The "Art" of Phase Noise Measurement

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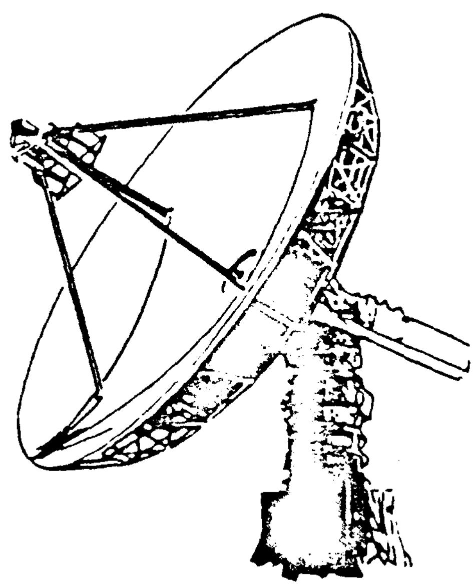
May 1983

RF & Microwave Measurement Symposium and Exhibition



HEWLETT PACKARD

INTRODUCTION

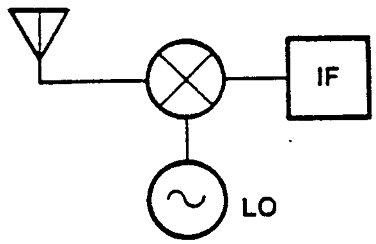
Measuring and specifying phase noise has become increasingly important as phase noise is the limiting factor in many RF and microwave systems, like Doppler radar and space telemetry systems or communication links. The complexity and subtleties of the measurement have earned it the reputation of being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

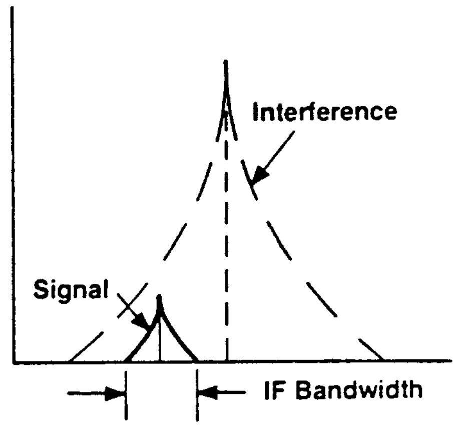
This paper gives an introduction to phase noise measurement focusing on the two most common and useful techniques, the "Two Source Phase Detector" and the "Delay Line Frequency Discriminator" methods. System limitations are pointed out as well as the potential sources of erroneous data when phase noise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

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

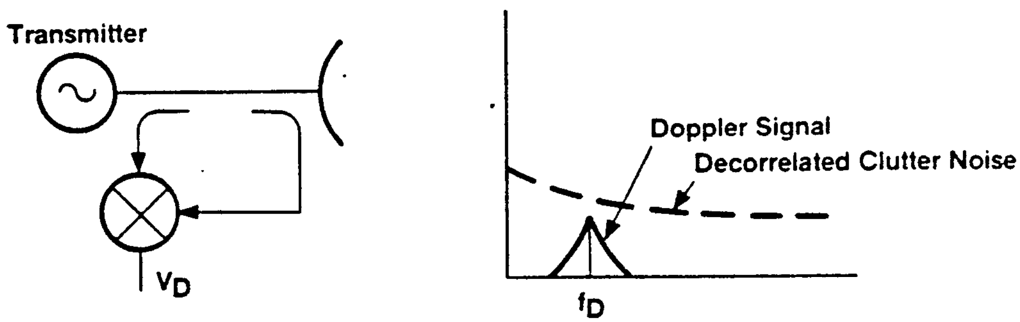
1. Why Measure Phase Noise?

Phase noise causes . . .

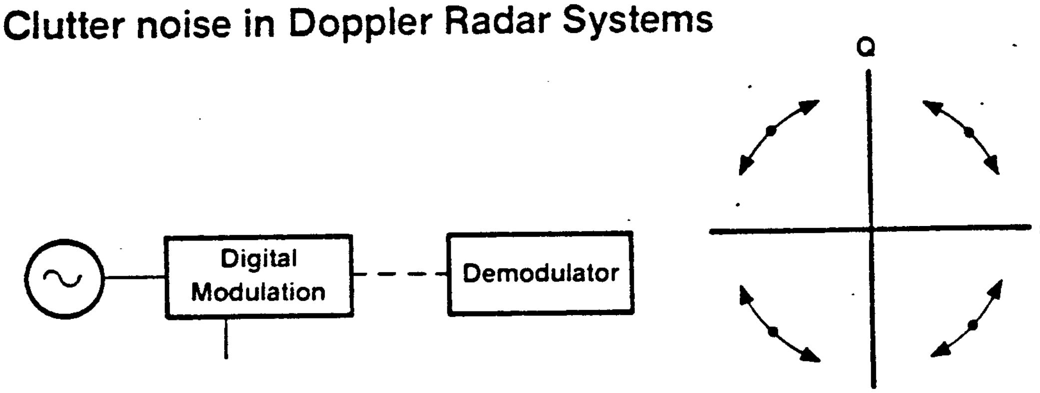




- **Lower receiver sensitivity in multi-signal environment**



- **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 interferring 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.

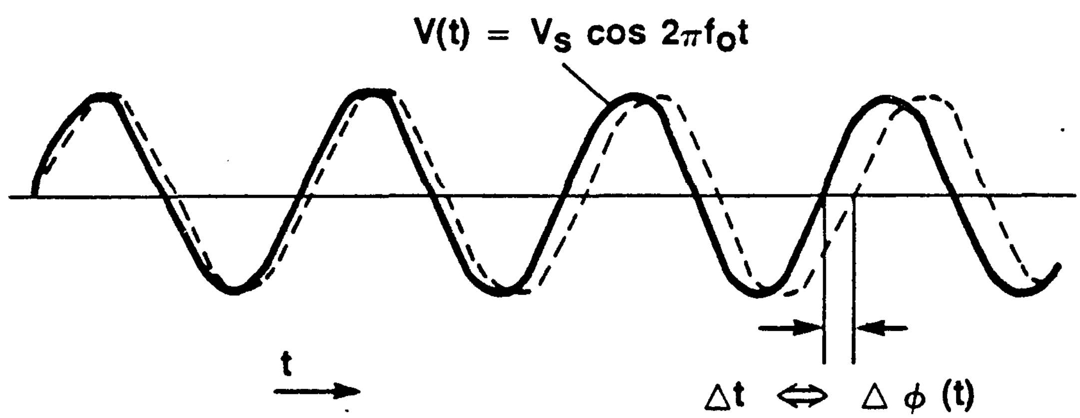
2. Basic Representation of Phase Noise

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.

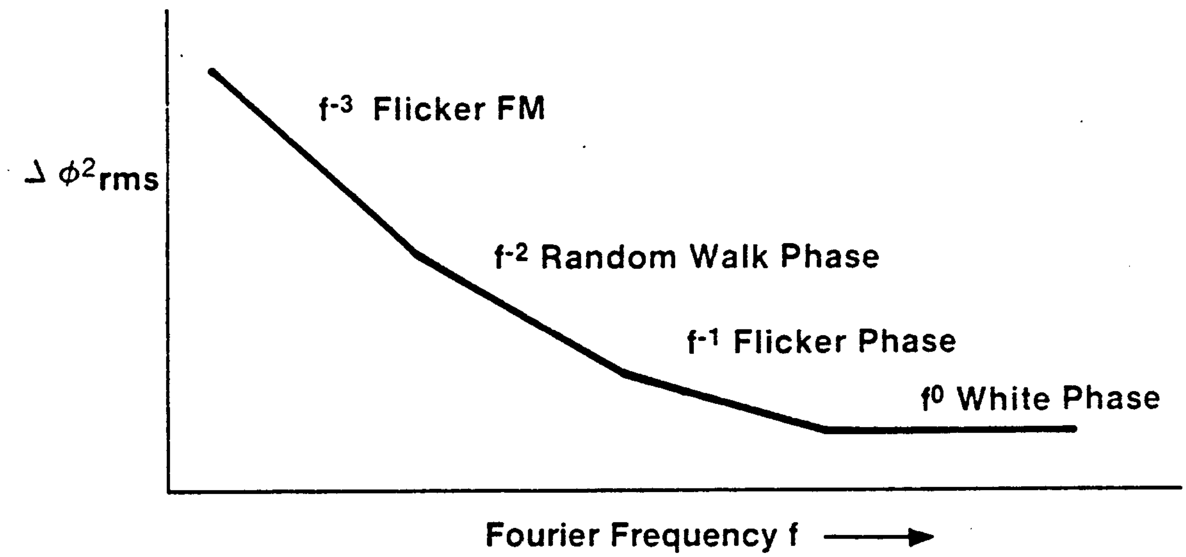
Phase or frequency fluctuations are the same physical phenomena. One can be derived from the other as angular frequency is the first derivative of phase with respect to time.

2.2 In the Frequency Domain:

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

Spectral Density of Phase Fluctuations $S_{\Delta\phi}$

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



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

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

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.

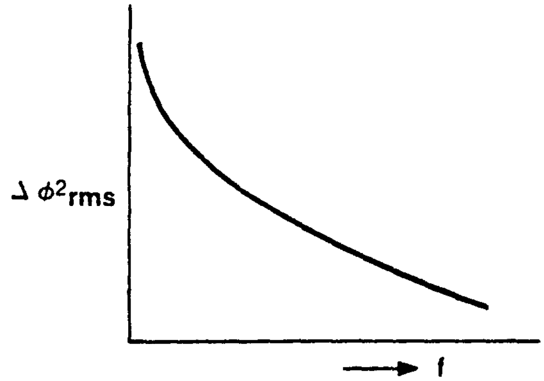
2.3 Relation of Phase Noise to Carrier Sideband Noise

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

$(\Delta\phi << 1)$

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

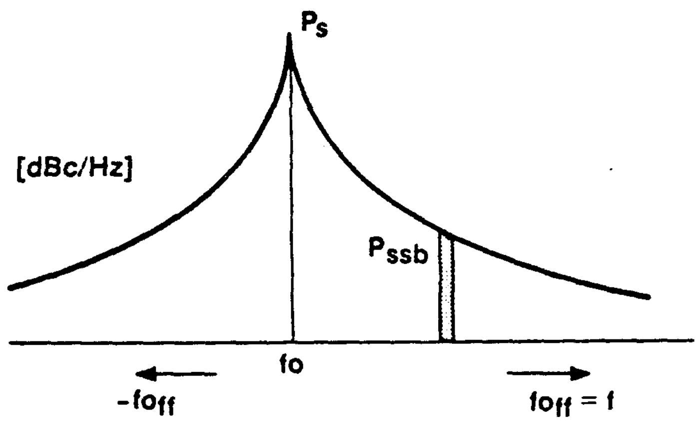


A common, though indirect, representation of phase noise is $\mathscr{L}(f)$, the ratio of the single sideband noise power in a 1 Hz bandwidth to the total carrier power specified at a given offset f of the carrier.

