Application Note

NOISE XT

PN9000 Automated Phase Noise Measurement System Application Note #2

1 INTRODUCTION

This Application Note describes the operational characteristics of the Europtest/Noise XT PN-9000 Automated Phase Noise Measurement System. The PN-9000 is a modular system which is configurable to support a wide variety of applications, reference sources, frequency coverages, and signal analysis requirements. The system comprises a Noise Demodulation Test Set with a selection of front end frequency coverage modules, a Programmable Ultra-low noise Reference, a Processing Module which supports a digitizer, user software, optional display, and a SCPI/IEEE-488.2 interface.

The system uses a Intel based PC as a bench top processor and display module with the signal conditioning, measurement and characterization system in a bench top chassis which can be rack mounted.

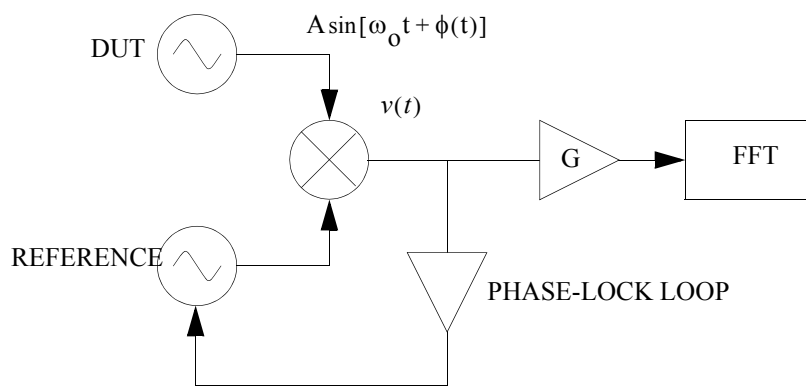## 2 THEORY OF OPERATION

Figure 1

Figure 1 shows a simplified block diagram of the PN-9000 system. The PN-9000 demodulates the phase fluctuations of the source under measurement by means of a quadrature locked low noise reference signal. A balanced mixer functions as a phase comparator and produces signal which comprises the time representation of the phase fluctuations of the source:

$$v(t) \;=\; K[\phi(t) - \phi_{ref}(t)]$$

1

where

K	demodulation factor of the mixer in volts/radian. This factor defines the 0 dBc reference of the phase noise spectrum.
$\phi(t)$	phase noise of the source to be measured.
$\phi_{ref}(t)$	phase noise of the reference signal

The reference is held in phase quadrature with respect to the source by means of the control voltage produced by the phase locked loop.

2.1 EFFECT OF THE PHASE LOCKED LOOP

The feedback action of the loop acts on the demodulated phase noise signal so that the frequency domain description of the effect is that of passing the output of the mixer through a high pass filter. If the impulse response of the filter is given by $h(t)$ and the corresponding transfer function is given by $H(f)$, then the corresponding time domain and frequency domain representations are:

TIME DOMAIN

$$v(t) = K[\phi(t) - \phi_{ref}(t)] \ h \otimes t)$$

where $\otimes$ is convolution

FREQUENCY DOMAIN:

$$S_v(f) = [K^2 S_\phi(f) + S_{\phi ref}(f)] \times H^2(f)$$

a result of the well known convolution theorems.

If the noise of the reference signal is much lower (perhaps 10 dB) at every frequency then that of the source to be measured, then the noise contribution to $S_v(f)$ can be neglected and:

$$S_v(f) = K^2 S_\phi(f) \times H^2(f)$$

To accurately characterize $S_v(f)$ in the region where $H(f)$ is not unity, it is necessary to compensate for the effect of the phase locked loop represented by $H(f)$. $H(f)$ can be either measured by injecting a white noise source, or calculated by measuring and setting the characterizing parameters of the loop. The PN-9000 sets and measures the necessary parameters to accurately correct for the effects of the loop.

2.2 CORRECTION FOR REFERENCE NOISE

Residual noise sources include the phase noise of the reference source and the noise of the measurement amplifiers. If the spectrum of the residual noise is known, then it is possible to make corrections for its influence on the measurement. This type of correction is generally applied only when the residual noise is no greater than the noise to be measured. For the table below, this corresponds to $\Delta = 0$ dB where:

Δ	Difference in dB between the residual noise and the noise of the source under test.
Correction	$10 \log (1 + 10^{-\Delta/10})$

Δ(dB)	0	1	2	3	4	5	6	7	8	9	10
Correction(dB)	-3	-2.54	-2.12	-1.76	-1.46	-1.19	-0.97	-0.79	-0.64	-0.51	-0.4

