}$$

If $f_2 = x.f_1$, then

$$L(f_2) = -6\text{ dB} + 20\log\frac{x\,\Delta f_{peak}}{f_m}$$

$$\frac{L(f_2)}{L(f_1)} = 20\log x$$

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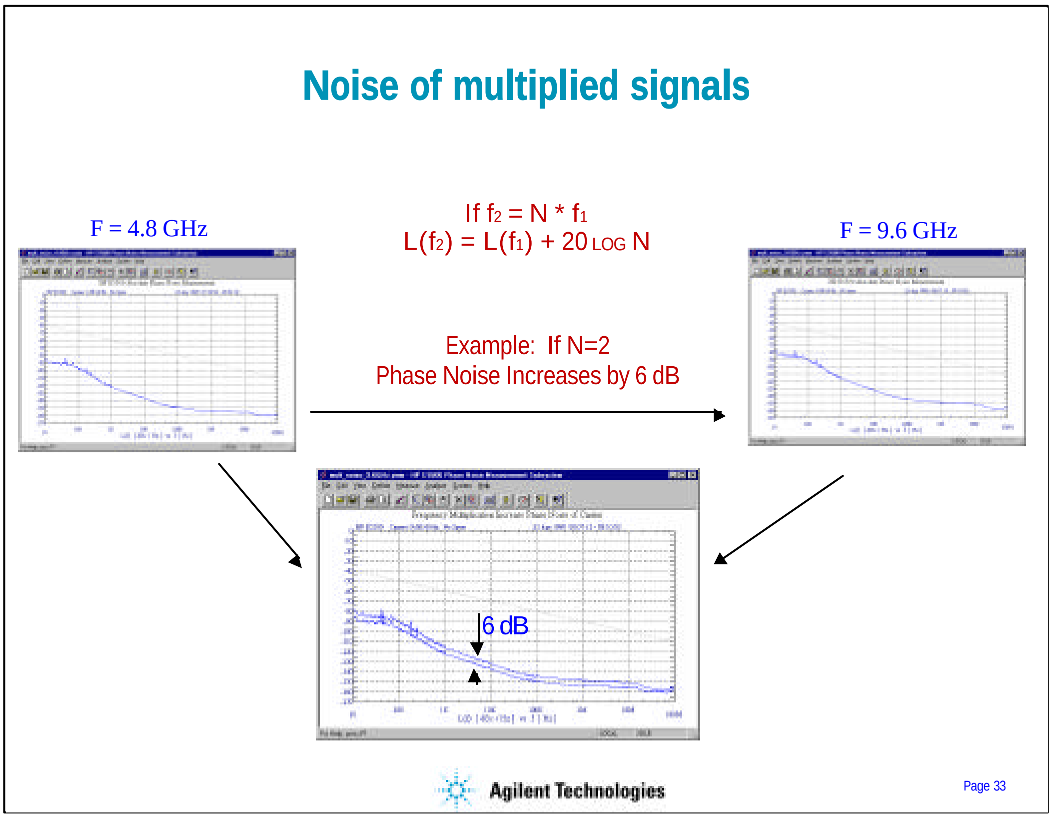
Another relationship to understand is the effect of frequency multiplication on the noise of a signal. Since phase noise (or frequency noise) can be thought of as continuous modulation signals, phase noise increases when a signal is multiplied. Essentially, for every doubling of frequency, the phase noise increases by 6 dB and conversely for every division by 2 of frequency, noise decreases by 6 dB.

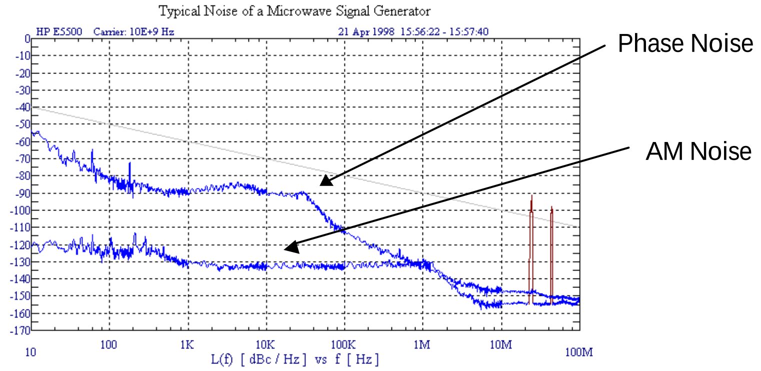


With this example, we have multiplied a 4.8 GHz signal to 9.6 GHz. As shown, the difference in the noise between the two is 6 dB.

AM Noise Power Spectral Density

$$M(f) = \frac{\text{Spectral Density of one AM Modulation Sideband}}{\text{Total Sideband Power}}$$

$$= \text{Single-Sideband AM Noise}$$



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AM noise, described here as M(f), is the power density of amplitude noise in a one hertz bandwidth relative to the carrier power. The example shown here indicates that while AM noise can often be considered to be much less than phase noise, there can offset frequencies where the AM noise can be equal to or even exceed the value of the phase noise.

Agenda

- Review Basic Phase Noise Concepts
- ● Define Noise Measurement Techniques
- Select an Appropriate Measurement Technique
- Optimizing Modern Noise Measurements
- Survey of Measurement Solutions Available

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This next section will discuss different measurement techniques used to obtain phase noise information.

Measurement Techniques

☞ Noise Measurement of Oscillators

- Spectrum Analyzer technique
- Phase Detector techniques
    * Discriminator method
    * PLL / reference source method

Additive Noise Measurement (two-port devices)

AM Noise Measurements

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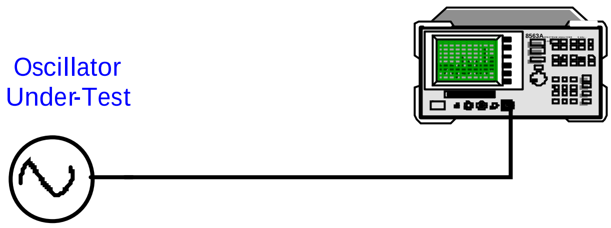
This section is separated into measuring the noise of oscillators or sources, measuring the residual or additive noise of two port devices, and the measurement of AM noise.

There are three dominant techniques used to measure the phase noise of oscillators: the direct spectrum analyzer approach, the discriminator technique and the PLL technique.

Direct Spectrum Measurement
Signal Analyzers



+ Easy to configure/use

– Measures total noise (phase noise + AM noise)
– Device drift limits close-to-carrier capability
– SA internal LO limits overall sensitivity

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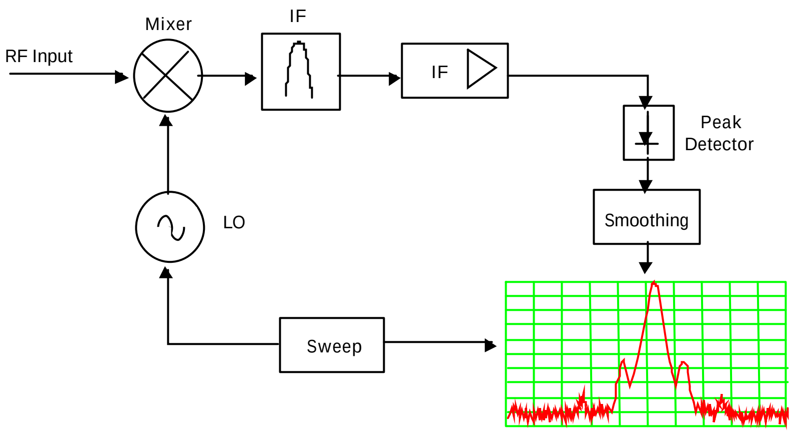
The most direct and probably the oldest method used to measure phase noise of oscillators is the direct spectrum analyzer method. Here the signal from the UUT is input into a spectrum analyzer tuned to the UUT frequency. The sideband noise power can be directly measured and compared to the carrier signal power to obtain L(f).

This approach actually measures the total sideband noise power (AM plus phase noise). If the AM noise is much less than the phase noise, the measurement can be considered to be a phase noise measurement. The sensitivity of this measurement approach is limited by the analyzers internal LO noise and the inability to track any signal drift limits the close-to-carrier measurement capability of the analyzer.

Direct Spectrum Measurement
Signal Analyzers



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Factors which must be accounted for when using a swept spectrum analyzer directly are the shape of the IF filter when the IF filter deviates from an ideal rectangular filter and the compensation for effects of the log amplifiers and peak detectors. Many spectrum analyzer now have an automatic correction function to take into account the equivalent noise bandwidth.

Measurement Accuracy Issues

- BW normalization (shape of IF filter)

- Amplitude accuracy (noise vs coherent signals)

- Inclusion of AM noise

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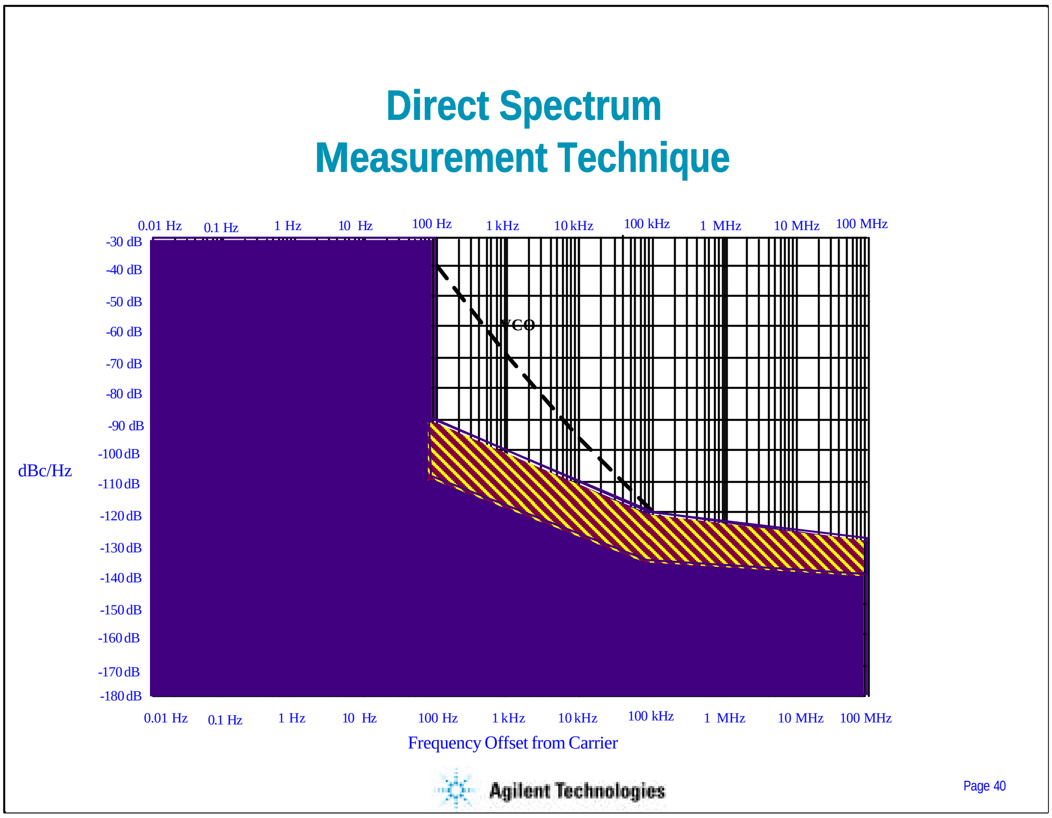
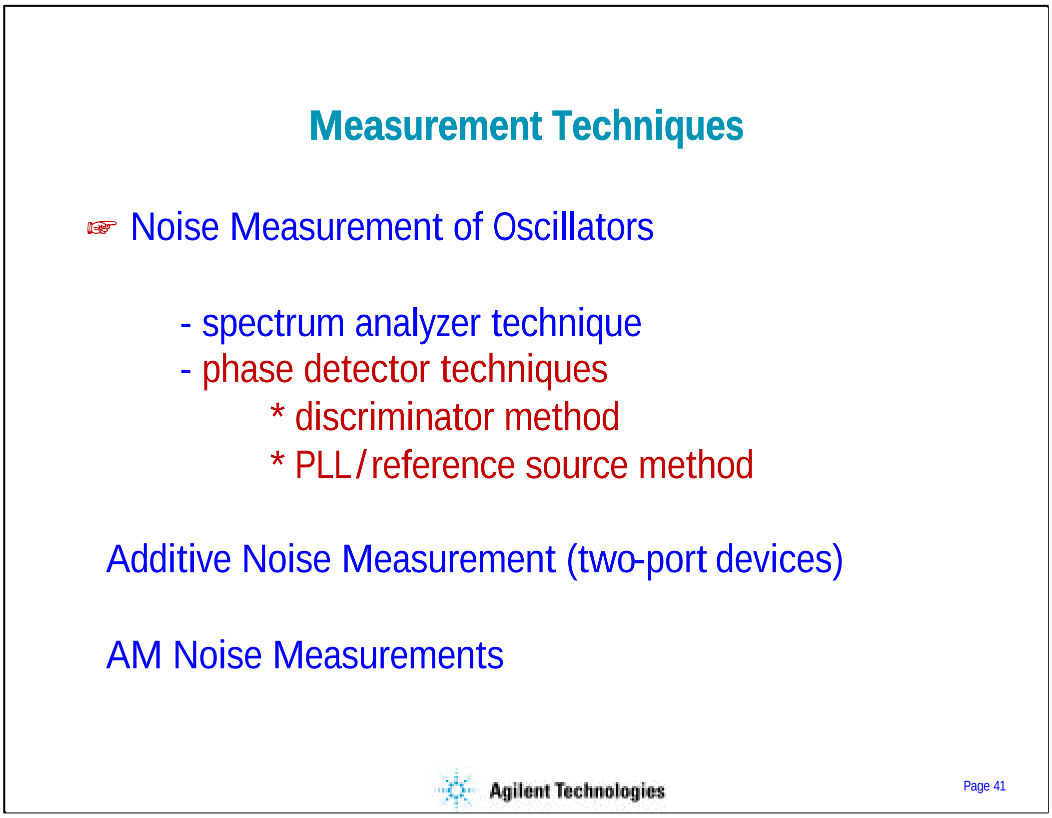


The available noise measurement range is shown here. The solid color area is typically not available when you use the direct spectrum technique. The boundaries tend to be slightly different for different analyzers and different frequency ranges.



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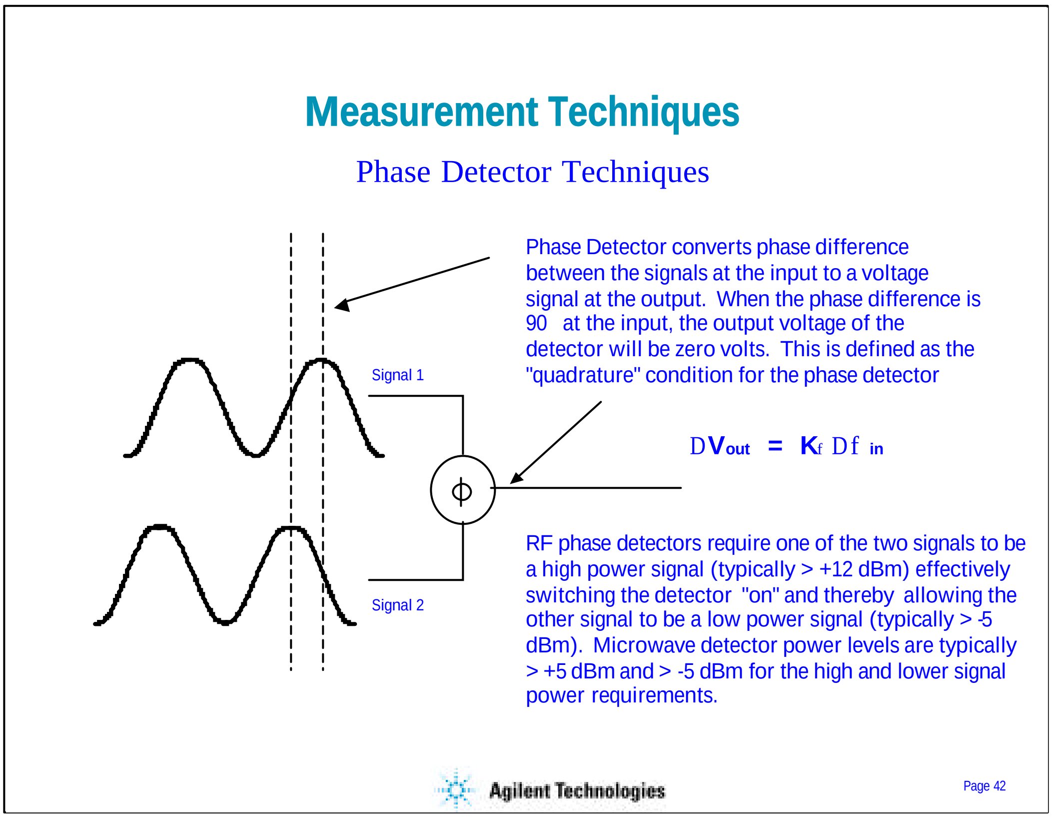
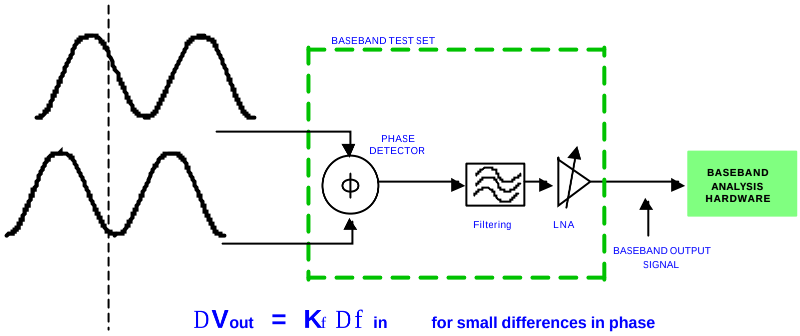


To separate phase noise from amplitude noise, a phase detector is required. The phase detector converts the phase difference of the two incident signals in to a voltage at the output of the detector. When the phase difference is set to 90 degrees (quadrature), the voltage output will be zero volts. Any phase fluctuation from quadrature will result in a voltage fluctuation at the output. When quadrature is not maintained, an error can be introduced into the results based on the amount of phase delta from quadrature. The error (dB) = 20 log [Cos(phase deviation from quadrature)]. Another expression is delta V = k(phi)*sqr(1-10((error dB/10).

Phase detectors (usually double balanced mixers) typically require large power signals at the input port to operate properly. One of the signals must be of high power to switch on the diodes in the detector allowing the other signal to be of lower power - typically greater than 0 dBm.

Measurement Techniques

Phase Detector Techniques



$\Delta V_{out} = K_\phi \Delta\phi_{in}$ **for small differences in phase**

Filtering	protects LNA from LO feedthrough & mixer sum products
LNA	improves sensitivity of baseband analyzers
Baseband Analysis	measures baseband voltage - frequency and magnitude

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For phase noise measurements, the phase difference between the two signals will be very small. This yields a small change in voltage output from the phase detector. The filtering block protects the LNA from LO feedthrough and mixer sum products. It also allows the measurement of low noise far from carrier even though the noise close-to-carrier may be very large.

The LNA improves the sensitivity of the baseband analyzers.

The baseband analyzer measures the voltage fluctuations to extract the desired magnitude and frequency information. This analysis hardware can vary from digitizers to FFT analyzers to swept RF spectrum analyzers depending on the flexibility of the phase noise measurement solution being used.

Phase Detector Techniques

- FM Discrimination

- Reference Source with PLL

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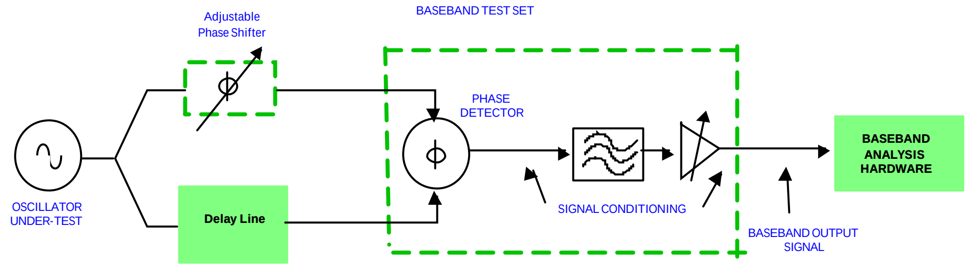
There are two different measurement techniques which use a phase detector (along with associated filters, LNA & baseband analyzers): 1) the FM discriminator measurement technique and 2) the PLL with reference source measurement technique.

FM Discriminator Measurement Technique

Delay Line Discriminator Method



$$\Delta V = K_\phi \, 2\pi\tau \, \Delta f \, \frac{\sin \pi f_m \tau}{\pi f_m \tau}$$

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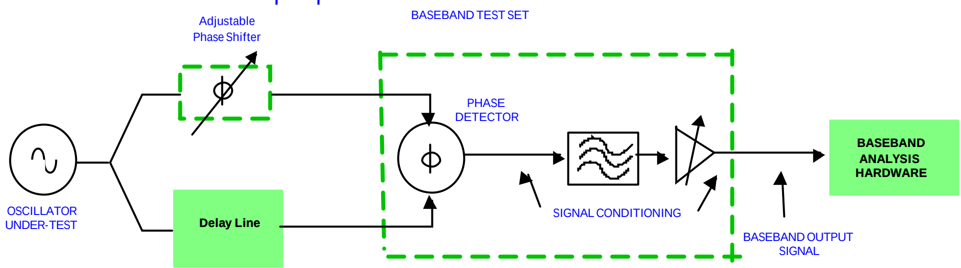
While there are many different implementations of frequency discriminators such as the cavity resonator, the RF bridge/delay line, and the delay line/mixer, we will concentrate this discussion on the delay line/mixer implementation.

The delay line converts frequency fluctuations of the source into phase fluctuations relative to the signal at the other port of the phase detector. The phase detector converts the phase fluctuations into voltage fluctuations for measurement and analysis.

The discriminator constant, Kd, is calculated from the phase detector constant, $K\phi$, and the amount of delay used. Note that the discriminator constant is independent of offset frequency f for $f<=1/2\pi.\tau$. Measurements at higher offset frequencies require correction for the sin(x)/x term of the discriminator constant.

