APPLICATION NOTE

Residual Phase Noise

Theory and Advanced Measurement Techniques using the N5511A Phase Noise Test System

KEYSIGHT

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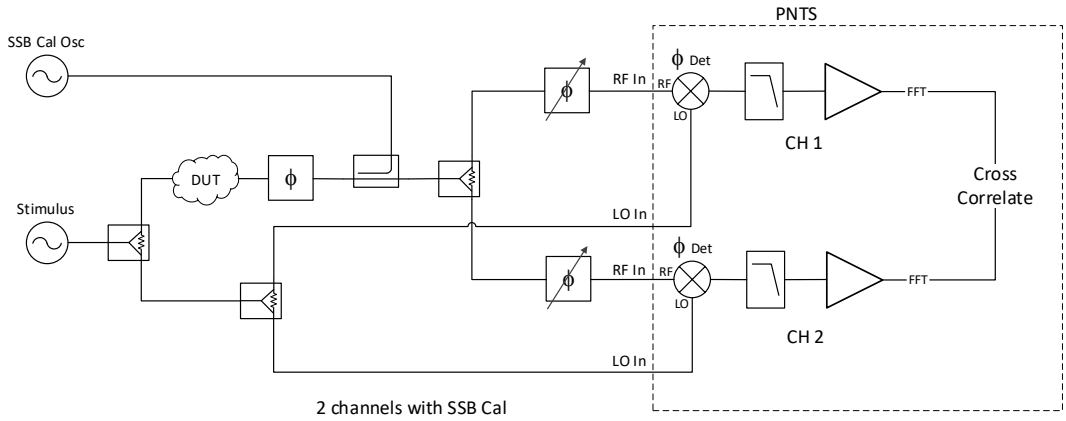
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Introduction



The challenges of Residual Phase Noise measurement are addressed by the N5511A Phase Noise Test System that achieves the industry's lowest noise floor using Cross Correlation of two RF/uW Analog Phase Detector outputs. Phase noise introduced by an Amplifier is termed additive or residual phase noise, while that of a Signal Source is termed absolute phase noise. Residual phase noise of an amplifier is typically much lower than the absolute phase noise of a Stimulus signal source driving it even for a high-quality phase noise measurement system like the N5511A. The challenge is then to be able to separate out the residual phase noise of the amplifier



(DUT) from the rest of the system contribution including the Stimulus that may have significantly higher absolute phase noise and AM noise. Even the DUT's own residual AM noise could be higher than its residual phase noise. To meet this challenge, two advanced techniques are incorporated as described in this application note, one for suppressing the Stimulus phase noise using a method of accurate delay matching, and a second one for significantly improving the AM rejection of the Phase detector.

The advanced noise suppression techniques are complemented with a Dual-Channel Cross-Correlation architecture with quadrature Phase Detection in the N5511A Residual Phase Noise test set up as shown above for a 2-port DUT that is non frequency-translating. The set up includes two delay line based adjustable phase shifters, three hybrid power splitters, and a coupler, all of which are integrated into the Keysight PNMI (Phase Noise Measurement Interface) Options K01 (1.4 GHz) and K26 (26 GHz). The Stimulus is generally provided by a laboratory-grade signal generator with excellent (low) phase noise characteristics such as the E8257D Analog PSG or similar. The Stimulus drives the DUT whose output is then phase adjusted before feeding the RF port of the Phase Detector in each channel. The Stimulus also directly feeds the LO port of the Phase detector in each channel through a hybrid power splitter. The Phase Shifters in the RF paths are adjusted such that the RF and LO signals are in quadrature i.e. 90° in relative phase. Under such conditions it can be shown that the phase noise of the Stimulus is cancelled at the Phase Detector's down converted output, while the residual phase noise contributed by the DUT appears at that output. The noise contributions from the Phase Detectors and the down converted baseband signal path components are greatly suppressed by taking the spectral cross correlation across the two

channels. One of the challenge is to ensure that the relative delay between the RF and LO paths caused by the DUT, adjustable phase shifters, and the connecting RF cables, is minimized or compensated, otherwise there loss of correlation of the Stimulus phase noise at the RF and LO inputs to the Phase detector that leads to its leakage at the output. The leaked Stimulus phase noise at the Phase Detector output increases with the offset frequency and has the undesired possibility of exceeding the level of the DUT residual phase noise beyond a certain offset frequency. The advanced noise suppression techniques include an indirect method of measurement of this uncompensated relative delay in-situ with the residual phase noise measurement and is covered in this application note.

Overview of This Application Note

The theory and techniques of Residual Phase Noise measurement are described, beginning with an introduction to Absolute and Residual Phase Noise, and an analysis of the Phase Detector operation including impact of AM noise. Detailed analysis of the Stimulus phase noise suppression is provided along with methods to determine the uncompensated delay that causes it. The Residual Phase Noise measurement highlighting the two methods for the Stimulus noise suppression are shown with N5511A PNTS + PNMI K26 set up for a CMD274 low noise GaAs amplifier. Finally, Keysight ADS Harmonic Balance based simulation model is provided to check out the impact of uncompensated delay and phase error on the Stimulus phase noise leakage and AM suppression. It shows full agreement with the theoretical equation that is derived and plotted using Matlab scripts all of which are included in the Appendix.

Phase Noise and Small Angle Limit

A noisy carrier with AM noise $A(t)$ and phase noise $\emptyset(t)$ is shown as:

$$v_n(t) = A(t) \cdot \cos\left[w_c t + \emptyset(t)\right] \quad (1)$$

Above can be depicted in complex notation as

$$v_n(t) = real\left[\, A(t) e^{jw_c t} e^{j\emptyset(t)}\right] \quad (2)$$

where

$A(t)e^{jw_c t}$ is the complex carrier signal at frequency w_c, with AM noise $A(t)$, and $e^{j\emptyset(t)}$ is the complex phase noise "phasor" describing phase disturbance or phase modulation of the complex carrier.

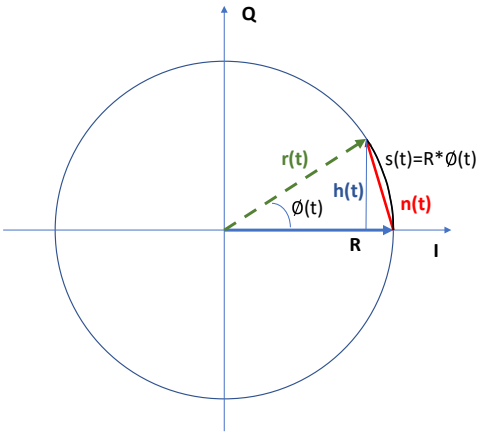


Figure 1a. Phase Noise can be quantified in 3 different ways: n(t)/R, h(t)/R, and s(t)/R = $\emptyset(t)$

IEEE defines the one-sided Phase Noise L(f) using the power spectral density of $\emptyset(t)$

$$\mathbf{L(f)} \equiv \frac{1}{2} S_{\emptyset}(f)$$

Here $\mathbf{f}$ is the offset frequency

Under small angle condition i.e., $|\emptyset(t)| << \pi/2$, L(f) is same as the half power spectral density of n(t)/R and h(t)/R.

Phase Noise of a signal source is its phase modulation due to noise. It is mathematically defined as a time-domain multiplicative noise i.e. a noise "Phasor" $e^{j\emptyset(t)}$ in time domain that is multiplied with the original signal $A(t)e^{jw_c t}$ of frequency w_c radians/s. $A(t)$ is the AM noise that is usually suppressed to a large extent by phase noise measurement systems, and this will be covered under the Phase Detector theory. $A(t) = R[1 + m(t)]$ where m(t) << 1 is the AM noise and R is the signal amplitude.

Physically phase noise could arise due to time-domain multiplication as in a mixer or bias and supply modulation, or with addition of thermal (white) noise that has both an AM and a PM noise component. Mathematically phase noise is a multiplicative "phasor" with the signal in time domain.

For the phase noise discussions here $A(t)$ s assumed to be a constant amplitude R as shown in Figure 1 that depicts the scaled phase noise "phasor" $Re^{j\emptyset(t)}$ and the various related components n(t), s(t) and h(t) than can be measured to quantify it. Here r(t) is the instantaneous position of this phasor and it makes an angle $\emptyset(t)$ with the real-axis I. The instantaneous phase $\emptyset(t)$ is related to the error vector n(t) and the perpendicular height h(t) as shown in Figure 1. The error vector n(t) is the instantaneous noise vector that can be directly measured from the spectrum of the signal e.g. using Keysight Spectrum Analyzers and translated to the actual phase noise plot using X-series Measurement App N9068EM0E. The Analog Phase Detector of the N5511A measures the perpendicular distance $h(t) = \sin[\emptyset(t)]$. The phase $\emptyset(t)$ can be measured directly using a complex signal digital Phase Demodulator as is done in the Keysight PNA-X with the Residual Phase Noise measurement option. Expanding Equation 1,

$$v_n(t) = A(t)[\cos(w_c t)\cos(\emptyset(t)) - \sin(w_c t)\sin(\emptyset(t))] \quad (3)$$

Under small angle condition of $|\emptyset(t)| << \frac{\pi}{2}$,

$$v_n(t) = A(t)[\cos(w_c t) - \sin(w_c t)\emptyset(t)] \quad (4)$$

For a narrow measurement bandwidth at offset frequency w_m,

$$\emptyset(t) = \emptyset_0 \sin(w_m t), \quad \emptyset_0 << \frac{\pi}{2}, \quad (5)$$

$$v_n(t) = A(t)[\cos(w_c t) - \frac{\emptyset_0}{2}\cos(w_c - w_m)t + \frac{\emptyset_0}{2}\cos(w_c + w_m)t] \quad (6)$$

From the above Equation we see there exists two sidebands at frequencies $(w_c - w_m)$ and $(w_c + w_m)$.

The power of each sideband relative to carrier is $\frac{\emptyset_0^2}{4}$ and is defined as the phase noise power $L_v(f_m)$ obtained directly from the signal spectrum under **small angle condition**. Note that $\omega_m = 2\pi f_m$.

Therefore,

$$L_v(f_m) = \frac{\emptyset_0^2}{4} \quad \text{defined for } \textbf{small angle condition} \quad (7)$$

The subscript v in $L_v(f_m)$ is a reminder that the phase noise is being derived from the signal spectrum under small angle conditions. Independent of whether the signal satisfies the small angle condition, the power of the phase modulation signal $\emptyset(t) = \emptyset_0 \sin(w_m t)$ is denoted as

$$S_{\emptyset}(f_m) = \frac{\emptyset_0^2}{2} \quad (8)$$

Using Equations (7) and (8) we see that under small angle condition

$$L_v(\mathrm{f}) \approx \frac{1}{2} S_{\emptyset}(f) \quad (9)$$

Note that $S_{\emptyset}(f) = \emptyset_{rms}^2(f)\frac{1}{BW}$ measured as rms phase in a resolution bandwidth BW centered at frequency f.

From Figure 1 it is apparent that when $|\emptyset(t)| << \frac{\pi}{2}$,

$$\emptyset(t) \cong \frac{n(t)}{R} \cong \frac{h(t)}{R}, \quad (10)$$

From Equations (9) and (10) $L_v(\mathrm{f})$ is therefore same as half of the power spectral density of n(t)/R and h(t)/R. Direct sideband spectral measurement of $v_n(t)$ provides power spectral density of n(t). The Phase Detector output of the N5511A directly provides the signal $h(t)$ when the RF and LO inputs to this detector are in Quadrature (90° relative phase).

IEEE Phase Noise Definition to Overcome Small Angle Limit

Historically phase noise spectrum was defined as $\mathrm{L_v(f) = S(f)}$, where $\mathrm{S(f)}$ is the power spectral density of the normalized noise n(t)/R. Here f is the frequency of n(t) and not the carrier. F is the offset from the carrier frequency F. $\mathrm{S(f)}$ is same as the normalized signal power spectral density of $\mathrm{v_n(t)/R}$ i.e. $\mathrm{v_n(t)}$ normalized to carrier power for f>0, and denoted as $\mathrm{S_v(f+F)}$. Note that $\mathrm{(f+F)}$ is the actual frequency of the signal spectrum as f is the offset from carrier frequency F.

We showed that $L_v(\mathrm{f}) \cong \frac{1}{2} S_{\emptyset}(f)$ only when rms[$\emptyset(t)$] is < 0.1 radians in the measurement bandwidth, equivalent to integrating $S_{\emptyset}(f)$ over the the measurement bandwidth.

In order to remove the difficulty when rms[$\emptyset(t)$] is >= 0.1 radians i.e. to define phase noise also for large angles, IEEE [1] defined the phase noise $\mathrm{L(f)}$ directly as:

$$\mathrm{L(f)} \equiv \frac{1}{2} S_{\emptyset}(f) \quad (11)$$

without any limits to the magnitude $|\emptyset(t)|$.

The IEEE [1] defines the Phase Noise $\mathrm{L(f)}$ as the half of the power spectral density of $\emptyset(t)$ with f as the frequency of the phase noise which is also the offset frequency measured from the carrier frequency F.

Therefore, when deriving $\mathrm{L(f)}$ from direct signal power spectral density measurement $\mathrm{S(f)}$ using $\mathrm{L(f) \cong S(f)}$, or when using analog Phase Detector of N5511A to measure h(t), we need to ensure the small angle condition of rms[$\emptyset(t)$] < 0.1 radians in the measurement bandwidth.

AM Noise

The AM noise $A(t)$ can be defined as

$$A(t) = R[1 + m(t)]$$

$$v_n(t) = R[1 + m(t)] \cdot \cos[w_c t + \emptyset(t)]$$

Taking a 1-tone example for $m(t)$

$$m(t) = m_o \cos(w_{am} t)$$

Expanding on the AM noise $A(t)$ the noisy carrier with phase noise $\emptyset(t)$:

$$v_n(t) = R[1 + m_o \cos(w_{am} t)] \cdot \cos[w_c t + \emptyset(t)] \quad (12)$$

$$v_n(t) = R\{\cos[w_c t + \emptyset(t)] + m_o \cos(w_{am} t) \cdot \cos[w_c t + \emptyset(t)]\}$$

$$v_n(t) = R\{\cos[w_c t + \emptyset(t)] + \frac{m_o}{2}\cos[(w_c + w_{am})t + \frac{\emptyset(t)}{2}] + \frac{m_o}{2}\cos[(w_c - w_{am})t + \frac{\emptyset(t)}{2}]\}$$

There are 3 terms: $\cos[w_c t + \emptyset(t)]$ which is the carrier with phase noise, and two AM sideband signals centered at frequencies $(w_c + w_{am})$ and $(w_c - w_{am})$ with amplitude $\frac{m_o}{2}$ and phase modulation $\frac{\emptyset(t)}{2}$. This phase modulation $\frac{\emptyset(t)}{2}$ power is much lower than that of the AM sideband carrier and therefore can be ignored relative to the sideband carriers. The resultant signal is

$$v_n(t) = R\{\cos[w_c t + \emptyset(t)] + \frac{m_o}{2}\cos[(w_c + w_{am})t] + \frac{m_o}{2}\cos[(w_c - w_{am})t]\} \quad (13)$$

The figure below depicts the AM and Phase Noise for 1-tone AM and Phase modulations at arbitrary frequencies w_{am} and w_m respectively. The phase noise term $\cos[w_c t + \emptyset(t)]$ is depicted by Equation 6.