2.3   THE PN-9000 PHASE NOISE TEST SET

The demodulation process shown in Figure 1 is implemented in the PN-9000 as a set of cooperating modules. The signal processing process is carried out by digitizing the demodulated noise and computing the Noise Power Spectrum. Figure 2 shows the PN-9000 module partitioning for the demodulation process.
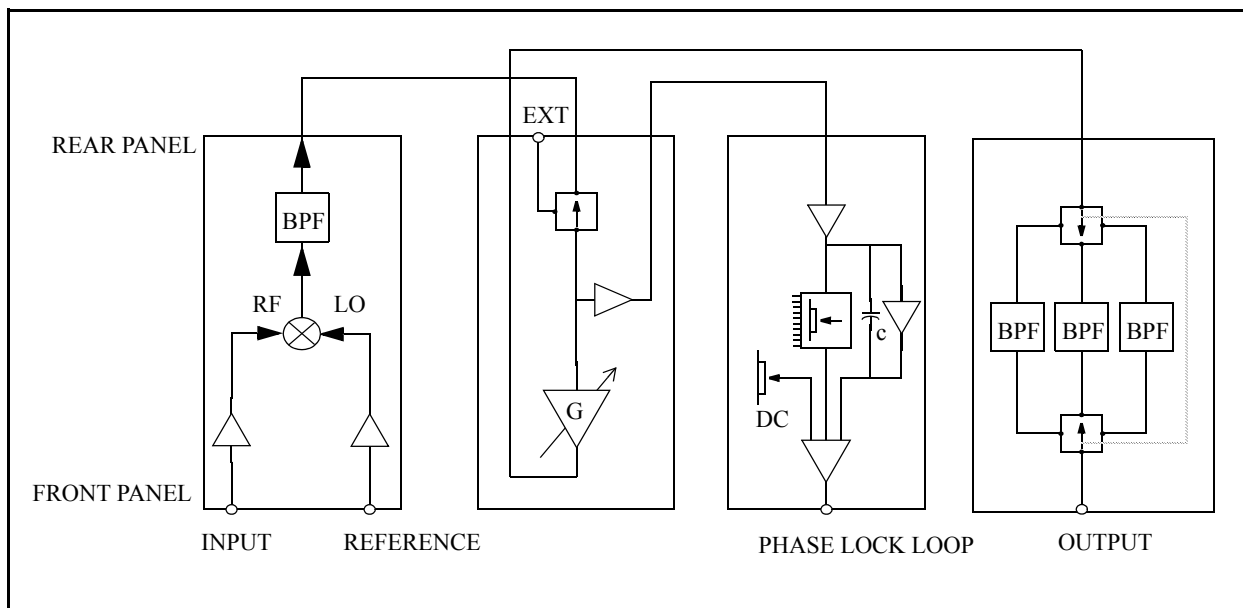

Figure 2

This measurement method can be thought of as having two components - Noise Demodulation and Spectral Analysis. The Noise Demodulation process requires initialization and calibration prior to actual measurements. This involves two aspects:

-   Calibration of the phase demodulator (V/rad)
-   Quadrature locking of the reference source

While these procedures are conceptually easy, practical implementation requires attention to aspects which may not be obvious but can cause immense difficulties in an operational sense. The PN-9000 is designed to allow for quick, accurate, and above all reliable automatic calibration procedures. Calibration of the phase demodulator and reference locking is achieved even with imperfect (non-sinusoidal)

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waveforms. The PN-9000 calibration system is able to accommodate a wide range of signal parameters automatically without needing operator intervention or compromising system accuracy. The following features support these robust capabilities:

- The system automatically measures the parameters it needs to configure itself, and automatically selects optimum configurations.
- The test system can adjust the level of the signal under test to accommodate both high and low level signals (-20 dBm).
- The phase locked loop control allows for setting of all parameters and the selection of an optimum set of parameters automatically.
- The phase locked loop maintains signal quadrature even when the source to be measured has substantial frequency drift.
- The reference source can be any type of programmable synthesizer. The PN-9000 has built in IEEE-488 command drivers to accommodate a wide variety of signal sources, standard and additional ones can be added if needed.

3   CALIBRATION

The calibration process involves measuring the demodulation factor of the phase comparator. This factor (K) depends on the amplitude of the signals provided by the sources. This measurement is made with the loop unlocked, and a beat frequency signal obtained at the output of the mixer. The PN-9000 requires that the frequency of the beat signal be in the range of 1 Hz to 100 kHz. The demodulation factor K is then given by the zero-crossing slope of the signal, and the zero level corresponds to the level obtained under quadrature lock. To measure the zero beat slope, the following three methods are generally used:

3.1   PEAK AMPLITUDE MEASUREMENT

For a pure sinusoidal beat signal having frequency F_b and amplitude A_b, the demodulation factor K is calculated as follows:

$$v(t) \; = \; A \sin[F_b t + \phi t]$$

Under phase lock:

$$F_b \; = \; v(t)$$
$$v(t) \; = \; A \sin(\phi(t))$$
$$\phi(t) << (1 \sin \phi(t) \approx \phi(t))$$
$$v(t) \; = \; A \sin(\phi(t))$$
$$K \; = \; A(\text{Volt/radian})$$

Calibration error introduced by this method is caused by harmonics of the beat signal.

3.2   ZERO CROSSING SLOPE

This method yields the most accuracy because the slope is actually measured in volts per radian. The PN-9000 analyzes the beat's temporal characteristics, calculates a polynomial which represents the signal

around the zero crossing $\left(\pm\frac{\pi}{8}\right)$, and then calculates the derivative of the polynomial which produces the calibration factor K. Figure 3 shows how the measurement is conducted.
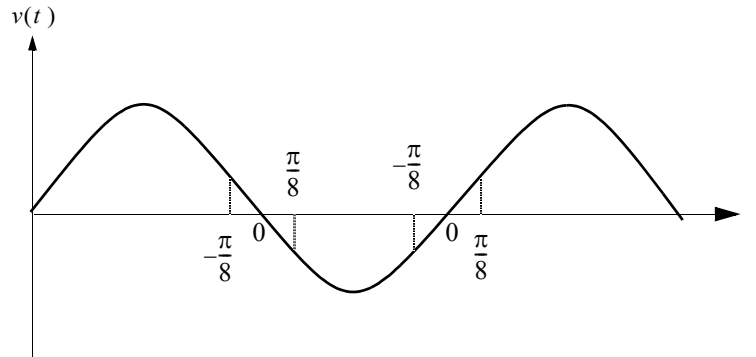

Figure 3

$$[v(t)]\left(\pm\frac{\pi}{8}\right) = ax^2 + bx + c$$

$$K = 2ax_o + b \quad (slope\ in\ x_o)$$

This method produces a valid measurement even if the beat signal is highly distorted. The PN-9000 uses this method of calibration for beat frequencies in the range of 1 Hz to 10 kHz. In the next section we show how the polarity of slope is selected.

3.3   SPECTRAL ANALYSIS FOR CALIBRATION

If the beat signal does not meet the requirements for the Zero Crossing Slope method, a spectral analysis of the beat signal must be conducted to determine the amplitude of the fundamental signal and its harmonic components. To reproduce the temporal characteristics of the signal, its harmonic content must be known. However it is insufficient to know just the level of the harmonics, the relative phase of the harmonics must also be known. Spectral analysis does not permit determination of this phase information, so this method pertains to two special cases:

- When the harmonics are in-phase, the temporal characteristic is "rectangular"
- When harmonics are in opposite phase, the temporal characteristic is "triangular"

These temporal characteristics are easily recognized and the PN-9000 can select either of these modes.

