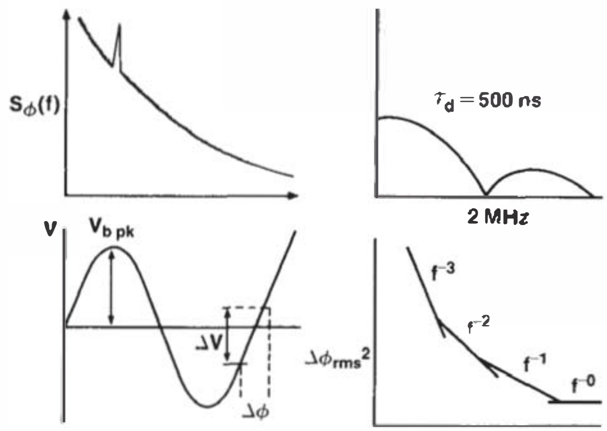
PHASE Noise

RF & Microwave Phase Noise Measurement Semin

HEWLETT PACKARD

WELCOME TO THE
HEWLETT PACKARD
RF and Microwave
Phase Noise Measurement Seminar



Welcome to the Phase Noise Measurement Seminar. Today, measuring and specifying phase noise has become increasingly important as phase noise is often the limiting factor in many RF and microwave systems. Both oscillators and devices have phase noise associated with them that must be measured.

PHASE NOISE CHARACTERIZATION OF SOURCES AND DEVICES

Typical Sources: Crystal Oscillators
YIG Oscillators
SAW Oscillators
DROs
Synthesizers
Cavity Oscillators

Typical Devices: Amplifiers
Multipliers
Dividers

WHAT ELSE?

We want to encourage interactive discussion throughout today's seminar so that we can all share your measurement problems and experiences. We hope to benefit, too, with new applications awareness, measurement conditions, and measurement technique. To start, we'd like a list from you of issues and concerns that you'd like to see resolved in this seminar and in particular the types of sources or devices that you must characterize and their frequency ranges.

SEMINAR AGENDA

I. Basis of Phase Noise
- ⇨ Why is Phase Noise important?
- What is Phase Noise?
- What causes Phase Noise?
- Quantifying Phase Noise

II. Measurement Techniques on Sources
1. Direct Spectrum Method
2. Heterodyne/Counter Method
3. Phase Detector Method
4. Frequency Discriminator Method
5. Summary of Source Measurement Techniques

III. Measurement of Two-Port Phase Noise (Devices)

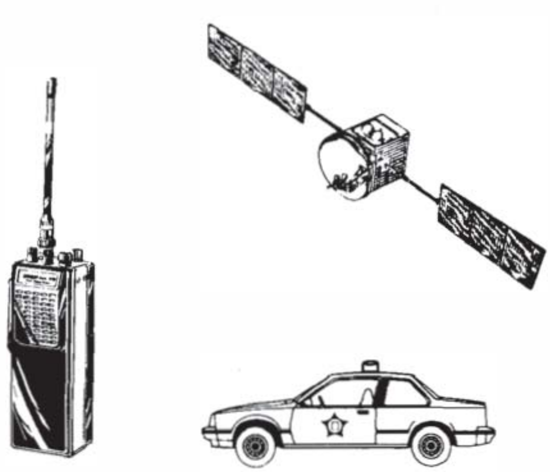
IV. Phase Noise Measurement on Pulsed Carriers

V. AM Noise Measurement

These are the topics that we'll discuss today. Where practical, we shall demonstrate the measurement concepts using both manual and automatic systems. We shall also try to look more at the practical aspects of these measurements; many mathematical derivations are left to the expertise in the list of references at the end of your handout. Also note that in the back of your handout is a glossary of symbols used in the seminar. (This glossary will also be useful in your further reading ofthe references, as some authors use different symbols for the same parameter.)

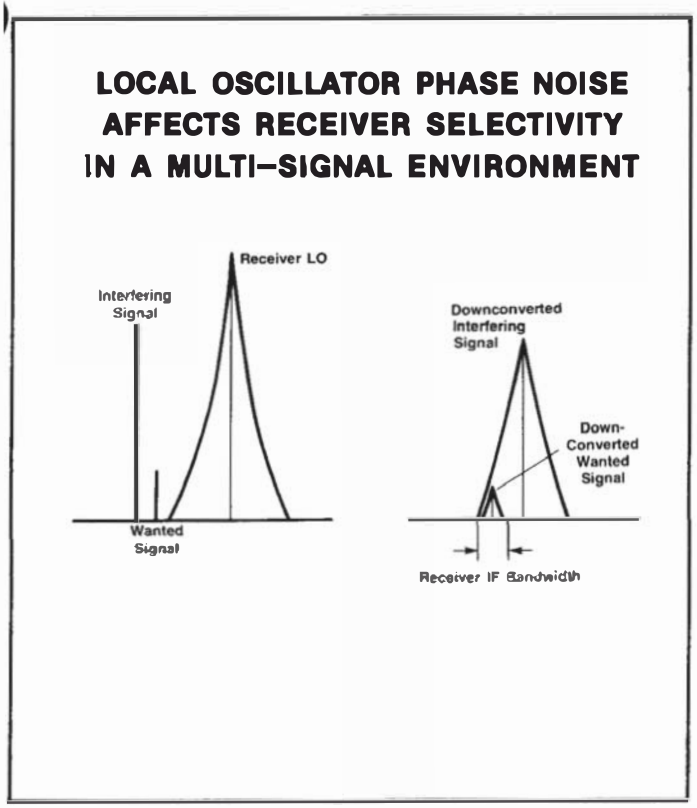
Why is Phase Noise Important?

IMPORTANCE OF PHASE NOISE VARIES WITH APPLICATION

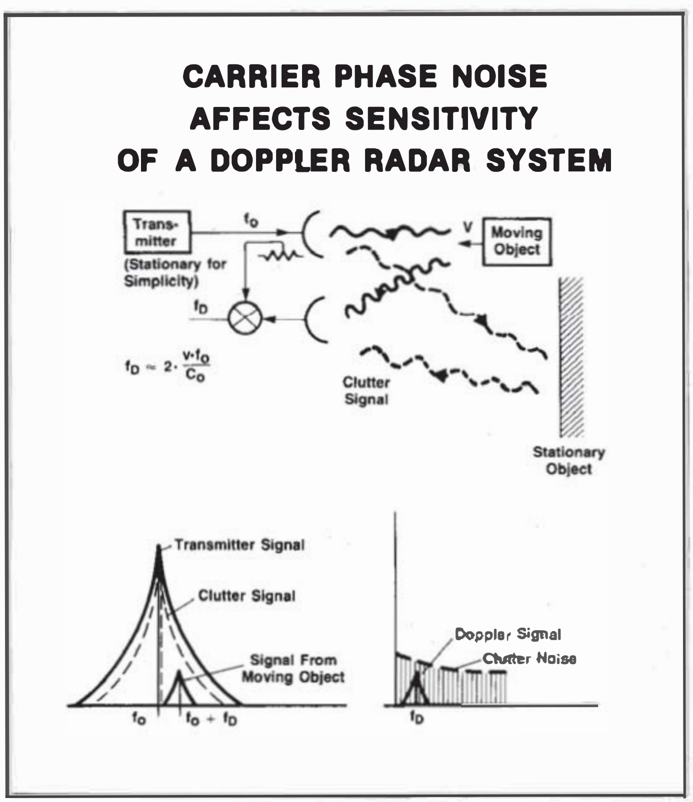


There is obviously a difference in the short-term stabilities of sources. But what is the reason to quantify this difference? Primarily, as we shall see further, short-term stability is often THE limiting factor in an application. The three applications shown here all REQUIRE a certain level of performance in short-term stability. Though the required level of performance differs, short-term stability is crucial in each application.

Short-term stability is not a parameter that comes for free with good design in other areas. In fact, it is one of the most expensive parameters to design for. Because these three applications require different levels of performance, it is very important to quantify and measure the required short-term stability, and to then choose the right source for the application.

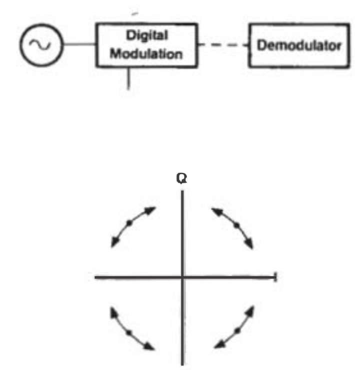


A high performance superheterodyne receiver serves as a good example for illustration. Suppose two signals are present at the input of receiver. These signals are to be down-converted to an IF where filters can separate the desired signal for processing. If the larger signal is desired, there should be no difficulty in recovering it. A problem may arise, however, if the desired signal is the smaller of the two. The phase noise of the LO is translated directly to the mixer products. The translated noise in the mixer may completely mask the smaller signal. Even though the receiver's IF filters may be sufficient to remove the larger signal, the smaller signal is no longer recoverable due to the LO phase noise. A noisy LO can degrade a receiver's dynamic range as well as selectivity.



Doppler radars determine the velocity of a target by measuring the small shifts in frequency that the return echoes have undergone. Unfortunately, the return signal includes much more than just the target echo. In the case of an airborne radar, the return echo also includes a large "clutter" signal which is basically the unavoidable frequency-shifted echo from the ground. The ratio of main beam clutter to target signal may be as high as 80 dB, which makes it difficult to separate the target signal from the main beam clutter. The problem is greatly aggravated when the received spectrum has frequency instabilities caused by phase noise either on the transmitter oscillator or the receiver LO. Such phase noise on the clutter signal can partially or totally mask the target signal, depending on its level and frequency separation from the carrier.

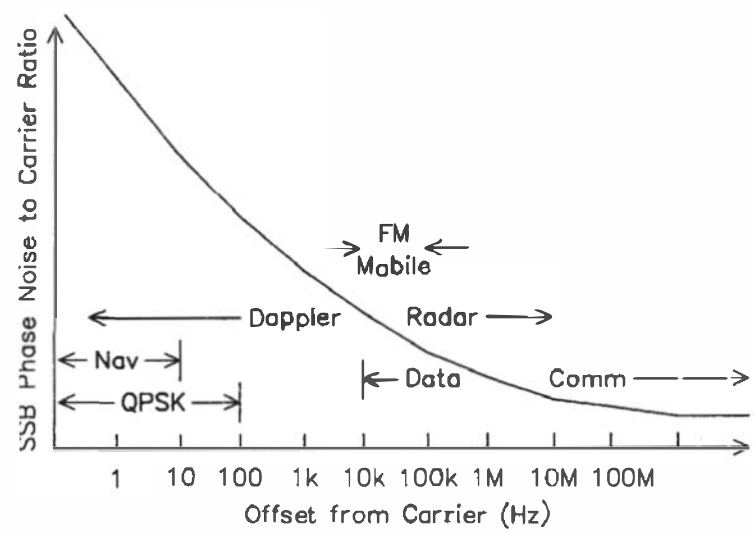
LOCAL OSCILLATOR PHASE NOISE AFFECTS THE BIT ERROR RATE OF A QPSK SYSTEM



State Diagram

In a Quadraphase Phase Shift Keying system, the IQ position of the information signal on the state diagram depends on the amplitude and phase information after demodulation. Amplitude noise affects the distance from the origin while phase noise affects the angular positioning. Close-in phase noise (or phase jitter in the time domain) on the system local oscillator affects the system bit-error rate.

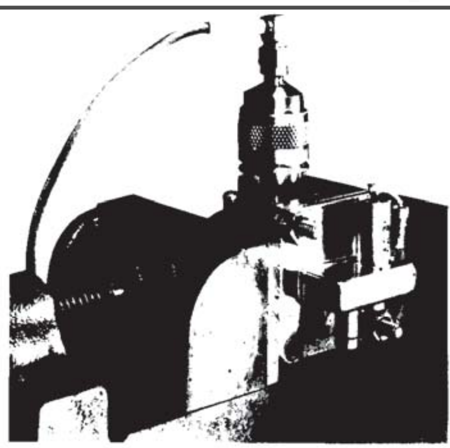
IS ALL NOISE IMPORTANT TO ALL OF THE PEOPLE ALL OF THE TIME?



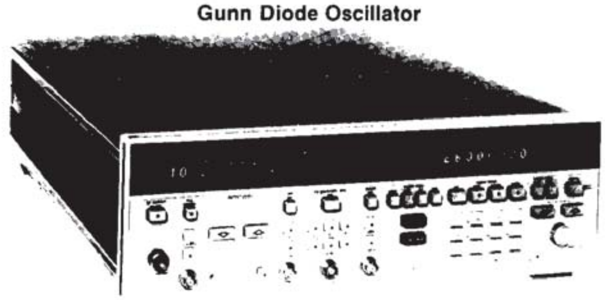
Phase noise is important in these applications, but where the noise is important (i.e., at what offset from carrier) differs. This graph shows some typical ranges offset frequencies where noise is important for differe applications. Because the range of offset frequencies where phase noise is important changes with the application, the type of source used also changes.

WHAT IS PHASE NOISE?

What is phase noise? What is a quantity with a statistical randomness?

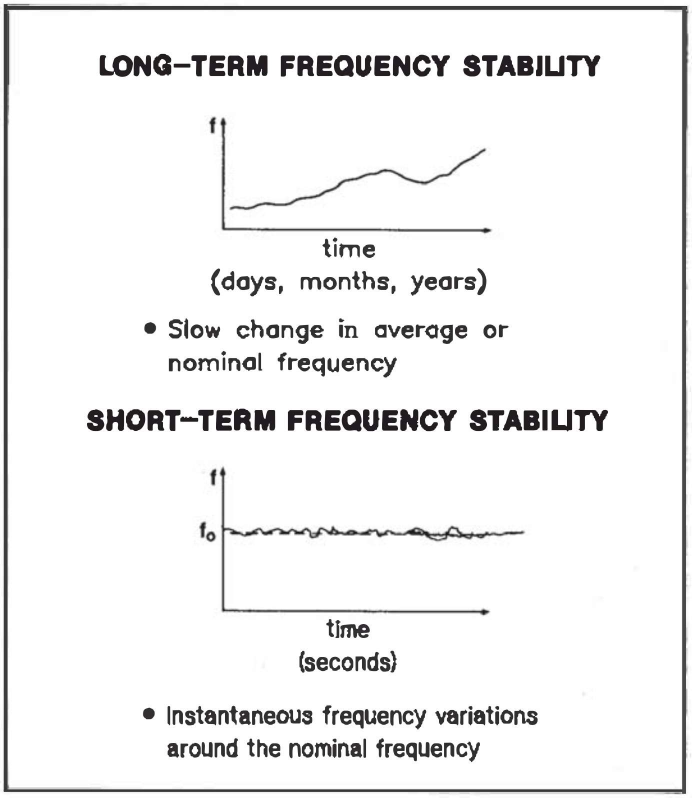


Gunn Diode Oscillator



Synthesized Source

Two very different sources with many differences in performance (and difference in cost!). One way they differ is in their frequency stability. This difference in frequency stability will definitely affect the type of application that they will be used in. What is frequency stability, and how can we describe the difference in frequency stability of these two sources?

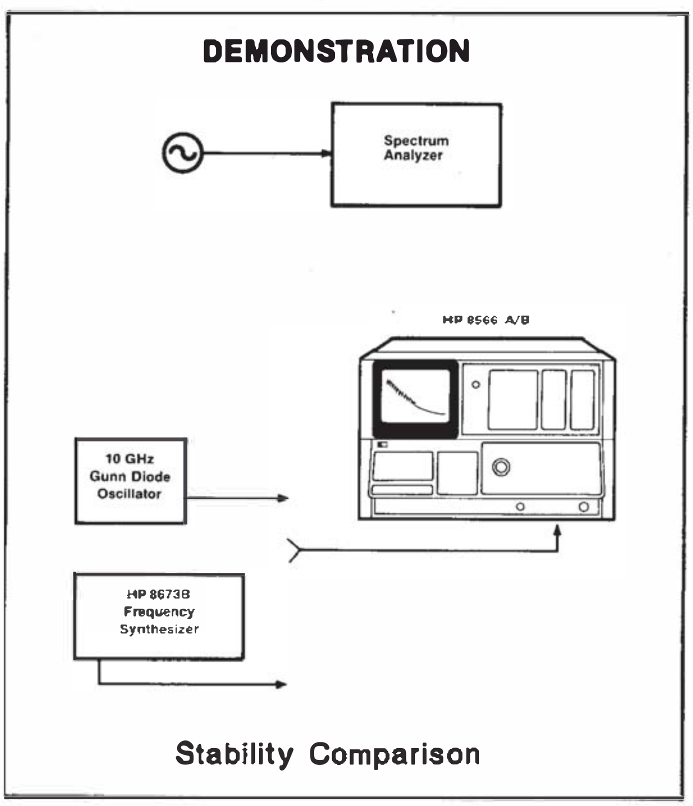


Frequency stability is generally defined by two parameters: long-term and short-term stability. It is commonly said that long-term frequency stability describes the variation in signal frequency that occurs over long time periods, and short-term stability refers to the variations that occur over time periods of a few seconds or less.

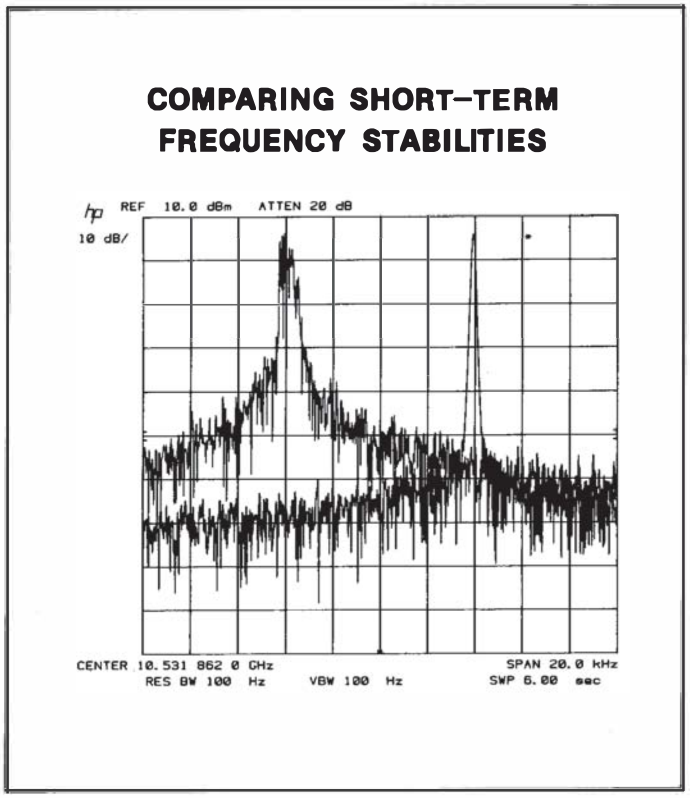
However, the dividing line between long-term and short-term stability is really a function of application. For example, in a communications system, all variations which are slower than the narrowest carrier or data clock tracking loop would be referred to as long-term, with the dividing line being a fraction of a second. On the other hand, a timekeeping system would observe day-to-day irregularities as short-term, with a dividing line corresponding to the length of a mission, which might be several days.

Long-term stability refers to slow changes in the average frequency with time due to secular changes in the resonator. It is usually expressed as a ratio, $\Delta f/f$ for a given period of time – hours, days, or even months. Long-term frequency stability is commonly called frequency drift, and is usually linear or sometimes exponential.

Short-term stability refers to changes in frequency which cannot be described as offset (static error) or drift, but are observed as random and/or periodic fluctuations about a mean. They are usually described in terms of variations about the nominal frequency that occur over time periods of a few seconds or less.

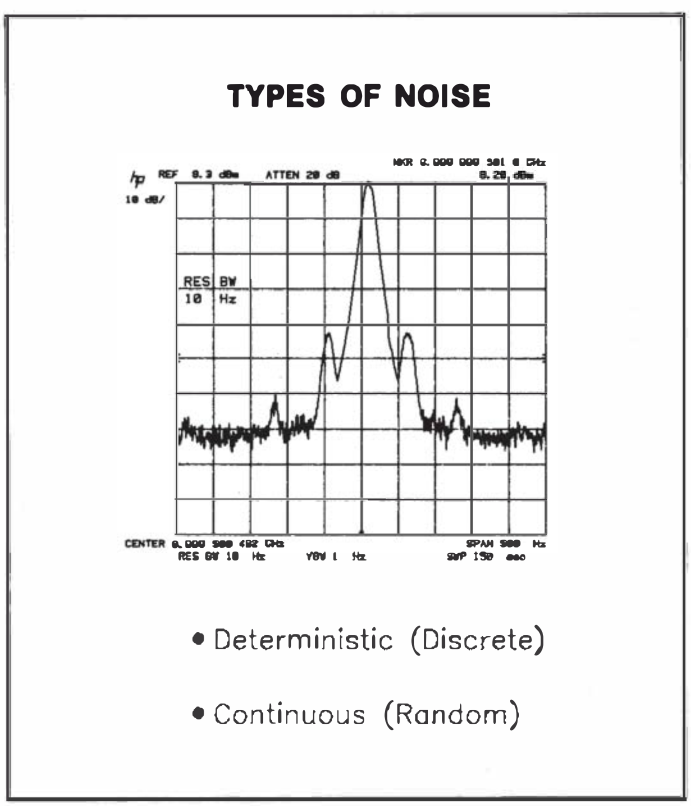


This demonstration illustrates the difference in the frequency spectrum of two sources with different short-term stability characteristics.



Short-term stability is more familiar to most of us in the frequency domain. Looking at a signal on an ideal spectrum analyzer (one with infinitely sharp filters and no short-term instability of its own), all of the signal's energy does not occur at a single spectral line, but rather some ofthe signal's energy occurs at frequencies offset from the nominal frequency.

Using a spectrum analyzer to observe our two example sources, it's obvious that the sources differ in short-term stability. How can we describe and quantify this difference? What units can we use to compare what we can visually see here? And once deciding upon units to use, how can we measure these values?



In discussing short-term stability, there are two "classes" of frequency variations – non-random (or deterministic) and random. The first, deterministic (or systematic, periodic, discrete, secular) are discrete signals which appear as distinct components on our ideal spectrum analyzer RF sideband spectrum. These signals, commonly called spurious, can be related to known phenomena in the signal source such as power line frequency, vibration frequencies, or mixer products.

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

IDEAL SIGNAL

$$V(t) = A_0 \sin 2\pi f_0 t$$

where

A_0 = nominal amplitude

f_0 = nominal frequency

REAL WORLD SIGNAL

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

where

$\varepsilon(t)$ = amplitude fluctuations

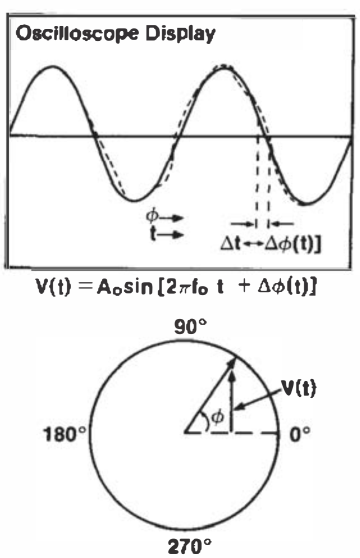
$\phi(t)$ = phase fluctuations

Before proceeding to the definitions of phase noise, let's get a more intuitive feel. If one could design a perfect oscillator, all signals could be described like this. In the frequency domain, this represents a signal with all energy at a single spectral line.

But in the real world, there's always a little something extra on your signal. Unwanted amplitude and frequency fluctuations are present on the signal. Note that the frequency fluctuations are actually an added term to the phase angle term of the equation of a signal. Because phase and frequency are related, you can speak equivalently about unwanted frequency or phase fluctuations.

IN THE TIME DOMAIN . . .

Phase Jitter



Angular Frequency

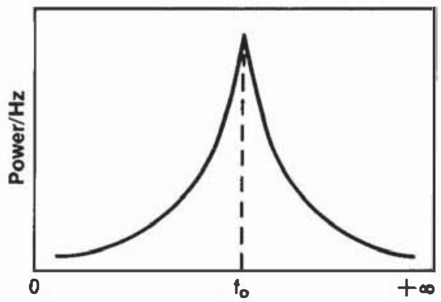
Concentrating first on the frequency fluctuations, let's see what these fluctuations would look like on a signal. In the time domain, phase is measured from a zero crossing, as illustrated by plotting the phase angle as the radius vector rotates at a constant angular rate determined by the frequency. Random noise processes affect the signal throughout its period.

Let's look on just one particular time in which the sine wave is perturbed for a short instant by noise. In this perturbed area, the ΔV and Δt (or Δφ) corresponds to another frequency. These perturbations repeat on each cycle at a recognizable, somewhat constant repetition rate. In fact, we will find that there is a significant amount of power in another signal whose period is the period of the perturbation shown.

Thus, in a sideband spectrum (rms power vs. frequency), we will observe a noticeable amount of power in the spectrum at the frequency corresponding to this perturbation, with an amplitude related to the characteristics of the perturbations. Thus, frequency variations, or phase noise since it is really instantaneous phase fluctuations, occur for a given instant of time within the cycle. How much time the signal spends at any given frequency is a statistically random phenomenon. This is best expressed as a spectral density function in the frequency domain.

IN THE FREQUENCY DOMAIN...

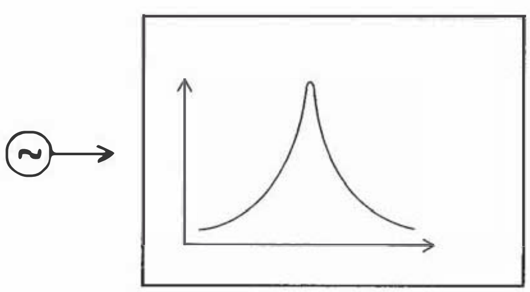
Power Spectral Density



Another way to think about phase noise is as a continuous spectrum of infinitely close phase modulation sidebands, arising from a composite of low frequency signals. A signal's stability can be described as power spectral density of phase fluctuations or frequency fluctuations (and later on we'll see the power spectral density of amplitude fluctuations).

ABSOLUTE (TOTAL) NOISE

- Specified on sources or complete system

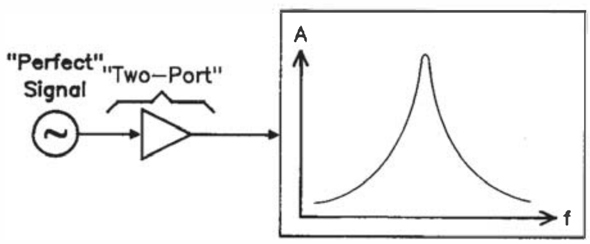


Discussions about phase noise can be divided into two topics: the total or "absolute" noise from an oscillator or system that generates a signal and the added or "two-port" noise that is added to a signal as it passes through a device or system.

Absolute noise measurements on the output signal of a system would include the noise that occurs when the signal is generated and the "two-port" noise added by the system signal processing devices.

TWO–PORT (RESIDUAL OR ADDITIVE) NOISE

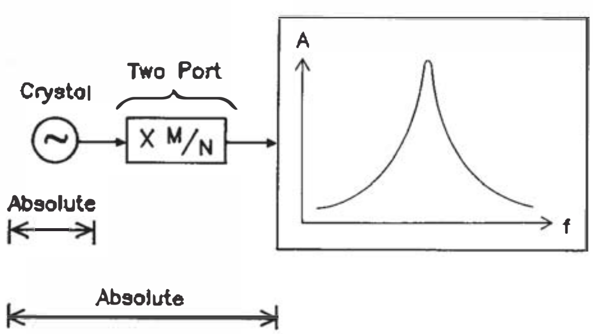
- Specified on devices or subsystem



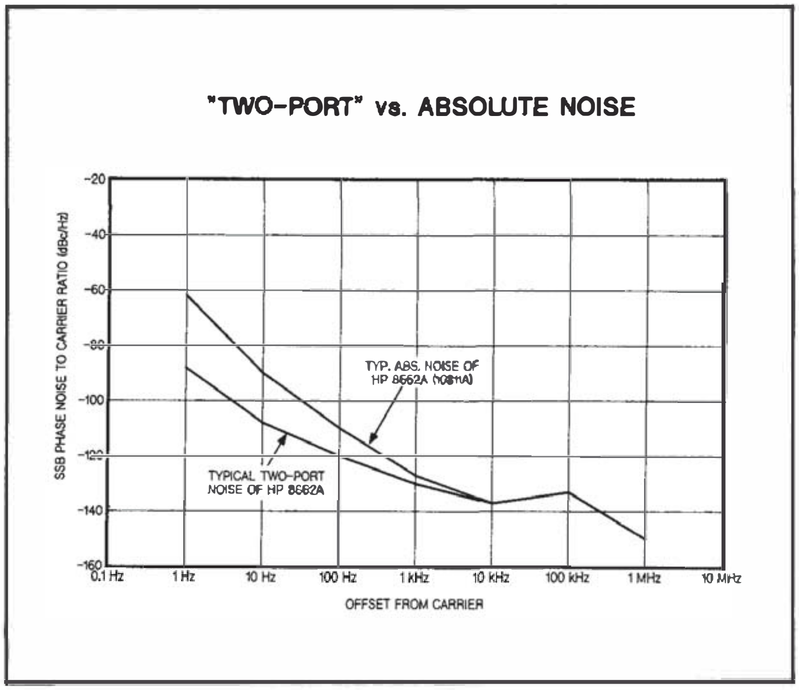
"Two-port" (or residual or additive) noise refers to the noise of devices (amplifiers, dividers, delay lines). Two-port noise is the noise contributed by a device, regardless of the noise of the reference oscillator used. One way to look at "two-port" noise is how much noise would be added by a device if a perfect (noise-less) signal were input to it. The name "two-port" emphasizes the contributed nature of the noise of devices.

"TWO–PORT" VS. ABSOLUTE NOISE

EXAMPLE: SYNTHESIZER



A "system" – such as a synthesizer – has both two-port and absolute noise associated with it. The reference signal of the synthesizer, comprising an oscillating element, has absolute noise. The synthesizer circuitry – phase lock loops, multipliers, dividers, etc. – have some two-port noise contribution. The integrated system, the synthesizer, also has a value for absolute noise, or all noise present at the output.



The absolute noise of the reference and the synt
and the two-port noise of the synthesizer are comp
here. Though the units have not been explained y
are still several important points about this graph
typically two-port noise of devices is less than the
absolute noise on sources, in particular at higher
frequencies. Second, even with a perfect reference
absolute noise of the system could never be below
port noise level.

What Causes Phase Noise?

THE BASICS OF NOISE GENERATION

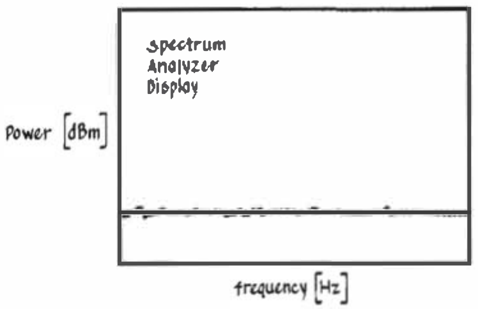
Thermal Noise?

Noise Figure?

Phase Noise?

In this section we will briefly look at the basics of noise generation. What are thermal noise and noise figure and how are they related to phase noise?

THERMAL NOISE



$N_p = KTB$

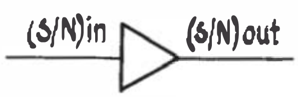
K = Boltzman's constant
T = Temperature K
B = Bandwidth

For T = 290K

$$N_p = -204 \frac{dB(Watts)}{Hz} = \frac{-174\ dBm}{Hz}$$

Thermal noise is the mean available noise power per HZ of bandwidth from a resistor at a temperature of TK. 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.

NOISE FIGURE



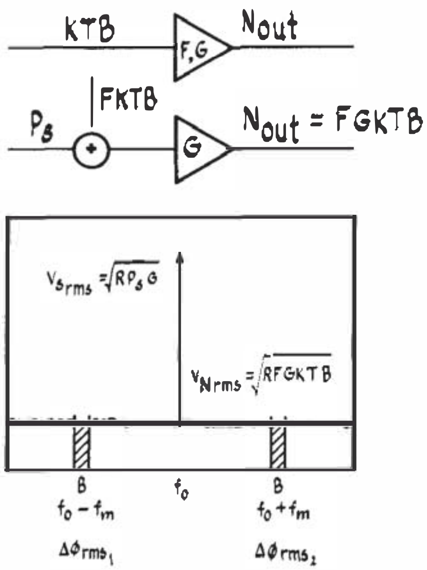
$$F \equiv \left.\frac{(S/N)in}{(S/N)out}\right|_{T_s = 290K}$$

$$F(dB) \equiv 10 \log \left.\frac{(S/N)in}{(S/N)out}\right|_{T_s = 290K}$$

What do thermal noise and noise figure have to do with phase noise?

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

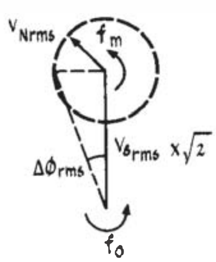
AMPLIFIER OUTPUT NOISE AS A FUNCTION OF THERMAL NOISE AND NOISE FIGURE



The noise power at the output of an amplifier can be calculated if its gain and noise figure are known. The noise at the 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.

USING PHASOR RELATIONSHIPS



For small $\Delta\phi_{rms}$

$$\Delta\phi_{rms} = \frac{V_{N_{rms}}}{\sqrt{2}\,V_{S_{rms}}} = \sqrt{\frac{FGKTB}{2P_sG}}$$

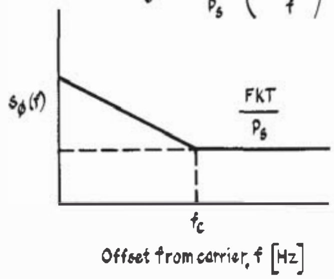
$$\Delta\phi_{rms}\ \text{total} = \sqrt{\Delta\phi^2_{rms_1} + \Delta\phi^2_{rms_2}} = \sqrt{\frac{FKTB}{P_s}}$$

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

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 $V_{N_{rms}}$ produces a $\Delta\varnothing_{rms}$ term. For small $\Delta\varnothing_{rms}$, $\Delta\varnothing_{rms} = V_{N_{rms}}/V_{S_{peak}}$. The total $\Delta\varnothing_{rms}$ can be found by adding the two individual phase components powerwise. Squaring this result and dividing by the bandwidth gives $S_\varnothing(f)$, 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.

ACTUAL PHASE NOISE

$$S_\phi(f) = \frac{FKT}{P_s}\left(1 + \frac{f_c}{f}\right)$$

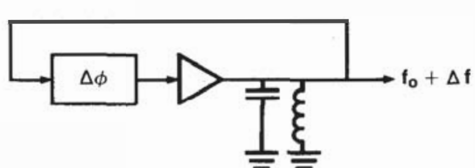


f_c = corner frequency

- Phase noise "flicker" appears $<f_c$
- Rule of thumb: "flicker" noise is ~−120 dBc/Hz at 1 Hz offset

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 floor at an empirically determined frequency f_c. For offset frequencies below f_c, $S_\phi(f)$ increases with f^{-1}.

NOISE PROCESSES IN AN OSCILLATOR

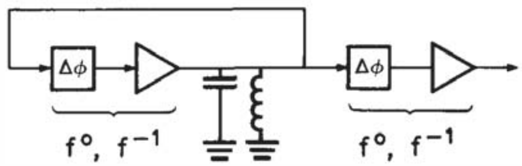


$$\Delta f = \Delta\phi \frac{f_0}{2Q}$$

Simple Feedback Model

In an oscillator, the white f^0 and flicker f^{-1} phase modulations cause even greater slopes of noise spectra. Let's see how that happens. First, add a resonator of some quality factor Q to the output of an amplifier. Second, connect the resonator output back to the amplifier input in the proper polarity for positive feedback. Third, consider the f^0 and f^{-1} of the amplifier to be represented by a phase modulator $\Delta\phi$ with a perfect amplifier. Next, any oscillator will shift *frequency* in response to a *phase* change anywhere in its loop, $\Delta f = \Delta\phi f_0/2Q$. Since f_0 and Q are constants, then phase modulation is converted directly to frequency modulation. This makes their spectral slopes 2 units more negative.

NOISE PROCESSES IN AN OSCILLATOR



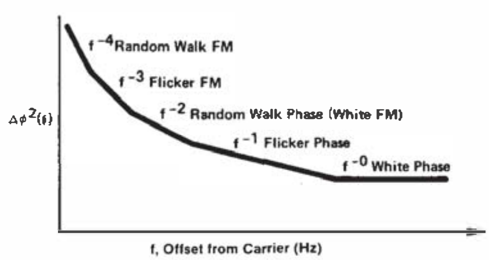
becomes f^{-2} , f^{-3}

thru $\Delta f = \Delta \varnothing \frac{f_o}{2Q}$

Simple Model with Buffer Amplifier

So 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 f^0 and f^{-1} noise slopes to the output signal.