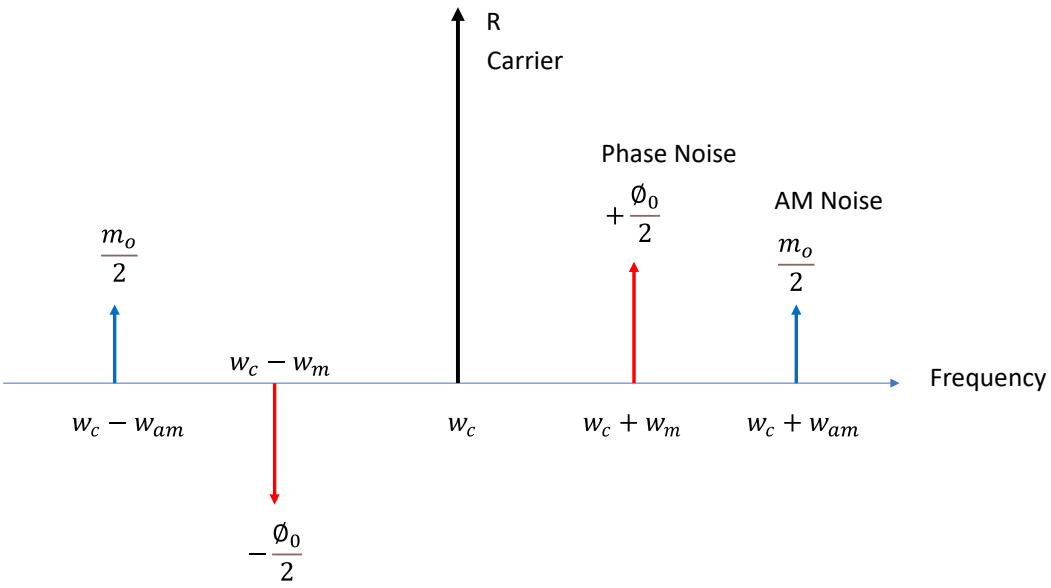


Figure 1b. AM and PM spurs with modulation frequencies w_{am} and w_m respectively

Analog Phase Detector

The N5511A incorporates RF and Microwave Phase Detectors (PD) for detecting phase noise in different frequency bands. They are double balanced mixer analog phase detectors that operate in phase quadrature wherein the "RF" Source signal is down converted to 0 Hz by mixing with a 90 degree phase locked "LO" signal as shown below for Absolute Phase Noise measurement. Quadrature phase ensures that the output signal is the demodulated relative phase variation between the RF and LO due to phase noise while the average or DC term is ideally zero. DC-nulling at the PD output is used to establish phase quadrature by a PLL in steady state. The AM noise of both the RF and LO are suppressed due to the DC output being ideally zero. The PD has a voltage conversion gain G_D under matched conditions, and it is dependent on the RF and LO drive levels. The output of the PD after down conversion of $v_n(t)$ is

$$s(t) = G_D * v_n(t) * \sin(w_c t)$$

$$s(t) = G_D * A(t)\left[\cos(w_c t) - \frac{\emptyset_0}{2}\cos(w_c - w_m)\,t + \frac{\emptyset_0}{2}\cos(w_c + w_m)\,t\right] * \sin(w_c t)$$

$$s(t) = G_D * A(t)\frac{\emptyset_0}{2}\sin(w_m t) \qquad \textbf{... after low pass filtering to remove } 2w_c\textbf{t terms}$$

$$s(t) = G_D * R[1 + m(t)]\frac{\emptyset_0}{2}\sin(w_m t), \qquad m(t) \ll 1 \; and \; \emptyset_0 \ll 1$$

$$s(t) = [G_D * R]\,\frac{\emptyset_0}{2}\sin(w_m t)$$

The PD constant K_D is defined as [output volts] / [input radians] in a measurement bandwidth. For the above 1-tone phase modulation, the input phase signal is $\emptyset_0 \sin(w_m t)$ and therefore K_D is:

$$K_D = [G_D * R]/2.$$

The N5511A accurately calibrates the PD constant K_D to convert the PD output voltage into radians.

In further analysis we ignore the terms G_D and R by assuming them to be 1 to simplify the equations and to easily track the noise signal amplitudes.

The simplified PD output signal is then

$$s(t) = \frac{\emptyset_0}{2}\sin(w_m t)$$

The PD operation is depicted in Figure 2 for the above simplified case, with G_D and R assumed 1, and K_D assumed ½ for simplicity.

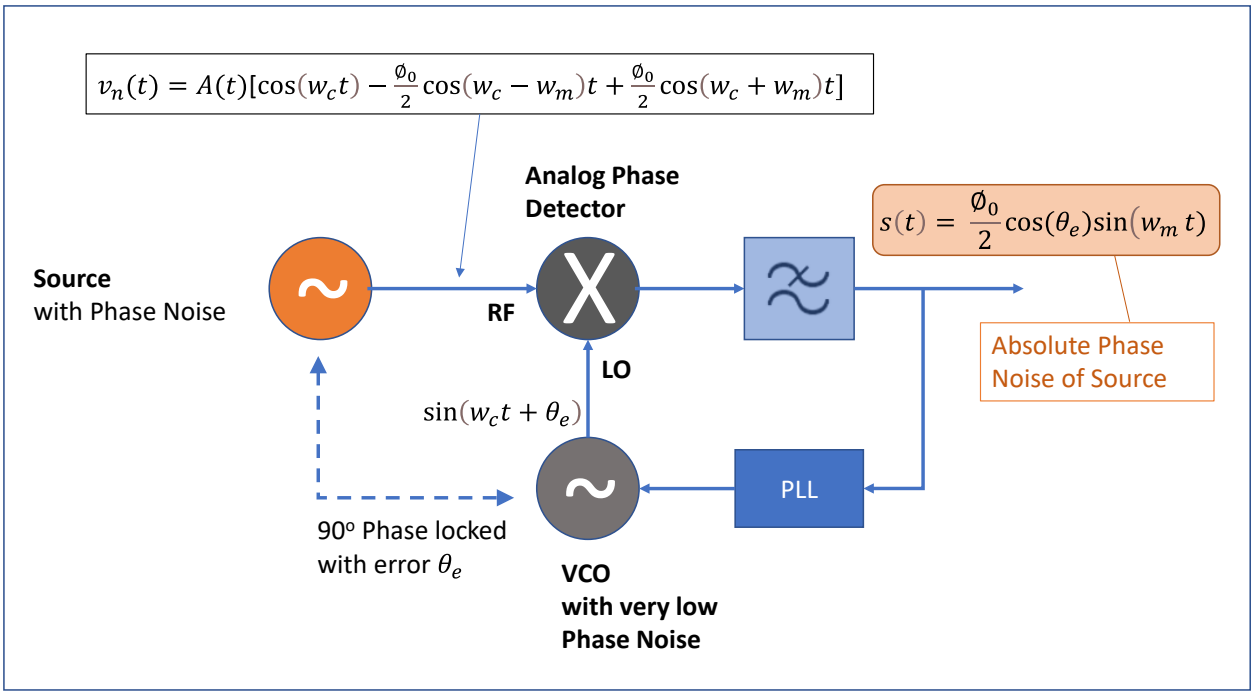


Figure 2. Analog Phase Detector operation for Absolute Phase noise. The AM noise is ignored i.e. $A(t) \approx R = 1$. The PD conversion voltage gain G_D is assumed 1. The PD constant is $K_D = [G_D * R]/2$ and is assumed ½ for simplicity.

With the introduction of quadrature phase error θ_e in the LO, the output of the PD after down conversion of $v_n(t)$ is

$$s(t) = v_n(t) * \sin(w_c t + \theta_e)$$

When measuring the phase noise, the PD output is divided by R*KD where KD is the detector gain constant that is obtained from calibration and assumed to be 1 here, and R is the input signal amplitude.

Ignoring the $m(t)$ AM term $A(t)$ of $v_n(t)$, and dividing the PD output by R,

$$s(t) = \left[\cos(w_c t) - \frac{\emptyset_0}{2}\cos(w_c - w_m)\,t + \frac{\emptyset_0}{2}\cos(w_c + w_m)\,t\right] * \sin(w_c t + \theta_e)$$

After low pass filtering to remove $2w_c$t terms, ignoring sign, and ignoring DC term of $\frac{1}{2}\sin(\theta_e)$ that is actually the suppressed AM signal $\frac{A(t)}{2}\sin(\theta_e)$ leaking out,

$$s(t) = \frac{\emptyset_0}{4}[\sin(-w_m t - \theta_e) + \sin(-w_m t + \theta_e)]$$

$$s(t) = \frac{\emptyset_0}{2}\cos(\theta_e)\sin(w_m t)$$

Absolute Phase Noise Conversion Gain

Absolute Phase Noise Conversion gain of the PD is the ratio of the down converted baseband output power to the input RF upper (or lower) sideband power:

$$Gc = [\frac{\emptyset_0}{2}\cos(\theta_e)]^2 / [\frac{\emptyset_0}{2}]^2 = \cos^2(\theta_e) \quad \text{...power gain of PD} \quad (14)$$

The common amplitude term R or $A(t)$ is not shown for the input and output signal amplitudes in the above equation.

AM Noise Conversion Gain

The PD operates in phase quadrature i.e. 90 degree relative phase between the RF and LO signals and with a quadrature phase error of θ_e. The Carrier with AM noise at the RF port is

$$v_n(t) = R\{\cos[w_c t + \emptyset(t)] + \frac{m_o}{2}\cos[(w_c + w_{am})t] + \frac{m_o}{2}\cos[(w_c - w_{am})t]\}$$

The LO is $\sin(w_c t + \theta_e)$

The PD output is

$$s(t) = v_n(t) * \sin(w_c t + \theta_e)$$

When measuring the AM noise, the PD output is divided by R*K_D where K_D is the detector gain constant that is obtained from calibration and assumed to be 1 here, and R is the input signal amplitude.

Dividing the PD output by R,

$$s(t) = \left[\cos(w_c t) + \frac{m_o}{2}\cos[(w_c + w_{am})t] + \frac{m_o}{2}\cos[(w_c - w_{am})t\right] * \sin(w_c t + \theta_e)$$

After low pass filtering to remove $2w_c$t terms, ignoring sign, and ignoring DC term of $\frac{1}{2}\sin(\theta_e)$ that is actually the suppressed AM signal $\frac{A(t)}{2}\sin(\theta_e)$ leaking out,

$$s(t) = \frac{m_0}{4}[\sin(w_{am}\, t + \theta_e) + \sin(-w_{am}\, t + \theta_e)]$$

$$s(t) = \frac{m_0}{2}\sin(\theta_e)\cos(w_{am}\, t)$$

Conversion gain is Ratio of PD baseband output power to the input RF upper (or lower) sideband power:

$$Gc = [\frac{m_0}{2}\sin(\theta_e)]^2 / [\frac{m_0}{2}]^2 = \sin^2(\theta_e) \quad \text{...power gain of PD} \quad (15)$$

The common amplitude term R is not shown for the input and output signal amplitudes in the above equation.

Comparison of AM and PM Conversion Gains

The AM and PM conversion gains of the PD as a function of the relative phase between the RF and LO inputs are shown in Figure 3. In the presence of a DC Error 'Verror' at the output of the PD, the DC-Null steady state of the PLL will no longer occur at 90°, but will be at a slight deviation from quadrature phase with an error 'ph' such that sin(ph) = -Verror/A where A is the maximum or peak dc output. As shown in Figure 4 when the DC Error is 17% of the full scale, there is 10 deg phase error, and the corresponding AM rejection is only 15 dB. Such amounts of DC errors are possible and reflect the degraded AM rejection of the PD in its presence. DC error arise from mixer self-detection and 2nd order nonlinearity.

Variation of Detection Gain with Quadrature Phase Error θ_e

Absolute Phase Noise conversion gain:

$$10\log\left[\cos^2(\theta_e)\right] \text{ dB}$$

AM Noise conversion gain:

$$10\log\left[\sin^2(\theta_e)\right] \text{ dB}$$

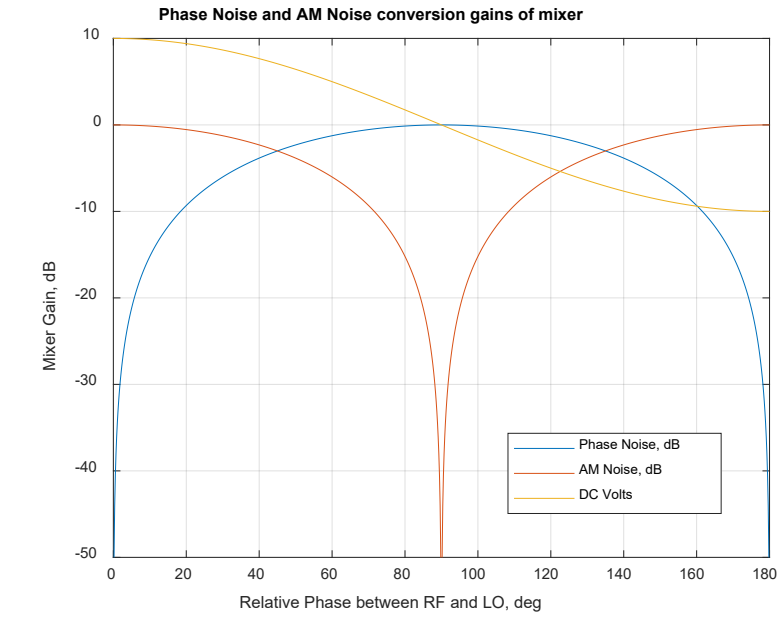


Figure 3. PD response for Phase Modulation and Amplitude Modulation. At 90 degrees or Quadrature Phase, the AM is rejected fully.

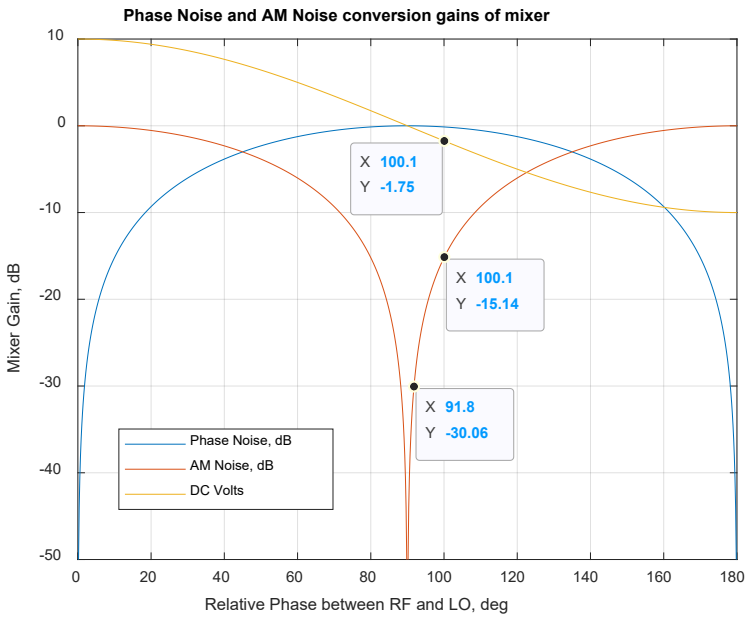


Figure 4. AM suppression with DC error. For a DC Error of 1.75V and a peak PD range of ±10V, there is 10° quadrature error and the AM suppression is only 15 dB. There is insignificant reduction for the Phase Noise conversion gain. The DC Error comes from 2nd order nonlinearity of the PD and from self-detection of RF and LO signals. The DC offsets of the remaining circuits like LNA etc can be calibrated out in the absense of signals at the factory or during test.