If the characteristic of the signal is uncertain, the exact calibration value lies between the two modes. The PN-9000 computes the mean value between the "rectangular" and "triangular" as its default method.

Test systems that use this calibration method but which do not distinguish between these cases introduce substantial calibration errors. The following example shows the errors which are introduced by ignoring the relative phase.
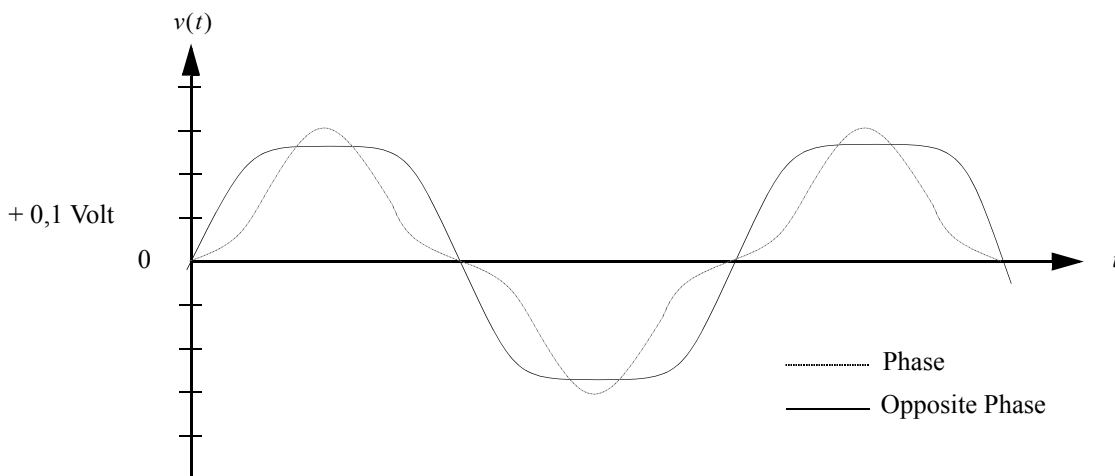

Figure 4

Suppose the level at the fundamental frequency is 0.3 v peak, and the 3rd harmonic is -20 dB, and the 5th harmonic is - 30 dB. Figure 4 shows the resultant waveforms for the two cases, in phase and opposite phase

The difference in slope at the zero crossings between the two cases is:

K = 0.434 volt/radian (Harmonics in phase)

K = 0.164 volt/radian (Harmonics in opposite phase)

The ratio of these calibration factors corresponds to a calibration uncertainty of 8.45 dB.

The following table shows the potential calibration errors due to 3rd order harmonics.

H3	ERROR
-30 dB	1.7 dB
-40 dB	0.6 dB
-50 dB	0.2 dB

3.4 ASYMMETRICAL SIGNALS

When the beat signal is relatively pure the difference between the slope of the positive going signal and the slope of the negative going signal is insignificant. For many actual signals this is not the case, and the beat signal obtained by mixing the source under test and the reference source is significantly asymmetric (see figure 5). Depending on the sense of the phase locked loop, the system aligns itself with respect to either the positive going signal or the negative going signal. The difference in slope therefore manifests itself in calibration errors in those systems which do not distinguish between the values K+ and K-.
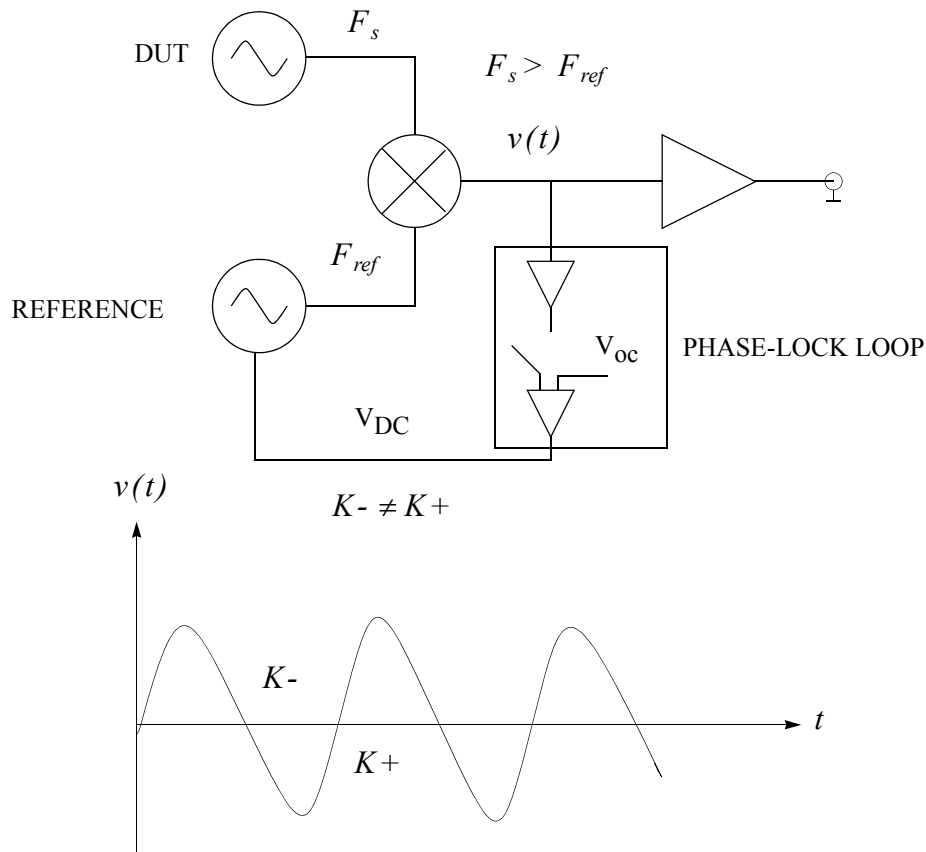

Figure 5

For example, if the fundamental signal has an amplitude of 0.3 volt peak and the 2nd harmonic is -20 dB, K+ = 0.359 volt/radian and K- = 0.239 volt/radian. The potential error is 3.5 dB depending on the sense of the loop. In order to avoid such errors, either the beat signal must be sinusoidal or the two possible

slopes must be measured individually and the sense of the loop determined. The PN-9000 measures K+ and K- individually and can choose the correct slope based on the sense of the phase locked loop.

