

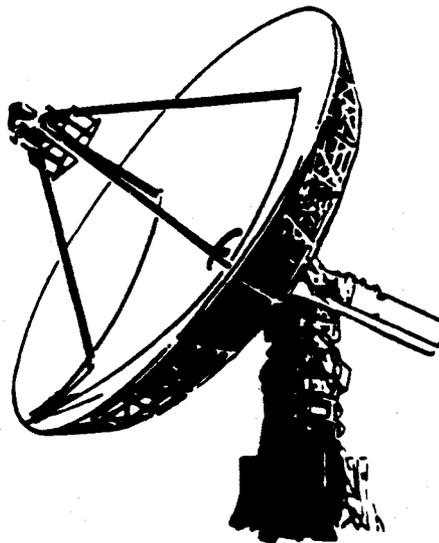
The "Art" of Phase Noise Measurement

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Measurement
Symposium
and
Exhibition**



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INTRODUCTION

Measuring and specifying phase noise has become increasingly important as phase noise is the limiting factor in many RF and microwave systems, like Doppler radar and space telemetry systems or communication links. The complexity and subtleties of the measurement have earned it the reputation of being more art than science.

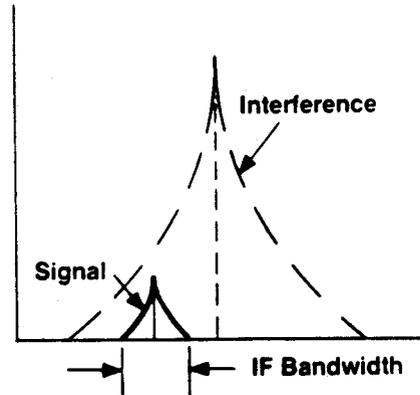
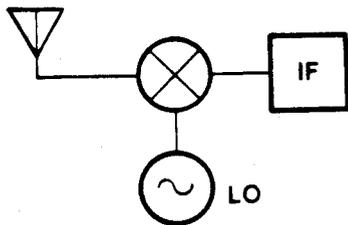
This paper gives an introduction to phase noise measurement focusing on the two most common and useful techniques, the "Two Source Phase Detector" and the "Delay Line Frequency Discriminator" methods. System limitations are pointed out as well as the potential sources of erroneous data when phase noise is measured in a bench set up.

The 11729C Carrier Noise Test Set is HP's latest contribution in this field, designed to simplify and automate the complex task. A table of comparisons will show where the 11729C and the various other HP solutions are optimally employed.

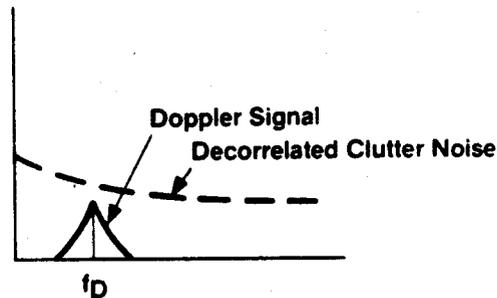
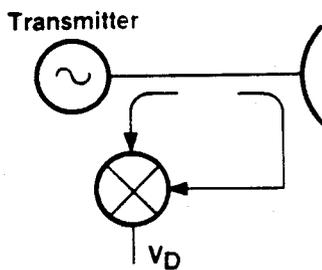


1. Why Measure Phase Noise?

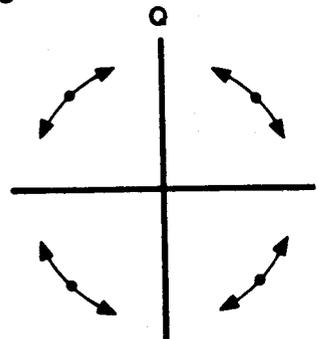
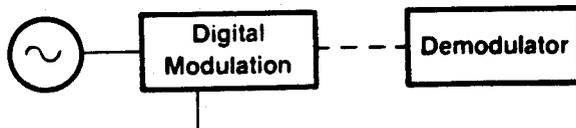
Phase noise causes . . .



- Lower receiver sensitivity in multi-signal environment



- Clutter noise in Doppler Radar Systems



- Phase errors in digital communication systems

Phase noise imposes fundamental limitations wherever a weak signal is processed in the presence of a strong interfering signal. In the above example, phase noise sidebands of the receiver local oscillator are transferred to the IF product of the strong interfering signal and cover up the weak wanted signal.

A similar situation exists with coherent Doppler radar. The strong interfering signal in this case is produced by reflections from large stationary objects. Phase noise side bands of this unwanted return signal are decorrelated by delay and potentially cover as clutter noise the weak Doppler signal.

Even in the world of digital data transmission phase noise is a limiting factor. Phase noise adds to overall system noise increasing Bit Error Rate and may cause a cycle slip in the carrier or data clock recovery.

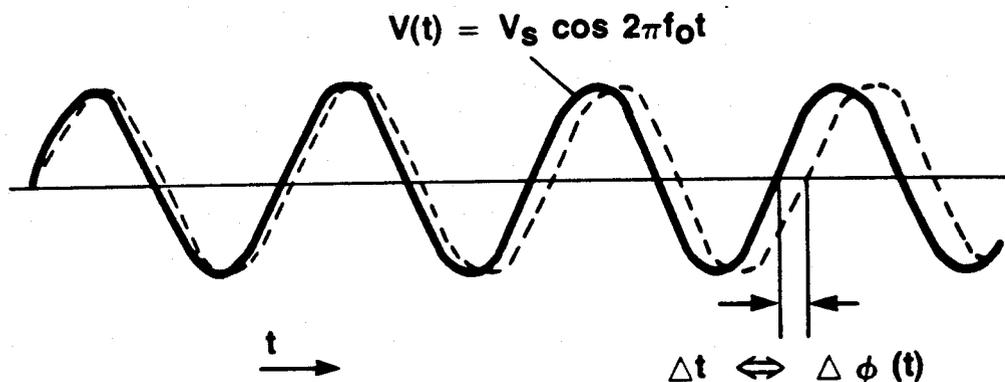
2. Basic Representation of Phase Noise

2.1 In the Time Domain:

$v(t)$ = Signal with random phase fluctuation $\Delta \phi (t)$

$$v(t) = V_s \cos [2\pi f_0 t + \Delta \phi (t)]$$

Oscilloscope Display



Frequency and phase are related by

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

In the time domain phase noise can be observed as phase jitter of the signal on an oscilloscope display or a time interval counter may detect the instantaneous time fluctuations of the zero crossings.

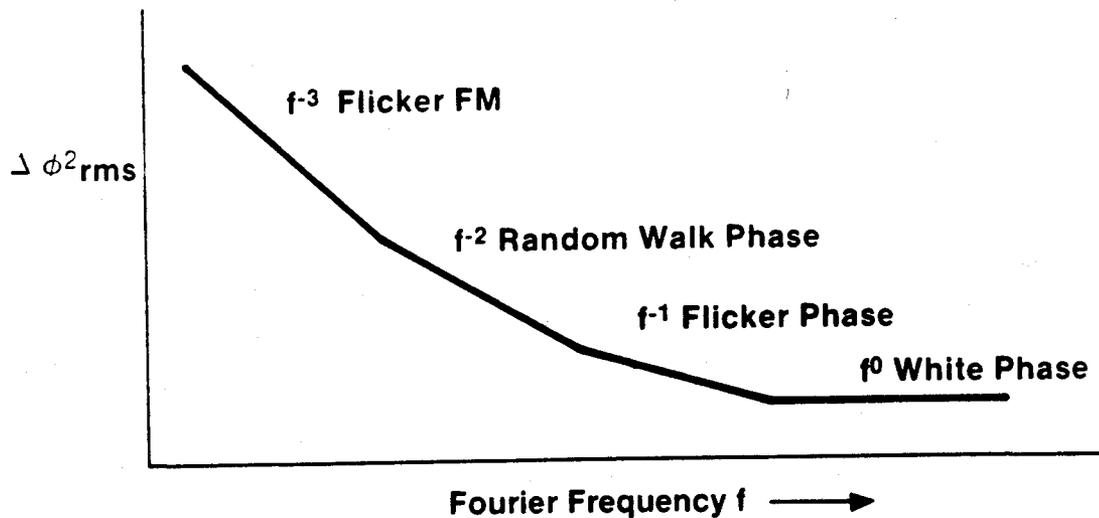
Phase or frequency fluctuations are the same physical phenomena. One can be derived from the other as angular frequency is the first derivative of phase with respect to time.

2.2 In the Frequency Domain:

$$F \Delta \phi(t) = \Delta \phi(f)$$

Spectral Density of Phase Fluctuations S_{ϕ}

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



Spectral Density of Frequency Fluctuations $S_{\Delta f}$

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

Spectral density distributions describe frequency or phase noise in the frequency domain. Terms like "White," "Flicker" and "Random Walk" refer to the slope of f^0 , f^{-1} , and f^{-2} of the density distribution. Multiplication by the Fourier frequency — corresponding to differentiation in the time domain — converts the spectral density of phase noise to the spectral density of frequency noise.

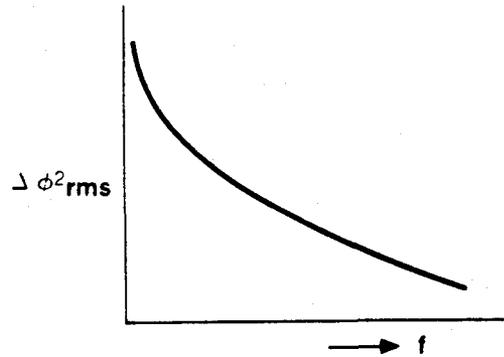
2.3 Relation of Phase Noise to Carrier Sideband Noise

The modulation of the signal phase manifests itself in modulation sidebands to the carrier at offset frequencies which are multiples of the modulation rate (Fourier frequency). The sideband levels are related to the magnitude of the phase deviation by Bessel terms. For small-angle modulation, only the first Bessel term is significant and the relation between phase deviation and sideband level is approximated by

For small angle modulation

$$(\Delta\phi \ll 1 \text{ radian})$$

$$\frac{P_{ssb}}{P_s} = \frac{1}{2} \Delta\phi_{rms}^2(f)$$

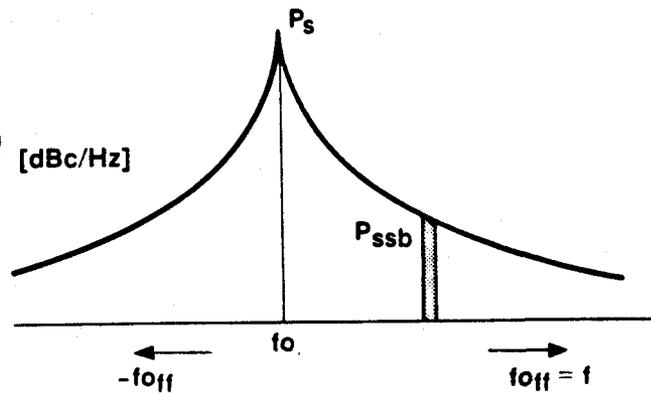


A common, though indirect, representation of phase noise is $\mathcal{L}(f)$, the ratio of the single sideband noise power in a 1 Hz bandwidth to the total carrier power specified at a given offset f of the carrier.

This representation is applicable only for very small phase deviations, sufficiently small to produce negligible higher order sidebands.