NOISE PROCESSES IN THE FREQUENCY DOMAIN



The resulting phase noise plot for an actual oscillator is as shown. The frequency domain response of a source would include terms like Random Walk, Flicker and White Phase Noise to describe the slope of spectral density for given offsets.

SEMINAR AGENDA

I. Basis of Phase Noise
- Why is Phase Noise important?
- What is Phase Noise?
- What causes Phase Noise?
- ⇨ Quantifying Phase Noise

II. Measurement Techniques on Sources
1. Direct Spectrum Method
2. Heterodyne/Counter Method
3. Phase Detector Method
4. Frequency Discriminator Method
5. Summary of Source Measurement Techniques

III. Measurement of Two-Port Phase Noise (Devices)

IV. Phase Noise Measurement on Pulsed Carriers

V. AM Noise Measurement

Quantifying Phase Noise

$\sigma_y^2(\tau)$

$\mathcal{L}(f)$

$S_\phi(f)$

QUANTIFYING PHASE NOISE

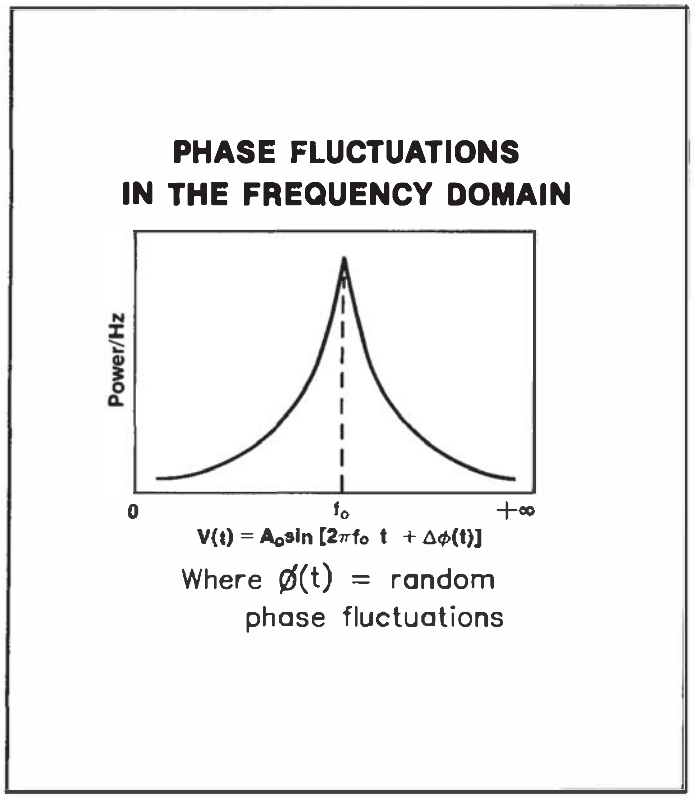
$S_{\Delta f}(f)$

$S_y(f)$

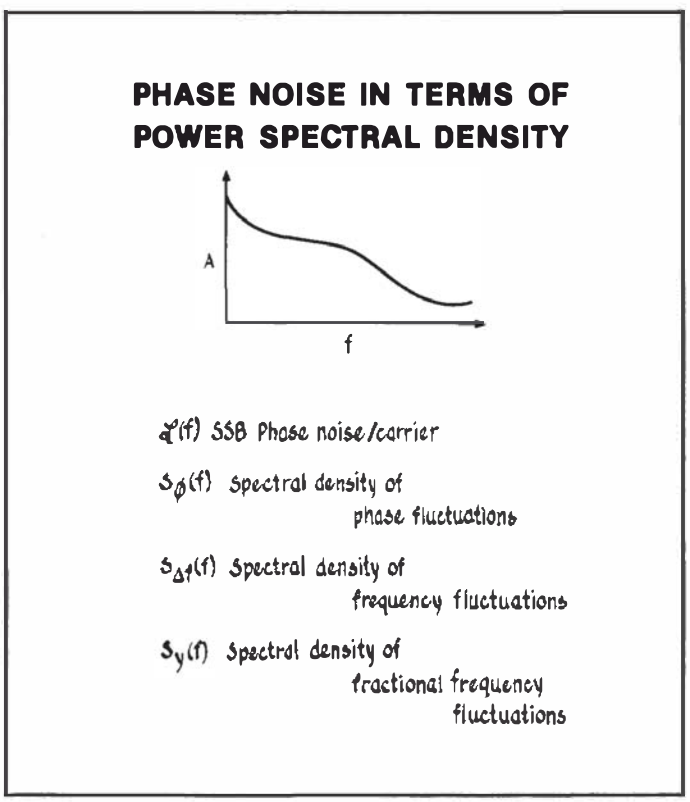
$\sigma_y(\tau)$

Δf_{res}

There are many different units used to quantify phase noise. In this section we will examine the most common ones, how they are derived and how they relate to one another.

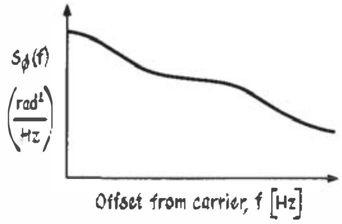


Due to random phase fluctuations, in the frequency domain a signal is no longer a discrete spectral line hut spreads out over frequencies both above and below the nominal signal frequency in the form of modulation sidebands. We need a way to quantify this frequency instability, or phase noise.



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 phase modulation of the carrier is actually equivalent to phase modulation by a noise source. The short-term instabilities are measured as low-level phase modulation of the carrier. Four units used to quantify the spectral density are shown.

$S_\phi(f)$ OR SPECTRAL DENSITY OF PHASE FLUCTUATIONS



Demodulate phase modulated signal with phase detector

$$\Delta V_{out} = K_\phi \Delta\phi_{in} \qquad K_\phi = V/rad$$

On spectrum analyzer

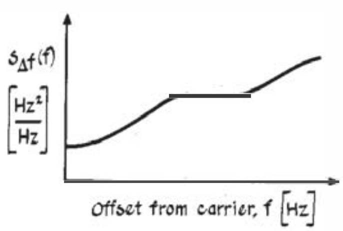
$$\Delta V_{rms}(f) = K_\phi \, \Delta\phi_{rms}(f) \; [V]$$

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

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

A measure of phase instability often used is $S_\varnothing(f)$, the spectral density of phase fluctuations on a per-Hertz basis. If we demodulate the phase modulated signal using a phase detector we obtain V_{out} as a function of the phase fluctuations of the incoming signal. Measuring V_{out} on a spectrum analyzer gives $\Delta V_{rms}(f)$ proportional to $\Delta\varnothing_{rms}(f)$. $S_{V_{rms}}(f)/K_\varnothing^2$ gives $\Delta\varnothing^2_{rms}(f)$ which is the spectral density an equivalent phase modulating source in rad²/Hz. This spectral density is particularly useful for analysis of phase noise effects on systems which have phase sensitive circuits such as digital FM communication links.

$S_{\Delta f}(f)$ OR SPECTRAL DENSITY OF FREQUENCY FLUCTUATIONS



$S_{\Delta f}(f)$ can be derived from $S_\phi(f)$

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

Transformed into the frequency domain .....

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

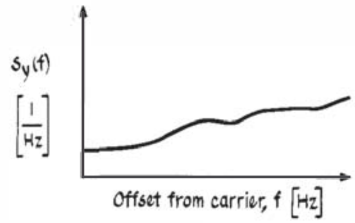
$$\Delta f^2_{rms}(f) = f^2 \Delta\phi^2_{rms}(f) \; [Hz^2]$$

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

$$S_{\Delta f}(f) = f^2 S_\phi(f) \left[\frac{Hz^2}{Hz}\right]$$

Another common term for quantifying short term frequency instability is $S_{\Delta f}(f)$, the spectral density of frequency fluctuations on a per-Hertz basis. $S_{\Delta f}(f)$ can be derived from $S_\varnothing(f)$ by transforming $\Delta f(t)$ from the time domain to the frequency domain by Laplace transform. This gives $\Delta f(f)_{rms} = f\Delta\varnothing(f)_{rms}$ or $\Delta f^2(f)_{rms} = f^2\Delta\varnothing^2(f)_{rms}$ which is the spectral density of frequency fluctuations in Hz²/Hz. Note $S_{\Delta f}(f) = f^2 S_\varnothing(f)$ [Hz²/Hz]. Caution must be taken when using $S_\varnothing(f)$ and $S_\varnothing(f)$ to compare the phase noise of sources at different frequencies.

$S_y(f)$ OR SPECTRAL DENSITY OF FRACTIONAL FREQUENCY FLUCTUATIONS



$S_y(f)$ is related to $S_\phi(f)$

$$y(t) = \frac{\Delta f(t)}{f_0} = \frac{\frac{1}{2\pi}\frac{d\Delta\phi(t)}{dt}}{f_0}$$

Transformed into the frequency domain .....

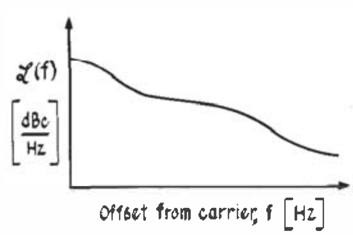
$$y(f) = \frac{2\pi f}{2\pi f_0}\Delta\phi(f)$$

$$y^2_{rms}(f) = \frac{f^2}{f_0^2}\Delta\phi^2_{rms}(f)$$

$$S_y(f) = \frac{f^2}{f_0^2}\frac{\Delta\phi^2_{rms}(f)}{B} = \frac{f^2}{f_0^2}S_\phi(f)\left[\frac{1}{Hz}\right]$$

$S_y(f)$, the spectral density of fractional frequency fluctuations allows direct comparison between sources of different carrier frequencies. $S_y(f)$ is also related to $S_\phi(f)$ and $S_{\Delta f}(f)$. Using the same Laplace transform approach on $\Delta f(t)/f_o$ we see that the spectral density of fractional frequency fluctuations is equal to the spectral density of frequency fluctuation divided by f_o^2.

SINGLE SIDEBAND PHASE NOISE $\mathcal{L}(f)$ OR SIDEBAND POWER WITH RESPECT TO CARRIER LEVEL

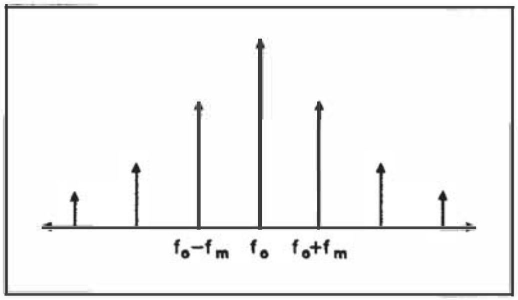


$$\mathcal{L}(f) = \frac{\text{Power Density (One Phase Modulation Sideband)}}{\text{Carrier Power}}\left[\frac{dBc}{Hz}\right]$$

$\mathcal{L}(f)$ can be derived from $S_\phi(f)$ using Phase Modulation Theory

$\mathcal{L}(f)$ is an indirect measure of noise energy easily related to the RF power spectrum observed on a spectrum analyzer. $\mathcal{L}(f)$ is defined as a the ratio of the power in one phase modulation sideband on a per-Hertz basis, to the total signal power. $\mathcal{L}(f)$ is usually presented logarithmically as a plot of phase modulation sidebands in the frequency domain, expressed in dB relative to the carrier per Hertz of bandwidth [dBc/Hz]. We will see that $\mathcal{L}(f)$ can be derived from $S_\varnothing(f)$ using phase modulation theory.

PHASE MODULATION AT f_m ...

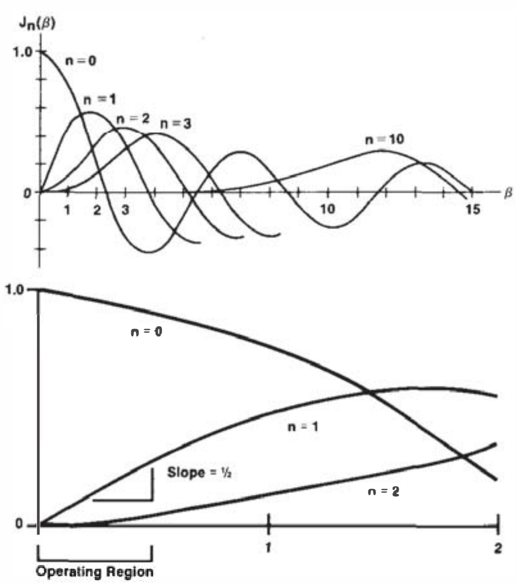


Produces sidebonds at f_m intervals from carrier.

Amplitude of carrier and sidebands ore determined by $\Delta\phi_{peak}$, the modulation index (β).

Phase modulation at a rate of f_m produces carrier sidebands spaced symmetrically about the carrier at intervals which are multiples of the modulation rate. The amplitude of the carrier and sidebands are determined by the modulation index(β) which is equal to $\Delta\phi_{peak}$.

BESSEL FUNCTIONS RELATE CARRIER AMPLITUDE TO SIDEBAND AMPLITUDE



Bessel functions relate the carrier amplitude to the sideband amplitude. Here we see Bessel coefficients for the first 10 sidebands as a function of the modulation index, $\Delta\varnothing_{peak}$. We can relate $S_{\varnothing}(f)$ to $\mathcal{L}(f)$ by making an important assumption. $\mathcal{L}(f)$ refers to the power density in one phase modulation sideband. We can see from the graph that the sideband power will be restricted to the first sideband only for modulation indices $<<1$ radian. For the same restriction, the relative amplitude of the carrier will always be one, and the slope of the Bessel function will be ½.

FOR $\Delta\phi_{peak}$ << 1 RADIAN

$$\frac{V_{ssbpk}(f)}{V_{spk}} = J_1 = \frac{1}{2}\Delta\phi_{pk}(f) \quad (\text{sinusoidal } \Delta\phi)$$

$$\frac{V_{ssb}^2(f)}{V_s^2} = \frac{P_{ssb}(f)}{P_s} = \frac{1}{4}\Delta\phi^2_{peak}(f)$$

$$\frac{P_{ssb}(f)}{P_s} = \frac{1}{4}\left[\sqrt{2}\,\Delta\phi_{rms}(f)\right]^2$$

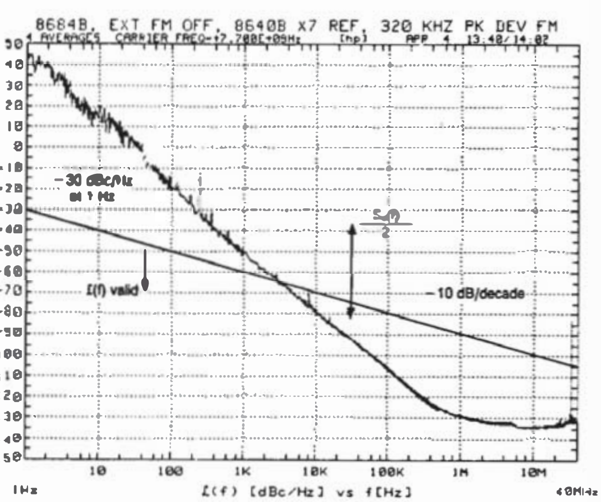
$$\left.\frac{P_{ssb}(f)}{P_s}\right|_{B = 1Hz} = \frac{1}{2}\,\frac{\Delta\phi^2_{rms}(f)}{B}\left[\frac{rad^2}{Hz}\right]$$

$$\left.\frac{P_{ssb}(f)}{P_s}\right|_{B = 1Hz} = \mathcal{L}_R(f) = \frac{1}{2}S_\phi(f)\left[\frac{rad^2}{Hz}\right]$$

Thus for $\Delta\varnothing_{peak} << 1$, we have $V_{ssb}(f)/V_s = \frac{1}{2}\Delta\varnothing_{peak}(f)$. Then $P_{ssb}(f)/P_s = \frac{1}{4}\Delta\varnothing^2_{peak}(f)$. Converting the peak phase deviation to an rms value and normalizing to a 1 Hz bandwidth we have $L_R(f) = \frac{1}{2}S_\varnothing(f)$.

REGION OF VALIDITY OF

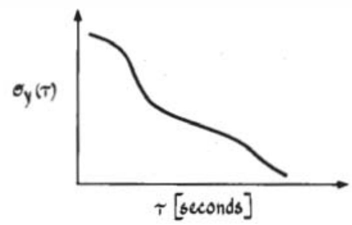
$$\mathcal{L}(f) = \frac{S_\phi(f)}{2}$$



Caution must be exercised when $\mathcal{L}(f)$ is calculated from the spectral density of the phase fluctuations because of the small angle criterion. This plot of $\mathcal{L}(f)$ resulting from the phase noise of a free running VCO illustrates the erroneous results that can occur if the instantaneous phase modulation exceeds a small angle. Approaching the carrier, $\mathcal{L}(f)$ is obviously increasingly in error as it reaches a relative level of +45 dBc/Hz at a 1 Hz offset (45 dB more noise power at a 1 Hz offset in a 1 Hz bandwidth that the total power in the signal.)

On this graph the 10 dB/decade line is drawn on the plot for a peak phase deviation of 0.2 radians integrated over any one decade of offset frequency. At approximately 0.2 radians the power in the higher order sidebands of the phase modulation is still insignificant compared to the power in the first order sideband which ensures the calculation of $\mathcal{L}(f)$ is still valid. Above the line the plot of $\mathcal{L}(f)$ is increasingly invalid and $S_\varnothing(f)$ must be used to represent the phase noise of the signal.

$\sigma_y(\tau)$ OR STANDARD DEVIATION OF FRACTIONAL FREQUENCY FLUCTUATIONS (TIME DOMAIN)



$$\sigma_y^2(\tau) = \text{Allan variance} \sim \frac{1}{2(M-1)} \sum_{K=1}^{M} (\bar{y}_{K+1} - \bar{y}_K)^2$$

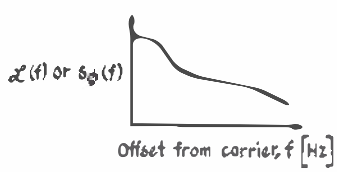
$\bar{y}$ = average $\frac{\Delta f}{f}$ over interval τ long

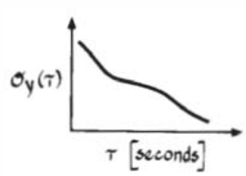
M = # of samples, τ = period of each sample

- Statistical measure

Frequency stability is also defined in the time domain with a sample variance known as the Allan variance. $\sigma_y(\tau)$ is the standard deviation of fractional frequency fluctuations $\Delta f/f_0$. A short τ will produce short-term information, while for a long τ, the short term instabilities will tend to average out and you will be left with longer-term information.

CONVERSION BETWEEN TIME AND FREQUENCY DOMAIN





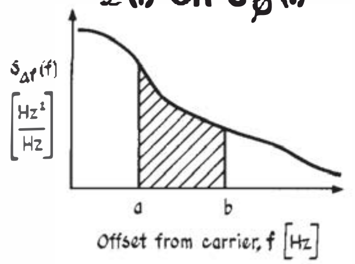
For example, for slope of $\mathcal{L}(f)$ as f^{-2}:

$$\sigma_y(\tau) = \frac{\sqrt{\mathcal{L}(f) f^2}}{f_0} \tau^{-1/2}$$

- Piecewise translatable
- Inverting axes

There are equations available to translate between the time and frequency domains. The translations apply to noise processes having particular slopes, and tend to reverse the independent variable axis. An example of one such equation for a slope of $\mathcal{L}(f)$ of f^{-2} is given here.

RELATING RESIDUAL FM TO $\mathcal{L}(f)$ OR $S_{\phi}(f)$



res FM = total rms frequency deviation within specified bandwidth

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

$$\text{res FM} = \int_a^b \sqrt{S_{\Delta f}(f)}$$

$$= \int_a^b \sqrt{\frac{f^2 S_{\phi}(f)}{B}} \; [\text{Hz}]$$

Residual FM is a familiar measure of frequency instability that is related to $S_{\phi}(f)$. Residual FM is the total rms frequency deviation with a specified bandwidth. Commonly used bandwidths are 50Hz to 3kHz, 300Hz to 3kHz, and 20Hz to 15kHz. Only the short-term frequency instability occuring at rates within the bandwidth is indicated. No information regarding the relative weighting of instability rates is conveyed. The presence of large spurious signals at frequencies near the frequency of the signal under test can greatly exaggerate the measured level of residual FM since the spurious signals are detected as FM sidebands.

II. Measurement Techniques on Sources

SEMINAR AGENDA

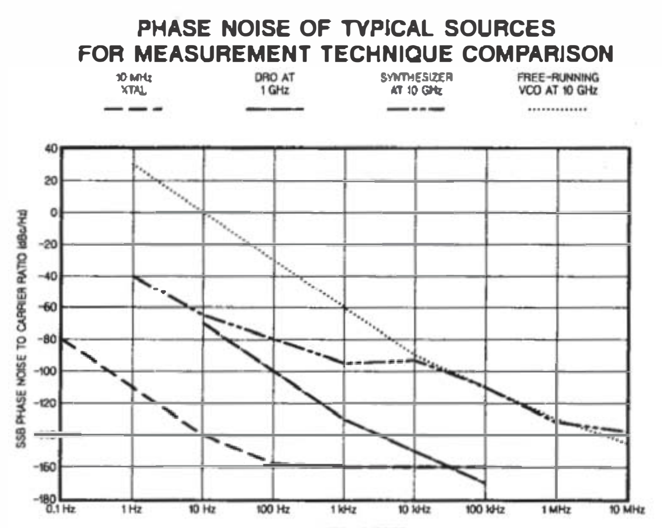
I. Basis of Phase Noise
- Why is Phase Noise important?
- What is Phase Noise?
- What causes Phase Noise?
- Quantifying Phase Noise

⇨ II. Measurement Techniques on Sources
1. Direct Spectrum Method
2. Heterodyne/Counter Method
3. Phase Detector Method
4. Frequency Discriminator Method
5. Summary of Source Measurement Techniques

III. Measurement of Two-Port Phase Noise (Devices)

IV. Phase Noise Measurement on Pulsed Carriers

V. AM Noise Measurement

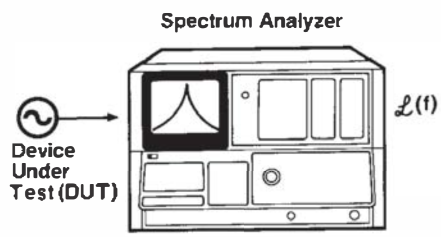


Here are four representative sources that will be used throughout the seminar in judging and comparing the capabilites of the four measurement methods.

1. Direct Spectrum Method

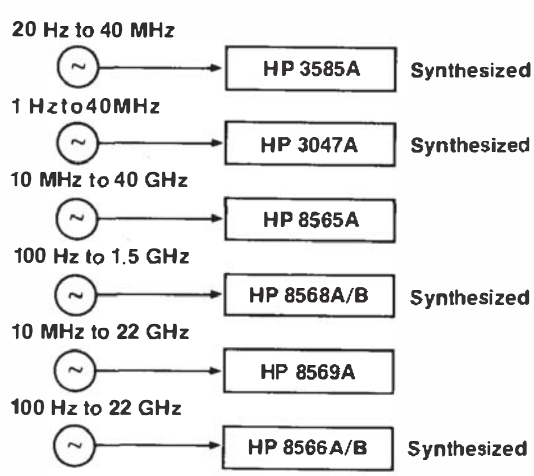
PHASE NOISE MEASUREMENT OF SOURCES

1. Direct Spectrum Analysis



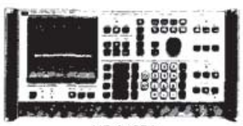
The simplest, easiest, and perhaps oldest method for phase noise analysis of sources is the direct spectrum technique. Here, the Device Under Test (DUT) is input into a spectrum analyzer tuned to the carrier frequency, directly measuring the power spectral density of the oscillator in terms of $\mathcal{L}$(f).

DIRECT SPECTRUM MEASUREMENT CHOICES



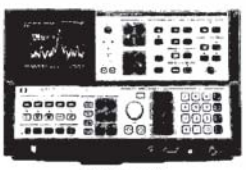
HP makes a number of high-quality spectrum analyzers that might be appropriate for direct spectrum analysis of sources. Analyzers covering from sub-Hertz to 22 GHz are available, to cover the frequency range of any DUT.

DIRECT SPECTRUM MEASUREMENT CHOICES



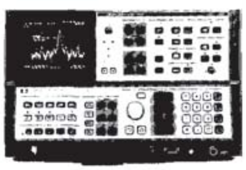
HP 3585A

- 20 Hz to 40 MHz
- 3 Hz RBW min. HP-IB
- −137 dBm to +30 dBm
- Synthesized LO



HP 8568A/B

- 100 Hz to 1.5 GHz
- 10 Hz RBW min. HP-IB
- −134 dBm to +30 dBm
- Synthesized LO



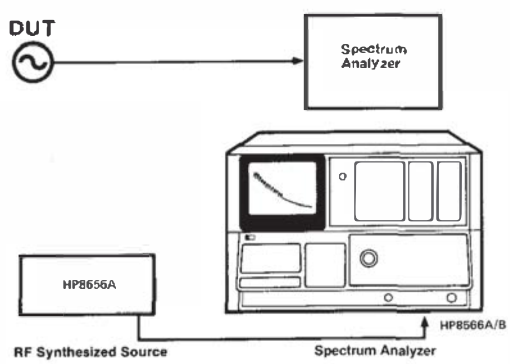
HP 8566A/B

- 100 Hz to 22 GHz
- 10 Hz RBW min. HP-IB
- −134 dBm to +30 dBm
- Synthesized LO

Of all the analyzers listed, the best choices for direct spectrum analysis are those listed here. These are all analog spectrum analyzers with synthesized LO's, and narrow resolution bandwidths.

As we will see later, spectrum analyzers with synthesized local oscillators offer the optimum performance for direct spectrum analysis of phase noise.

DEMONSTRATION



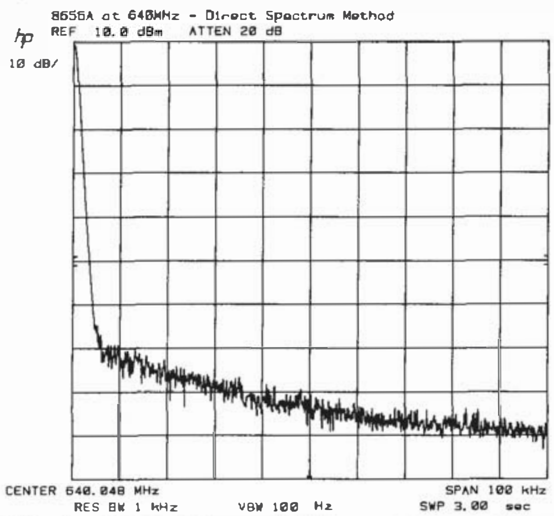
Direct Spectrum Method

This demonstration illustrates the measurement of the single sideband phase noise, $\mathcal{L}(f)$, of a source.

INTERPRETING THE RESULTS

$$\mathcal{L}(f) = \frac{P_{ssb}(f)}{P_s} \quad [dBc/Hz]$$

1. Measure carrier level P_s
2. Measure sideband level $P_{ssb}(f)$
3. Apply Corrections



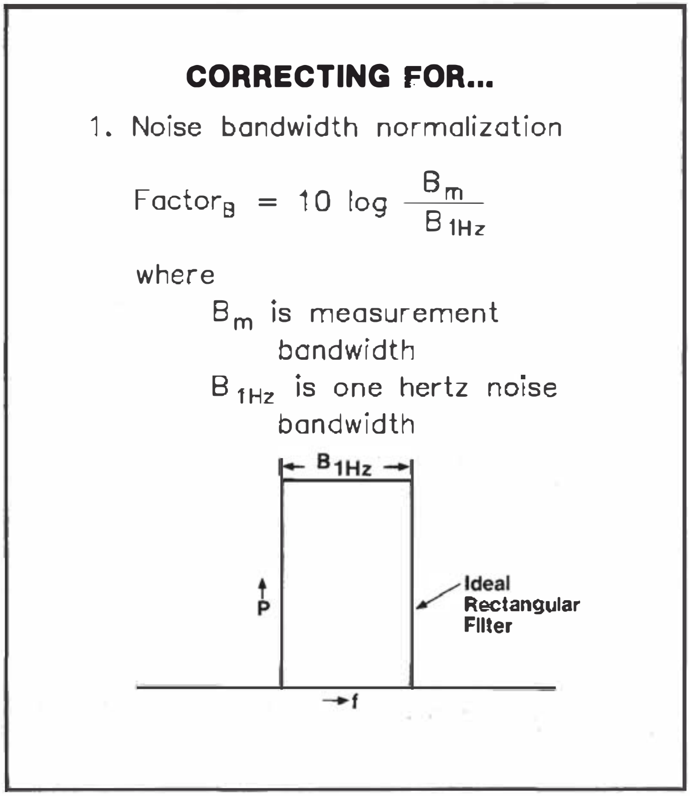
A typical spectrum analyzer display would look like this. What is $\mathcal{L}(f)$ at a 10 kHz offset for this test source?

Here, the power in the carrier P_s is read as +7 dBm. The marker reads −67 dBm, 10 kHz away from the carrier. P_{ssb} minus P_s is equal to −74 dBc. But is this equal to $\mathcal{L}(f)$?

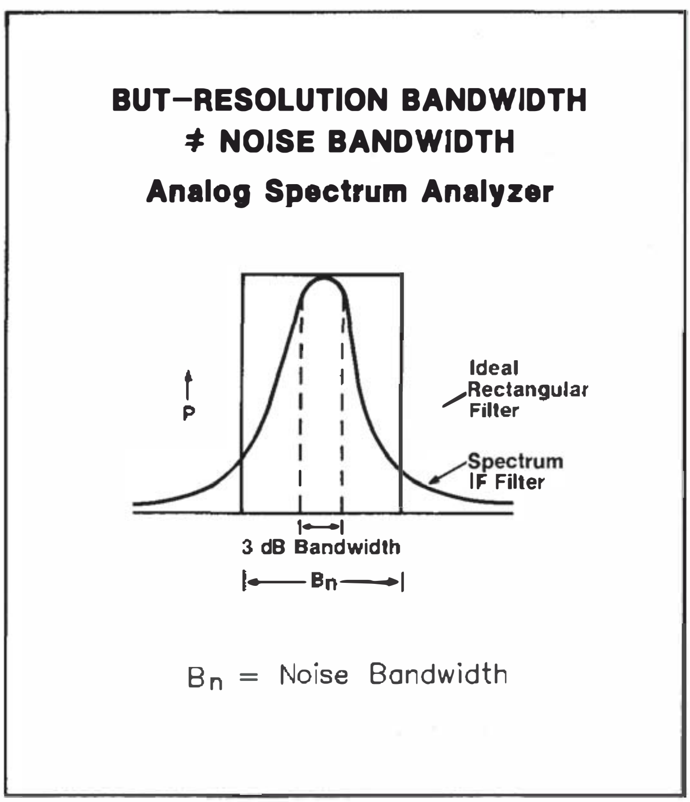
DIRECT SPECTRUM ANALYSIS
Correction Factors

1. Noise bandwidth normalization
2. Effect of spectrum analyzer circuitry

In the direct spectrum technique, there are two correction factors that must be used on our value for P_{ssb} minus P_s in order to yield $\mathcal{L}(f)$. First, $\mathcal{L}(f)$ is specified in a one hertz bandwidth, and our measurement bandwidth on the previous slide was not one hertz. Typically, in the direct spectrum technique, it's impossible (and very impractical in terms of time) to make measurements in a one hertz bandwidth. We also must be careful that our measurement circuitry does not effect the quantity we are trying to measure, and that it measures it accurately.



Normalizing to a 1 Hz noise bandwidth from a given measurement bandwidth is a simple matter, involving a simple power relationship. Looking back two slides, what is our measurement bandwidth?



But $\mathcal{L}(f)$ requires us to normalize to an equivalent 1 Hz noise bandwidth. A spectrum analyzer's 3 dB resolution bandwidth is not its equivalent noise bandwidth. (Noise is defined as any signal which has its energy present over a frequency band significantly wider than the spectrum analyzer resolution BW; ie, any signal where individual spectral components are not resolved.) The noise bandwidth is defined as the bandwidth of an ideal rectangular filter having the same *power response* as the actual IF filter. How do we find this equivalent noise bandwidth?

FINDING EQUIVALENT NOISE BANDWIDTH (B_n)

1. First approximation...

   Noise $B_n \cong 1.2 \times$ nominal $B_{3\ dB}$

2. Second approximation...

   Noise $B_n \cong 1.2 \times$ measured $B_{3\ dB}$

3. Most accurate...

   Measure equivalent B_n

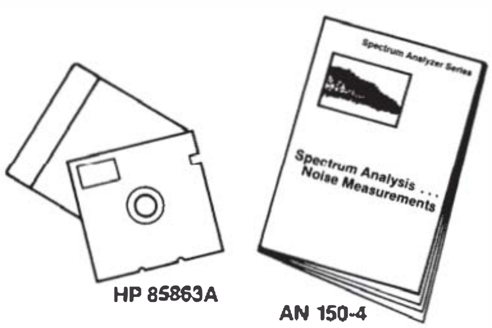
Depending on the desired accuracy, there are three ways to derive the equivalent noise bandwidth. In order of increasing accuracy . . .

For a first approximation, most Hewlett-Packard spectrum analyzers have a noise bandwidth approximately 1.2 times the nominal 3 dB resolution bandwidth setting. For our example, our correction factor for a nominal 300 Hz resolution bandwidth setting is 25.6 dB.

For a more accurate estimate of the correction factor to equivalent noise bandwidth, you could measure the actual resolution bandwidth of the spectrum analyzer, using an accurate signal source and using the analyzer in zero scan. (A typical synthesized HP spectrum analyzer has a specified 3 dB IF bandwidth accuracy of ±10% to ±20%.)

Of course, the most accurate way to determine equivalent noise bandwidth is to carefully measure the actual power response of the IF bandwidth.

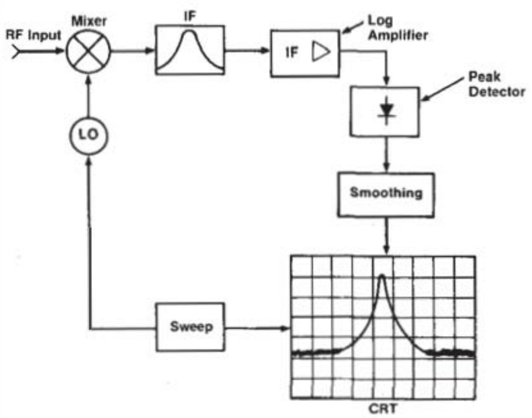
FINDING EQUIVALENT NOISE BANDWIDTH



Details on the effect of a spectrum analyzer on noise, including how to measure noise equivalent bandwidth, can be found in AN 150-4. Or, for a simpler solution, there is available software for some spectrum analyzers to automatically measure equivalent noise bandwidth.

CORRECTING FOR...

2. Response of Analog S/A to Random Noise



Actual Noise = Measured Noise + 2.5 dB

In addition to the noise power bandwidth correction, another correction is needed when measuring random noise as the spectrum analyzer's detection circuitry is calibrated for accuracy in measuring discrete signals (sinusoids). In most analog spectrum analyzers, there is a logarithmic IF amplifer followed by a peak detector. A peak or envelope detector used to measure random noise results in a reading lower than the true rms value of the average noise (typically about 1.05 dB lower). The log shaping tends to amplify noise peaks less than the rest of the noise signal, resulting in a detected signal which is smaller than its true rms value. The correction for the log display mode combined with the detector characteristics gives a total correction for HP analyzers of 2.5 dB, which should be *added* to any random noise measured in the log display.