The measurement process undertaken by the PN-9000 is as follows:

The loop is unlocked and the sense of the control with respect to the reference source is determined. The control is positive if the frequency is increased by increasing the control voltage. The control is negative if the frequency is decreased by increasing the control voltage.

The frequency of the reference source is set at a value lower than that of the source under test.

The correct slope is selected according to the sense of the reference control.

The sense of the phase locked loop is thus determined and the beat signal need not be purely sinusoidal as the asymmetry of the beat signal no longer poses a problem.

4 PHASE LOCKING THE REFERENCE

The PN-9000 phase locks a reference source by constructing a phase locked loop with proportional and integral gain (2nd order). All necessary parameters are programmable and the system automatically verifies that parameters are compatible. The system compensates for the obtained transfer function so that the measurement of the source under test is not degraded. Figure 6 shows how phase lock is obtained.
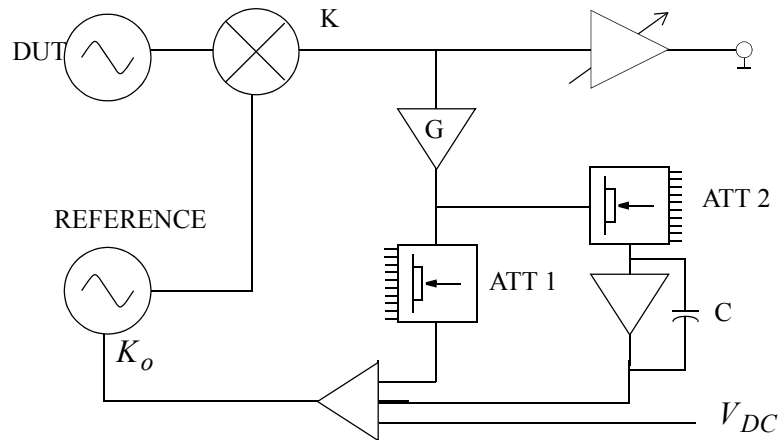

Figure 6

A detailed analysis of the transfer function based on parameter values is given in Comstron Application Note 93-0001. The parameters that contribute to the transfer function are as follows:

K	demodulation constant of the phase comparator
G	gain of the loop amplifier
ATT1	programmable attenuator which controls proportional gain
ATT2	programmable attenuator which controls integral gain
K_0	slope of the control voltage for the reference source
F_m	bandwidth of the voltage control for the reference source

The constants K and K_o are measured automatically, the gain G is known, and the system programs ATT1 and ATT2 to adjust the phase lock bandwidth to a chosen value.

4.1   TRANSFER FUNCTION

The following example shows how the system locks a reference source and compensates for the resultant transfer function.

> Assume a source to be tested has a power output level of +3dBm
>
> The measured demodulation constant K = 0.2 volt/radian.
>
> The loop amplifier gain G = 20 dB
>
> The control slope of the reference source K_o = 500 Hz/volt
>
> The frequency control bandwidth F_m = 250 Hz

The system is requested to produce a phase lock bandwidth of 250 Hz. The system calculates and sets the attenuator values ATT1 = 0.25, ATT2 = 0.157 to produce a phase lock with the transfer function shown in Figure 7.
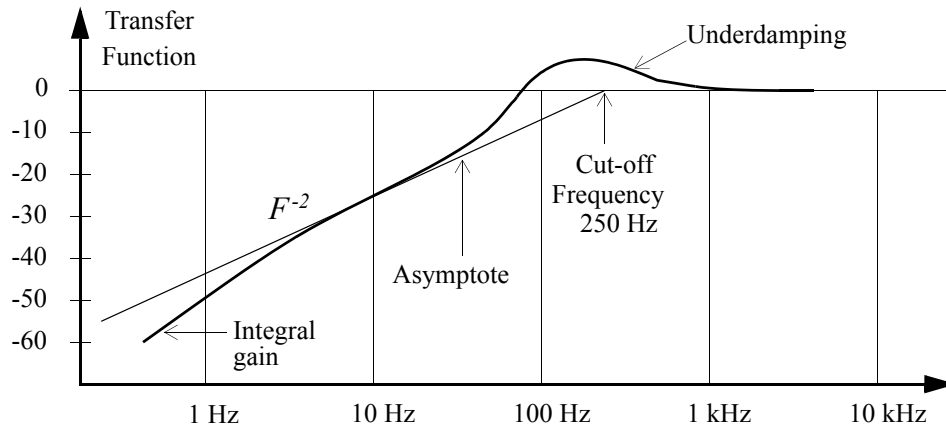

Figure 7

ATT1 is set to obtain a noise bandwidth F_b = 250 Hz

ATT2 is set to achieve a damping factor $\zeta > 5$.

$$\zeta = \frac{1}{2\left(\dfrac{F_b}{F_{rc}}\right)} \qquad \text{where } F_{rc} = 16\text{Hz} \times \text{ATT2}$$

In this example the underdamping at 250 Hz is produced by the phase shift of the frequency control input. The system corrects for this result since the frequency control bandwidth is known.

4.2   TRACKING BANDWIDTH

For this example the reference source has a frequency control slope of 500 Hz/volt. If we assume it voltage control range to be ± 5 volts, the resulting tracking bandwidth is ± 2500 Hz. Thus if the frequency

of the source to be tested remains within 2500 Hz of the center frequency of the reference source, the system maintains phase lock.