AM at Phase Detector Output

AM component of the Source signal appears at the output of the PD through these mechanisms:

- Direct AM detection
  - The PD output has the baseband AM modulation component $\frac{A(t)}{2}\sin(\theta_e)$ as derived earlier. This occurs because of a DC term $\sin(\theta_e)$ at the PD output resulting from Quadrature phase error θ_e in the PD during the phase noise detection down conversion.
  - 2nd order nonlinearity also contributes to the DC error at the PD output, thereby contributing to the quadrature phase error θ_e that results in the AM component $\frac{A(t)}{2}\sin(\theta_e)$ at the PD output.
  - The output AM spectrum is an attenuated version of the input AM spectrum. The attenuation factor is $\sin(\theta_e)$.
- Second Order Nonlinearity
  - This is a square-law nonlinearity phenomenon producing 2nd order intermodulation (IM2)
    - mostly at the RF input stage of a mixer PD,
    - and to a lower extent at the IF output stage.
    - It is also produced by RF leaking into the LO port resulting in self-detection of the RF signal by the mixer.
  - When the LO duty cycle is not 50%, it has a DC component that feeds the RF input stage IM2 component to the IF output.
  - The IM2 produced at the output stage of a mixer is cancelled in a balanced circuit, but imbalance due to imperfections will result in some of it appearing at the IF load.
  - The IM2 spectrum is from the squaring of the source signal, and only a part of it is a scaled version of the AM component of the source. The other part is twice the frequency but 6 dB lower in power. It is derived next.

Direct AM Detection

With LO to RF leakage signal amplitude ρ*R (denoted in terms of RF signal amplitude), there is a DC term at the PD output (normalized to R)

$$v_{dc} = \rho * \sin(w_c t + \theta_{leak}) * \sin(w_c t)$$

After low pass filtering, and leakage phase shift $\theta_{leak} \ll \frac{\pi}{2}$

$$v_{dc} \approx \frac{\rho}{2}$$

This creates a quadrature phase error θ_e where $\sin(\theta_e) = \frac{\rho}{2}$ radian when the system adjusts the LO phase for zero output of the PD, and it results in the AM of the RF signal leaking out based on the DC term $\frac{A(t)}{2}\sin(\theta_e)$. The output AM signal is (after dividing by R for the phase noise calibration).

A part of the DC term also comes from the 2nd order nonlinearity of the PD.

$$s_{am2}(t) = \frac{A(t)}{2R}\frac{\rho}{2} = [1 + m(t)]\frac{\rho}{4}$$

Typically, $\frac{\rho}{4}$ is around -20 dB which is the AM suppression of the PD.

Second Order Nonlinearity

The PD gain for the RF input stage or IF output stage can be modeled as

$$y(t) = a_1 x(t) + a_2 x^2(t) + \cdots$$

When $x(t)$ is a 2-tone signal at frequencies f1 and f2, the $a_2 x^2(t)$ term produces an IM2 component at |f1-f2| frequency. When produced at the RF input stage, it appears at the PD output due to a DC component in the LO arising from duty cycle error. The PD mixer 2-Tone SOI (second order intercept point) specification determines the output IM2 level based on the input RF power.

A 1-tone AM (or PM) modulation at frequency w_{am} produces an upper and lower sideband around the carrier at a distance of w_{am} as shown earlier. There are now 3 tones with relative separations of w_{am} and $2w_{am}$. With the $a_2 x^2(t)$ operation, there will be a DC term, and two tones at the PD output at frequencies of w_{am} and $2w_{am}$. The power of the signal at $2w_{am}$ is $\frac{m_o}{2}$ (<<1) relative to that at w_{am} and can therefore be ignored.

With RF to LO leakage factor γ, there is an AM detection component at the PD output due to self-mixing:

$$s_{am1}(t) = v_n(t) * [\gamma v_n(t)]$$

$$s_{am1}(t) = \frac{\gamma}{2} A^2(t)$$

Residual Phase Noise

Residual phase noise is associated with a 2-port device (DUT) and is a measure of the additional phase noise the device contributes to its input signal when measured at its output. A mathematical model is shown below. Typically, if the DUT is an Amplifier or a non-frequency translating device, then the DUT's contribution is called Additive Phase Noise. If the DUT is a frequency translating device like a PLL Synthesizer with an input reference frequency signal and an output synthesized frequency, then the DUT's contribution is called Residual Phase Noise. In this document we use the term Residual phase noise to include Additive phase noise as well. It is equivalent to the Absolute Phase Noise measured at the DUT output when it is driven by an ideal noise-free signal at the input. However, as this is not possible, we need to de-embed the input Absolute Phase Noise from the output Absolute Phase Noise to compute the Residual phase noise. In the N5511A, this is achieved using a method wherein the input Stimulus phase noise is simply cancelled when measuring the DUT output phase noise using a PD as described next.

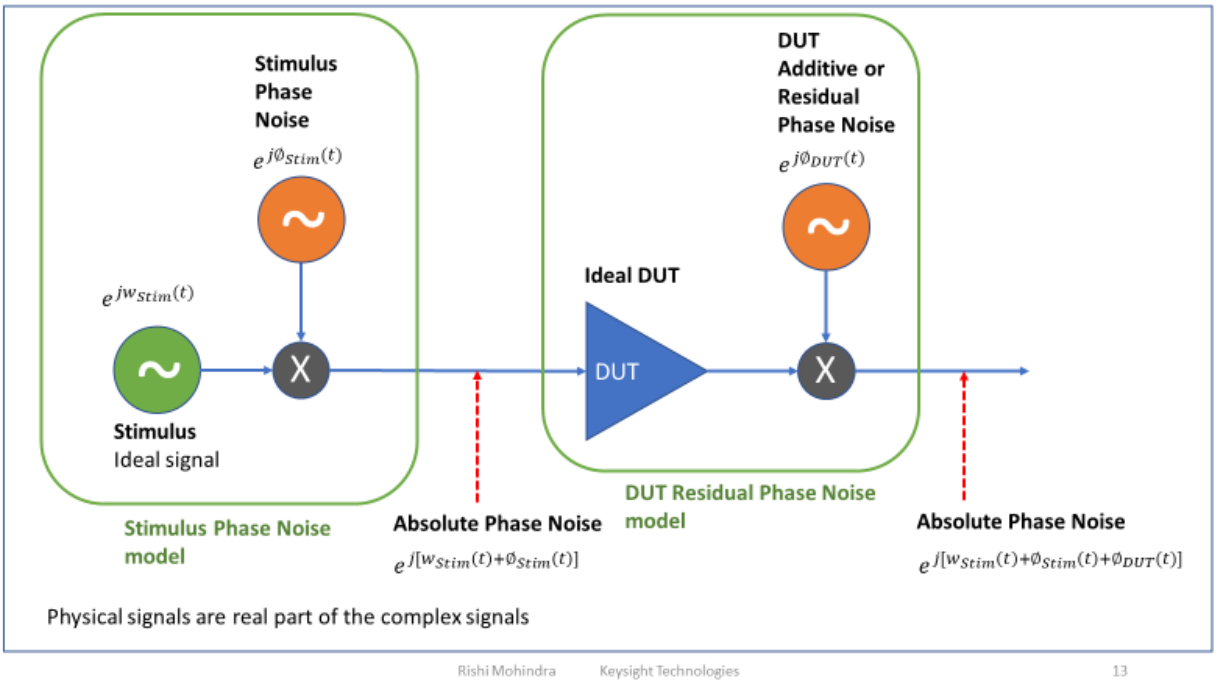


Figure 5. Absolute and Residual phase noise

Using a quadrature phase detection method as shown below, the Stimulus is mixed with a 90° phase shifted version of itself, and the resultant output is zero independent of the Stimulus phase noise.

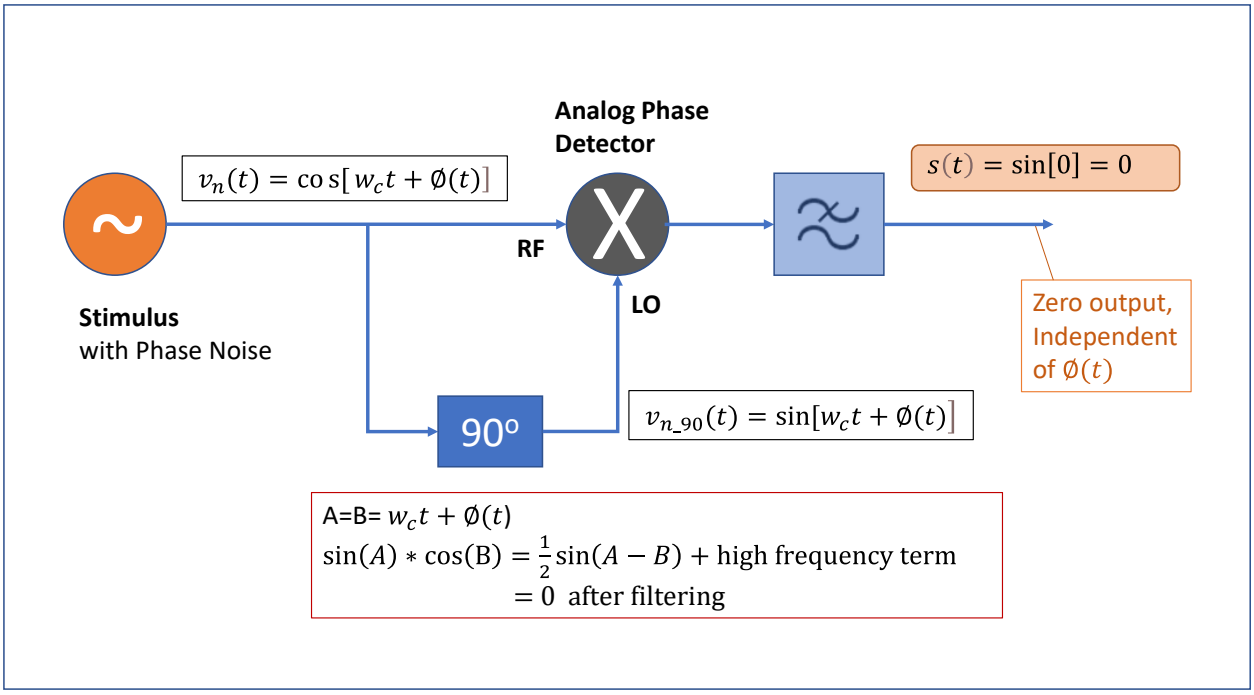


Figure 6. Perfect cancellation of Stimulus phase noise at PD down converted output.

If we insert a DUT in one of the arms of the PD, then only the residual phase noise of the DUT will appear at the output of the PD as shown below. The Absolute phase noise of the Stimulus is completely suppressed.

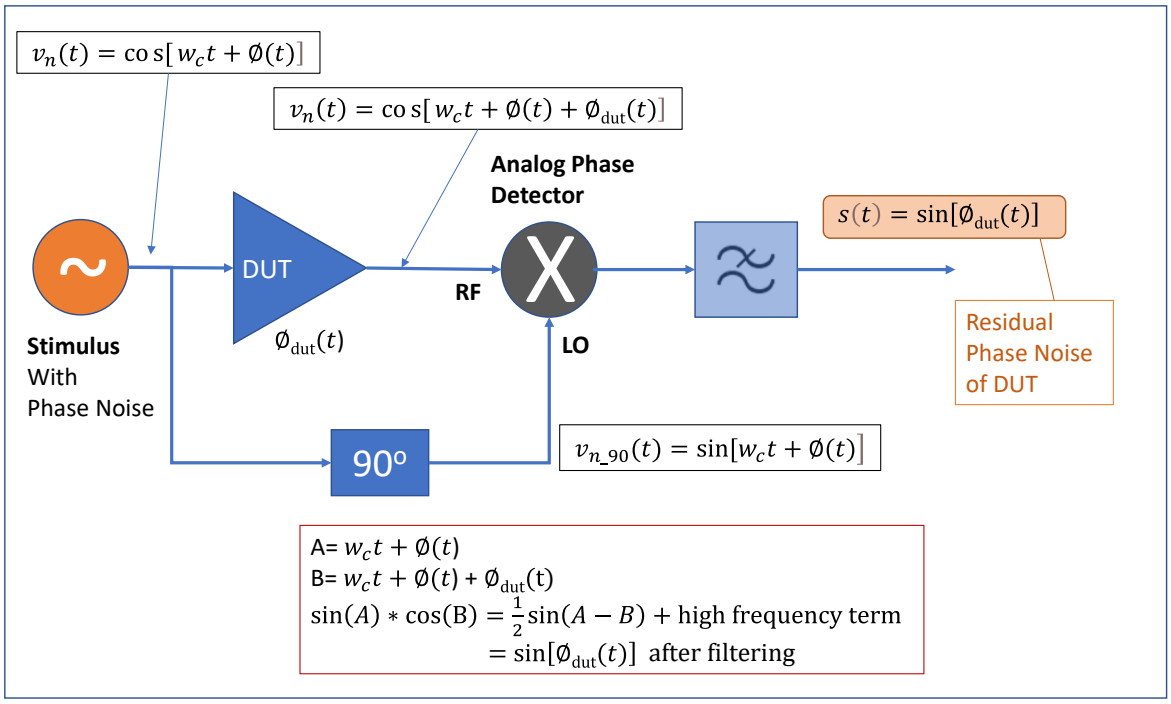


Figure 7. DUT residual phase noise measurement. Only the DUT's residual phase noise appears at the PD down converted output while the Stimulus phase noise is completely rejected.