TYPICAL MEASUREMENT RESULTS

$$\mathcal{L}(f) = \frac{P_{ssb}(f)}{P_s} \quad [\text{dBc/Hz}]$$

$$\mathcal{L}(f) = P_m - (\text{Factor}_B) + C_n - P_s$$

	10 kHz		100 kHz
P_m = Measured sideband level	___	dBm	___
B_{3dB} = Resolution bandwidth ___ Hz			
Noise power bandwidth			
$= B_n = B_{3\,dB} \times 1.2 =$ ___ Hz			
$\text{Factor}_B = 10 \text{ Log } (B_n) =$	___	dB	___
C_n = Random signal correction	2.5	dB	2.5
P_s = Carrier level	___	dBm	___
$\mathcal{L}(f) =$	___	$\frac{\text{dBc}}{\text{Hz}}$	___

DIRECT SPECTRUM MEASUREMENTS

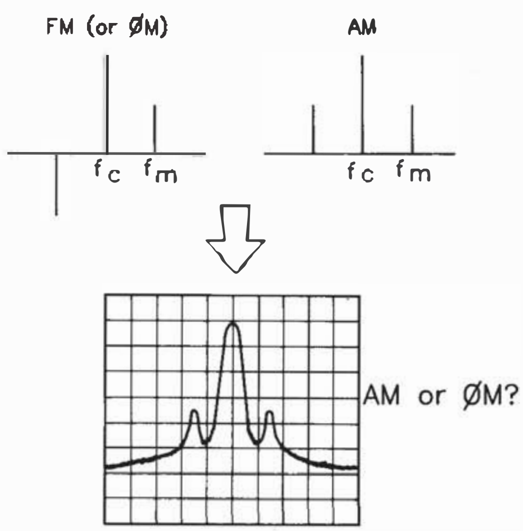
AREAS OF CAUTION

1. AM << ØM
2. Range and resolution
3. Sweep time

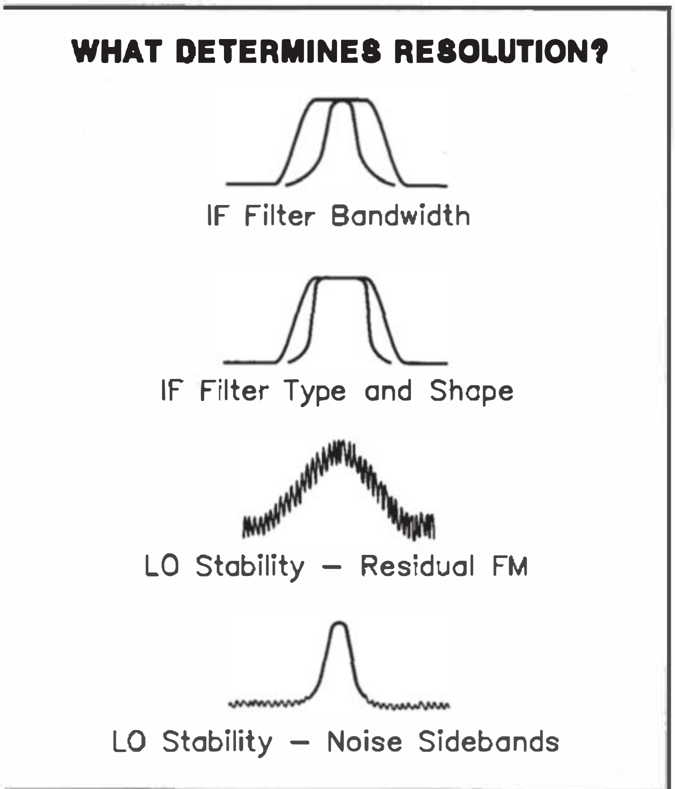
That was so easy there's got to be a catch. The direct spectrum technique is indeed the simplest of all phase noise measurement techniques, but is has several limitations as listed here.

DIRECT SPECTRUM MEASUREMENT LIMITATIONS

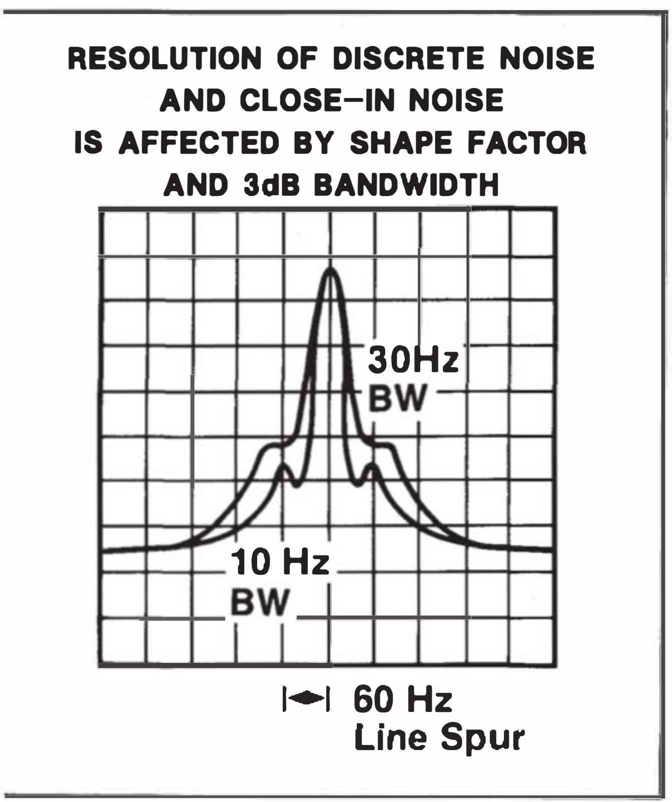
Cannot separate AM and ØM



A spectrum analyzer cannot distinguish amplitude modulation from phase modulation or frequency modulation. The definition of $\mathcal{L}(f)$ is the power measured in one *phase modulation* sidehand divided by the total carrier power. In order for a valid number for $\mathcal{L}(f)$, the sideband power measured must be phase modulation sidebands only. If the source has high AM noise, this condition will not be met. Thus, for accurate phase noise measurements using the direct spectrum method, the source's AM noise must be <<ØM noise.

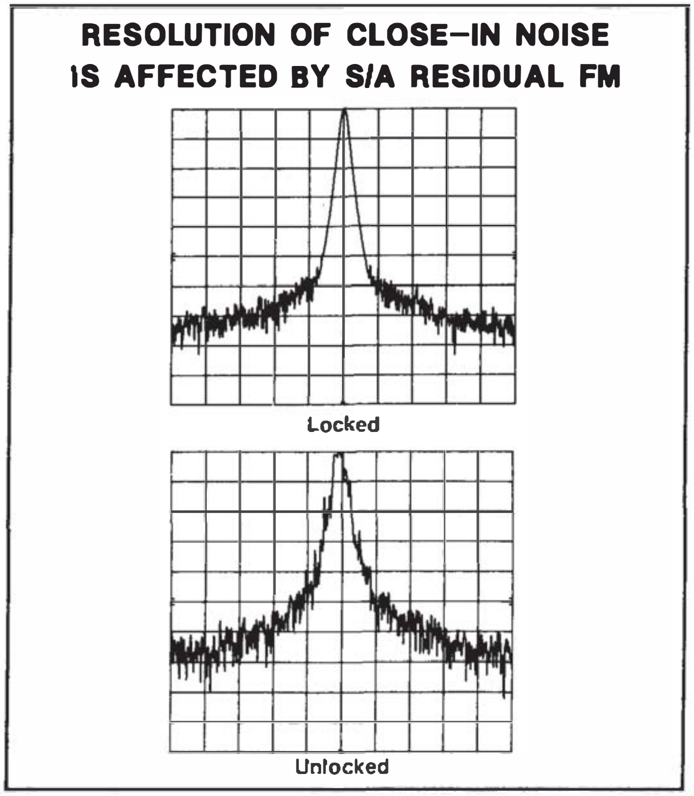


The primary limitations to the direct spectrum method are related to the spectrum analyzer's dynamic range and resolution. From the spectrum analyzer simplified block diagram, what elements of the spectrum analyzer would affect range and resolution?

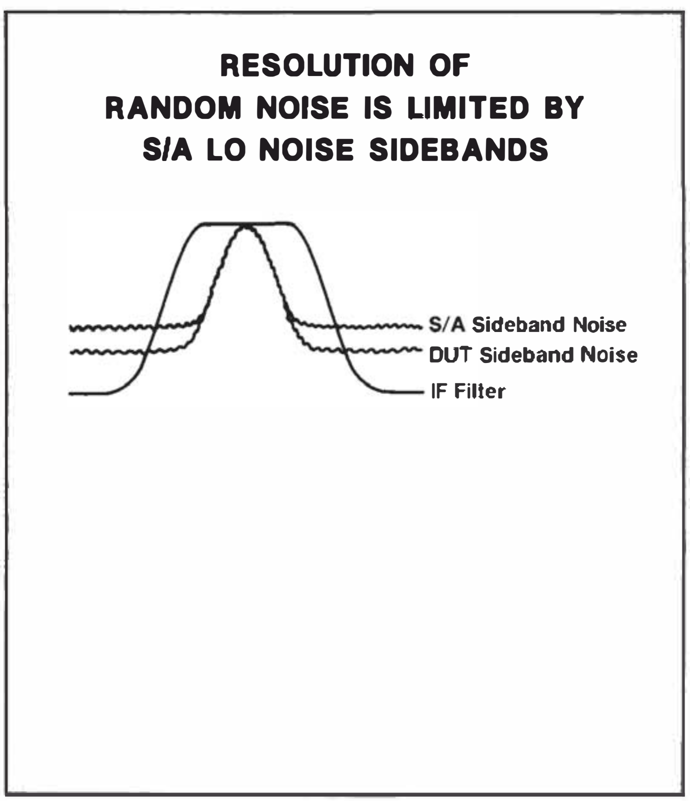


In measuring random noise, the 3 dB BW and shape factor can limit the resolution of discrete noise. As an example, a typical source hasline spurious at 60 and 120 Hz. If the spectrum analyzer has a minimum resolution BW of 300 Hz, these discrete signals will not be resolved, but instead will appear incorrectly as a high level of close-in noise.

Even with a narrow resolution bandwidth, the shape factor of the spectrum analyzer will limit the resolution of close-in noise, or any noise close to a high level spur. For example, let's say we chose a 10 Hz RES BW, with a shape factor of 11:1. If the source's phase noise 100 Hz away from the carrier was down − 100 dBc (a very reasonable number for an RF carrier), the filter width 60 dB down is 110 Hz, and the noise will be hidden under the skirt of the response to the carrier.



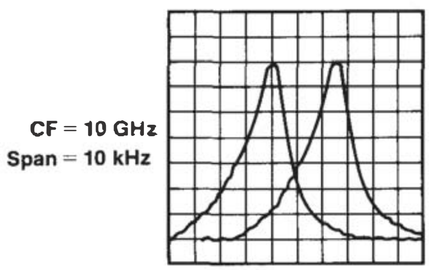
The other factor in resolution is related to the frequency stability of the analyzer's local oscillator. As a first measure of frequency stability, we can look at the LO's residual FM. It would be inappropriate to allow an analyzer to have a resolution BW narrow enough to observe the analyzer's own instability. This means that the analyzer's residual FM dictates the minimum resolution bandwidth allowable. (Which, in turn, determines the minimum spacing of equal amplitude signals.) In general, only spectrum analyzers with synthesized local oscillators are very useful for direct spectrum analysis of phase noise.



We already saw that residual FM only gives a general feel for the level of a signal's instability. Even after stabilization techniques of the S/A local oscillator minimizes residual FM, there is still sideband noise associated with the LO. The noise characteristic of the LO is transferred to all incoming signals onto the IF. The phase noise on the local oscillator can mask the noise on the DUT that we might otherwise be able to see, if we considered only the 3 dB bandwidth and the shape factor. This sideband noise gives the better shape factor of the square-topped filter less actual value in resolving close-in or low level signals.

DIRECT SPECTRUM MEASUREMENT LIMITATIONS

- Cannot measure close-in to a drifting source

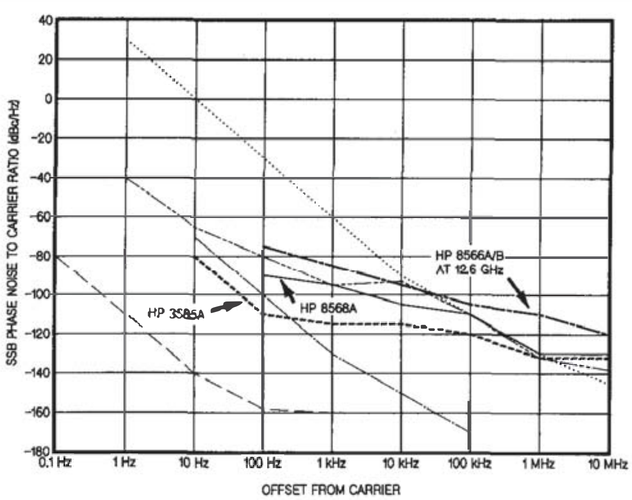


weeptime (10 KHz span, 100 Hz $B_{3\,dB}$) = 3 sec.

Direct Spectrum Analysis is useful for measuring sources with relatively high phase noise, but the source must be fairly stable (long-term wise). That is, the direct spectrum technique cannot measure close-in to a drifting source. The closer-in you desire to look, the narrower the resolution bandwidth, and the longer the sweeptime. On a free-running source with even fairly good drift characteristics, this drift during the sweep on the spectrum analyzer will yield incorrect phase noise data.

For example, a good cavity oscillator at 10 GHz will typically drift 30 kHz/min. Spanning as much as 50 kHz (for an offset frequency >10 kHz from the carrier) with a 100 Hz resolution bandwidth will typically take at least 20 seconds. By that time, the DUT will have drifted 10 kHz!

NOISE FLOORS OF SEVERAL HP SPECTRUM ANALYZERS



Here are some typically obtainable spectrum analyzer noise floors as a function of frequency, as well as the noise of some typical sources. As you can see, the direct spectrum technique isn't extremely sensitive, it is most useful for measuring stable sources with relatively high phase noise at low offsets.

DIRECT SPECTRUM METHOD

ADVANTAGES

- Very flexible
- Simple to use
- Displays $\mathcal{L}(f)$ directly
- Accurately displays discrete signals simultaneously

DISADVANTAGES

- Can't measure sources with high AM
- Can't measure low noise sources
- Can't measure close-in on a drifting source

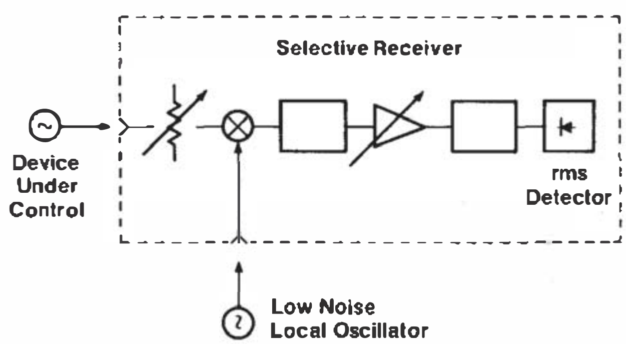
Direct Spectrum Analysis is optimum for measuring a source with relatively high noise and low drift.

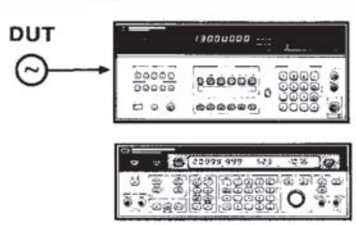
In summary, the direct spectrum technique for measuring the phase noise on a carrier is an easy, simple technique. It displays $\mathcal{L}$ (f) directly, and also accurately displays the discrete signals simultaneously. Unfortunately, it can't be used to measure very clean (spectrally pure) sources, nor can it measure 'noisier' sources that have either high AM or drift. It is however, perfect for measuring the phase noise on a stable (typically synthesized or phase-locked) source with relatively high noise sidebands.

DIRECT SPECTRUM METHOD

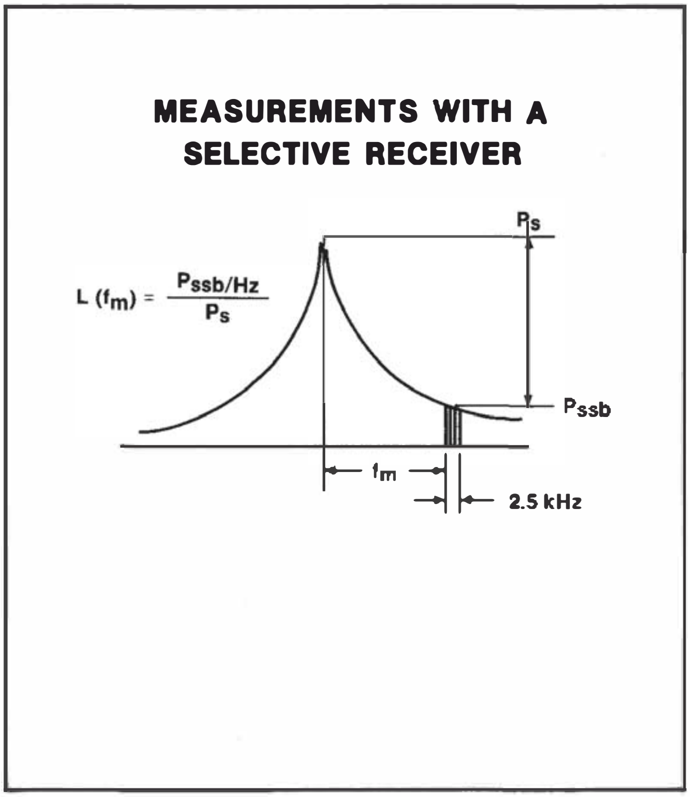
Selective Receiver

HP 8901B





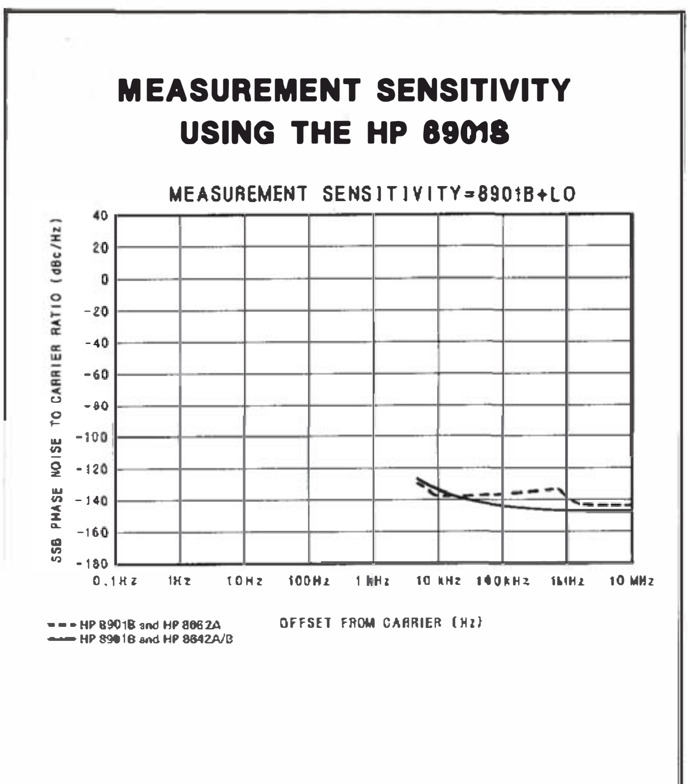
To improve measurement sensitivity an alternate direct spectrum technique can be used which consists of a selective receiver and a low phase noise local oscillator (LO). HP offers a very selective receiver, the HP 8901B Modulation Analyzer (with Option 030 High Selectivity) and a number of low noise signal generators which can be used as the external LO.



The measurement procedure for making phase noise measurements with a selective receiver is very simple. You just make selective power measurements at the carrier frequency and at the desired offset from the carrier, calculate the difference between the readings and convert the result to a 1 Hz bandwidth.

The 8901B with the high selectivity option (Option 030) makes selective power measurements from 10 MHz to 1300 MHz, over 115 dB dynamic range with over 90 dB of rejection and ±0.5 dB accuracy. The analyzer's measurement bandwidth is 2.5 kHz.

The 8901B's measurement performance can be easily extended up to 18 GHz or 26.5 GHz with either the HP 11793A Microwave Converter and a microwave local oscillator, or the HP 11729B Carrier Noise Test Set and HP 8662A or HP 8663A Synthesized Signal Generator.



The measurement sensitivity of the 8901B Modulation Analyzer with an external LO is very good. The noise floor of the analyzer is less than −150 dBc/Hz. You can see on the plot that for offsets less than 10 kHz the HP 8901B with the HP 8662A/8663A provides the lowest phase noise performance and for offsets greater than 10 kHz the HP 8901B with the HP 8642A/B provides the most measurement sensitivity.

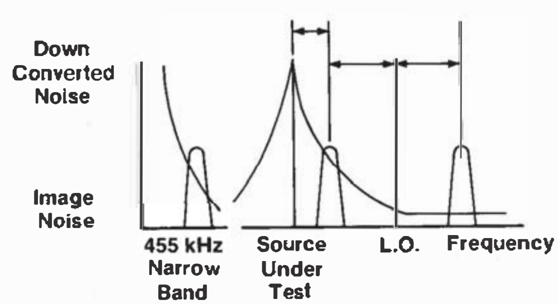
SIMPLE MEASUREMENT PROCEDURE

1. Measure carrier level P_S
2. Set a reference
2. Measure sideband level P_{SSB}
4. Display results in dBc or dBc/Hz
5. Apply image correction

TOTAL TEST TIME < 10 SECONDS

The 8901B modulation analyzer will measure and display the side-band level in seconds. The 8901B's effective noise bandwidth (34 dB for 2.5 kHz filter) can be corrected for by the instrument, providing direct readings in dBc/Hz if desired. Because the analyzer does not have a tuned front end, a correction must be made for the image signal.

IMAGE CORRECTION



When the external LO is set to a given frequency, the analyzer converts the desired and image signals into the IF passband. The resulting signal in the passband is the sum of these two signals. When making phase noise measurements, if the noise at the image frequency (910 kHz away) is of comparable amplitude to the noise at the offset of interest, the image noise will sum into the IF and the result will be affected. This typically happens at large offsets where white thermal noise predominates, and the image adds 3 dB to the result. To calculate the image correction either use the C_I plot or use the C_I equation.

$$C_I = -10 \, \mathrm{Log}_{10}[10^{N_I/10} + 1]$$

where

C_I = Image correction

$$N_I = \frac{\text{Image noise level}}{\text{offset noise level}} \text{ dB}$$

TYPICAL MEASUREMENT RESULTS

$$\mathcal{L}(f) = \frac{P_{SSB}}{P_S} \text{ [dBc/Hz]}$$

$$\mathcal{L}(f) = P_M + C_I$$

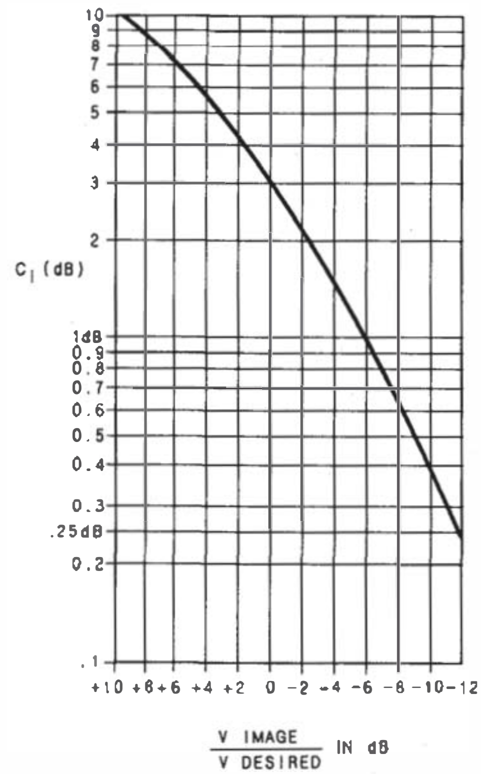
where

P_M = Measured sideband level relative to the carrier converted to dBc/Hz (the analyzer will display this value) ______ dBc/Hz

C_I = Image Correction ______

$\mathcal{L}(f)$ = ______ dBc/Hz

Image Correction



DIRECT SPECTRUM MEASUREMENTS USING A SELECTIVE RECEIVER

Areas of Caution

1. AM << ØM
2. Offset
3. Resolution

1. The 8901B Modulation Analyzer measures the total noise power in the measurement bandwidth and therefore cannot distinguish between phase noise and AM noise.

2. Measurements as close as 5 kHz from the carrier can be made before IF feedthrough begins to saturate the IF amplifiers and the rms detector.

3. In addition, any discrete signals such as spurious sidebands, may be interpreted incorrectly by this method. This is because the filter noise bandwidth correction factor (34 dB for 2.5 kHz filter) applies only to distributed noise throughout the filter passband. Discrete signals therefore will be measured as 34 dB lower than the actual value.

DIRECT SPECTRUM METHOD USING SELECTIVE RECEIVER

ADVANTAGES

- Fast/Easy
- Low noise floor
- Displays $\mathcal{L}(f)$ directly
- Easily automated
- Moderate cost

DISADVANTAGES

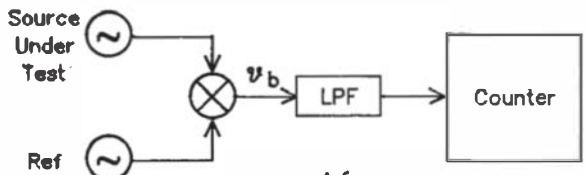
- Can't measure sources with high AM
- 2.5 kHz resolution
- Offsets > 5 kHz
- Correct for image noise

In summary, the direct spectrum technique using a selective receiver is fast and easy to use. There is no need to phase-lock signals. The data is displayed conveniently, either in dBc or dBc in a 1 Hz bandwidth. Noise at a given offset is measured and displayed in real-time (measurement rate = 5 readings/second). The analyzer only requires an exteral LO and testing can be fully automated using an external controller. Although the analyzer can not separate AM and PHiM noise, total SSB noise, regardless of type, is often what is important when analyzing communication systems.

2. Heterodyne/Counter Method

PHASE NOISE MEASUREMENT OF SOURCES

2. Heterodyne/Counter method



$$y(t) = \frac{\Delta f_o}{f_o}$$

$$\sigma^2_y(\tau) = \left\langle \frac{(\bar{y}_{k+1} - \bar{y}_k)^2}{2} \right\rangle$$

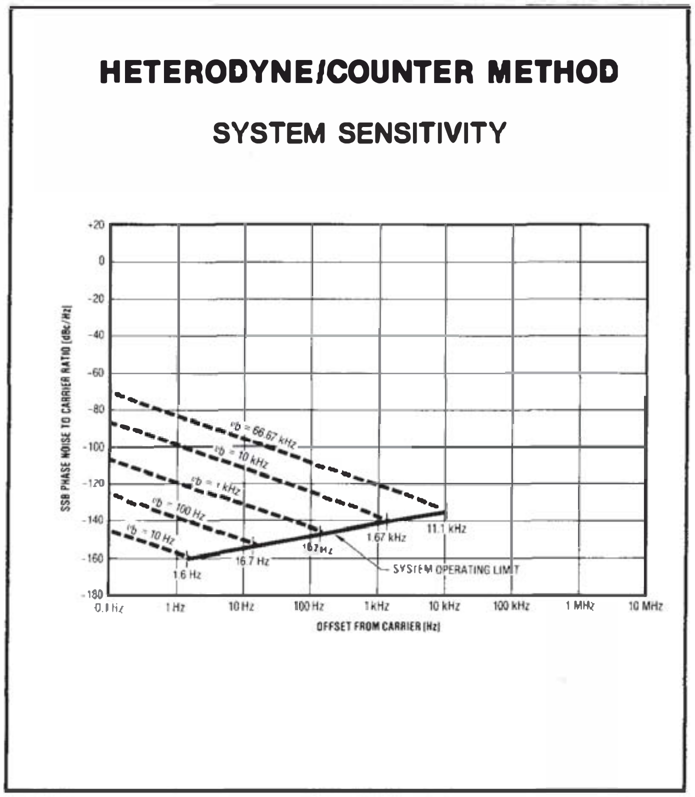
- Statistical measure of source instability

The second technique for making short-term stability measurements on sources is commonly called the 'time-domain' technique (or heterodyne frequency counter technique). The Device Under Test and the Reference are downconverted with a mixer to a low IF frequency, or beat frequency, ν_b. Then a high resolution counter repeatedly counts the IF signal, with the time period between each measurement held constant. This allows several calculations of the fractional frequency difference, y, over the time period used. A special variance of these differences, called the Allan variance (after its inventor), can then be calculated. The square root of this variance is called $\sigma(\tau)$, where τ is the time period used in the measurement. The whole process is generally repeated for several different time periods, or τ, and $\sigma(\tau)$ is plotted vs. τ as an indication of the signal's short-term frequency stability. Note that $\sigma(\tau)$ is a statistical measure of many samples.

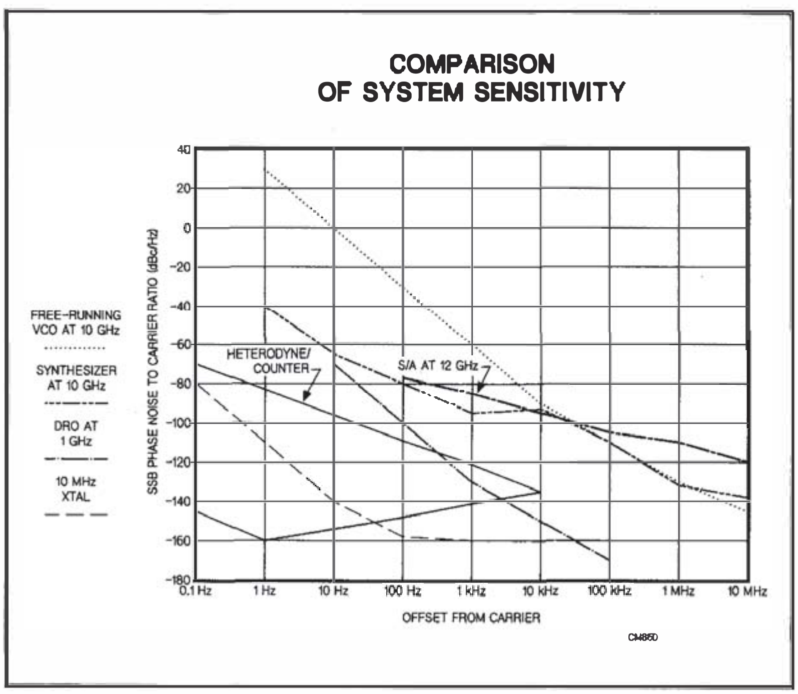
HETERODYNE/COUNTER METHOD MEASUREMENT CONSIDERATIONS

- Sources must be offset
- Source drift << V_b
- Requires low noise reference
- Offsets < 10KHz
- Noise must fall rapidly with increasing fm

Note that because it is really only the IF frequency that is effected by the instrumentation, the time-domain technique is essentially a very broad-band method, easily usable to 18 GHz. There are, however, several important limitations of this method. One, the sources must be offsettable, with good resolution to set up a fairly low frequency beat note. Second, the sources must be fairly stable. In order for the measured beat note frequency to be indicative of the short-term stability of the DUT, the long-term stability must be small relative to the beat note frequency. Third, since the mixer measures the combined short-term instability of the DUT and the reference, the reference must be lower in noise than the DUT. Fourth, the effective offset frequencies that the time domain technique can measure are limited by the dead time and sampling speed of the counter. In common implementations, this limits the measurement to offset frequencies <10 kHz from the carrier. Finally, the time domain technique always gives an accurate measure of $\sigma(\tau)$. However, for an accurate measure of phase noise, the noise of the DUT must be falling rapidly as a function of offset frequency. We'll look at this limitation more closely in a moment.

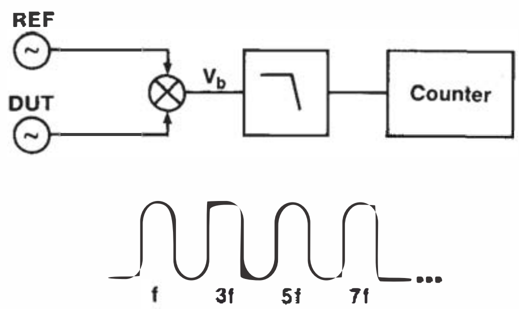


The time domain technique is a very sensitive measurement method for measuring close-in phase noise. It was the first measurement technique that was usable for measuring at equivalent offset frequencies of less than 1 Hz from the carrier (it's typically usable to an offset frequency of 0.01 Hz!). Because of its inherent reading in the time domain, this technique is very useful where the application problem is best thought of in the time-domain, or where very close-in noise is critical (such as measurements on time standards for navigation systems or for digital communications applications).



For comparison, the sensitivity of the direct sp technique and the time domain technique are co here. The contribution of the time-domain techni close-in measurements is easily seen.

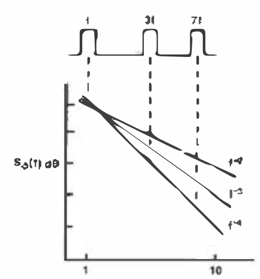
HETERODYNE/COUNTER METHOD
LIMITATIONS FOR MEASURING SYNTHESIZED SOURCES



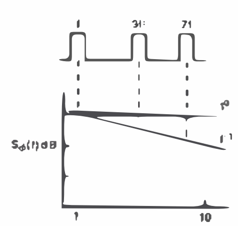
Sampling the counter in a prescribed manner results in a digital filtering process.

Returning to the limitations of the time-domain method, remember that it requires sources to be very stable (ie, primary or secondary standards or sources locked to standards). However, it also has certain limitations for measuring synthesizers. This arises from the fact that sampling the counter in a prescribed manner results in a digital filtering process on the noise of the DUT. A digital filter, as well as having the filtering effect at f, also has the equivalent filter response at 3f, 5f, 7f, . . .

AN INFINITE DIGITAL FILTER INTEGRATES IN NOISE AT OTHER OFFSETS



offset from carrier, f (Hz)



offset from carrier, f (Hz)

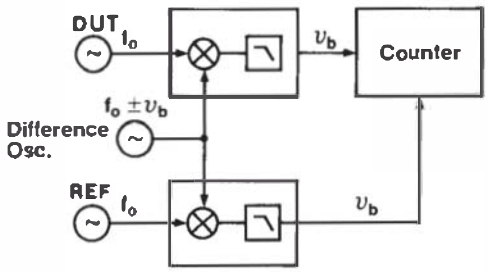
This infinite digital filter integrates in the noise at other offset frequencies. If the noise is falling as f^{-4}, f^{-3}, or f^{-2}, the integrated noise from the infinite digital filter is not significant. The total noise reading will be higher, but within normal error limits.

However, if the noise is falling only as f^{-1} or f^{0}, this noise contribution from the other responses of the filter will indeed contribute significantly to the measured value, resulting in an unacceptable error.

In most designs of synthesizers, the 'synthesizer knee' usually occurs at offsets >10 kHz from the carrier, the useable range of the time domain technique. However, care should be taken if the bandwidths of the synthesized DUT result in f^{-1} or f^{0} noise components at offsets <10 kHz.

HETERODYNE/COUNTER METHOD

DUAL MIXER TIME DIFFERENCE (FOR NON-OFFSETTABLE SOURCES)



- Reference must have equal or lower noise than DUT
- Difference oscillator must have lower f^0 noise

For non-offsettable sources (such as primary frequency standards), there is also a dual-mixer time difference configuration for the time-domain method. Here, a common difference oscillator produces two beat signals, which are statistically sampled and compared. In this configuration, the reference must still have equal or lowe noise than the DUT. However, due to the common nature of the difference oscillator, most of its noise is correlated out, except for any f^0 noise component.