4.3   ACQUISITION BANDWIDTH

For this example, in order to achieve lock the frequency deviation must not be greater than 250 Hz.

4.4   PHASE LOCK BANDWIDTH

The phase lock bandwidth depends on the frequency stability of the sources. The frequency stability of unknown sources can be easily estimated using a frequency counter - a few consecutive frequency measurements provides an estimate of frequency fluctuations. In order to ensure loop stability the system imposes a minimum bandwidth value of two percent of the control slope of the reference source.

For the example the minimum bandwidth would be $K_o/50 = 500/50 = 10$ Hz.

5   DOUBLE DOWN CONVERSION

It is often more convenient to down-convert a microwave frequency source under test in two steps rather than directly lock a microwave reference source. The particular method used depends on reference sources available. Figure 8 shows the generic double down conversion scheme.
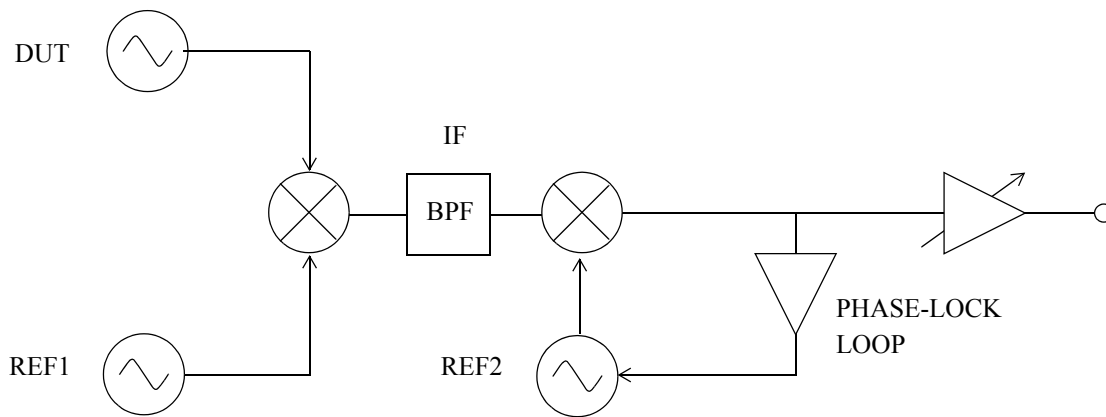

Figure 8

The second reference source, REF2 is locked on the down-converted signal IF. Three different configurations are possible depending on whether the reference sources REF1 and REF2 are wide band synthesizers or fixed frequency oscillators. The calibration and phase lock processes are the same as for the normal single conversion method.

The advantages of this method for microwave measurements include:

- It is not necessary to have a microwave reference source with exactly the same frequency as the source to be tested.
- The microwave source need not have lock capability
- The frequencies of the microwave sources are not the same so unintended synchronization of the two sources is unlikely.
- Phase lock can easily be achieved regardless of the microwave frequency to be tested.
- The reference sources of the PN-9000 can be used on the LO channel of the mixer. Since the signal to be tested is on the RF channel its level can be as low as -20 dBm.
- For discrete frequency coverage the implementation is compact, inexpensive, and reliable.

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5.1   REF1 SYNTHESIZER; REF2 OSCILLATOR

This double down conversion configuration (Figure 9) allows for conversion of all frequencies down to a fixed frequency IF provided that the synthesizer has adequate frequency range and resolution.
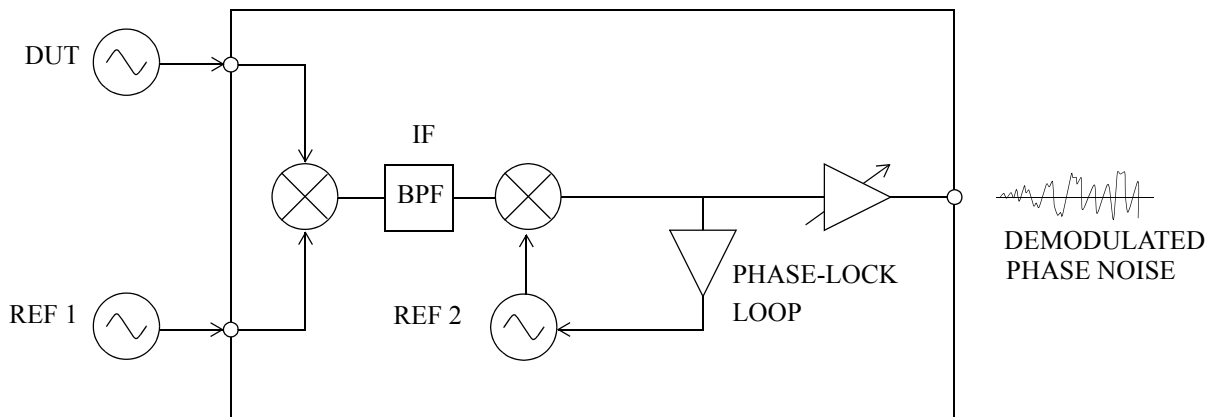

Figure 9

For example, REF1 is a microwave synthesizer 10 MHz to 18 GHz with resolution of 1 HZ. REF2 can then be a voltage controlled crystal reference oscillator (VCXO) with fixed frequency of 10 MHZ, 100 MHZ, etc.

Advantages:

- The synthesized down-converter need not be locked
- Phase lock is always the same regardless of the frequency to be measured
- REF2 is included in the PN-9000
- Phase lock is incorporated in the PN-9000

Disadvantages:

- The phase noise of the microwave synthesizer reference limits the residual noise of the measurements.