Definition of $\mathcal{L}(f)$

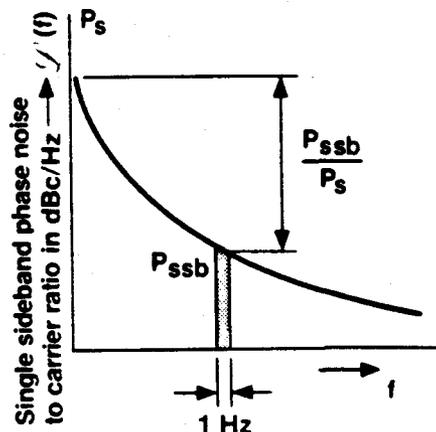
$$\mathcal{L}(f) = \frac{P_{ssb}(\text{per 1 Hz})}{P_s} \text{ [dBc/Hz]}$$



Relation between \mathcal{L} , S_ϕ , $S_{\Delta f}$:

for $\Delta\phi \ll 1$ radian

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



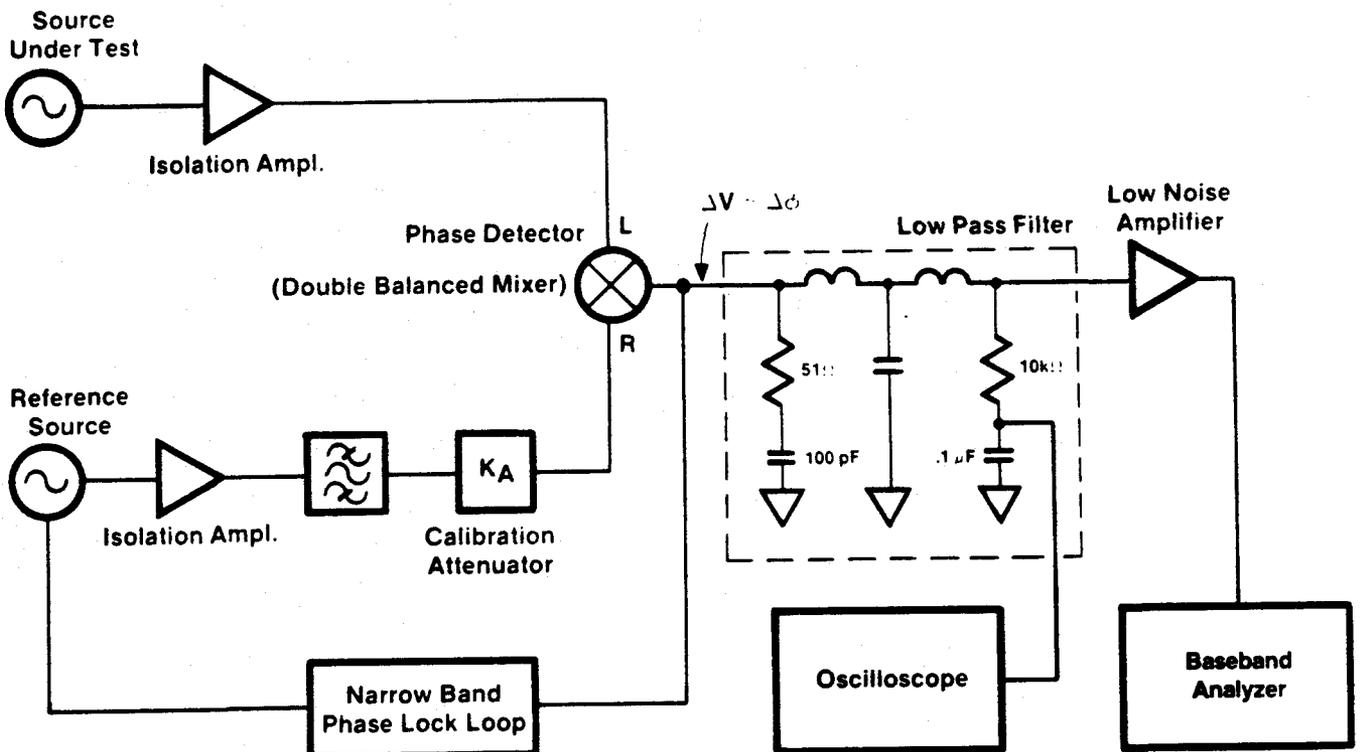
3. Phase Noise Measurement Techniques

- Direct Spectrum Analysis
- ▶ • Two Sources and Phase Detector
- Frequency Discriminator
- ▶ • Delay Line Frequency Discriminator
- Carrier Suppression (Ondria Bridge)
- Heterodyne Period Counter

From the multitude of techniques to measure phase or frequency noise two basic methods are singled out and explained in some detail on the following pages: the "Two Sources and Phase Detector" technique and the "Delay Line Frequency Discriminator" technique.

Both methods are relatively simple to implement and are broadband solutions. They also compliment each other well. The first method measures phase noise; high sensitivity is obtainable and two sources are needed. The second method measures frequency noise, has inherently lower sensitivity close in but no second source is required.

3.1 Two Sources and Phase Detector Technique



$$S_{\phi}(f) = \Delta \phi_{rms}^2(f) = 1/2 \frac{\Delta V^2_{rms}(1 \text{ Hz})}{V^2_{B \text{ rms}}}$$

for $\Delta \phi \ll 1 \text{ rad.}$

$$\mathcal{L}(f) = 1/2 S_{\phi}(f) = 1/4 \frac{\Delta V^2_{rms}(1 \text{ Hz})}{V^2_{B \text{ rms}}}$$

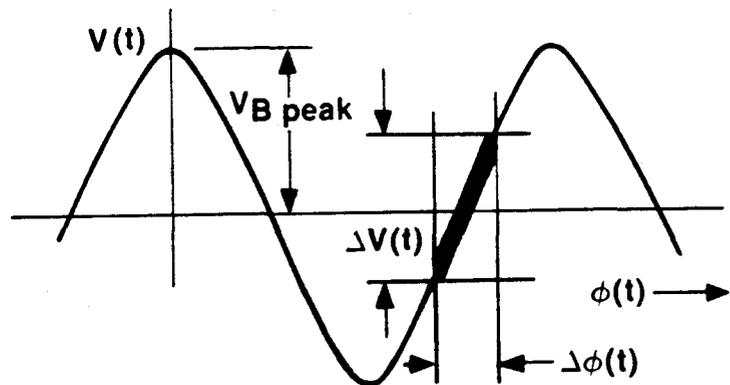
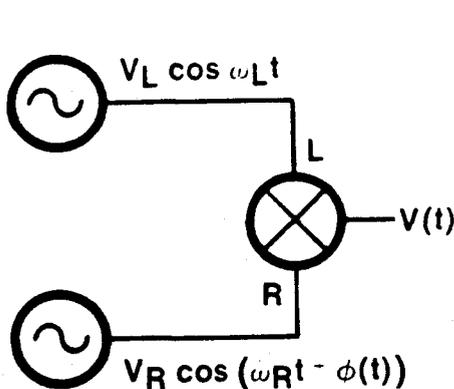
$S_{\phi}(f)$
 $\mathcal{L}(f)$
 (for $\Delta \phi \ll 1 \text{ rad.}$)

With the LO and RF input in phase quadrature steady state wise, the mixer IF output generates a signal proportional to the phase difference of the two sources. In linear operation the phase detector constant equals the peak voltage of the sinusoidal beat signal produced when either source is frequency offset.

Most sources would drift out of quadrature over the period of measurement. A narrow band phase lock loop automatically forces the two signals into phase quadrature. At rates less than the loop bandwidth the sources now track each other; at rates higher than the loop bandwidth the phase fluctuations are unaffected. The two isolation amplifiers should prevent injection locking of the sources.

Phase noise measured on the IF port represents the rms-sum of the noise contributions of each source. For definite data on the source under test, the phase noise of the reference source should be either negligible or well characterized. In the absence of any information on both sources, one can at least state that neither source noise is worse than the measured data. If 3 unknown sources are available, 3 measurements with 3 different source combinations yield sufficient data to calculate accurately each individual noise level.

3.1.1 The Mixer as Phase Detector



$$V(t) = \underbrace{K_L V_R}_{V_B \text{ peak}} \cos((\omega_R - \omega_L)t + \phi(t)) + \cancel{K_L V_R \cos((\omega_R + \omega_L)t + \phi(t))} + \dots$$

$$V(t) = V_B \text{ peak} \cos((\omega_R - \omega_L)t + \phi(t))$$

K_L = Conversion Loss
 $V_B \text{ peak}$ = Peak Voltage of Beat Signal

for $\omega_L = \omega_R$ and $\phi(t) = (k+1) 90^\circ + \Delta\phi(t)$

$$\Delta V(t) = (\pm) V_B \text{ peak} \sin \Delta\phi(t)$$

for $\Delta\phi \ll 1$ radian

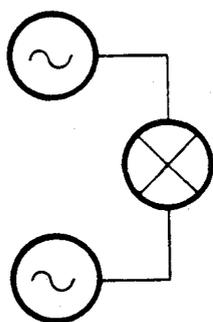
$$\Delta V(t) = (\pm) K_\phi \Delta\phi(t) \quad K_\phi = V_B \text{ peak} \left[\frac{\text{Volts}}{\text{rad}} \right]$$

The frequency conversion of signals applied to a mixer is physically based on one signal gating the other or on the nonlinear conductance of the mixer diodes. Either case represents a multiplying process. Among many terms, the mixer also generates a product resulting in the difference frequency of the input signals.

With a sinusoidal input at the RF port and assuming linear operation (linear relation between RF and IF voltage), the beat signal is also a sine wave with a peak voltage $V_B \text{ peak}$. The phase slope of this beat signal at zero crossings equals $\pm V_B \text{ peak}$.

Setting both input signals to the same frequency and at a phase offset of 90° phase fluctuations $\Delta\phi(t)$ result in voltage fluctuations $\Delta V(t)$ which for small phase deviations are linearly related. The mixer acts as a phase detector with a phase detector constant of K_ϕ equal to $V_B \text{ peak}$.

3.1.2 Calibration with Beat Signal



$$\Delta V(f) = K_{\phi} \Delta \phi(f)$$

$$K_{\phi} = V_B \text{ peak} \\ = \sqrt{2} V_B \text{ rms}$$

$$\Delta \phi_{\text{rms}}(f) = \frac{1}{K_{\phi}} \Delta V_{\text{rms}}(f) = \frac{1}{\sqrt{2} V_B \text{ rms}} \Delta V_{\text{rms}}(f)$$

$$S_{\phi}(f) = \Delta \phi^2_{\text{rms}}(f) = 1/2 \frac{\Delta V^2_{\text{rms}}(1 \text{ Hz})}{V^2_{B \text{ rms}}}$$

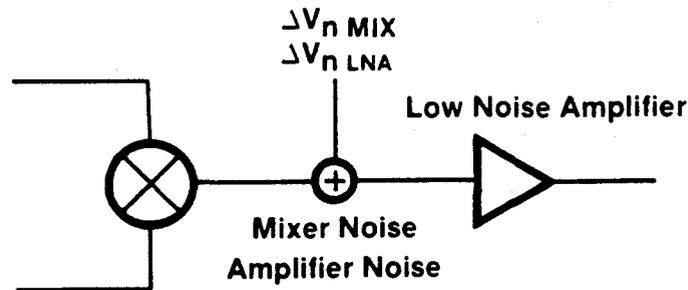
for $\Delta \phi \ll 1$ $\mathcal{L}(f) = 1/2 S_{\phi}(f) = 1/4 \frac{\Delta V^2_{\text{rms}}(1 \text{ Hz})}{V^2_{B \text{ rms}}}$

- | | | | |
|------------------|---|---|---|
| $\mathcal{L}(f)$ | = | $\Delta V^2_{\text{rms}} \text{ dBm}$ | Phase noise spectrum in dBm corrected for bandwidth and analyzer characteristics. |
| | | $-V^2_{B \text{ rms}} \text{ dBm}$ | Beat signal in dBm. |
| | | $-K_A \text{ dBm}$ | Attenuation (if beat signal was attenuated in calibration). |
| | | -6 dB | Accounts for rms value of beat signal (3 dB) and conversion of $S_{\phi}(f)$ to $\mathcal{L}(f)$ (3 dB). |
| | | $+2.5 \text{ dB}$ | Accounts for log amplifier and video filtering (averaging) of noise (if analog spectrum analyzer is used). |

Assuming $V_B \text{ peak}$ as the phase detector constant dictates linear operation of the mixer. The calibration of the spectrum analyzer for $\mathcal{L}(f)$ or $S_{\Delta \phi}(f)$ can then be read from the above equations.