Stimulus Phase Noise Leakage at Phase Detector Output

In a practical set up, the 90° phase shift in the LO path is achieved using a Line Stretcher based Delay Line whose electrical delay τ is adjusted such that the associated phase shift is an odd multiple of 90 degree at the carrier frequency w_c. In practice the delay τ is adjusted so that there is a DC-null at the PD output which occurs when the relative phase between the RF and LO ports is an odd multiple of 90 degree. Therefore,

$$\boldsymbol{w_c \tau = n\frac{\pi}{2}} \quad \text{...for odd n} \qquad (15a)$$

Unfortunately, the delay τ produces an exact 90° (or odd multiple) quadrature phase shift only at w_c and not at the offsetted frequencies $w = w_c + w_m$. With increasing offsetted frequency $w = w_c + w_m$ a frequency dependent phase error θ_{em} is produced in the quadrature phase shifter delay line and some amount of the Stimulus phase noise starts to appear at the PD output, increasing with the offset frequency w_m. It is depicted in the figure below for the case of a 1-tone phase noise spur at a modulation frequency w_m. The equations for the Stimulus leakage are derived in the next section. The relation between the frequency dependent quadrature phase error θ_{em} and the offset frequency w_m is shown below.

$$(w_c + w_m)\tau = n\frac{\pi}{2} + \theta_{em} \quad \text{...for odd n}$$

$$\theta_{em} = w_m \tau = n\frac{\pi}{2}\frac{\boldsymbol{w_m}}{\boldsymbol{w_c}} \quad \text{...for odd n}$$

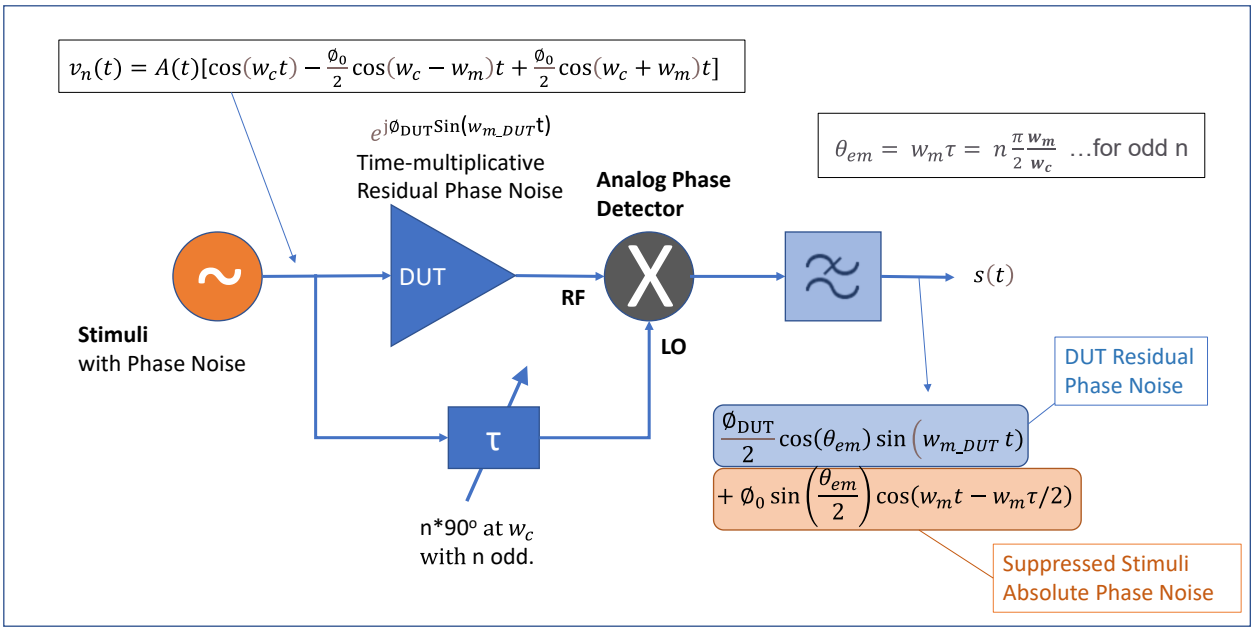


Figure 8. Practical system for DUT residual phase noise measurement. Using a practical phase shifter based on an adjustable delay line, Stimulus phase noise leaks at the PD output, increasing with w_m. The DUT's residual phase noise appears at the PD down converted with a relatively flat gain over offset frequency w_m. The small frequency dependent quadrature phase error is $\theta_{em} = w_m\tau = n\frac{\pi}{2}\frac{w_m}{w_c}$ for odd n, and the $\cos(\theta_{em})$ term is ~1, and can be ignored.

Derivation for Stimulus Leakage

Referring to Figure 8, and from Equation (6) for 1-tone phase modulation of Stimulus, and ignoring the DUT phase noise, the PD output is

$$r(t) = v_n(t) * v_n(t-\tau)$$

$$r(t) = \left[\cos(w_c t) - \frac{\emptyset_0}{2}\cos(w_c - w_m)\,t + \frac{\emptyset_0}{2}\cos(w_c + w_m)\,t\right] * \left[\cos(w_c(t-\tau)) - \frac{\emptyset_0}{2}\cos(w_c - w_m)\,(t-\tau) + \frac{\emptyset_0}{2}\cos(w_c + w_m)(t-\tau)\right]$$

After low pass filtering to remove $2w_c$t terms, and ignoring $\emptyset_0^2$ terms that are very small,

$$r(t) = \frac{1}{2}\cos(w_c\tau) - \frac{\emptyset_0}{4}\cos(w_c\tau + w_m t - w_m\tau) + \frac{\emptyset_0}{4}\cos(w_c\tau - w_m t + w_m\tau) - \frac{\emptyset_0}{4}\cos(w_c\tau - w_m t) + \frac{\emptyset_0}{4}\cos(w_c\tau + w_m t)$$

$$r(t) = \frac{1}{2}\cos(w_c\tau) + \frac{\emptyset_0}{4}[-\cos(w_c\tau + w_m t - w_m\tau) + \cos(w_c\tau - w_m t + w_m\tau) - \cos(w_c\tau - w_m t) + \cos(w_c\tau + w_m t)]$$

Putting $w_c\tau = \frac{\pi}{2}$ and [later $w_c\tau = n\frac{\pi}{2}$ for odd n], and with

$\cos\left(\frac{\pi}{2} - \theta\right) = \sin(\theta)$,

$\cos\left(n\frac{\pi}{2} - \theta\right) = \sin(\theta)$, n=1, 5, 9, ...

$\cos\left(n\frac{\pi}{2}-\theta\right) = -\sin(\theta)$, n=3, 7, 11, ...

and ignoring DC term,

$$r(t) = \frac{\emptyset_0}{4}[-\sin(-w_m t + w_m \tau) + \sin(w_m t - w_m \tau) - \sin(w_m t) + \sin(-w_m t)]$$

$$r(t) = \frac{\emptyset_0}{2}[\sin(w_m t - w_m \tau) - \sin(w_m t)]$$

$$r(t) = \emptyset_0 \cos\left(w_m t - \frac{w_m \tau}{2}\right) \sin\left(\frac{w_m \tau}{2}\right)$$

$$r(t) = \emptyset_0 \sin\left(\frac{\pi}{2}\frac{w_m}{2w_c}\right) \cos(w_m t + \theta_\tau)$$

$$r(t) = \emptyset_0 \sin\left(\frac{\theta_{em}}{2}\right) \cos(w_m t + \theta_\tau)$$

Stimulus PN Suppression

The Stimulus PN Suppression is defined here as the ratio of the Stimulus phase noise power leaking at the PD output in the Residual Phase Noise set up, to that of the Absolute phase noise power of the Stimulus at the PD output under a similar set up but with a noise-free quadrature LO. The latter is a hypothetical set up for a regular down conversion of the Stimulus signal $v_n(t)$ using an ideal quadrature LO of $\sin(w_c t)$ as below:

$$s(t) = v_n(t) * \sin(w_c t)$$

$$s(t) = \left[\cos(w_c t) - \frac{\emptyset_0}{2}\cos(w_c - w_m)\, t + \frac{\emptyset_0}{2}\cos(w_c + w_m)\, t\right] * \sin(w_c t)$$

After low pass filtering to remove $2w_c$t terms,

$$s(t) = \frac{\emptyset_0}{2}\sin(w_m)\, t$$

The Stimulus PN Suppression is the ratio of the power of $r(t)$ and $s(t)$:

Suppression = $[\frac{\emptyset_0^2}{2}\sin^2(\frac{\pi}{2}\frac{w_m}{2w_c})]/[\frac{\emptyset_0^2}{8}]$

Suppression = $4\sin^2(\frac{\pi}{2}\frac{w_m}{2w_c})$

For phase shifts of odd multiples of 90 degrees i.e. 90, 270, 450, ... degrees i.e. for n*90, n=1, 3, 5, ...

Suppression = $4\sin^2(n\frac{\pi}{2}\frac{w_m}{2w_c})$ ...power ratio

Suppression in dB = $10log[4\sin^2\left(n\frac{\pi}{2}\frac{w_m}{2w_c}\right)]$ **(15b)**

The Stimulus PN Suppression in dB is plotted in Figure 9 for odd values of 'n' in the x-axis.

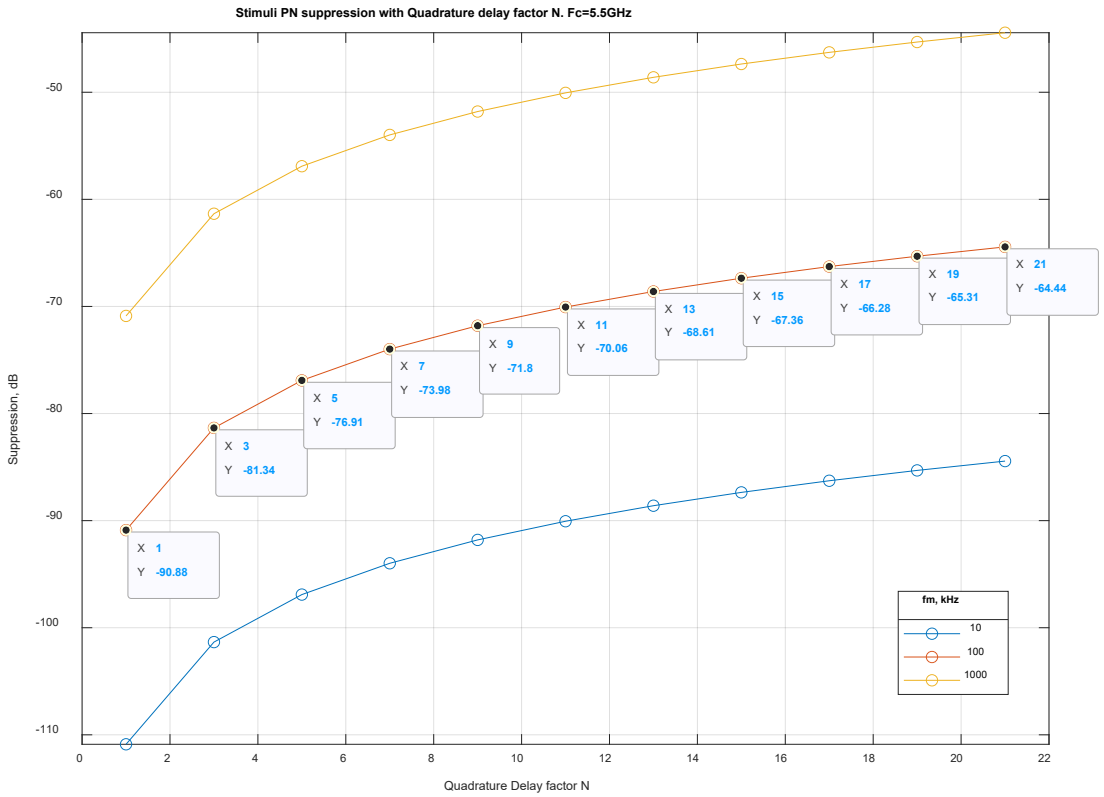


Figure 9. Suppression of Stimulus phase noise at PD output vs odd Quadrature Delay multiplier 'n' from 1 to 21. The offset frequencies are 10 kHz, 100kHz and 1 MHz. Carrier frequency is 5.5 GHz.

The Stimulus PN Suppression is plotted in Figure 10 with the relative Quadrature Delays on the x-axis. The odd 'n' values are converted to the corresponding relative Quadrature Delay using the equation:

Quadrature Delays $\tau = \frac{n\pi}{2w_c}$

Relative delay τ is given by

$\boldsymbol{w_c\tau = n\frac{\pi}{2}}$ ...for odd n (15a)

Stimulus phase noise suppression varies with n, $\boldsymbol{w_m}$and $\boldsymbol{w_c}$ and is given by

Suppression in dB = $\boldsymbol{10log[4\sin^2\left(n\frac{\pi}{2}\frac{w_m}{2w_c}\right)]}$ (15b)

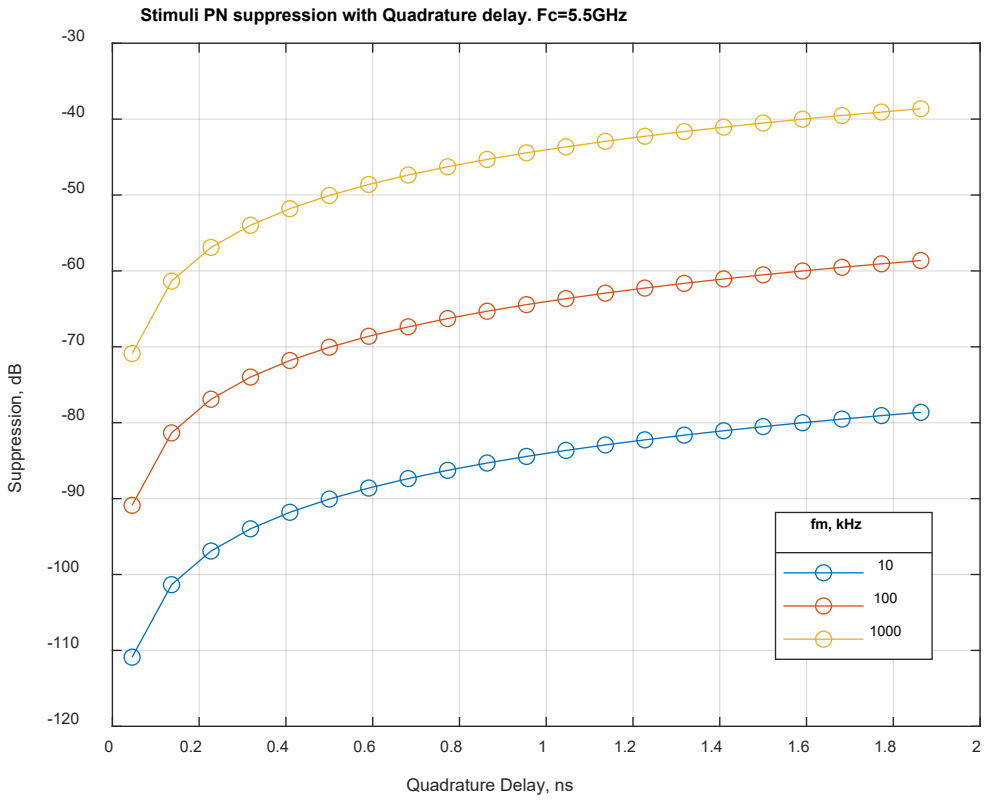


Figure 10. Suppression of Stimulus phase noise at PD output vs Quadrature Delay in ns, for 10 kHz, 100 kHz and 1000 kHz offsets at 5.5 GHz carrier.

The spectrum of the suppressed Stimulus phase noise are plotted below for 5.5 GHz and 500 MHz carriers, for the Keysight E8257D 'PSG' signal source. It determines the noise floor at the far offset frequencies e.g. > 1 MHz offsets. For closer offsets, the AM noise of the Stimulus and DUT can have a major influence, as shown later.