HETERODYNE/COUNTER METHOD

ADVANTAGES

- wide input frequency range
- good sensitivity close-to-carrier

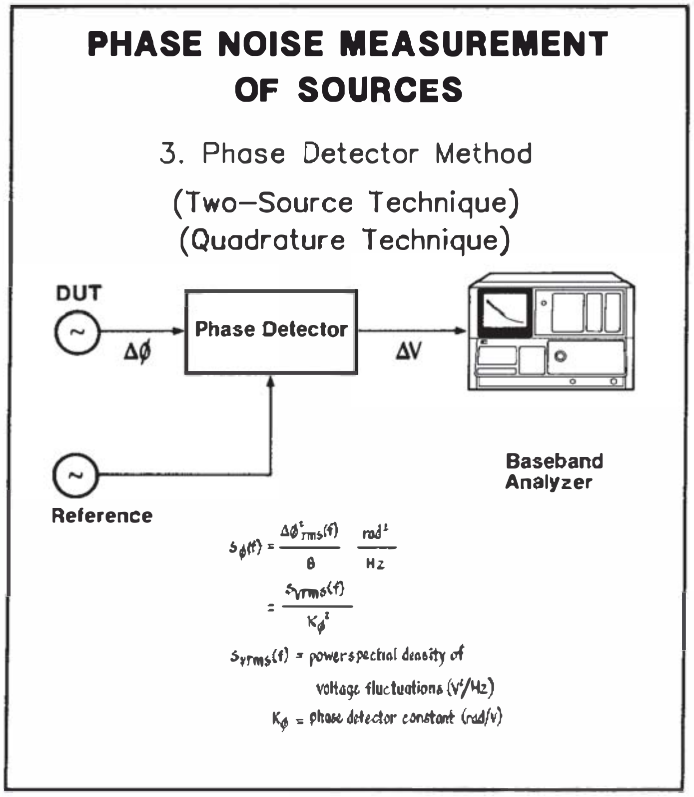
DISADVANTAGES

- requires two sources
- sources must be offset
- can't measure at offsets >10 kHz
- limitations for measuring synthesized sources

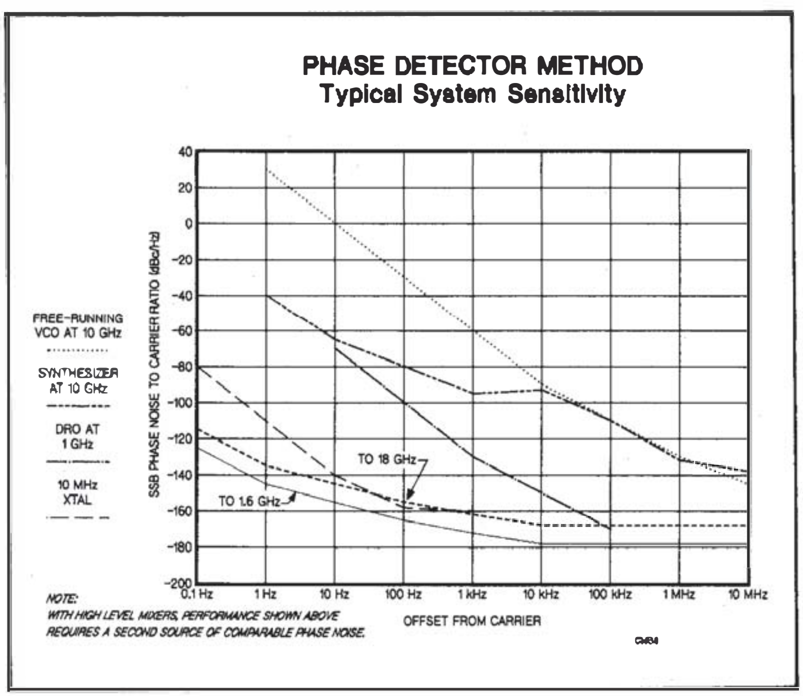
In summary, the time domain technique is a very sensitive measurement method for close-to-the-carrier analysis of signals from 5 MHz to 18 GHz. Its disadvantages have made it useful primarily for measuring frequency standards.

Notes

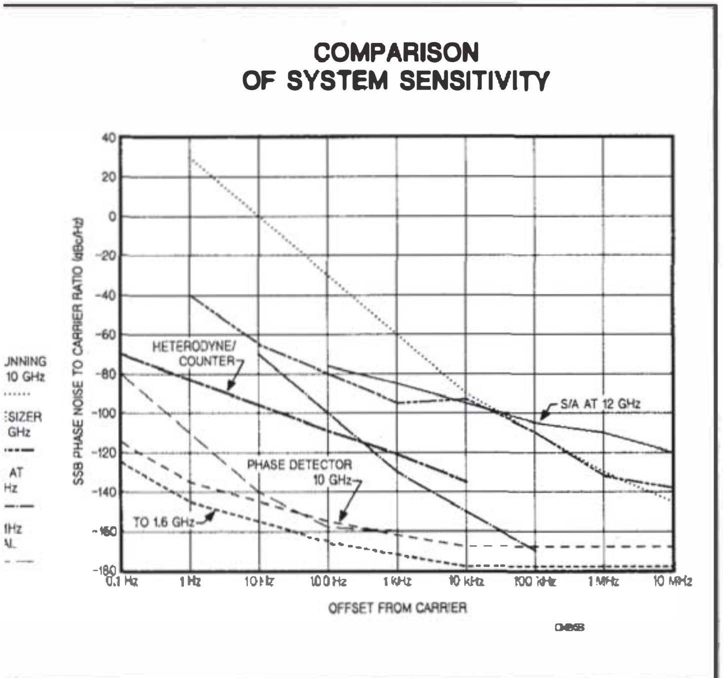
3. Phase Detector Method



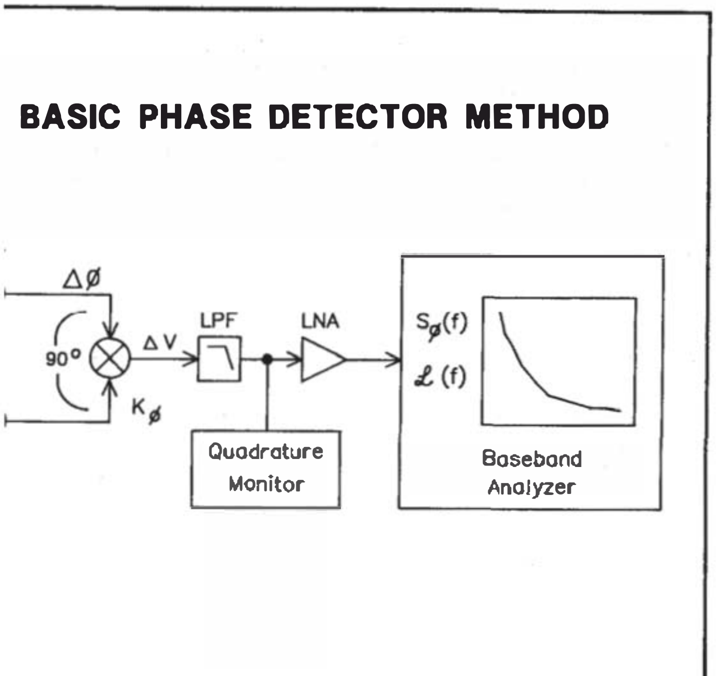
The third method of phase noise measurement is commonly called the phase detector method (also called the two-source technique or the quadrature technique). This technique is becoming increasingly popular, as we will see that it is over-all the most sensitive and broadband technique.



The sensitivity of the phase detector method is s here. The phase detector method provides good se over the entire offset frequency range; it can be u measure high quality standards (good close-in noi state-of-the-art free-running oscillators (low broa noise). The sensitivity shown can be obtained wit level mixers; note that typically using a lower fre mixer yields higher sensitivity. We will see later another component in the system is often the real factor in a real-life application.



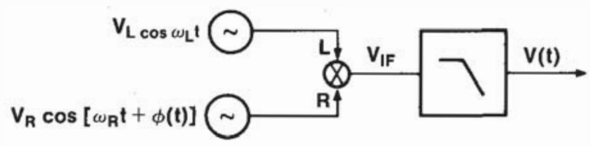
We can now compare the sensitivity of three of our measurement methods. Notice that the phase detector has the overall lowest noise floor. For this reason, with new implementations, it is enjoying increased use.



The basic phase detector method is shown here. At the heart of this method is the phase detector. Two sources, at the same frequency and in phase quadrature, are input to a double-balanced mixer used as a phase detector. The mixer sum frequency ($2f_0$) is filtered off, and the difference frequency is 0 Hz with an average voltage output of 0 V. Riding on this dc signal are ac voltage fluctuations proportional to the combined phase noise of the two sources. The baseband signal is often amplified, and input to a baseband spectrum analyzer. In practice, a scope or dc voltmeter is used as a quadrature monitor.

In the phase detector method, the mixer selection is important to the overall system performance. The mixer must cover the frequency range of the DUT, and the IF port must be dc coupled and should have a flat frequency response. Since the noise floor sensitivity is related to the mixer input levels, high level mixers yield better performance. However, be careful to match mixer drive requirements to available source power. Before going on with the method itself, we first need to understand the operation of a mixer as a phase detector.

DOUBLE–BALANCED MIXER USED AS A PHASE DETECTOR



$$V_{IF}(t) = K_L V_R \cos\left[(\omega_R - \omega_L)t + \phi(t)\right] + K_L V_R \cos\left[(\omega_R + \omega_L)t + \phi(t)\right] + \ldots$$

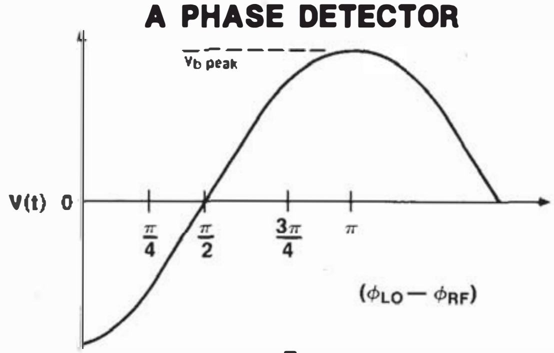
$$V(t) = K_L V_R \cos\left[(\omega_R - \omega_L)t + \phi(t)\right]$$

where

K_L = mixer efficiency

Two signals are input to a double balanced mixer. The phase term, ∅(t), represents any phase fluctuation that is not common to both signals. The IF output of the mixer will be the sum and difference frequencies; the sum frequency is filtered off, leaving the difference frequency. (The sum frequency is filtered off to prevent this large signal from overloading the LNA or baseband analyzer in the system. The filter also provides a crude termination for the mixer so that instead of just reflecting the sum frequency at the mixer output it tends to terminate the sum at the mixer output.)

DOUBLE–BALANCED MIXER USED AS A PHASE DETECTOR



$$V(t) = K_L V_R \cos\left[(\omega_R - \omega_L)t + \phi(t)\right]$$

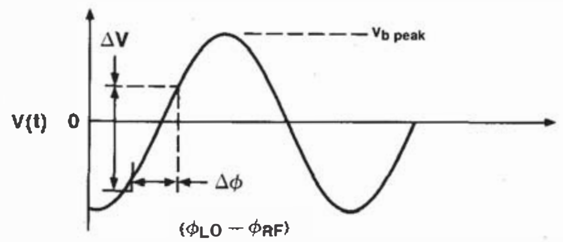
$$\text{if } \omega_R = \omega_L$$

$$V(t) = K_L V_R \cos\left[\phi(t)\right]$$

$$= V_{bpeak} \cos\left[\phi(t)\right]$$

Let the peak amplitude of V(t) be defined as V_{bpeak} (peak voltage of the beat signal, equal to $K_L V_r$). Now, to operate a mixer as a phase detector, the two input signals must be at the same frequency. This yields an average difference frequency of 0 Hz, and the mixer outputs a signal proportional to the cosine of the phase difference between the two signals.

DOUBLE-BALANCED MIXER USED AS A PHASE DETECTOR



$$V(t) = V_{bpeak} \cos[\phi(t)]$$

$$\text{let } \phi(t) = (2K+1)90^\circ + \Delta\phi(t)$$

$$\Delta V(t) = \pm V_{bpeak} \sin \Delta\phi(t)$$

for $\Delta\phi_{peak} < \cong 0.2$ radian, $\sin \Delta\phi(t) \cong \Delta\phi(t)$

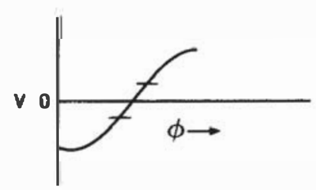
$$\Delta V(t) = \pm V_{bpeak} \Delta\phi(t)$$

$$\Delta V(t) = K_\phi \Delta\phi(t)$$

K_ϕ = phase detector constant (volts/radian) = V_{bpeak}

If we also force the two input signals to be nominally 90° out-of-phase (phase quadrature), the instantaneous voltage fluctuations out of the mixer (phase detector) will be directly proportional to the input instantaneous phase difference. For small phase deviations, we operate the mixer only in the linear region around 0 V dc, and the output voltage is related to the phase difference between the two input signals by a constant, K_ϕ, called the phase detector constant, in Volts/radian.

IMPORTANCE OF QUADRATURE



$$\text{error(dB)} = 20 \log\left[\cos(\text{magnitude of phase deviation from quadrature})\right]$$

or

allowable deviation from $0V_{dc}$ =

$$K_\phi \sqrt{1 - 10^{\frac{\text{error(dB)}}{10}}}$$

Since one of our assumptions in operating this system is that we are operating near the "zero" crossing of the phase detector, it is important that the input sources to the phase detector stay in good quadrature. Deviation from quadrature will result in an error as shown here.

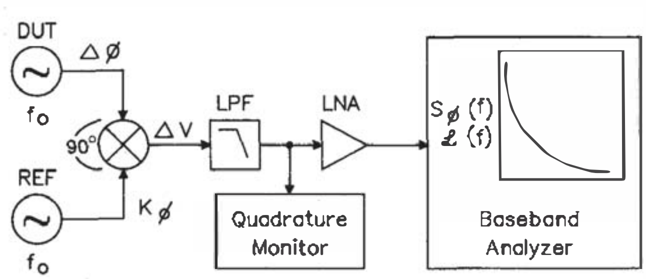
PHASE DETECTOR METHOD— PROCEDURE

1. Set–Up
2. Calibrate
3. Establish Quadrature and Measure
4. Corrections

The phase detector method measures voltage fluctuations directly proportional to the combined phase fluctuations of the two input sources. To best understand the phase detector method, and how system noise floor is achieved, it is useful to go through the actual measurement procedure. There are basically four steps.

BASIC PHASE DETECTOR METHOD

1. Set–up

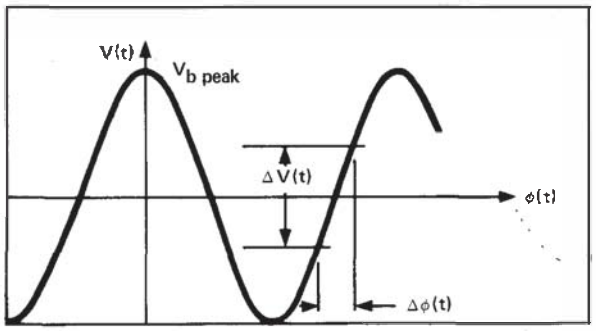


This is the "basic" set-up with the DUT and th signals equal in frequency and 90° out of phase (quadrature) at the phase detector.

PHASE DETECTOR METHOD PROCEDURE

2. Calibration
   A. Beat note method

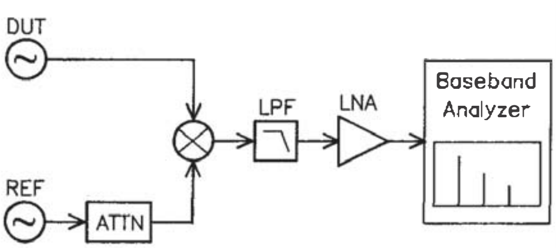
Offset one source to obtain a beat note (IF)



$$\Delta V = K_{\emptyset} \Delta \emptyset$$

After set-up, you must calibrate the phase detector. A common method of calibration is called the 'beat note method'. One of the input sources is offset from the frequency of the other to produce an IF signal. From this IF signal, we hope to find the slope of the sine wave at the zero crossings, which is equal to $K_{\emptyset}$.

PHASE DETECTOR METHOD BEAT NOTE CALIBRATION

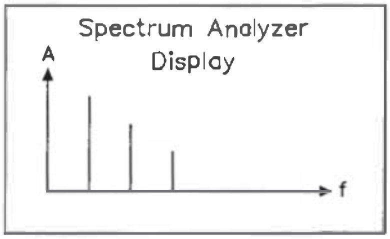


$$K_{\emptyset} = V_{b\,peak}$$

Typically, the levels to the mixer are adjusted such that the IF is a sine wave (usually by adding attenuation to the low signal path or R port). If the IF is a sine wave, mathematics tells us that the slope of a sine wave at the zero crossings is equal to the *peak amplitude.* Defining V_{bpeak} as the peak amplitude of the beat note or calibration signal, $V_{bpeak} = K_{\emptyset}$.

Another good reason to add attenuation during calibration is to avoid overloading the LNA (low noise amplifier) or the baseband spectrum analyzer during calibration. During the actual measurement of the noise sidebands, the LNA is designed to amplify lower level signals, not high beat notes. Also, we will see that in practice, best accuracy is obtained if the spectrum analyzer settings are not changed during calibrate and measure. This can be accomplished by appropriate setting of the calibration attenuation.

PHASE DETECTOR METHOD BEAT NOTE CALIBRATION



Known:

1) $K_\phi = V_{b_{peak}}$; $\Delta V(f) = K_\phi \Delta\phi$
2) measured value is $V_{b_{rms}}$

Therefore:

$$K_\phi = \sqrt{2}\ V_{b_{rms}}$$

$$\Delta V(f) = \sqrt{2}\ V_{b_{rms}}\ \Delta\phi(f)$$

On our baseband spectrum analyzer, we will see the beat note (and of course harmonically related components). A spectrum analyzer will measure V_{brms}. Therefore, $K_\varnothing$ is equal to $\sqrt{2}V_{brms}$. Defining our noise output voltage spectrum as $\Delta V(f)$, we see that $\Delta V(f)$ is equal to $\sqrt{2}\,V_{brms}$ times the phase fluctuations.

OTHER CALIBRATION TECHNIQUES

B. Non - Sinusoidal Beatnote

(IF = $A\sin\omega t + B\sin 3\omega t + C\sin 5\omega t + \ldots$)

$K_\phi = A - 3B + 5C - \ldots$

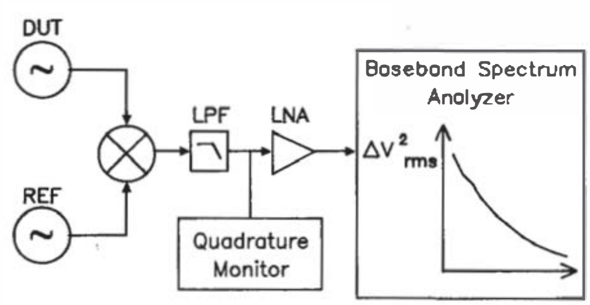
C. Known Sideband

There are also other calibration techniques for the phase detector method. One removes the limitation of having to have a sinusoidal beat note. In this calibration method, $K_\varnothing$ is computed from the amplitudes of all the IF signals.

The phase detector can also be calibrated with a known level sideband.

PHASE DETECTOR METHOD – PROCEDURE

3. Establish quadrature and measure



- Spectrum Analyzer measures ΔV^2_{rms}

After calibration (finding K_ϕ), the beat note is removed (sources set to identical frequencies) and the phase is adjusted to set the sources in phase quadrature. Now the output of the phase detector is a baseband signal, measured by the spectrum analyzer, measuring $\Delta Vrms^2$. How do we relate these voltage fluctuations to phase noise?

INTERPRETING THE RESULTS

KNOWN:

(1) $\Delta V(f) = K_\phi \Delta\phi(f)$ (Phase Detector)

(2) $\Delta V(f) = \sqrt{2}\, V_{brms} \Delta\phi(f)$ (From Calibration)

(3) $S_\phi(f) \equiv \dfrac{\Delta\phi^2_{rms}(f)}{B}$ used to measure $\Delta\phi_{rms}$

$\dfrac{rad^2}{Hz}$ (Definition)

THEREFORE:

$$\therefore\ S_\phi(f) = \Delta\phi^2_{rms}(f) = \left[\frac{\Delta V_{rms}(f)}{\sqrt{2}\, V_{brms}}\right]^2$$ (in 1Hz Bandwidth)

$$S_\phi(f) = \tfrac{1}{2}\,\frac{\Delta V^2_{rms}(f)}{V^2_{brms}}$$

for $\Delta\phi_{pK} \ll 1rad$

$$\mathcal{L}(f) = \tfrac{1}{2} S_\phi(f) = \tfrac{1}{4}\,\frac{\Delta V^2_{rms}(f)}{V^2_{brms}}$$

The phase detector method measures phase fluctuations. Therefore, the output of the phase detector is most naturally expressed in $S_\phi(f)$, which is the fundamental measure of phase noise. We can also express the output in terms of $\mathcal{L}(f)$.

INTERPRETING THE RESULTS (LOGARITHMICALLY)

$$\mathcal{L}(f)[\text{dbc}] = 10 \log \left[\frac{1}{4} \frac{\Delta V^2_{rms}(f)}{V^2_{brms}} \right] =$$

$$20 \log \Delta V_{rms} - 20 \log V_{b\,rms} + 10 \log \frac{1}{4}$$

$\mathcal{L}(f) = P_{ssb}(\text{dBm})$ power of noise spectrum in 1 Hz Bandwidth

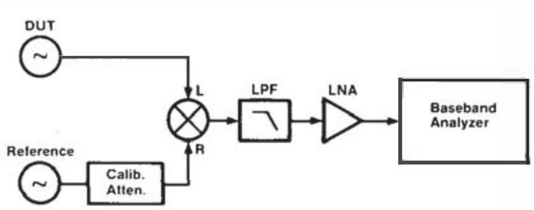
$-P_b$ (dBm) power of calibration beat note

-6(dB) conversion for rms value of beat signal and for $S_\phi(f) \rightarrow \mathcal{L}(f)$

Let's express these equations logarithmically, to correspond to the power readings on the spectrum analyzer. $\mathcal{L}$ (f) is as shown here; $S_\phi(f)$ (in dB) would be P_{ssb} - P_b - 3 dB.

CORRECTION FACTORS

1. Calibration Attenuation
2. Bandwidth Normalization
3. S/A Effects

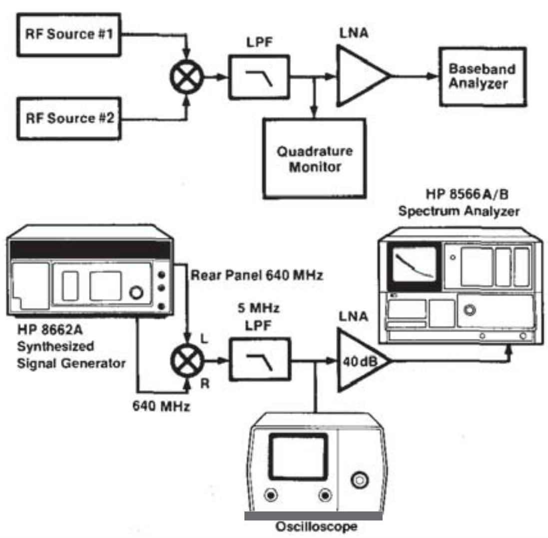


As well as the 6 dB 'correction' to the two measured readings of P_{ssb} and P_b, in practice there are other correction factors to take into account.

The BW normalization correction (because the noise sidebands are rarely ever measured in a 1 Hz bandwidth) and the correction for the response of analog spectrum analyzers to noise are already familiar. The correction for 'calibration attenuation' arises from the need to have a sinusoidal IF signal.

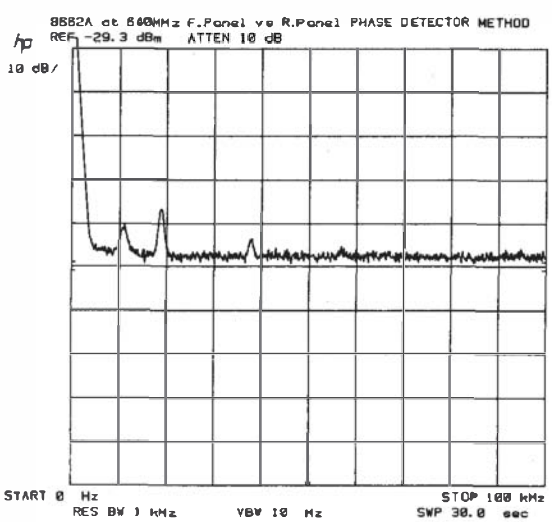
If attenuation is added in the R path of the phase detector, any amplitude change on the input will be translated to the output. Thus, V_{bpeak} measured during calibration (K_ϕ) will be low by the amount of the added attenuation. For example, if V_{bpeak} was measured as -40 dBm, and if 30 dB of attenuation had been inserted during calibration (but removed during measurement), then the actual level of K_ϕ during a noise measurement would be -10 dBm.

DEMONSTRATION



Phase Detector Method

TYPICAL MEASUREMENT RESULTS



"640 MHz Rear Panel vs. 640 MHz Front Panel"

Phase Detector Method

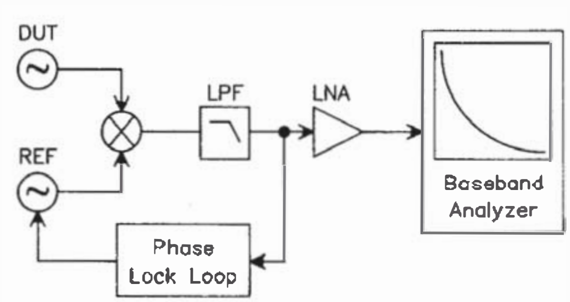
INTERPRETING the RESULTS

$$\mathcal{L}(f)[dBc] = 10 \log \frac{P_{ssb}}{P_s} = 10 \log P_{ssb} - 10 \log P_s$$

	10kHz	100kHz
= P_m power of measured noise	—	—
− P_b power of actual cal signal	—	—
− CAL.ATTEN. attenuated cal power	—	—
− 6dB	—	—
− Bandwidth NORMALIZATION	—	—
+ S/A CORRECTION (if applicable)	—	—
$\mathcal{L}(f)$ =	—	—

Putting in all correction factors leads to our final equation.

"REAL-LIFE" MEASUREMENT



In an actual system, the sources do not stay in quadrature by themselves. They are forced to remain in a 90 degree phase relationship by the use of a second order phase lock loop in a feedback path to one of the oscillators The error voltage out of the phase lock loop is applied to one of the sources, forcing it to track the other in phase. Note that this requires one of the sources to be electronically tunable.

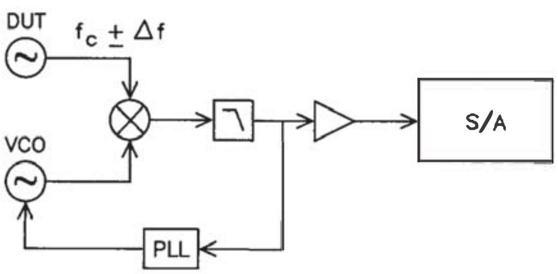
LHR – LOOP HOLDING RANGE

(DEF)

- LHR = Tuning range of source used as the loop VCO

(EFFECT)

- Amount that source under test can drift and still maintain lock at the phase detector



Since the phase lock loop (PLL) affects the measurement data so significantly, it is important to understand these effects. There are two important parameters of the PLL – Loop Holding Range and Loop Bandwidth.

The Loop Holding Range is defined as the tuning range of the source used as the loop VCO. In practice, this translates to how far the source under test can drift, and the PLL can still maintain quadrature at the phase detector. For example, if the loop VCO has only 25 kHz of electronic tuning range, it might be very difficult to try to lock up and track a 10 GHz microwave GaAs FET oscillator.

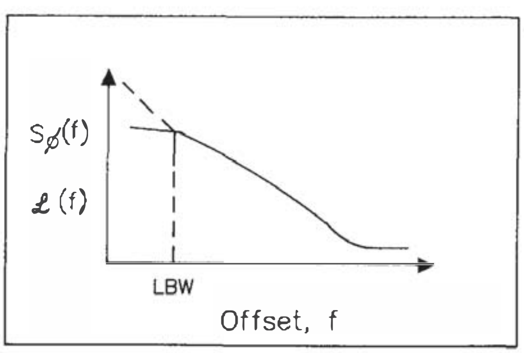
LBW – LOOP BANDWIDTH

(DEF)

- LBW = The rates of phase fluctuations that the loop can track

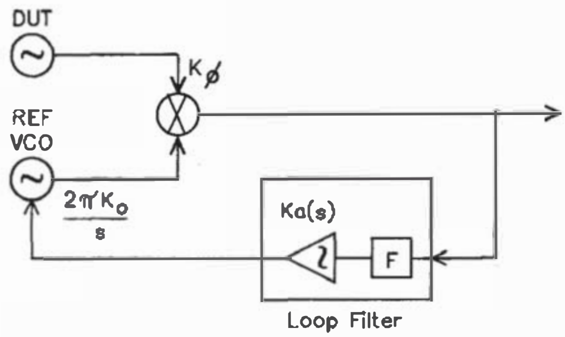
(EFFECT)

- Determines offset frequencies where correction factor is needed



The Loop Bandwidth is defined as the rates of phase fluctuations that the PLL can track. Outside of the LBW, the phase of the reference and the phase of the DUT are not correlated. In practice, LBW defines the offset frequencies where the measured phase noise data out of the phase detector represents the noise of the DUT.

WHAT DETERMINES LOOP BANDWIDTH?



Second order phase lock loop model

K_ϕ = Phase slope V/rad (phase detector gain factor)

K_o = VCO slope (Hz/V)

F = Loop gain stage

$K_a(s)$ = Loop amplifier gain (function of freq.)

Since the phase lock loop is such an important part (the phase detector method, it is useful to understand n about it, and in particular, what determines the actua loop bandwidth. This simplified block diagram of a sec order phase lock loop (second order indicating the integrator in the feedback path) shows several essenti components. K_ϕ is determined by the mixer and the signal input levels. K_o is a function of the source used the loop VCO. F and $K_a(s)$ are dependent on the desigr the phase lock loop; often, if F is fixed, it's value will b included in the value for $K_a(s)$.

WHAT DETERMINES LOOP BANDWIDTH?

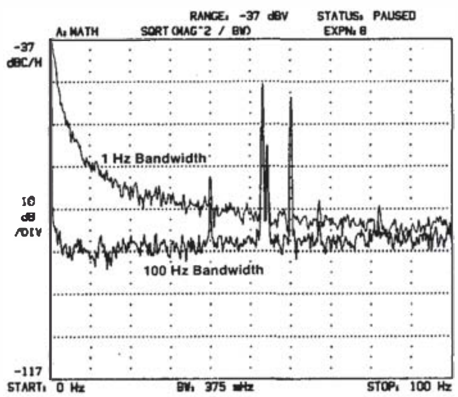
$$G_{OL} = \frac{K_\phi F K_a(s) 2\pi K_o}{s}$$

$$LBW = f\big|_{G_{OL} = 1}$$

$$\longrightarrow \boxed{\therefore LBW = K_\phi F K_a(s) K_o}$$

The open loop gain is defined as the products of the gains around the loop. The loop bandwidth is defined : that frequency that makes the open loop gain equal tc

EFFECTS of LBW



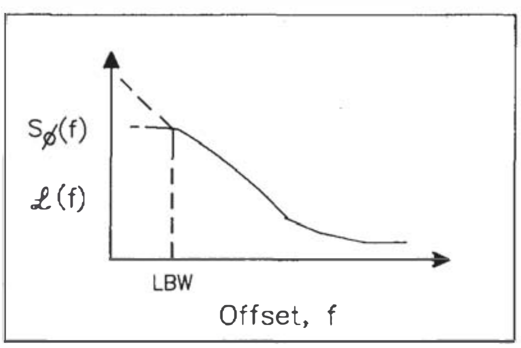
Noise plots showing effects of loop bandwidth setting

The effect of the loop bandwidth on the measured data is easily demonstrated. This spectrum analyzer display is from 0 to 100 Hz. The upper trace is the detected noise spectrum output when the phase lock loop had a 1 Hz LBW. The lower trace clearly shows the noise suppression within the PLL BW when the DUT was locked to the reference in a 100 Hz LBW.

What loop bandwidth is needed for a given source? The bandwidth of the phase lock loop must be chosen large enough to keep the phase detector in its linear range; ie, such that the phase deviations are less than about 0.2 radians. Thus, for low noise synthesizers, with low close-in noise, it is possible to lock up a 10 GHz source with less than a 1 Hz loop bandwith. If the close-in noise is high, or if there are many close-in spurious such as line-related spurs (which can also drive the mixer out of its linear range), a wider bandwidth loop must be chosen to reduce the close-in noise. The noisier the source, the wider the loop bandwidth needed. (In practice, if a source has very high-level close-in noise, it will be easier to make the measurement using a frequency discriminator.)

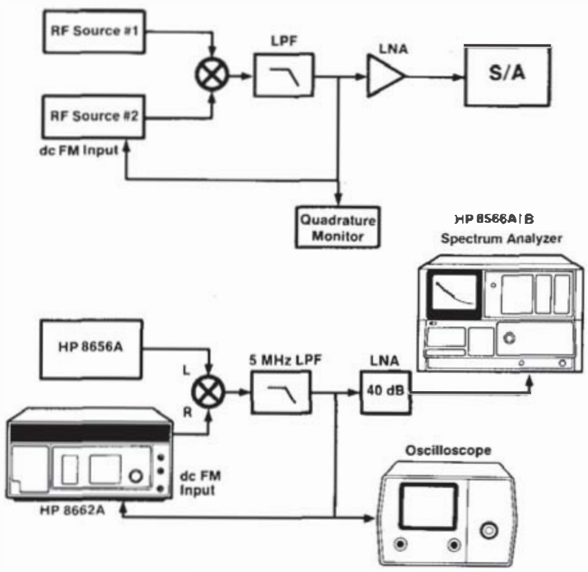
WHAT'S THE PRICE OF THE PLL?

- Additional correction factor needed at offsets < LBW



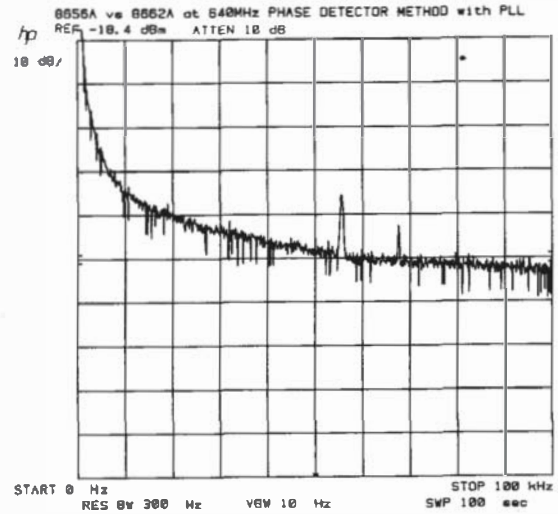
Not only does the phase lock loop add circuit complexity to the measurement, it also exacts a penalty. Because a phase lock loop forces one source to track the other source *in phase,* it also effectively suppresses the detected phase noise of the test source within the phase lock loop bandwidth. (Remember, the phase detector outputs a voltage proportional to the phase difference of the two input signals. If the phase of one source tracks the other, the output delta phase will be less.) Thus, when making phase noise measurements with the phase detector method, uncorrected measurements are valid only outside the loop bandwidth.

DEMONSTRATION



Phase Detector Method with PLL

TYPICAL MEASUREMENT RESULTS



HP 8656A at 640 MHz

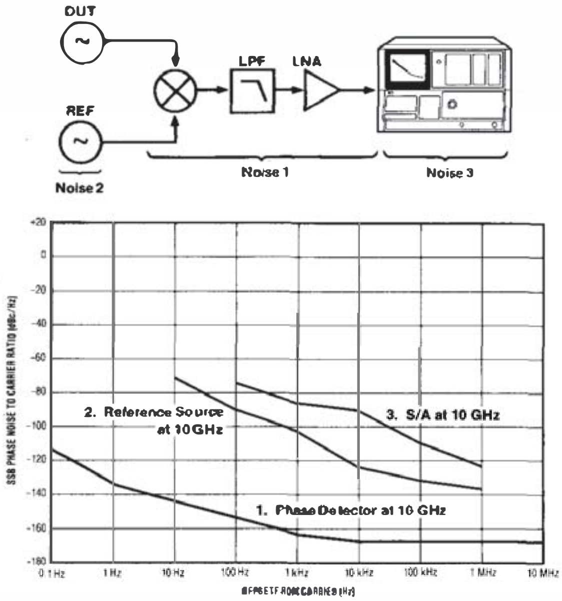
Phase Detector Method with PLL

INTERPRETING the RESULTS

$$\mathcal{L}(f)[dBc] = 10\log \frac{P_{ssb}}{P_s} = 10\log P_{ssb} - 10\log P_s$$

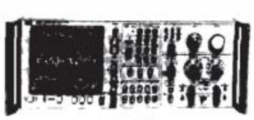
	10kHz	100kHz
= P_m power of measured noise	___	___
- P_b power of actual cal signal	___	___
- CAL.ATTEN. attenuated cal power	___	___
- 6dB	___	___
- Bandwidth NORMALIZATION	___	___
+ S/A CORRECTION (if applicable)	___	___
$\mathcal{L}(f)$ =	___	___

WHAT SETS SYSTEM NOISE FLOOR IN PHASE DETECTOR METHOD?