5.2   REF1 DISCRETE FREQUENCIES; REF2 SYNTHESIZER

This conversion configuration (Figure 10) uses a first reference source comprising a limited number of discrete frequencies. Every signal to be tested is down-converted into a frequency range which can be covered by the frequency synthesizer associated with REF2.
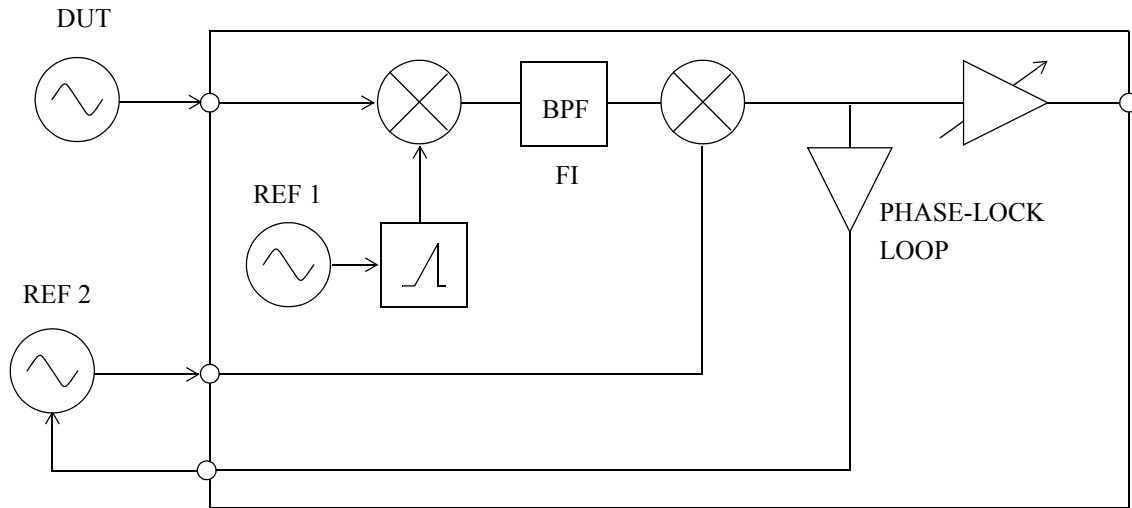

Figure 10

For example, REF1 is a frequency comb in the range of 1 to 18 GHz with 400 MHz spacing. REF 2 is an RF synthesizer covering a frequency range up to 400 MHz. REF 2 is locked by the PN-9000 through its FM DC input.

Advantages:

- REF1 can be extremely low noise
- The RF synthesizer can be a standard synthesizer provided that it have the required frequency range and resolution, a DC FM input, and a phase noise level compatible with the measurement requirements.
- The synthesizer can follow considerable frequency drift in the source to be measured.

5.3   REF1 AND REF2 PROVIDE DISCRETE FREQUENCIES

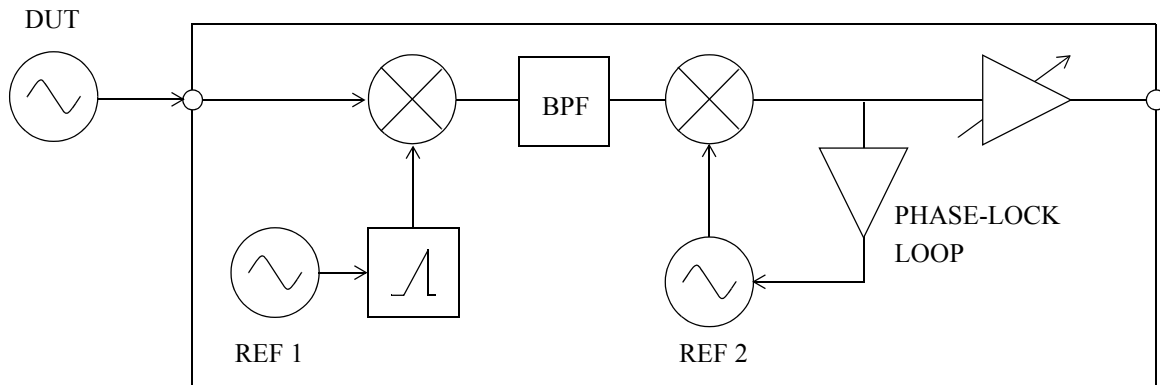This configuration (Figure 11) uses two references set at discrete frequencies.

Figure 11

For example, REF1 could be 1 to 18 GHZ with 400 MHz spacing and REF2 100 MHz to 400 MHz with 100 MHz spacing. This would allow measurements to be made over a wide frequency range (e.g. 1 to 18 GHz) but at a limited number of frequencies (e.g. 100 MHz steps).

Advantages:

-   Both references can be extremely low noise
-   The cost of the references is much lower than using a synthesizer for either reference

Disadvantages:

-   The measurements can be performed only at a limited number of frequencies.

6   EXTERNAL MIXER

For frequencies that are outside the range of the PN-9000 (e.g. higher than 18 GHz), and external mixer can be used for conversion. The mixer can either be used to shift to DC or to an IF frequency, and therefore its output can be connected either to the external input of the amplifier section as shown in Figure 2, or to the RF inputs of the PN-9000 mixer section.

6.1   DIRECT VOLTAGE OUTPUT

The reference is set at the same frequency as that of the signal to be tested, in essence a single down conversion method, and the IF output of the mixer connected to the external input of the amplifier section (Figure 12).
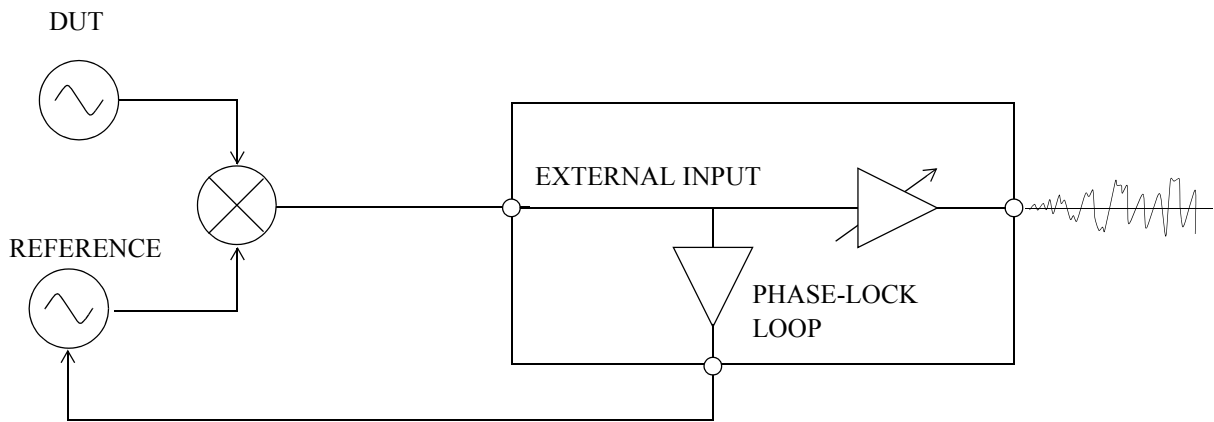