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

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

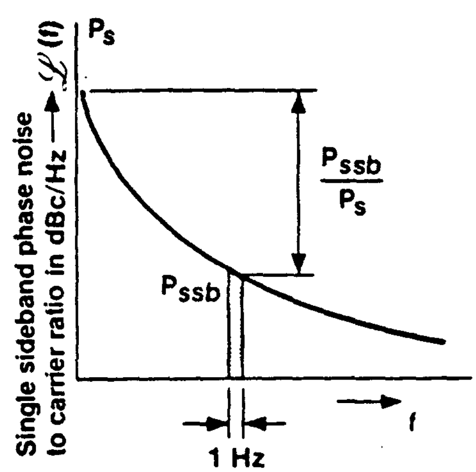
$$\mathscr{L}(f) = \frac{P_{ssb}(\text{per 1 Hz})}{P_s} \quad [\text{dBc/Hz}]$$



Relation between $\mathscr{L}$, $S_{\Delta\phi}$, $S_{\Delta f}$:

for $\Delta\phi << 1$

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



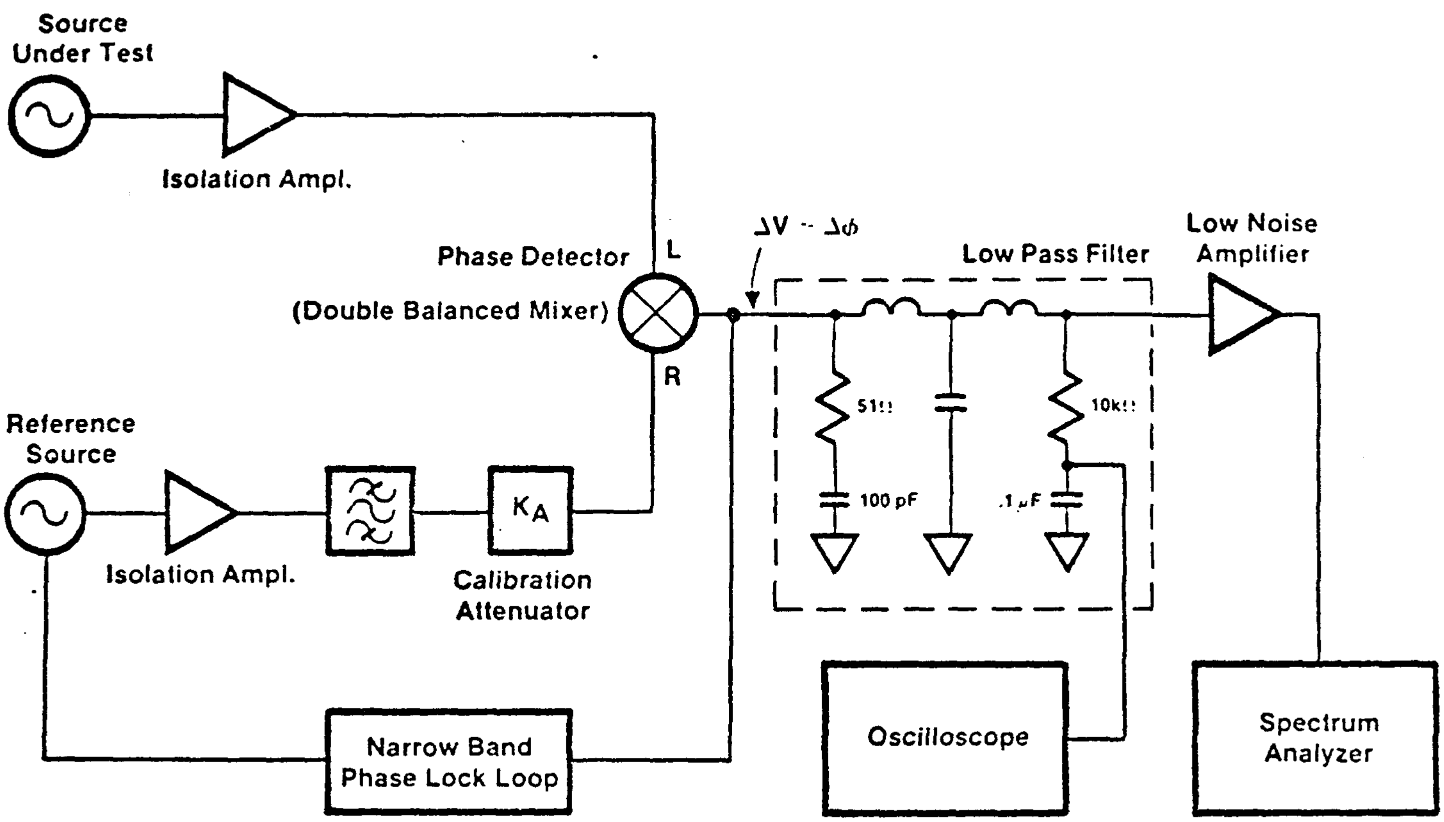
3. Phase Noise Measurement Techniques

- 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 ou and explained in some detail on the following pages: the "Two Sources and Phase Detector" technique an the "Delay Line Frequency Discriminator" technique.

Both methods are relatively simple to implement and are broadband solutions. They also complimer each other well. The first method measures phase noise; high sensitivity is obtainable and two sources ar needed. The second method measures frequency noise, has inherently lower sensitivity close in but n second source is required.

3.1 Two Sources and Phase Detector Technique



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

$\mathcal{L}(f)$

(for $\Delta\phi << 1$)

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

for $\Delta\phi << 1$

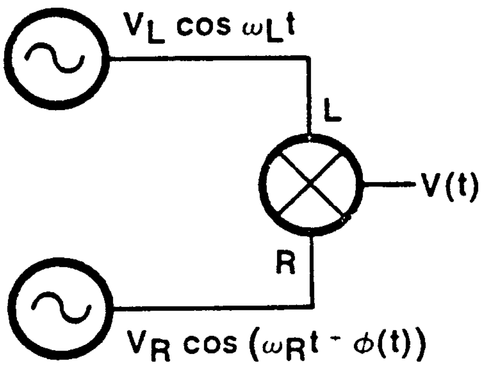
$$\mathcal{L}(f) = 1/2\ S_{\Delta\phi}(f) = 1/4\ \frac{\Delta V^2_{rms}(1\ Hz)}{V^2_{B\ rms}}$$

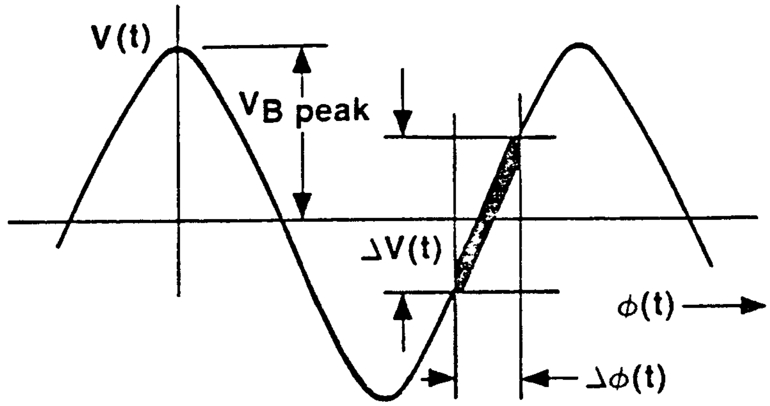
With the LO and RF input in phase quadrature steady state wise, the mixer IF output generates a signal proportional to the phase difference of the two sources. In linear operation the phase detector constant equals the peak voltage of the sinusoidal beat signal produced when either source is frequency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





$$V(t) = \underbrace{K_L V_R}_{V_{B\,peak}} \cos\big((\omega_R - \omega_L)\,t + \phi(t)\big) + \cancel{K_L V_R \cos\big((\omega_R + \omega_L)\,t + \phi(t)\big)} + \ldots$$

$$V(t) = V_{B\,peak} \cos\big((\omega_R - \omega_L)\,t + \phi(t)\big)$$

K_L = Conversion Loss

$V_{B\,peak}$ = Peak Voltage of Beat Signal

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

$$\Delta V(t) = (\pm)\, V_{B\,peak} \sin \Delta\phi(t)$$

for $\Delta\phi \ll 1$

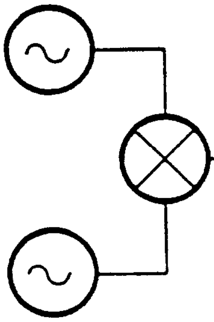
$$\Delta V(t) = (\pm)\, K_\phi\, \Delta\phi(t) \qquad K_\phi = V_{B\,peak} \left[\frac{\text{Volts}}{\text{rad}}\right]$$

The frequency conversion of signals applied to a mixer is physically based on one signal gating th 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 inpu signals.

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

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 a a phase detector with a phase detector constant of K_ϕ equal to $V_{B\,peak}$.

3.1.2 Calibration with Beat Signal



$$\Delta V(f) = K_\phi \, \Delta \phi (f) \qquad K_\phi = V_{B\ peak} = \sqrt{2} V_{B\ rms}$$

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

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

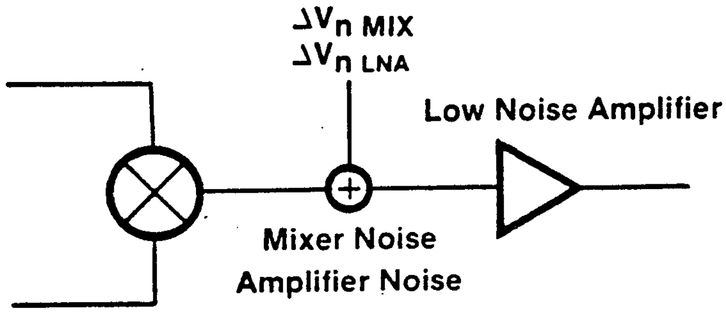
for $\Delta \phi << 1$ $$\mathcal{L}(f) = 1/2 \; S_{\Delta\phi}(f) = 1/4 \; \frac{\Delta V^2_{rms}(1\ Hz)}{V^2_{B\ rms}}$$

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

Assuming $V_{B\ peak}$ 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.