FM Discriminator Measurement Technique

- Discriminator noise floor (delay length, max offset freq)
- Effect of DUT AM noise on the discriminator noise floor
- DUT output power



$$\Delta V = K_\phi \, 2\pi\tau \, \Delta f \, \frac{\sin \pi f_m \tau}{\pi f_m \tau}$$

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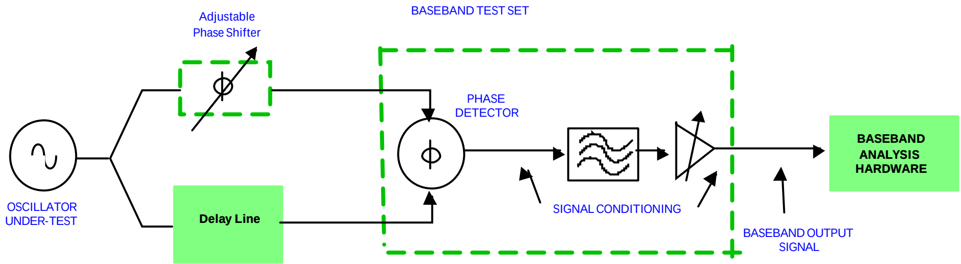
For the discriminator technique, the measurement noise floor and the maximum offset frequency are determined by the amount of delay used. Other factors which can affect the measured results are the presence of high AM noise from the UUT and low output power of the UUT.

AM noise is normally suppressed 20-30 dB by the phase detector (relative to the phase noise present). Since phase noise is suppressed close-to-carrier by 60-100 dB - due to the effects of the discriminator - the relative contribution from AM noise will be enhanced. Oscillators which use GaAs output amplifiers are at more risk due to the high AM noise contribution of GaAs devices.

If the UUT output power is not sufficient to overcome power losses in the signal splitter and the delay line, the power presented to the phase detector will be less. Problems ranging from an increase in the discriminator noise floor to the phase detector not functioning may result. Adding low noise amplifiers prior to the signal splitter or prior to the phase detector solve the low power problem, but the additive noise of the amplifiers will cause the discriminator noise floor to increase.

FM Discriminator Measurement Technique

Discriminator Method
with Frequency Downconversion



- Discriminator noise floor (delay length, max offset freq)
- DUT output power
- Effect of DUT AM noise
- Downconverter Noise

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When the carrier frequency increases, the losses in the delay line can become prohibitive. A low noise downconverter can be used to translate the carrier frequency to the nominal frequency range where the discriminator technique works the best. The noise of the downconverter may limit the noise measurement of the UUT far-from-carrier where the sensitivity of the discriminator is the best.

FM Discriminator Measurement Technique

Discriminator Sensitivity

$$S_n(f) = \frac{S_{Vrms}(f)}{K_d^2} \quad \left[\frac{Hz^2}{Hz}\right]$$

$$S_f(f) = \frac{S_{Vrms}(f)}{f^2 K_d^2}$$

The sensitivity of the discriminator technique is approximately equal to the phase detector system sensitivity at $f = 1/(2\pi\tau)$ and tipped up by 20 dB/decade for $f < 1/(2\pi\tau)$.

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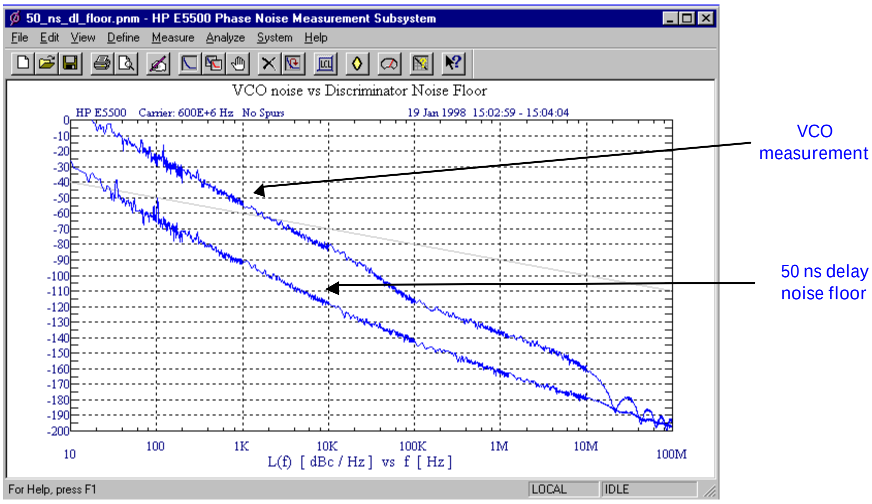
These equations show the why the noise floor increases by 20 dB/decade as the offset frequency gets smaller (closer-to-carrier). Increasing the delay improves the noise floor closer-to-carrier but reduces the power to the phase detector. It also reduces the maximum offset frequency that can be measured with no sin(x)/x correction.

FM Discriminator Measurement Technique
Discriminator Noise Floor



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FM Discriminator Measurement Technique

Discriminator Constant K_d

$$\Delta V = K_\phi \, 2\pi\tau \, \Delta f \, \frac{\sin \pi f_m \tau}{\pi f_m \tau}$$

$$\Delta V = K_\phi \, 2\pi\tau \, \Delta f \qquad \text{for } f_m < 1/(2\pi\tau)$$

The calibration of the system from the discriminator constant, Kd, is valid up to an offset frequency of $f = 1/(2\pi\tau)$ and to $f = 1/(2\tau)$ with correction for the sin x / x term.

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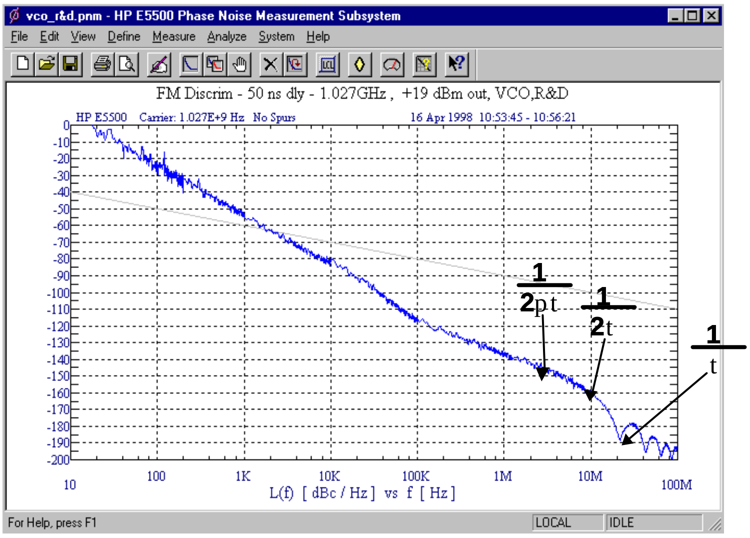
Most implementations for the discriminator technique do not provide corrections for the sin(x)/x response within the discriminator constant. This is mostly due to not knowing the exact amount of delay used for this measurement. For a specific delay implementation, you can correct the measured data for this region of offset frequencies.

FM Discriminator Measurement Technique

Typical Measurement Result using 50ns delay line



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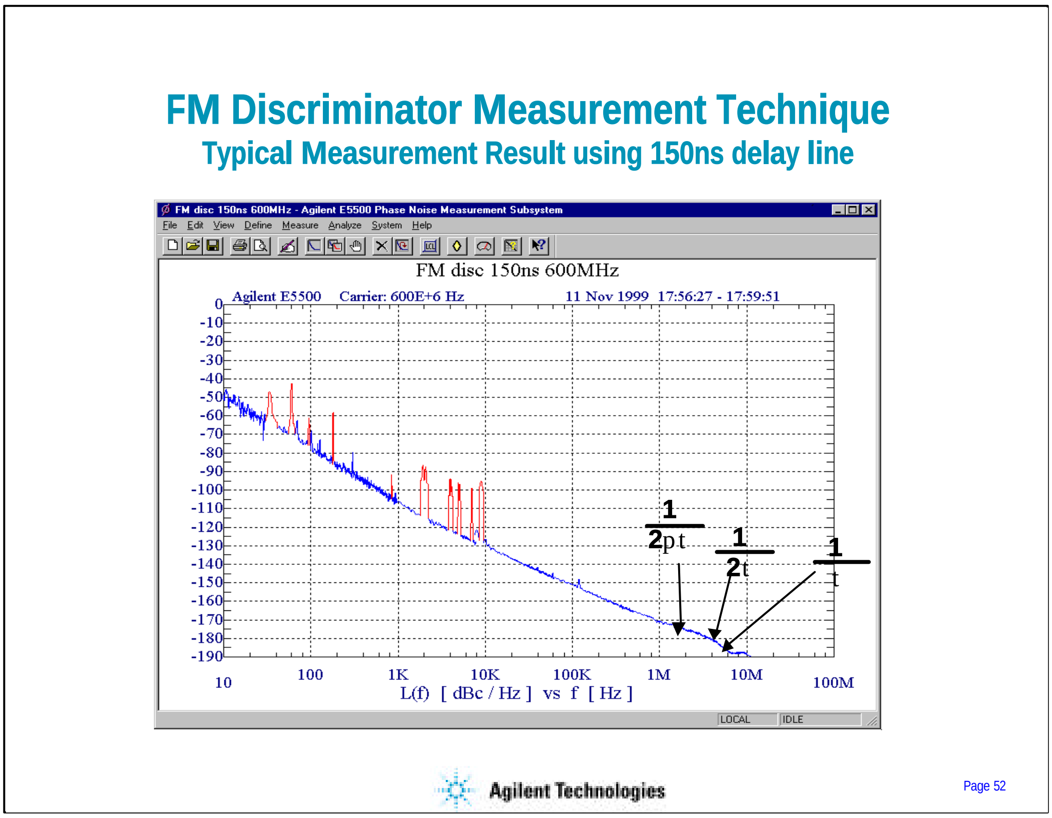
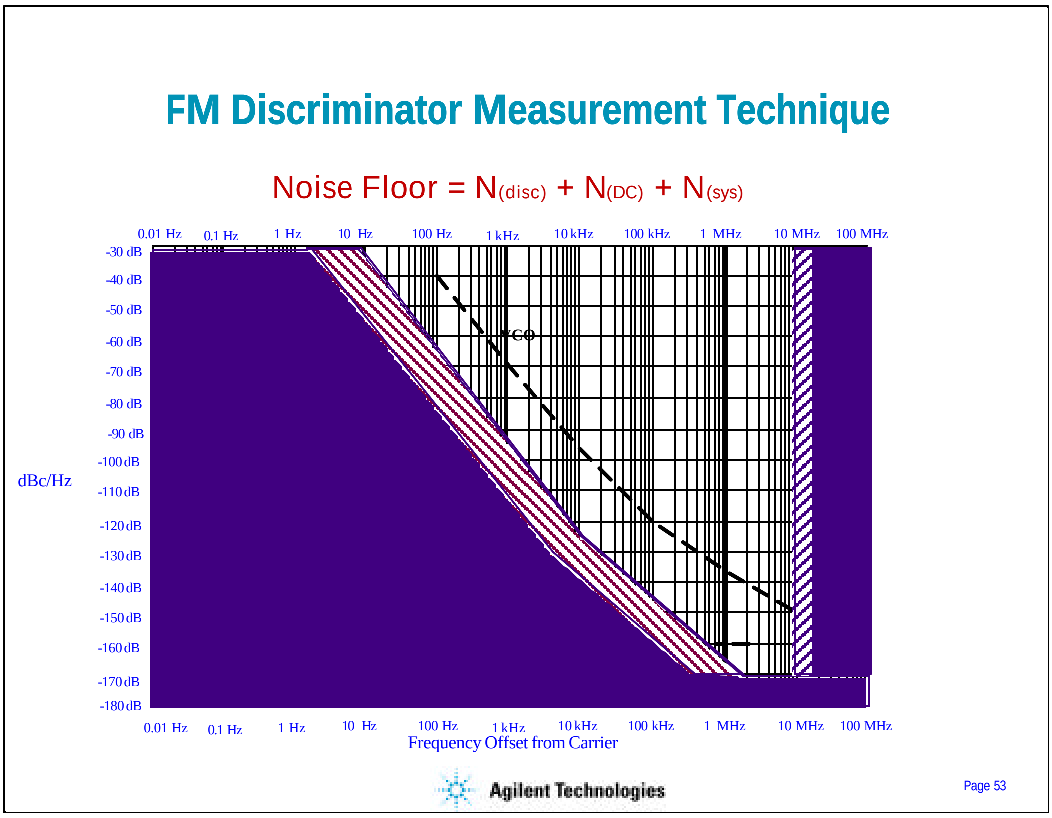
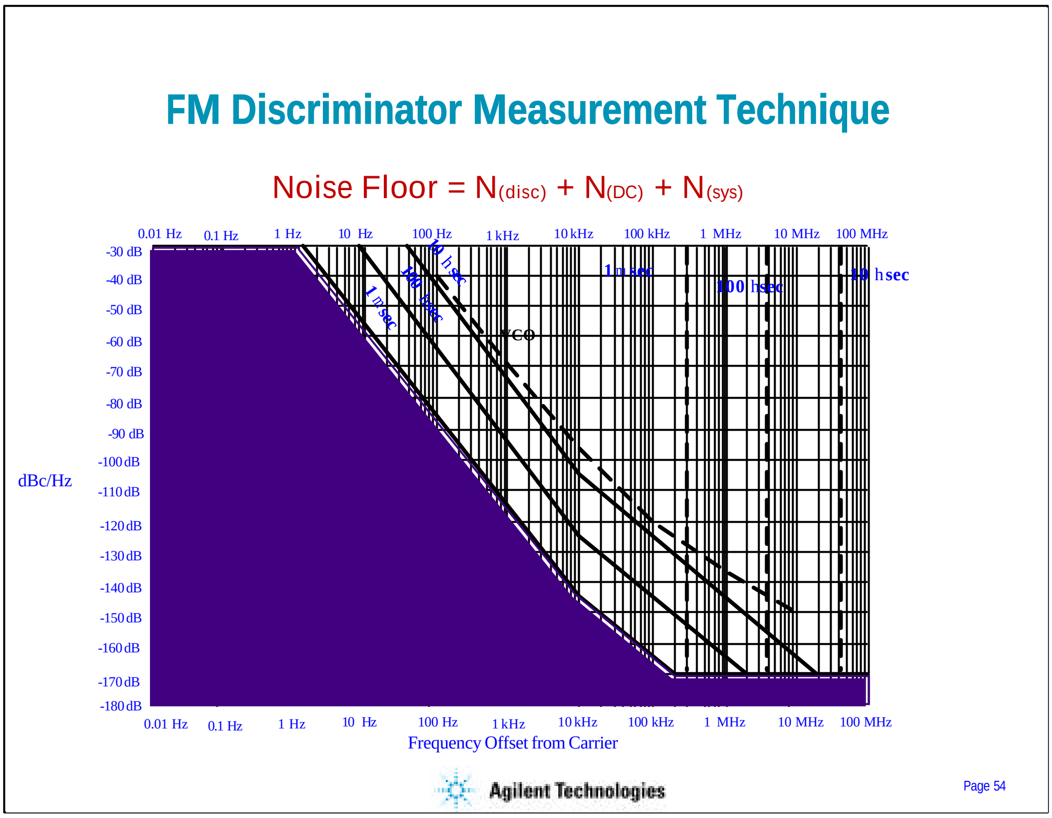
For a 50 ns delay line, the 1/2πτ, 1/2τ, and 1/τ offset frequencies are indicated here. A correction for 1/2τ is upward about 4 dB.



This slide shows another example of a high quality VCO at 600 MHz being measured using a 150 nS delay line. The signal level into the input splitter was quite large allowing the phase detector to be driven optimally and leading to an excellent sensitivity at the 1/2 π tau offset.



The overall noise floor of a discriminator measurement will be the noise floor of the discriminator plus the noise of the measurement system. The noise of a downconverter will limit measurements for offset frequencies far from carrier only. The noise floor of the discriminator dominates other contributors for close-to-carrier measurements. The solid color areas on this graph are not available for this technique.



If the amount of delay is changed, the noise floor of the discriminator will change. For a 10 nsec delay, the offset frequency where the discriminator noise floor intersects the phase detector noise floor is approximately 16 MHz ($1/2\pi\tau$). For a delay of 100 nsec or a delay of 1 usec, the noise floor closer-to-carrier improves. While longer delays improve the noise floor, the eventual power loss in the delay line will exceed the source power available and cancel further improvement. Also, longer delay lines limit the maximum offset frequency that can be measured.

FM Discriminator Measurement Technique Accuracy Issues

- measurement of Kd
- measurement accuracy of baseband analyzer
- quadrature maintenance
- frequency response of test set
- frequency discriminator flatness
- high frequency mismatch/flatness (> 1 MHz)
- impact of AM noise present

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These are the major factors which can cause measurement inaccuracy. An uncertainty budget of +/- 2 dB for offsets < 1 MHz and +/- 4 dB for offsets < 100 MHz comprise limits based on the ability to verify this uncertainty.

For offset frequencies < 1 MHz, the error budget is based on an assumed +/- 1 dB uncertainty for the measurement of Kd and +/- 1 dB for the RSS combination of the uncertainty of the instrumentation error, quadrature maintenance, discriminator flatness, and the overall frequency response of the baseband test set.

For offset frequencies > 1 MHz, the additional +/- 2 dB of uncertainty is attributed to the high frequency mismatch uncertainties and amplitude flatness of the hardware configuration.

The presence of AM noise can also cause measurement inaccuracy but is not accounted for within the normal uncertainty budgets.

FM Discriminator Measurement Technique

Summary

Advantages:

- ✓ Sensitivity matches free-running VCO characteristics
- ✓ Requires only VCO-under-test
- ✓ Low noise floor far-from-carrier

Disadvantages:

- ✗ Poor sensitivity close-to-carrier
- ✗ May be difficult to implement with large values of delay (signal loss)
- ✗ Limited offset frequencies
- ✗ AM noise effects "enhanced"

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FM Discrimination is a good match for most VCO's. There are, however, limitations to consider and address for any specific VCO.

Phase Detector Techniques

- FM Discrimination

☞ - Reference Source with PLL

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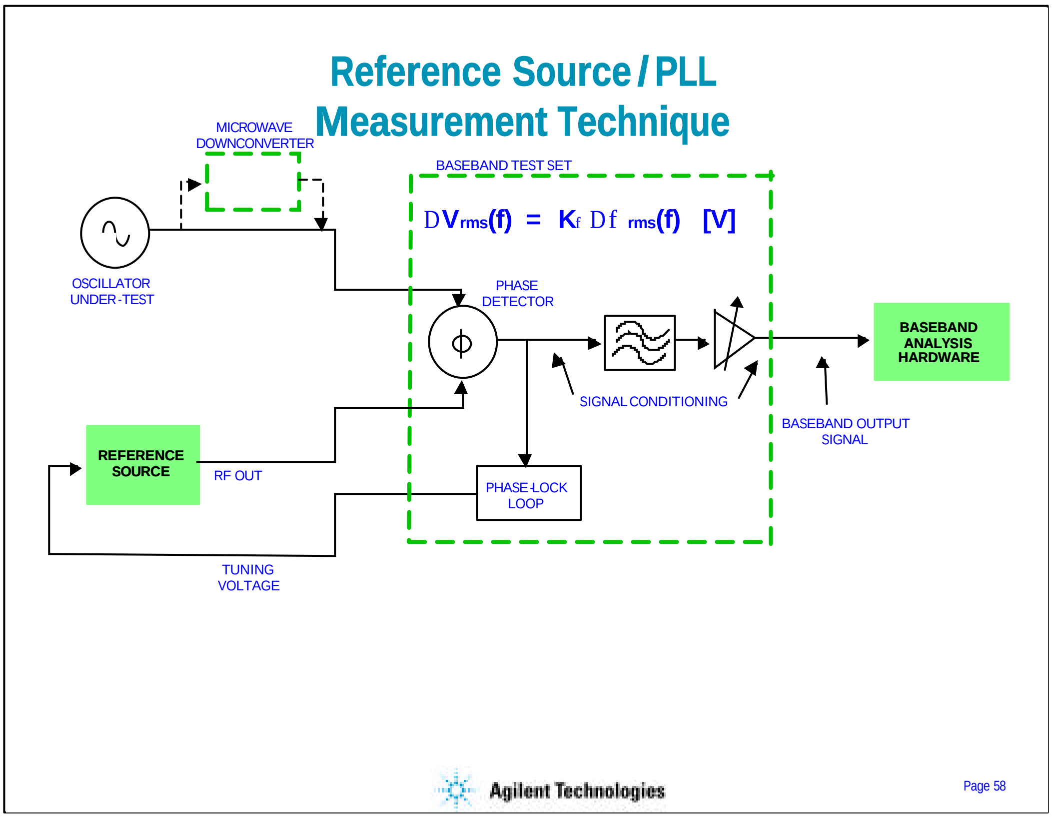
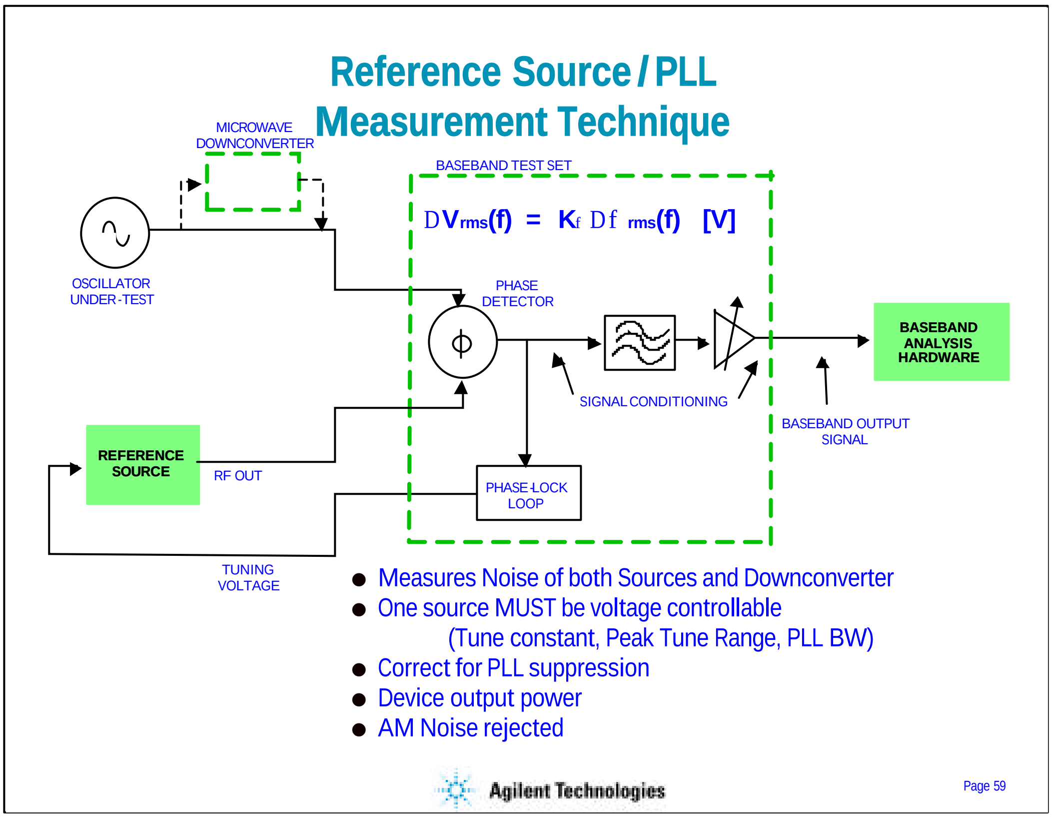
The next measurement approach which uses the phase detector is the phase-lock-loop with reference source technique.