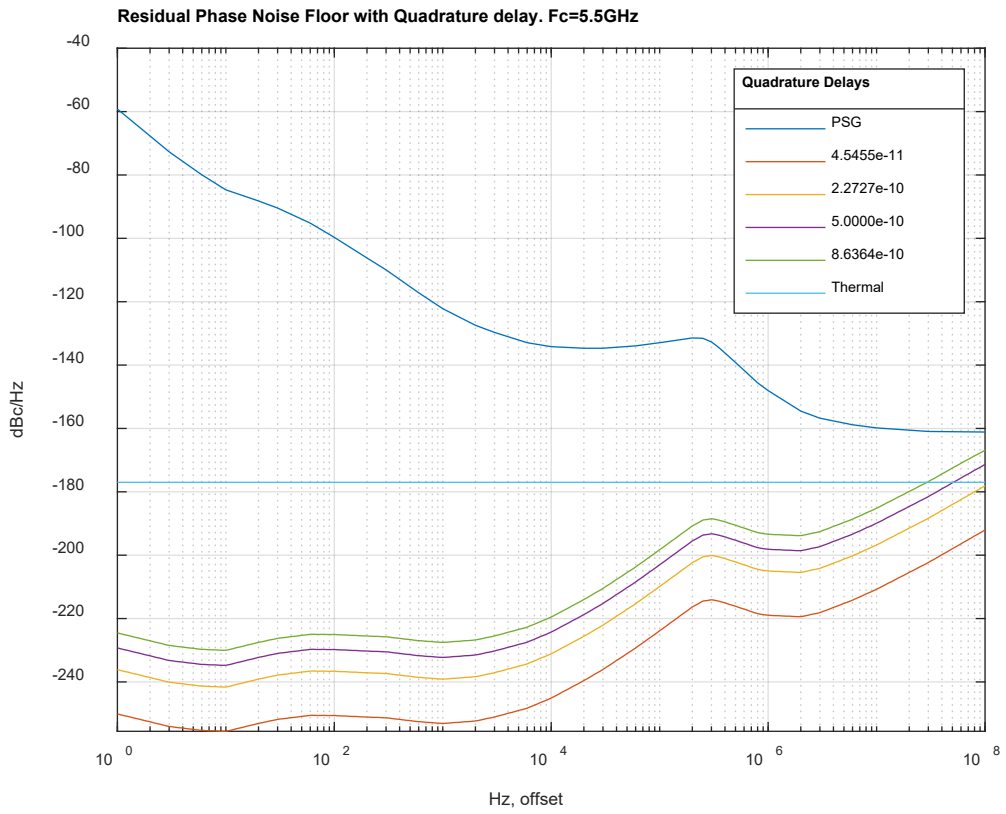


Figure 11. Keysight E8257D 'PSG' Stimulus Phase Noise at PD output, at 5.5 GHz. The Quadrature Delays are for n = 1, 5, 11 and 19. Overlaid are the -177 dBc/Hz thermal phase noise floor and the E8257D phase noise.

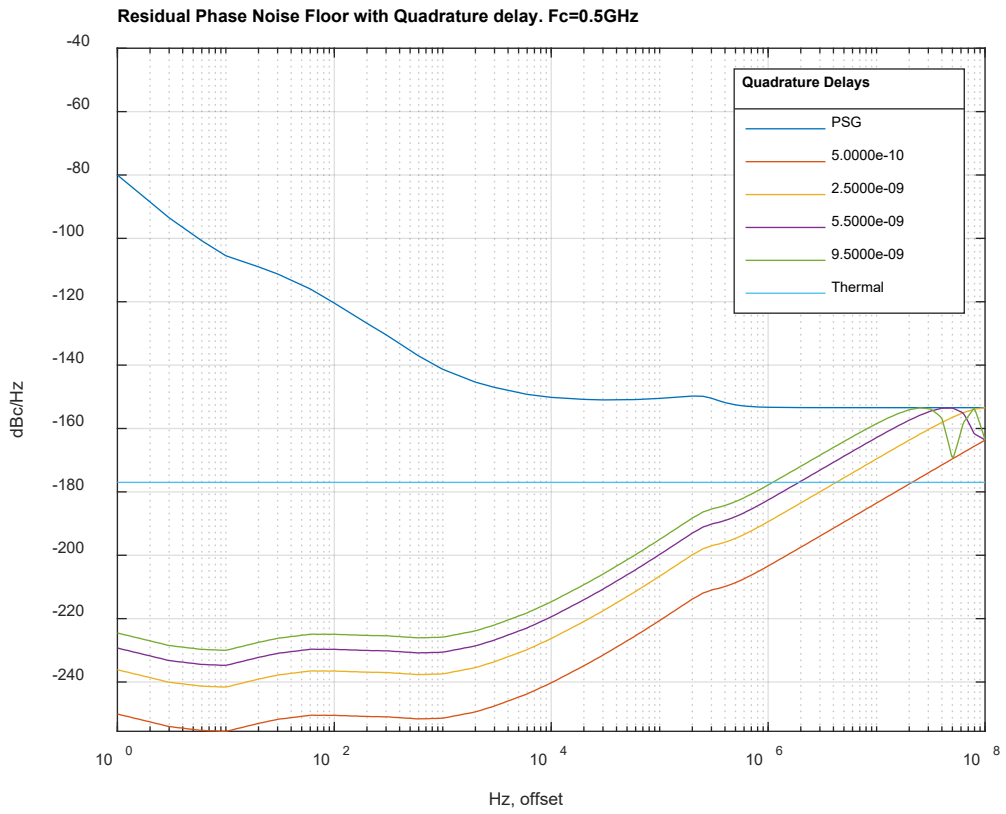


Figure 12. Keysight E8257D 'PSG' Stimulus Phase Noise at PD output, at 500 MHz. The Quadrature Delays are for n = 1, 5, 11 and 19. Overlaid are the -177 dBc/Hz thermal phase noise floor and the E8257D phase noise.

Impact of LO and RF Self-Detection on the Stimulus and DUT AM Noise Suppression

Leakage between the LO and RF ports creates a DC error due to self-detection in the PD. DC error also arises from the 2nd (even) order nonlinearity in the PD. These lead to an error in the Quadrature phase setting that is based on DC-nulling at the PD output done by adjusting the delay of a phase shifter line stretcher. Due to this phase error θ_e, only a small amount of additional Stimulus phase noise leaks at the PD output as shown by the equation in the figure below, and as derived in the Appendix. However, the major impact of this phase error θ_e is reduced suppression of the Stimulus and DUT AM noise that appears at the PD output, as depicted in the figure below. The AM noise amplitude varies with $\sin(\theta_e)$ at the PD output as derived earlier in the section on AM Noise.

Stimulus AM noise suppression varies with Set-point Quadrature Phase Error $\boldsymbol{\theta_e}$

AM Suppression in dB = $10log[4\sin^2(\theta_e)]$

In addition to the set-point phase error of θ_e at carrier frequency $\boldsymbol{w_c}$, there is also an increasing phase error θ_{em} with offset frequency w_m due to the delay in the line stretcher. It has a major impact on the Stimulus phase noise leakage at the PD output. The suppression of the AM and PM noise at the PD output are shown in Figure 13 below. As the DUT signal is absent at the LO port of the PD, the DUT AM noise is 6 dB less than the Stimulus AM noise. However, if the PD internal circuit is designed to suppress the AM noise at the LO port, then Stimulus AM noise at the PD output will also be lower (up to 6 dB). The AM noise suppression is not sensitive to the offset frequency dependent quadrature phase error θ_{em} arising from the relative delay as shown below for the 1-tone sinusoidal AM modulation case.

Stimulus AM signal:

$$v_n(t) = R\{\cos[w_c t + \emptyset(t)] + \frac{m_o}{2}\cos[(w_c + w_{am})t] + \frac{m_o}{2}\cos[(w_c - w_{am})t]\}$$

PD output is

$$r(t) = v_n(t) * v_n(t - \tau)$$

With a set-point phase error of θ_e there is an error τ_e in the phase shifter delay given by

$w_c(\tau + \tau_e) = \frac{\pi}{2} + \theta_e$ where $w_c\tau_e = \theta_e$

Replacing τ with $\tau' = (\tau + \tau_e)$ and then substituting $w_c\tau_e$ with θ_e it can be shown that

$$r(t) = m_o\sin(\theta_e)\cos(w_m t + w_m\tau/2)$$

The AM Suppression is $[m_o\sin(\theta_e)]^2 / [\frac{m_o}{2}]^2 = 4\sin^2(\theta_e)$ ...power ratio

AM Suppression in dB = $10log[4\sin^2(\theta_e)]$ (16)

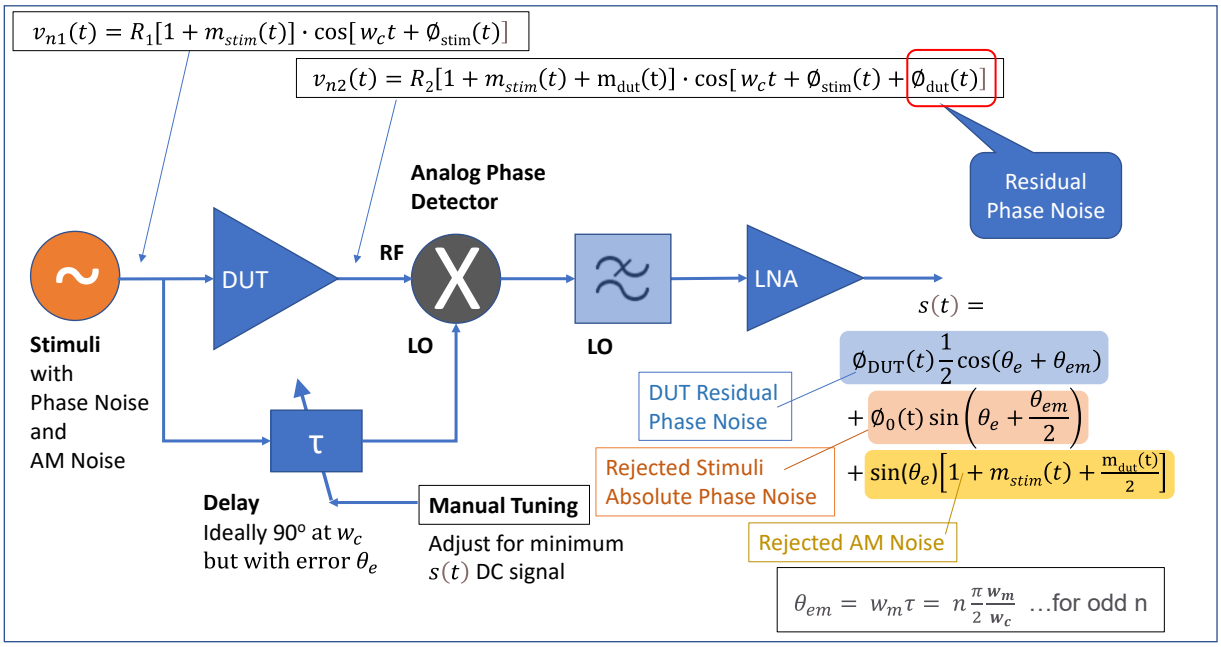
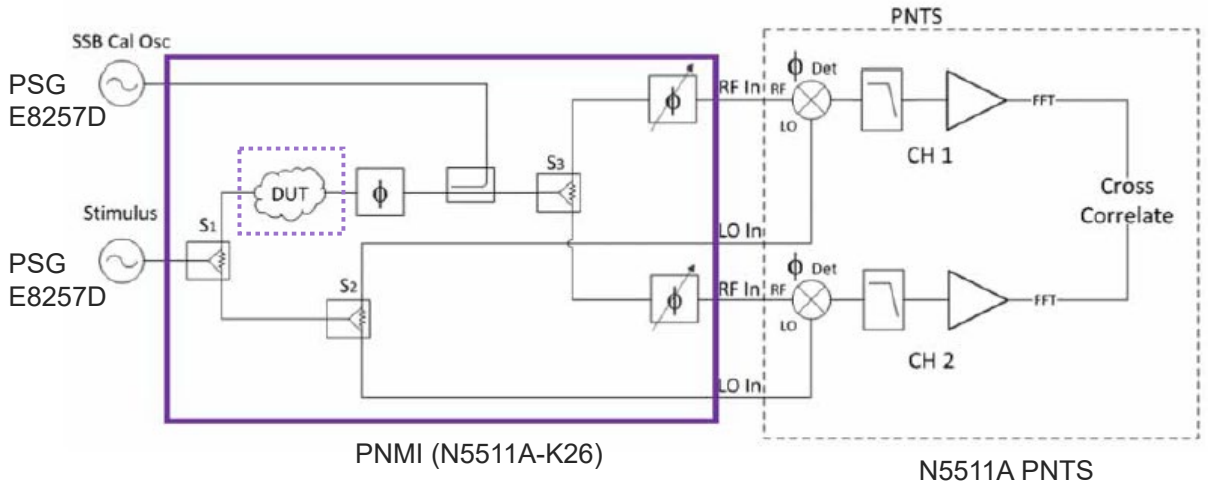


Figure 13. Set-point phase error $\boldsymbol{\theta_e}$ (due to DC-nulling error) has a major impact on the Stimulus and DUT AM noise at the PD output. The offset frequency dependent phase error θ_{em} (due to relative delay) has a major impact on the Stimulus phase noise at the PD output. The $\cos(\theta_e + \theta_{em})$ term is ~1, and can be ignored for the DUT residual phase noise.

Advanced Techniques for Residual Phase Noise Measurement Using N5511A and PNMI

The N5511A User Guide explains in detail how to measure the residual phase noise. In this section two advanced techniques are described for suppressing the measurement noise floor to improve the accuracy of the measured DUT residual phase noise and to enable usage of Stimulus with higher AM and PM noise. The first technique utilizes in-situ measurement and compensation of the relative quadrature delay to ensure it is minimized and identical for both channels in the Dual-Channel Cross-Correlation set up of the N5511A. This minimizes the Stimulus phase noise leakage at the PD output.

The second technique is for setting the Quadrature Phase Shifters more accurately to improve the AM suppression of the PD and requires an additional step after the initial DC-nulling adjustment. The usual method is to adjust the line stretchers (phase shifters) so that the PD output signal for each channel is at a DC-null which corresponds to a 90° relative phase shift between the RF and LO signals. However, due to self-detection in the PD and due to its 2nd order nonlinearity, there is a DC bias at its output, leading to a phase error at the DC-null setting. This phase error degrades the AM suppression of the PD as was shown earlier. Leakage of Stimulus and DUT AM down converted noise at the PD output increases the residual phase noise measurement noise floor. Note that the DC-nulling adjustment is done before the calibration of the PD gain, and before the residual phase noise measurement.



Tips: For these two techniques and in general with the PNMI usage, we can interchange the RF and LO signals at the PD input ports instead of using the typical set up depicted above. It is done to get a higher drive level for the LO port compared to the RF port and is the case when the DUT output power is higher than the Stimulus power minus approximately 3 dB of the internal splitter. The LO port of the PD has significant AM suppression compared to the RF port that is more linear. If the DUT has significant AM noise that could affect the measurements, it may be advantageous to feed the DUT signal to the LO port instead of the RF port. If the Stimulus AM noise is more than that of the DUT, then the DUT should be connected to the RF port. The SSB Cal Osc should be in the RF path so as not to suppress the AM component of the composite RF signal, to make the calibration more accurate. It is good practice to measure the Stimulus AM noise and subtract 30 dB to get an idea of the AM dependent residual phase noise floor with the improved AM suppression technique.

Relative Delay Estimation and Compensation

Spur Method

Using the theoretical suppression of the Stimulus phase noise as a function of the Quadrature delay (equation 15a and 15b), we determine the actual Quadrature Delay from the measured suppression of the Stimulus phase noise. The Stimulus phase noise suppression measurement is done in-situ during the DUT residual phase noise measurement without having to physically measure the delays from the Stimulus to the RF and LO ports of the PD that otherwise would have required using a Network Analyzer with TDR capability.