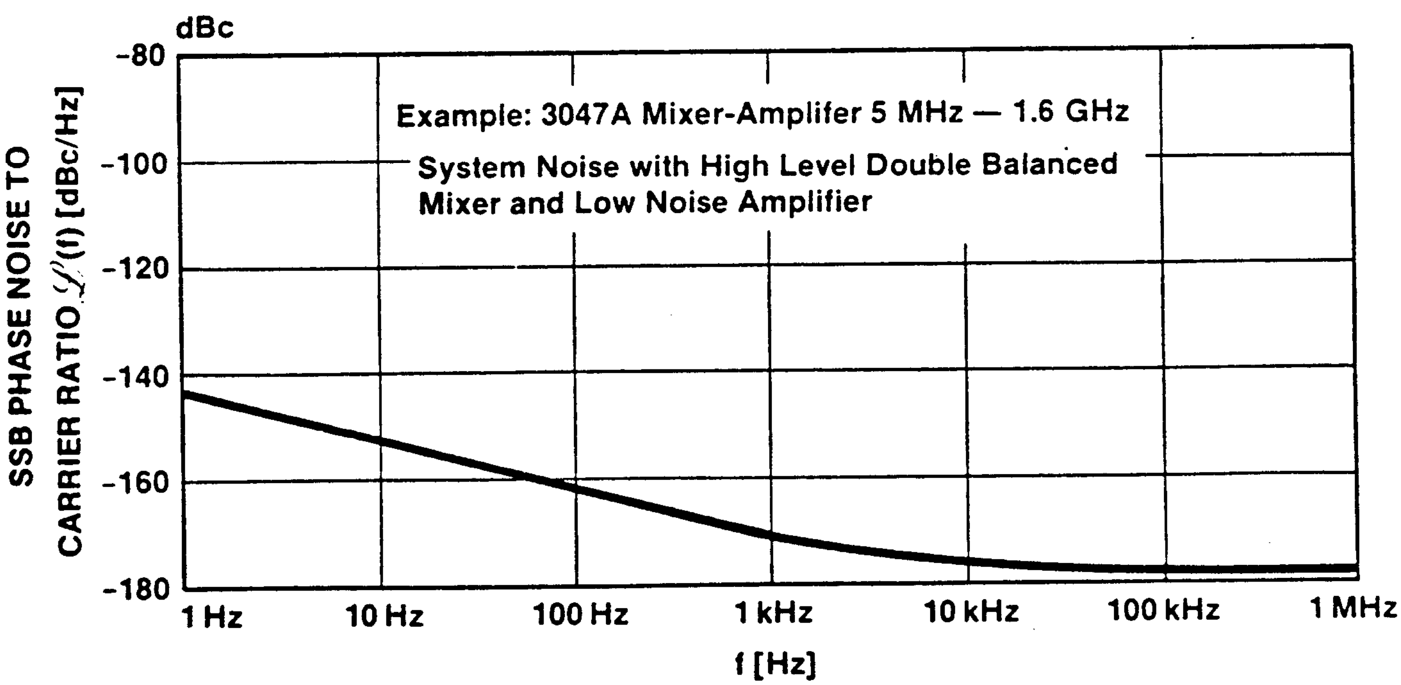
Other calibration methods are based on the characterization of the phase slope, on calibrated PM or FM sidebands on one of the two sources or on the injection of a calibrated spurious signal.

3.1.3 System Noise Floor



$$\mathcal{L}\ \text{system}\ (f) = \frac{1}{4}\ \frac{\Delta V^2_{n\ MIX} + \Delta V^2_{n\ LNA}}{V^2_{B\ rms}}$$

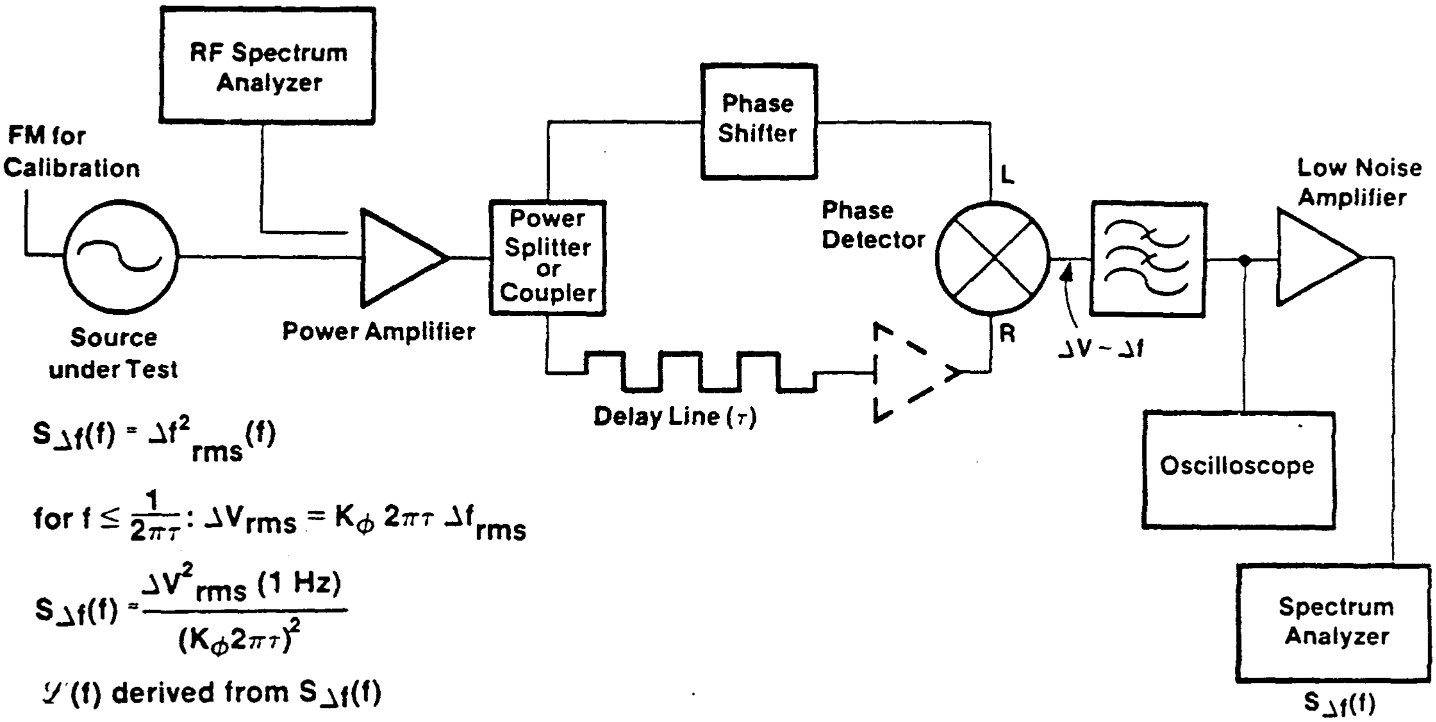
The noise floor of the measurement system is established by the equivalent amplifier and mixer noise (shot noise, flicker noise) at the IF port reduced by the phase detector gain K_ϕ ($V_{B\ peak}$ for linear operation). System sensitivity is therefore maximized by using a high level mixer, selecting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



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

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

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

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

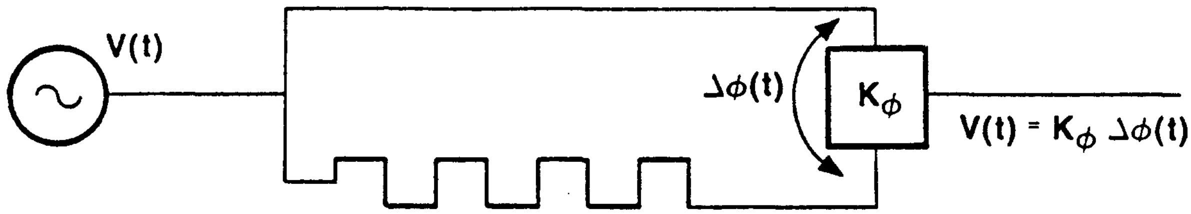
for $\Delta\Phi \ll 1$: $\mathcal{L}(f) = \frac{1}{2}\frac{1}{f^2} S_{\Delta f}(f) = \frac{1}{2}\frac{1}{f^2}\frac{\Delta V^2_{rms}\,(1\ \mathrm{Hz})}{(K_\phi 2\pi\tau)^2}$

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



Delay time τ

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

$V(t) = V_o \cos\left(2\pi f_o t + \frac{\Delta f}{f} \cos 2\pi f t\right)$

Carrier with sinusoidal FM

f_o = Carrier frequency

f = FM rate

Δf = FM deviation

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

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

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

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

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

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

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

Amplitude Response:

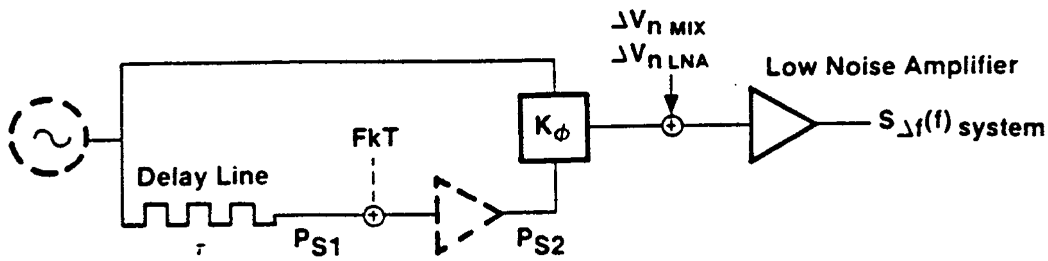
$$\Delta V = K_\phi \, 2 \frac{\Delta f}{f} \sin \pi f \tau = K_\phi \, 2\pi\tau \, \Delta f \, \frac{\sin \pi f \tau}{\pi f \tau}$$

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

$$\Delta V = \underbrace{K_\phi \, 2\pi\tau}_{} \, \Delta f$$

$K_D \left[\frac{V}{Hz}\right]$ = 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): $K_\phi = V_{B\ peak} = \sqrt{2\, P_{S2}\, L \times 50\Omega}$

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.