Within this technique, another source is used to provide the reference phase signal for the phase detector. The phase-lock-loop is used to control either of the two sources and establish phase quadrature at the phase detector. This means that one of the two sources used in this approach must have DC voltage control capability. The phase noise that is measured at the phase detector is the sum of the mean square phase fluctuations. If one source has much lower phase noise characteristics, then the measurement results will reflect the higher phase noise of the other source.

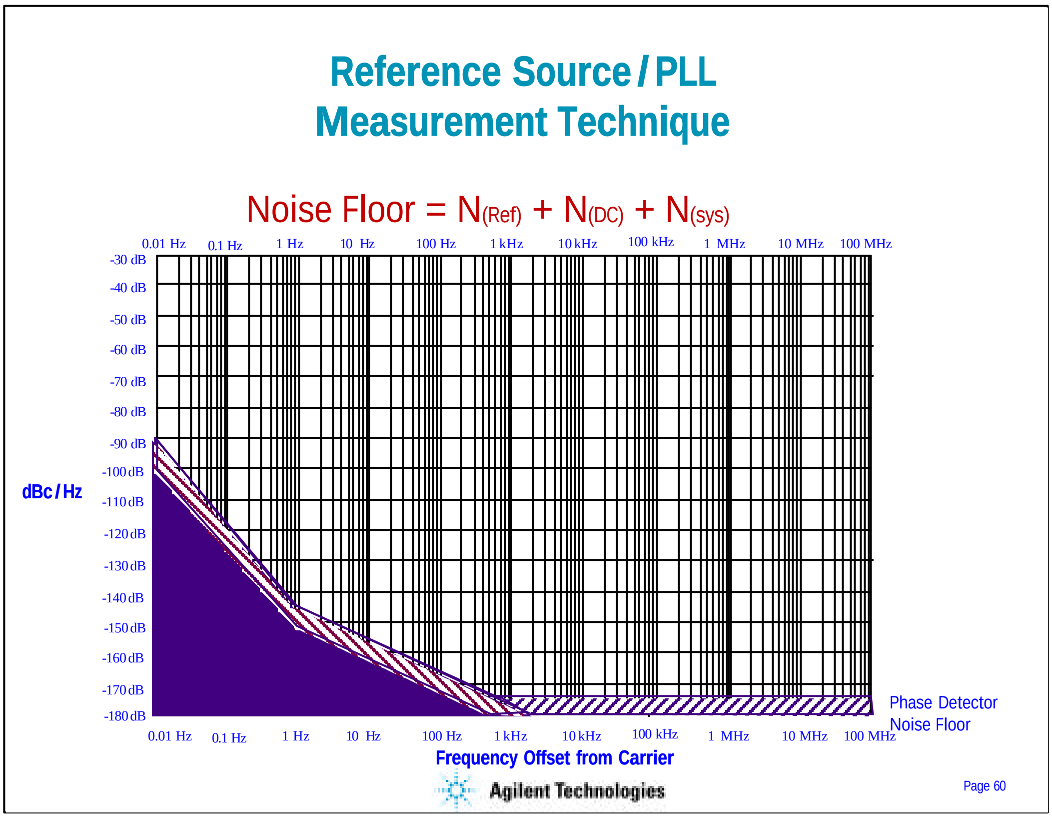
Like the previous FM discriminator measurement technique, low noise downconverters can be used to translate the carrier frequency of the UUT to a lower IF frequency for measurement purposes. Reference sources such as RF signal generators have much lower phase noise characteristics in the < 1 GHz range than any microwave signal generator may have at the carrier frequency of the UUT. The noise of the downconverter, when used, many times becomes a limiting factor in the overall system measurement noise floor.



For noise measurements within the PLL bandwidth, the measured results are suppressed by the PLL. Since the characteristics of the PLL are known, a correction factor can be developed and applied to results within the closed loop bandwidth to achieve accurate results.

Low power levels from the UUT or the reference source can cause the phase detector noise floor to rise or the phase detector to not operate. Low noise amplifiers prior to the phase detector can help solve this issue but the residual noise of the amplifiers will add to the phase detector noise floor. An increase in the system noise floor is an issue only when the noise of the UUT is about the same magnitude as the noise floor itself.

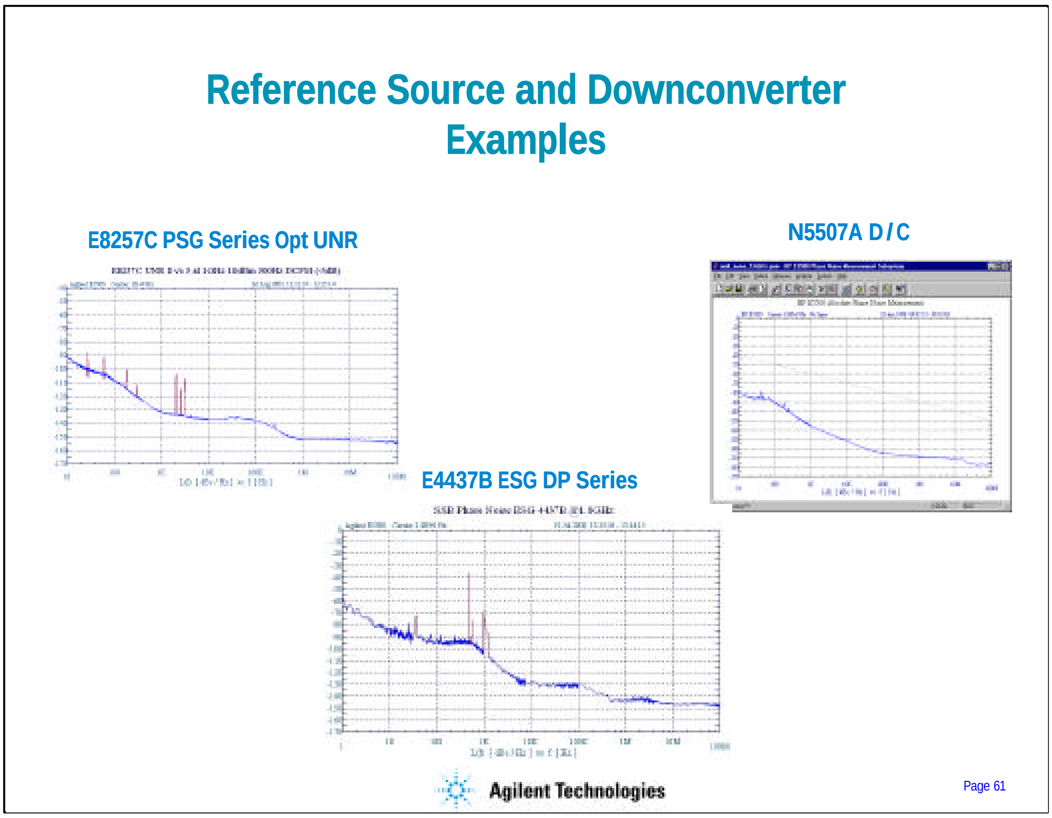
The normal rejection for AM noise is about 20-30 dB with a double-balanced mixer used for the phase detector.



The overall noise floor of this measurement approach is the combination of the noise of the system (N(sys) = phase detector, LNAs, baseband analyzers) plus the noise of reference source(N(ref)) plus the noise of the microwave downconverter (N(DC)).

This graph shows the typical noise of the system with no reference sources or downconverters. The variation in the overall noise floor is a function of signal power from the UUT and the additive noise of any internal amplifiers used in front of the phase detector.

The rise in noise for close-to-carrier offset frequencies is due to the 1/f noise of the phase detector and the LNA's.



Here are some examples of a couple of different RF signal generators and a microwave downconverter. Notice that neither of the two sources have both low noise close-to-carrier AND low noise far-from-carrier. This trade-off issue exists for all commercially available RF signal generators.

Reference Source Noise

HOW MUCH LOWER SHOULD THE REFERENCE BE?

$\mathcal{L}$ dut - $\mathcal{L}$ 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

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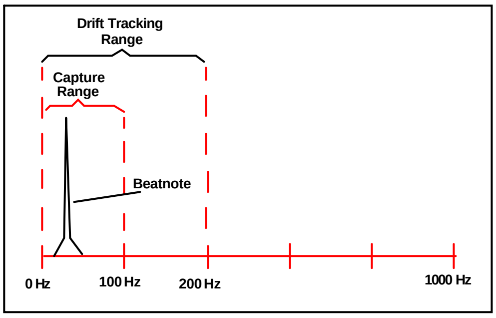
When the noise of the reference source (or downconverter) is the same as the UUT noise, the measured noise will be 3 dB higher than either of the two signals. If the reference source is more that 15 dB lower, the measured noise will - for all practical purposes - be the noise of the UUT.

Voltage Controlled Source

- Peak Tune Range (PTR) = Tune constant (K_{VCO}) x Tune Range (V)
- Capture Range = 10% of PTR
- Drift Tracking Range = 20% of PTR



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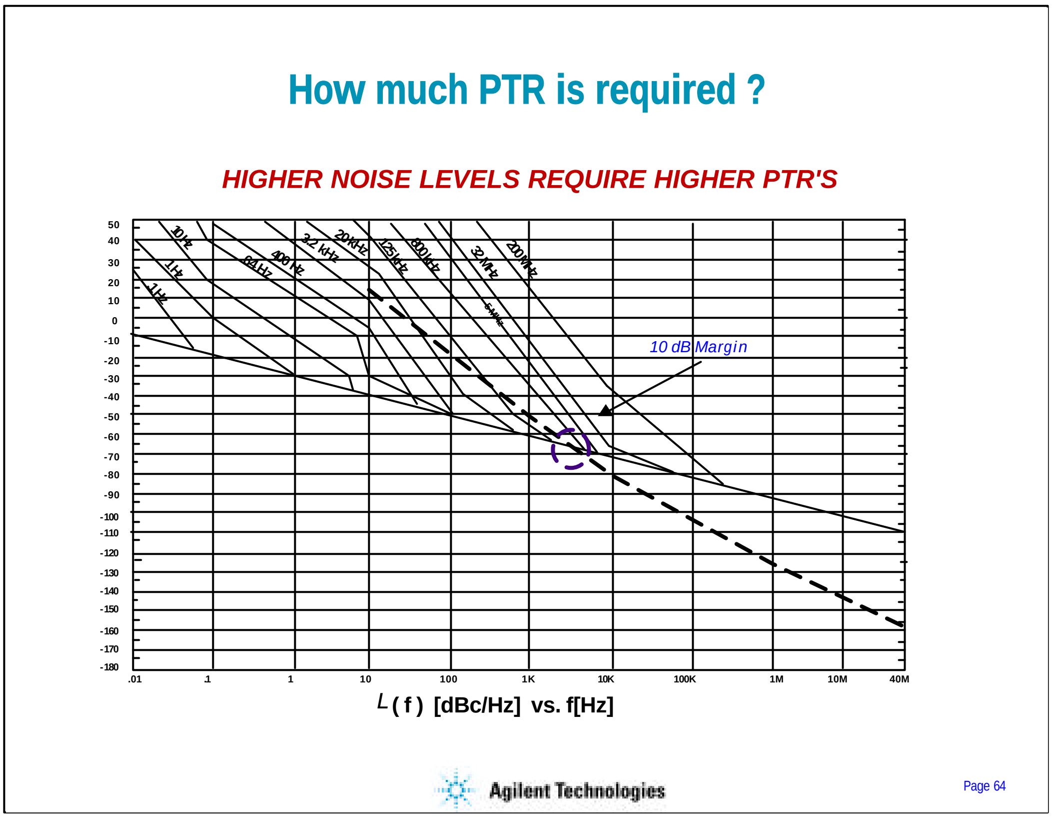
One of the two sources used in the PLL with reference source technique must be voltage controllable. The tune constant for the source with voltage control will be in Hz/V. The tunable frequency range or peak tune range (PTR) is the source tune constant in Hz/V multiplied by the available voltage range (typically < 10 V).

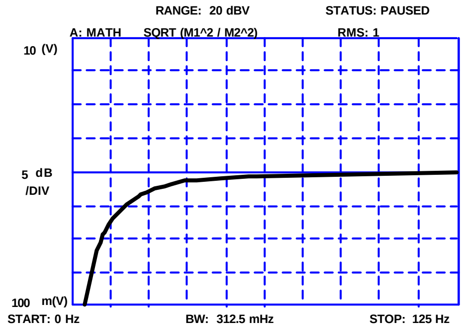
The frequency difference of the two sources at the phase detector inputs be less than 10% of the PTR for the PLL to close. Once the PLL is closed, it will track frequency drift difference between the two sources up to about 20% of the PTR before losing phase-lock.



The amount of PTR necessary to allow the PLL to close and the measurement to be completed is a function of the noise of the UUT. Higher noise levels require higher PTR's.

The general rule is to choose a PTR from this graph when you have about a 10 dB margin above the expected noise at the crossing of the small angle line. In this example, a 10 dB margin above the noise where it crosses the small angle line would indicate that a minimum 5 MHz PTR should be used for this particular device.

PLL Suppression



Suppression Correction: Theoretical (assumes linear Kvco)
Verification (measures actual PLL suppression)
None (no measurement < PLL BW)

PLL Suppression Verification: - detects errors in calibration constants
- improves suppression correction accuracy confidence

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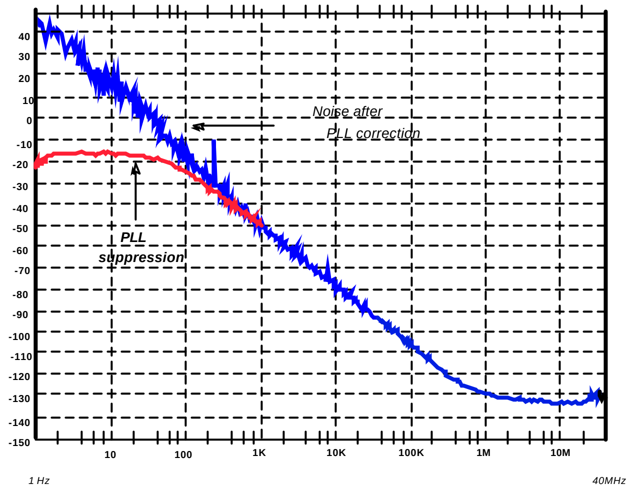
The PLL suppresses noise for frequencies less than the closed loop bandwidth. The parameters for the PLL determine the magnitude and bandwidth of the closed loop suppression (loop filters, Kvco, Kϕ, PTR). If these loop parameters are known, then a correction factor can be calculated for the closed loop bandwidth and applied to the measured data to provide accurate noise measurement results.

Suppression verification measures the actual loop suppression characteristics and compares the results to the "theoretical" value. This process detects errors in the detector constant, the tune constant linearity, limited VCO tune port bandwidth conditions, and injection locking conditions. It therefore improves confidence in the entire measurement, not just inside the PLL bandwidth.

Of course, if the minimum offset frequency of the measurement system is always greater than the closed loop bandwidth, no correction is necessary. Some measurement solutions always correct for loop suppression and others never do.

PLL Suppression Correction



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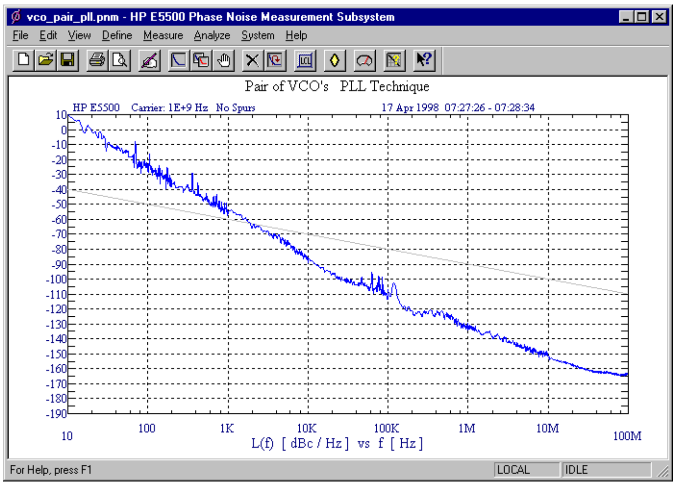
This graph is an example of PLL Suppression correction. If a correction value is not applied to the actual measured noise for offset frequencies less than the PLL BW, then the results will be suppressed as compared to the actual noise of the source being measured.

Example VCO PLL Measurement



Two Vco's:
K_{vco} = 35 MHz/V
PTR = 210 MHz

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Here is an example of a measurement of two VCO's. The nominal tune constant of the reference VCO is 35 MHz/V. The PTR of 210 MHz is a result of using a 6 volt tune range. Notice that the only maximum offset frequency limitation is that of the measurement system itself.

PLL with Reference Source Technique

Advantages

- ✓ Lowest overall noise floor available
- ✓ Wide offset frequency range
- ✓ Wide input (carrier) frequency range
- ✓ Provides AM suppression

Disadvantages

- ✗ Requires two sources
- ✗ High drift rate sources require high PTR

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In summary, the PLL technique has the best overall system sensitivity for oscillator measurements and generally considered the first choice among measurements techniques for sources.

Measurement Accuracy Issues

- measurement of K_ϕ
- measurement accuracy of baseband analyzer
- quadrature maintenance
- frequency response of test set
- loop suppression correction
- high frequency mismatch/flatness (> 1 MHz)

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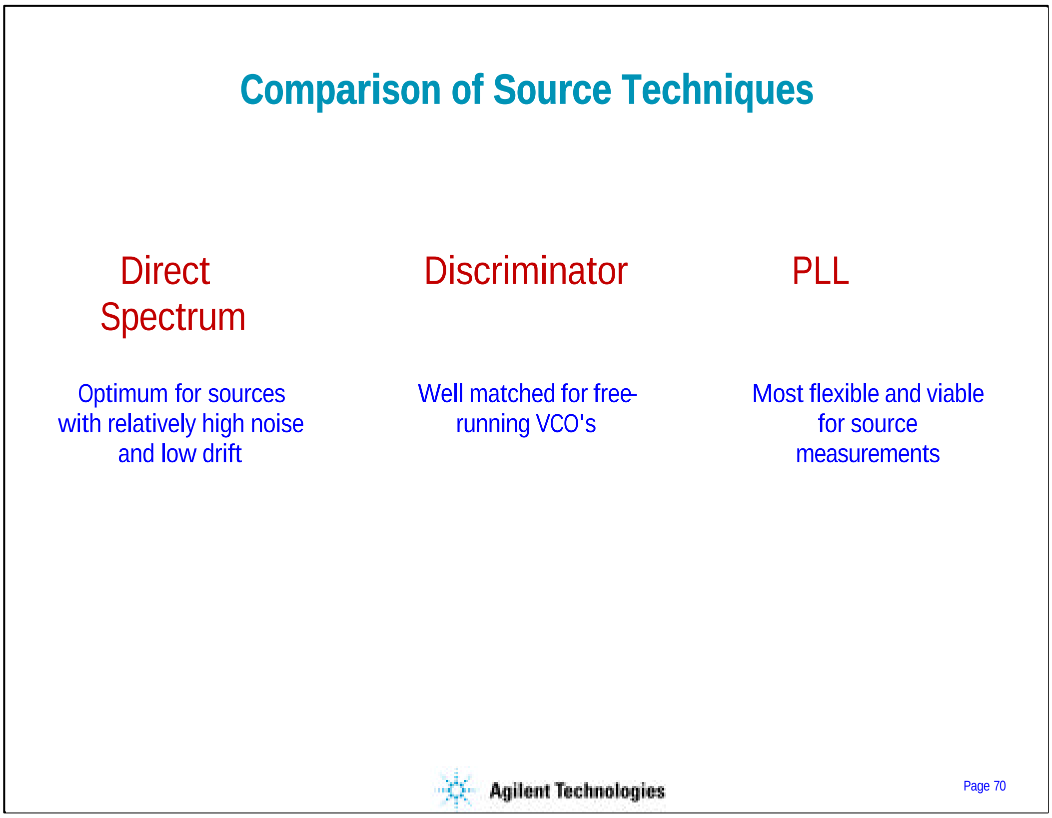
These are the major factors which can cause measurement inaccuracy. A typical uncertainty budget of +/- 2 dB for offsets < 1 MHz and +/- 4 dB for offsets < 100 MHz comprise an RSS summation of these factors.

For offset frequencies < 1 MHz, the error budget is based on an assumed +/- 1 dB uncertainty for the instrumentation error of Kphi and +/- 1 dB for the RSS combination of the uncertainty of the baseband analyzer, quadrature maintenance, and the overall frequency response of the baseband test set.

For offset frequencies > 1 MHz, the additional +/- 2 dB of uncertainty is attributed to the high frequency mismatch uncertainties and amplitude flatness of the hardware configuration.



For source measurements, the PLL technique offers the most flexible and viable solution while the discriminator technique and the direct spectrum techniques are good solutions for specific types of oscillators.