A known amount of a phase noise spur is created at the Stimulus to measure its suppression at the PD output during the residual phase noise measurement of the DUT. From the N5511A based residual phase noise measurement plot of the DUT the level of the spur is detected in dBc (not dBc/Hz) and its suppression is computed relative to the input level. From this suppression value the relative delay between the RF (DUT) path and LO path is computed using Equation 15.

The Stimulus phase noise suppression is measured by creating a phase noise spur at known relative level at offset frequency w_m at the Stimulus, and then measuring the corresponding PD output spur level in dBc during the residual phase noise measurement of the DUT using the N5511A PNTS. Then using Equation 15a/b we determine the value of n and the Delay τ. If n>>1 which could be the case in a practical system that is wired up initially, the relative Delay τ is then reduced by adding (or subtracting) suitable cable lengths in either the DUT (RF) or LO paths, whichever electrical length is less (more). Creating a known amount of spur is easy using the Keysight E8257D signal source as the Stimulus. We enable internal phase modulation with 0.1 rad peak with 100kHz sinewave modulation. This is within the small angle limit and the 0.1 rad peak setting shows up as -26 dBc spur at the 100 kHz offset. The relative spur level is computed as $20\log\left(\frac{\emptyset_0}{2}\right)$ where $\emptyset_0$ = 0.1 rad. A measured example is shown in Figure 14.

Equation 15a/b has been tested in the lab using a line stretcher in series with the DUT to vary the value of 'n' and measure the corresponding suppression of a 100 kHz PM spur of the E8257D Stimulus in a N5511A residual phase noise set up for 5.5 GHz. It follows the values of the Markers in Figure 9 rather accurately.

For a dual-channel cross-correlation system, the relative quadrature delay measurement can be done separately for each channel to determine their individual relative delays especially when 'n' starts becoming small such that very small delay adjustments are required compared to the total delay of the cable and DUT. Each channel may have an n that is slightly different from each other (e.g. difference of 2) due to cable length mismatches. With cross-correlation the estimated n will be an average for the two channels. A final cross-correlation measurement is done to determine the output Spur level and to confirm that the lowest 'n' value has been attained for dual-channel mode so that the Stimulus phase noise leakage at the PD output is well below the residual phase noise of the DUT. Typically, DUTs have flat or decreasing residual phase noise at far offsets (> 5 MHz). If the measured residual phase noise is increasing with offset frequency, it is caused by insufficient suppression of the Stimulus phase noise.

The N5511A PNTS, PNMI-K26 and E8257D (Stimulus and SSB Cal Osc) was used to measure the residual PN of a CMD274 low noise GaAs amplifier at 5.5 GHz. A -26 dBc spur was used at 10 kHz, 100 kHz and 1 MHz offsets in separate residual phase noise measurements that are plotted below. Dual-Channel Cross-Correlation was used. For the initial wiring, their respective suppression was measured as -30, -40 and -70 dB as shown in Figure 15, closely corresponding to a relative Quadrature Delay of 5 ns based on the theoretical model of Equation 15a/b as plotted below in Figure 14 with the 3 markers on the right side.

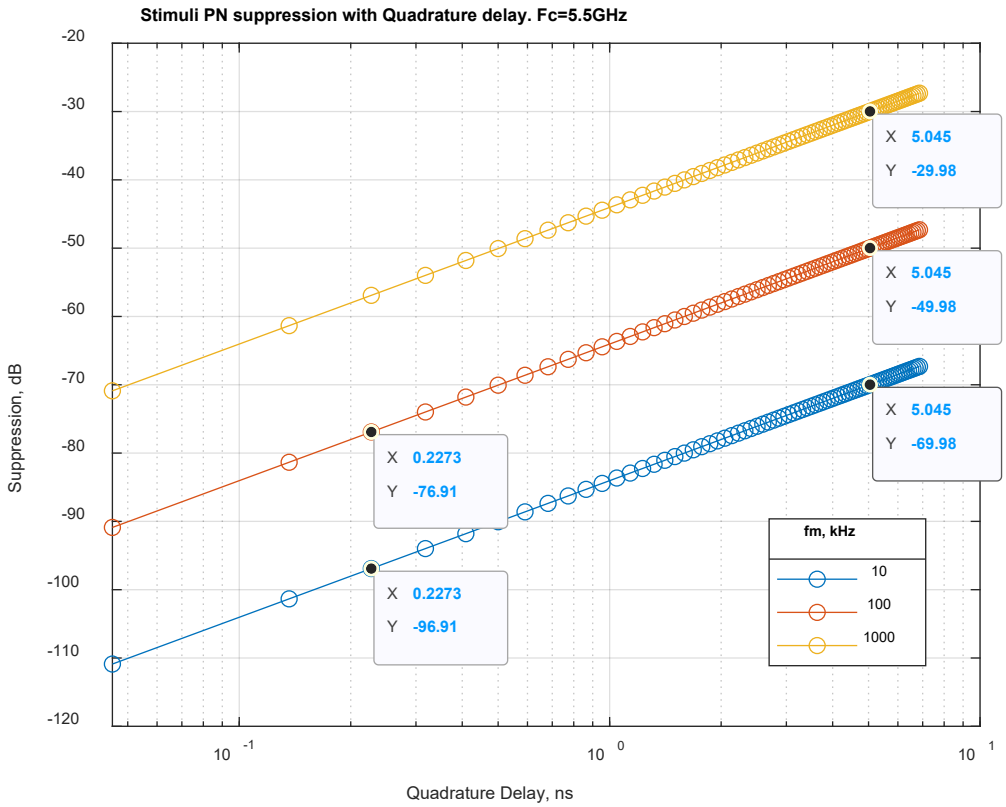
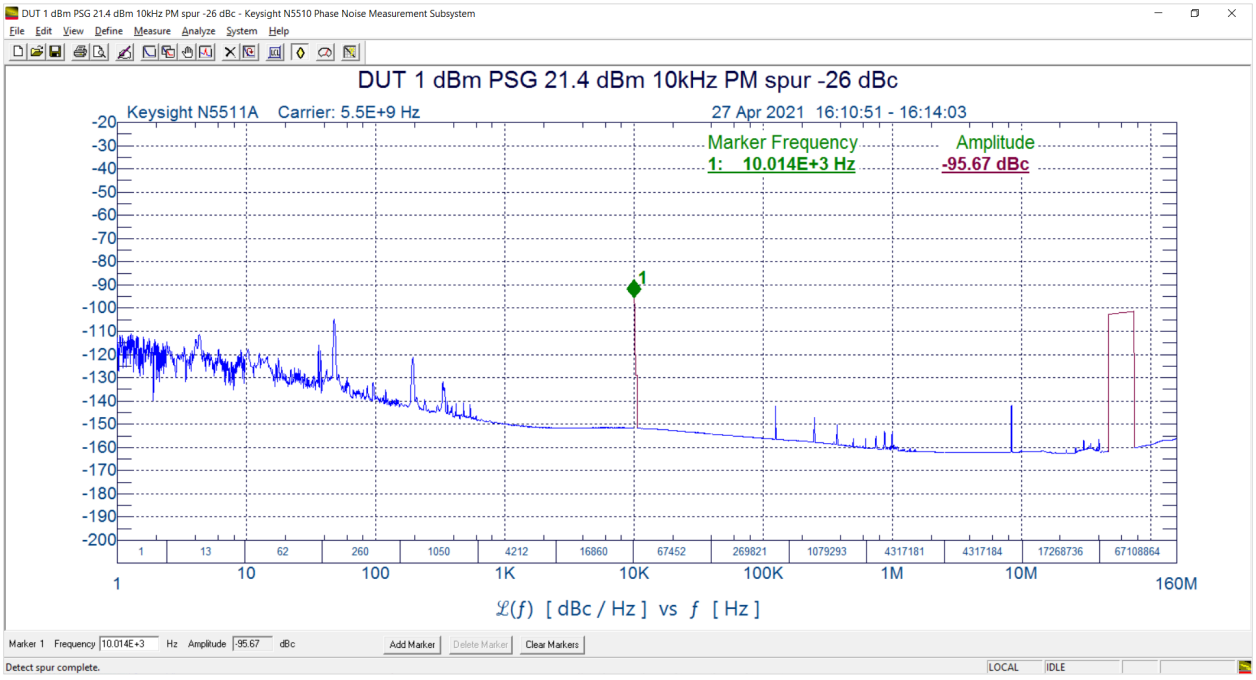


Figure 14. Suppression of Stimulus phase noise at PD output versus Quadrature Delays, for various Offset Frequencies of 10 kHz, 100 kHz and 1 MHz. Markers are places close to measured points. Those on the left side are after delay matching, with relative delay of 0.22 ns corresponding to n=5. The traces begin at n=1.



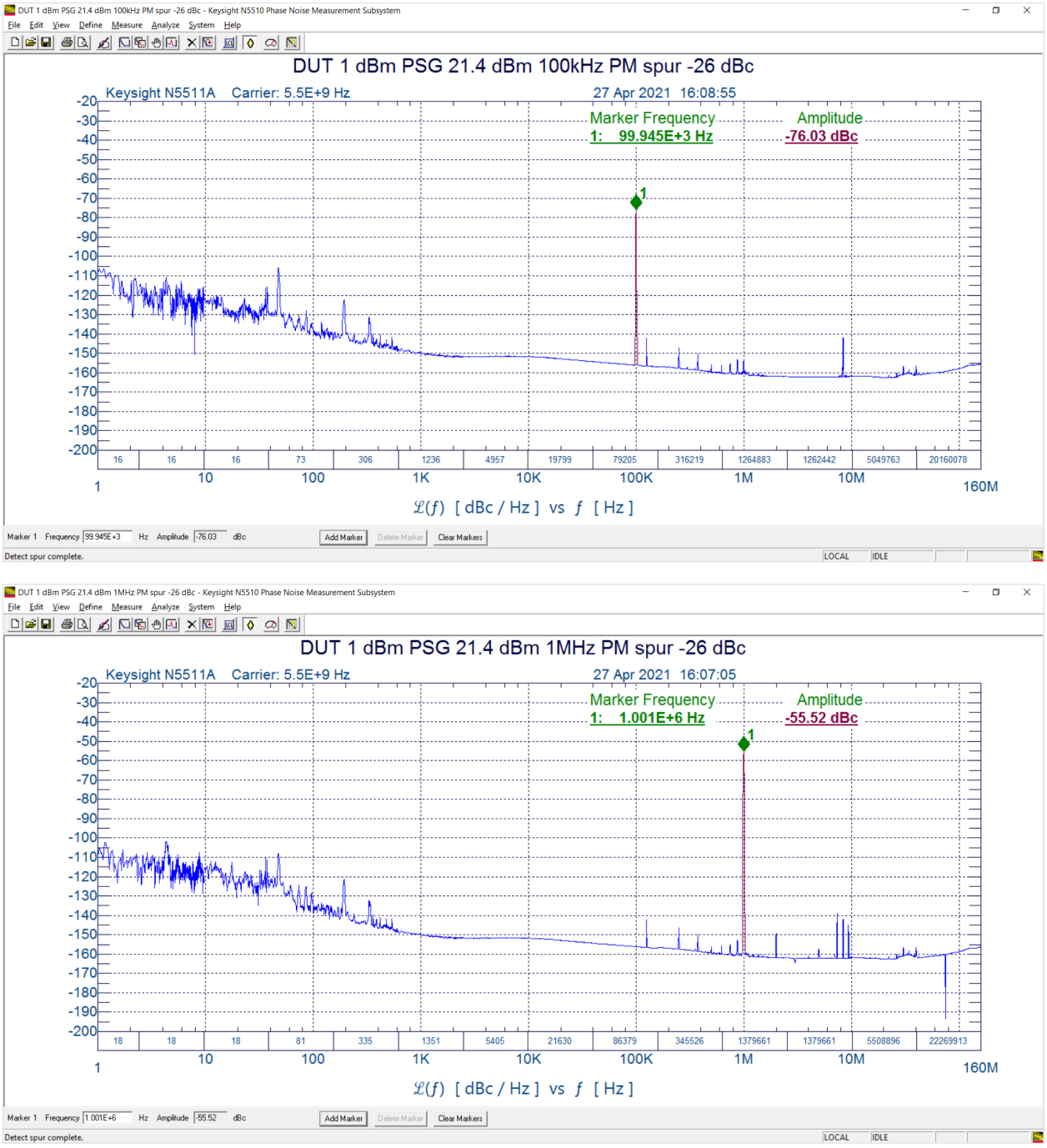


Figure 15. A -26 dBc spur was used at 10 kHz, 100 kHz and 1 MHz offsets in separate residual phase noise measurements that are plotted above for the initial wiring. Their respective suppression are -30, -40 and -70 dB based on the difference between the Marker1 value and -26 dB. This corresponds to approximately 5ns relative Quadrature Delay using the graph in Figure 14.

After matching the relative Quadrature Delays, the suppression of the spurs was improved by 30 dB as shown by the Dual-Channel measurements below in Figure 16 for 10 kHz and 100 kHz offsets. The suppression for 10 kHz and 100 kHz spurs are -77.7 and -100.6 dB respectively. The corresponding relative Quadrature Delay value is 0.22 ns as shown by the markers on the left side

in Figure 14 for the theoretical model. Individual channels of the N5511A were not optimized separately for the relative quadrature delays because the residual phase noise plot had already become ‘flat’ at the high frequency offsets (>50 MHz) indicating that it was dominated by the DUT and not the Stimulus.

When minimizing the relative Quadrature Delay, it is not necessary to obtain n=1 or identical values of n for both channels if the Stimulus phase noise leakage at the PD output is well below the DUT residual phase noise.

- Flat or decreasing residual phase noise plot at far offsets e.g. > 1 MHz indicates good suppression of Stimulus phase noise, ensuring its leakage is well below the DUT residual phase noise
- In addition to above, the ripple at the far offset should be very small. They potentially arise from the Stimulus leakage in presence of significant relative Quadrature Delay between RF and LO ports of the PD.
- Increasing residual phase noise plot at far offsets e.g. > 1 MHz indicates poor suppression of Stimulus phase noise, with its leakage level going above the DUT residual phase noise at the frequency where the plot tilts upwards.