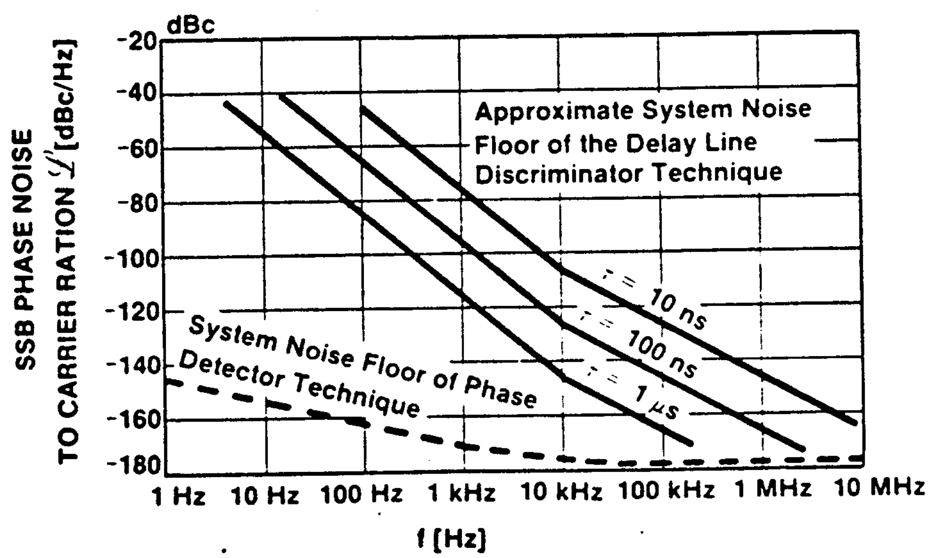
$$S_{\Delta f}(f)_{system} = \frac{\Delta V^2_{n\ MIX}}{(K_\phi 2\pi\tau)^2} + \frac{\Delta V^2_{n\ LNA}}{(K_\phi 2\pi\tau)^2} + \frac{\Delta\Theta^2_{RF}}{(2\pi\tau)^2}$$

$$\mathcal{L}(f) = \frac{1}{2}\frac{1}{f^2} S_{\Delta f}(f)_{system} \qquad \text{for } \Delta\phi << 1$$

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

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, L(f) increases with f^{-2} and f^{-3} towards the carrier. This means good sensitivity far out (limited by the sinx/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 $\Delta\Theta^2_1$ (or $\mathcal{L}_1$). The delay line discriminator-interpreting $\Delta\Theta$ as frequency noise enhances this equivalent phase noise by a factor of $1/2\pi f\tau$.

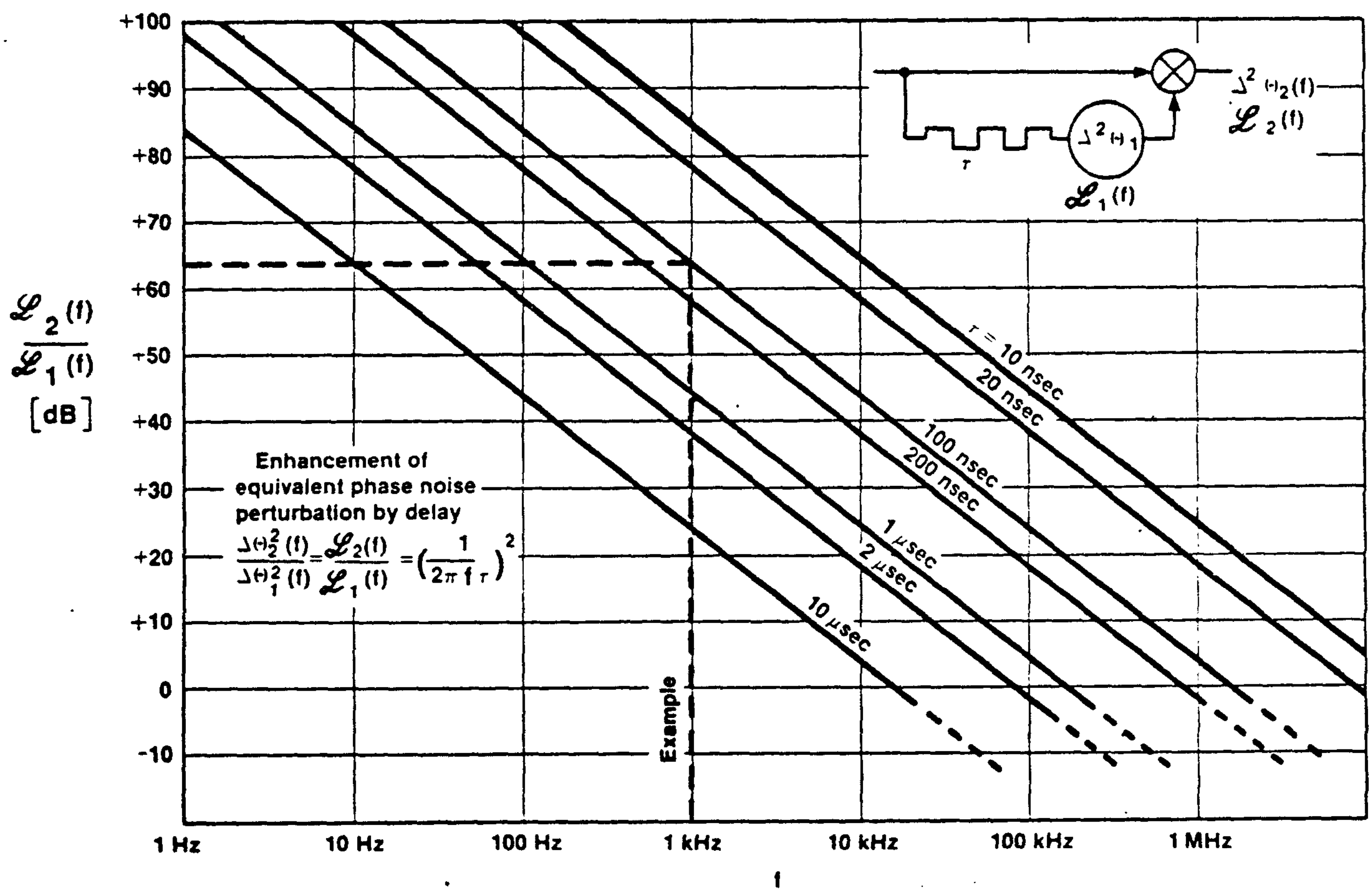
Example:

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

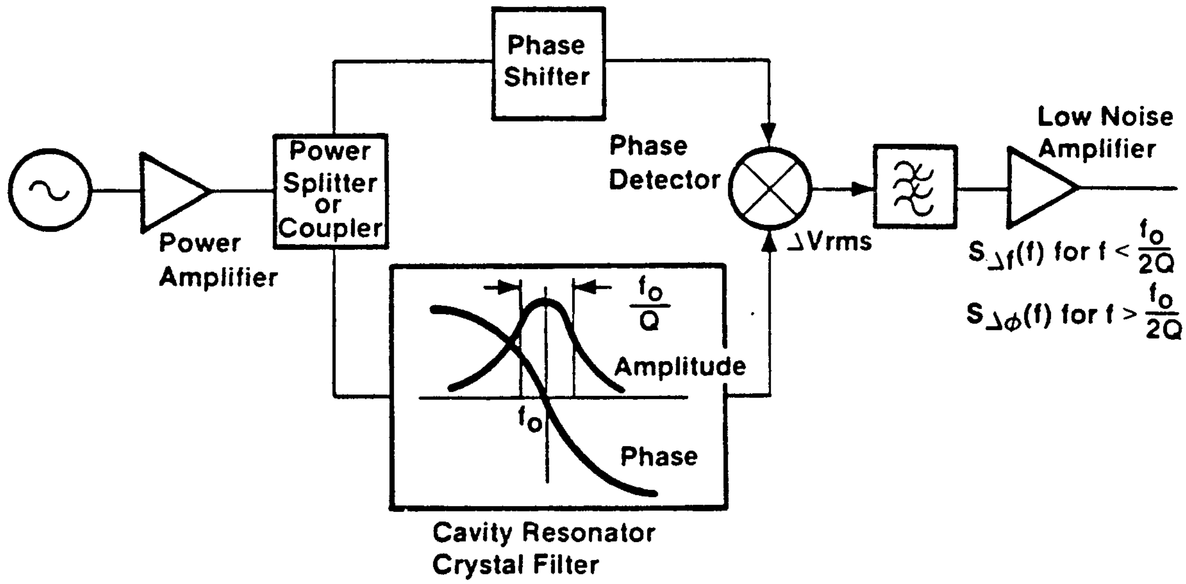
With a 100 nsec delay time the resulting system noise floor is

$$\mathcal{L}_2 = \mathcal{L}_1 \left(\frac{1}{2\pi f\tau}\right)^2.$$

$$= -160 \text{ dBc} + 64 \text{ dB} = -96 \text{ dBc/Hz}$$



3.2.3 Frequency Discriminator with Resonator



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

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

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

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

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 $Q/\pi f_o$ for Fourier frequencies less than half the bandwidth. Inside $f_o/2Q$, the system therefore measures the spectral density of frequency noise with a discriminator constant of $K_\phi\, 2Q/f_o$.