Measurement Techniques

Noise Measurement of Oscillators
- Spectrum analyzer technique
- phase detector techniques
- discriminator method
- PLL / reference source method

☞ Additive Noise measurement (two-port devices)

AM Noise measurements

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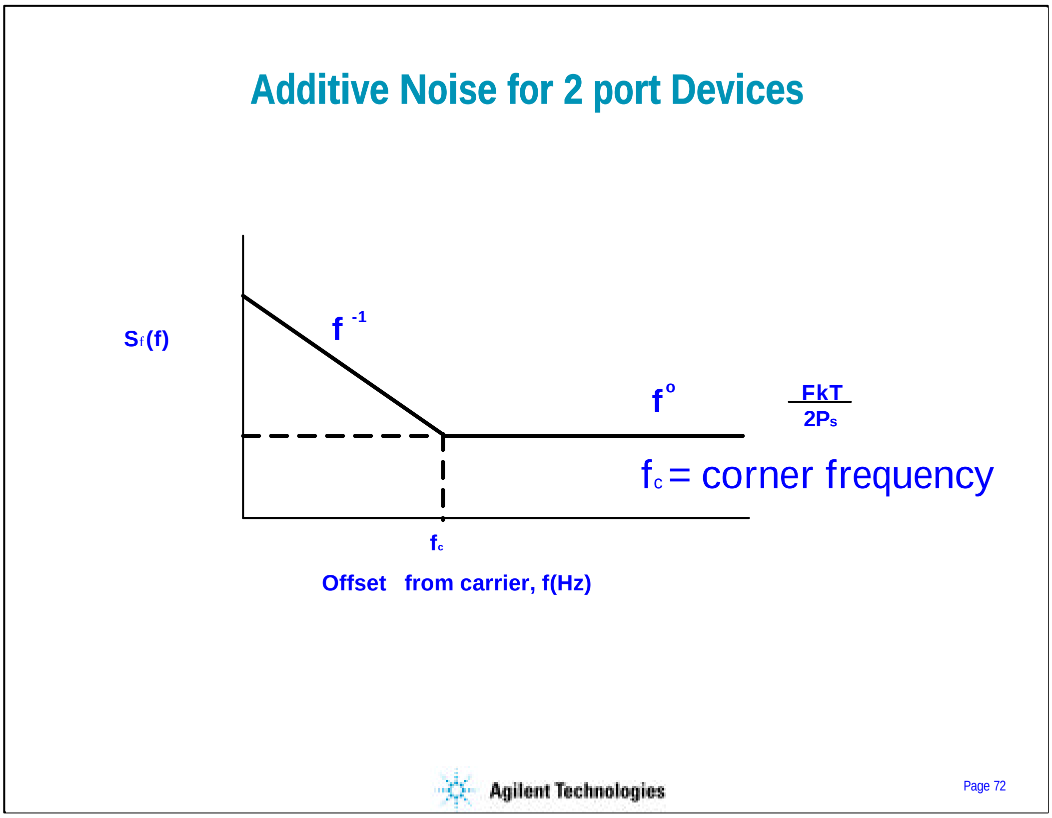
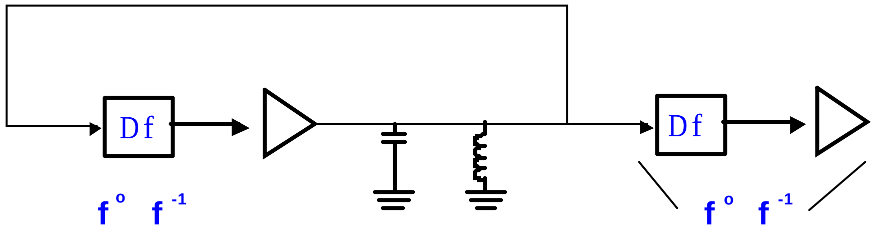
This next section will focus on the measurement of additive or residual phase noise for two-port devices.



Remember the typical 1/f noise curve for an amplifier? The two components which comprise additive noise is the flat, white noise component and the 1/f noise component. Both of these noise component are part of the device's "noise figure".

Noise Processes in an Oscillator



becomes f^{-2} & f^{-3}

thru $\Delta\phi = \Delta f \dfrac{f_o}{2Q}$

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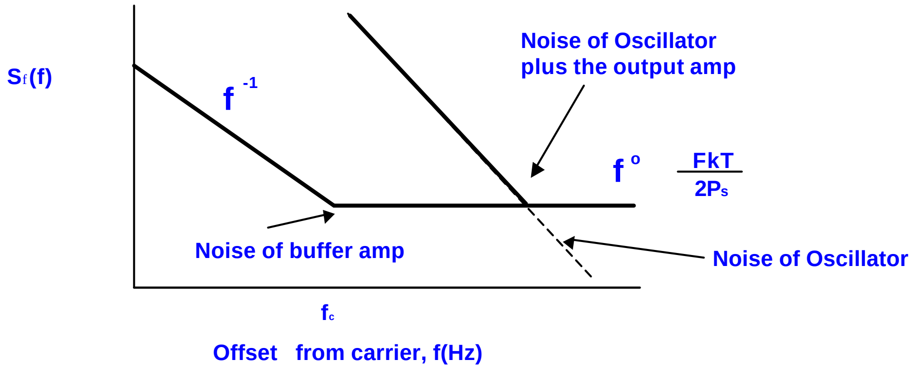
The additive noise of amplifiers, when used in conjunction with resonators to create stable oscillators can appear as higher order noise within the oscillator's noise characteristics (close-to-carrier) as well as the broadband far-from-carrier noise characteristics.

Additive Noise for Two-port Devices



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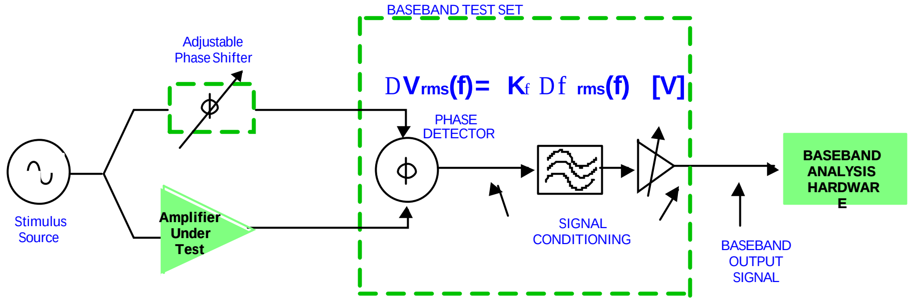
The noise of a buffer amplifier adds to the noise of an oscillator by limiting the far-from-carrier noise characteristics of the oscillator/buffer amplifier combination.

Additive Noise Measurements



- Additive PM noise can be less than the PM noise from device or stimulus source
- System noise floor same as PLL method
- Time delay of both paths ideally need to be equal (but in quadrature)
- Additive PM noise can be less than AM noise from device or stimulus source

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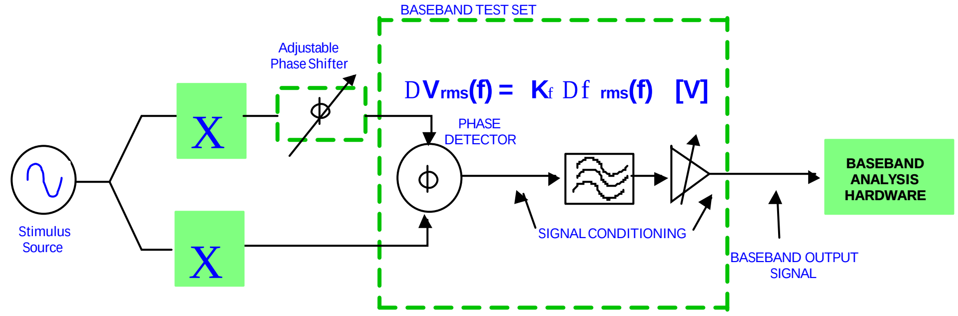
Two-port phase noise characterization requires a stimulus source to provide the input signal reference. The goal of the measurement is to have the noise of the source common to both signal paths and arrive correlated at the phase detector. This requires the two signal paths lengths to be as equal as possible. The phase shifter is used to establish quadrature of the two signals at the phase detector.

With the noise of the source is correlated at the phase detector, the remaining phase noise difference is the noise of the two-port device.

Additive Noise Measurements
Frequency Conversion



- Additive PM noise can be less than the PM noise from device or stimulus source
- System noise floor same as PLL method
- Time delay of both paths ideally need to be equal (but in quadrature)
- Additive PM noise can be less than AM noise from device or stimulus source
- Frequencies at the phase detector must be equal

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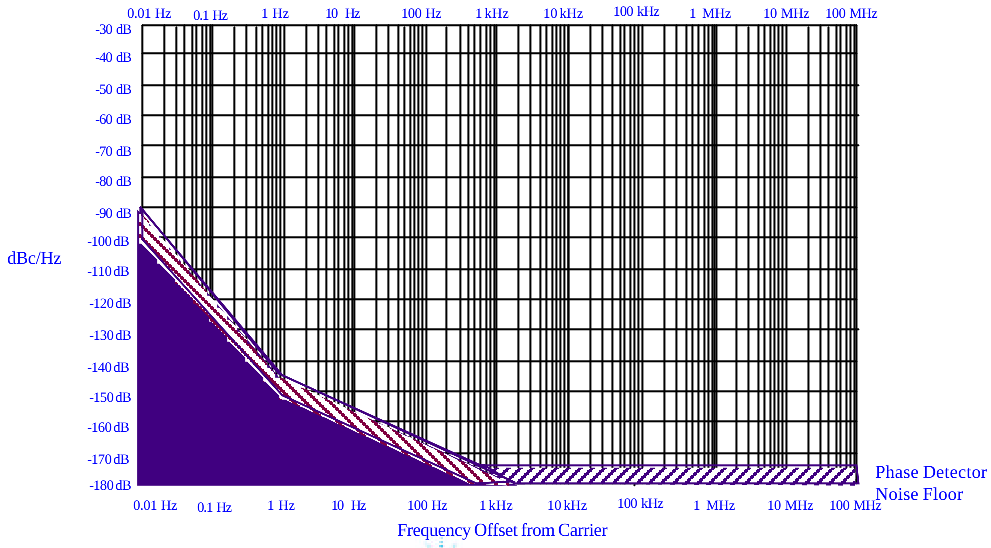
If the output frequency of the UUT does not equal the input frequency, then it is not possible to measure the two-port noise of a single device. When a like device (with similar or less noise) is placed in the reference path, then the noise of both devices can be measured. As in the PLL method for oscillators, the sum of the noise of the two devices is measured, therefore both devices are at least as quiet as the measurement result everywhere. At any particular offset, one device is at least 3 dB quieter than the measurement result.

Additive Noise Measurements

Noise Floor = $N_{(sys)}$



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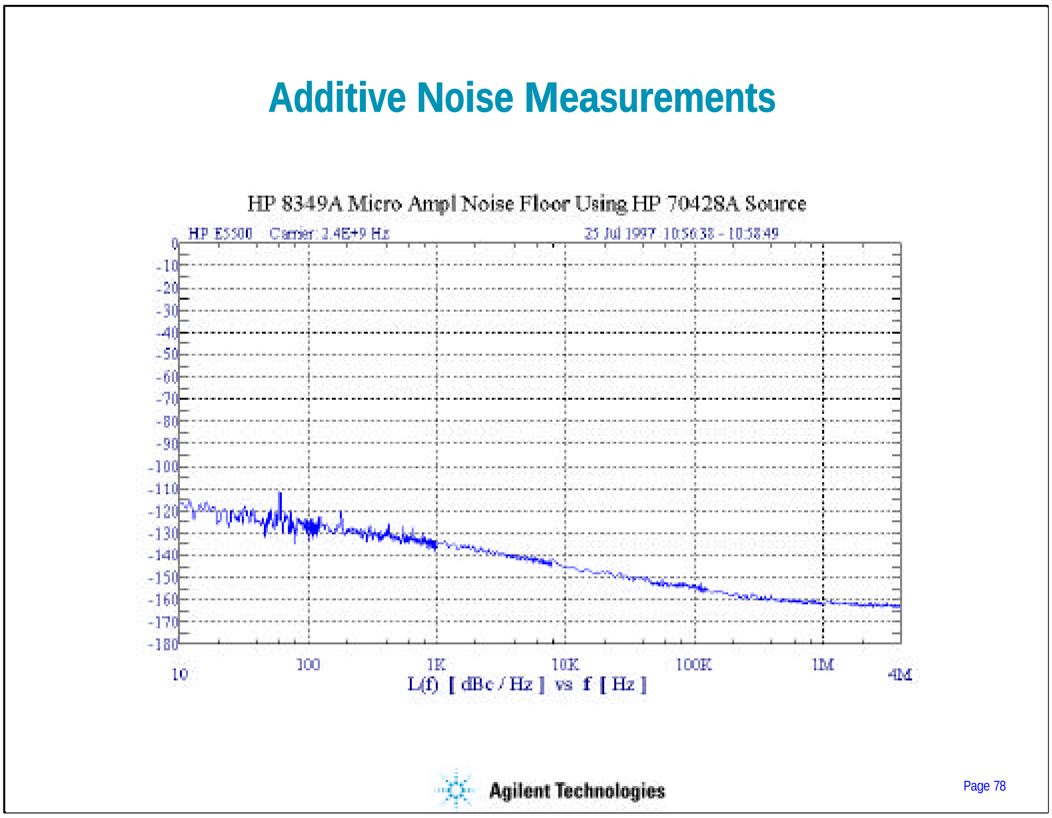
The overall noise floor of this measurement approach is the combination of the noise of the system (phase detector, LNAs, baseband analyzers)

The rise in noise for close-to-carrier offset frequencies is due to the 1/f noise of the phase detector and the LNA's.



Here is a typical CW residual measurement of a microwave amplifier at a carrier frequency of 2.4 GHz.

Measurement Accuracy Issues

- measurement of K_ϕ
- measurement accuracy of baseband analyzer
- quadrature maintenance
- frequency response of test set
- high frequency mismatch/flatness (> 1 MHz)

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These are the major factors which can cause measurement inaccuracy. A typical uncertainty budget of +/- 2 dB for offsets < 1 MHz and +/- 4 dB for offsets < 100 MHz comprise an RSS summation of these factors.

For offset frequencies < 1 MHz, the error budget is based on an assumed +/- 1 dB uncertainty for the measurement of Kphi and +/- 1 dB for the RSS combination of the uncertainty of baseband analyzer, quadrature maintenance, and the overall frequency response of the baseband test set.

For offset frequencies > 1 MHz, the additional +/- 2 dB of uncertainty is attributed to the high frequency mismatch uncertainties and amplitude flatness of the hardware configuration.

Measurement Techniques

Noise Measurement of Oscillators
- Spectrum analyzer technique
- phase detector techniques
    - discriminator method
    - PLL / reference source method

Additive Noise measurement (two-port devices)

☞AM Noise measurements

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The last measurement technique to discuss is the one for AM noise measurements.

AM Noise Measurement Technique

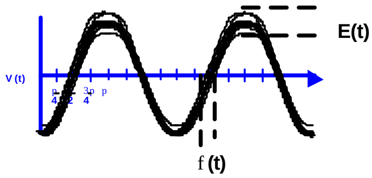
REAL WORLD SIGNAL

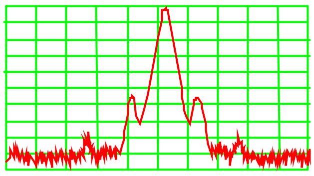
$V(t) = [A_0 + E(t)] \sin[2\pi f_0 t + \phi(t)]$

Where

$E(t)$ = random amplitude fluctuations

$\phi(t)$ = random phase fluctuations





- AM-PM Conversion
- AM noise may degrade phase noise floor

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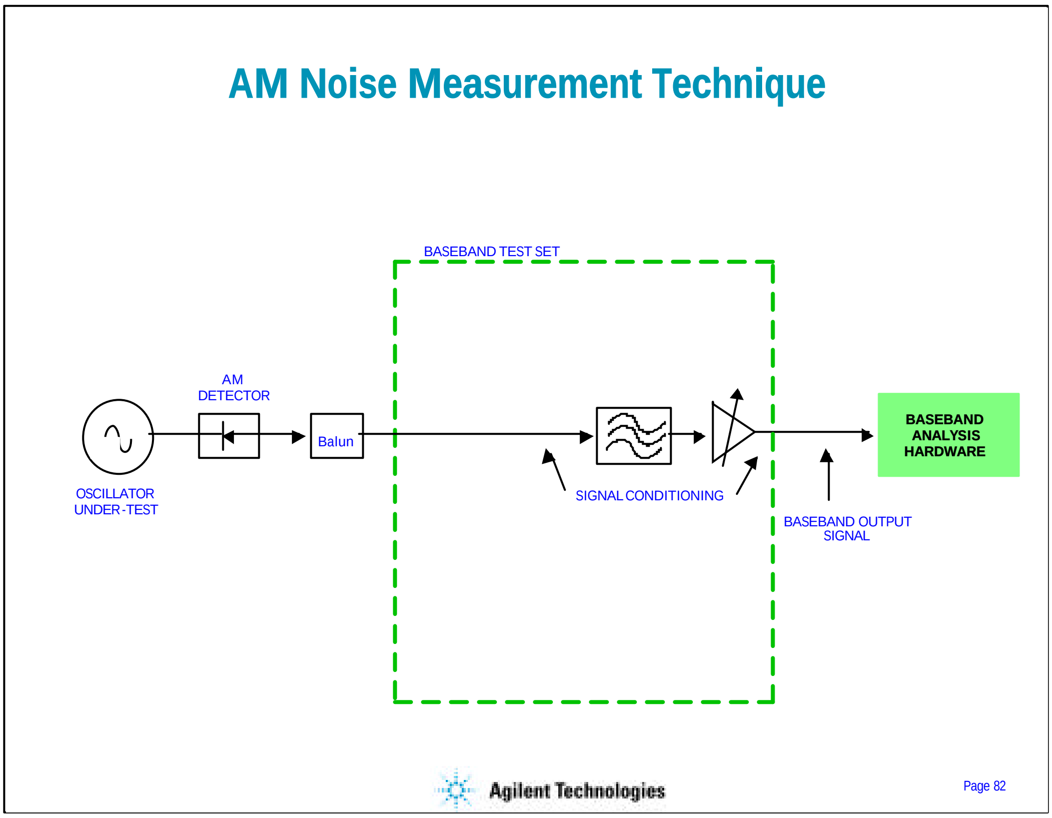
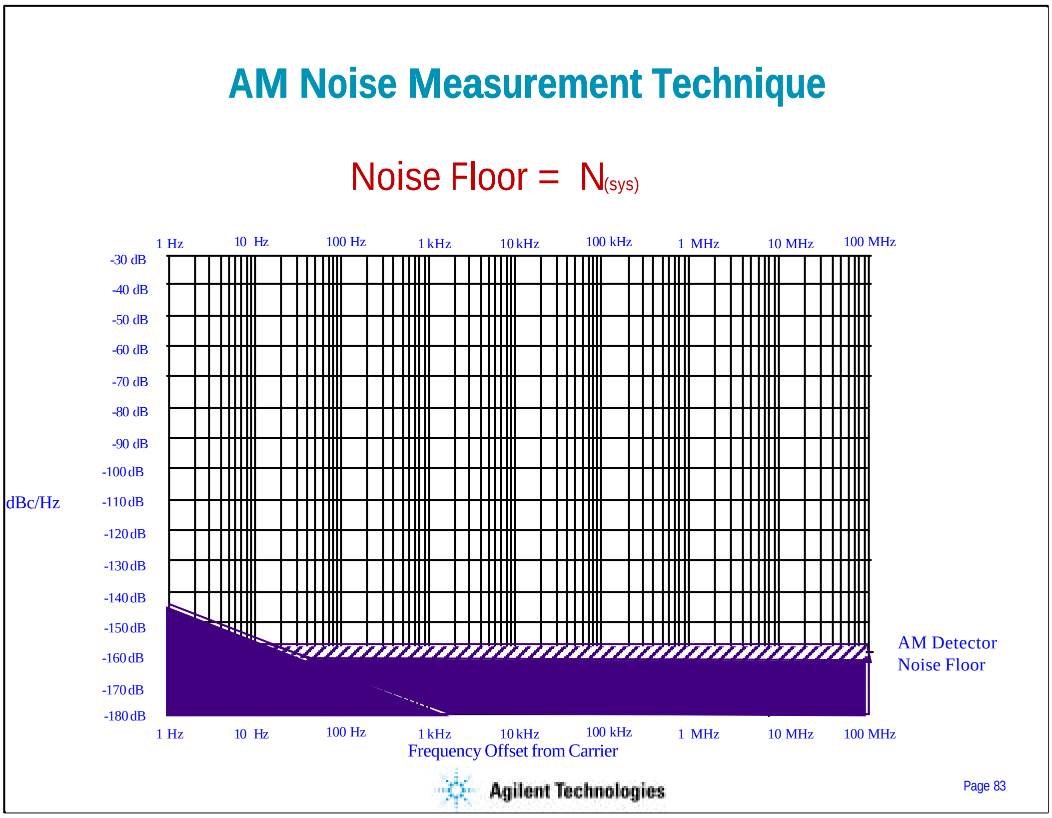
Real world signals have both phase noise characteristics and amplitude noise characteristics. Many amplifiers have AM-PM conversion sensitivity characteristics and the presence of amplitude noise at the input can be converted to phase noise at the output of the amplifier.

In addition to the AM-PM conversion issues, AM noise can cause inaccurate phase noise measurements in some measurements techniques.