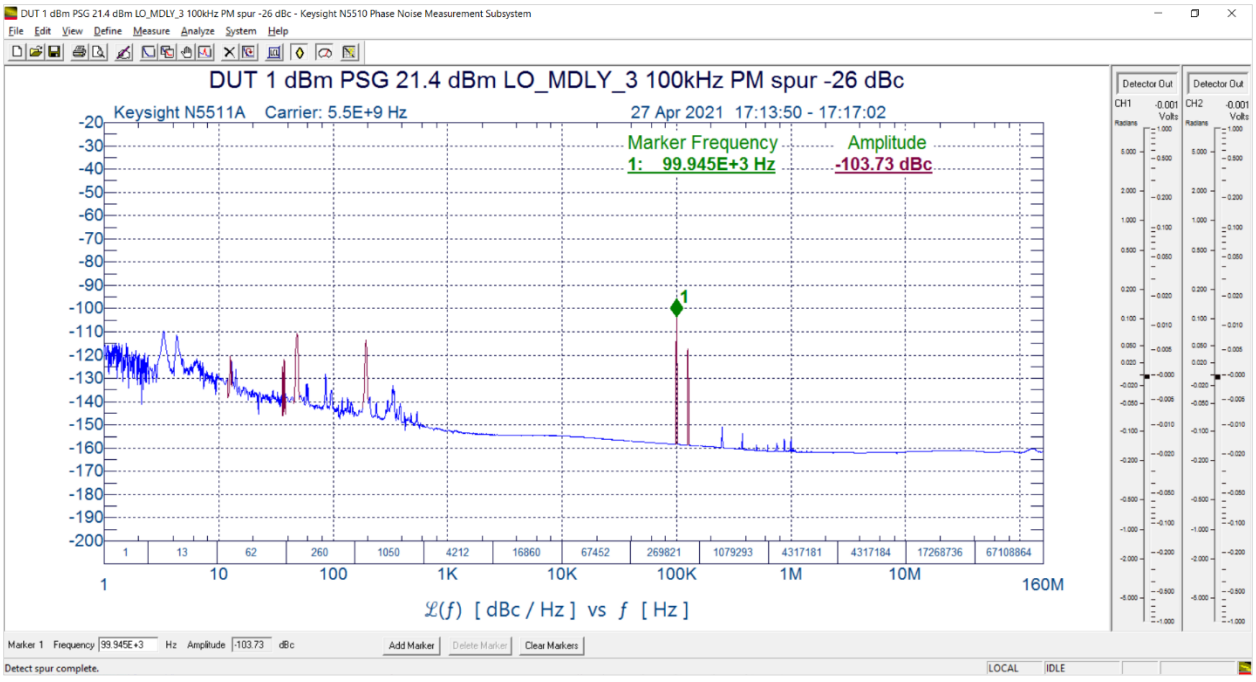
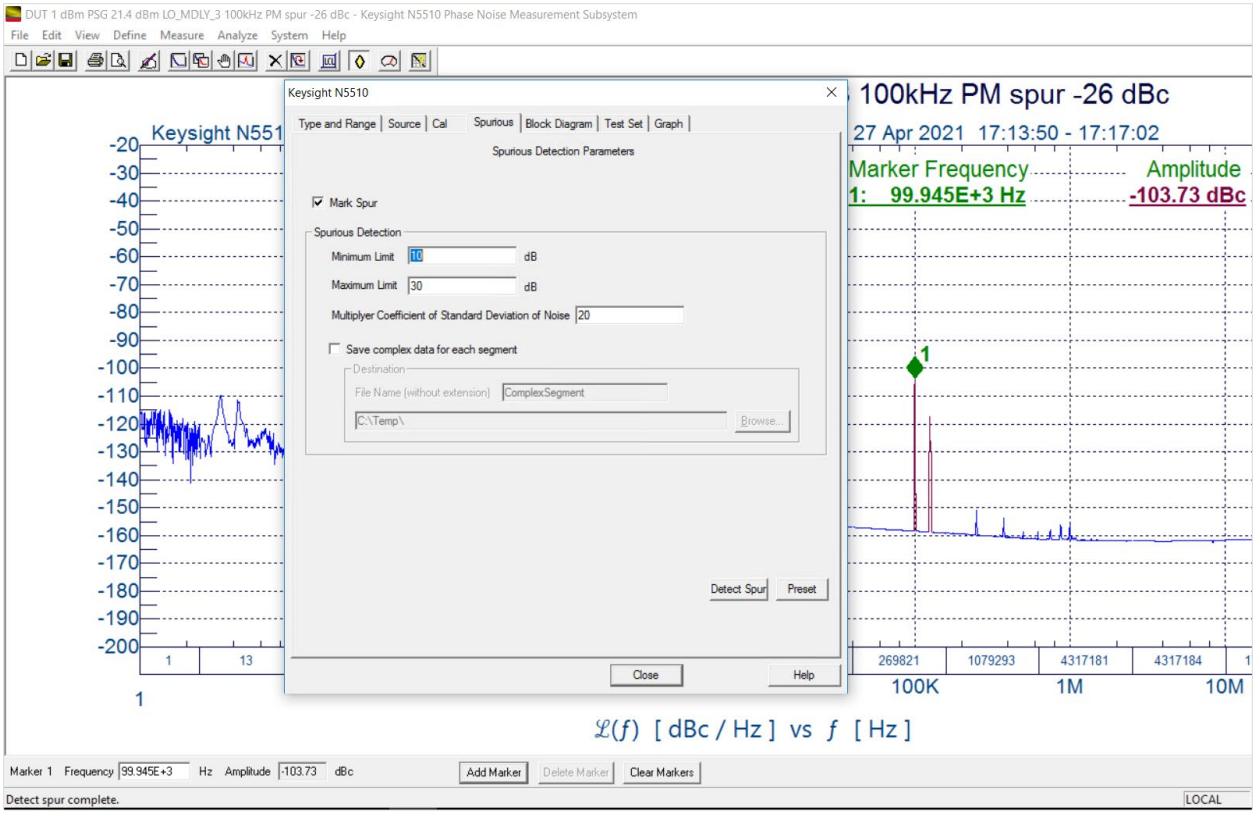




Figure 16. After reducing the relative delay to a lower value corresponding to n=3, the 100 kHz spur suppression is now -103.7-(-26) = -77.7 dB. The phase noise floor is flat at 160 MHz offset due to the small value of n. The suppression for the 10 kHz spur is -126.6-(-26) = -100.6 dB.

Enabling Spur Measurements in dBc

To enable Spur measurement in dBc, check the box ‘Mark Spur’ in the UI’s ‘Spurious’ tab, enable and select a Marker in the measurement window, and enter the required offset frequency for the Marker. The Marker will show up at the peak level of the Spur and the offset frequency entered will change to the nearest FFT-point. The set-up is shown below for -26 dBc spur at 100 kHz offset at the Stimulus, appearing at a suppressed level of -103.7 dBc at the PD output. To re-compute the trace after changing the UI parameters related to the Spur detection after a measurement has completed, click the Spur icon shown below with the blue arrow.



If the noise deviation is σ [dB] and the Peak height is Hpk [dB] (for both single & dual channel measurement), then the Peak is regarded as Spur if one of the following conditions is met:

Hpk > Maximum Limit

If Hpk in the range between Minimum and Maximum, Hpk > Coef * σ

We can change Min/Max and/or Coef parameters to change the sensitivity for detecting Spurs. Once detected, they change from dBc/Hz to dBc, along with a change in Peak trace color.

Note: The spurs are measured accurately by the N5511A. However, there may be some amount of PM/AM conversion in the DUT which could produce small errors due to limited AM suppression of the PD. It should not have a significant impact on the computed value of 'n' related to the Quadrature Delay. Using the AM Suppression technique during the Quadrature Phase adjustment as described later, will alleviate this issue.

Peaks and Null Method

In another method, we measure the residual phase noise using Stimulus with a high wideband phase noise to determine the relative Quadrature Delay and the corresponding odd value of n. Optionally we can use a wideband OFDM signal or generate wideband white phase noise at the Stimulus using IQ modulation. We look at the Peak and Nulls of Eqn 15a/b in the residual phase noise measurement plot and determine the value of odd 'n' and the corresponding relative quadrature delay. The steps are as follows:

The odd value of 'n' is determined by looking at the 1st 'null' frequency offset or 1st 'peak' frequency offset in the residual phase noise plot which is measured at the carrier frequency Fc:

1st null is at offset Fnull = 4*Fc/n, so n = 4*Fc/Fnull (17)

1st peak is at offset Fpeak = 2*Fc/n, so n = 2*Fc/Fpeak (18)

From n, the corresponding relative Quadrature Delay is $\tau = \mathrm{n}\frac{\pi}{2w_c} = \frac{n}{4Fc}$ (19)

Example

In the Residual Phase Noise measurement shown below (black trace) at Fc=500 MHz carrier using a relatively high phase noise Stimulus of the E8257D, there is a 1st null at Fnull=100 MHz offset (approximately), indicating that n= 4Fc/Fnull = 20 using Eqn (17). From Eqn (1), $\tau = \mathrm{n}\frac{\pi}{2w_c} = \frac{n}{4Fc} = 10ns$ approximately. The actual odd value of n is either 19 or 21 that is closest to 20. The approximate value of 10ns has been confirmed from the Network Analyzer S21_Delay measurements shown below in Figure 18 for the RF and LO path delays from the Stimulus input port of the PNMI-K06. At the 500 MHz marker point the time difference between the two traces is roughly 10 ns.

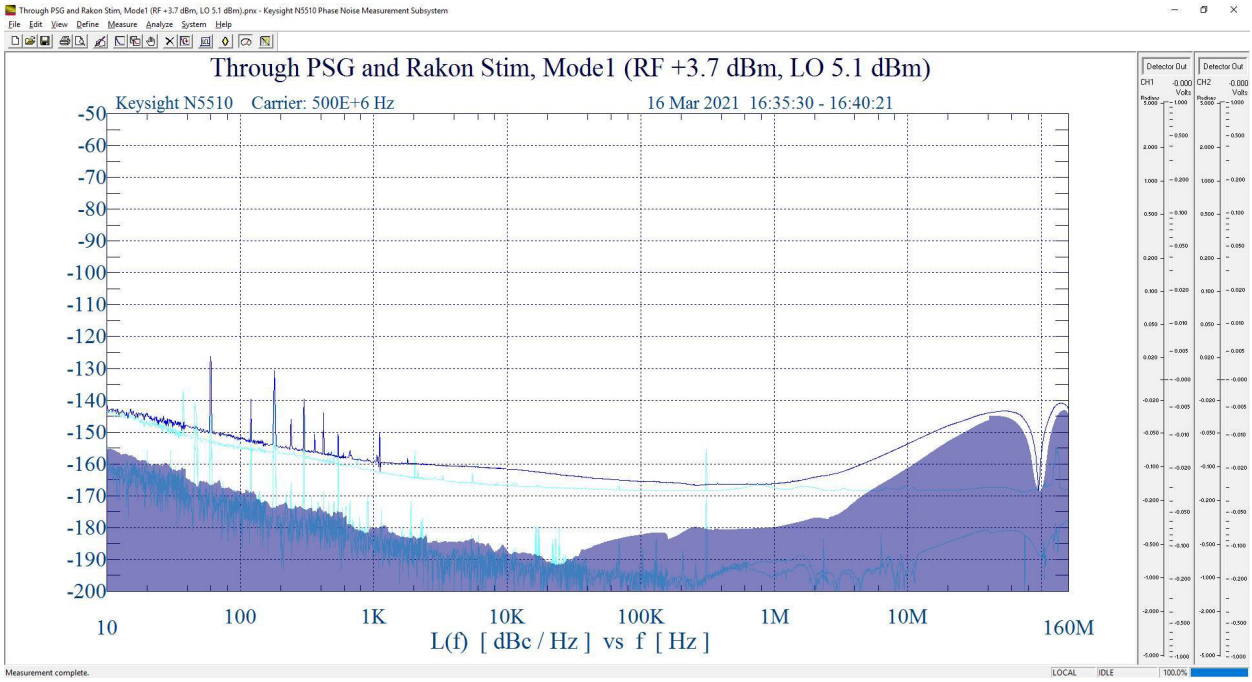


Figure 17. Peaks and Nulls in residual phase noise with high wide band Stimulus phase noise. In the black trace using the E8257D, the 1st null at approximately 100 MHz shows a relative delay of 10ns. The lower blue trace is the residual phase noise floor using an ultra-low phase noise 500 MHz Rakon OCVCSO as Stimulus. The middle blue trace is the DUT residual phase noise using Rakon source as Stimulus.

The 500 MHz Residual PN plot shown above in Figure 17 used an ultra-low phase noise 500 MHz Rakon OCVCSO as Stimulus. The bottom blue trace is the measured residual phase noise floor with the Rakon stimulus, middle blue trace is the Rakon Stimulus + DUT (Amplifier), while the top black trace is a PSG E8257D Stimulus residual noise floor without any DUT i.e., a through

connection for the DUT. The relative delay mismatch is 10ns for the black trace. The Rakon based noise floor also starts increasing due to the delay mismatch, but it is well below the DUT (Amplifier) residual phase noise.

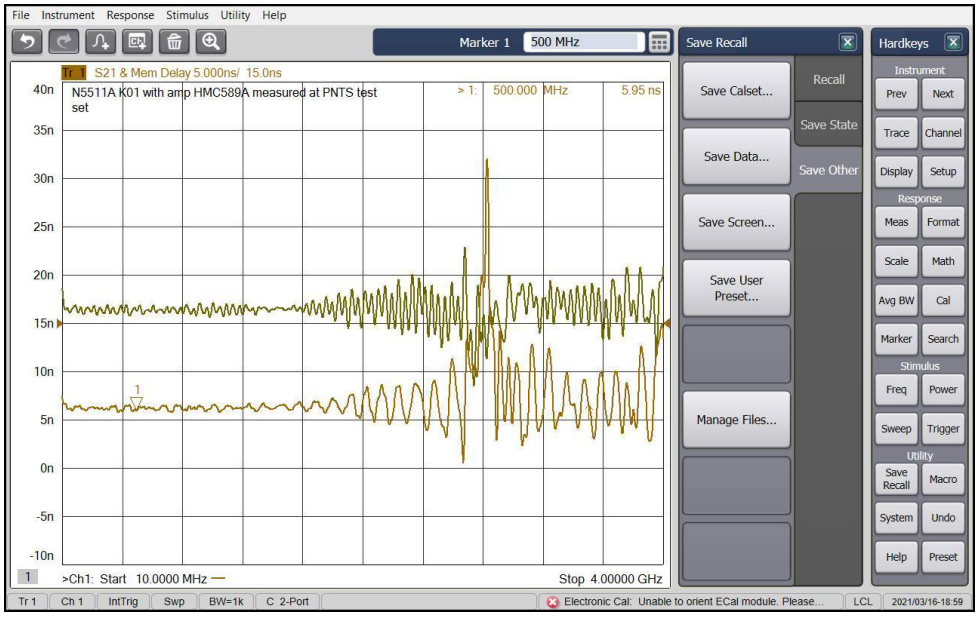


Figure 18. Relative delay measurements using Network Analyzer S21_delay. A more reliable way especially for a bandpass response of the PNMI is to use the "Enhanced Time Domain Analysis" with TDR option in the Network Analyzer, instead of using S21_delay format or enabling time domain transform of complex (linear) S21. The "Enhanced Time Domain Analysis" uses stepped CW Frequency technique to resolve frequency domain response into time domain and for a conforming system for delay measurement, the first frequency point of the frequency sweep must be same as the frequency step size.

Lowest Values of 'n' for N5511A

As max offset is limited to 160 MHz in the N5511A and we need to have at least 1 peak, so minimum 'n' that can be detected is the smallest odd integer greater or equal to 2Fc/160 MHz.

- Fc=500 MHz, then minimum odd n that can be detected is >= 2*500 MHz/160 MHz i.e. 7
- Fc=5 GHz, then minimum odd n that can be detected is >= 2*5 GHz/160 MHz i.e. 63

For the 1st Null frequency at 160 MHz offset:

- Fc=500 MHz, then minimum odd n that can be detected is >= 4*500 MHz/160 MHz i.e. 13
- Fc=5 GHz, then minimum odd n that can be detected is >= 4*5 GHz/160 MHz = 125

The above examples are plotted below in Figure 19 and 20. Figure 19 below shows some examples of Peak and Null plots for Fc=500 MHz, n=1,2,3,11 (corresponding Quadrature Delays shown in the legend).