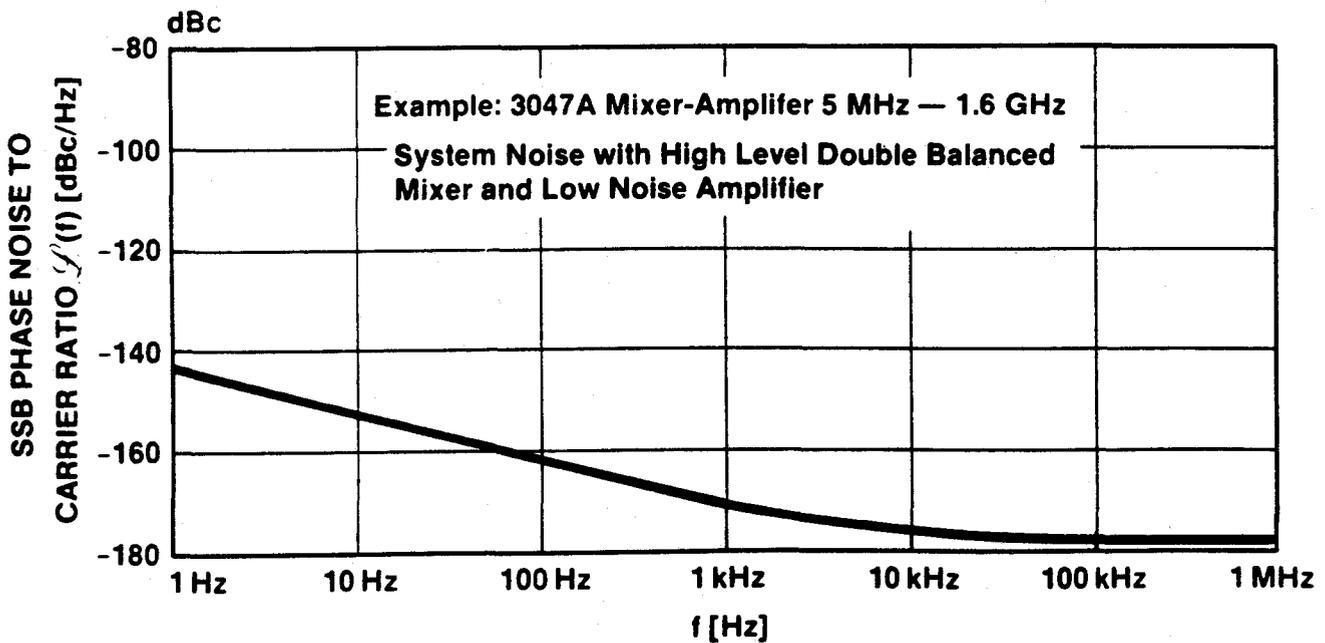
Other calibration methods are based on the characterization of the phase slope, on calibrated PM or FM sidebands on one of the two sources or on the injection of a calibrated spurious signal.

3.1.3 System Noise Floor



$$\gamma_{\text{system}}(f) = \frac{1}{4} \frac{\Delta V_{n \text{ MIX}}^2 + \Delta V_{n \text{ LNA}}^2}{V_{B \text{ rms}}^2}$$

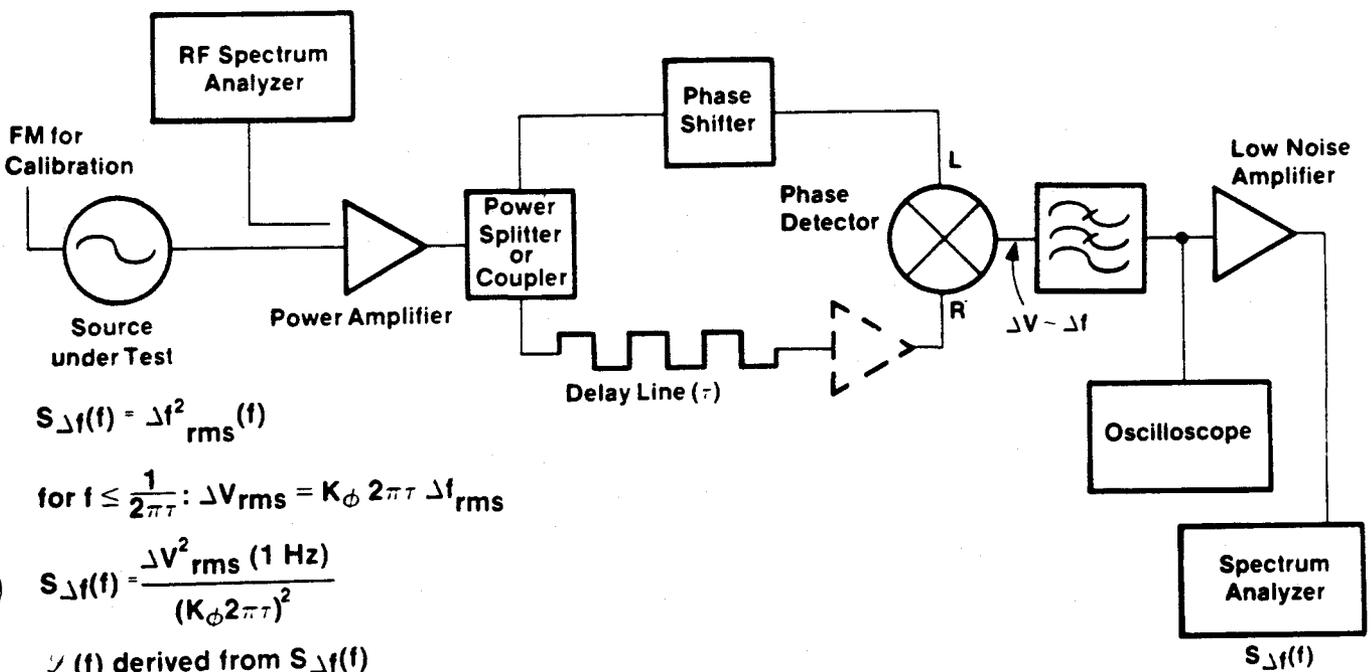
The noise floor of the measurement system is established by the equivalent amplifier and mixer noise (shot noise, flicker noise) at the IF port reduced by the phase detector gain K_{ϕ} ($V_{B \text{ peak}}$ for linear operation). System sensitivity is therefore maximized by using a high level mixer, selecting a mixer with low flicker noise and following the mixer with a low noise amplifier.



3.1.4 Potential Error Sources

- **Suppression or peaking of noise near or below the phase lock loop bandwidth.**
- **Injection locking**
- **Deviation from phase quadrature (resulting in an effectively lower K_{ϕ})**
- **Non-linear operation of mixer (if linearity was assumed in calibration)**
- **Distortion of RF-Signal (resulting in a deviation of K_{ϕ} from VB peak)**
- **Saturation of the preamplifier or spectrum analyzer in calibration or by high spurious signals (e.g. line spurious)**
- **Change of impedance interface while going from calibration to measurement**
- **Insufficient cancelation of common source noise in a residual noise test**
- **AM-noise (in particular AM noise of the common source in a residual noise test)**
- **Closely spaced spurious (e.g. line multiples) misinterpreted as phase noise**
- **Noise injected by peripheral circuitry (power supply, phase lock loop)**
- **Noise injected by peripheral instrumentation (e.g., oscilloscope, DVM)**
- **Microphonic noise.**

3.2 Delay Line Frequency Discriminator Technique



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

$$\text{for } f \leq \frac{1}{2\pi\tau} : \Delta V_{rms} = K_{\phi} 2\pi\tau \Delta f_{rms}$$

$$S_{\Delta f}(f) = \frac{\Delta V^2_{rms}(1 \text{ Hz})}{(K_{\phi} 2\pi\tau)^2}$$

$\mathcal{V}(f)$ derived from $S_{\Delta f}(f)$

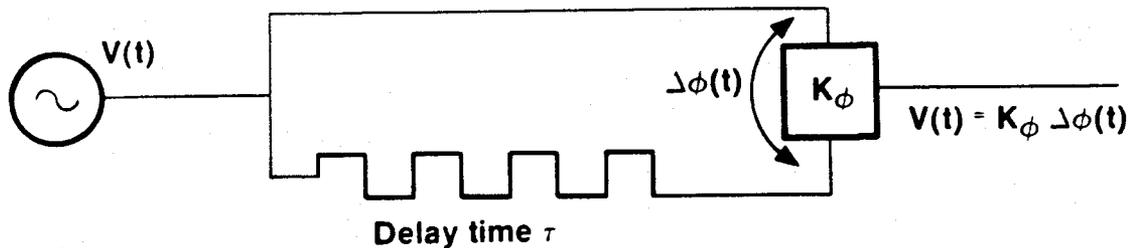
$$\text{for } \Delta\phi \ll 1 : \mathcal{V}(f) = \frac{1}{2} \frac{1}{f^2} S_{\Delta f}(f) = \frac{1}{2} \frac{1}{f^2} \frac{\Delta V^2_{rms}(1 \text{ Hz})}{(K_{\phi} 2\pi\tau)^2}$$

A delay line and a mixer operating as a phase detector have the combined effect of a frequency discriminator. The delay line transforms any frequency fluctuation into a phase fluctuation and the mixer with the L and R inputs at 90° offset linearly converts the phase fluctuations into voltage fluctuations at the IF port. As derived on the next page, the equivalent frequency discriminator constant K_D is proportional to τ and K_{ϕ} .

The frequency to phase conversion has a $\sin x/x$ characteristic with a null at $f = 1/\tau$. Measurements are therefore limited to $f < 1/2\tau$ (with correction).

The system can be calibrated by characterizing the frequency slope or K_{ϕ} and τ , or more conveniently by adding FM sidebands to the source under test (or a substitute) and determining their level as reference level with an RF spectrum analyzer.

3.2.1 Theory of Delay Line Frequency Discriminator



$$f(t) = f_0 + \Delta f \sin 2\pi ft$$

$$V(t) = V_0 \cos(2\pi f_0 t + \frac{\Delta f}{f} \cos 2\pi ft)$$

Carrier with sinusoidal FM

f_0 = Carrier frequency

f = FM rate

Δf = FM deviation

Phase Difference on Phase Detector $\Delta\phi(t)$

$$\Delta\phi(t) = 2\pi f_0(t - \tau) + \frac{\Delta f}{f} \cos 2\pi f(t - \tau) - 2\pi f_0 t - \frac{\Delta f}{f} \cos 2\pi ft$$

$$\Delta\phi(t) = -2\pi f_0 \tau + \frac{\Delta f}{f} (\cos 2\pi f(t - \tau) - \cos 2\pi ft)$$

$$\Delta\phi(t) = -2\pi f_0 \tau + 2 \frac{\Delta f}{f} \sin \pi f \tau \sin \left[2\pi f \left(t - \frac{\tau}{2} \right) \right]$$

Phase Detector Output: $V(t) = K_\phi \Delta\phi(t)$

for $2\pi f_0 \tau = (2k+1)\frac{\pi}{2}$ (Phase Quadrature):

$$V(t) = K_\phi 2 \frac{\Delta f}{f} \sin \pi f \tau \sin 2\pi f \left(t - \frac{\tau}{2} \right)$$

Transfer Function:

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

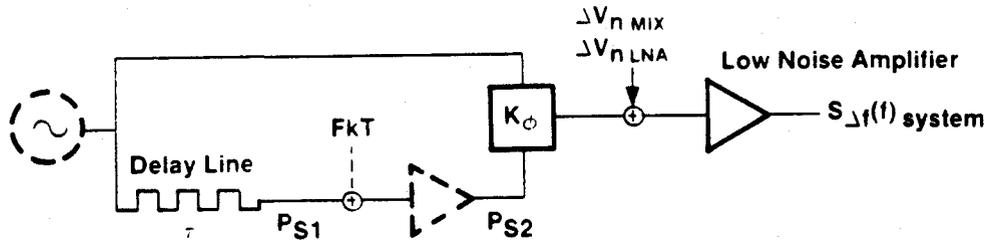
$$\text{for } f \leq \frac{1}{2\pi\tau}: \quad \frac{\sin \pi f \tau}{\pi f \tau} \approx 1$$

$$\Delta V = K_\phi 2\pi \tau \Delta f$$

$$\underbrace{K_\phi 2\pi \tau}_{K_D \left[\frac{V}{\text{Hz}} \right]} = \text{Frequency Discriminator Constant} = K_\phi 2\pi \tau$$

3.2.2

System Noise Floor



Phase Noise of RF Amplifier (if used): $\Delta\theta^2_{RF} = \frac{FkT}{P_{S1}} \left(1 + \frac{f}{f_c}\right)$

K_ϕ (in linear operation): $K_\phi = V_{B \text{ peak}} = \sqrt{2 P_{S2} L \times 50\Omega}$

with L = Conversion Loss (Power) of Mixer

Mixer noise and baseband amplifier noise are interpreted by the system as frequency noise. This also holds for phase noise introduced by an RF amplifier. The amplifier might have been added to boost the signal level lowered by the lossy delay line.