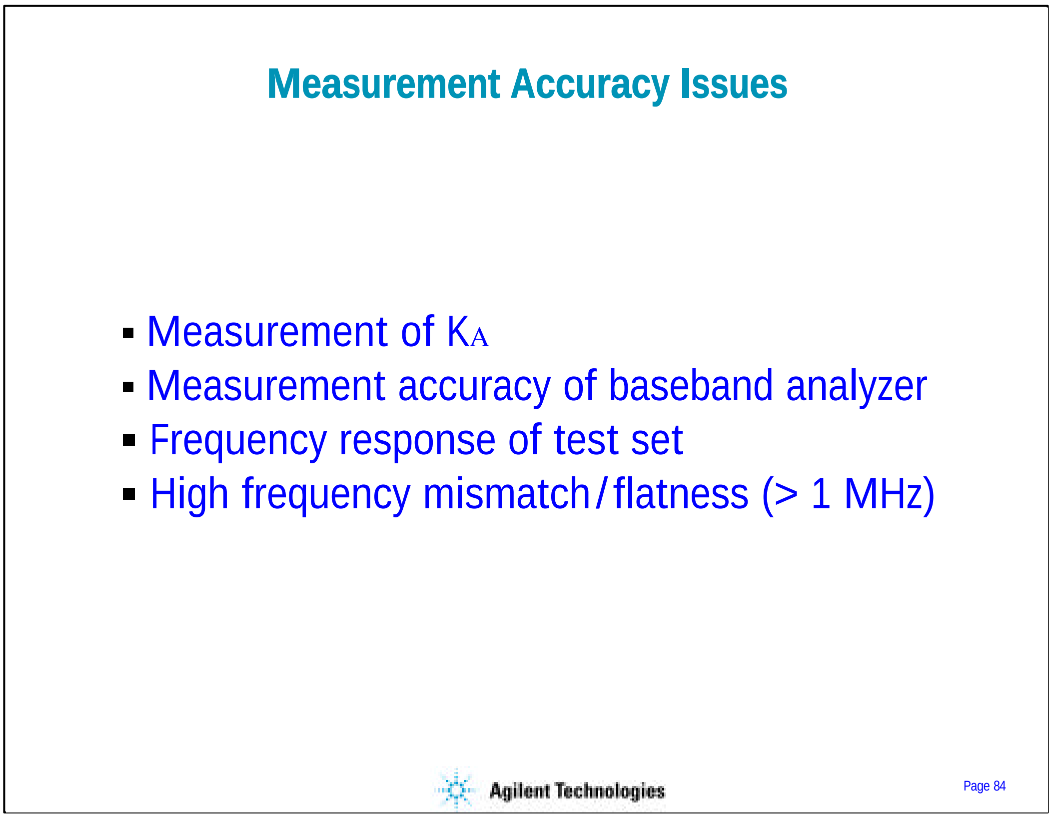


The measurement technique for AM noise uses a diode detector (AM detector) to convert amplitude noise fluctuations to voltage fluctuations which are measured by the baseband analyzer.

A balun is highly recommended when using an external AM detector to eliminate ground loop spurious signals which can be measured by the system but are not part of the signal from the UUT.



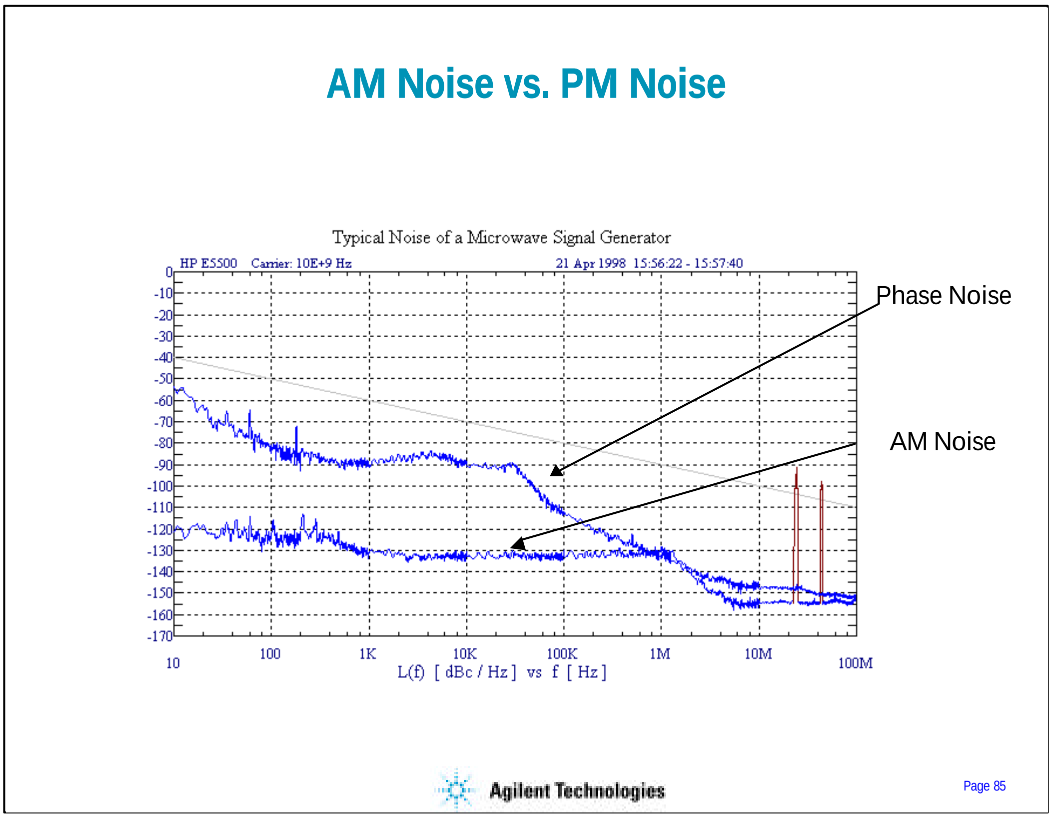
The noise floor for AM measurements is a function of the carrier power and the sensitivity of the AM detector. Most diode detectors have typical sensitivities in the -150 to -160 dBc/Hz range. The close-to-carrier frequency limit is due to the blocking capacitor used for the AM detector.



These are the major factors which can cause measurement inaccuracy. A typical uncertainty budget of +/- 2 dB for offsets < 1 MHz and +/- 4 dB for offsets < 100 MHz comprise an RSS summation of these factors.

For offset frequencies < 1 MHz, the error budget is based on an assumed +/- 1 dB uncertainty for the measurement of K_A and +/- 1 dB for the RSS combination of the uncertainty of the baseband analyzer and the overall frequency response of the baseband test set.

For offset frequencies > 1 MHz, the additional +/- 2 dB of uncertainty is attributed to the high frequency mismatch uncertainties and amplitude flatness of the hardware configuration.



Shown here is a typical example of an AM noise measurement of a microwave synthesizer as compared to the phase noise measurement of that synthesizer. Notice that at some offset frequencies, the AM noise of the source is equal to the phase noise of the source.

Agenda

- Review Basic Phase Noise Concepts
- Define Measurement Techniques
- ● Select an Appropriate Measurement Technique
- Understand Measurement Trade-offs
- Survey of Measurement Solutions Available

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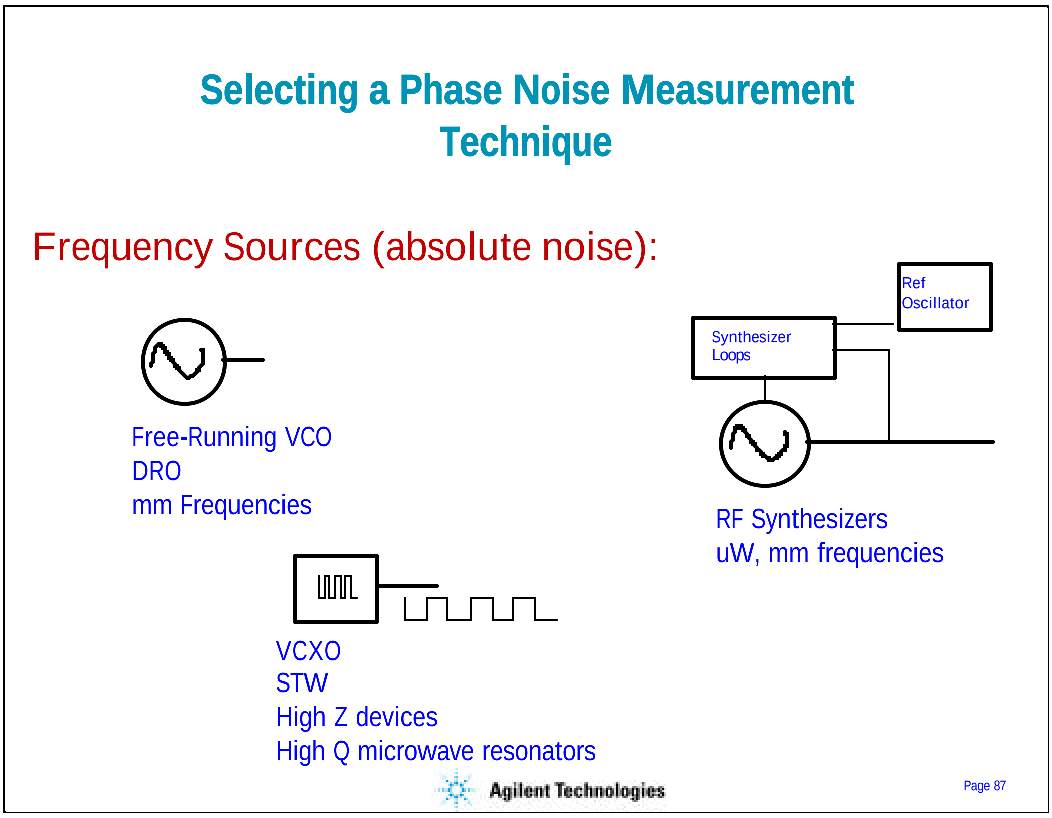
Once the measurement techniques are understood, the next step is to pick a technique which is appropriate for specific devices. This section will focus on selecting a measurement technique for the different types of oscillators or sources which are typically measured.

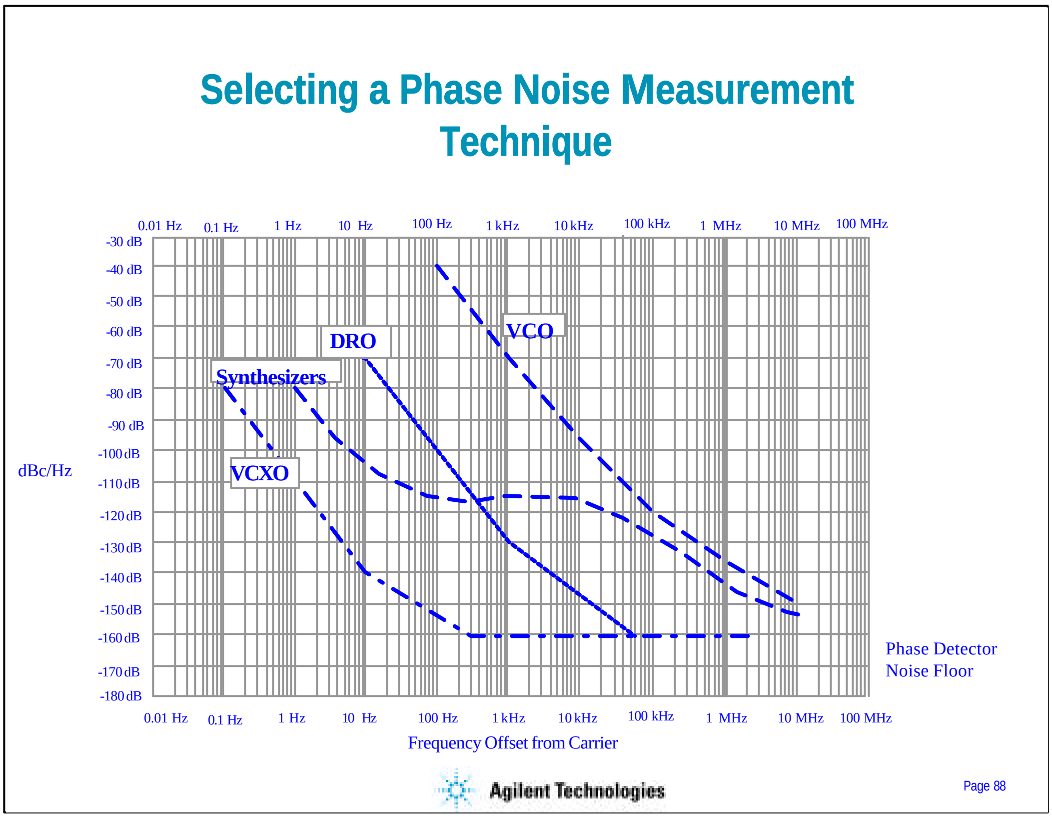


Oscillators or frequency sources can be placed into three general categories:

1) free running oscillators (VCO's, DRO's, YIG)

2) synthesizers (combines free running oscillators with low noise reference oscillators and synthesis pll's)

3) low noise to ultra low noise devices (VCXO, STW, SAW, High Q)



As seen from this graph, these devices vary significantly in their phase noise characteristics.

Selecting a Phase Noise Measurement Technique

	PLL	PLL FM Discriminator Direct Spectrum
mm	Synthesizers	VCO DRO Gunn
uWave	Synthesizers	VCO DRO Gunn
RF	Synthesizers High Q resonators	VCO DRO Synthesizers
IF	VCXO TCXO High Q resonators	
	Close-to-carrier	Far-from-carrier

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This table approximates the noise characteristics of the different types of oscillators as a function of close-to-carrier Vs far-from-carrier offset frequencies and places them into categories which are a function of carrier frequency range.

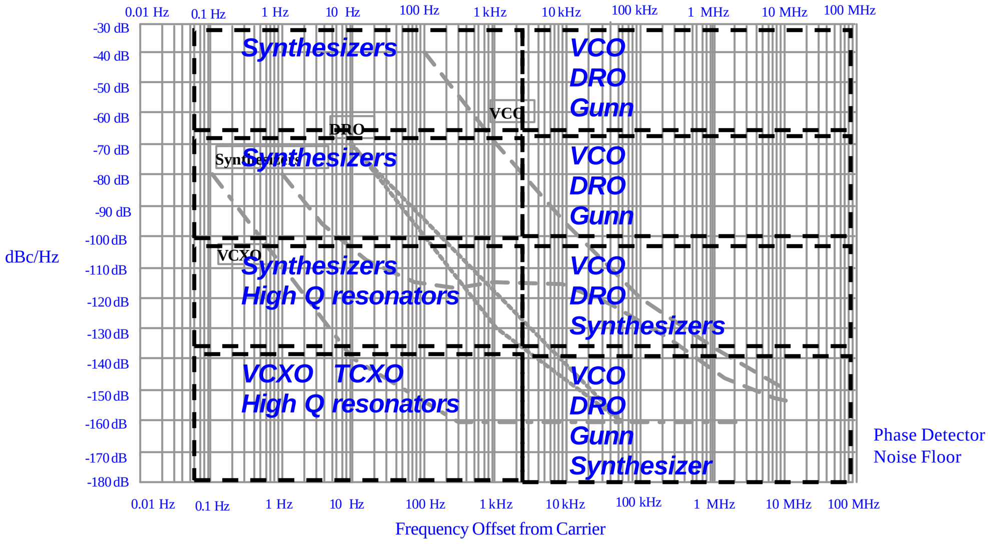
Low frequency devices such as VCXO's and TCXO's tend to be the lowest noise devices close-to-carrier.

Conversely, free running oscillators tend to be the highest noise devices close to carrier.

If you need to make close-to-carrier measurements that can be generalized as low noise, you should be using the PLL/reference source technique.

If your interest in phase noise information is mostly far-from-carrier information, then all three of the measurement techniques can be used, depending upon the required measurement noise floor.

Selecting a Phase Noise Measurement Technique



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High Q Resonators

Issues:
- Ultra low noise close-to-carrier
- RF, uWave, mm frequencies
- No voltage control

Solution:
- Comparison Oscillator of like performance
- Implement a specific "PLL interpolation oscillator"

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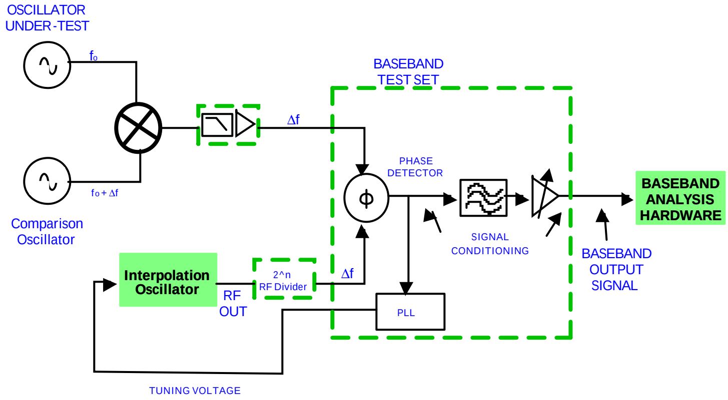
High Q oscillators such as VCXO's and TCXO's tend to have ultra low phase noise close-to-carrier. But these devices are of low frequency and have some amount of voltage control allowing them to measured with the PLL/reference source technique.

There are high Q oscillators which have ultra low noise characteristics, carrier frequencies in the microwave band, and have no voltage control. These devices require a comparison oscillator of like performance and a specific implementation of a reference PLL interpolation oscillator.

High Q, High Frequency PLL Measurement Technique



- Measures Noise of both high frequency oscillators
- Δf defines maximum offset frequency
- RF divider provides low noise floor
- Interpolation oscillator provides "fine" frequency resolution

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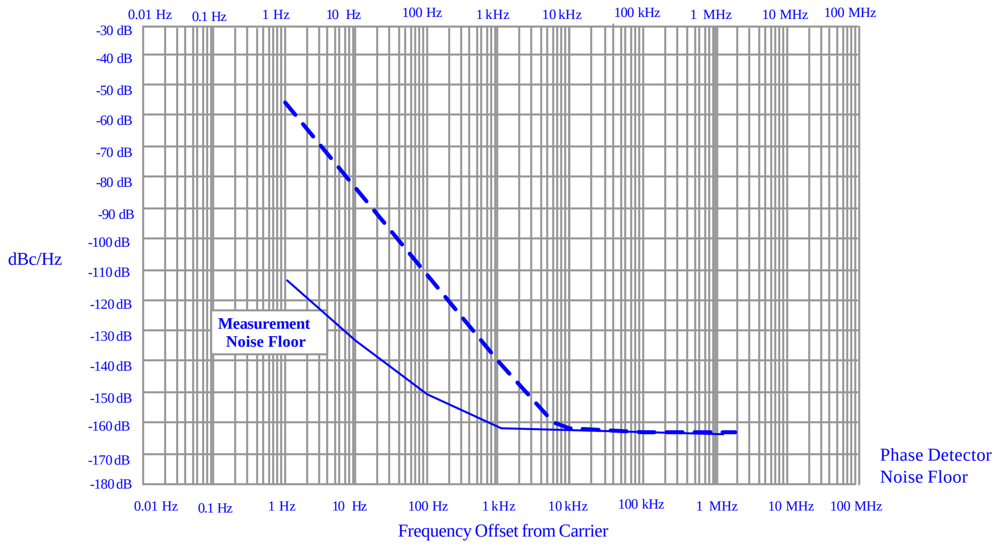
The comparison oscillator - which is of like performance compared to the UUT - has a carrier frequency which is different from the UUT carrier frequency by approximately 10 MHz.

The UUT and the comparison oscillator signals are mixed down to the difference frequency. This IF frequency can be filtered and amplified if necessary before routing it to the phase detector.

The interpolation oscillator consists of a standard RF signal generator whose output feeds a low noise RF divider and has EFC tuning control for the PLL. The RF output of the signal generator is set to a low RF value such as 320 MHz and the divider is set to a divide-by-32 value producing a "tunable" 10 MHz signal for the other port of the phase detector.
(EFC tuning control allows an external voltage signal to be applied to the internal low noise reference of the signal generator. The Agilent 8662A signal generators are examples of generators with effective EFC control.)

Microwave, High Q Oscillator Results



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The measurement noise floor for this approach is shown here in comparison to the low noise of the UUT. The RF divider provides an approximate noise floor far-from-carrier of about -165 dBc/Hz. The close-to-carrier noise floor is the RF signal generators phase noise reduced by 30 dB.

The advantage of using a signal generator is the flexibility to set the interpolation frequency to exactly the difference frequency between the UUT and the comparison oscillator.

mm Source Frequency Measurement Techniques

Issues:
- Very high carrier frequencies (waveguide vs coax)
- No standard mm downconverters

Solution:
- Use "harmonic mixers" to downconvert mm signal
- Useful for PLL or FM discriminator technique

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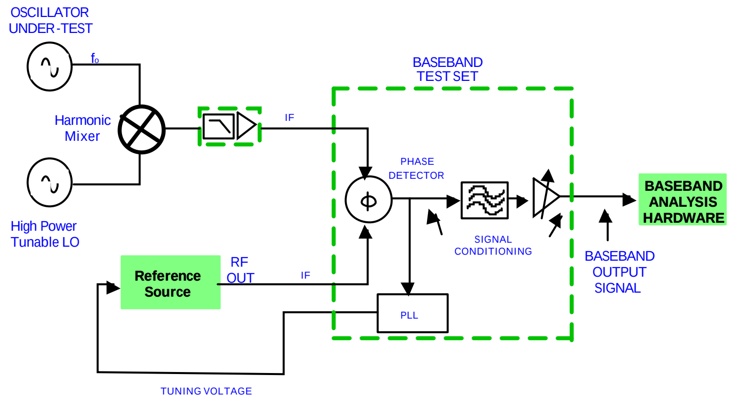
The measurement of frequencies greater than 26 GHz can be difficult. Standard mm frequency downconverters do not exist and most of the mm range is split into waveguide bands.

The solution to this difficult measurement is to use standard waveguide harmonic mixers to translate the mm frequency to a lower RF/IF frequency. This approach works well for either the PLL/reference source technique or the discriminator technique.

mm Frequency
PLL Measurement Technique



- LO Noise sets overall system noise floor
- Harmonic mixer may need "bias" for odd harmonics
- IF gain necessary due to low power IF signal from harmonic mixer

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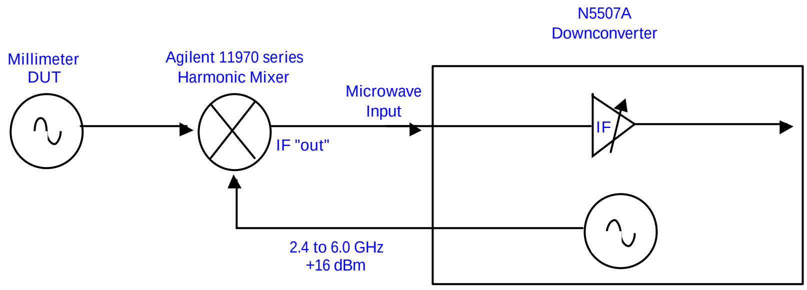
Shown here is the generalized approach. The tunable LO drives the harmonic mixer to produce a harmonic frequency which is close to the mm frequency of the UUT. The IF frequency must be amplified before it is delivered to the signal input of the phase detector. Not shown here is the bias circuitry necessary to allow the harmonic mixer to use either "odd" or "even" harmonics to produce a frequency close to the mm carrier frequency.