Figure 12

The phase lock section of the PN-9000 then controls the reference source REF1

6.2   IF OUTPUT

The frequency of the reference source is offset, and multiple down conversion is employed. Depending upon the IF frequency provided, either double conversion or single conversion can be applied to the IF signal. Figure 13 shows an example where the IF can be as high as 18 GHz so that double conversion must be used.
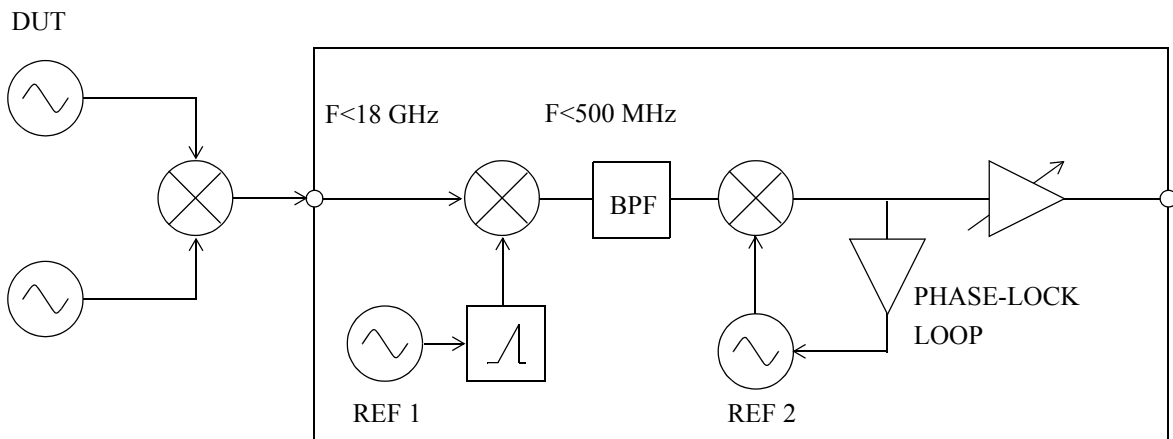

Figure 13

7   DELAY LINE METHOD

For sources which exhibit only mid-quality phase noise, the delay line demodulation method can be used. The source to be tested is connected to the LO channel of the mixer through a phase shifter which allows adjustment to phase quadrature with respect to the RF channel. The same source is connected to the RF channel through a delay line. This method is shown in Figure 14.
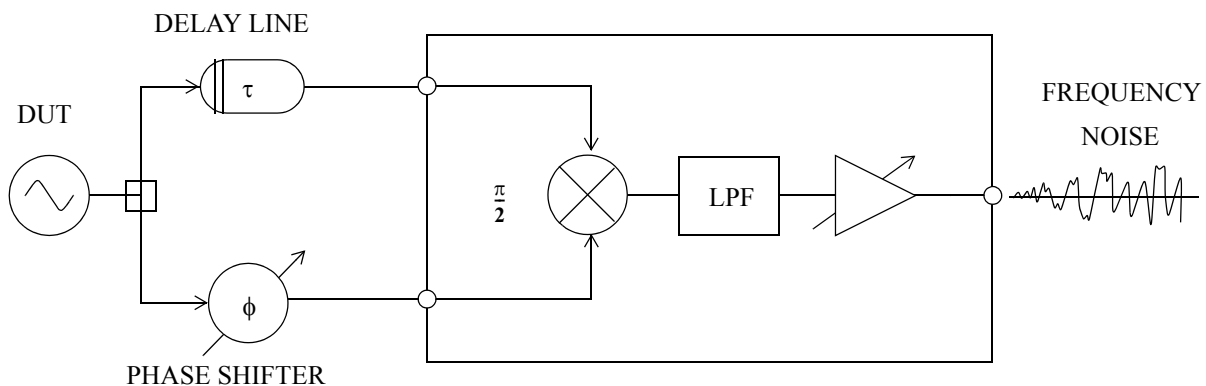

Figure 14.

The output level of the mixer is proportional to the frequency fluctuations of the source provided that the phase of the two channels are maintained at a relationship of $\frac{\pi}{2}$.

$$v(t) = K 2\pi\tau \left(\frac{\sin(\pi f \tau)}{(\pi f \tau)} \right) \times \Delta f(t)$$

where:

K	the demodulation constant of the mixer
τ	is the delay of the line
$\Delta f(t)$	is the source frequency noise

$$\text{for } f < \frac{1}{10\tau} \qquad \frac{\sin(\pi f \tau)}{(\pi f \tau)} \approx 1$$

$$v(t) = K 2\pi\tau \times \Delta f(t)$$

$$v(t) = K_f \times \Delta f(t)$$

The sensitivity of this demodulation method is proportional to the delay τ. The bandwidth ($1/10\tau$) is inversely proportional to τ. It is often difficult to balance the two and find an acceptable compromise. The spectral analysis must also transform the frequency noise measured by this method into the equivalent spectral density of phase noise. The advantages and disadvantages of this method are further described in Application Note 93-0001.

8   COMPONENTS AND SYSTEM NOISE

Measuring noise added by components such as frequency dividers and mixers or systems such as amplifiers and down-converters is becoming increasingly important because of the considerable noise level introduced by them compared to the noise produced by frequency sources.