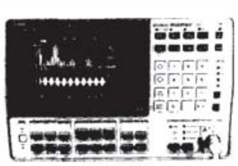
If we've rejected the direct spectrum method because the noise floor was not low enough, how can we use these broadband spectrum analyzers as the baseband analyzer in the phase detector technique and still get a low noise floor?

PHASE DETECTOR METHOD BASEBAND SPECTRUM ANALYZER CHOICES



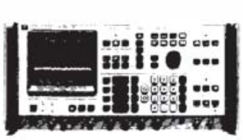
HP 3582A

- 20 mHz to 25 kHz
- 2–channel FFT
- Noise source for loop characterization
- rms averaging
- 1 Hz BW normalization



HP 3561A

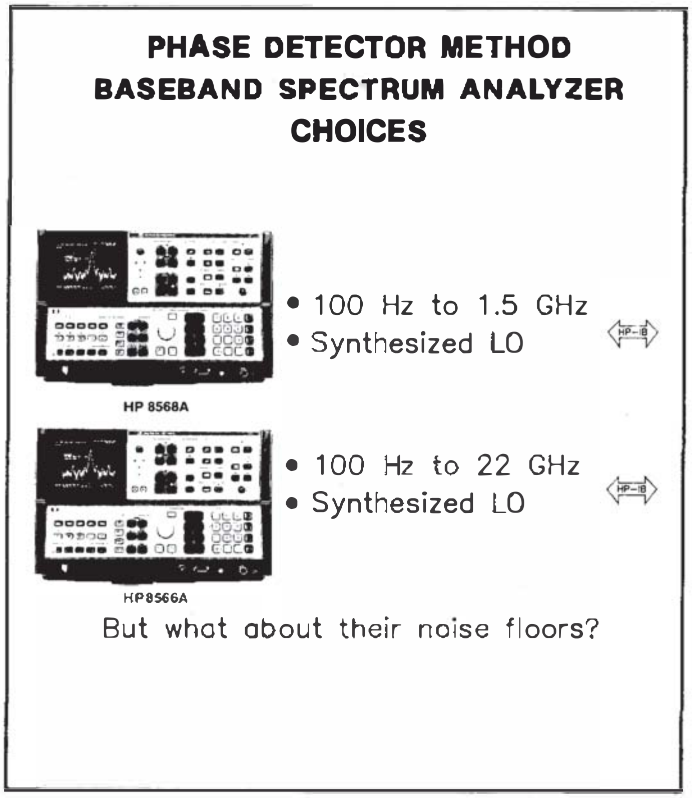
- 20μHz to 100 kHz
- Single channel FFt
- Time and frequency displays
- Noise source for loop characterization
- rms averaging



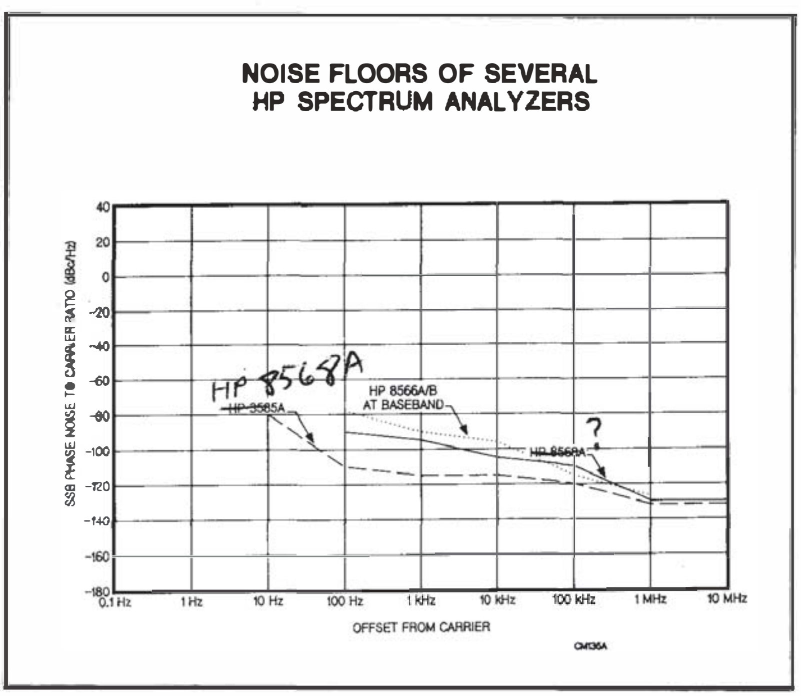
HP 3585A

- 20 Hz to 40 MHz
- Synthesized LO
- Tracking generator for loop characterization
- Digital and analog averaging
- 1 Hz BW normalization

For use as the baseband spectrum analyzer, three choices are very popular, depending on offset range of interest, and whether or not one or two baseband analyzers are used.



Also very popular for use as the baseband spectrum analyzer are these two broadband analog spectrum analyzers. Because these analyzers are often already needed in the test station for other measurements, use of them also as the baseband analyzer leverages the equipment requirements. But what about the required noise performance?

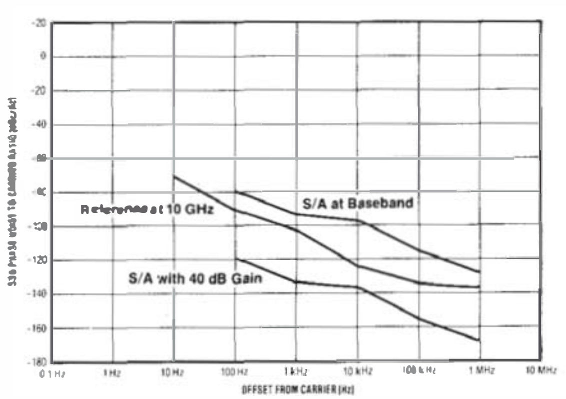


Here are the noise floors of the local oscillators i several of the HP spectrum analyzers listed. Of co general, the noise of a higher frequency spectrum analyzer is higher than that of a lower frequency spectrum analyzer.

WHY DOESN'T A BROADBAND SPECTRUM ANALYZER'S NOISE FLOOR LIMIT SENSITIVITY?

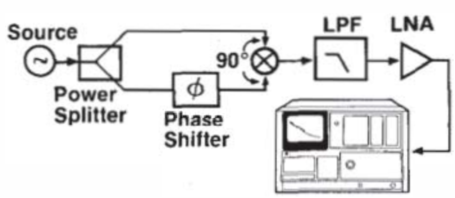
Phose detection gives added sensitivity;

1. Removol of carrier allows LNA use and moximum S/A dynamic range.
2. Translates noise to bosebond where S/A is most sensitive.



Several factors combine to decrease the effective noise floor (or increase the effective sensitivity) of the analyzers. In general, using HP's implementations of the phase detector method, and any of HP's synthesized LO spectrum analyzers, the noise floor of the combined measurement system will be limited by the noise of the *reference source*, not the noise floor of the analyzer used.

MEASUREMENT OF PHASE DETECTOR METHOD SYSTEM NOISE FLOOR

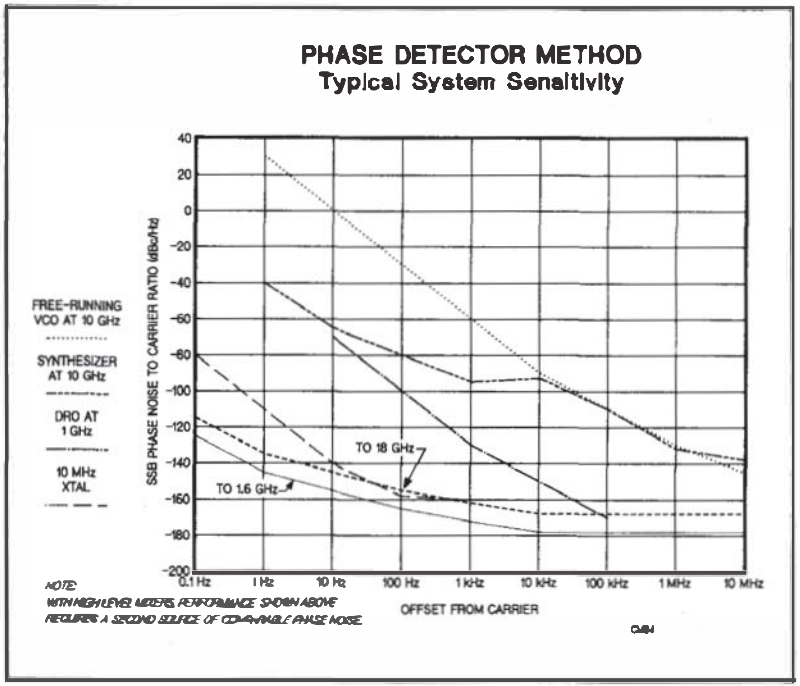


Equal delay in both paths means signal's fluctuations still correlated at the mixer

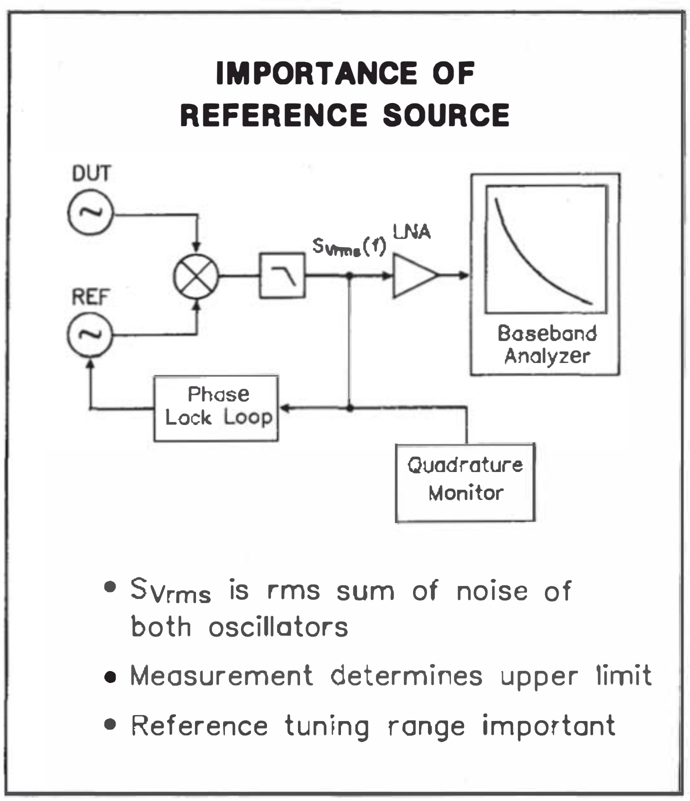
CAUTIONS:

- AM noise of source must be sufficiently rejected
- Path delays must be equal to prevent decorrelation of source noise.

The noise floor of the phase detector method is fairly easily measured. It is important to minimize any delay difference between the two paths, so that the noise of the source will be well-correlated out. This leaves the noise contributions of the mixer and LNA to be measured.



The sensitivity of the phase detector method is s
here. The phase detector method provides good ser
over the entire offset frequency range; it can be us
measure high quality standards (good close-in noi
state-of-the-art free-running oscillators (low broac
noise). The sensitivity shown can be obtained with
level mixers; note that typically using a lower frec
mixer yields higher sensitivity.



The most critical component (or at least usually the hardest to obtain) of the phase detector method is probably the reference source. Since the spectrum analyzer measures the rms sum of the noise of both oscillators, the most important criterion for choosing a reference source is that its phase noise be less than what is being measured. The measured noise sets an upper limit; the measured noise will be the maximum noise of either source and at any particular offset frequency the noise of one of the sources will be at least 3 dB lower.

Though the absolute phase noise is the most important criteria for selection as a reference for the phase detector method, its AM noise, if very high, will also affect the accuracy of the measurement. Also, in general, it is desired to have the reference source act as the loop VCO, as the noise of the DUT can change if it is being used in a dc FM mode. If the reference is being used as the loop VCO, its electronic tuning range will in many cases determine whether or not a measurement can be made on a given device. Many low-noise sources that might be suitable for use as a reference in a phase detector measurement either do not have electronic tuning (typically dc FM) or they do not have sufficient tuning range.

IMPORTANCE of REFERENCE SOURCE

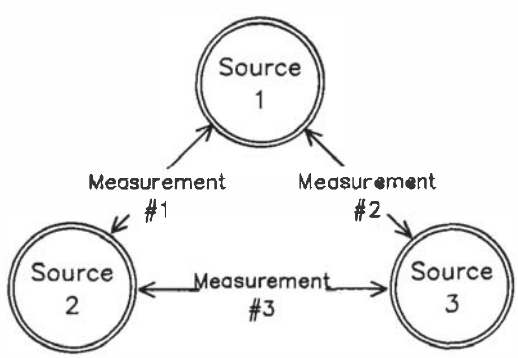
HOW MUCH LOWER SHOULD THE REFERENCE NOISE BE?

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

$\mathcal{L}_{dut} - \mathcal{L}_{ref}$ (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

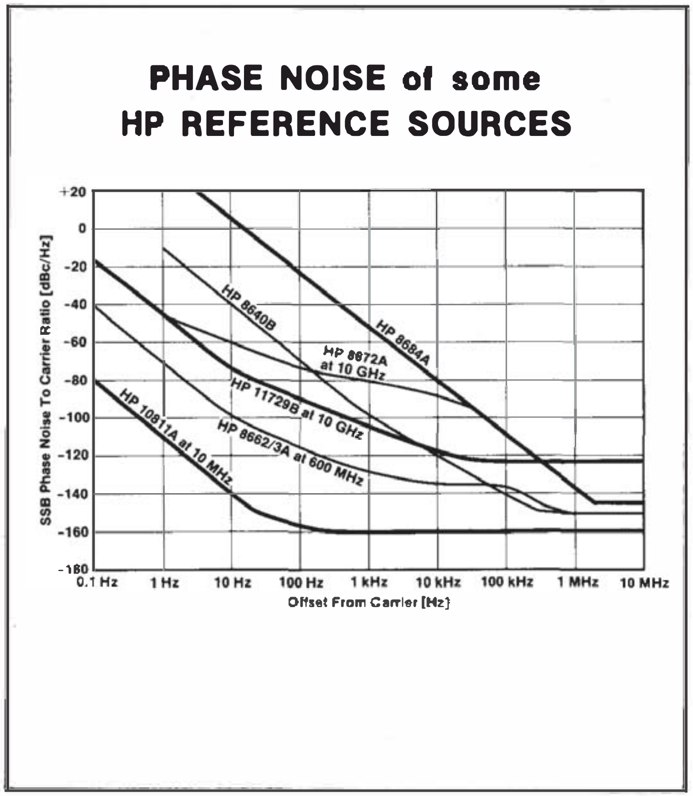
A standard margin of 10 dB is usually sufficient to ensure the measurement results are not significantly affected. If a reference source with low enough phase noise to measure the full offset range is not available, several alternatives are available. One option is to use several reference sources with sufficiently low noise at specific offset ranges. Another method would be to use a reference source comparable to the source under test so that the measurement results can be attributed equally to the noise from each source. (The assumption is made that both sources have equal noise, and 3 dB is subtracted from the measured value.)

THREE–SOURCE COMPARISON



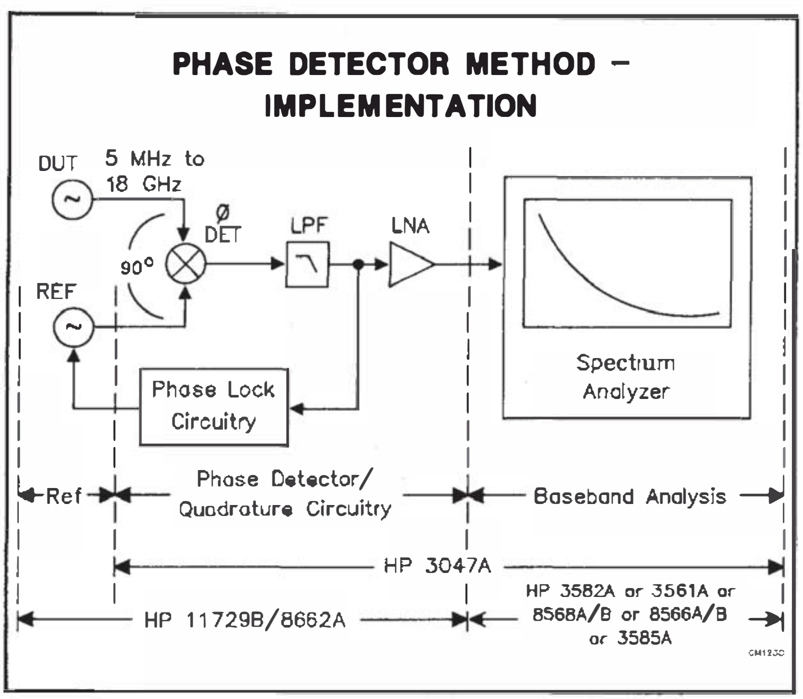
Determines noise of each source provided noise of source is comparable (3 to 6 dB difference).

If three comparable sources are available (within about 3 dB of each other and all tunable), it is possible to make three pair-wise measurements and separate out the noise from each source. If one of the sources is appreciably lower (approximately 3 to 6 dB lower) than the others, its lower noise performance will still be indicated, but its actual noise information cannot be accurately separated out from the higher noise of the other devices.



Choice of a reference source depends on the frequency o the DUT, the required noise performance, and whether or not the reference needs to operate as the loop VCO. (Since phase noise is probably the single most expensive parameter in a source, its useful to determine how good a reference is required.)

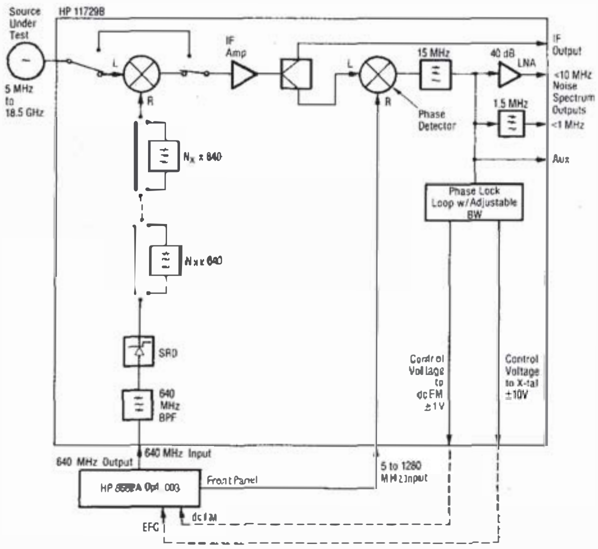
Here are some of the best HP reference sources. For RF applications, the HP 8640A/B features very good broadband performance for measuring free-running sources. The HP 8662A and 8663A provides the lowest overall phase noise of any commercially available broadband source, for DUTs to 2.56 GHz. For microwave applications, the cavity tuned HP 8684A has low broadband noise and 10 MHz of dc FM. The HP 8670 family (and related HP 8340 family) are synthesizers with good close-in noise. And the lowest overall noise performance at microwave is provided by the HP 11729B/ 8662A.



Moving next to available instrumentation, it is convenient to break up the required components three essential pieces; the reference section, the p detector/quadrature circuitry, and the baseband a section. We will look at available instrumentatio three of these areas.

PHASE DETECTOR METHOD

SIMPLIFIED HP 11729B BLOCK DIAGRAM



One HP implementation of the phase detector method is the HP 11729B Carrier Noise Test Set, which embodies all the little bits and pieces of the phase detector method, including the ability to generate a very low noise microwave reference signal, to make measurements easier. Not a single box solution, it is always used with one or more RF sources and a baseband analyzer.

It represents a modified phase detector method in that it utilizes a dual-downconversion to baseband. A low noise 640 MHz reference signal (available from the HP8662A or 8663 Option 003) is multiplied to microwave with a step recovery diode. A single combline is mixed with the microwave DUT, to an IF in the range of 5 to 1280 MHz. The double balanced mixer phase compares the IF signal with the front panel output signal of the HP 8662A. A variable bandwidth phase lock loop is used to generate error voltages, to electronically tune the HP 8662A and maintain phase quadrature at the phase detector.

EFFECT OF MULTIPLICATION ON THE NOISE OF A SIGNAL

from Bessel algebra for small m:

$$\frac{V_{ssb}}{V_s} \cong J_1 \cong 1/2\,\frac{\Delta f_{peak}}{f_m} \cong 1/2\,\Delta\phi_{peak}$$

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

If $f_2 = x f_1$, then

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

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

To be able to determine the noise floor of the HP 11729B/8662A at any frequency, we need to understand the effect of multiplication on the noise of a signal. Since phase noise (or frequency noise) can be thought of as continuous modulation sidebands, phase noise increases when a signal is multiplied.

COMPUTING HP 11729B/8882A SYSTEM NOISE

$$\mathcal{L}\ \text{system (dBc)} = 10 \log \left(N^2 \times 10^{\frac{\mathcal{L}1}{10}} + 10^{\frac{\mathcal{L}2}{10}} + 10^{\frac{\mathcal{L}3}{10}}\right)$$

N = Harmonic of 640 MHz reference

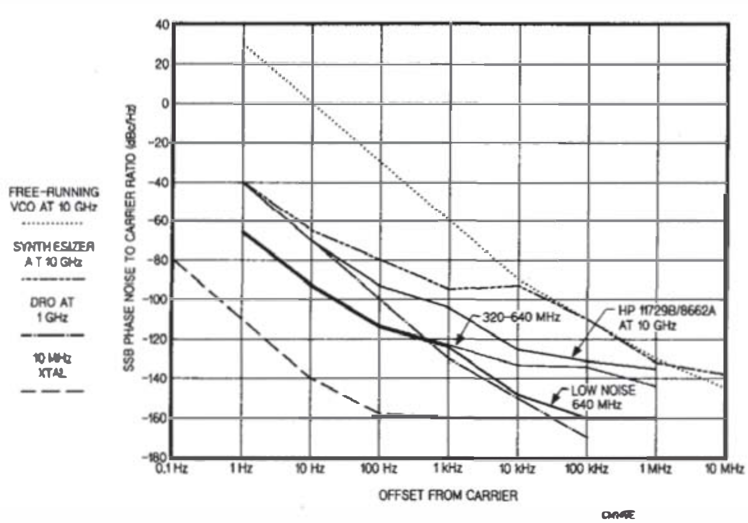
$\mathcal{L}_1$ = Absolute SSB Phase Noise of 640 MHz reference (dBc/Hz)

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

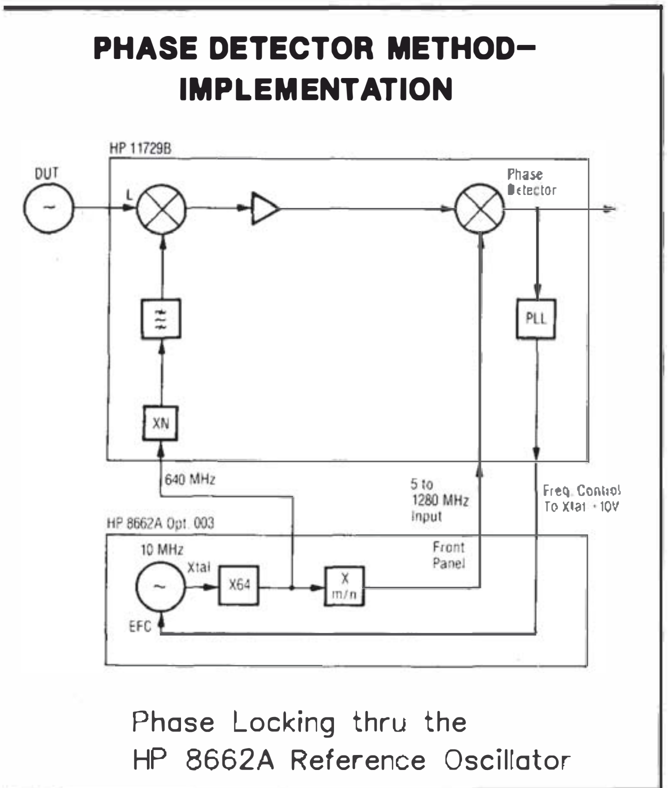
$\mathcal{L}_3$ = Two–port noise of HP 11729B (dBc/Hz)

There are three noise contributions in the HP 11729B – the phase noise on the 640 MHz signal (available with HI 8662A Option 003), the noise on the tunable signal at the phase detector, and the two-port noise of the HP 11729B. System noise will depend on the frequency of the DUT, and thus how many times the 640 MHz signal must be multiplied.

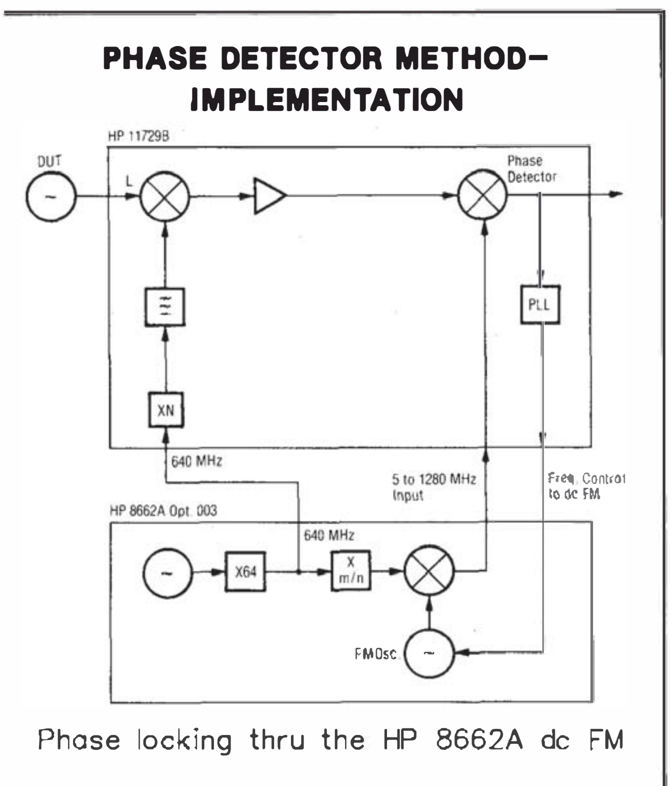
TYPICAL HP 11729B/8662A SYSTEM SENSITIVITY



The 640 MHz signal is used for multiplication of its lower broadband noise level. Note that for multiplication factors, the broadband noise on tl 640 MHz signal can add significantly to the tota noise. At higher multiplications (carriers greate about 8 GHz), the multiplied noise of the 640 MI is dominant. This multiplication scheme results 30 dB lower noise at 10 kHz than that available microwave sources such as the HP 8670 or HP 8: families. (At 10 kHz from a 10 GHz carrier, the I and HP 8340 have typical noise of about −93 dB HP 11729B/8662A has typical noise of −123 dBc this offset!)

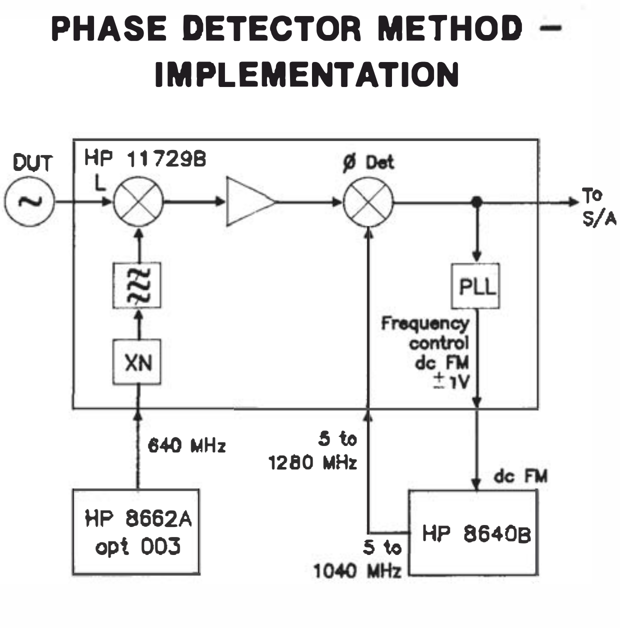


Thus, the HP 11729B/8662A provides the lowest noise floor for test signals to 18 GHz. It also provides the loop VCO needed for the phase detector method. Depending on the source under test and how wide of phase lock loop bandwidth will be needed to maintain the phase detector in its linear range, there are several ways to phase lock. First, for very stable sources, phase lock can be establishes through the electronic frequency control (EFC) of the HP 8662A internal 10 MHz reference oscillator, using the HP 11729B ±10 V phase lock loop control voltage output. This yields a loop holding range at 10 GHz of 1 kHz, and a maximum loop bandwidth of about 1 kHz also.



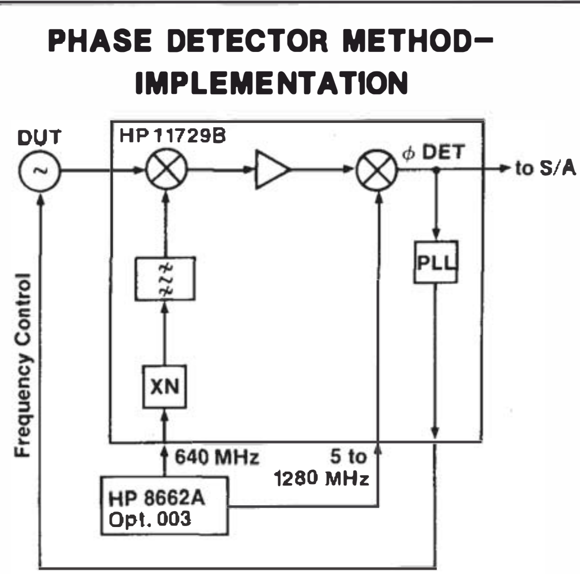
For stable free-running sources (cavities, fixed frequency free-running sources), lock via the HP 8662A dc FM with the ±1 V loop control voltage. This yields a maximum LHR of 200 kHz, and a maximum loop bandwidth of about 50 kHz.

Note that since obtainable HP 8662A dc FM deviation is a function of frequency (and can drop as low as 25 kHz), the obtainable LHR and LBW is dependent on the frequency of the DUT and the resultant IF frequency. If in a situation where the allowable deviation is only 25 kHz and lock cannot be established, first change the DUT frequency if possible to yield a new IF. If this cannot be done, go to another method of phase locking.



Phase locking using a third source as the loop VCO

For very unstable sources, a third RF source can be used as the loop VCO. In particular, the HP 8640B is a good choice because it features wide dc FM deviation (up to 10 MHz).



Phase locking using the frequency control of the DUT

A fourth choice for phase locking is via the tuning port of the DUT. Since microwave oscillators typically have much more tuning range than RF sources, this method can be useful for locking very unstable sources. However, since the phase noise of a source can be affected by tuning, it is usually preferred to make the measurement without using the DUT as the VCO.

In general, if the source is unstable to the point where phase locking via a 3rd RF source or locking via the DUT seems the only way to lock, it might be easier to use the frequency discriminator method, in particular if very close-in measurements are not desired.

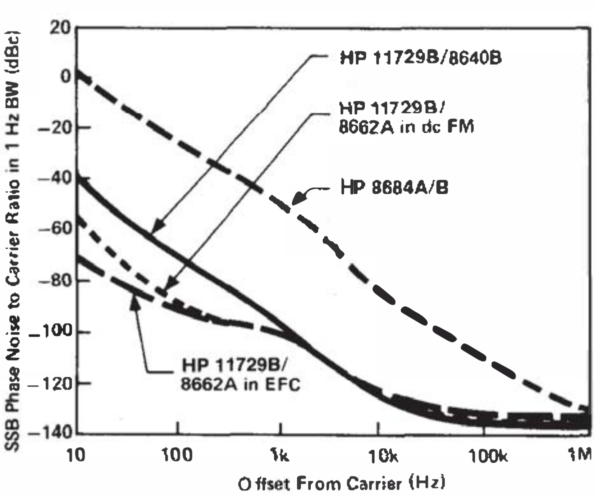
Also check measurement set-up; see if line spurs can be reduced with clean power supplies.

COMPARISON OF LOOP VCO'S AND LOOP BANDWIDTH (LBW)

VCO	Tuning Range	≈ Max LBW
HP 8662A EFC	$f_{DUT}/10^7$	1 kHz
HP 8662A dc FM	±25 to ±200 kHz	50 kHz
HP 8640B dc FM	±5 MHz	100 kHz
DUT	>10 MHz	150 kHz

Summarizes the options for phase locking and their typical performance.

EFFECT of METHOD of PHASE LOCKING on SYSTEM NOISE FLOOR



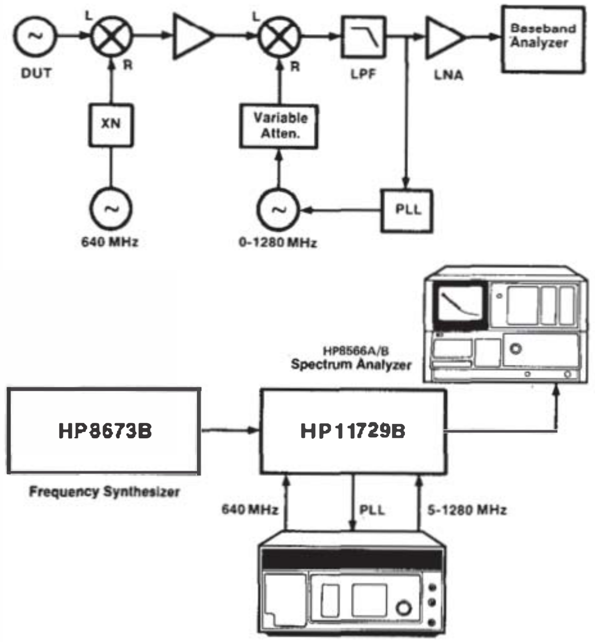
Since the method of phase locking uses different sources, the system noise also changes. In general, choosing a loop VCO with more tuning range increases system noise. However, since it is usually noisier sources that require more loop bandwidth, the increased system noise typically does not limit the measurement.

PHASE DETECTOR METHOD— PROCEDURE

1. Set—Up
2. Calibrate
3. Establish Quodrature and Measure
4. Corrections

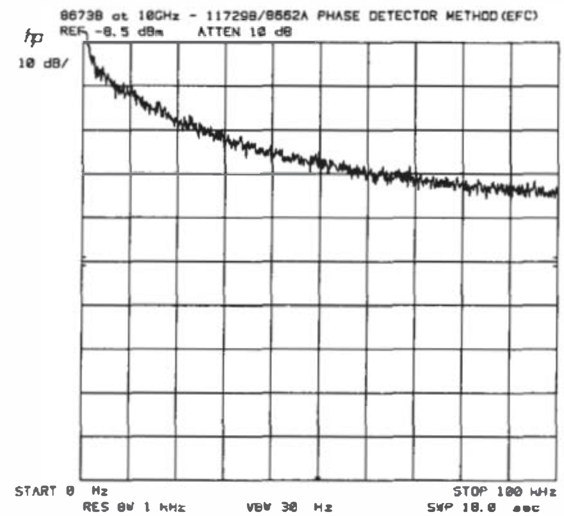
To see how to use the HP 11729B/8662A in the phase detector method to make a phase noise measurement, let's review the procedure.

DEMONSTRATION



Phase Detector Method
Using the HP 11729B/8662A

TYPICAL MEASUREMENT RESULTS



HP 8673B at 10 GHz

Phase Detector Method
Using the HP 11729B/8662A

INTERPRETING the RESULTS

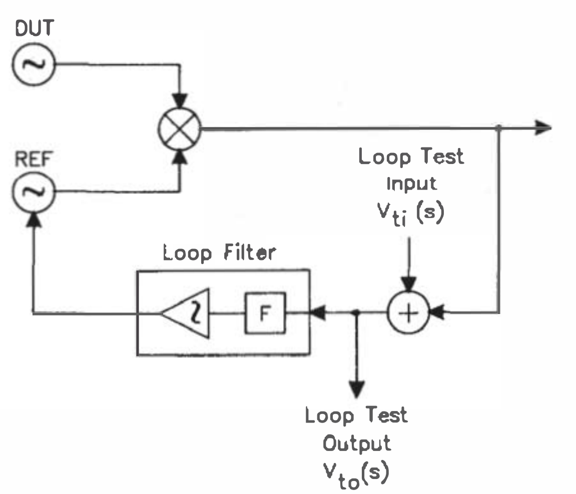
$$\mathcal{L}(f)[dBc] = 10 \log \frac{P_{ssb}}{P_s} = 10 \log P_{ssb} - 10 \log P_s$$

	10kHz	100kHz
$= P_m$ power of measured noise	—	—
$- P_b$ power of actual cal signal	—	—
- CAL.ATTEN. attenuated cal power	—	—
- 6dB	—	—
- Bandwidth NORMALIZATION	—	—
+ S/A CORRECTION (if applicable)	—	—
$\mathcal{L}(f) =$	—	—

BUT WHAT IF I WANT TO MAKE MEASUREMENTS AT OFFSETS INSIDE THE LOOP BANDWIDTH?