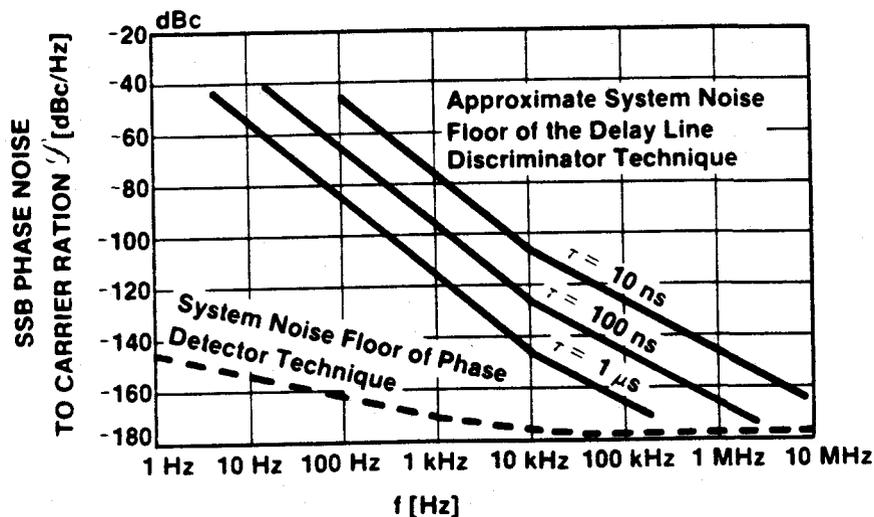
$$S_{\Delta f(f)} = \frac{\Delta V^2_{n \text{ MIX}}}{(K_\phi 2\pi\tau)^2} + \frac{\Delta V^2_{n \text{ LNA}}}{(K_\phi 2\pi\tau)^2} + \frac{\Delta\theta^2_{RF}}{(2\pi\tau)^2}$$

$$\mathcal{L}(f) = \frac{1}{2} \frac{1}{f^2} S_{\Delta f(f)} \quad \text{for } \Delta\phi \ll 1$$

$$\mathcal{L}(f) = \frac{1}{2} \left(\frac{1}{2\pi f\tau}\right)^2 \left(\frac{\Delta V^2_{n \text{ MIX}}}{K^2_\phi} + \frac{\Delta V^2_{n \text{ LNA}}}{K^2_\phi} + \Delta\theta^2_{RF} \right)$$

Mixer, baseband amplifier and RF amplifier noise all have a flicker noise characteristic close in and show white noise farther out. The contributions of mixer and baseband amplifier are reduced by K_ϕ which points out the benefit of the RF amplifier as it maximizes K_ϕ . A better solution, though, is to employ a power amplifier before the signal split and eliminate the need for an RF amplifier and with it $\Delta\theta_{RF}$.

Since $S_{\Delta f}$ has a flicker and then white noise characteristic, $\mathcal{L}(f)$ increases with f^{-2} and f^{-3} towards the carrier. This means good sensitivity far out (limited by the $\sin x/x$ response) but poor sensitivity as compared with the phase detector method close in.



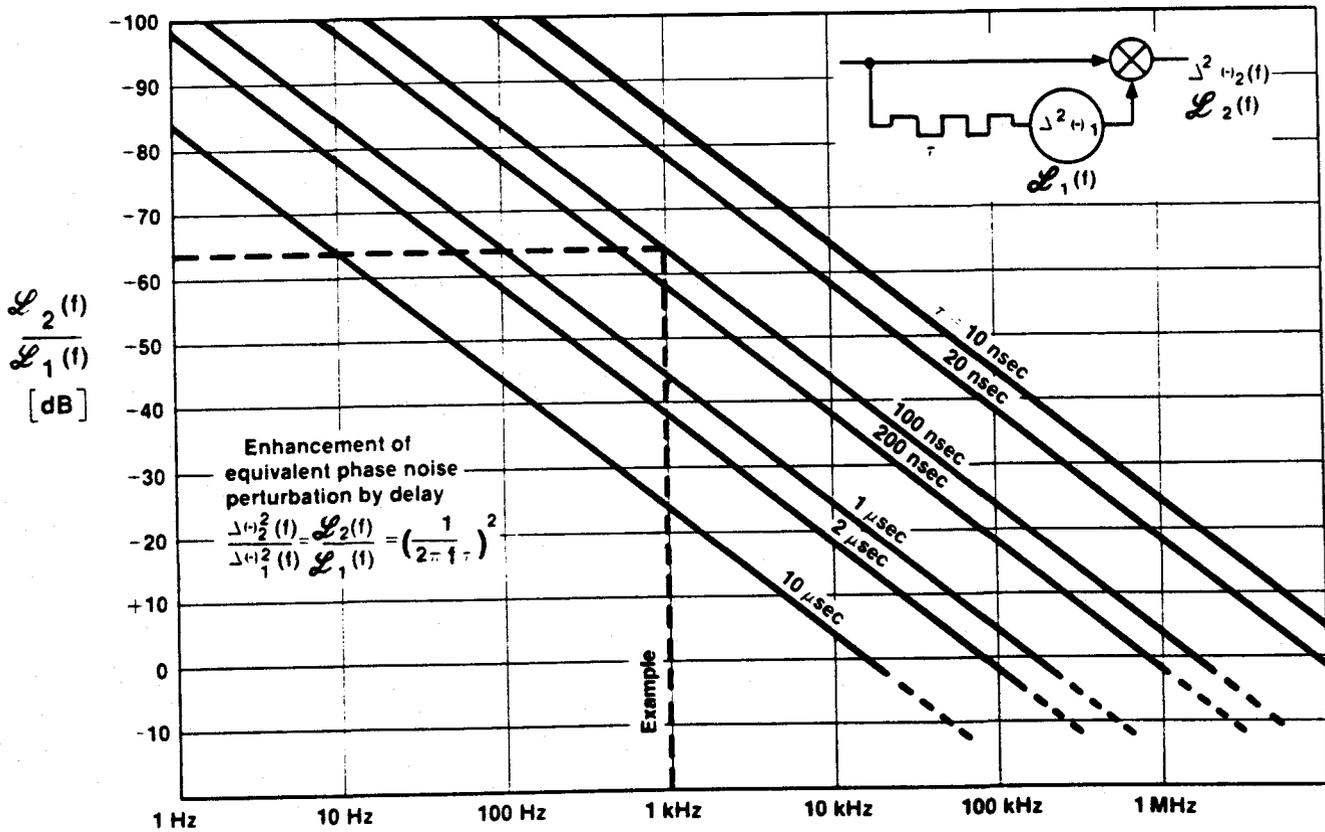
A different way of estimating the system noise floor of the delay line discriminator technique is to lump all noise contributions into an equivalent phase noise perturbation $\Delta\theta_1$ (or \mathcal{L}_1). The delay line discriminator interpreting $\Delta\theta$ as frequency noise enhances this equivalent phase noise by a factor of $1/2\pi f\tau$.

Example:

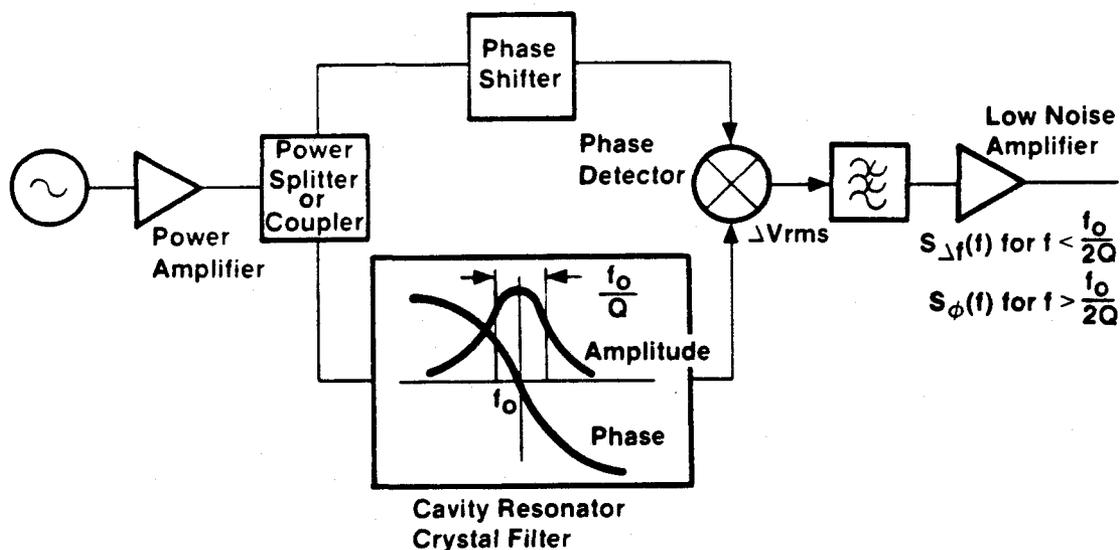
Phase noise of the mixer at 1 kHz $\mathcal{L}_1(1 \text{ kHz}) = -160 \text{ dBc/Hz}$.

With a 100 nsec delay time the resulting system noise floor is

$$\begin{aligned} \mathcal{L}_2 &= \mathcal{L}_1 \left(\frac{1}{2\pi f\tau} \right)^2 \\ &= -160 \text{ dBc} + 64 \text{ dB} = -96 \text{ dBc/Hz} \end{aligned}$$



3.2.3 Frequency Discriminator with Resonator



Inside Resonator Bandwidth ($f < \frac{f_0}{2Q}$):

$$\text{Equivalent Delay Time } \tau = \left. \frac{d\phi}{d\omega} \right|_{f_0} = \frac{Q}{\pi f_0}$$

$$K_D = K_\phi 2\pi\tau = K_\phi \frac{2Q}{f_0}$$

$$\Delta V_{rms}(f) = K_\phi \frac{2Q}{f_0} \Delta f_{rms}(f)$$

Although not a broadband solution, it can be advantageous to replace the delay line with a cavity resonator or crystal filter. Group delay ($d\phi/d\omega$) produces an equivalent τ of $Q/\pi f_0$ for Fourier frequencies less than half the bandwidth. Inside $f_0/2Q$, the system therefore measures the spectral density of frequency noise with a discriminator constant of $K_\phi 2Q/f_0$.

Outside Resonator Bandwidth ($f > \frac{f_0}{2Q}$):

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

Outside $f_0/2Q$, the system may be viewed as 2 signals applied to a phase detector with the sideband noise of one signal stripped off by the cavity resonator. The system then measures $\Delta \phi_{rms}^2$ with a phase detector constant K_ϕ .

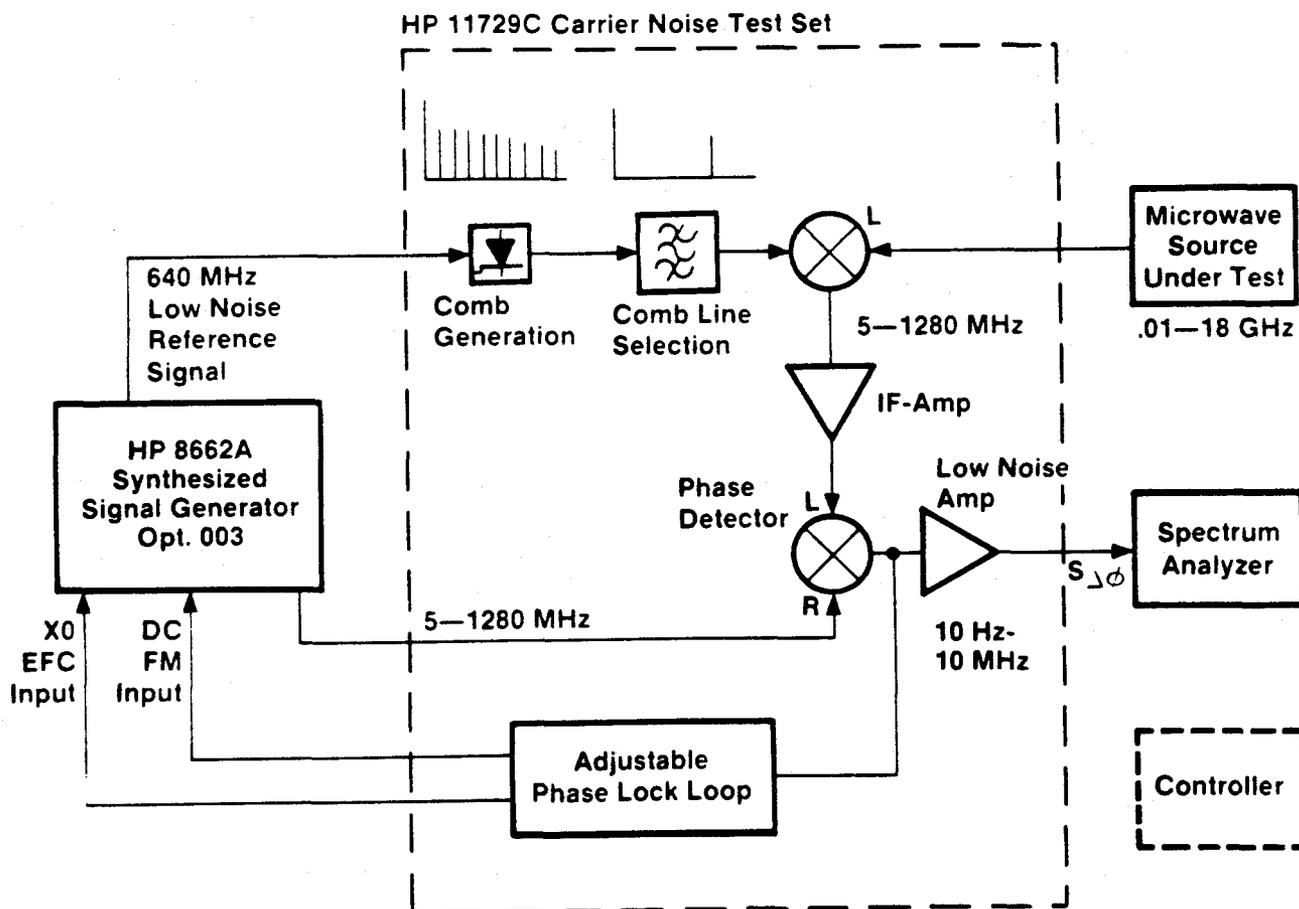
4. **HP Phase Noise Measurement Solutions**

- **HP 11729C/8662A Phase Noise Test System
(phase detector mode)**
- **HP 11729C Carrier Noise Test Set
(delay line discriminator mode)**
- **HP 3047A Phase Noise Measurement System**
- **HP 11740A Microwave Phase Noise
Measurement System**

Following is a comparison of HP phase noise measurement solutions with particular focus on the 11729C Carrier Noise Test Set, HP's most recent contribution to phase noise measurement.