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

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

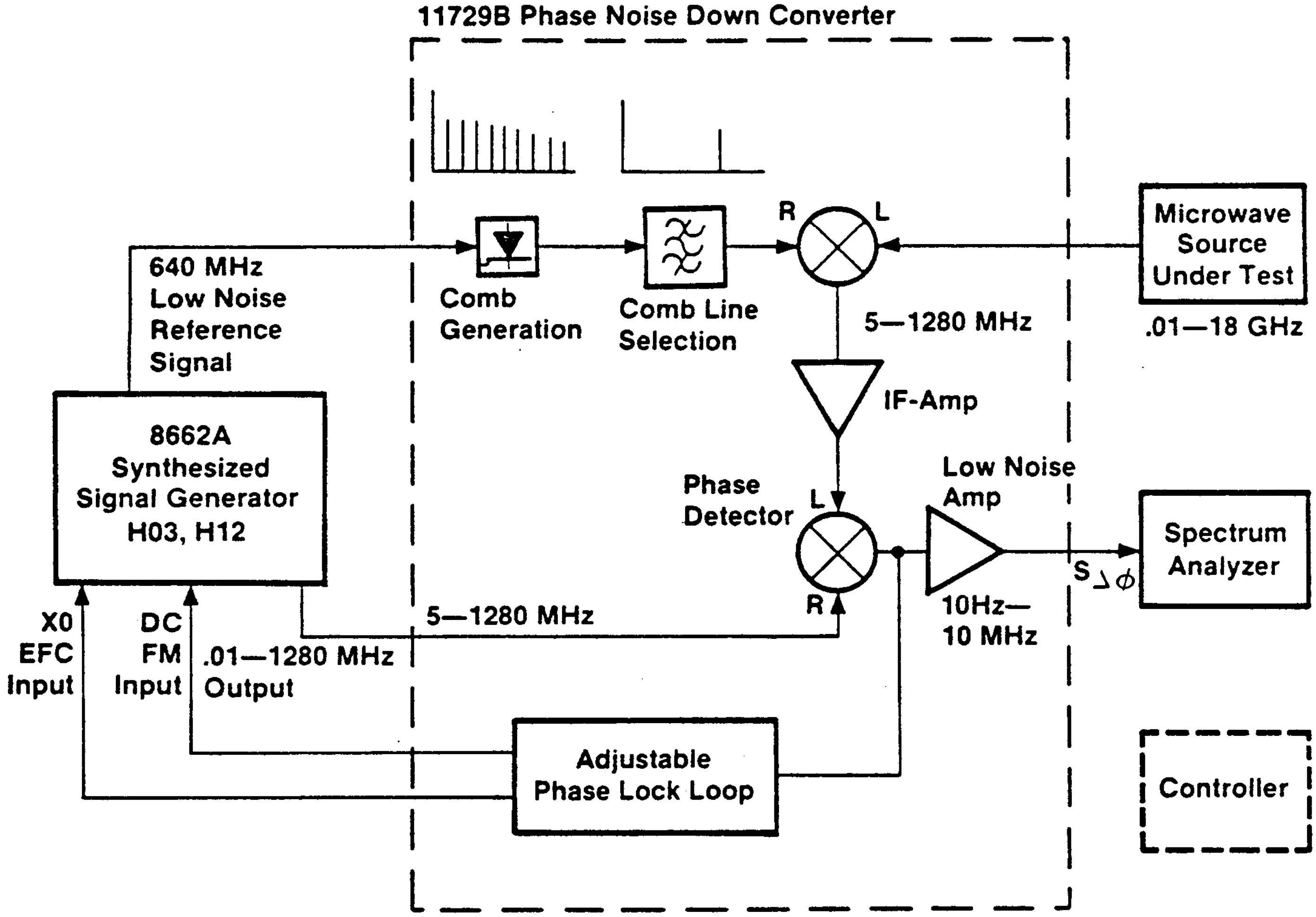
Outside $f_o/2Q$, the system may be viewed as 2 signals applied to a phase detector with the sideband noise of one signal stripped off by the cavity resonator. The system then measures $\Delta\phi^2_{rms}$ with a phase detector constant K_ϕ.

4. HP Phase Noise Measurement Solutions

- 11729B Phase Noise Test System
- 3047A Spectrum Analyzer System
- 3047A/11729B Phase Noise Test System
- 5390A Frequency Stability Analyzer

Following is a comparison of HP phase noise measurement solutions with particular focus on the 11729B Phase Noise Down Converter, HP's most recent contribution to phase noise measurement.

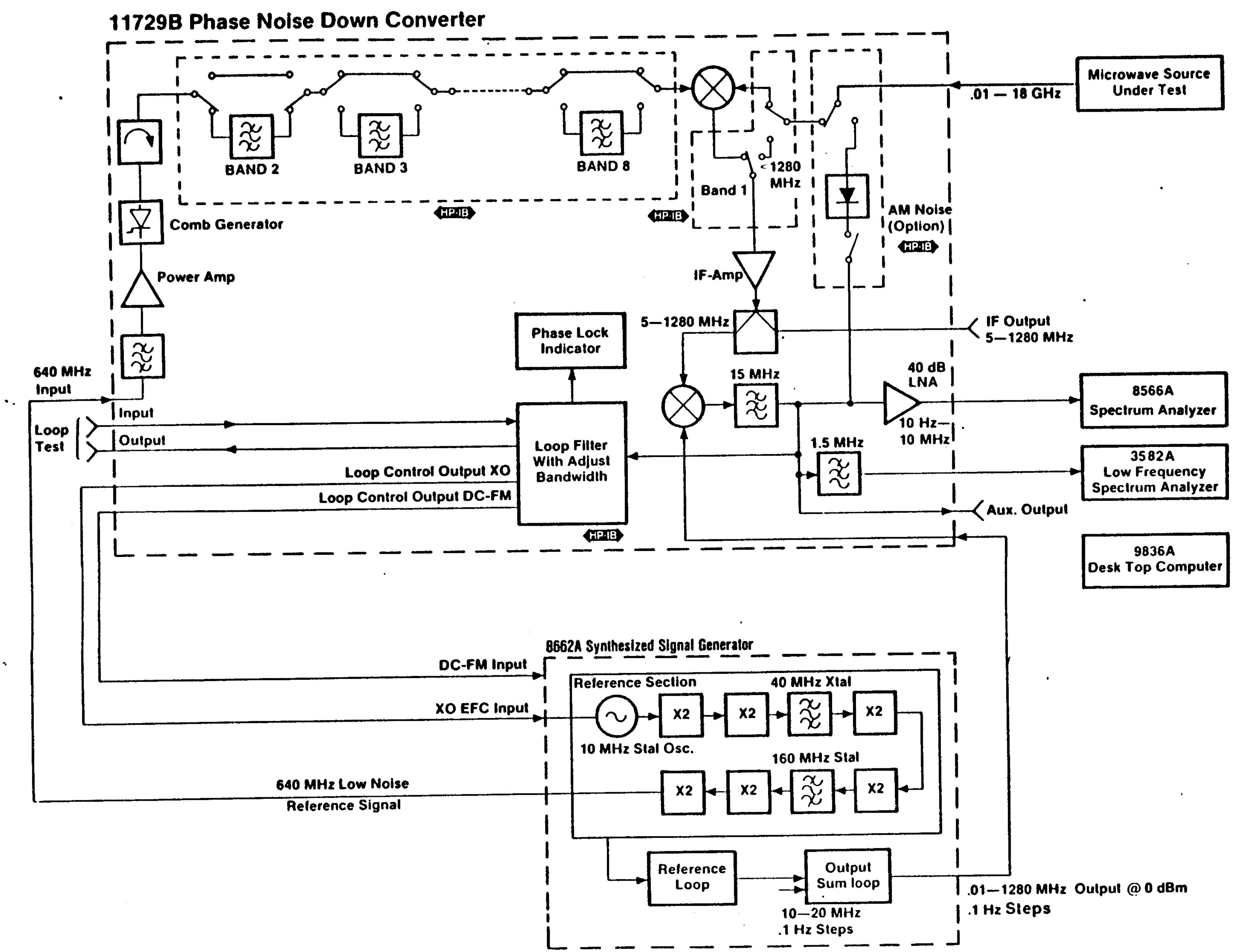
4.1 11729B Phase Noise Test System
Principle of Operation



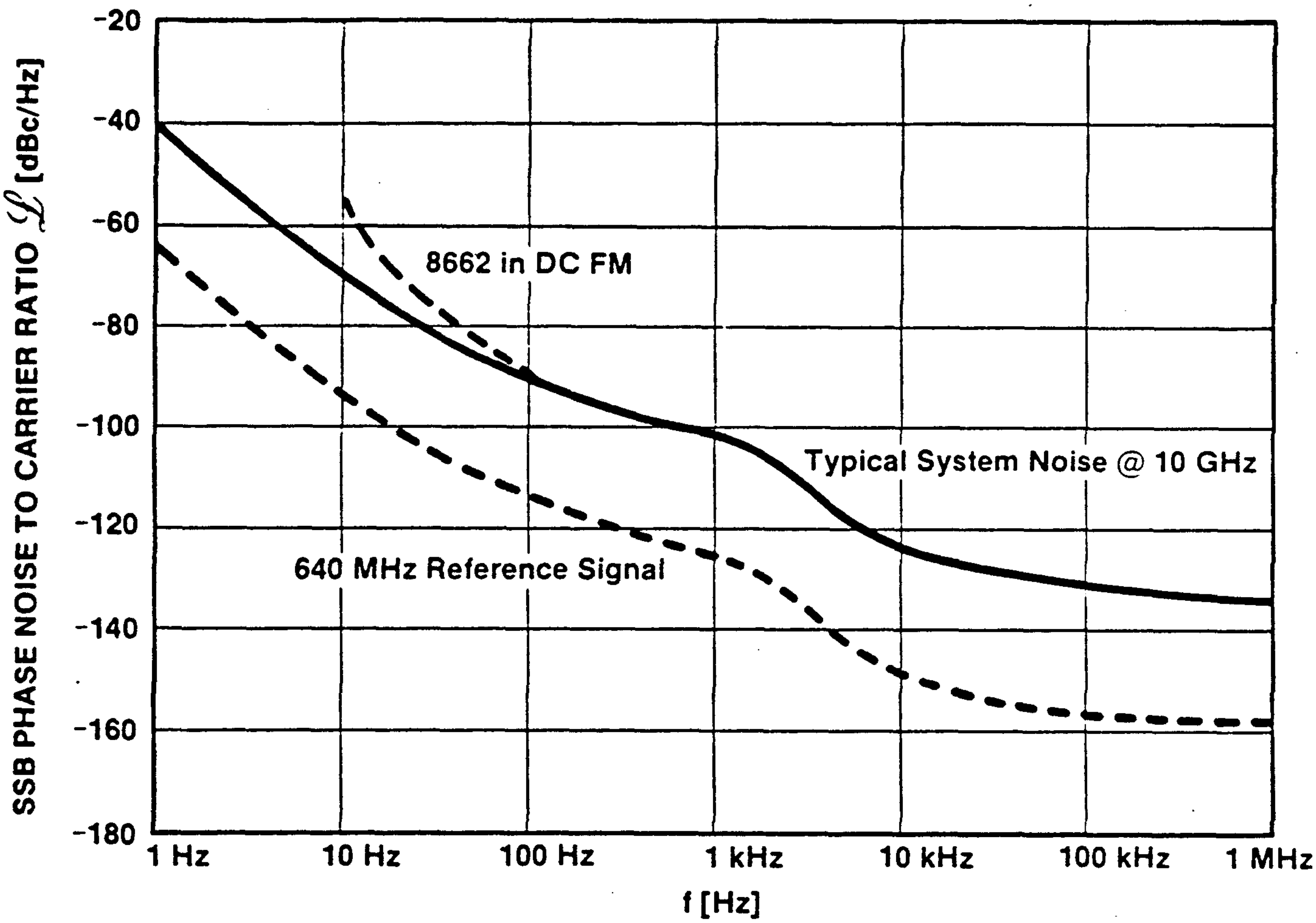
The 8662A Synthesized Signal Generator provides a state of the art low noise reference signal at 640 MHz to the 11729B Phase Noise Down Converter. This signal is applied to a step recovery diode multiplier which generates a comb of signals spaced by 640 MHz ranging beyond 18 GHz. A switched in filter selects an appropriate combline. 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 8662A (.01 to 1280 MHz). Phase quadrature is enforced by phase locking the 8662A to the signal under test either thru the limited electronic control of the 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.