Microwave systems in particular require these measurements since GaAs technology produces much higher flicker noise than Si technology does. For GaAs a typical flicker cutoff frequency with respect to white noise is 100 MHz whereas the equivalent in Si technology is less than 10 kHz. Thus it is necessary to evaluate their effect on low noise signals which are acted upon by these components or systems.

Two measurement configurations are needed depending on whether the output frequency of the device under test is the same as the input frequency.

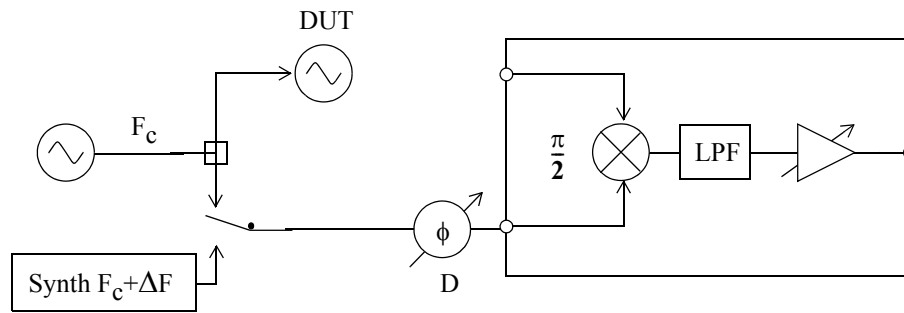## 8.1   OUTPUT F = INPUT F

Figure 15

Figure 15 shows the generic test setup. For this case the DUT is replaced by a direct line from the signal source. The phase shifter D is needed to adjust quadrature for the demodulation mixer.

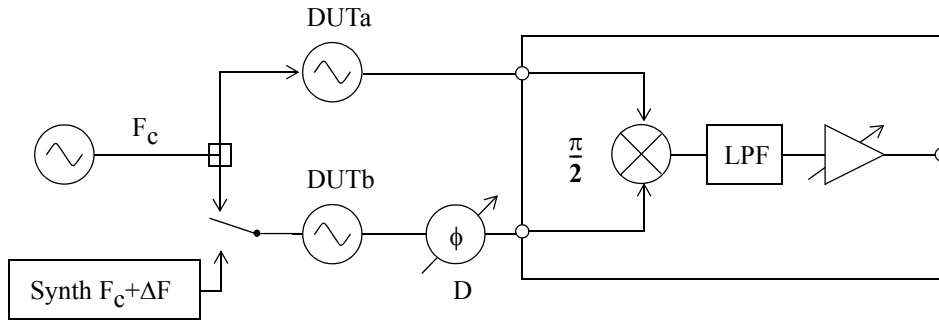## 8.2 OUTPUT F ≠ INPUT F

Figure 16

Figure 16 illustrates the second test situation. In this case two DUTs with identical characteristics must be inserted in the two channels of the mixer (DUTa, DUTb). The measured noise which results from this method is the sum of the noise of the two DUTs.

8.3 CALIBRATION

The first method uses a reference source with a different frequency to obtain a beat signal at the output of the mixer and thus use the standard calibration method.

The second method requires a strictly sinusoidal signal (harmonics < 30 dBc) for the RF signal. If this requirement is met, the PN-9000 can be calibrated without using a second reference source. The unit can be calibrated on the DC level produced by the mixer when both of its channels are in phase by adjusting the phase shifter.

8.4 RESIDUAL NOISE

The residual noise of the source must be accounted for since it is not completely compensated for at the output of the mixer. A constant factor of attenuation is obtained and the residual noise of the measurement is equal to the phase noise of the source from which the constant factor is deducted. If the differential delay between the two channels is introduced by the DUT, the factor is calculated as follows:

$$A = 20\log 2\sin(\pi f\tau)$$

where

f	offset from the center frequency
τ	differential delay

For example if τ = 1 microsecond, f = 10 kHz then A = 24 dB

The differential delay can either be measured, or the attenuation factor calculated by modulating the source frequency and measuring the rejection.

9   AMPLITUDE NOISE

A reference signal is not needed for this measurement. The signal to be tested directly drives the PN-9000 which is equipped with a wide-band amplitude detector (10 MHz to 18 GHz). Figure 17 shows the configuration.
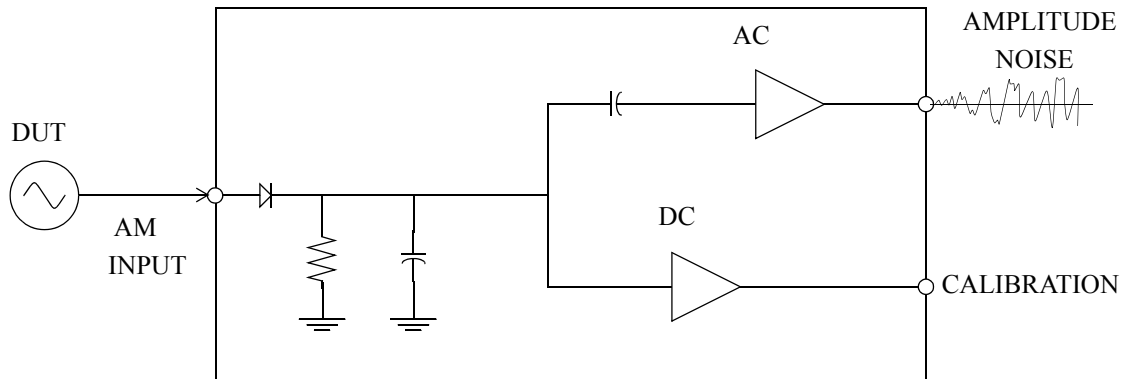
Figure 17

Calibration is conducted by modulating the amplitude of the source to be tested. The demodulation slope of the detector is then calculated.

The PN-9000 aids in this sometimes difficult procedure by allowing calibration on CW levels. The PN-9000 calculates the demodulation slope for this continuous detection.