4.1

HP 11729C/8662A Phase Noise Test System Principle of Operation (Phase Detector Method)

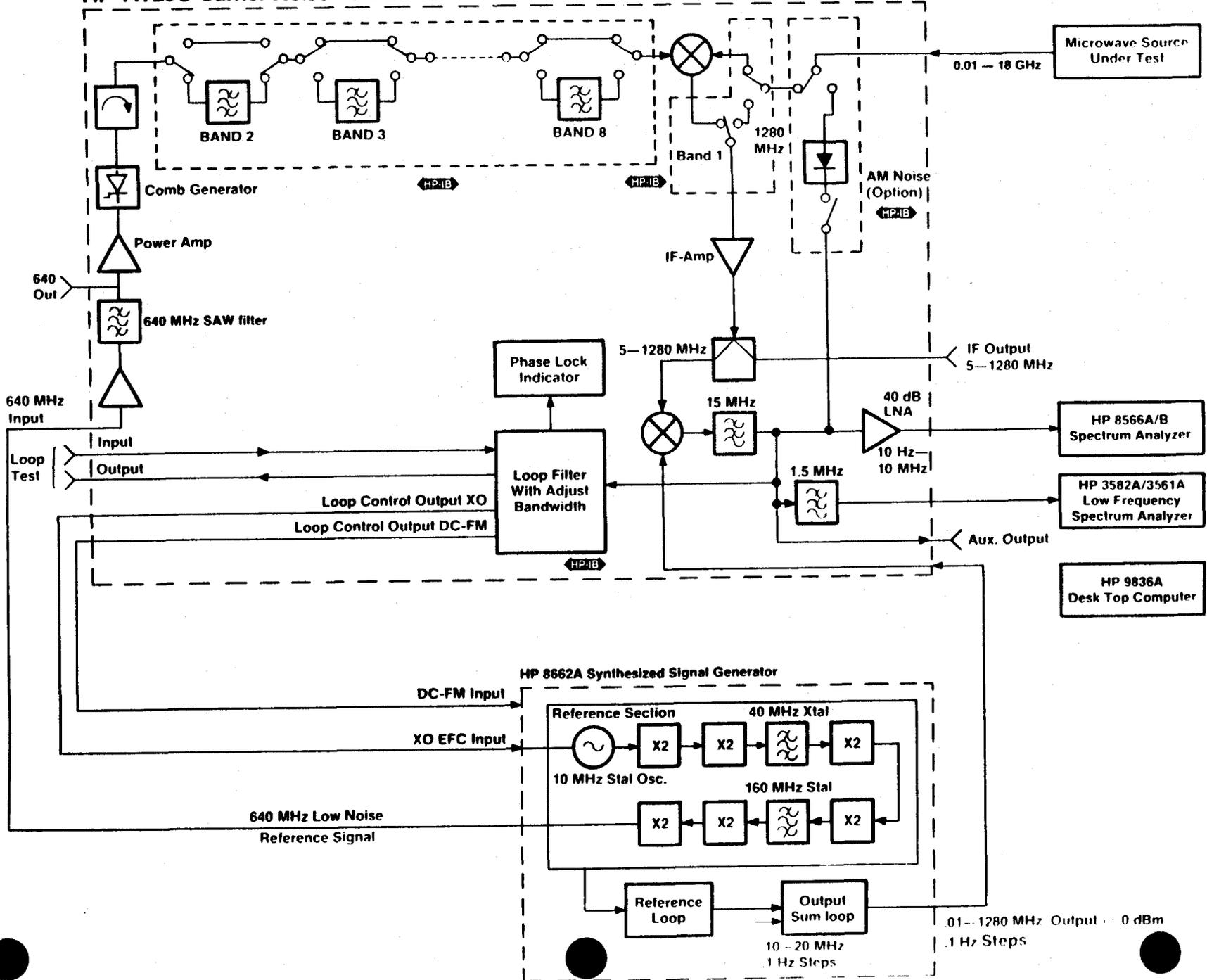


The HP 8662A Synthesized Signal Generator provides a state-of-the-art low noise reference signal at 640 MHz to the HP 11729C Carrier Noise Test Set. This signal is applied to a step recovery diode multiplier which generates a comb of signals spaced by 640 MHz ranging beyond 18 GHz. A switched in filter selects an appropriate comb line. This microwave reference signal down converts the microwave source under test into an IF range of 5 to 1280 MHz.

The double balanced mixer phase compares the IF signal with the output signal of the HP 8662A (.01 to 1280 MHz). Phase quadrature is enforced by phase locking the HP 8662A to the signal under test either thru the limited electronic control of the HP 8662A's 10 MHz crystal oscillator or thru its DC-FM port. The loop bandwidths can be set over a 4 decade range.

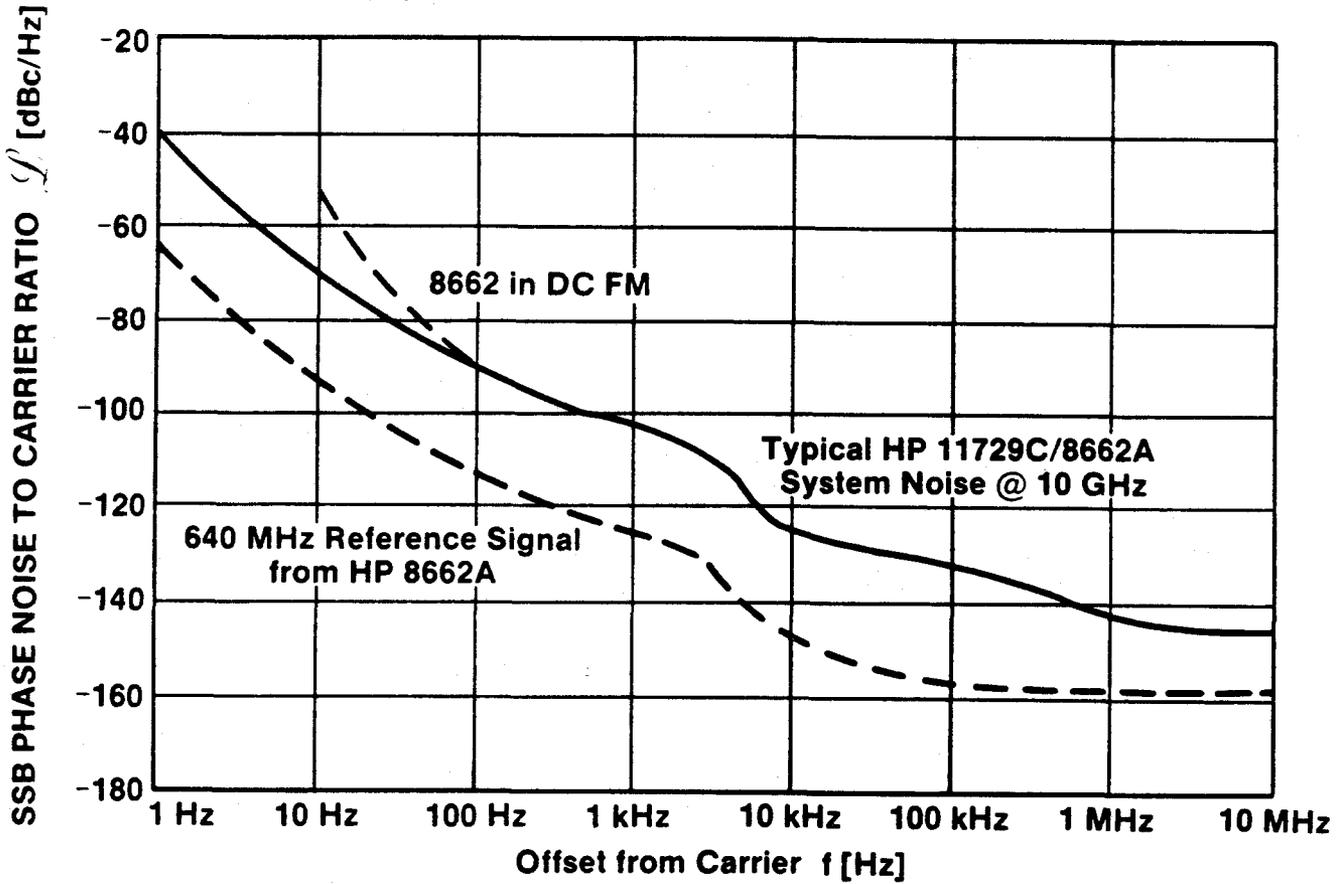
A first order phase lock loop is enabled by the capture feature making lock acquisition easy. Then the second order phase lock loop (with very high dc gain) maintains lock and phase quadrature. To enable phase noise measurements within the loop bandwidth, the phase noise suppression can be characterized via the loop test ports.

HP 11729C Carrier Noise Test Set



4.1.1

Noise Floor of HP 11729C/8662A Phase Noise Test System (Phase Detector Method)



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

N = Multiplication number of the 640 MHz signal

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

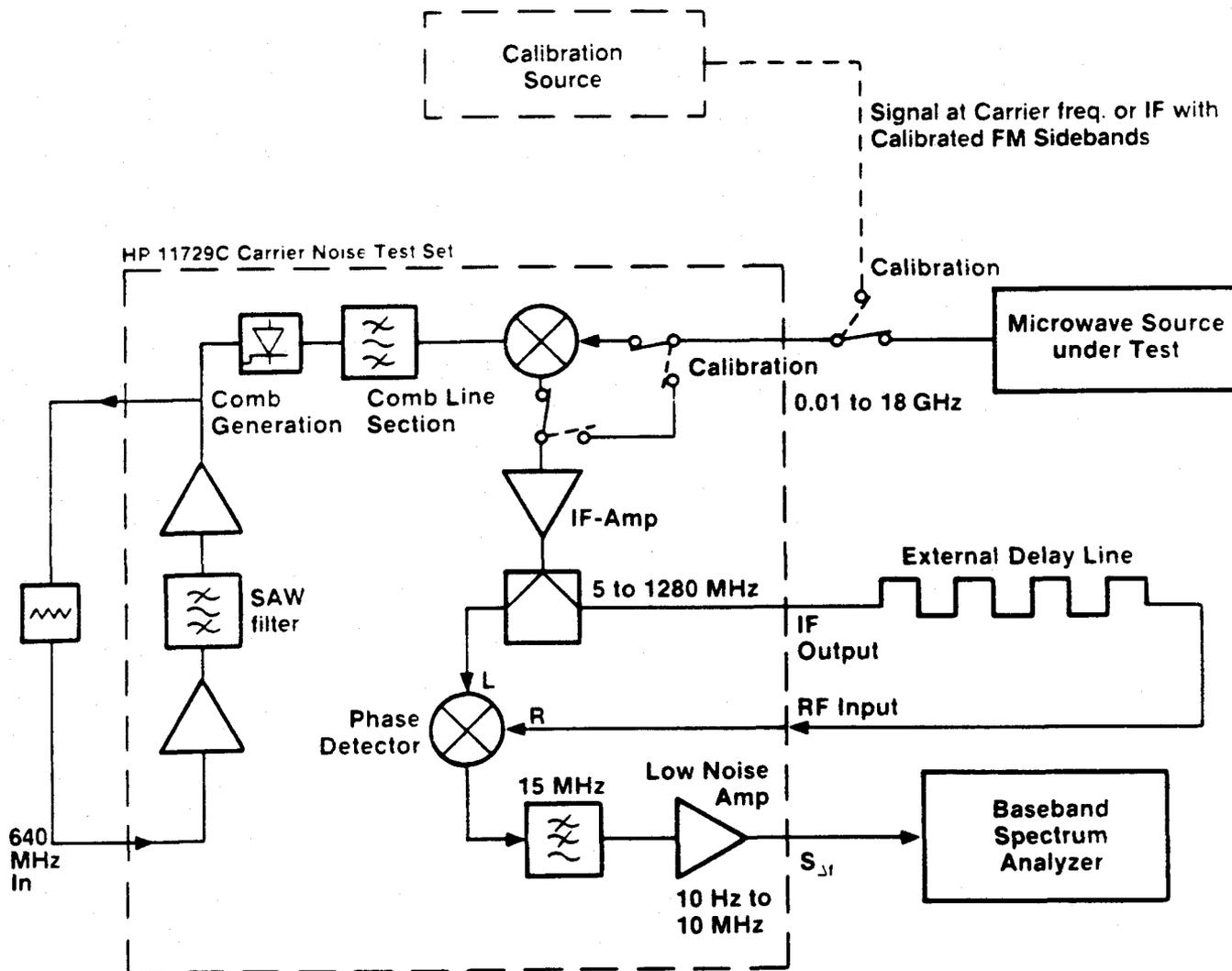
\mathcal{L}_2 = Absolute SSB Phase Noise of the 5 to 640 (1280) MHz tunable signal (dBc/Hz).