Even with all the right components, and careful design and technique, the primary limitation to the phase detector method seems to be close-in measurement. The phase detector method has theoretically good close-in sensitivity. How can we make use of this?

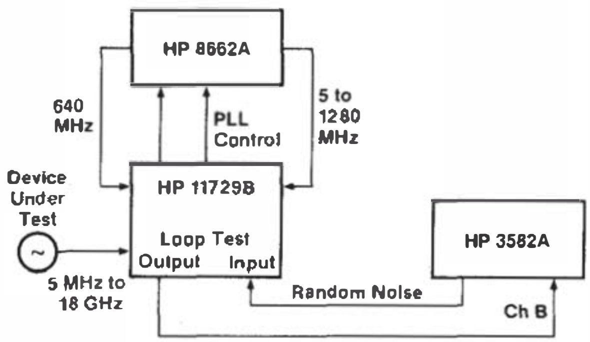
PHASE LOCK LOOP CHARACTERIZATION



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

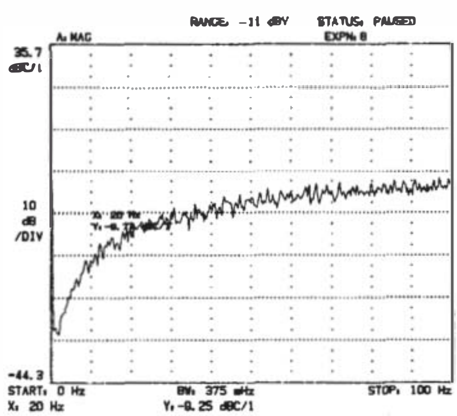
It is possible to characterize the effect of the phase lock loop on noise, and then to add to the measured value the amount of the loop noise suppression. A phase lock loop can be characterized by injecting flat noise into the loop, and then measuring the response of the loop.

SYSTEM SET-UP FOR LOOP CHARACTERIZATION



For loop characterization, the HP 11729B has loop test ports. A random noise source or tracking generator (or variable sine wave) is applied to the loop test input, and the loop response can be traced out at the loop test output. The HP 3582A's or HP 3561A's random noise source works well for this. Note that the noise source in the HP 3582A or 3561A is band-limited; thus, the HP 3582A can be used to characterize loops only 25 kHz wide, the HP 3561A, loops 100 kHz wide.

PHASE LOCK LOOP CHARACTERIZATION



Typical loop filter transfer function

Here is a typical phase lock loop filter transfer function. The display yields two important pieces of information. First, the LBW can be determined, which designates the offset frequencies for which an uncorrected phase noise measurement can be made. In this example, loop bandwidth is about 100 Hz. Any noise measured at offsets greater than 100 Hz do not need correction. Second, the amount of noise suppression as a function of offset frequency can be measured and then used to correct the measured value of noise. For example, at a 20 Hz offset, the loop suppresses the noise by about 10 dB. The level of phase noise measured at 20 Hz would need to be corrected by this 10 dB for a valid measurement.

INTERPRETING THE RESULTS

$\mathcal{L}(f) = P_m(f)$ measured noise power

− Bandwidth normalization

− P_b measured cal signal

− Cal attn

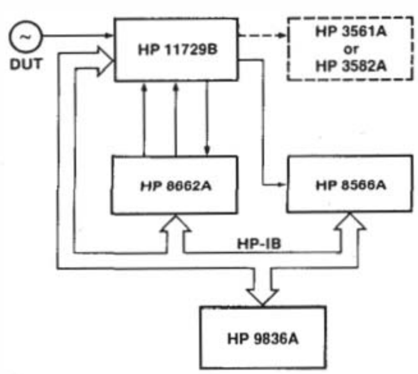
− 6dB

+ S/A correction
(if applicable)

+ Loop noise suppression
(if applicable)

We now have our final equation for obtaining $\mathcal{L}(f)$ values from the phase detector method.

AUTOMATIC PHASE NOISE MEASUREMENTS



SYSTEM ADVANTAGES

- Improved productivity and accuracy
- Hard copy output
- Flexibility

Since the HP 11729B and HP 8662A are HP-IB programmable, an automatic phase noise measurement system is easily set-up with the baseband spectrum analyzer of choice. As mentioned, a common system is the HP 11729B/8662A/8566A and the HP 3561A if offsets close-to-the-carrier are desired.

AUTOMATIC PHASE NOISE MEASUREMENT SYSTEM
HP 3047A

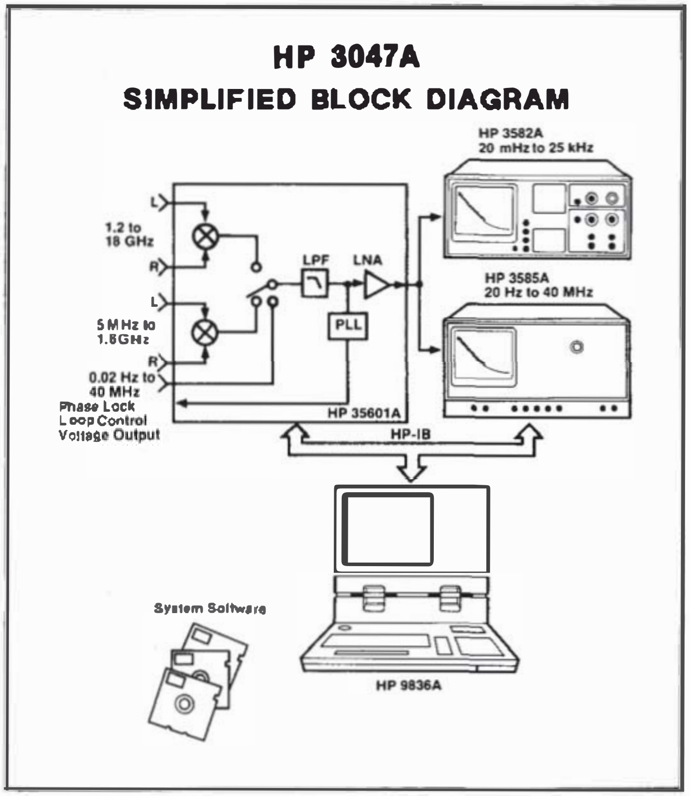
Automatic Calibration and Measurement of $\mathcal{L}(f)$. $S_{\phi}(f)$, $S_{\Delta f}(f)$, $S_y(f)$, $m(f)$

Offset Range: 0.02 Hz to 40 MHz
Carrier Range: 5 MHz to 18 GHz
and above
Measurement Accuracy: ± 2 dB

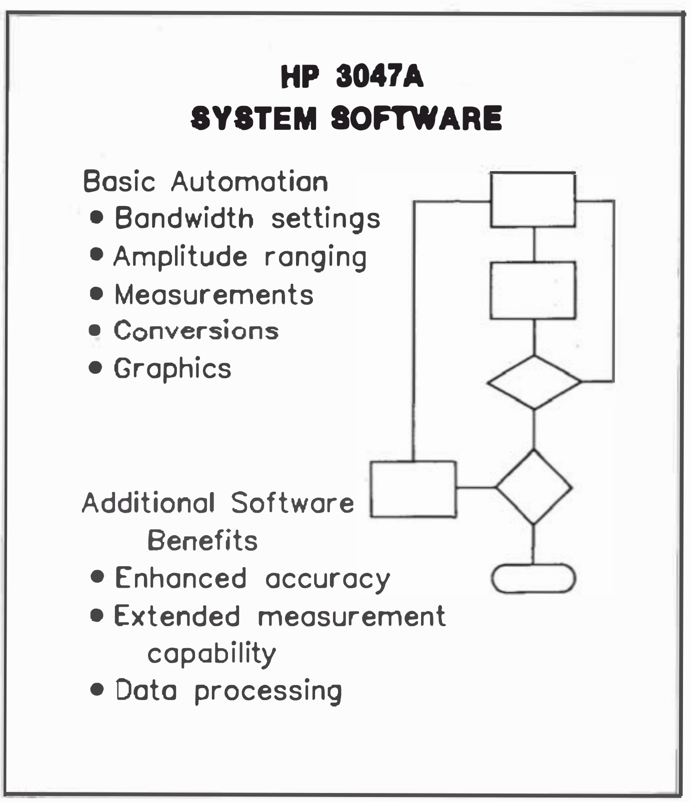
AUTOMATIC PHASE NOISE MEASUREMENT SYSTEM
HP 3047A

1. Direct spectrum mode
2. Noise sideband mode
   (20 Hz to 40 MHz)
3. Phase noise mode
   (5 MHz to 18 GHz)

A fully configured hardware/software system for phase noise measurements is found in the HP 3047A Phase Noise Measurement System. It can be used in three measurement modes, but is most powerful in the phase noise mode.



The HP 3047A consists of the HP 3582A and HP 3585A Spectrum analyzers, the HP 35601A Spectrum Analyzer Interface, the HP 9836A Desktop Computer, and system software and documentation. The HP 35601A contains a microwave phase detector, an RF phase detector, baseband signal processing circuitry, and a powerful, variable bandwidth phase lock loop.



The HP 3047A system software not only takes the manual steps out of the measurement, it also enhances the accuracy and data processing capability of the measurement.

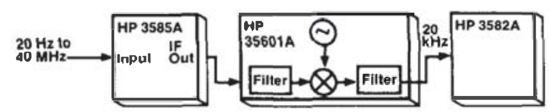
HP 3047A SYSTEM SOFTWARE

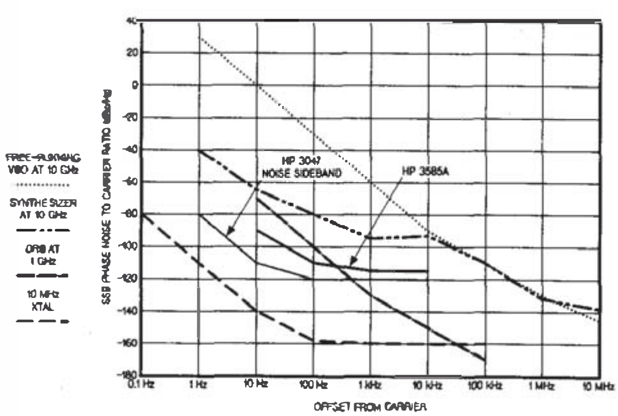
POWERFUL SYSTEM SOFTWARE RAISES ACCURACY AND VERSATILITY

- Signal path calibration
- Phase detector calibration
- Mixer dc offset calibration
- Phase lock loop characterization
- Active marker
- Line drawing
- Integrated noise
- Three source comparison

Increased accuracy is obtained with the 3047A's extensive 'self-characterization'. The signal path is calibrated from the phase detector, thru the low pass filters and amplifiers, and including the phase lock loop response. The 3047A uses an algorithm to correct for phase lock loop response, allowing accurate measurements many decades inside the loop bandwidth.

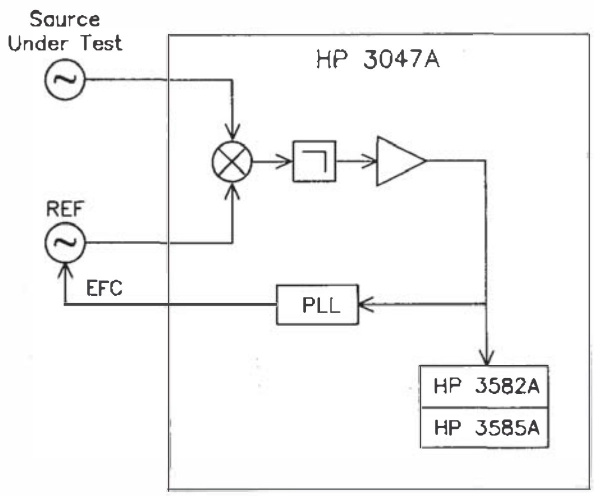
HP 3047A < 40 MHz NOISE SIDEBAND MODE





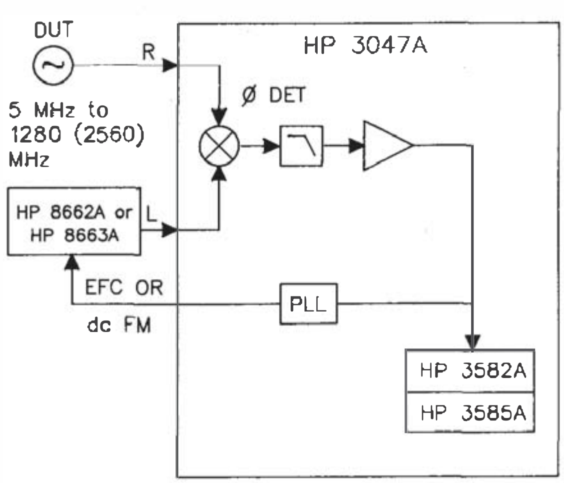
The simplest use of the HP 3047A software is in the direct spectrum mode, or the 40 MHz noise sideband mode. In the noise sideband mode, an internal source is used as the reference, improving the sensitivity over the direct spectrum capability of the HP 3585A. It is useful on sources to 40 MHz.

CONFIGURING THE HP 3047A (PHASE DETECTOR METHOD) WITH ARBITRARY REFERENCE SOURCE



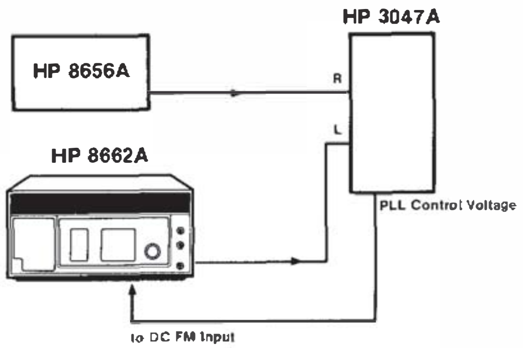
The most powerful configuration of the HP 3047A is in the phase noise mode. The standard HP 3047A does not include a reference source. For state-of-the-art measurements, a second device under test or user-defined reference source (one of which must be tunable) can be used as a reference in the phase detector method.

AUTOMATIC PHASE NOISE MEASUREMENTS ON RF SOURCES HP 3047A/8662A OR HP 8663A



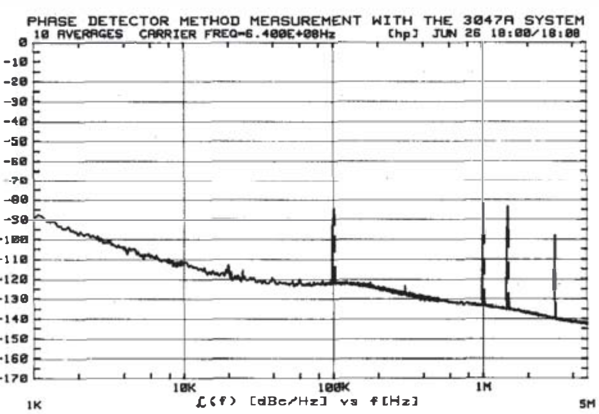
For RF measurements, the HP 8662A or HP 8663A is an excellent choice as a reference source, yielding a system for measurement of sources at frequencies up to 2.56 GHz.

DEMONSTRATION



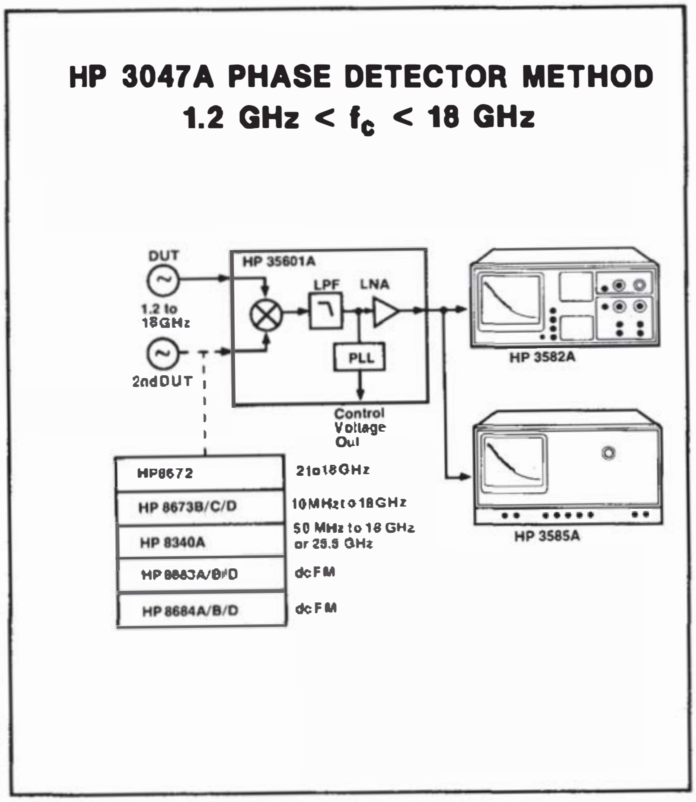
Phase Detector Method Using the HP 3047A

TYPICAL MEASUREMENT RESULTS

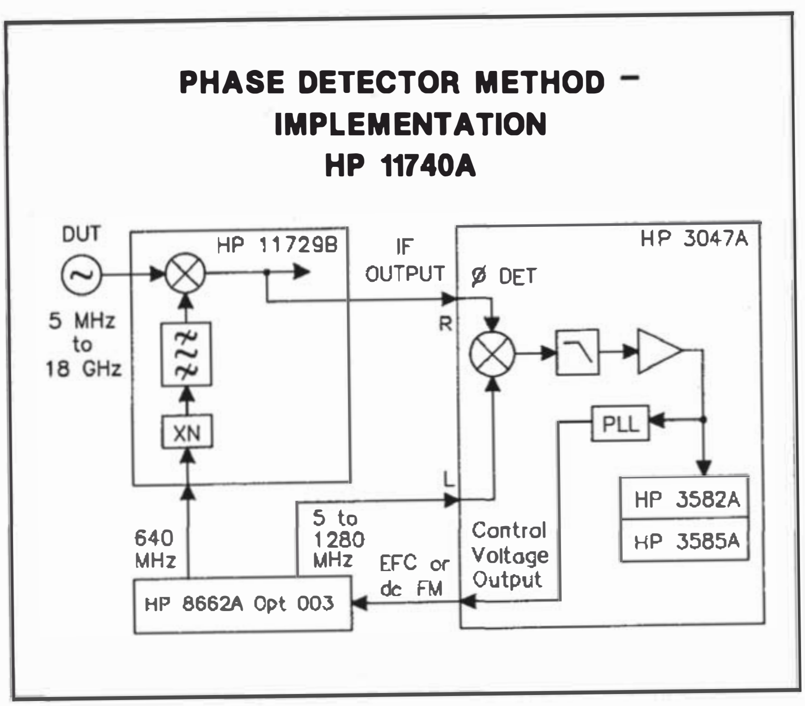


HP 8656A ot 640 MHz

Phase Detector Method
Using the HP 3047A



For microwave measurements, a number of HP sources could be used as the reference, depending on the noise floor desired. This source would be controlled manually.



The HP 11740A Microwave Phase Noise Measu System is a complete Automatic System for phas characterization of sources, 5 MHz to 18 GHz. It c of the HP 11729B Carrier Noise Test set, the HP 8 HP 8663A) Synthesized signal generator, the HP Spectrum Analyzer System, the HP 9836A Deskt Computer with required operating system and m (ordered separately), and system software, docum and warranty.

The HP 11740A adds the necessary hardware a software integration of the critical low noise micr reference source into the standard HP 3047A Spe Analyzer System, with an overall system specific and warranty. All standard features and software capabilities of the HP 3047A discussed on earlier are retained.

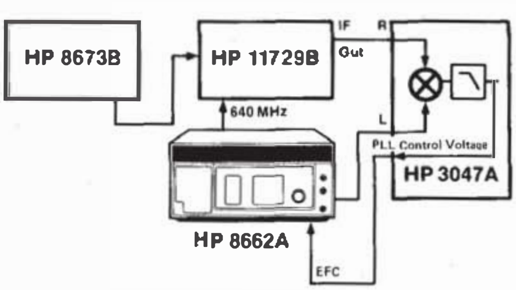
The options permit deletion of an instrument al owned or upgrading to a higher performance instr

In this configuration, the HP 11729B is used to upconvert the HP 8662A, and to downconvert the microwave DUT. The phase detector, baseband sig processing and phase lock loop of the HP 3047A ar employed to complete the phase detector method implementation.

HP 11740A SYSTEM ADVANTAGES

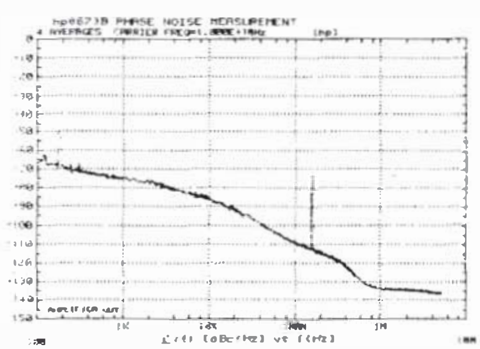
- Built-in low noise microwave reference source and VCO
- Complete system specifications
- User-friendly, menu driven software
- System documentation and support

DEMONSTRATION



Phase Detector Method
Using the HP 11740A

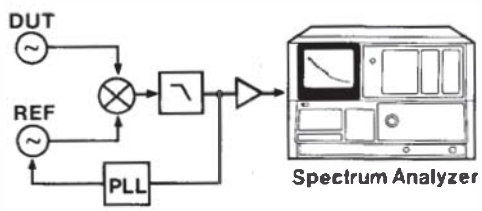
TYPICAL MEASUREMENT RESULTS



Phase Detector Method
Using the HP 11740A

PHASE DETECTOR METHOD

CONSIDERATIONS IN SYSTEM ACCURACY



CONTRIBUTOR	TYPICAL UNCERTAINTY (±dB)
1. Spectrum analyzer accuracy (relative)	0.4–1.5
2. Attenuator accuracy of calibration	0.4
3. Mismatch uncertainty (associated with setting CW reference level	0.15
4. Uncertainty in ±2.5 dB correction	0.20
5. Accuracy of measured IF noise bandwidth	0.20
6. Relative IF bandwidth gains	0.05
7. Analyzer frequency response	0.25
8. Phase detector flatness	0.20
9. Baseband signal processing flatness	0.50–1.0

Here's a very detailed isolation of potential error sources. Of course, the largest potential error is the relative accuracy of the spectrum analyzer, with the accuracy of the calibration step, and the baseband signal processing (low pass filters, LNA) flatness also very important.

PHASE DETECTOR METHOD

CONSIDERATIONS IN SYSTEM ACCURACY (CONTINUED)

CONTRIBUTOR	TYPICAL UNCERTAINTY (±dB)
10. Random error due to randomness of noise	0.5
11. Mixer dc offset/quadrature maintenance	0.03
12. Noise floor contribution ($\mathcal{L}_{ref} = \mathcal{L}_{dut}$ −15dB)	0.2
13. Accuracy of loop characterization	0.2

OUTSIDE LOOP BANDWIDTH

	Minimum	Maximum
LINEAR SUMMATION	3.08	4.68
RSS	0.85	1.53

INSIDE LOOP BANDWIDTH (one decade)

	Minimum	Maximum
LINEAR SUMMATION	3.28	4.88
RSS	0.86	1.53

Now, depending on whether the errors are simply linearly summed, or the root sum of the squares is taken, or maybe some combination of the two, a phase noise measurement using the phase detector method can be made with better than ±2 dB accuracy.

PHASE DETECTOR METHOD GENERAL SYSTEM PRECAUTIONS

- Non-linear operation of the mixer will result in a calibration error
- Saturation of the amplifier or S/A in calibration or by high spuriour signals (e.g. line spurs)
- Distortion of RF-signal yields deviation of K_ϕ from V_b peak
- Supression or peaking of noise near the phase lock loop bandwidth
- Impedance interfaces should remain unchanged between calibration and measurement
- Injection locking of the DUT can occur
- Deviation from phase quadrature (results in lower K_ϕ)
- Closely spaced spurious can be misinterpreted as phase noise with insufficient S/A resolution and averaging
- Noise injected by peripheral circuitry (e.g. power supply) can be a dominant contributor to phase noise
- Vibration can excite significant noise in DUT

The phase detector, the PLL, the reference source – all are critical components in the phase detector method. Measurement procedure and system set-up can also be very important. Here is a list of things to watch out for, and possible sources of error.

PHASE DETECTOR METHOD

ADVANTAGES

- Lowest overall noise floor
- Wide range of offset frequencies
- Wide input frequency range
- Some AM suppression
- Easily automated

DISADVANTAGES

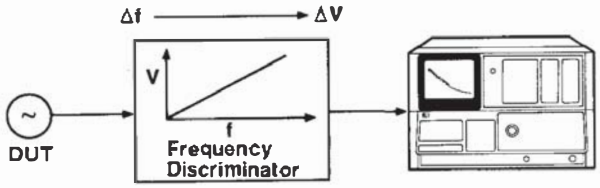
- Requires two sources
- Complexity
- Close-in measurements require additional correction
- Difficulty with locking up high drift rate sources

In final summary, the phase detector method is enjoying increasing popularity because it has the lowest overall system sensitivity. It is also the most complex method, but with careful system design, it can be a very general-purpose solution.

4. Frequency Discriminator Method

PHASE NOISE MEASUREMENT OF SOURCES

4. Frequency Discriminator method



$$S_{\Delta f}(f) = \frac{\Delta f^2_{rms}(f)}{B} \left[\frac{Hz^2}{Hz}\right] = \frac{S_{vrms}(f)}{K_d^2}$$

$S_{vrms}(f)$ = power spectral density of the voltage fluctuations (V^2/Hz)

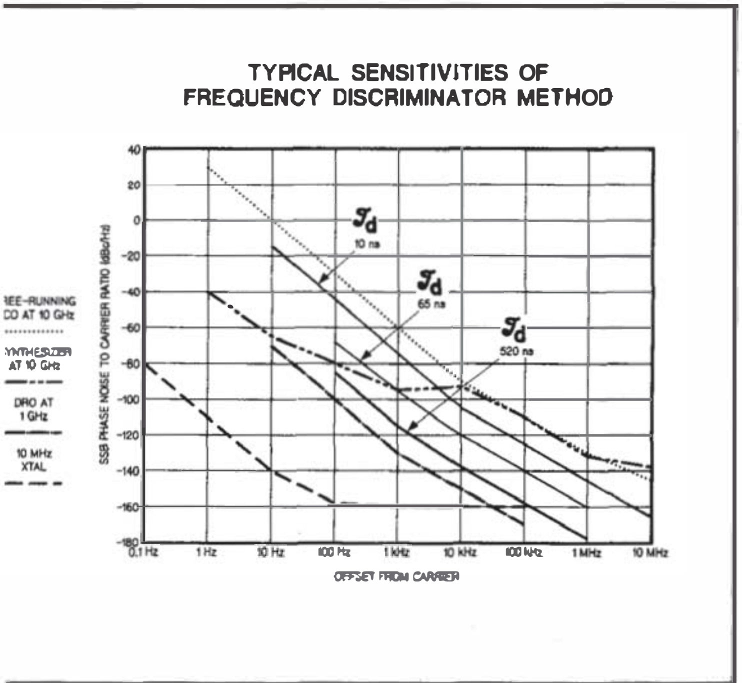
K_d = frequency discriminator constant (V/Hz)

The fourth method of phase noise measurement on sources is called the frequency discriminator method (also sometimes called the one-oscillator method). In this method, the frequency fluctuations of the source are translated to baseband voltage fluctuations which can then be measured by a baseband analyzer. K_d, the frequency discriminator constant in V/Hz, is defined as the translation factor between the frequency fluctuation at the input of the discriminator and the corresponding voltage output. The fundamental output of the frequency discriminator can be described by $S_{\Delta f}(f)$, the spectral density of frequency fluctuations.

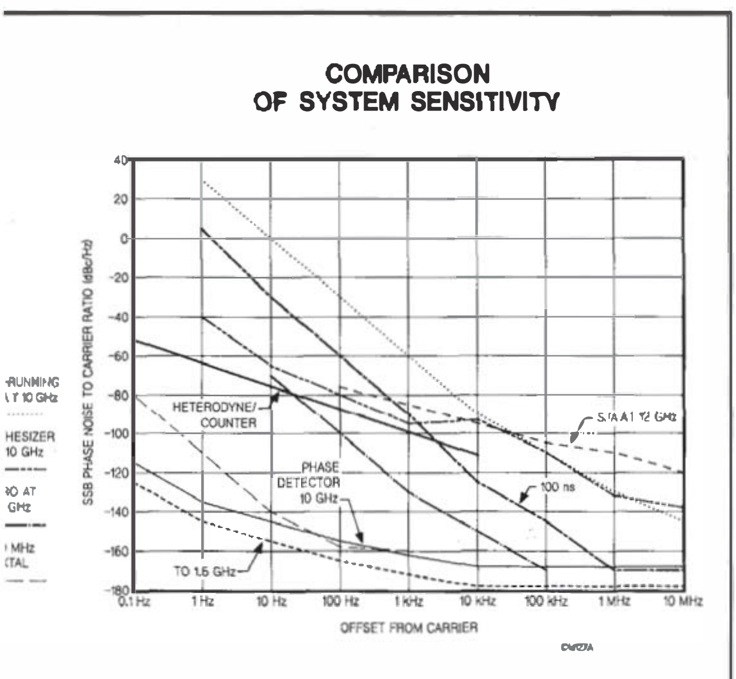
FREQUENCY DISCRIMINATORS

- Delay Line/Mixer
- RF Bridge/Delay Line
- Cavity Resonator
- Slope Detector and Ratio Detector
- Dual Delay Line (Cross–Spectrum Analysis)

There are several common implemenations of frequency discriminators, each with its own advantages and disadvantages. For example, a cavity resonator used as a frequency discriminator can yield very high sensitivity, but typically very narrow input bandwidth. We will concentrate our discussion on the delay line/mixer implementation.

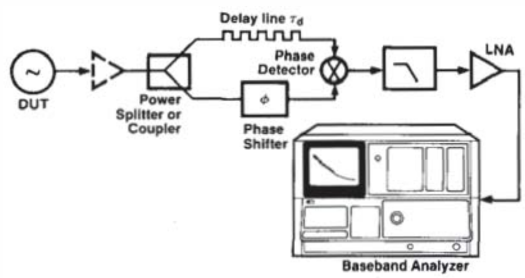


A delay line/mixer used as a discriminator has typica sensitivity as shown; notice that the sensitivity is a function of the delay time. The frequency discriminato method has very good broadband sensitivity; however, because of the inherent relationship between frequency and phase, the sensitivity of the discriminator method degrades as $1/f^2$ as the carrier under test is approached. Because this slope follows the noise characteristic of fre running sources, and because this technique does not require a second source for downconversion, it is useful measuring sources with large, low rate phase instabilities. But because of its high close-in sensitivity is not very useful for measuring very stable sources clo to the carrier.



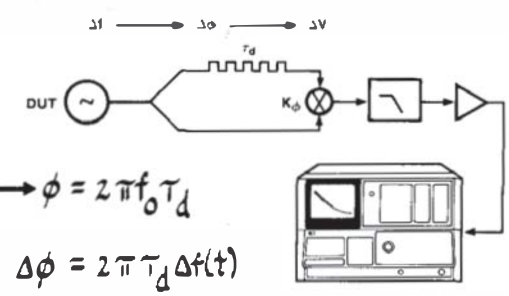
This graph then compares all four measurement methods with the four "standard" sources that have bee plotted on all previous sensitivity graphs. Remember th the phase detector method sensitivity is really only a theoretical value given a perfect reference source. In a practical system, the noise floor of the reference sets the system noise floor.

DELAY LINE/MIXER USED AS A FREQUENCY DISCRIMINATOR



In the 'delay line/mixer' implementation of a frequency discriminator, the DUT is split into two channels. One channel, sometimes called the non-delay or reference channel, is applied directly to one port of a double-balanced mixer, which will be operated as a phase detector. This channel is also referred to as the local oscillator channel since it drives the mixer at the prescribed impedance level (the usual LO drive). The other channel is delayed through some delay element, and then input to the other port of the mixer.

MEASURING $S_{\Delta f}(f)$



$$\rightarrow \phi = 2\pi f_0 \tau_d$$

$$\Delta\phi = 2\pi \tau_d \Delta f(t)$$

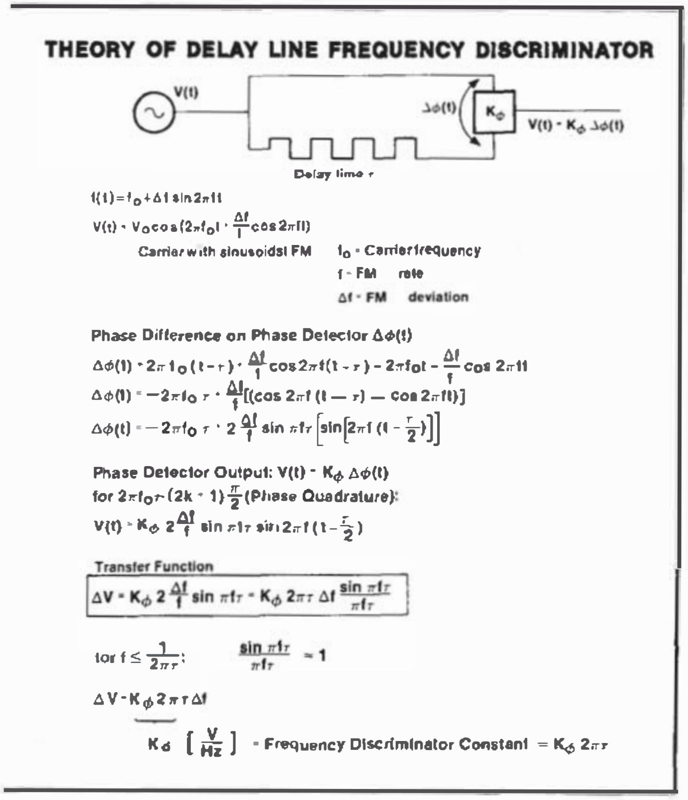
$$\Delta V_{rms} = K_\phi \Delta\phi_{rms}$$

$$\Delta V_{rms} = K_\phi 2\pi \tau_d \, \Delta f_{rms}(t)$$

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

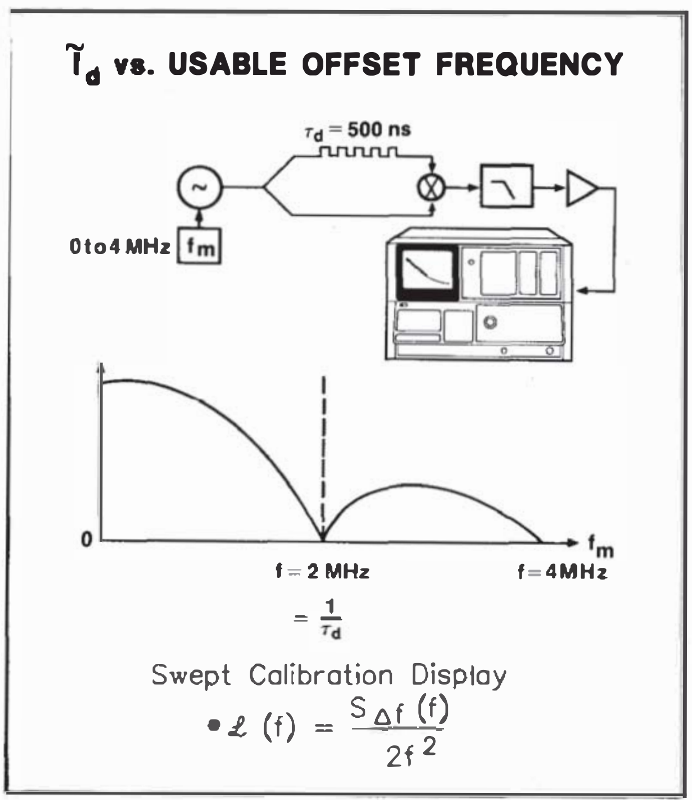
$$= \left. \frac{\Delta V^2_{rms}(f)}{(K_\phi 2\pi \tau_d)^2} \right|_{B = 1\,Hz}$$

Note that the delay line can serve as a frequency to phase transducer. That is, the nominal frequency arrives at the phase detector at a certain phase. If we change the frequency passing through the delay line, then the amount of phase shift incurred in the fixed delay time will be proportional to the frequency. This delay element uncorrelates the noise of the source arriving at the phase detector in the two paths. The phase detector then converts the phase fluctuations into their voltage equivalent for measurement.