11729B Phase Noise Down Converter
Microwave Source Under Test
.01 — 18 GHz
BAND 2
BAND 3
BAND 8
HP-IB
Comb Generator
Power Amp
Band 1
< 1280 MHz
AM Noise (Option)
IF-Amp
5—1280 MHz
IF Output 5—1280 MHz
Phase Lock Indicator
15 MHz
40 dB LNA
10 Hz—10 MHz
8566A Spectrum Analyzer
1.5 MHz
3582A Low Frequency Spectrum Analyzer
Aux. Output
9836A Desk Top Computer
640 MHz Input
Loop Test
Input
Output
Loop Filter With Adjust Bandwidth
Loop Control Output XO
Loop Control Output DC-FM
8662A Synthesized Signal Generator
DC-FM Input
XO EFC Input
Reference Section
10 MHz Stal Osc.
X2
40 MHz Xtal
160 MHz Stal
640 MHz Low Noise Reference Signal
Reference Loop
Output Sum loop
10—20 MHz .1 Hz Steps
.01—1280 MHz Output @ 0 dBm .1 Hz Steps

4.1.1 Noise Floor of 11729B Phase Noise Test System



$$\mathscr{L}_{system} = 10 \log \left(N^2 \times 10^{\frac{\mathscr{L}_1}{10}} - 10^{\frac{\mathscr{L}_2}{10}} - 10^{\frac{\mathscr{L}_3}{10}} \right)$$

N = Multiplication number of the 640 MHz signal

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

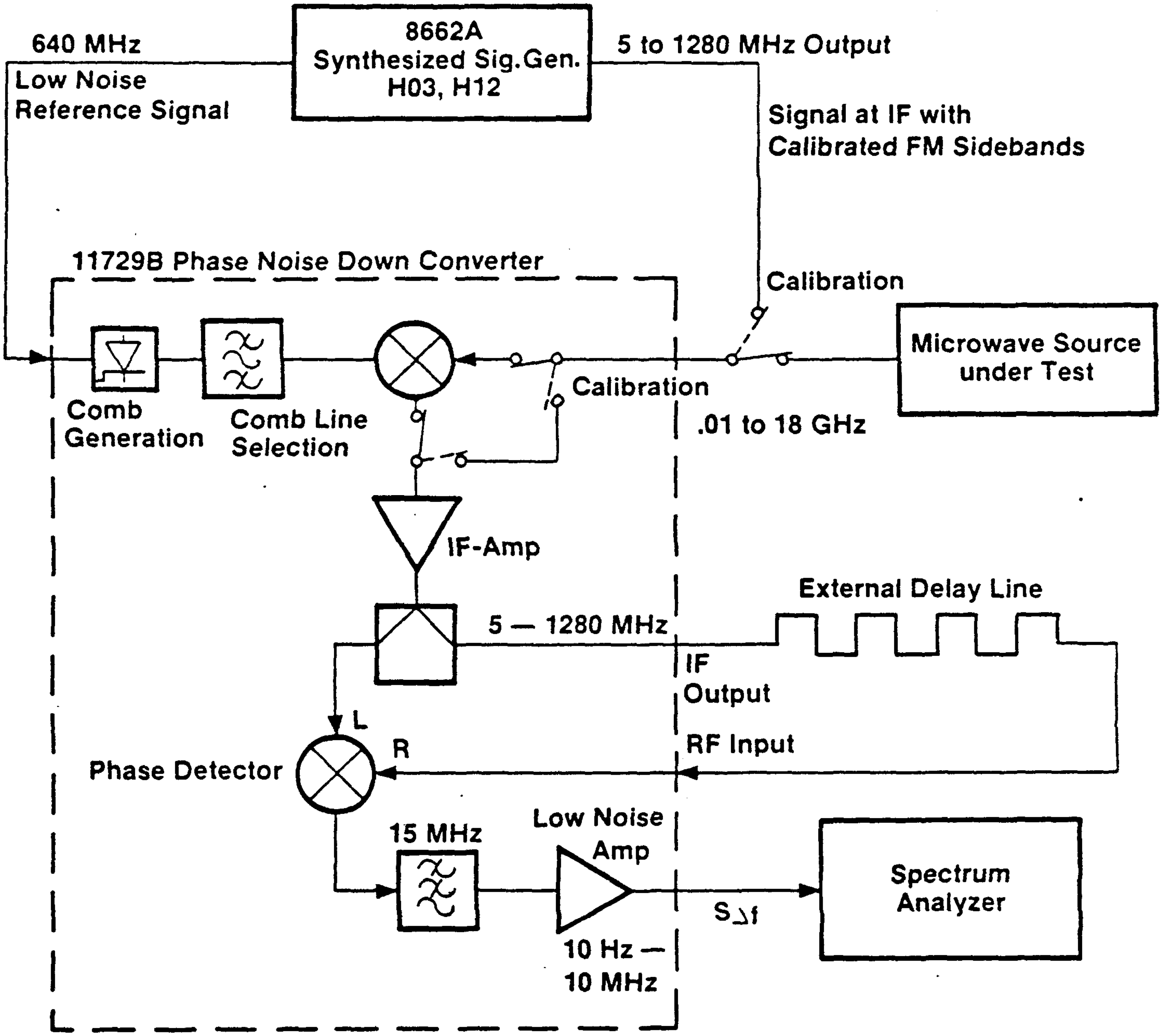
$\mathscr{L}_2$ = Absolute SSB Phase Noise of the 5 to 640 (1280) MHz tunable signal (dBc/Hz).

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

Offset from carrier	Typical 640 MHz low noise output	Typical 320-640 MHz tunable 8662A output	Typical *Residual* noise of 11729B @ 5 GHz	Typical *Residual* noise of 11729B @ 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	-147 dBc	-141 dBc
1 MHz	-159 dBc	-146 dBc	-149 dBc	-143 dBc

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

4.1.2 11729B Operating with Delay Line



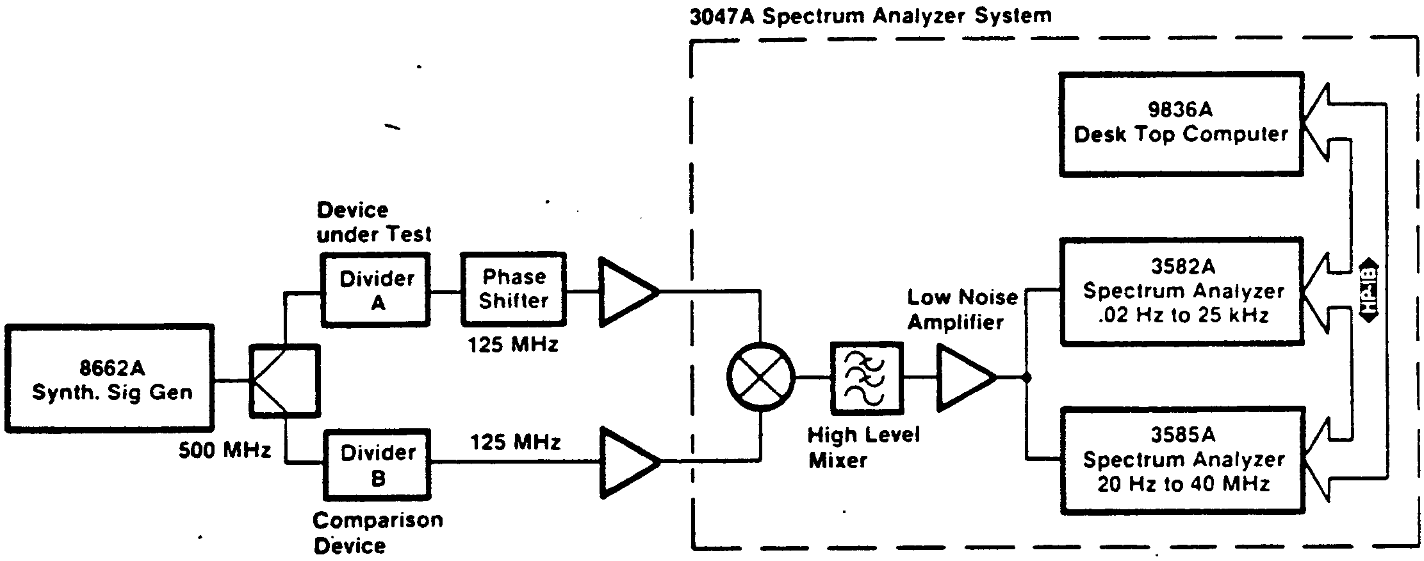
- **Low noise down conversion of microwave signal under test**
- **Delay line operating at IF (5—1500 MHz)**
- **Calibration signal provided by 8662A**
- **Phase quadrature indication.**

The 11729B may also be used in a delay line 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. The 8662A with FM sidebands and set to the IF signal frequency is a convenient calibration tool.