\mathcal{L}_3 = Two-port noise of HP 11729C (dBc/Hz).

Offset from carrier	Typical 640 MHz low noise output	Typical 320-640 MHz tunable 8662A output	Typical Two-port noise of HP 11729C @ 5 GHz	Typical Two-port noise of HP 11729C @ 10 GHz
1 Hz	-64 dBc	-64 dBc	-99 dBc	-93 dBc
10 Hz	-94 dBc	-94 dBc	-112 dBc	-106 dBc
100 Hz	-114 dBc	-114 dBc	-125 dBc	-119 dBc
1 kHz	-126 dBc	-125 dBc	-132 dBc	-126 dBc
10 kHz	-149 dBc	-136 dBc	-137 dBc	-131 dBc
100 kHz	-159 dBc	-136 dBc	-148 dBc	-142 dBc
1 MHz	-159 dBc	-146 dBc	-153 dBc	-147 dBc

For most microwave applications, \mathcal{L}_1 , the SSB phase noise of the 640 MHz reference signal, is the dominating factor in setting the noise floor of the system as its phase noise is multiplied up to microwave frequencies.

HP 11729C Operating in Discriminator Mode

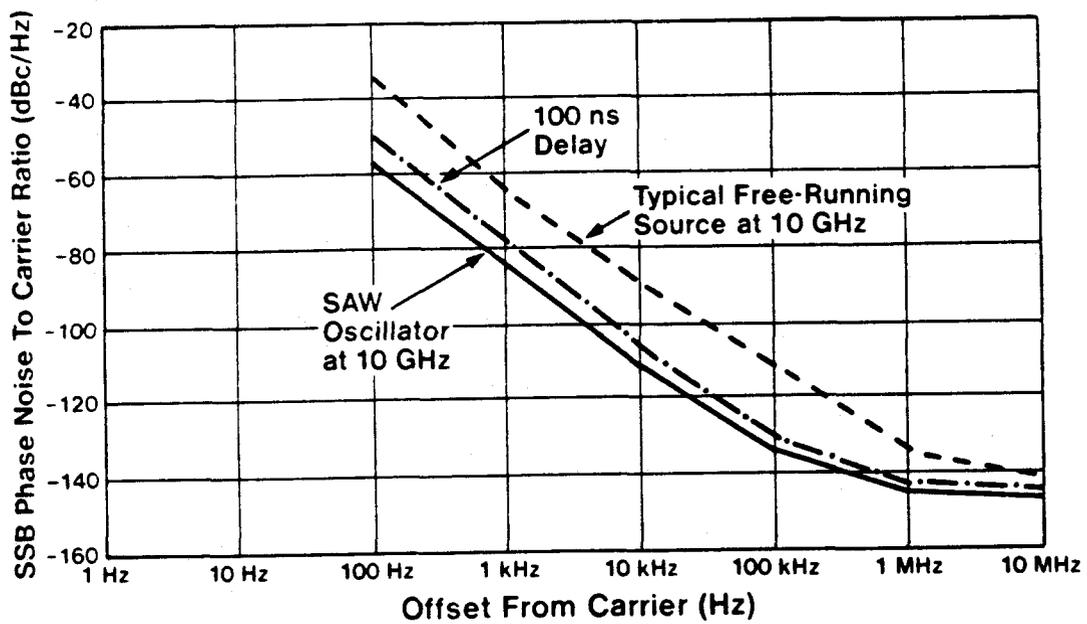


- Low noise down conversion of microwave signal under test
- Delay line operating at IF (5—1500 MHz)
- Calibration signal provided by HP 8662A
- Phase quadrature indication.

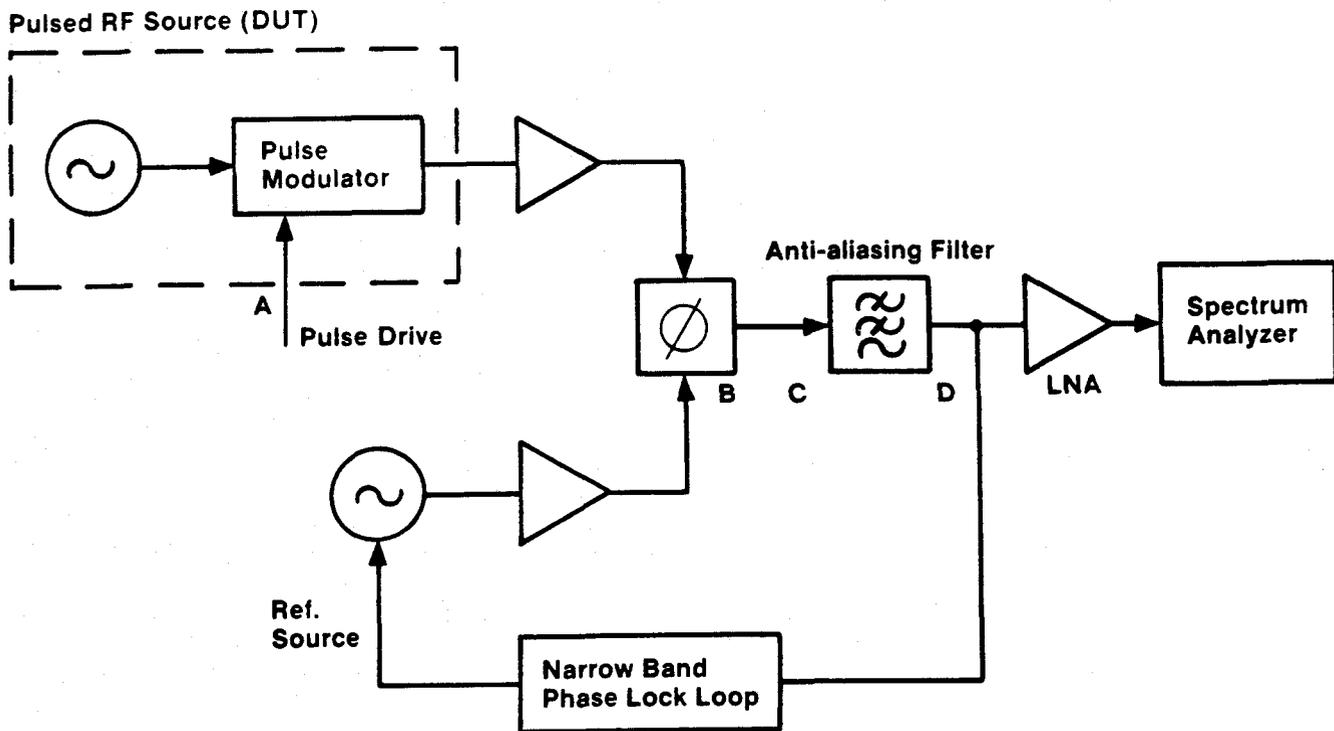
The HP 11729C may also be used in a delay line discriminator mode. The microwave reference signal generated by the 640 MHz signal down converts the source under test into the IF range. The power splitter and auxiliary IF output allows one to simply insert an appropriate delay line externally.

Delaying the signal at IF means lower delay line losses and also a wider choice of delay devices, like e.g. SAW delay line. Any signal generator with FM sidebands at IF or RF can serve as a calibration tool.

4.1.3. **Sensitivity of HP 11729C**
(Delay Line Frequency Discriminator Mode)



Phase Noise Measurement on Pulsed RF Signals



The baseband signal of a pulsed RF signal phaselocked to a reference source equals the baseband signal of an unpulsed RF signal gated by the pulse envelope (see Fig. A, B, C). The CW noise spectrum $\mathcal{L}(f)$ is multiplied by the envelope function of the pulse,

$$h(t) = \frac{\tau}{T} + \frac{2}{\pi} \sum_{n=1}^{\infty} \frac{(-1)^n}{n} \sin \frac{n\pi\tau}{T} \cdot \cos \frac{n2\cdot\pi t}{T}$$

Observed in the frequency domain, this gating process means convolving the two spectra to obtain the pulsed baseband spectrum (see Fig. C). The resulting spectrum has $\mathcal{L}(f)$ folded around multiples of the pulse repetition frequency (PRF). The relevant part of this spectrum is yielded by the DC term of the pulse envelope function, τ/T . It shows that the level of the unpulsed phase noise spectrum $\mathcal{L}(f)$ is reduced by the square of the duty cycle.

$$\mathcal{L}_{\text{pulsed}}^{(f)} = \left(\frac{\tau}{T}\right)^2 \mathcal{L}_{\text{unpulsed}}^{(f)}$$

A lowpass filter, the anti-aliasing filter, with a corner frequency of $< \text{PRF}/2$ needs to be employed to eliminate all the aliasing products (See Fig. D). It also serves to filter any phase transients associated with the pulse modulator which otherwise would limit the low noise amplifier.

The system sensitivity decreases with decreasing duty cycle. A sample and hold circuit in the baseband circuit, driven by the pulse signal, would improve the system sensitivity for low duty cycles. A simpler way to reduce system noise in the pulse off period is to gate either the LO or the R drive of the mixer-phase detector.

A side effect of phase locking to a pulsed RF signal is the reduction of the effective loop bandwidth by τ/T . For duty cycles $< 5\%$, it becomes increasingly important to have a well balanced mixer-phase detector.

Time Domain

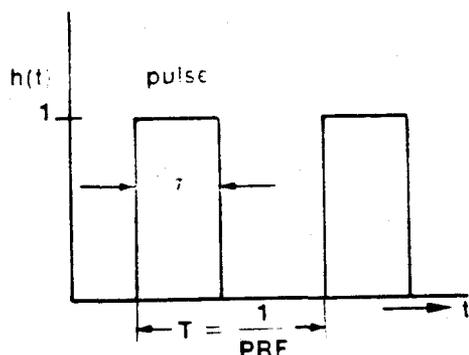
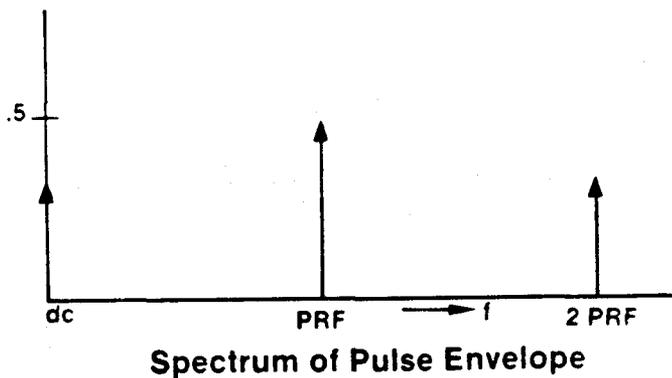