A more rigorous derivation of a delay line/mixer used as a frequency discriminator is shown here. The important equation is the final magnitude of the transfer response. The sinusoidal output term of the mixer responds as sin $\pi f\tau/\pi f\tau$. This means the output response is periodic in $\omega = 2\pi f$, and will have peaks and nulls, with the first null at $1/\tau_d$. To avoid having to compensate for the sin x/x response, measurements are typically made at modulation frequencies well away from the null, ½ or less. It is possible to measure to offset frequencies out to and beyond the null by scaling the measurement results using this function, but the sensitivity of the system gets poor near the nulls.

Also note that the final transfer function (assuming measurements made $< \frac{1}{2}\tau_d$) is independent of carrier frequency. Thus, a 50 nsline at 10 GHz has the same sensitivity as a 50 ns line at 500 MHz.



To see the effect of the delay element on input frequencies, note that if the differential delay between the two channels is zero, there is no phase difference at the detector output as a swept CW signal is applied to the system. Here, the detected output interference display is shown when a swept frequency CW signal is applied to a system which has a differential delay of 500 ns. (The signal amplitudes are here assumed to be almost equal.) Since there is a two-channel note, there is a null in the response every 360 degreess $1/\tau_d$ Hz as shown.

Lastly, the conversion to $\mathcal{L}(f)$ is shown again here. Because the frequency discriminator outputs $S_{\Delta f}(f)$ directly, when the sensitivity of the frequency discriminator system is plotted as a function of phase variations $\mathcal{L}(f)$ or $S_{\phi}(f)$, the offset frequency squared term, f^2, in the denominator indicates that system sensitivity will increase by 20 dB per decade as the offset frequency of the measurement decreases. The sensitivity gets better until it equals the sensitivity of the phase detector at an offset frequency of $1/(2\pi\tau)$.

FREQUENCY DISCRIMINATOR METHOD PROCEDURE

1. Measure Power Levels
2. Calibrate
3. Establish Quadrature and Measure
4. Corrections

To better understand the frequency discriminator method, let's go through a typical measurement procedure on our test source. There are basically four steps in the delay line/mixer implementation: 1) set-up and measure power levels, 2) calibrate, 3) establish quadrature and measure, and 4) corrections.

2. CALIBRATION METHOD #1

measure τ_d and K_ϕ

$$S_{\Delta f}(f) = \frac{\Delta V^2_{rms}(f)}{(K_\phi 2\pi\tau_d)^2} = \frac{\Delta V^2_{rms}(f)}{K_d^2} \left[\frac{Hz^2}{Hz}\right]$$

1. τ_d is a function of delay element used

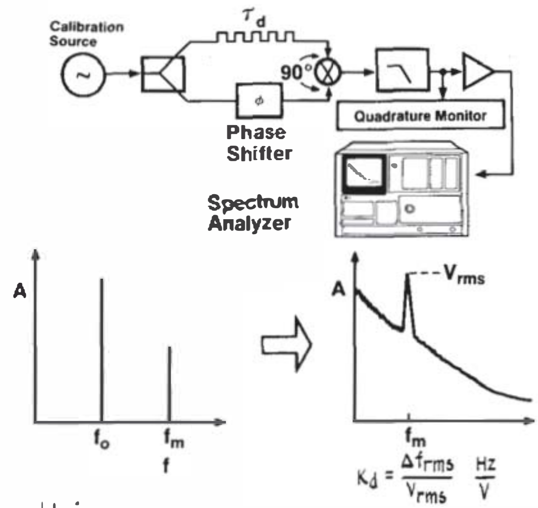
   $\tau_d \cong$ 1.5ns/foot for cables with polyethelene dielectric

2. $K_\phi = V_{bpeak}$ for sinusoids

One calibration (ie, establish the transfer coefficient from frequency fluctuations to voltage fluctuations) method is to determine K_d from τ_d and $K_\varnothing$. τ_d is a function of the length and type of delay element used; τ_d is approximately 1.5 ns/foot for cables with polyethelene dielectric. $K_\varnothing$, or the phase detector constant, can be determined by establishing a sine wave out of the mixer. The slope of the sine wave at the zero crossings ($K_\varnothing$) is equal to the peak amplitude of the signal (for linear mixer operation). This calibration method requires an accurate measure of the delay, τ_d. This can be determined by measuring the carrier frequency difference, Δf_0 between two consecutive zero crossings on the quadrature monitor ($\Delta f_0 = \frac{1}{2\tau_d}$).

2. CALIBRATION METHOD #2

Find composite K_d by evaluating response of system to a well-characterized input.

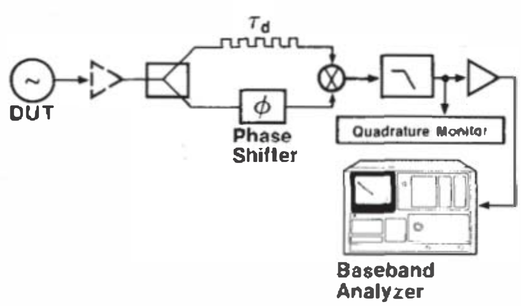


Using:

- Known sideband/signal or
- Known β or
- Known deviation

A more common calibration method is to determine the composite K_d by evaluating the response of the system (delay line, mixer, low pass filter and LNA) to a known input. This can be done by setting up a single FM tone with known sideband/carrier ratio, or by setting up FM such that the carrier is nulled, yielding a known modulation index, or setting up a known deviation (with the system in phase quadrature). The response of the system to this known value can then be used as a reference level.

3. ESTABLISH QUADRATURE AND MEASURE



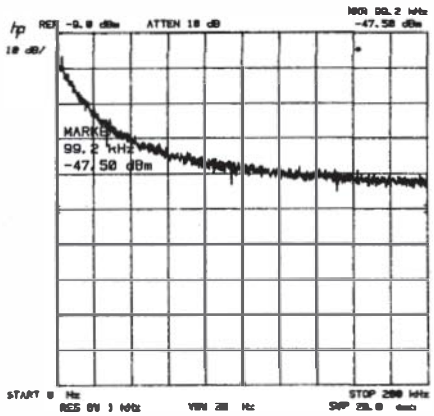
A. Establish Quadrature by:
   1.) Using phase shifter
   2.) Using line stretcher
   3.) Varying DUT frequency

B. Measure $S_{V_{rms}}(f)$ on spectrum analyzer

After calibration, the input signals to the phase detector are re-adjusted for the phase quadrature condition (90 degrees out of phase) if necessary. There are several ways to establish quadrature. First, you can vary the frequency of the DUT slightly (perhaps a couple of MHz is sufficient), until the average voltage out of the phase detector is 0V dc (measured on a quadrature monitor such as an oscilloscope or DVM). If the DUT frequency cannot be changed, a line stretcher can be added to the fixed delay path and adjusted until quadrature is established. Typically, no more than a few inches of variable length are needed. If this is awkward, manual or digital phase shifters can also be used.

Once quadrature is established, the spectral density of the voltage fluctuations can be measured on the baseband analyzer. Typically, the delay line is broadband enough such that even if the source drifts, the two inputs to the phase detector will remain in the quadrature condition for the duration of the measurement. However, quadrature should be monitored during the measurement, in case the DUT frequency changes significantly.

4. CONVERT TO PROPER UNITS AND CORRECTIONS



ΔV^2_{rms} in 1Hz Bandwidth = sideband level ____ ΔV^2_{rms}

− normalization to 1Hz ____

− S/A correction ____

$$S_{\Delta f}(f) = \frac{\Delta V^2_{rms}(f)}{K_d^2} \quad \frac{Hz^2}{Hz} =$$

After measuring the spectral density of the voltage fluctuations, the measured values are converted to the proper units and corrected. First, since $S_{\Delta f}(f)$ is defined on a per Hertz basis, the measured noise is normalized to a 1 Hz noise bandwidth. The 2.5 dB spectrum analyzer correction is taken into account if necessary. But how do we convert the measured noise (in dBm) into units of Hz^2/Hz?

COMPUTING $S_{\Delta f}$ (f) FROM S/A DISPLAY

1. Establish Known cal. sideband/carrier = $\mathcal{L}_{cal}$ = ____ dBc

2. Translate to $\mathcal{L}_R$ = antilog $\mathcal{L}_{cal}/10$ = ____ $\frac{Watt}{Watt}$

3. Find Δf_{cal} in rmsHz

   $\Delta f_{cal} = f_m \sqrt{2\mathcal{L}_R}$ = ____ rmsHz

4. Translate Δf_{cal} to dBHz

   Δf_{cal} (dBHz) = 20log Δf_{cal} = ____ dBHz

5. Measure P_{cal} = ____ dBm

6. Find scale factor

   SF(dB) = Δf_{cal} (dBHz) − P_{cal} (dBm) = ____ dBHz/dBm

7. Measure P_{noise} (in 1Hz Bandwidth)

8. $S_{\Delta f}(f)$ (dB) = P_{noise} + SF

9. $S_{\Delta f}(f) \left(\frac{Hz^2}{Hz}\right)$ = antilog $\frac{S_{\Delta f}(f)\,(dB)}{10}$ = ____ $\frac{Hz^2}{Hz}$

If units of $S_{\Delta f}(f)$ are desired, the data manipulations are shown here. They look more complicated then they are, because the data has to be translated from log form to absolute terms and back again several times.

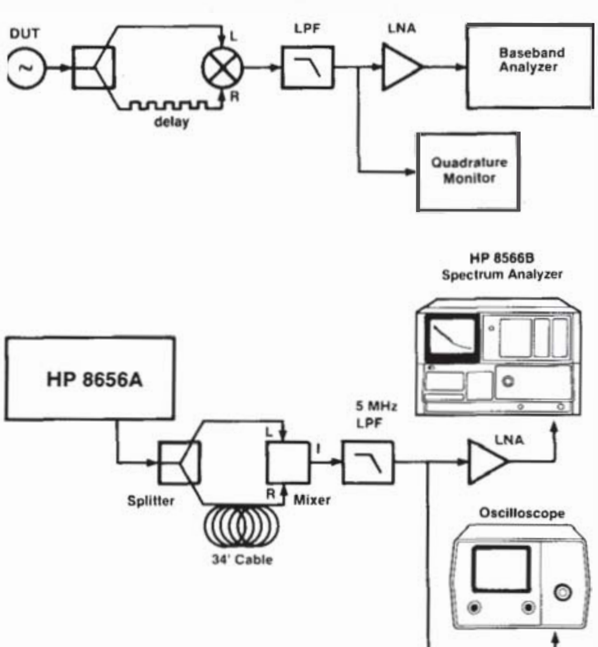
CONVERT TO $\mathcal{L}(f)$ IF DESIRED

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

	10 kHz		100 kHz
$\mathcal{L}(f)$ = measured noise level	—	(dBm)	—
- detected calibration level	—	(dBm)	—
- set sideband to carrier ratio	—	(dBc)	—
Bandwidth Normalization	—	(dB)	—
S/A CORRECTION (if applicable)	—	(dB)	—
$- 20 \log f_m/f_{cal}$	—	(dB)	—
$\mathcal{L}(f)$ =	—	(dBc/Hz)	—

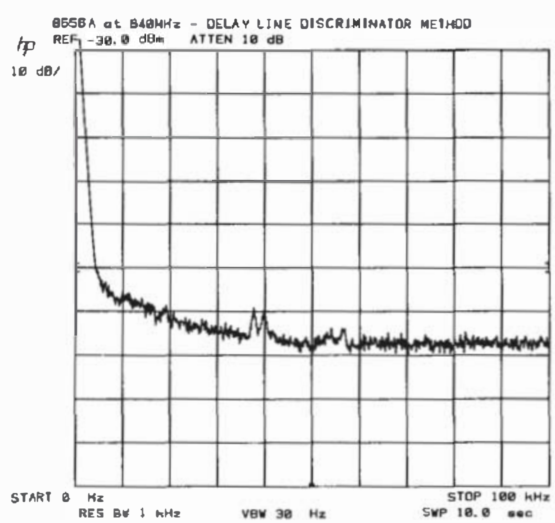
Since we are using a spectrum analyzer (which measures power) as our baseband analyzer, it is much easier to translate the measured voltages into $S_{\phi}(f)$ or $\mathcal{L}(f)$, as shown here. Note the correction factor to translate the calibration term from the calibration offset frequency to a given measurement frequency. This term comes from the $1/f^2$ in the translation of $S_{\Delta f}(f)$ to $\mathcal{L}(f)$.

DEMONSTRATION



Delay Line Mixer
Frequency Discriminator Method

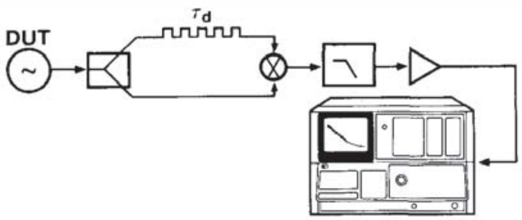
TYPICAL MEASUREMENT RESULTS



HP 8656A at 640 MHz

Delay Line/Mixer
Frequency Discriminator Method

CONSIDERATIONS IN SYSTEM ACCURACY



CONTRIBUTOR	TYPICAL UNCERTAINTY (±dB)
1. Accuracy of power level measurements (relative)	0.05
2. Spectrum analyzer accuracy (relative)	0.4–1.5
3. Attenuator accuracy of calibration	0.06
4. Mismatch uncertainty associated with setting CW reference level	0.15
5. Uncertainty in ±2.5 dB correction	0.20
6. Accuracy of measured IF noise bandwidth	0.20
7. Relative IF bandwidth gains	0.05
8. Analyzer frequency response	0.25

Here's a fairly detailed analysis of potential error sources, with an estimation of typical values. The major contributor to system accuracy is the accuracy of the baseband analyzer; this is kept small by using the analyzer in a relative mode (ie, by measuring the noise relative to the calibration level). The flatness of the baseband signal processing section can also contribute significantly to overall system accuracy. Better accuracy can be achieved if this contribution is measured and then compensated for.

CONSIDERATIONS IN SYSTEM ACCURACY (Continued)

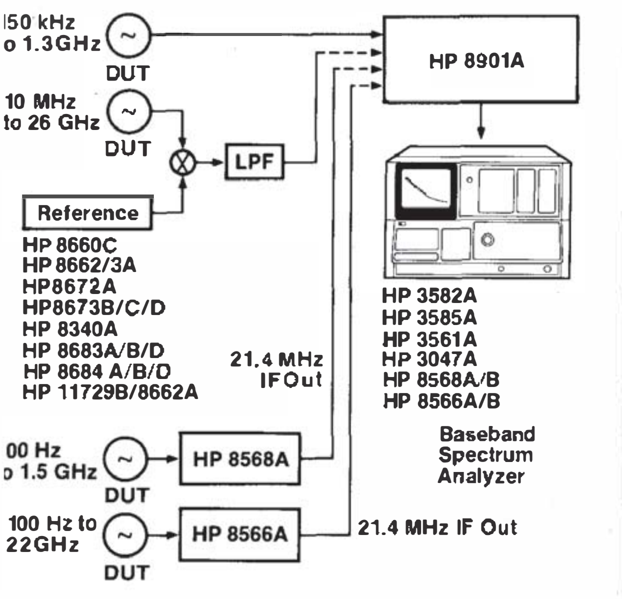
CONTRIBUTOR	TYPICAL UNCERTAINTY (±dB)
9. Frequency discriminator flatness	0.30
10. Baseband signal processing flatness	0.50–1.0
11. Random error due to randomness of noise	0.50
12. Mixer dc offset/quadrature maintenance	0.03
13. Accuracy of setting mod index or sideband carrier	0.10

TOTAL UNCERTAINTY (±dB)

	Minimum	Maximum
Linear Summation	2.79	4.39
RSS	0.84	1.53

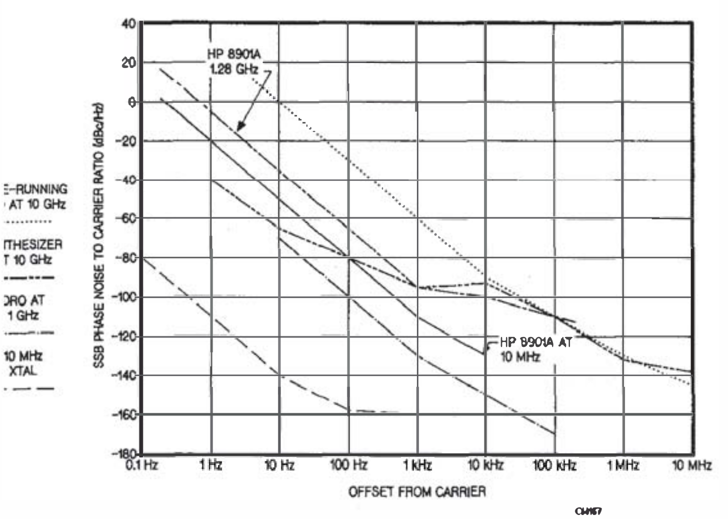
Taking all these inaccuracies into account, an overall system accuracy can be determined.

USING THE HP 8901A MODULATION ANALYZER AS A FREQUENCY DISCRIMINATOR



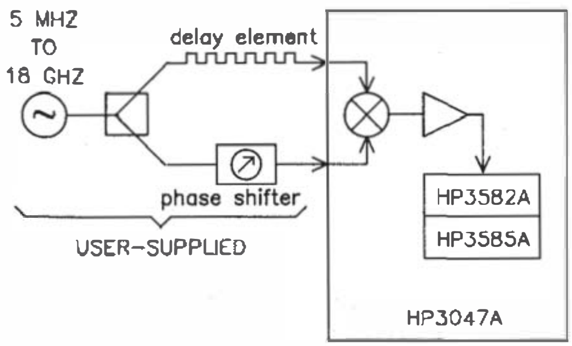
The simplest frequency discriminator implementation is the HP 8901A. It can be used directly with a baseband spectrum analyzer for sources to 1.3 GHz, or higher frequency sources can be first downconverted and then input. Alternatively, the IF output from the HP 8568A/B or HP 8566A/B can be used to drive the HP 8901A.

TYPICAL SENSITIVITY OF THE HP 8901A USED AS A FREQUENCY DISCRIMINATOR



The HP 8901A, combined with a downconverter if necessary, is a simple solution providing sufficient sensitivity for many sources.

FREQUENCY DISCRIMINATOR METHOD – IMPLEMENTATION HP 3047A



- discriminate at carrier frequency

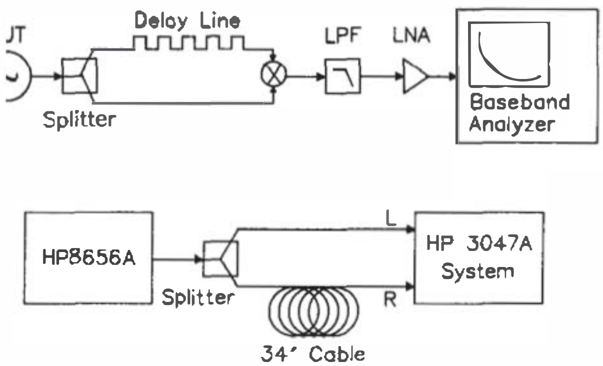
For better sensitivity, the HP 3047A can be used in a frequency discriminator mode (delay line/mixer used as a frequency discriminator). The DUT is externally split, one side is delayed, and both the delayed and the reference channel are input to the system. The software leads the user thru the measurement, allowing calibration on a known sideband, or inputting of the discriminator constant K_d.

ADVANTAGES OF HP 3047A SYSTEM USED IN FREQUENCY DISCRIMINATOR METHOD

- Automatic calibration, conversion of units
- Hard–copy output
- Sensitivity follows spectra of free–running source
- Only one source required

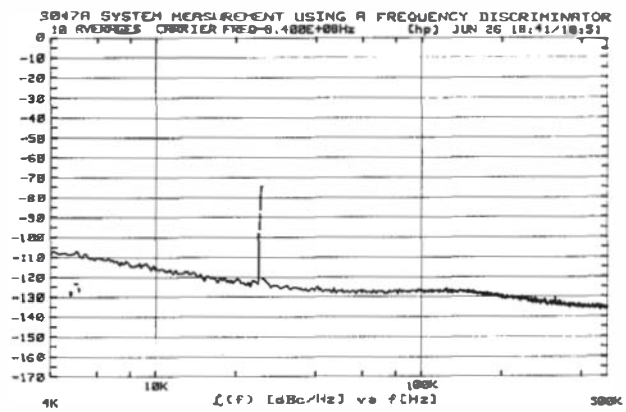
Using the HP 3047A in the frequency discriminator mode allows the data collection and translation to be done automatically. The HP 3047A contains both a microwave phase detector and an RF phase detector.

DEMONSTRATION



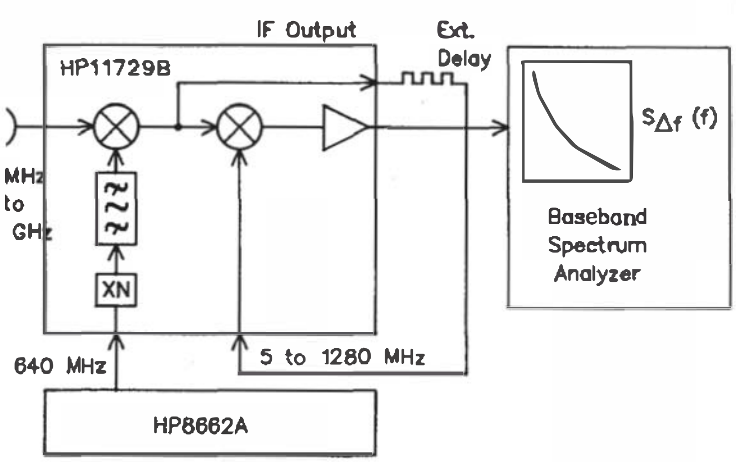
Delay Line/Mixer
Frequency Discriminator Method
Using the HP 3047A System

TYPICAL MEASUREMENT RESULTS



HP 8656A at 640 MHz
Delay Line/Mixer
Frequency Discriminator Method
Using the HP 3047A System

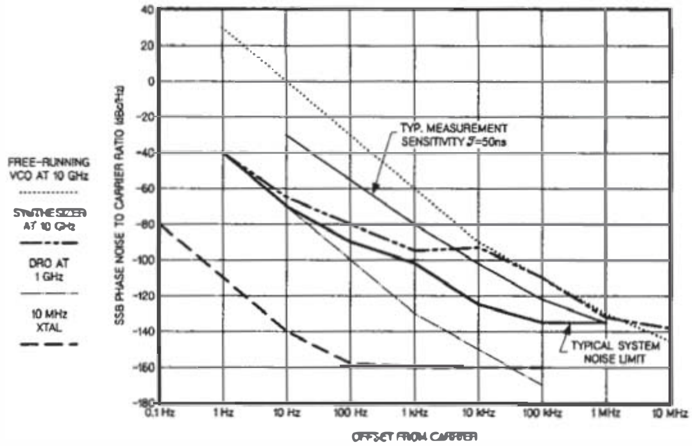
FREQUENCY DISCRIMINATOR METHOD– IMPLEMENTATION HP 11729B/8662A



Discriminate at HP11729B IF

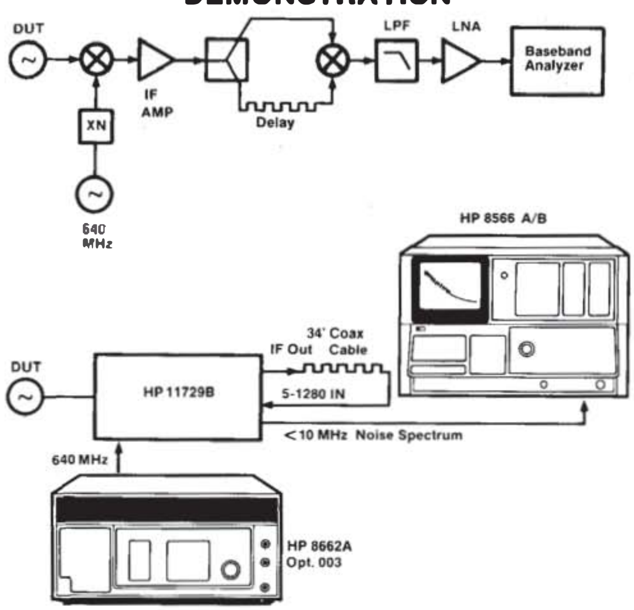
The HP 11729B can also be used in the frequency discriminator mode. A single, fixed frequency 640 MHz signal is still required. The DUT is downconverted, and the discriminator placed at the IF. The necessary power splitter and phase detector are already provided. The onl external piece of hardware needed is a simple length of cable, which can be as inexpensive as a piece of RG 55 cable, since the highest operating frequency is less than 1.3 GHz. The HP 8662A with internal FM modulation ca be used as a convenientcalibration signal.

TYPICAL SENSITIVITY OF HP 11729B/8662A IN FREQUENCY DISCRIMINATOR METHOD



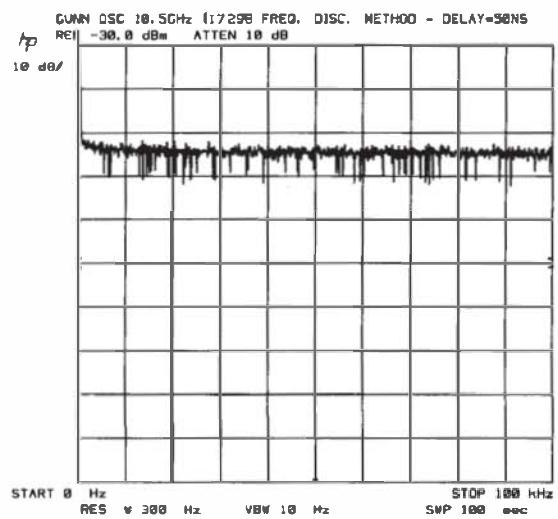
Typical achievable system sensitivity is shown
34 ft.-length of coaxial cable was used as the dela
element. Of course, longer delay lines would yield
increased sensitivity, but reduced offset frequenc

DEMONSTRATION



Delay Line/Mixer
Frequency Discriminator Method
Using the HP 11729B/8662A

TYPICAL MEASUREMENT RESULTS



Delay Line/Mixer
Frequency Discriminator Method
Using the HP 11729B/8662A

CONVERT TO $\mathcal{L}(f)$ IF DESIRED

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

	10 kHz		100 kHz
$\mathcal{L}(f)$ = measured noiselevel	—	(dBm)	—
- detected calibration level	—	(dBm)	—
- set sideband to carrier ratio	—	(dBc)	—
Bandwidth Normalization	—	(dB)	—
S/A CORRECTION (if applicable)	—	(dB)	—
$- 20 \log f_m/f_{cal}$	—	(dB)	—
$\mathcal{L}(f) =$	—	(dBc/Hz)	—

HP 11729B/8662A IN FREQUENCY DISCRIMINATOR METHOD

ADVANTAGES

- Increased sensitivity can be obtained with long line without problem of loss
- HP 8662A can be used for calibration
- Built in quadrature monitor
- Broadband discriminator
- Delay line at IF less expensive to implement than delay line at microwave

DISADVANTAGES

- Limited to noise floor of multiplied 640 MHz
- Quadrature set by varying DUT frequency or length of line

The HP 11729B/8662A in the frequency discriminator mode provide an easy solution for measurements on unstable or free-running sources. It can be used to measure sources to 18 GHz, yet the discriminator always operates less then 1.3 GHz. The one disadvantage of HP 11729B/8662A implementation over a true 'single source' technique is that the measurement is limited to the noise floor of the multiplied 640 MHz signal.

FREQUENCY DISCRIMINATOR METHOD

ADVANTAGES

- Requires only one source
- Low broadband noise floor
- Sensitivity matches free running VCO characteristic
- AM Suppression
  ≈30 dB to 1.5 GHz
  ≈15 dB > 1.5 GHz

DISADVANTAGES

- Poor close–in sensitivity
- May be difficult to implement delay–line discriminator at microwave due to loss of line and microphonics
- Other sensitive implementations at microwave are usually narrowband

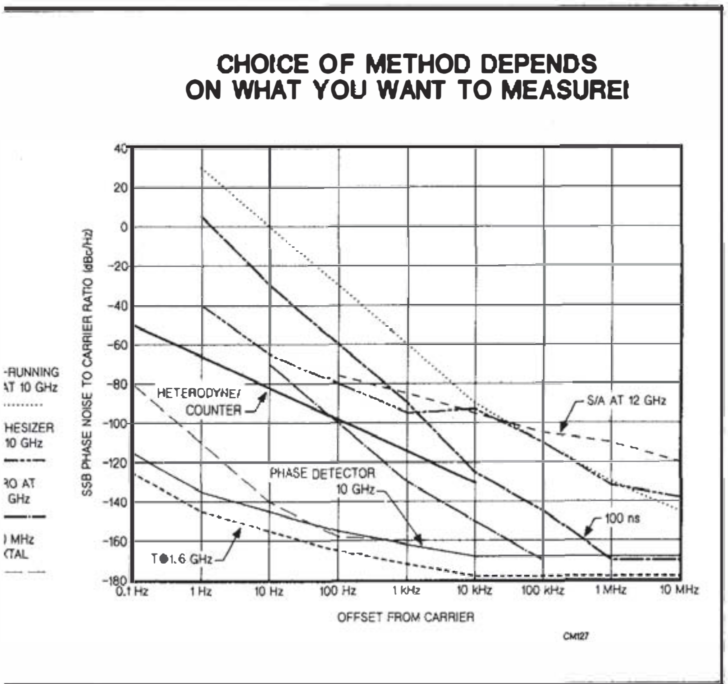
In summary, the primary advantages of the frequency discriminator method are shown here. In practice, it may be difficult to implement the delay line discriminator at microwave frequencies, as longer delay will improve the sensitivity but eventually the loss in the delay line will exceed the source power available and cancel any further improvement. Also longer delay lines limit the maximum offset frequency that can be measured.

5. Summary of Source Measurement Techniques

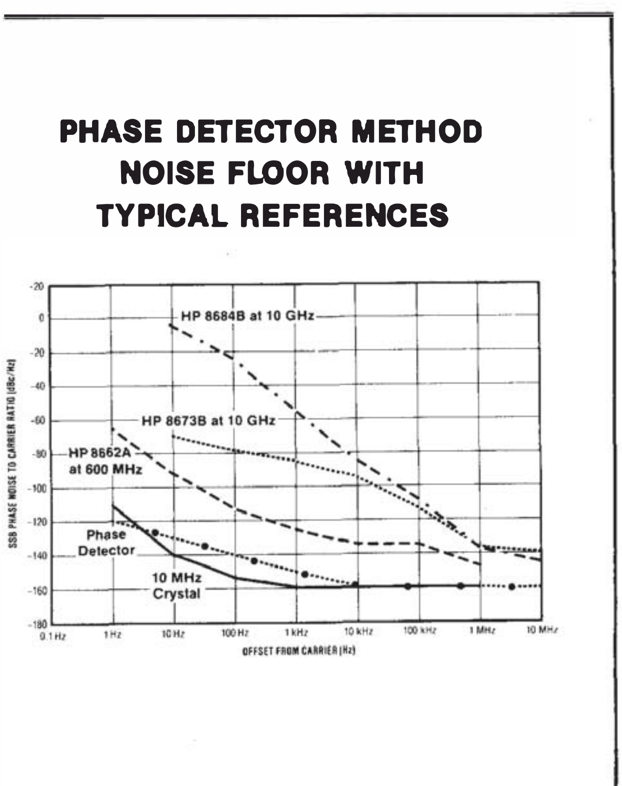
SUMMARY

Q: So now I know all about Phase Noise.
But which method do I use?

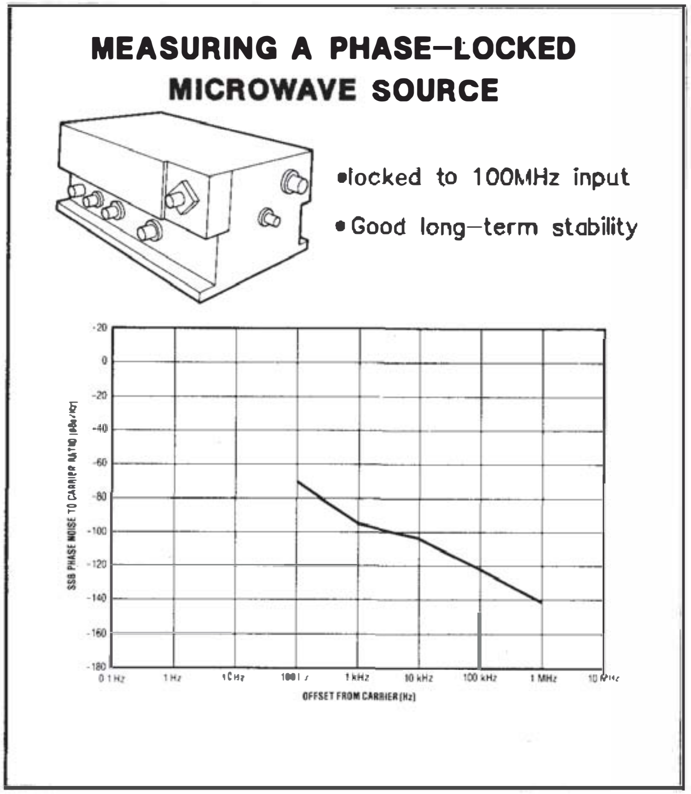
A: That depends!



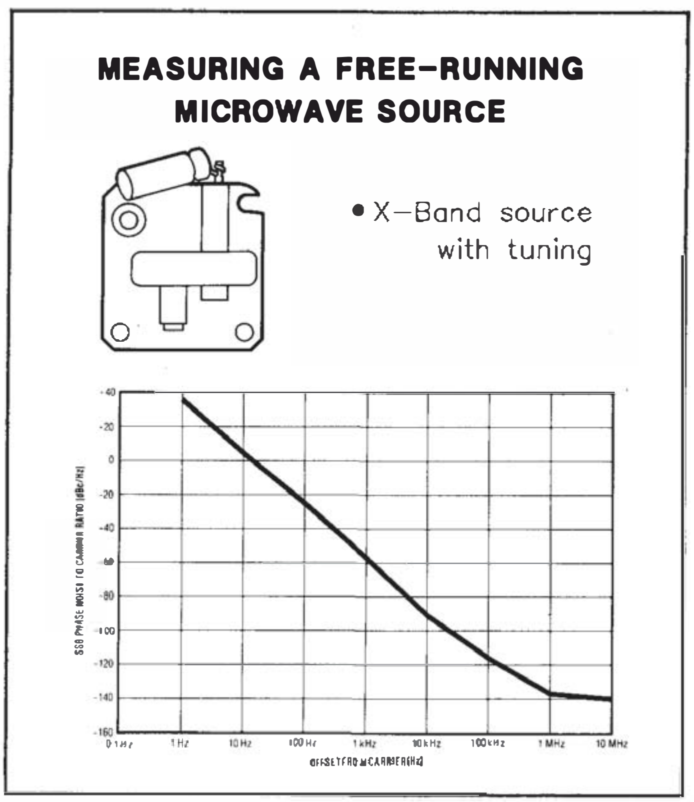
The first criterion for deciding on a particular method c phase noise measurement is how low the source or device noise is that is to be measured. Here again is a comparison of the sensitivities of the four phase noise measurement methods. Direct spectrum method is easiest if the source is stable and the noise is fairly high; time domain has limited usefulness. Frequency discriminator sensitivity follows the spectra of free-running sources; the phase detector method can have the lowest overall *sensitivity*, depending on the reference source used.



To determine usable sensitivity in the phase detector method, the reference source must be specified. Here are low noise sources usable for references, from 10 MHz to 10 GHz.

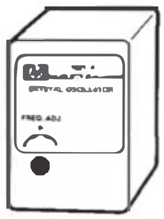


Given what you know about the performance and complexity of the measurement methods, which method would you recommend to test the phase noise of a phase-locked microwave source with this level of expected phase noise?

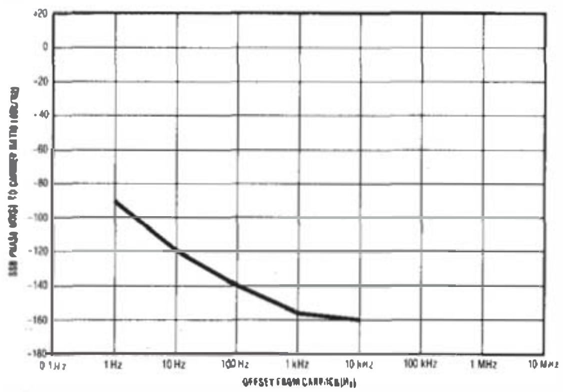


Which technique would you choose for this Gunn diode VCO? Why?