The reference source can be a typical RF signal generator.

mm Frequency
PLL Measurement Technique



- High power LO internal to the 26 GHz downconverter
- Harmonic mixer "bias" internal to 26 GHz downconverter for odd harmonics
- IF amplifier internal to the 26 GHz downconverter
- Downconverter firmware "auto-selects" odd or even harmonic and LO frequency

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The Agilent N5507A microwave downconverter simplifies the mm measurement. The internal LO is tunable in 600 MHz steps and delivers up to +16 dBm to the LO port of the harmonic mixer. The IF signal is routed to the input of the downconverter and directly to the 45 dB variable gain IF amplifier before being routed to the phase detector input. Not shown here is the internal bias circuitry which automatically biases the harmonic mixer when "odd" harmonics are necessary.

High Output Impedance Measurement Techniques

Issues:

- Low noise close-to-carrier
- No "standard" references
- High output impedance not compatible with 50 ohm measuring system
- Unique device packaging
- Low output power (dBm)

Solutions:

- Use high impedance, high frequency probe
- Implement "specific" buffer circuit

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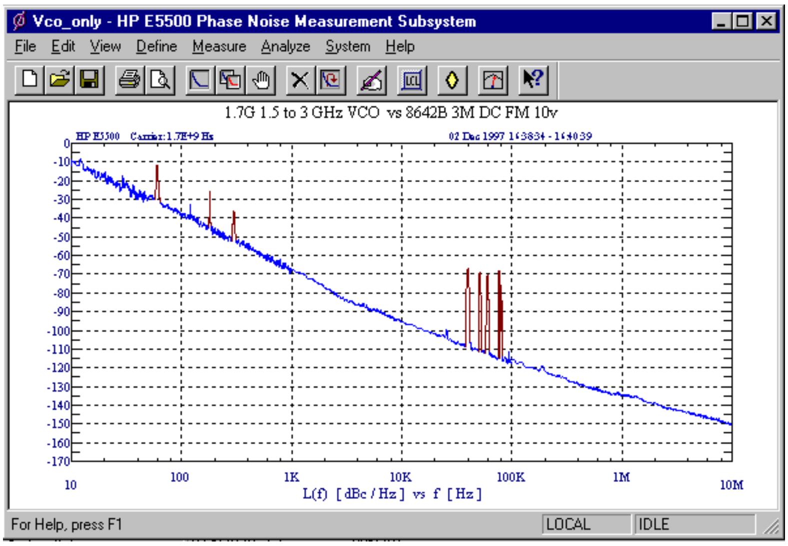
Another example of a device which adds complexity to a measurement is when the UUT has a low power, high impedance output. Generally these devices are of very low noise close-to-carrier where the PLL technique must be employed. The high output impedance of the UUT is not compatible with a 50 ohm measurement system (the device may stop oscillating due to the high current load). Complicating matter even more are the issues of device packaging and low output power.

There are two possible solutions to this:

1) use a high frequency, high input impedance probe

2) implement a specific buffer circuit to translate the high impedance to 50 ohms.

VCO Measurement - 50 Ohm system



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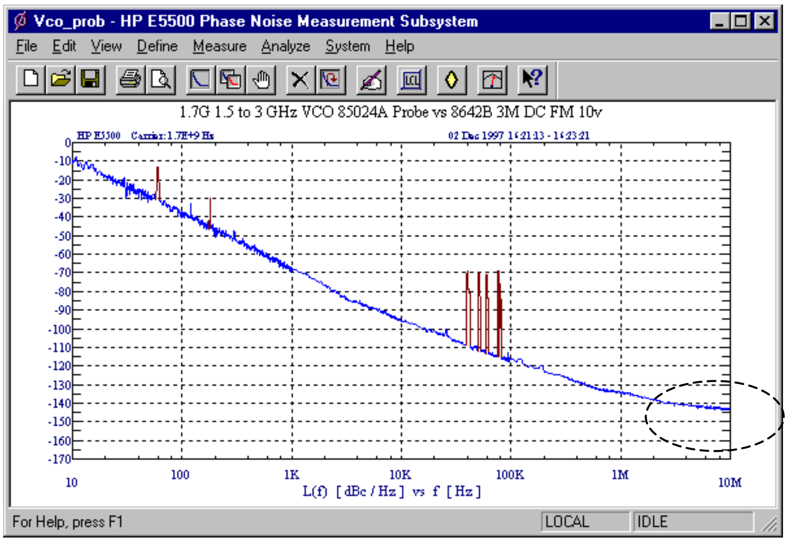
Here is an example of a high frequency VCO measured with a 50 Ohm measurement system.

VCO Measurement - 85024A Probe



Probe additive noise
vs.
VCO Noise

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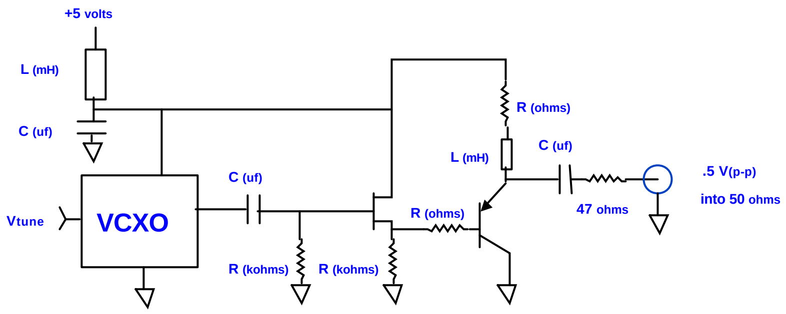
Here is the same device measured by using a high frequency, high input impedance active probe. Notice the difference of noise at the far-from-carrier offset frequencies. This is where the noise of the VCO is approaching the additive noise of the probe.

Simple High-Z Buffer Circuit



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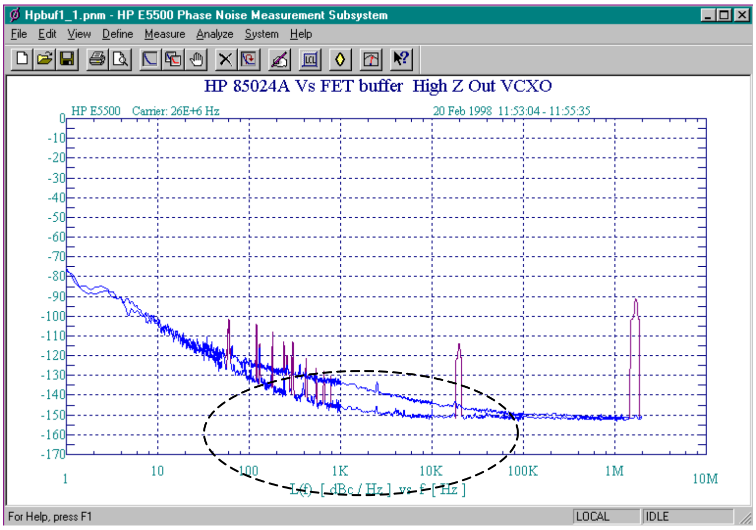
A high output impedance device requires a buffer circuit with a high input impedance. Here is a typical example of a high impedance buffer circuit. The FET is used to keep the loading factor to a minimum while the PNP transistor is used to provide a 50 ohm output to the measurement system. Since each UUT is specific to a package style, this circuit should be implemented locally.

VCXO Measurements

Probe Vs Buffer Circuit



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Measuring this high output impedance device with the two alternative solutions shows a difference in noise performance in the 10 Hz to 100 KHz offset frequency range. This difference is due to the higher additive noise of the active probe for low carrier frequencies. The far-from-carrier noise is limited because of the low output power of the UUT.

Agenda

- Review Basic Phase Noise Concepts
- Define Measurement Techniques
- Select an Appropriate Measurement Technique
- ● Understand Measurement Trade-offs
- Survey of Measurement Solutions Available

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In this next section we will be discussing how to optimize the overall noise measurement.

Optimizing Modern Noise Measurements

- Measurement Speed
  (sacrifice measurement information)

- Measurement Information
  (sacrifice measurement speed)

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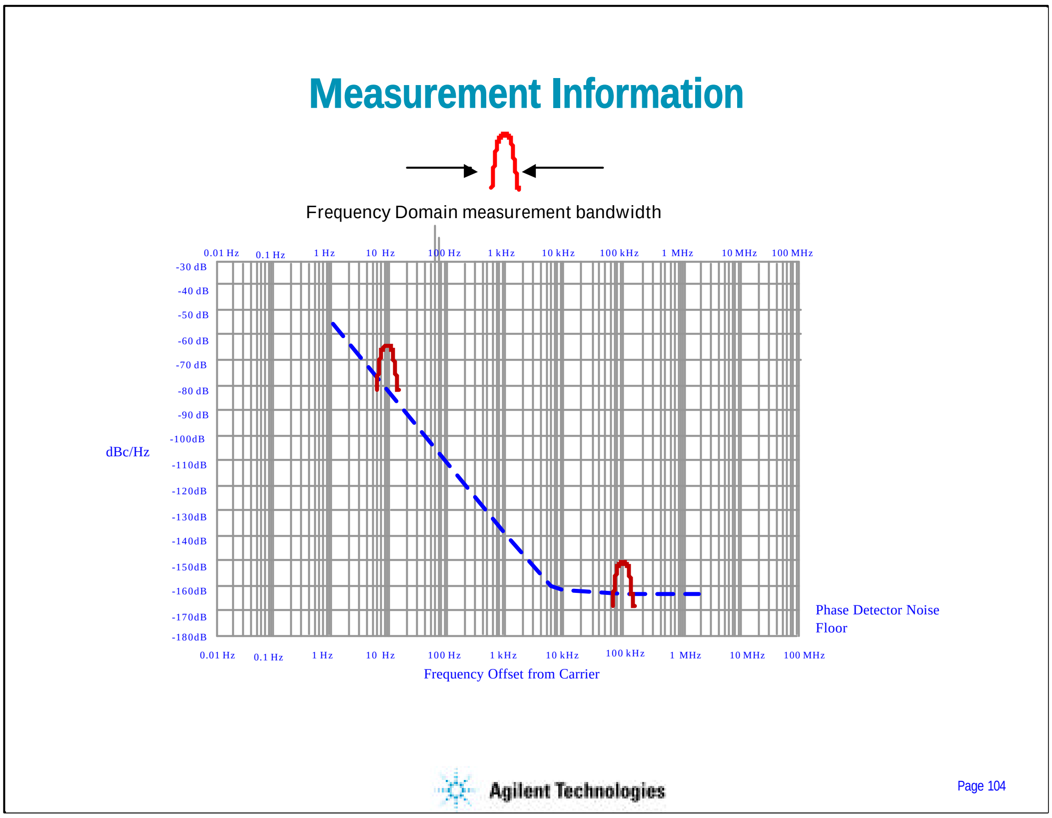
There are many aspects to a phase noise measurement which one could optimize.

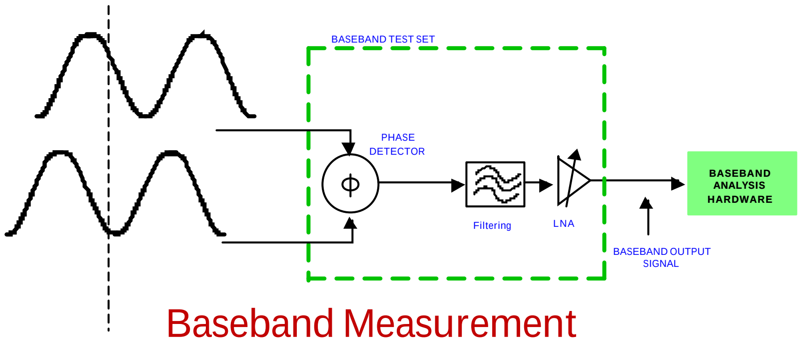
Optimizing to measure as fast as possible has the unfortunate tradeoff of possibly sacrificing measurement information. And conversely, optimizing a measurement for information requires more data, which takes more time, which means the measurement takes longer.

Optimizing a measurement to be very easy to perform usually means trading off the ability to re-configure the measurement hardware to achieve better measurement sensitivity, or to do other measurement types. This is usually because very sensitive measurements tend to be more expensive than less sensitive measurements however, recent introductions of new signal analysers have made significant changes to this.



Phase noise measurement information comes in the form of a spectral distribution of noise data (per unit bandwidth) and spurious frequency data. The measurement of these items are a function of the "effective"frequency resolution bandwidth. For offset frequencies far-from-carrier, the "effective" resolution bandwidth can be much larger than the resolution bandwidth for offset frequencies close-to-carrier. The narrower the bandwidth, the more time required to gather the measured data and conversely the wider the bandwidth, the less time required.

Measurement Information



Baseband Measurement

Digitizer	collects time domain information, must be converted to frequency domain information
FFT analyzer	collects time domain information and converts it to frequency domain information directly
Swept RF signal analyzer	collects frequency domain information directly

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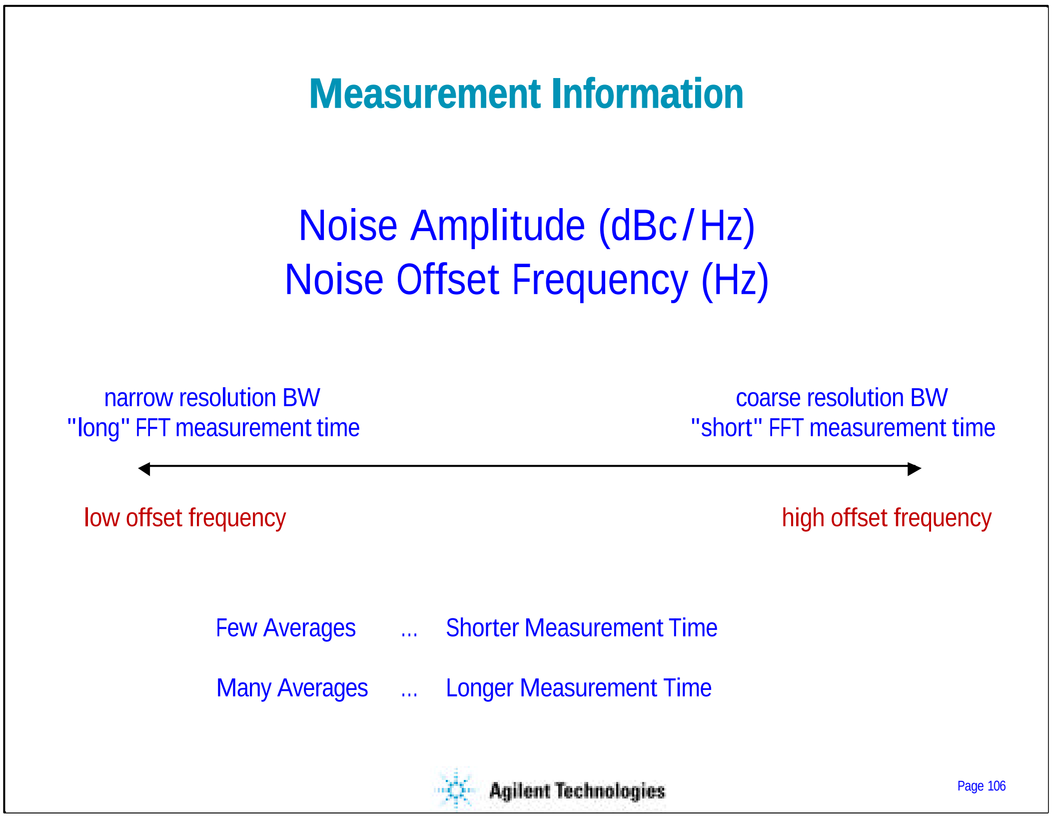
To measure the baseband signal, there are generally three approaches used in modern solutions - using a digitizer card, using an FFT analyzer, and using an RF spectrum analyzer.

A digitizer and FFT analyzers measure time domain information. The information from a digitizer must be converted to frequency domain information using an FFT conversion process. An FFT analyzer collect the time domain information and converts it directly to frequency domain information using its own FFT conversion process. Both of these solutions have an upper offset frequency limit depending on the digitization capability used.

An RF signal analyzer can also measure baseband noise. This is a different measurement than measuring noise through the direct spectrum measurement technique. RF signal analyzers have a minimum input frequency limit.

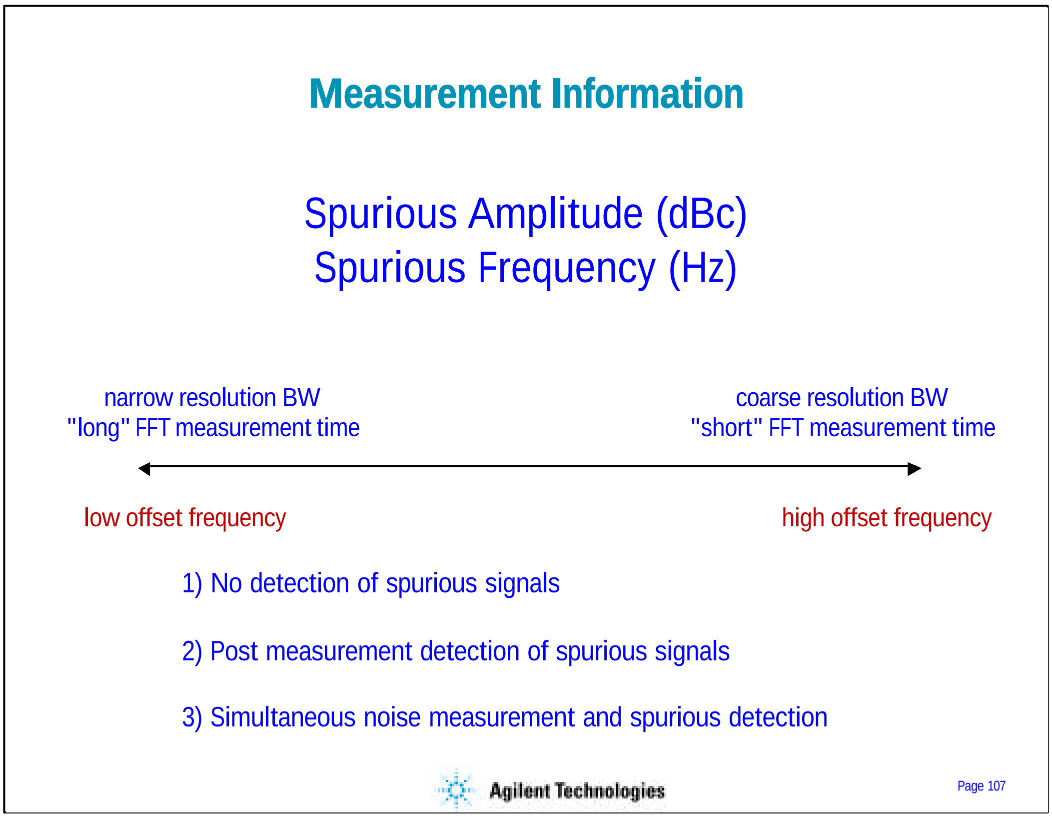


When digitizers and FFT analyzers are used to measure the noise information, the relationship of frequency resolution and time is the underlying issue with "effective" resolution bandwidth and the time required to gather the measured data.

For FFT analysis, narrow frequency resolution requires long time records where Ttotal = 1/fmin (fmin is the minimum frequency resolution required for the information desired).

For swept RF signal analyzers, this main issue is the resolution bandwidth of the signal analyzer.

Averaging of measured data is generally required to reduce the variation of the measured data - especially because we are interested in noise information.



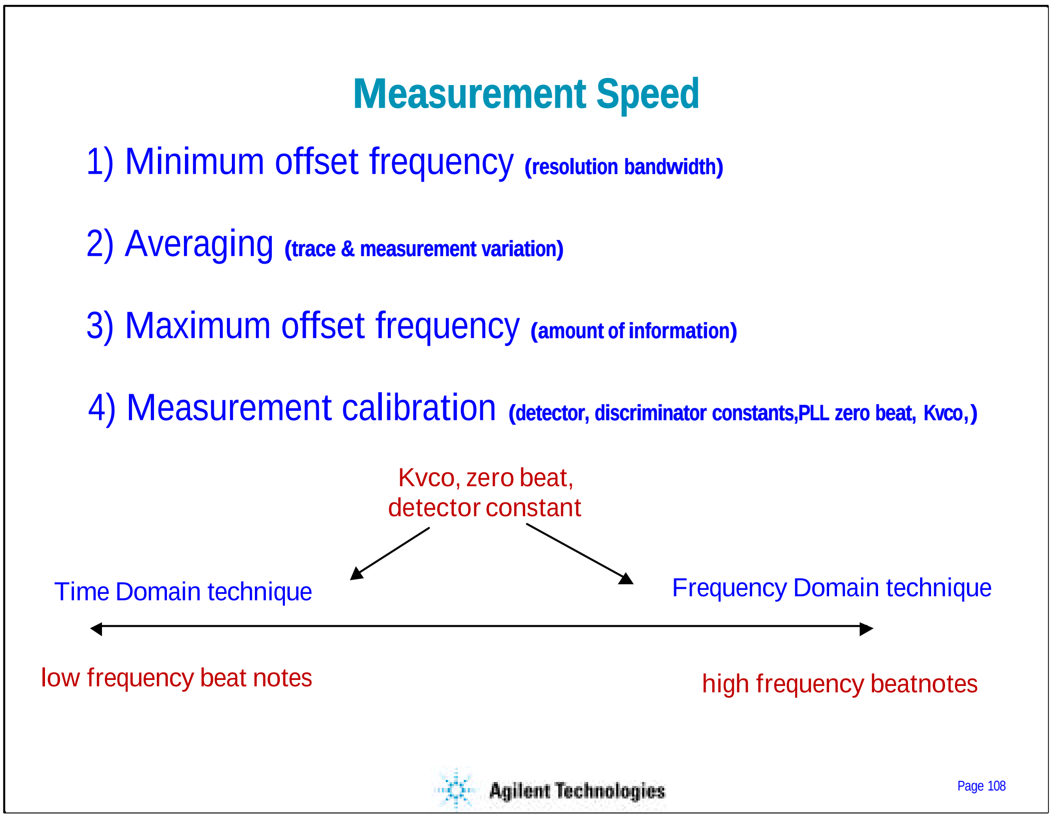
The ability to identify spurious signal - apart from noise - is also a function of measured frequency resolution. Wide "effective" resolution bandwidths allow many spurious signals to go un-noticed while narrow resolution bandwidths can increase the overall measurement time significantly. The magnitude of spurious signals resolved is a function of averaging - low level spurious signals required much more averaging than does high level spurious signals.