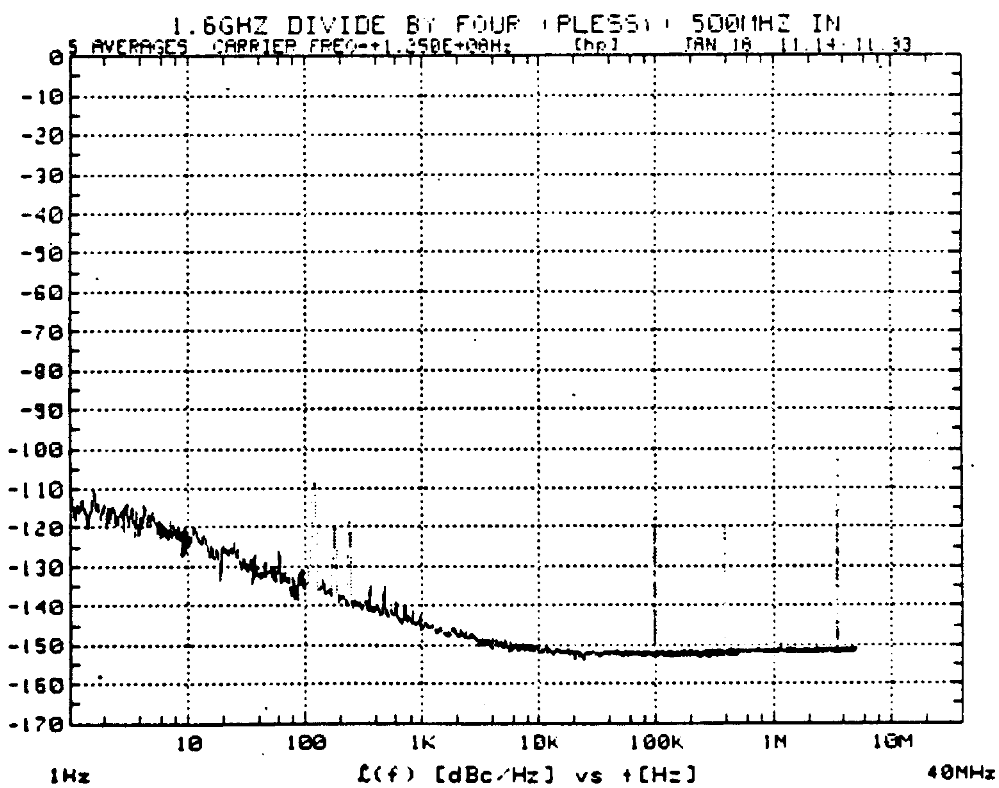
3047A Spectrum Analyzer System

Application Example:
Residual phase noise measurement on frequency divider

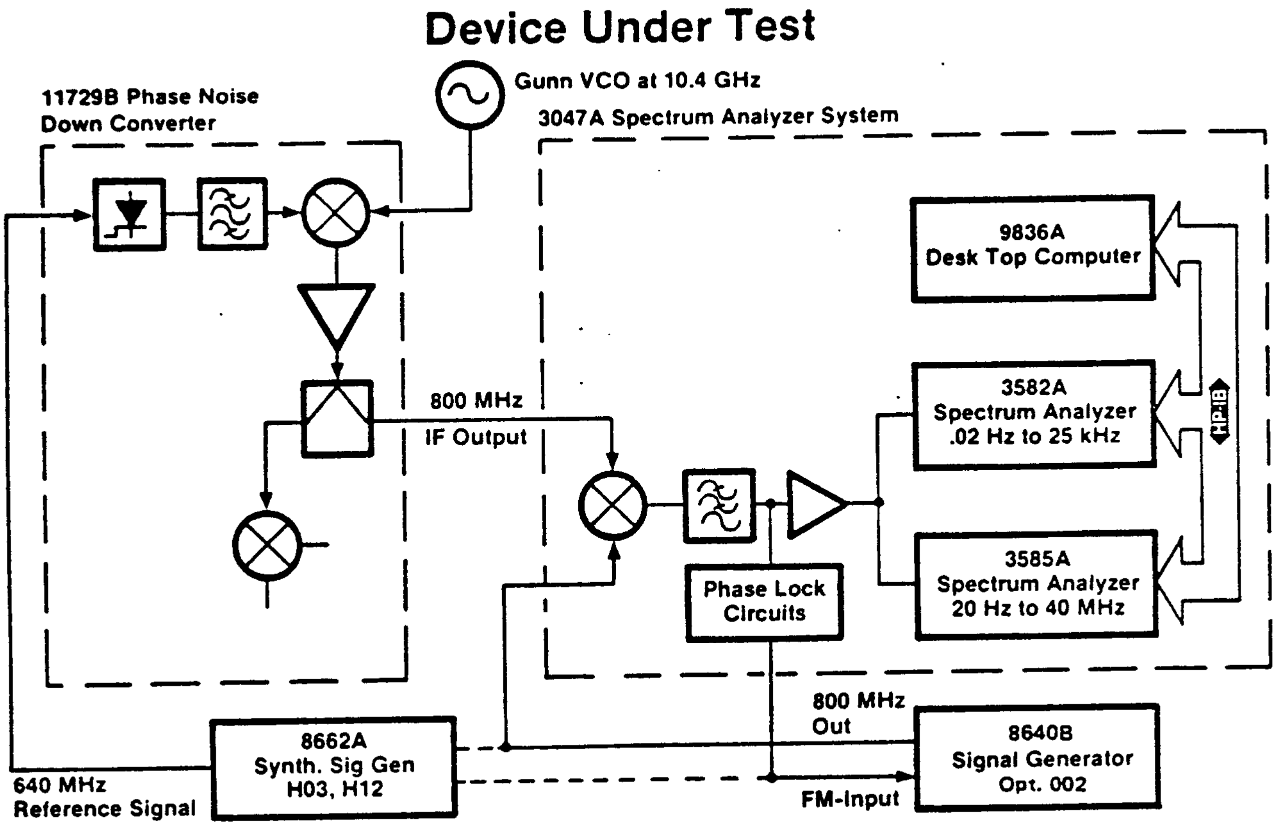


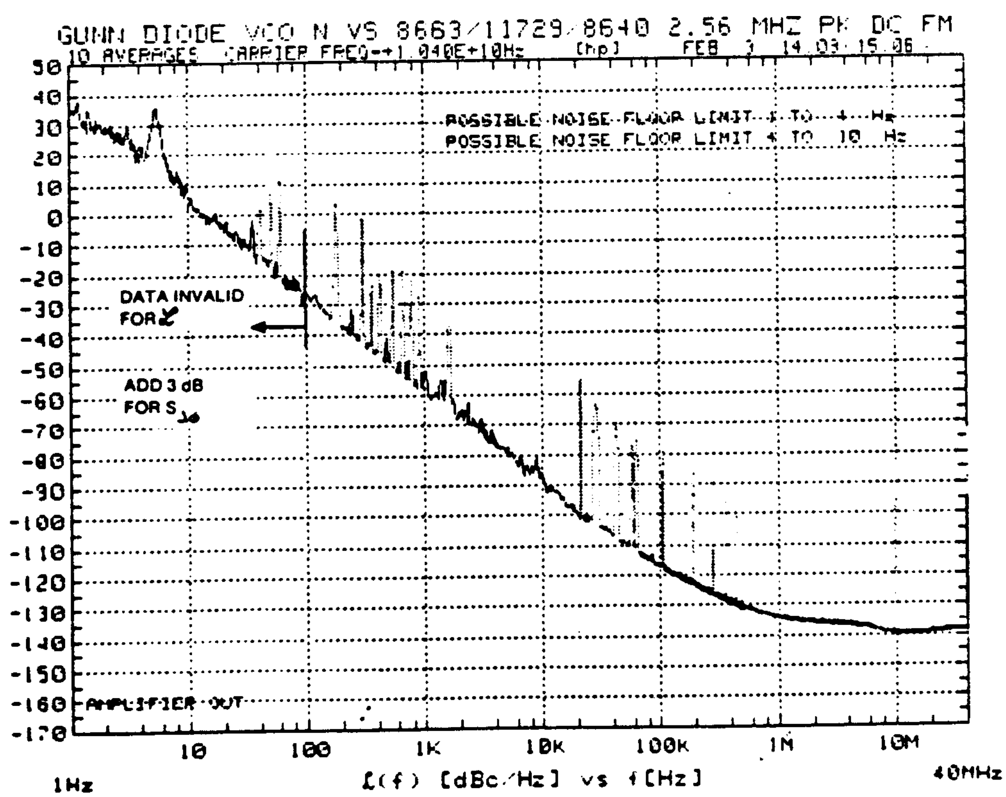
The 3047A Spectrum Analyzer System represents a high sensitivity phase detector system with controllable phase lock circuits. Extensive software automates the calibration and measurement procedures.

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.