MEASURING A HIGH-QUALITY LOW FREQUENCY OSCILLATOR



- High stability 10 MHz crystal
- Good long-term stability (<ppm/day)



What about this high quality quartz oscillator?

GENERAL SYSTEM CONSIDERATIONS

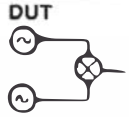
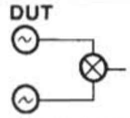
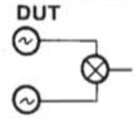
or

how to make things easier with careful set-up

- Heat sinking of source
- Microphonic isolation of source
- Reduce frequency pulling effects on source with circulator
- Choose cable to reduce microphonics
- The RFI of the environment
- "Clean" power supplies
- Resonant frequency of building!?

With a better understanding of the fundamentals and mathematics of phase noise measurement, don't forget that the real world is still out there! Experience in making phase noise measurement quickly teaches that careful set-up and system environment are critical to good measurements.

Phase Noise Measurement Systems Comparison

	HP 11729B Phase Noise Test System	HP 3047A Phase Noise Measurement System	HP 3047A/11729B μωPhase Noise Measurement System
	HP 8566A/B HP 8568A/B HP 3585A HP 3582A HP 3561A Spectrum Analyzers	HP 3585A Spectrum Analyzer	HP 3585A Spectrum Analyzer
	HP 11729B Phase Noise Test Set	HP 3582A Spectrum Analyzer	HP 3582A Spectrum Analyzer
	HP 8662A or HP 8663A Synthesized Sig Gen	HP 35601A S.A. Interface	HP 35601A S.A. Interface
		HP 8662A HP 8663A HP 8640A HP 8672A HP 8673B Reference Sources	HP 11729B Phase Noise Test Set
			HP 8662A or HP 8663A Synthesized Sig Gen
	Controller	HP 9836A Controller	HP 9836A Controller
Principle of Operation	DUT  Reference	DUT  Comparable Source	DUT  Reference
Frequency Range	.005 — 18 GHz	.005 — 18 GHz	.005 — 18 GHz
Low Noise Microwave Reference Signal	yes	no	yes
System Noise Spec at 10 GHz	-120 dBc at 10 kHz -132 dBc noise floor	-160 dBc at 10 kHz* -160 dBc noise floor	-120 dBc at 10 kHz -132 dBc noise floor (with 11729B)
Noise Spectrum (offset frequency)	<10 MHz	<40 MHz	<10 MHz (<40 MHz)
Two-port Noise	no	yes	yes
Delay Line Mode	yes	yes	yes
AM-Noise (Microwave)	yes	w/external detector	yes
Manual Operation	yes	no	yes
Automatic Operation (HP-IB)	yes	yes	yes
Software	User Provided	extensive	extensive

*Specification shown requires a second, user supplied source of comparable phase noise.

This compares some of the performance parameters of HP's two-source phase noise measurement systems.

III. Measurement of Two Port Phase Noise

SEMINAR AGENDA

I. Basis of Phase Noise
- Why is Phase Noise important?
- What is Phase Noise?
- What causes Phase Noise?
- Quantifying Phase Noise

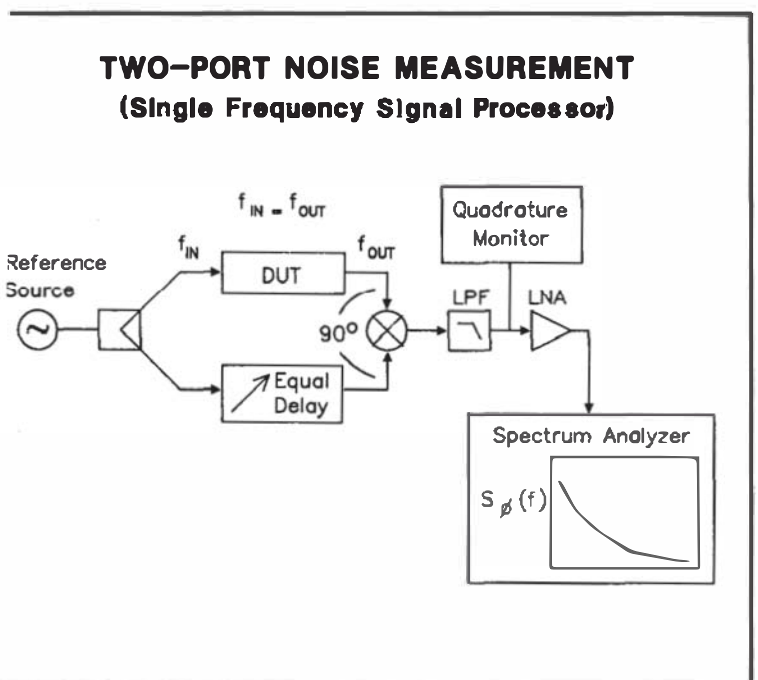
II. Measurement Techniques on Sources
1. Direct Spectrum Method
2. Heterodyne/Counter Method
3. Phase Detector Method
4. Frequency Discriminator Method
5. Summary of Source Measurement Techniques

⇨ III. Measurement of Two-Port Phase Noise (Devices)

IV. Phase Noise Measurement on Pulsed Carriers

V. AM Noise Measurement

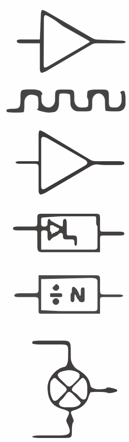
Turning to phase noise measurements on devices, our selection of measurement method is significantly easier. Again, the phase noise added or contributed by a device is called two-port, or residual, or additive noise.



Two-port phase noise characterization is done with a modified frequency discriminator method. Actually, it's done exactly like the noise floor test for the frequency discriminator or phase detector method. A common reference source is split and applied to two channels simultaneously. Since the goal is to have the noise of the source common to both channels and arrive correlated at the phase detector, the delay in the two paths is held as close to the same as possible. The signals are set in quadrature with a phase shifter, and the noise contribution of the DUT can be measured.

TYPICAL DUT'S WHERE TWO-PORT NOISE IS MEASURED

- TWT's
- SAW delay lines
- amplifiers
- multipliers
- dividers
- mixers

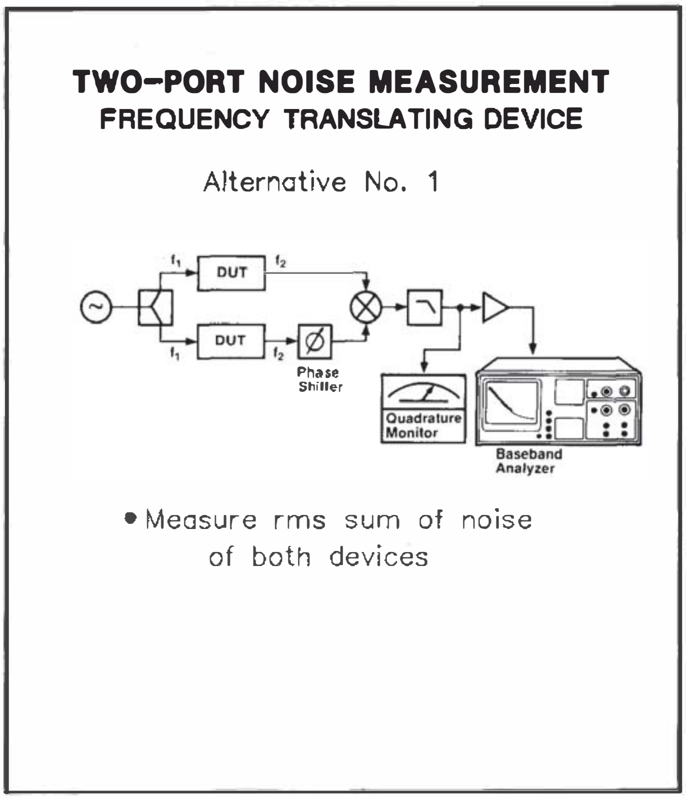


Two-port noise measurements are made on a wide variety of sources, with measurements on TWT's one of the most common, and measurements on SAWs gaining as SAW technology advances.

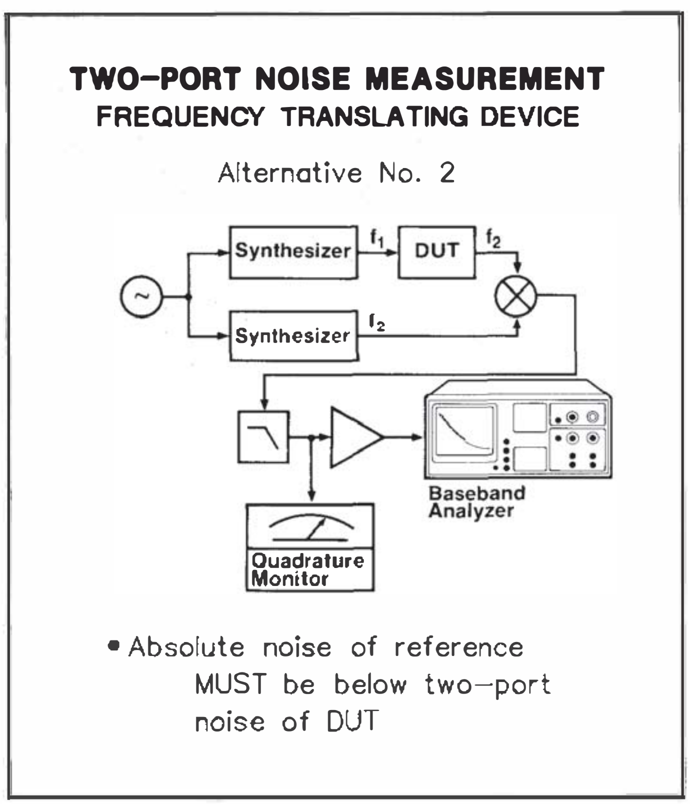
TWO-PORT NOISE MEASUREMENT CONCERNS/LIMITATIONS

1. AM << ØM
2. Typical AM rejection
   ≈ 30 dB to 1.5 GHz
   ≈ 15 dB > 1.5 GHz
3. Noise floor same as phase detector method
4. Time delay in unknown path
5. Filtering effects of unknown path

Some of the areas of caution for a two-port measurement are the same as for an absolute measurement. However, because of the nature of two-port noise, some new concerns arise. Typically, two-port noise of devices is much, much less than the absolute noise of sources, and it does not increase linearly as a function of carrier frequency (nor does it necessarily follow the typical f^{-2} and f^{-3} relationships of sources as the carrier is approached). Also, the two-port AM noise of a device might be of the same order of magnitude as its phase noise. This means even more attention must be paid to system noise floor. For example, typically a good quality source is used as the reference, even though ideally all source noise will be correlated out.

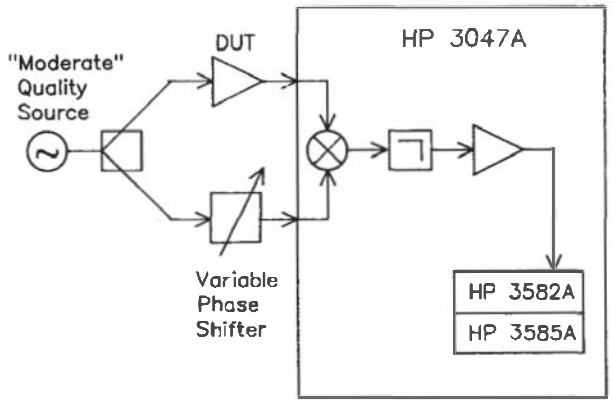


If the output frequency of the DUT does not equal the input frequency, then it is impossible to directly measure the two-port noise with a single measurement. Commonly, a similar DUT (with assumed similar noise) is placed in the reference path as well. The rms sum of noise of both devices is then measured. As in the phase detector method when using a similar source as a reference, the measurement sets an upper limit on the noise of each device. At each offset frequency, one of the devices is at least 3 dB better. And if three devices with similar noise can be measured, the noise of one of the devices can be calculated.



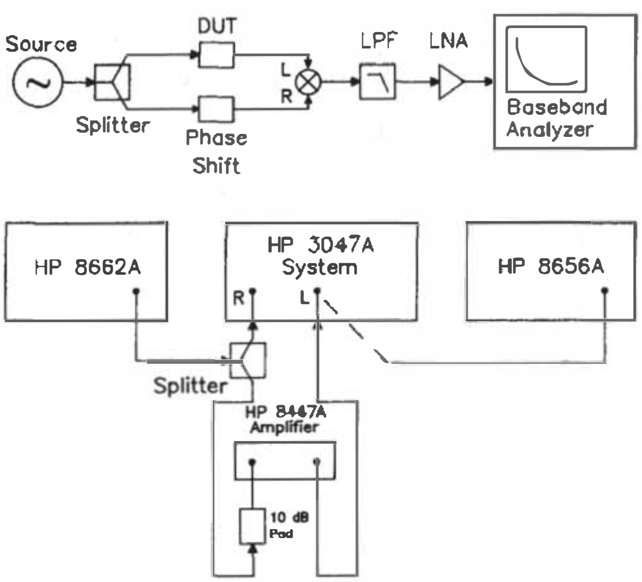
Another alternative to measuring the two-port noise of a frequency translating device is to use a synthesizer in each path, instead of a second DUT. The limitation here is that the absolute noise of the synthesizers must be below the two-port noise of the DUT. This might be workable sometimes at RF frequencies. However, for microwave frequencies, the absolute noise of sources is typically much higher than the two-port noise of devices.

MAKING TWO-PORT PHASE NOISE MEASUREMENTS WITH THE HP 3047A



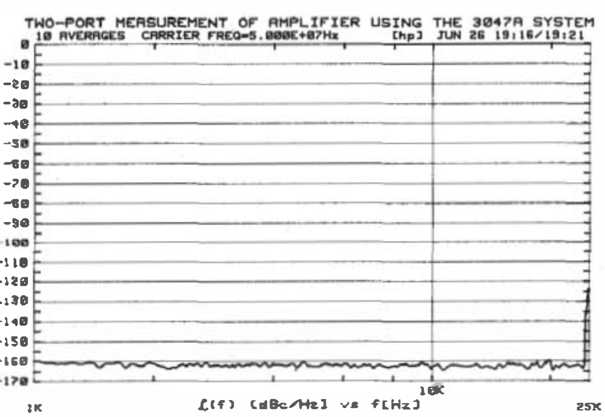
For two-port phase noise measurements, HP has really only one solution – the HP 3047A Phase Noise Measurement System. The HP 3047A has been widely and successfully used for two-port noise measurements, as its high level mixers yield a low system noise floor. (The HP 11729B can not be used for two-port microwave phase noise measurements because the microwave mixer is not dc coupled on the IF, nor is there access to both ports of the mixer.)

DEMONSTRATION



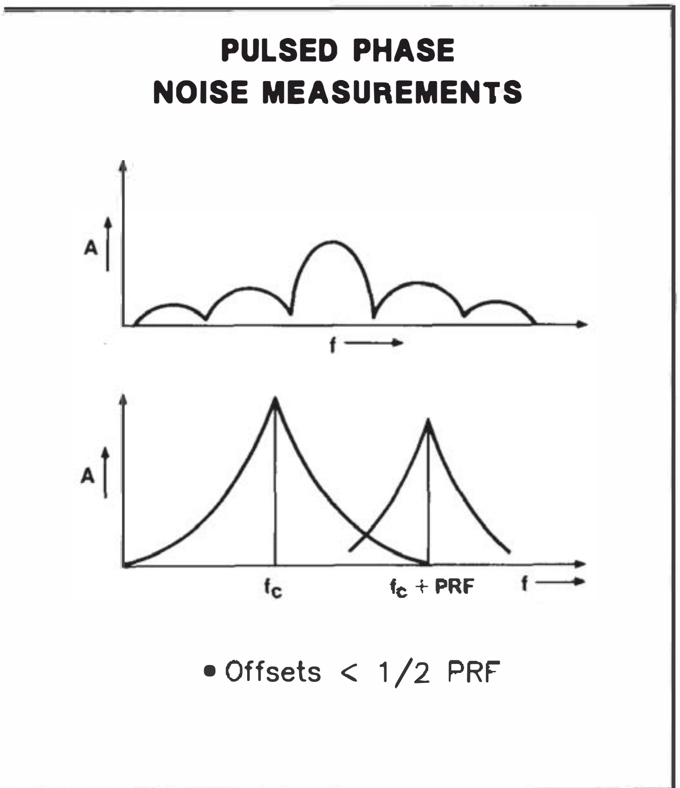
Two-Port Measurement Using the HP 3047A System

TYPICAL MEASUREMENT RESULTS

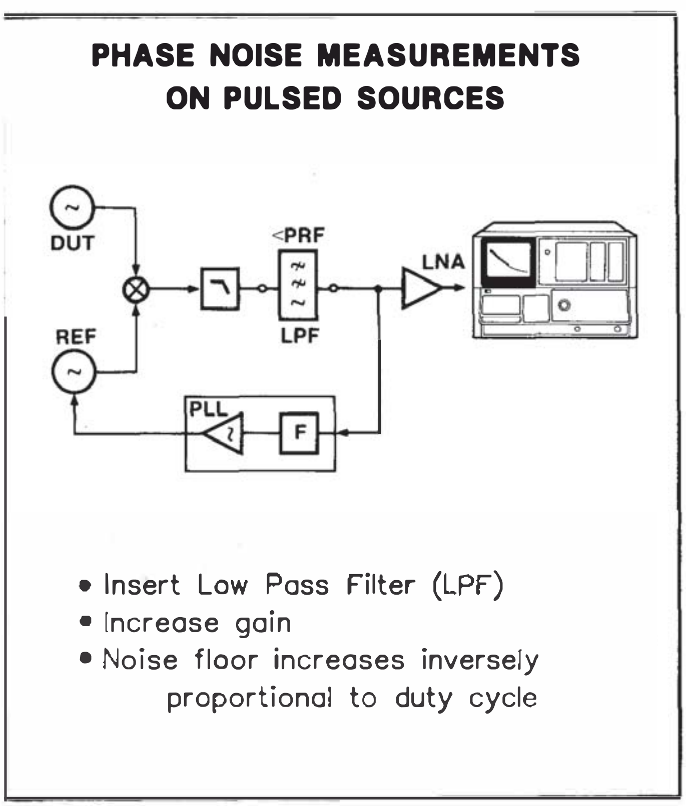


Two-Port Phase Noise Measurement Using the HP 3047A System

IV. Phase Noise Measurements on Pulsed Carriers

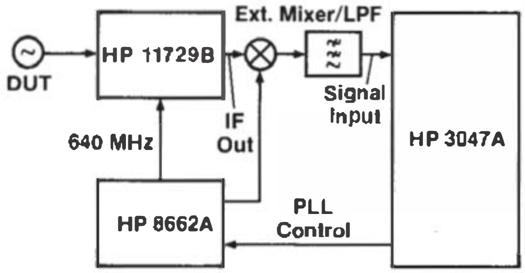


Making phase noise measurements on pulsed sources brings up a new set of complications. First, it is important to recognize that, since the noise spectrum is repeated on each spectral line, phase noise measurements are only valid to less than ½ the PRF. (If the first PRF line was equal to the carrier in amplitude, a phase noise measurement at ½ the PRF would yield noise data 3 dB too high.)



Frequency discriminators do not lend themselves easily to pulsed phase noise measurements, though some work has been successfully done (eg, by F. Labarr at TRW). For HP's hardware implementations, pulsed phase noise measurements can be done with some limitations with the phase detector method. In general, it is necessary to insert a LPF to remove the PRF feedthrough from feeding thru to the LNA, baseband analyzer, or phase lock loop. Also, because the test signal is now only present part of the time, it is usually necessary to increase the gain in the phase lock loop to maintain quadrature. Finally, since the reference is on all the time (but the DUT pulsed), the system noise floor increases as a function of duty cycle.

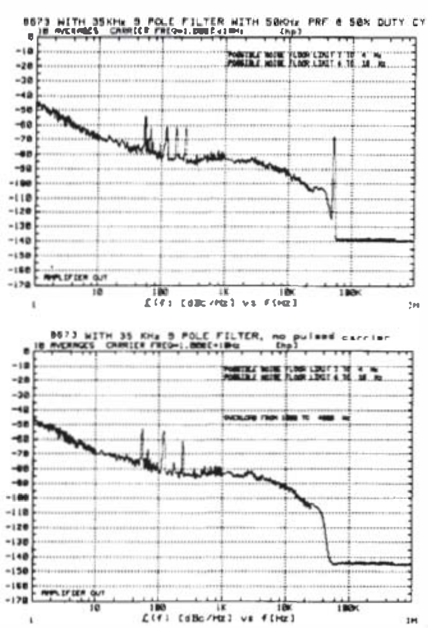
PULSED PHASE NOISE MEASUREMENTS ON SOURCES WITH HP 11729B/8662A/3047A



- Select K_o, external mixer
- Calibrate on pulsed signal
- Duty cycles down to ≈ 20%

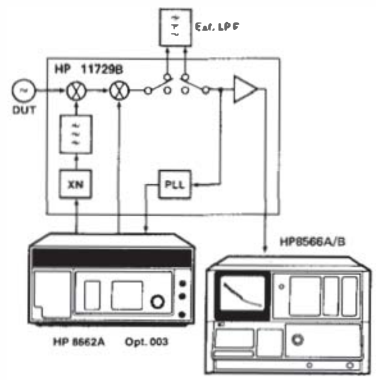
It is possible to make phase noise measurements on pulsed sources with the HP 11729B/8662A/3047A. In order to use the standard units without internal modifications, an external phase detector is used at the HP 11729B IF output. The L port drive is provided by the front panel signal of the HP 8662A. After the phase detector, a low pass filter removes the sum mixing products and the PRF lines, and the resulting signal is applied to the HP 3047A Signal Input port. This configuration allows measurement on pulsed sources with duty cycles down to about 20%.

TYPICAL PERFORMANCE OF PULSED HP 11729B/8662A/3047A



Measured output from an HP 11740S system (HP 11729B/8662A/3047A). A 10 GHz source, 400 kHz PRF, with 50% duty cycle was measured with a 250 kHz LPF inserted to remove the PRF. The measured values on the pulsed source are indistinguishable from the measured values when operating the source in CW.

PULSED PHASE NOISE MEASUREMENTS WITH HP 11729B/8662A



- External LPF switched in
- Duty cycles down to <5%
- Disables out-of-phase lock indicator
- Rising noise floor and decreasing offset frequencies as a function of duty cycle

For measurements on pulsed sources with lower duty cycles, the HP 11729B/8662A can be used. It is necessary to slightly reconfigure the HP 11729B so that the LPF can be placed before the low noise amplifier and the phase lock loop. (There are ports at the rear panel of the HP 11729B for the LPF; however, internal cable rerouting must be done to switch the LPF into the signal path.) Also, for very low duty cycles, a small external voltage must be applied to the 'Auxiliary Noise Spectrum Output' to adjust the balance of the diodes in the phase detector. With this configuration, phase noise measurements on pulsed sources with duty cycles down to <5% have been successfully made.

Again, it's important to remember that noise floor increases as a function of duty cycle. Also, at very low duty cycles, eventually the noise floor of components in the HP 11729B will dominate (in particular, the noise of the IF amplifier).

WHAT ABOUT LOWER DUTY CYCLES?

- Sample and hold
- Modified delay line technique (TRW)

For lower duty cycles, there has been experimentation with Sample and Hold circuitry, and a modified delay line technique. However, these are not implemented in HP instrumentation.

PULSED "TWO–PORT" PHASE NOISE MEASUREMENTS WITH HP 3047A

- Switch out LNA
- Use high power inputs
- Calibrate on pulsed signal
- Duty cycles down to ≈ 10%

The HP 3047A can also be used for pulsed two-port phase noise measurements. Typically, it is necessary to switch out the internal HP 3047A low noise amplifier. If calibration is done on a pulsed signal, no corrections to the measured data needs to be made. Successful measurements have been made to duty cycles less than 20%.

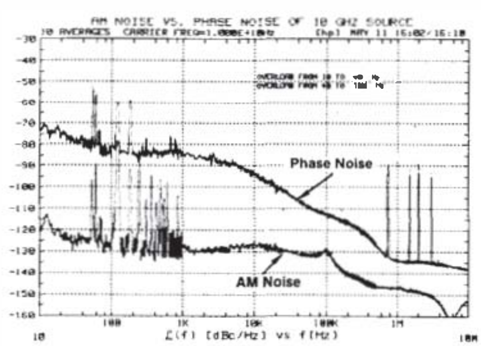
V. AM Noise Measurement

AM NOISE MEASUREMENTS

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

Corresponding to the $\mathcal{L}(f)$ definition, $\mathcal{M}(f)$ is defined as the noise power in one AM modulation sideband, divided by the total signal power, in units of dBc/Hz.

AM NOISE vs. PHASE NOISE

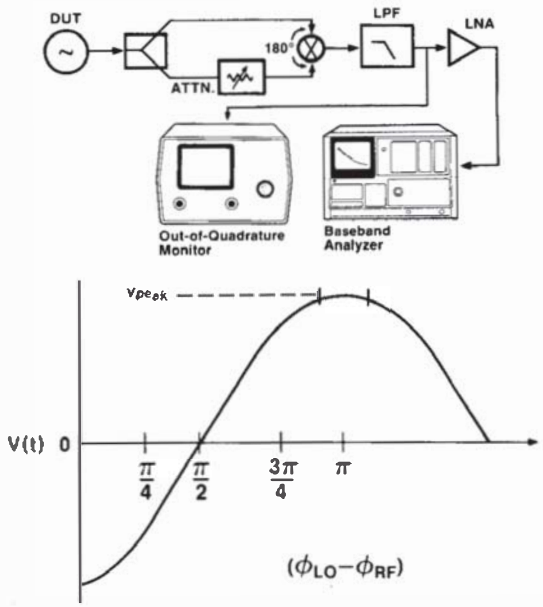


Though AM noise usually measured on signals, in practice phase noise is more important (both technically and economically) than AM noise for three reasons.

1. The majority of high capacity (ie, expensive) communication systems use angle modulation.
2. AM noise in complex signal sources is usually <<phase noise.
3. The FM noise or phase noise on the Local Oscillator is directly transferred to the input signal, whereas the effects of the LO AM noise may be greatly mitigated by the use of balanced mixers.

AM NOISE MEASUREMENTS

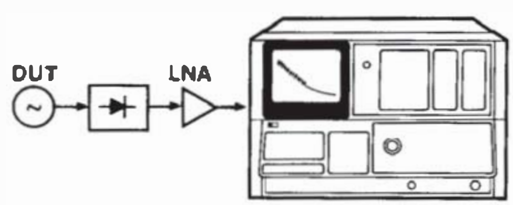
Method #1 – Using a Double-Balanced Mixer as an AM Detector



There are two common methods of AM noise measurement; the first is the corollary to the phase detector method of phase noise measurement. Here, a double-balanced mixer is operated 180 degrees out of phase (out-of-quadrature condition), to yield maximum sensitivity to AM fluctuations and minimum sensitivity to phase fluctuations.

AM NOISE MEASUREMENTS

Method #2 – Using An AM Detector



The second method (the corollary of the direct spectrum technique) simply measures the AM fluctuations with an AM detector, and then looks at the baseband fluctuations.

AM NOISE MEASUREMENT PROCEDURE

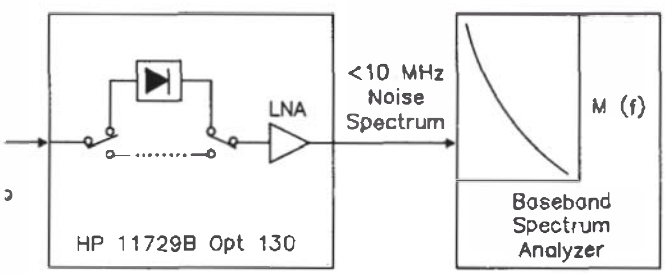
1. Calibrate
   on known level sideband/carrier ratio, in out-of-quadrature condition
2. Measure V_{noise}
3. Corrections

$\mathcal{M}(f) =$ V_{noise} (dBm)
− V_{cal} (dBm)
+ sideband/carrier cal
+ ATTN.
(used in method #1)
− $10 \log B_n$
+ S/A effects

dBc/Hz

Both techniques have similar procedures, and are fairly straightforward.

AM NOISE MEASUREMENTS



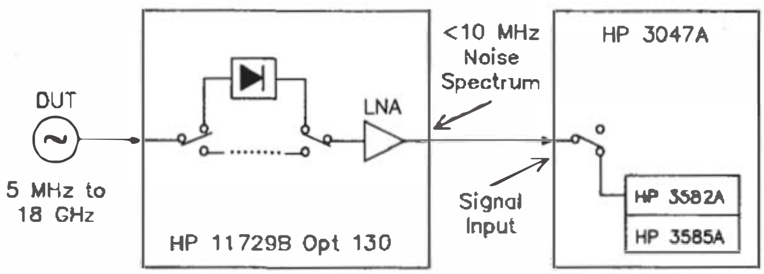
ch-Top" system with HP 11729B Opt. 130/8662A

) Calibrate with HP 8662A w/AM Sidebands

) Measure

Direct AM noise measurements can be made with the HP 11729B Option 130. The HP 8662A with known AM sidebands is a convenient calibration source.

AM NOISE MEASUREMENTS

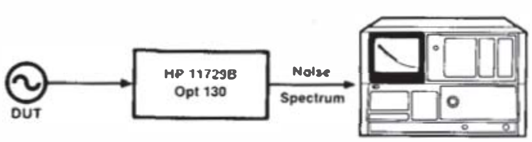


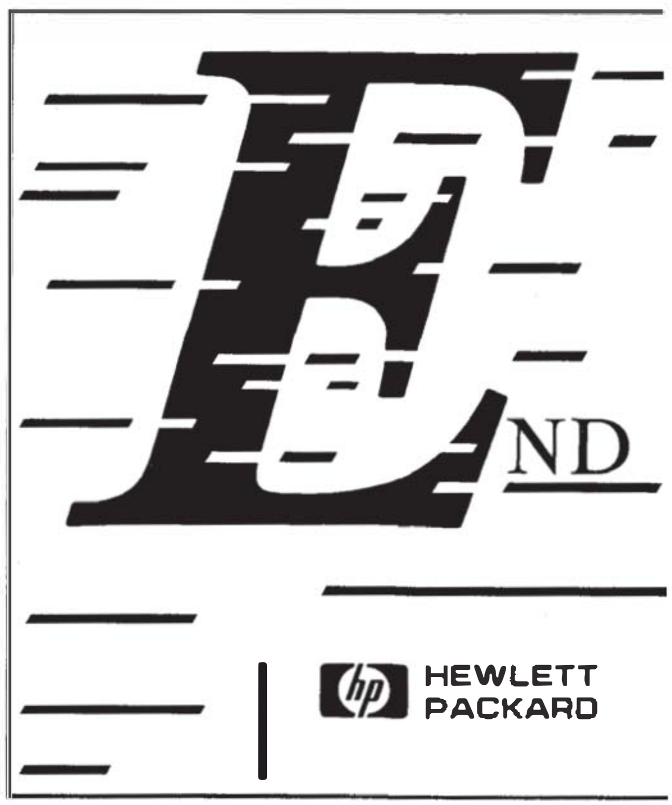
Automatic system with HP 11729B Opt 130/8662A/3047A

1) Select "Phase Noise w/o voltage control"
2) Calibrate with HP 8662A w/AM sidebands
3) Measure

The HP 11729B Option 130 can also be used automatically with the HP 3047A software. If prop calibrated, the system software for phase noise measurements can also be used for AM noise measurements.

DEMONSTRATION





GLOSSARY OF SYMBOLS

SYMBOL	UNITS	DESCRIPTION
f_o	Hz	Carrier frequency
f	Hz	Fourier frequency (sideband, offset, baseband)
f_m	Hz	Modulating frequency
f_c	Hz	Corner frequency
$f(t)$, $\nu(t)$	Hz	Instantaneous frequency
$\Delta f(t), \delta f(t), \delta\nu(t)$	Hz	Instantaneous frequency fluctuations
$\Delta\phi(t), \delta\phi(t)$	rad	Instantaneous phase fluctuations
τ_d	sec	Delay time
K_ϕ	Volts/rad	Phase detector constant
K_d	Volts/Hz	Frequency discriminator constant
$S_\phi(f)$	rad^2/Hz	Spectral density of phase fluctuations
$S_{\Delta f}(f)$	Hz^2/Hz	Spectral density of frequency fluctuations
$S_y(f)$	Hz^{-1}	One-sided spectral density fractional frequency fluctuations
$L(f)$	dBc/Hz	Single-sideband signal to carrier ratio
res FM	Hz	Residual FM
$M(f)$	dBc/Hz	Spectral density of AM noise
V_{bpeak}	Volts	Peak voltage of sinusoidal beat signal
V_{ssb}	Volt	Single sideband voltage
V_s	Volt	Signal voltage
$S_v rms$	$Volts^2/Hz$	Power spectral density of voltage fluctuations
β	(rad)	Modulation index
J_n		Bessel coefficient
$\sigma^2_y(\tau)$		Allan variance
τ	sec	Period of sample

M		Number of time domain averages
y		$\Delta f/f_o$ over interval τ long
N_p	Watts	Noise power
k	Joules/Kelvin	Boltzman's constant 1.38×10^{-23}
F		Noise Figure
G		Gain
B	Hz	Bandwidth
B_n	Hz	Noise bandwidth
ω	rad/sec	Angular velocity
LBW	Hz	Loop Bandwidth
LHR	Hz	Loop Holding Range

References

Hewlett-Packard Application Notes

PN11729B-1 Phase Noise Characterization of Microwave Oscillators – Phase Detector Method

AN 246-2 Measuring Phase Noise with the HP 3585A Spectrum Analyzer

AN 207 Understanding and Measuring Phase Noise in the Frequency Domain

AN 150-4 Spectrum Analysis ..... Noise Measurements

AN 283-1 Applications and Measurements of Low Phase Noise Signals Using the HP 8662A Synthesized Signal Generator

AN 270-2 Automatic Phase Noise Measurements with the HP 8568A Spectrum Analyzer

Other References

Scherer, D., "Generation of Low Phase Noise Microwave Signals", HP RF and Microwave Measurement Symposium, May, 1981

Scherer, D., "The 'Art' of Phase Noise Measurements", HP RF and Microwave Measurement Symposium, May, 1983

Scherer, D., "Design Principles and Test Methods for Low Phase Noise RF and Microwave Sources", HP RF and Microwave Measurement Symposium, October, 1978

Temple, R., "Choosing a Phase Noise Measurement Technique – Concepts and Implementations", HP RF and Microwave Measurement Symposium, February, 1983

Fischer, M., "An Overview of Modern Techniques for Measuring Spectral Purity", Microwaves, Vol. 18, No. 7, pp. 66-75, July 1979

Fischer, M. "Frequency Stability Measurement Procedures", Proceedings, Eighth Annual PTTI Meeting, Goddard Space Flight Center, Code 250, Greenbelt, Maryland, 1976, pp. 575-618

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

Shoaf, J.H., Halford, D., Risley, A.S., "Frequency Stability Specification and Measurement: High Frequency and Microwave Signals, N.B.S. Technical Note 632, National Bureau of Standards, Boulder, CO, 1973

Ashley, R.J. et al, "Measurement of Noise in Microwave Transmitters", IEEE Transactions on Microwave Theory and Techniques, Vol. MTT-25, No. 4, pp. 294-318, April, 1977

Lance, Seal, Hudson, Mendoza, and Halford, "Phase Noise Measurements Using Cross-Spectrum Analysis", Conference on Precision Electromagnetic Measurements, Ottowa, Canada, June, 1978

Lance, Seal, Mendoza, Hudson, "Phase Noise Characteristics of Frequency Sources", Ninth Annual PTTI Meeting, Goddard Space Flight Center, Greenbelt, Maryland, November, 1977

Lance, Seal, Labaar, "Phase Noise Measurement Systems", TRW

Lance, Seal, "Phase Noise and AM Noise Measurements in the Frequency Domain at Millimeter Wave Frequencies", TRW

Labaar, F., "A New Type of Delay Line FM Discriminator", TRW

Ondria, F.G., "A Microwave System for Measurements of AM and FM Noise Spectra", IEEE Transactions on Microwave Theory and Techniques, Vol. MTT-16, pp. 767-781, September, 1968

Robbins, W.P., "Phase Noise in Signal Sources", IEEE Telecommunications Series No. 9, 1982