Fig A: Pulse Envelope

Frequency Domain



Spectrum of Pulse Envelope

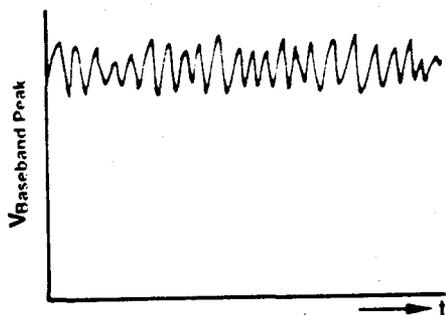
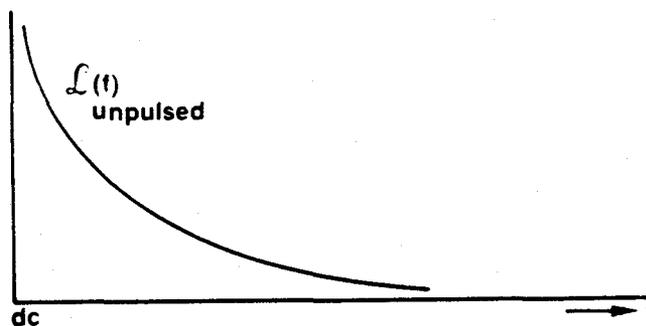


Fig B: Unpulsed Baseband Signal (Unpulsed Phase Noise)



Unpulsed Phase Noise Spectrum

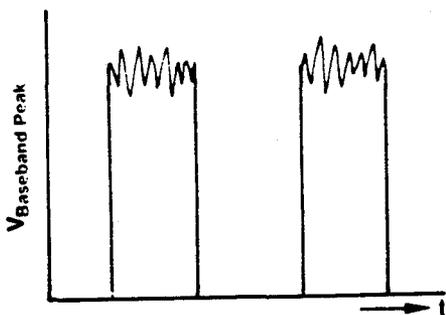
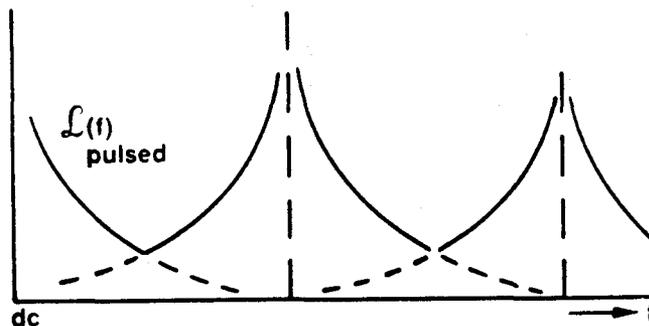


Fig C: Pulsed Baseband Signal before Low Pass



Pulsed Phase Noise Spectrum before Low Pass

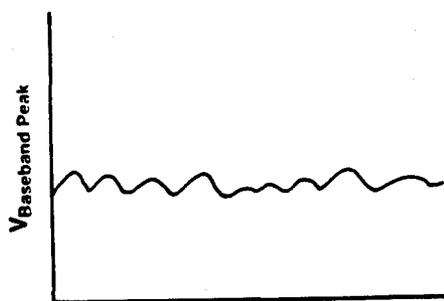
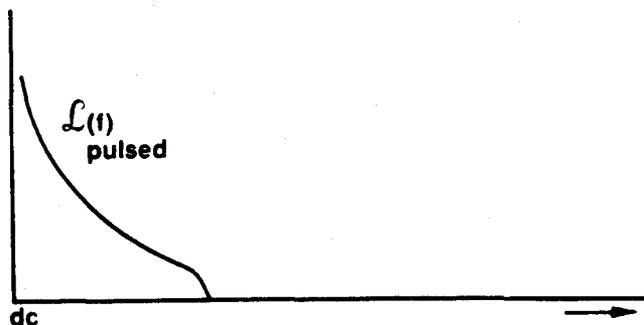
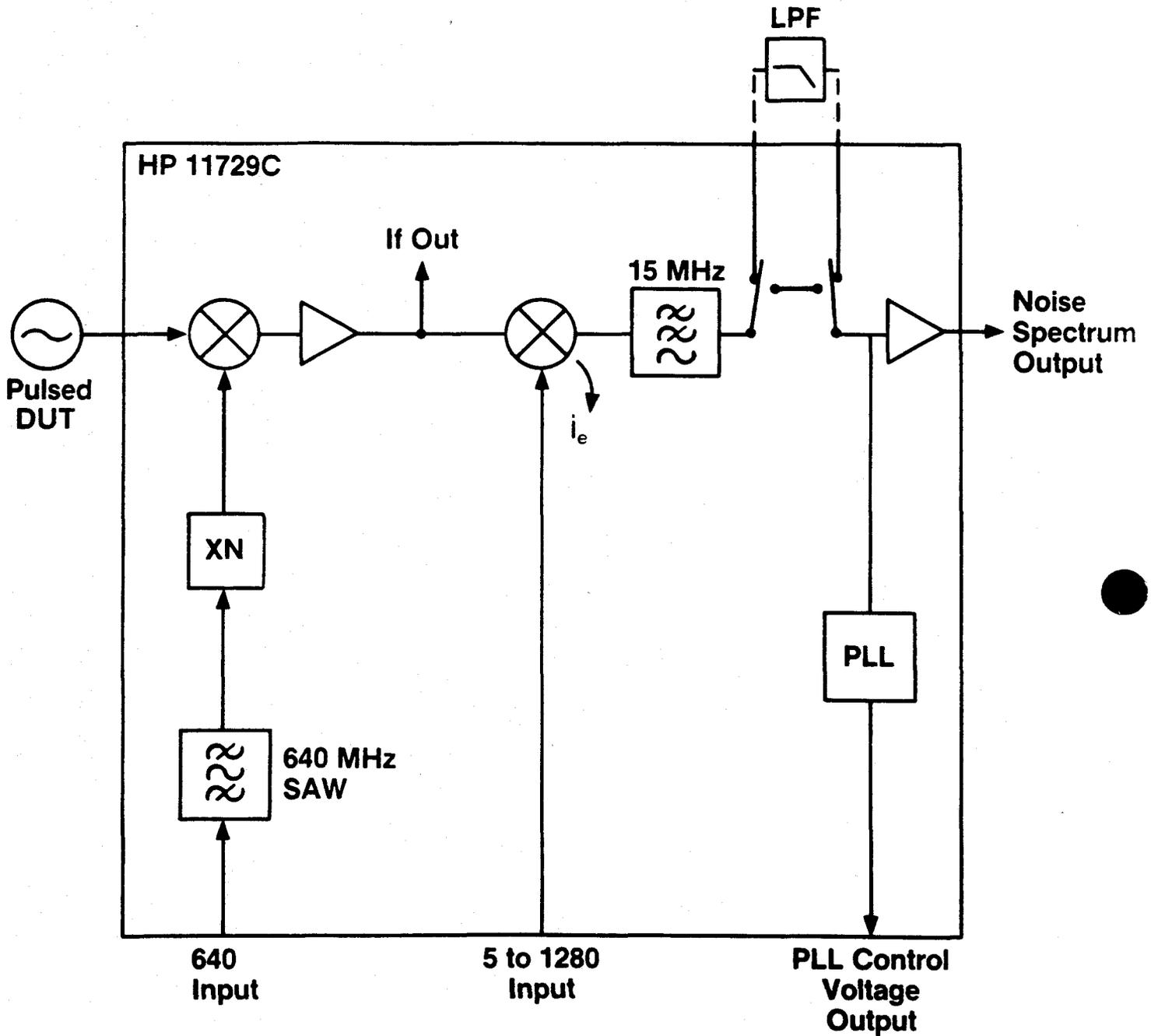


Fig D: Pulsed Baseband Signal after Low Pass



Pulsed Phase Noise Spectrum after Low Pass

4.2.2. Pulsed Phase Noise Measurements
with the HP 11729C (Phase Detector Mode)

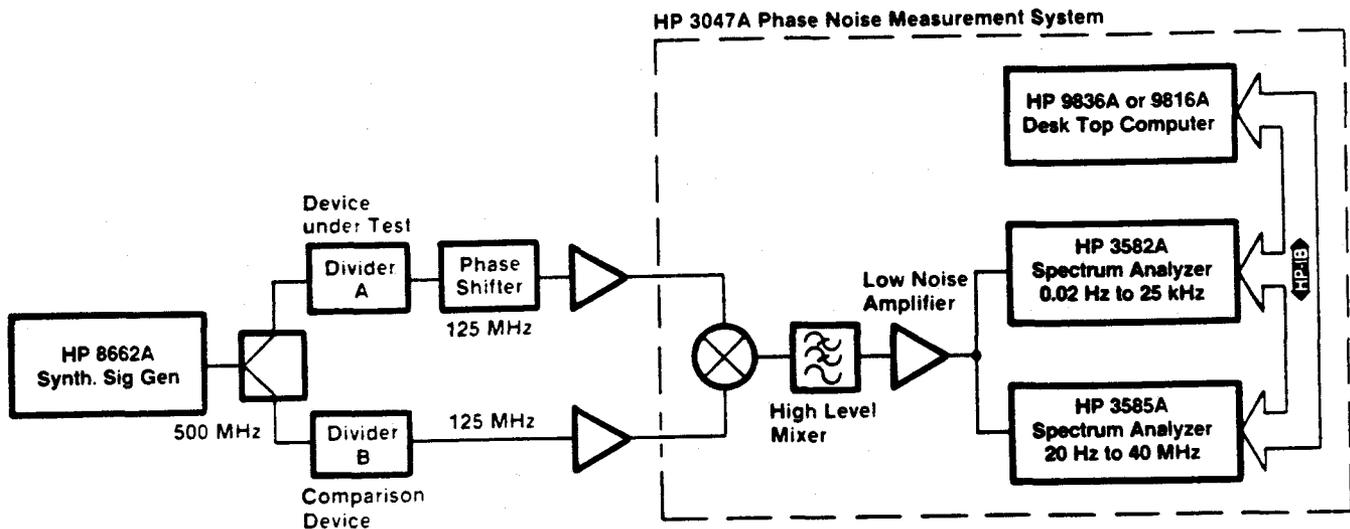


Selecting pulse mode (phase noise) on the HP 11729C applies an error current to the mixer/phase detector and switches in a user-supplied low pass filter to remove the pulse repetition frequency feedthrough.

HP 3047A Phase Noise Measurement System

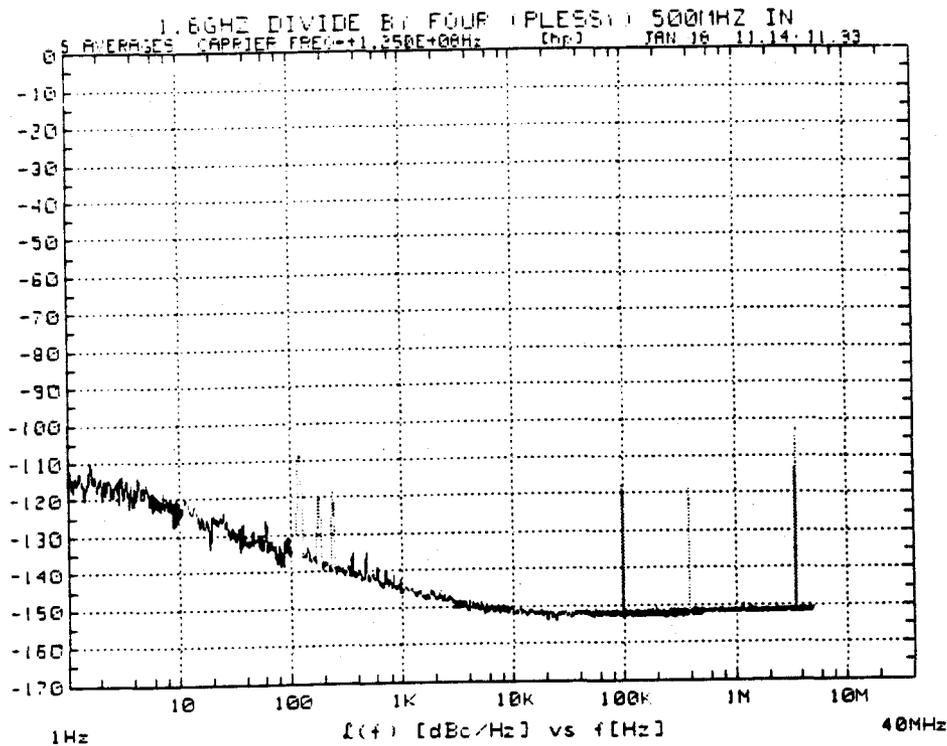
Application Example:

Two-port phase noise measurement on frequency divider

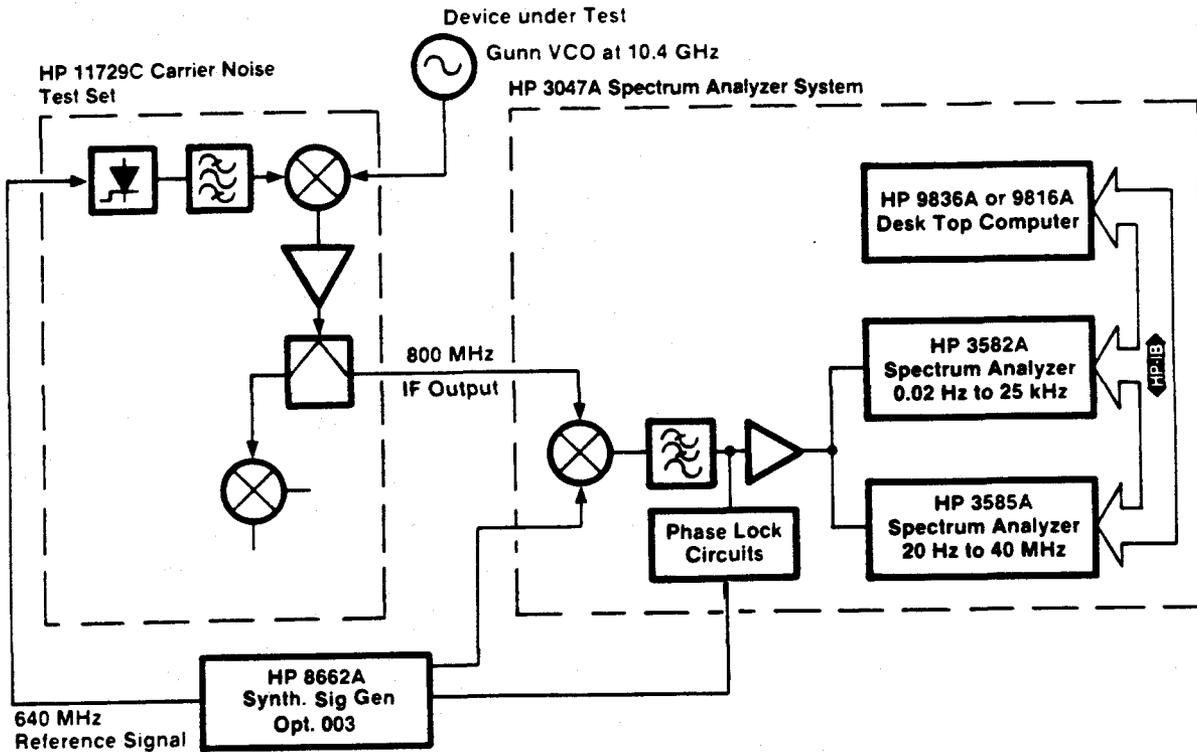


The HP 3047A Phase Noise Measurement System represents a high sensitivity phase detector system with controllable phase lock circuits. Extensive software automates the calibration and measurement procedures.

Determining the phase noise of frequency dividers is an example of a residual phase noise measurement. Here, the phase noise of the 8662 is largely canceled and the system will measure the rms sum of noise of both dividers.

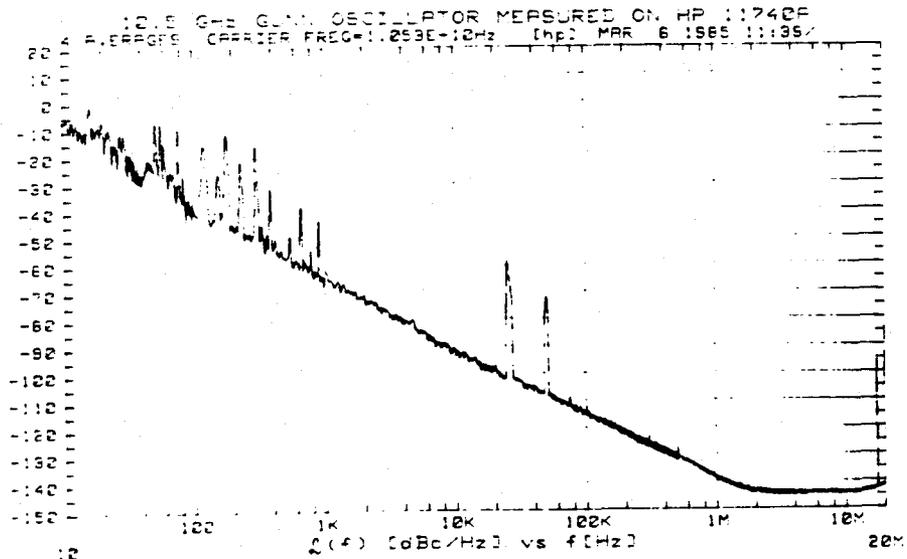


HP 11740A Microwave Phase Noise Measurement System

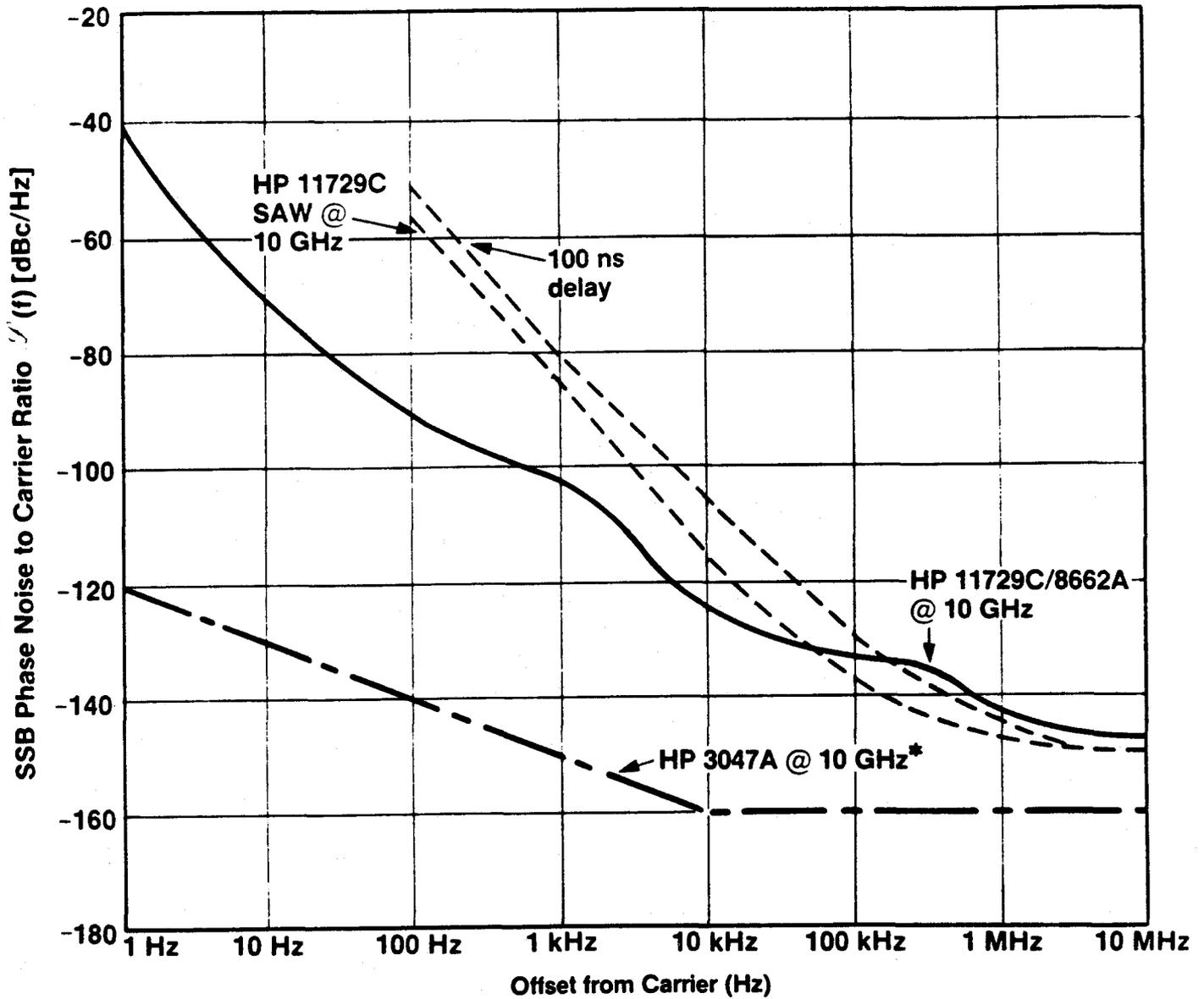


In the HP 11740A Microwave Noise Measurement System, the HP 11729C provides a broadband and low noise microwave reference signal eliminating the need for a second comparable source.

The HP 3047A phase detector system then compares the IF output of the HP 11729C with either the 5–1280 MHz output of the HP 8662A or any other suitable and tunable IF source.

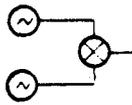
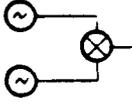
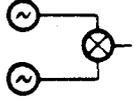
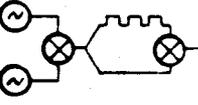


Comparison of System Sensitivities



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

4.6 Phase Noise Measurement Systems Comparison

	HP 11729C/8662A Phase Noise Test System (phase detector mode)	HP 3047A Phase Noise Measurement System	HP 11740A Microwave Phase Noise Measurement System	HP 11729C Carrier Noise Test Set (frequency discriminator mode)
	<div style="border: 1px dashed black; padding: 5px; margin-bottom: 5px;"> HP 8566A HP 8566A B HP 3585A HP3582A Spectrum Analyzer </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 11729C Carrier Noise Test Set </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 8662A Synthesized Sig Gen </div> <div style="border: 1px dashed black; padding: 5px; margin-bottom: 5px; width: fit-content; margin-left: auto;"> Controller </div>	<div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 3585A Spectrum Analyzer </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 3582A Spectrum Analyzer </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 35601A S.A. Interface </div> <div style="border: 1px dashed black; padding: 5px; margin-bottom: 5px;"> User-supplied Reference HP 8662A HP 8663A HP 8340A HP 8672A HP 8673B </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 9836A or 9816A </div>	<div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 3585A Spectrum Analyzer </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 3582A Spectrum Analyzer </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 35601A S.A. Interface </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 11729C Carrier Noise Test Set </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 8662A or 8663A Synthesized Sig Gen </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 9836A or 9816A </div>	<div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> HP 11729C Carrier Noise Test Set </div> <div style="border: 1px dashed black; padding: 5px; margin-bottom: 5px;"> HP 8566A/B HP 8568A/B HP 3561A Spectrum Analyzer </div> <div style="border: 1px solid black; padding: 5px; margin-bottom: 5px;"> Controller </div>
Principle of Operation	Source Under Test  Reference	Source Under Test  User-Supplied Reference	Source Under Test  Reference	Source Under Test  Reference
Frequency Range	10 MHz — 18 GHz	5 MHz — 18 GHz	5 MHz — 18 GHz	10 MHz — 18 GHz
Low Noise Microwave Reference Signal	yes	no	yes	yes
Typical System Noise at 10 GHz	123 dBc at 10 kHz -145 dBc noise floor	-170 dBc at 10 kHz* -170 dBc noise floor	-123 dBc at 10 kHz -145 dBc noise floor (with HP 11729C)	-112 dBc at 10 kHz -145 dBc noise floor
Offset Frequency	<10 MHz	<40 MHz	<10 MHz (<40 MHz)	<10 MHz
Two-port Noise	no	yes	yes	no
AM-Noise (Microwave)	yes	w/external detector	yes	yes
Manual Operation	yes	no	yes	yes
Automatic Operation (HP-IB)	yes	yes	yes	yes
Software	user provided	extensive	extensive	user provided

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

GLOSSARY OF SYMBOLS

FkTB	[mW]	Available noise power in bandwidth B
f_o	[Hz]	Carrier frequency
f_c	[Hz]	Corner frequency of flicker noise
f	[Hz]	Fourier frequency (sideband-, offset-, modulation, baseband-frequency)
f(t)	[Hz]	Instantaneous frequency
Δf(t)	[Hz]	Instantaneous frequency fluctuation
K_φ	[volt/rad]	Phase detector constant
K_D	[volt/Hz]	Frequency discriminator constant
L(f)	[dBc/Hz]	Ratio of single sideband phase noise to total signal power in a 1 Hz bandwidth f Hertz from the carrier
P_s	[mW]	Signal power
P_{ssb}	[mW]	Power of single sideband
Q		Quality factor of resonator
S_{Δf(f)}	[Hz²/Hz]	Spectral density of frequency fluctuations
S_{φ(f)}	[rad²/Hz]	Spectral density of phase fluctuations
t	[sec]	Time
v(t)	[volt]	Instantaneous voltage
V_{B peak}	[volt]	Peak voltage of sinusoidal beat signal
Δφ(t)	[rad]	Instantaneous phase fluctuation
Δθ(t)	[rad]	Instantaneous phase perturbation
τ	[sec]	Delay time

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