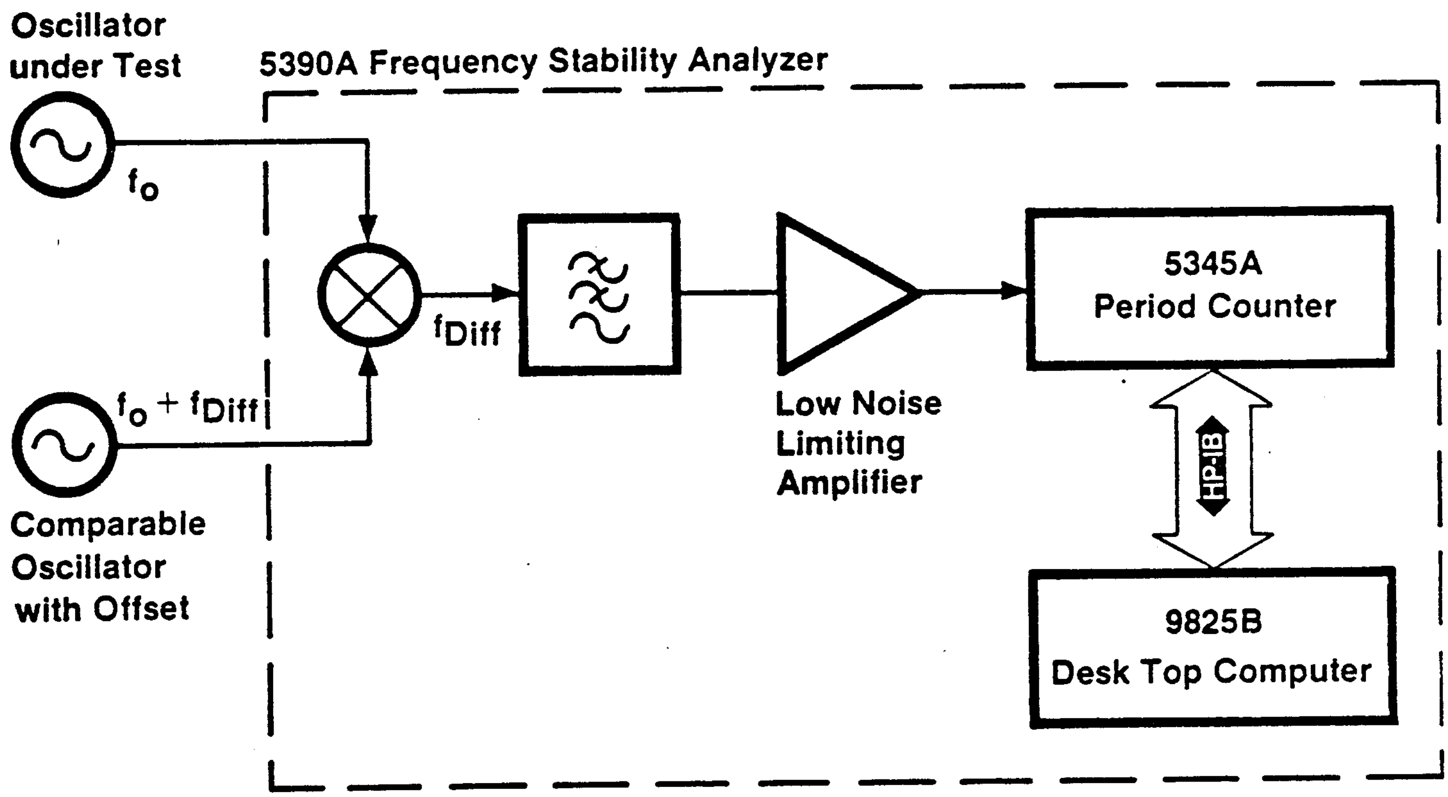


4.3 3047A/11729B Phase Noise Test System





5390A Frequency Stability Analyzer

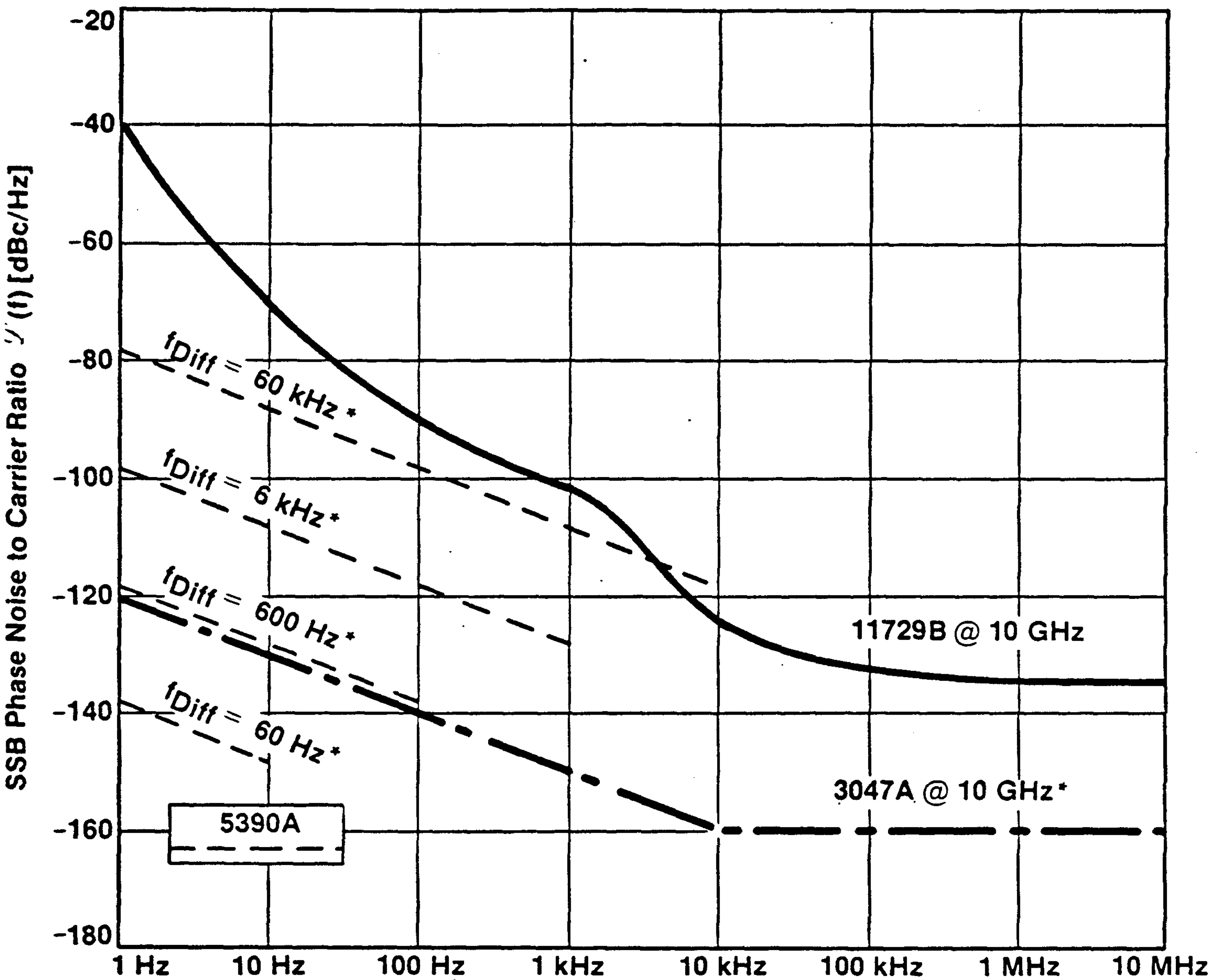


$$\mathscr{L}\text{system}(f) = -173 + 20 \log f_{Diff} - 10 \log f$$

The 5390A Frequency Stability Analyzer measures frequency noise in the time domain (Allan Variance by means of a period counter. To achieve sufficient resolution, heterodyning is a necessity and requires a second comparable source offset by the desired difference frequency.

The 9825B converts the time domain data into spectral densities. The 5390A has a very impressive noise floor close in, e.g., <−140 dBc/Hz at 10 Hz with a difference frequency of 60 Hz. Beyond 100 Hz, it quickly loses sensitivity compared to other methods.

4.5 Comparison of System Sensitivities



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

4.6 Phase Noise Measurement Systems Comparison

	11729B Phase Noise Test System	3047A Spectrum Analyzer System	3047A/11729B Phase Noise Test System	5390A Frequency Stability Analyzer
	8566A 8568A 3585A 3582A Spectrum Analyzers 11729B Phase Noise Down Converter 8662A Synthesized Sig Gen Controller	3585A Spectrum Analyzer 3582A Spectrum Analyzer 35601A S.A. Interface 9845B or 9836A	3585A Spectrum Analyzer 3582A Spectrum Analyzer 35601A S.A. Interface 11729B Phase Noise Down Converter 8662A Synthesized Sig Gen 9836A Controller	5345A Counter 5358A Meas. Stor. Plug-in 10830A Mixer, IF Amp. 9825B Controller
Principle of Operation	Source under Test Reference	Source Under Test Comparable Source	Source Under Test Reference	Source Under Test Counter Comparable Source w/offset
Frequency Range	.005 — 18 GHz	.005 — 18 GHz	.005 — 18 GHz	500 kHz — 18 GHz
Low Noise Microwave Reference Signal	yes	no	yes	no
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)	-140 dBc at 1 Hz* -126 dBc at 1 kHz
Noise Spectrum	<10 MHz	<40 MHz	<10 MHz (<40 MHz)	<10 kHz
Residual Noise (Measurement of Amplif., Dividers)	no	yes	yes	no
Delay Line Mode	yes	yes	yes	no
AM-Noise (Microwave)	yes	w/external detector	yes	no
Manual Operation	yes	no	yes	interactive applicatio program
Automatic Operation (HP-IB)	yes	yes	yes	yes
Software	planned	extensive	extensive for 3047 planned for 11729	yes

***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
$\Delta f(t)$	[Hz]	Instantaneous frequency fluctuation
K_ϕ	[volt/rad]	Phase detector constant
K_D	[volt/Hz]	Frequecy discriminator constant
$\mathscr{L}(f)$	[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
$S_{\Delta f}(f)$	[Hz^2/Hz]	Spectral density of frequency fluctuations
$S_{\Delta\phi}(f)$	[rad^2/Hz]	Spectral density of phase noise
t	[sec]	Time
v(t)	[volt]	Instantaneous voltage
$V_{B\ peak}$	[volt]	Peak voltage of sinusoidal beat signal
$\Delta\phi(t)$	[rad]	Instantaneous phase fluctuation
$\Delta\Theta(t)$	[rad]	Instantaneous phase perturbance
τ	[sec]	Delay time

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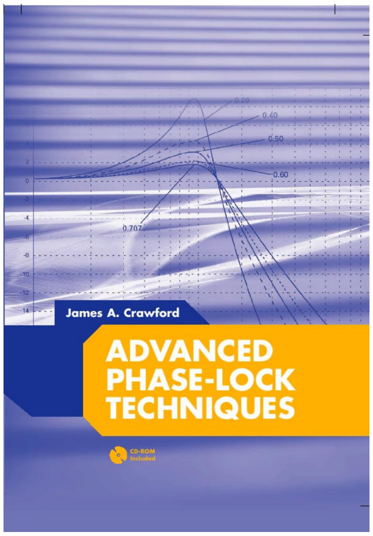

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