Within modern phase noise measurement solutions, there are three approaches used to identify spurious signals:

1) no detection - treat everything as noise

2) treat everything as noise, post process the data to identify spurious only if asked to do so

3) always identify spurious signals and present them at the same time that noise data is presented.

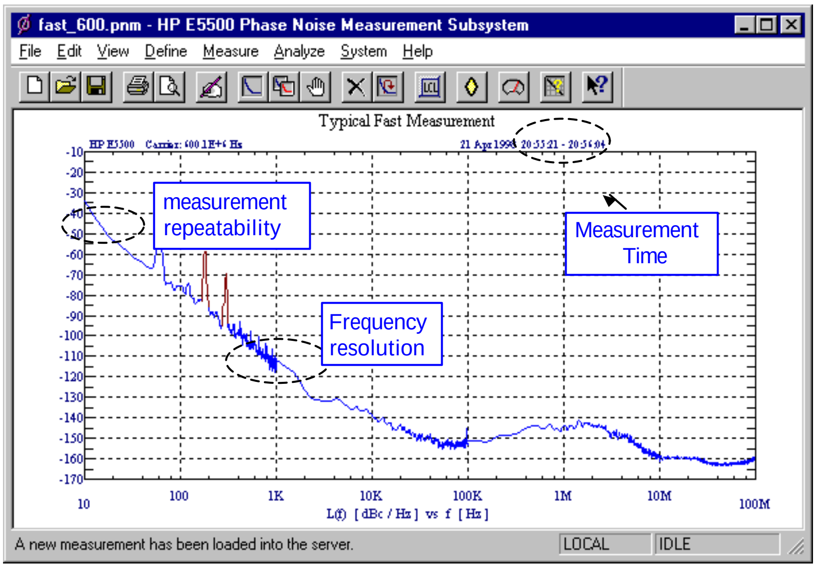


The factor which influences measurement speed the most is the minimum offset frequency selected since this determines the longest time record or the narrowest resolution bandwidth. Averaging is the next highest contributor to long measurement times since it impacts how much the noise trace varies and how much measurement to measurement variation exists. The maximum offset frequency determines the overall quantity of information selected.

Measurement calibration can add time to the overall measurement, especially if the carrier frequencies are low, and the PTRs are very small. For most RF and microwave carrier frequencies, the measurement calibration time is small compared to the total measurement time. For crystal oscillators, the measurement calibration time can exceed the actual measurement time - even when faster techniques are used to collect the measurement calibration information.

Measurement Speed

"fast" measurements



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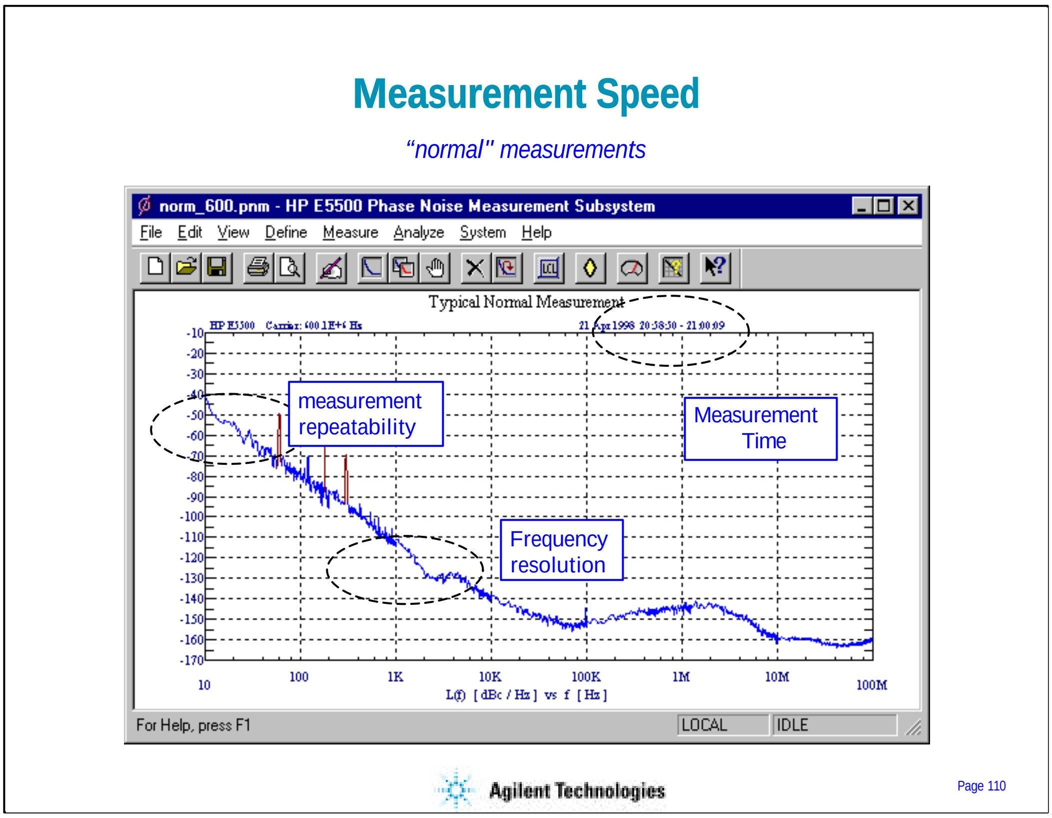
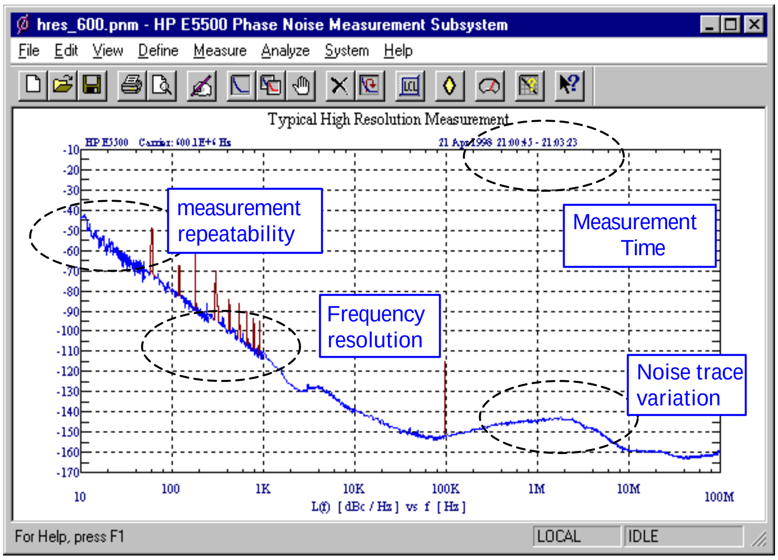
Fast measurements provide a quick look at the noise characteristics of the UUT. These measurements use minimal averaging and wide resolution bandwidths. The variation of the noise trace can be 5-10 dB and the measurement to measurement variation can be 10-15 dB. These attributes are especially noticeable at the measurement segment transition boundaries.



Normal measurements provide better frequency resolution and more averaging than a fast measurement. Measurement repeatability and trace variation is reduced to less than 3 dB typically at the price of longer measurement times.

Measurement Speed

"High Resolution" measurements



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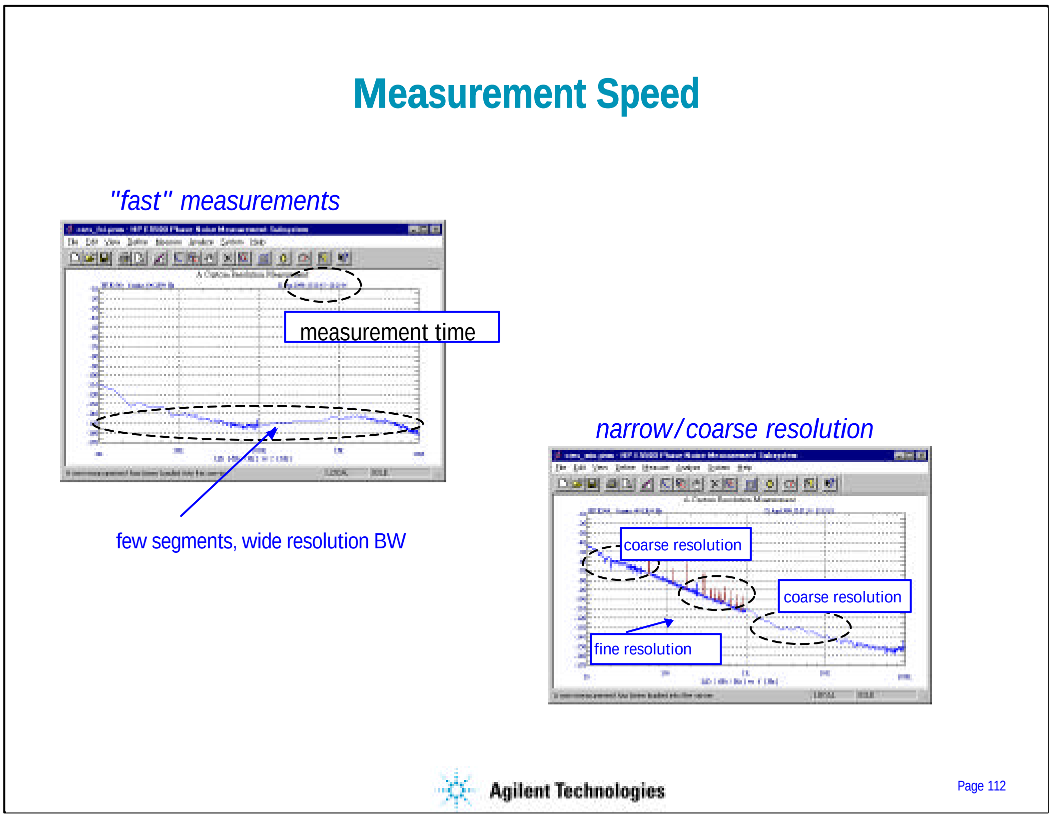
High resolution measurements provide the best resolution of spurious signals and more averaging for typical standard measurements. Measurement time increases significantly while measurement repeatability is typically much less than 1 dB and trace variation is small.



Custom measurements should allow the user to make specific trade-offs of resolution and averaging as required by the information that is needed. This type of measurement may seek to reduce measurement time or to increase frequency resolution in specific offset frequency ranges.

Agenda

- Review Basic Phase Noise Concepts
- Define Measurement Techniques
- Select an Appropriate Measurement Technique
- Understand Measurement Trade-offs
- ● **Survey of Measurement Solutions Available**

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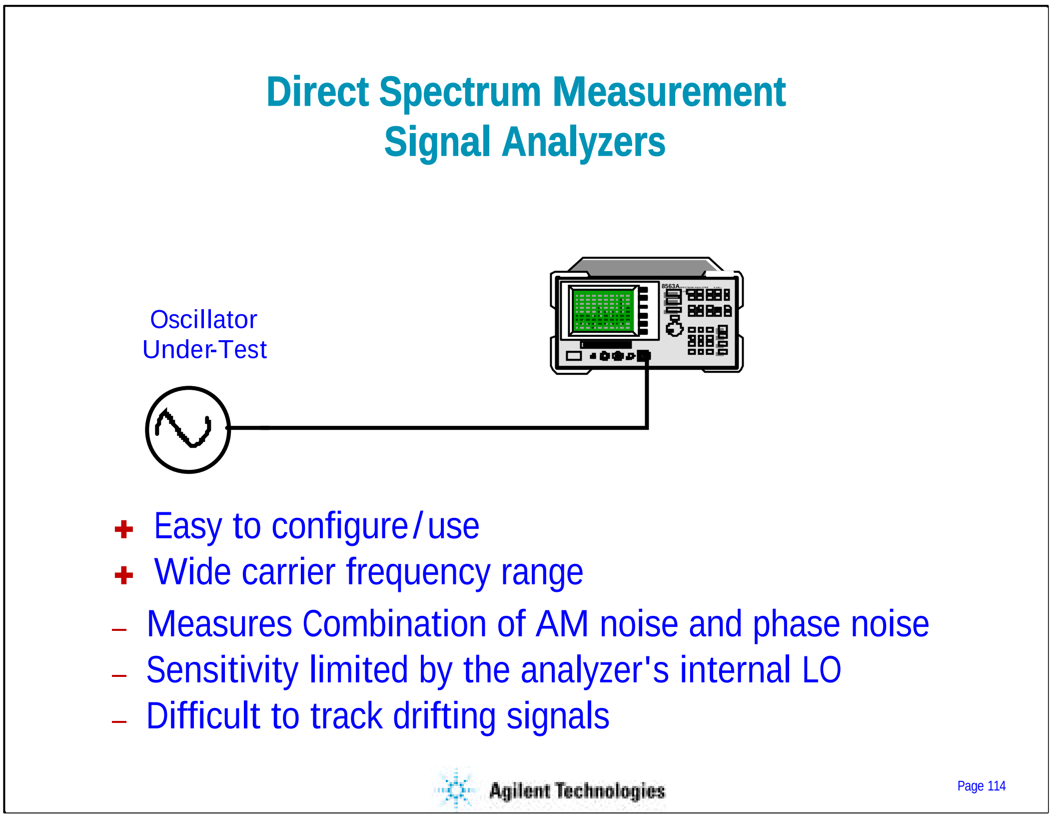
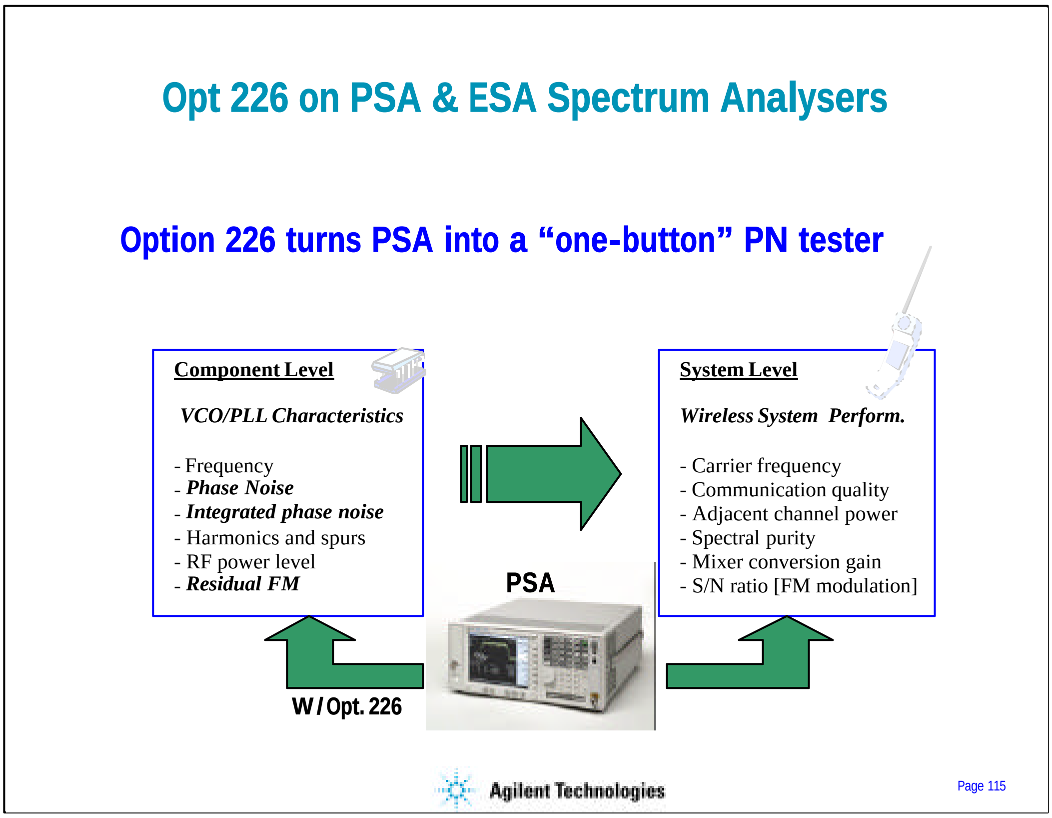
This last section introduces the various phase noise measurement solutions available from Agilent Technologies.



While easy to use for a wide range of carrier frequencies, the spectrum analyzer can only measure the combination of absolute AM noise and phase noise with the overall sensitivity being limited by the noise of the internal LO. Other measurement types such as residual noise and AM noise are not possible.


Opt 226 on PSA & ESA Spectrum Analysers
Option 226 turns PSA into a "one-button" PN tester
Component Level
VCO/PLL Characteristics
- Frequency
- Phase Noise
- Integrated phase noise
- Harmonics and spurs
- RF power level
- Residual FM
System Level
Wireless System Perform.
- Carrier frequency
- Communication quality
- Adjacent channel power
- Spectral purity
- Mixer conversion gain
- S/N ratio [FM modulation]
PSA
W / Opt. 226
Agilent Technologies
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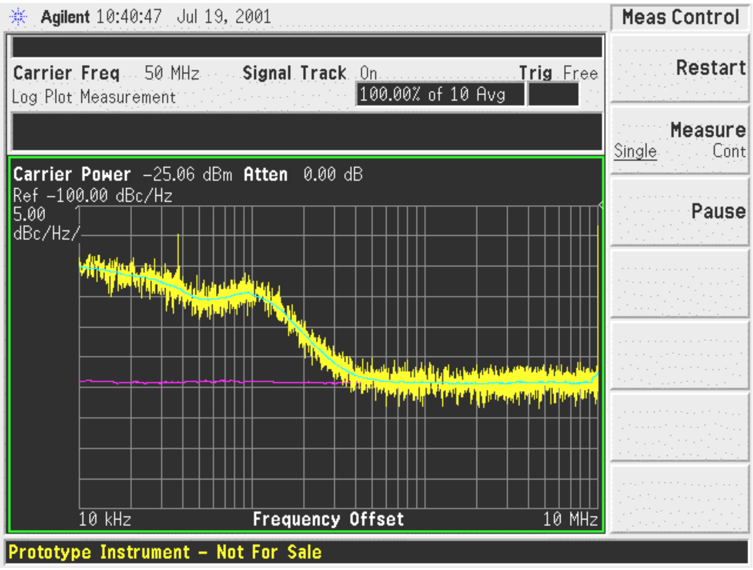
Opt 226 on PSA & ESA Spectrum Analysers



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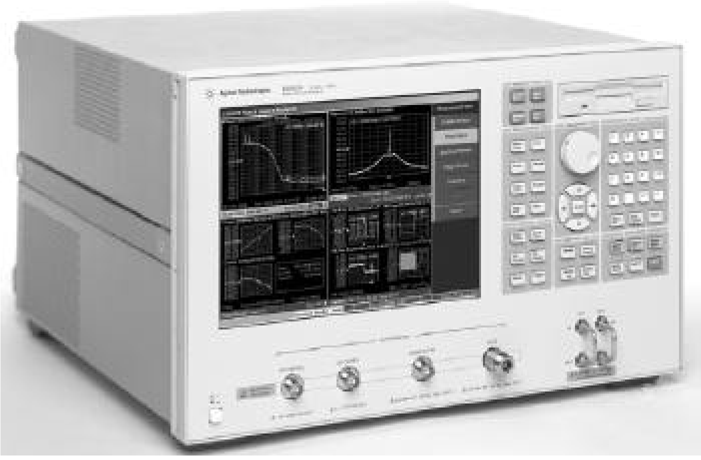
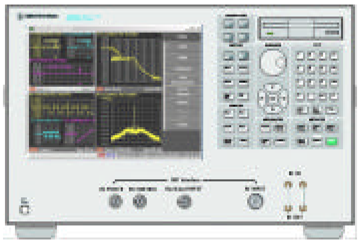
E5052A Signal Source Analyser



E5052A Signal Source Analyser

- One-box Analyzer for Signal Source Test
- World's Best Class Performance of
  - Phase Noise Measurement
  - Transient Measurement
  - Built-in Ultra Low Noise DC Sources



- Ease of Use
  - Multiple Window Operation
  - PC Connectivity and Automation

... Dramatically Improves Your Design and Test Productivity

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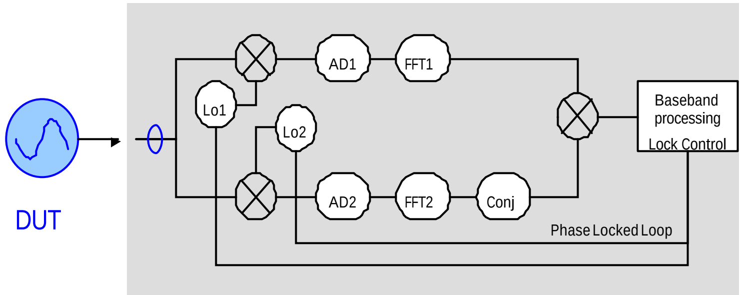
Signal Source Analyser – Key Technology

- *Why Low Phase Noise without External Clean Source ?*
- "Cross-Correlation Method " - The Signal Source Analyzer has two independent LOs, and cancels the phase noise of those using a cross-correlation method. With this technique, it is possible to measure lower phase noise beyond the limitation of LOs phase noise and other noise processes in the independent path of the system.
- *Why So fast Phase Noise/Transient Measurement ?*
- "Use of 100MSa/s A/DC with Hardware Stepped FFT"



Signal Source Analyzer – E5052A

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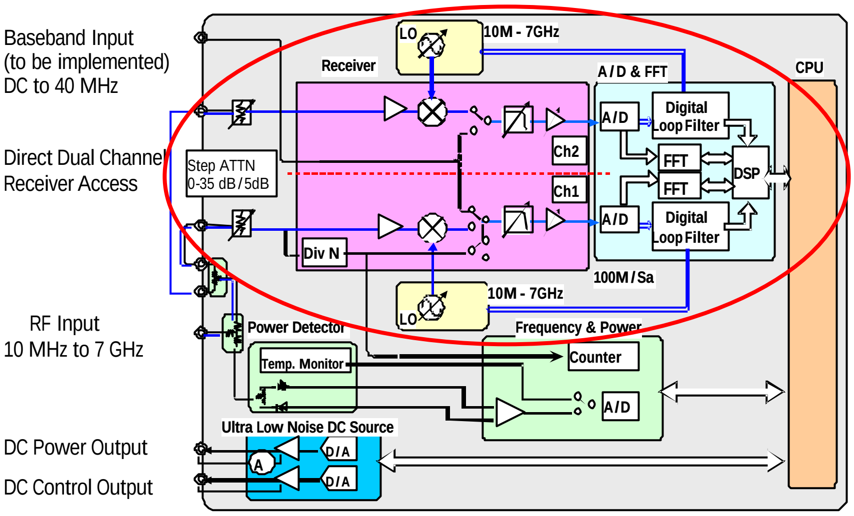
SSA's Block Diagrams and Signal Paths



January, 2004

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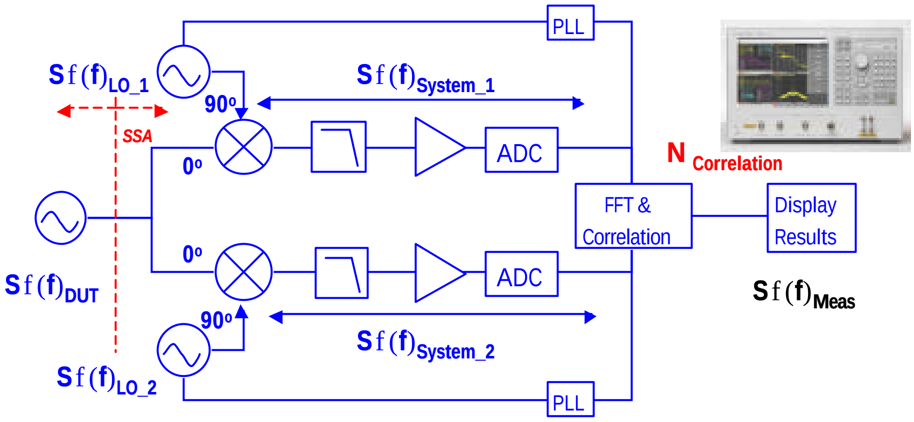
SSA's Two-Channel Cross-Correlation Technique



$$Sf(f)_{Meas} = Sf(f)_{DUT} + \frac{Sf(f)_{LO\_1} + Sf(f)_{LO\_2} + Sf(f)_{System\_1} + Sf(f)_{System\_2}}{\sqrt{N_{Correlation}}}$$

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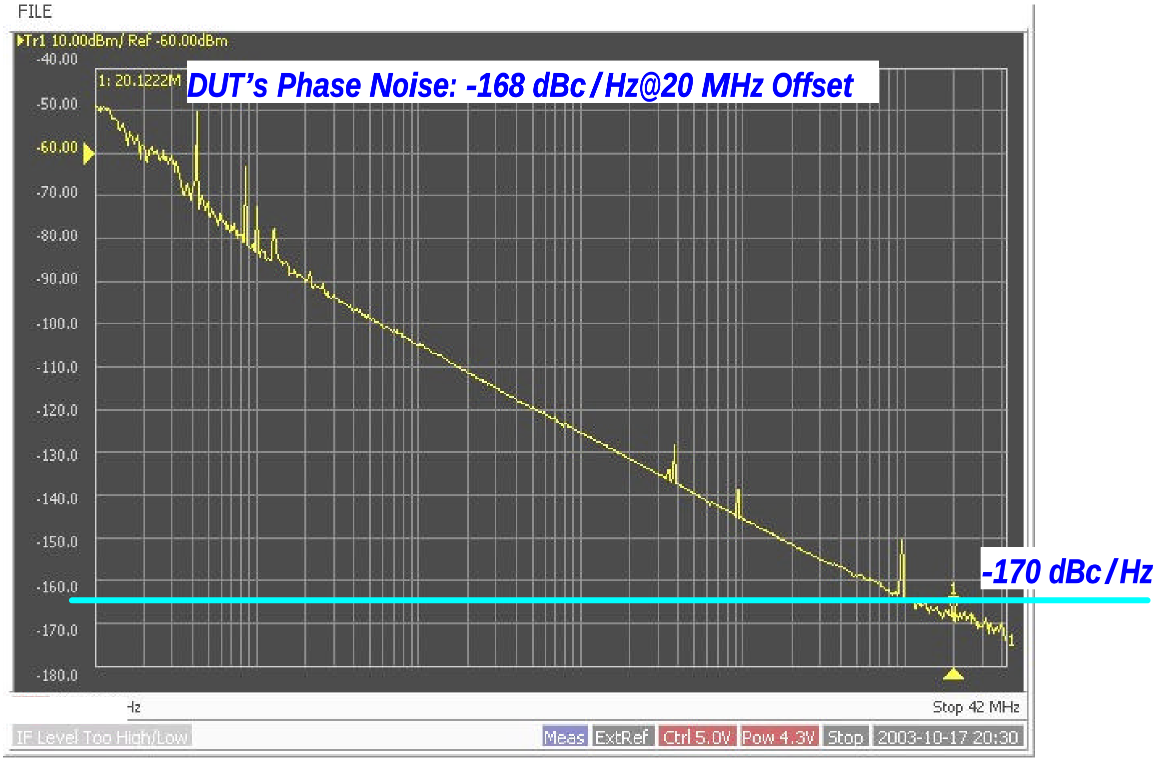
Example: 900 GSM VCO Measurement



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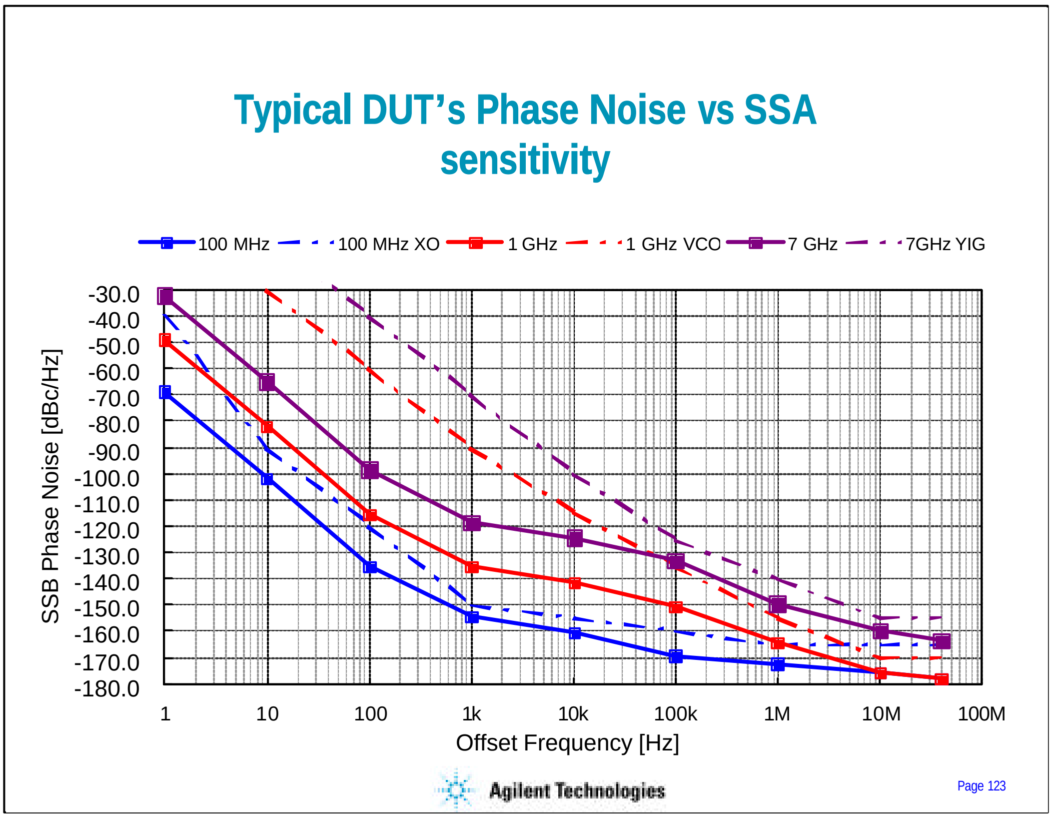
Typical DUT's Phase Noise vs SSA sensitivity
100 MHz
100 MHz XO
1 GHz
1 GHz VCO
7 GHz
7GHz YIG
SSB Phase Noise [dBc/Hz]
-30.0
-40.0
-50.0
-60.0
-70.0
-80.0
-90.0
-100.0
-110.0
-120.0
-130.0
-140.0
-150.0
-160.0
-170.0
-180.0
1
10
100
1k
10k
100k
1M
10M
100M
Offset Frequency [Hz]
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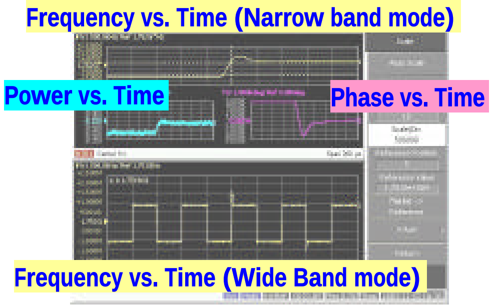
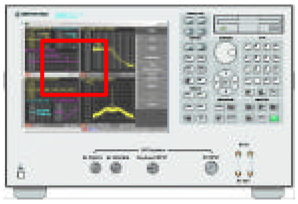
Signal Source Analyzer
Transient Measurement

- Simultaneous Measurements
  - Frequency Transient
  - Phase Transient
  - Power Transient
- Exceptional High Sampling Rate of 10 nsec
- External Hardware Trigger
- Video & Pre-Trigger to Observe Before/After the Event



Example: GSM PLL Synthesizer Measurement



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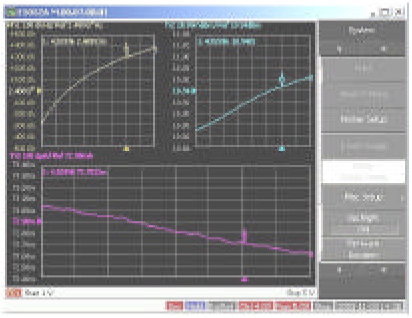
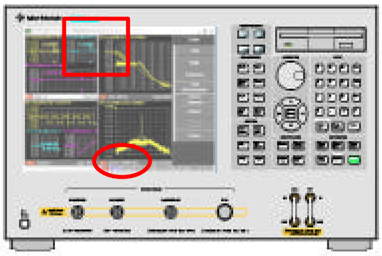
Signal Source Analyzer's
Frequency, Power & DC Current Measurements

- Simultaneous Measurements
  - Frequency vs. DC Tune or DC Power Voltage
  - RF Power vs. DC Tune or DC Power Voltage
  - DC Consumption Current (@ DC Power)
  -



- Built-in Ultra Low Noise DC Sources for Accurate Oscillator Characterization
  - 1nV/SQRT(Hz) for DC Tune Voltage
  - 10nV/SQRT(Hz) for DC Power Voltage



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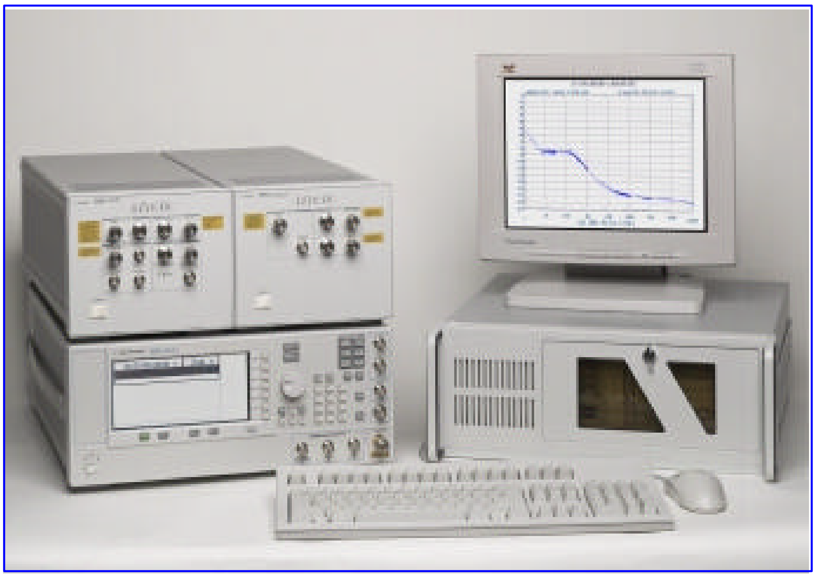
E5505A Phase Noise System



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The Agilent E5505A provides the complete set of measurement types previously discussed. As shown the offset frequency coverage is 0.01Hz to 2 MHz but with the addition of a low cost spectrum analyser such as the E4401B the upper offset range is extended to 100 MHz. Measurement speed is excellent due to the use of the 5 Msample/s digitiser handling the data below 2 MHz, a typical 10 Hz to 100 MHz measurement cycle taking ~ 10 secs to complete.

E5505A Phase Noise System

Oscillator measurements:

- FM discriminator
- PLL w/reference source
- AM noise
- CW and pulsed signals

Residual measurements

- CW and pulsed

Baseband noise measurements

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The E5500 series of phase noise measurement systems offer the complete set of measurement types we have discussed today for both CW and pulsed signals.

E5505A Phase Noise System

- Fast measurement speed
- Wide Offset measurement range (0.01 Hz to 100 MHz)
- Wide carrier frequency range (50 kHz to 110 GHz)
- Easy-to-use
- Configuration flexibility
- Measurement flexibility
- Low overall cost-of-measurement
- Full remote control possible over LAN/Serial interfaces
- Standard programmable interface

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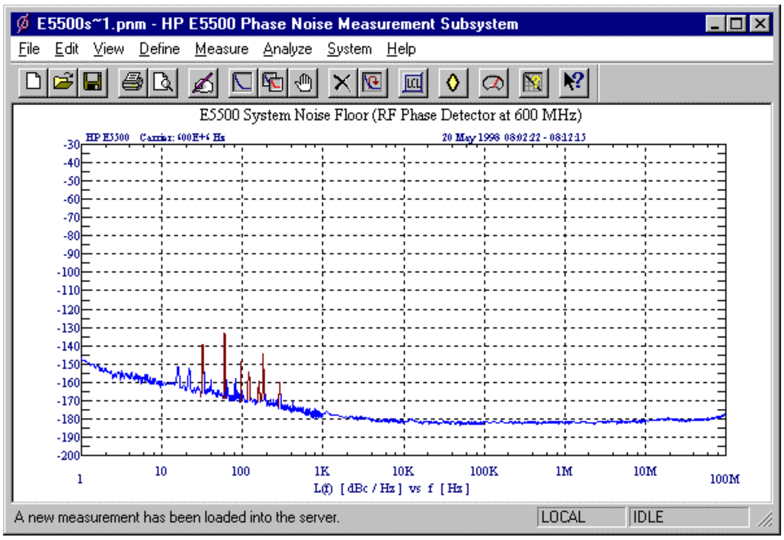
In addition to the complete set of measurement types and techniques, the Agilent E5505A extends the offset measurement range to 100 MHz, the carrier frequency range down to 50 KHz and up to 110 GHz and provide a very easy-to-use graphical user-interface.

With the standard programmable interface, this solution can easily be integrated into any existing production or other ATE environment. And since measurements can be configured with many existing Agilent spectrum analyzers and sources, the overall cost of measurements can be lowered.

E5505A Phase Noise System

Typical Noise Floor



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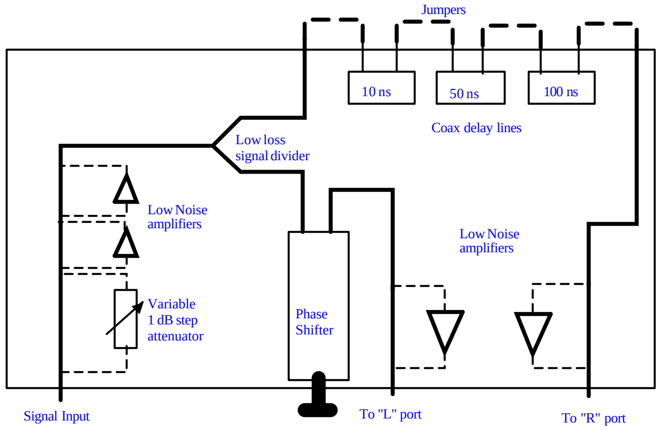
The typical noise floor of the measurement system without a reference sources is lower than -180 dBc/Hz for offsets greater than 10 KHz.

E5505A Phase Noise System

Measurement convenience accessories available:

e.g. 2 GHz Phase shifter test set: 70429A Opt K16



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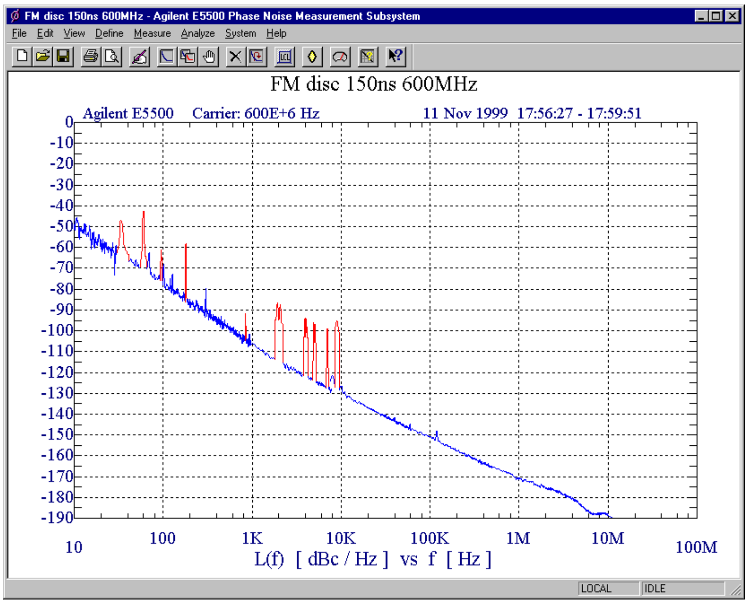
Measurement convenience accessories such as the Agilent 70429A Opt K15 and K16 phase shifter test sets provide a simple implementation for various measurement types such as the discriminator technique as well as residual measurement in the RF frequency range.

This particular test set includes 4 low noise amplifiers (5 MHz to 2 GHz), a variable 1 dB step attenuator, a low loss signal divider, 3 different lengths of coax delay which can be used for discriminator measurements, and an RF phase shifter (500 MHz to 2 GHz) with coax jumpers allowing flexible configurations to be used. For residual measurements, the delay lines would not be used, being replaced by the UUT.

600 MHz Example Measurement



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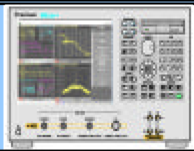
"Phase Noise Only" Measurement Comparison

	E5052A Signal Source Analyzer	E5505A Phase Noise System
Phase Noise Measurement Capabilities	 Absolute CW Phase Noise w/o the need for ext. reference source	Most flexible phase noise measurements: 2-port (residual) and absolute phase noise on CW and Pulsed carriers. AM noise, baseband noise, Phase and time jitter plus identification of deterministic (Spurious) signals
Frequency range	10MHz to 7 GHz or 110 GHz[1]	50 kHz to 1.6G, or 6 G, or 18G, or 26.5G, or 110 GHz[1]
Offset Frequency Range	1 Hz to 40 MHz	0.01 Hz to 100 MHz 
Noise Floor	-178 dBc/Hz [3]	-180 dBc/Hz [2]
Reference Source	Phase noise of integrated source is much better than any of Agilent's signal generators, such as PSG series.	Requires Customer-choice of Reference Source
More	Simplest of Ease of Use Exceptional High Speed Measurement	Most flexible measurements and equipment configuration for insight into designs

[1] Above 26.5 GHz requires 11970 series harmonic mixer with N5507A down converter
[2] Actual noise floor limited by the reference source and/or down converter used
[3] Integrated reference source and cross-correlation technique enhances phase noise floor by 20*LOG(N)

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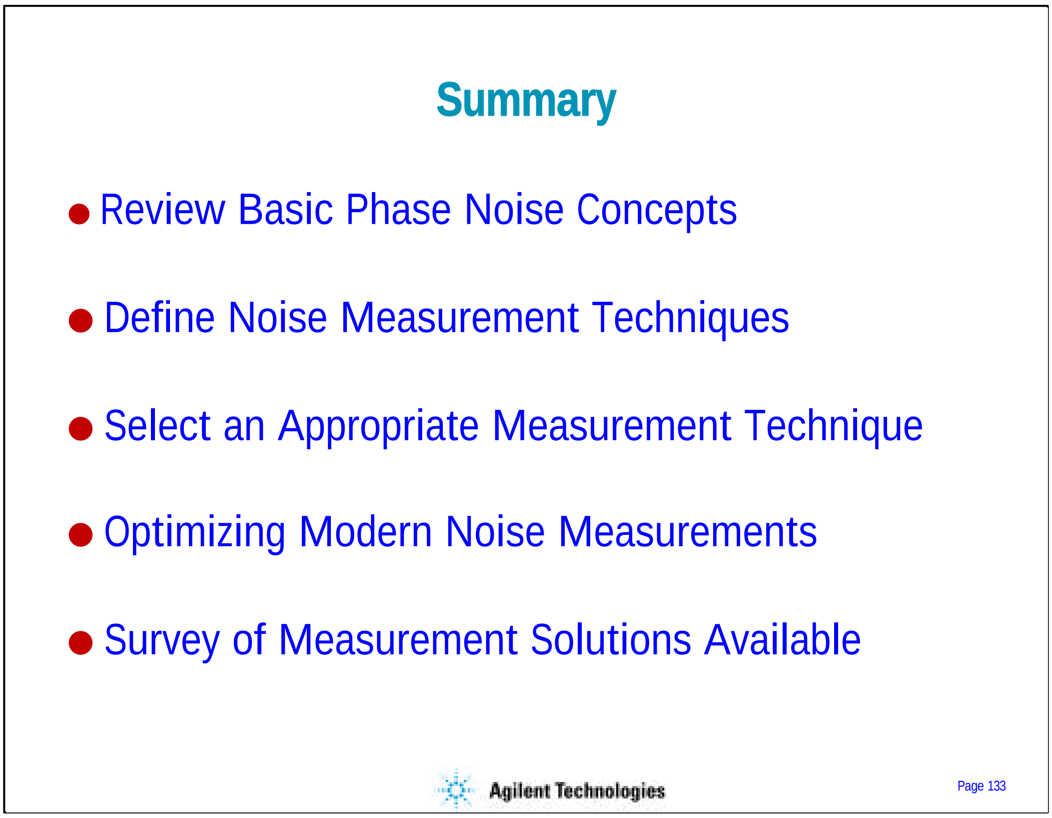

This a good time to ask specific questions which have not been addressed today.