If no null or peak occurs, then we use the previous method i.e. measuring the Spur suppression and looking up the corresponding relative delay or 'n.' However, with reasonable cable lengths it is easy to get peaks and nulls especially when the relative quadrature delay has not been minimized.

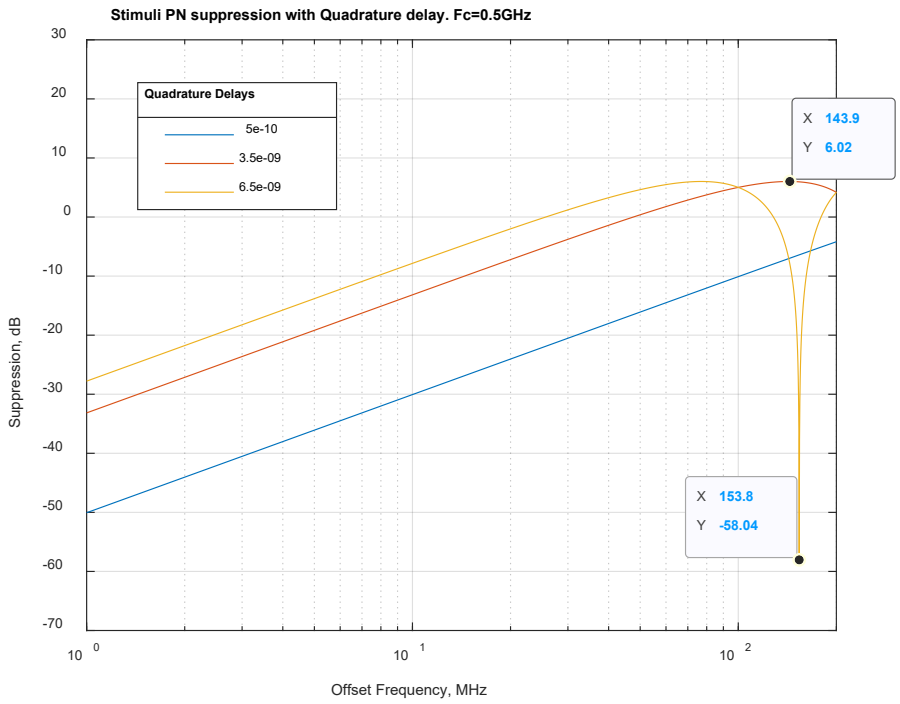


Figure 19. Suppression of Stimulus phase noise at PD output versus Offset Frequency, for relative quadrature delays corresponding to n = 1, 7, 13 at 500 MHz carrier

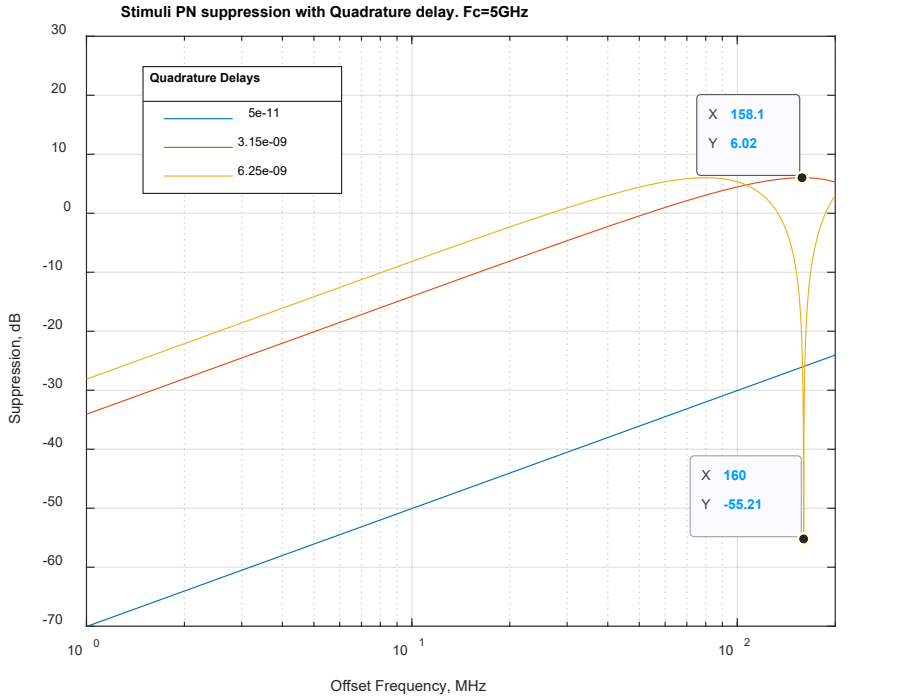


Figure 20. Suppression of Stimulus phase noise at PD output versus Offset Frequency, for relative quadrature delays corresponding to n = 1, 63, 125 at 5 GHz carrier.

AM Suppression Improvement

As explained earlier, improvement of Stimulus and DUT AM Noise suppression at PD Output helps in lowering the residual phase noise measurement floor. This is especially important in situations when the Stimulus or DUT have AM noise that is approximately 6 dB or more higher than the residual phase noise of the DUT in which case a measured 16 dB AM Rejection of the PD would be insufficient and result in significant error in the measurement. The 2nd technique improves the AM suppression of the PD beyond the typical measured value of 16 dB and is described below.

Using the Quadrature Monitoring Meters of the N5511A, initially the DC voltage at the PD output is monitored in presence of the RF and LO signals and the approximate Quadrature Phase is established using DC-nulling. Due to presence of DC Error arising from the mixer 2nd order nonlinearities (IP2) and self-detection, the quadrature setting is not optimal, and the AM will not be suppressed much. To obtain the true Quadrature Phase setting, AM modulation of the Stimulus is turned on e.g. using a sinusoidal signal at a baseband angular modulation freq w_m. The demodulated tone spectrum is monitored at the PD output "Monitor port" of N5511A at the baseband freq w_m and the relative Delay (line stretcher or phase shifter) is adjusted slightly away from the DC-Null point to minimize the tone level. This achieves the optimum AM suppression and the true Quadrature phase setting. After optimizing the AM suppression, the AM modulation is turned OFF, and the regular Residual Phase Noise measurements are done. For the 100 kHz AM spur we may be able to improve the PD suppression by 15-30 dB at 5.5 GHz compared to the DC-null setting.

The improvement in the Residual Phase Noise floor based on this technique is shown in Figure 22, while the improvement in measurement of the DUT residual phase noise is shown in Figure 23, using the E8257D as the Stimulus.

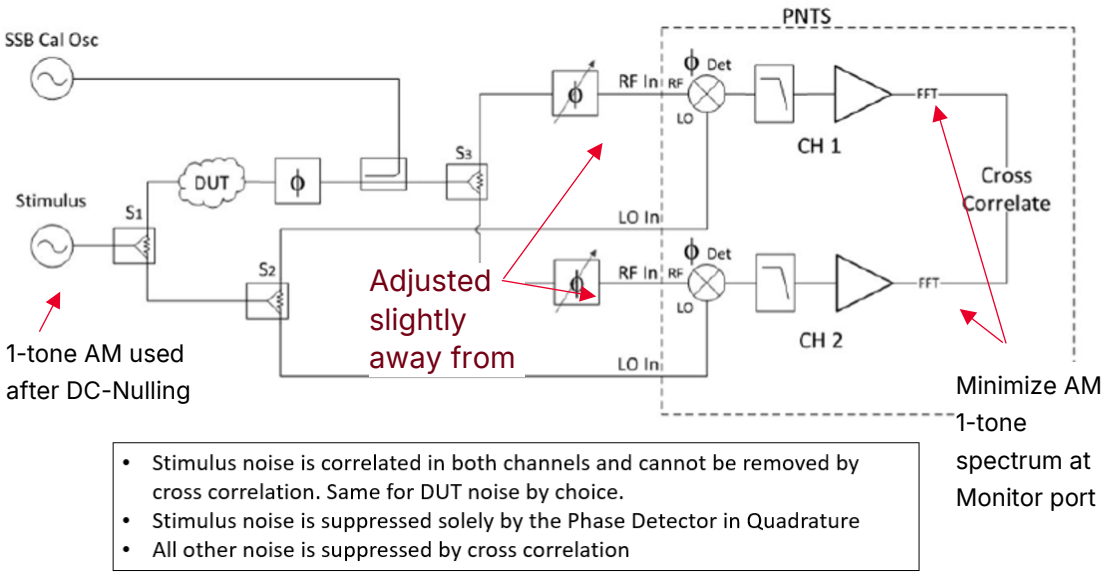


Figure 21. Dual-channel cross correlation using two References. 1% AM at 100 kHz modulation frequency is enabled at the Stimulus (E8257D) corresponding to 20*log(0.1)-6 = -46 dBc AM spur. This spur is monitored at the PD output for each channel at the Monitor port using a Spectrum Analyzer and its level minimized by adjusting the respective phase shifter slightly from the DC-null setting.

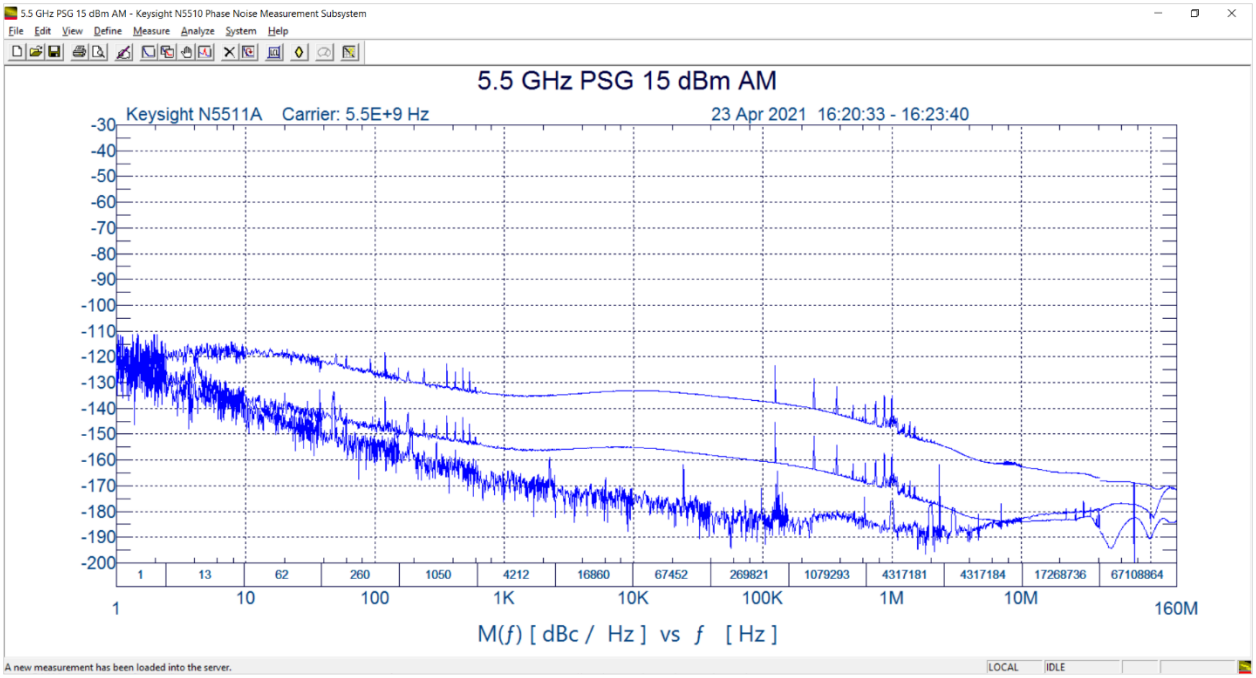


Figure 22. Improving the AM Rejection in N5511A for Residual Phase Noise Measurement. Plot shows the Residual Phase Noise Floor of the system before and after optimization for AM Suppression. Overlaid is the AM Noise plot of the E8257D Stimulus that largely determines the non-optimized Residual Phase Noise floor.

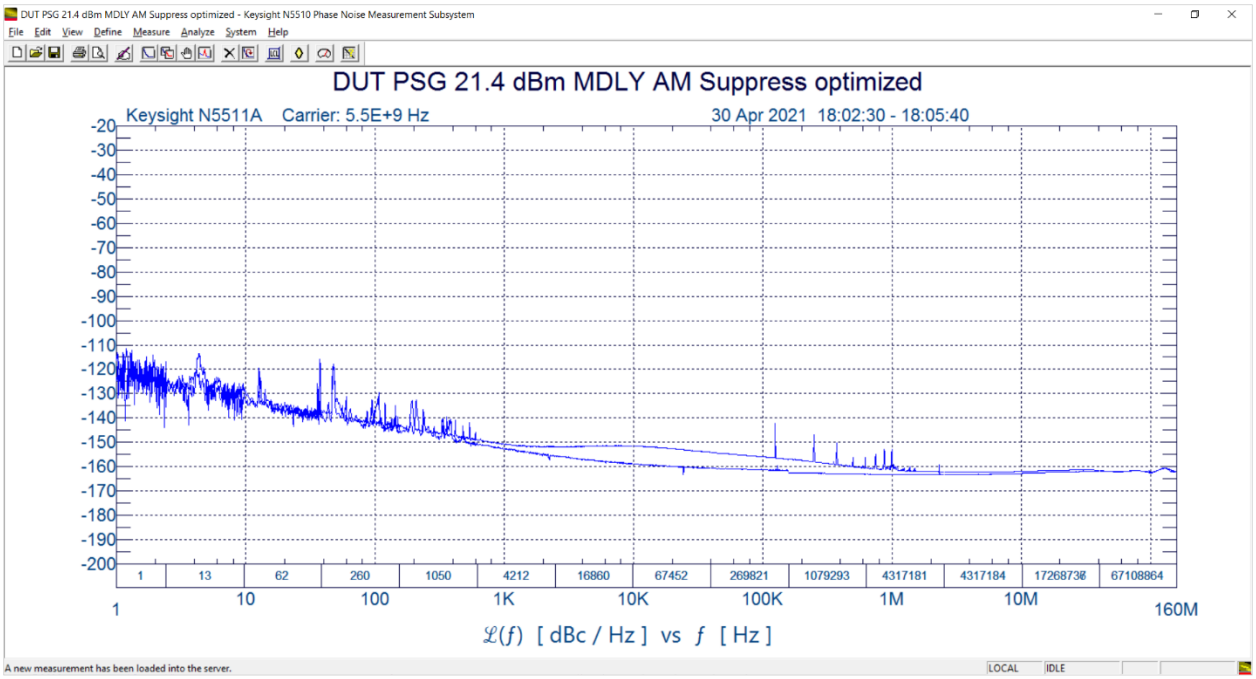


Figure 23. Residual Phase Noise of the DUT before and after optimization for the PD AM Suppression. It shows that the E8257D with a high AM noise can still be used as the Stimulus if we optimize the system for AM suppression.

Enabling Out-of-Lock Condition for Improved AM Suppression

When the Phase Shifters are adjusted slightly from their DC-null setting to minimize the AM Spur at the PD output, there will be some amount of DC error and the f/w will show an Out-of-Lock condition and not proceed further. To overcome this, we have to check the box 'Ignore out-of-lock conditions' shown below in the UI.

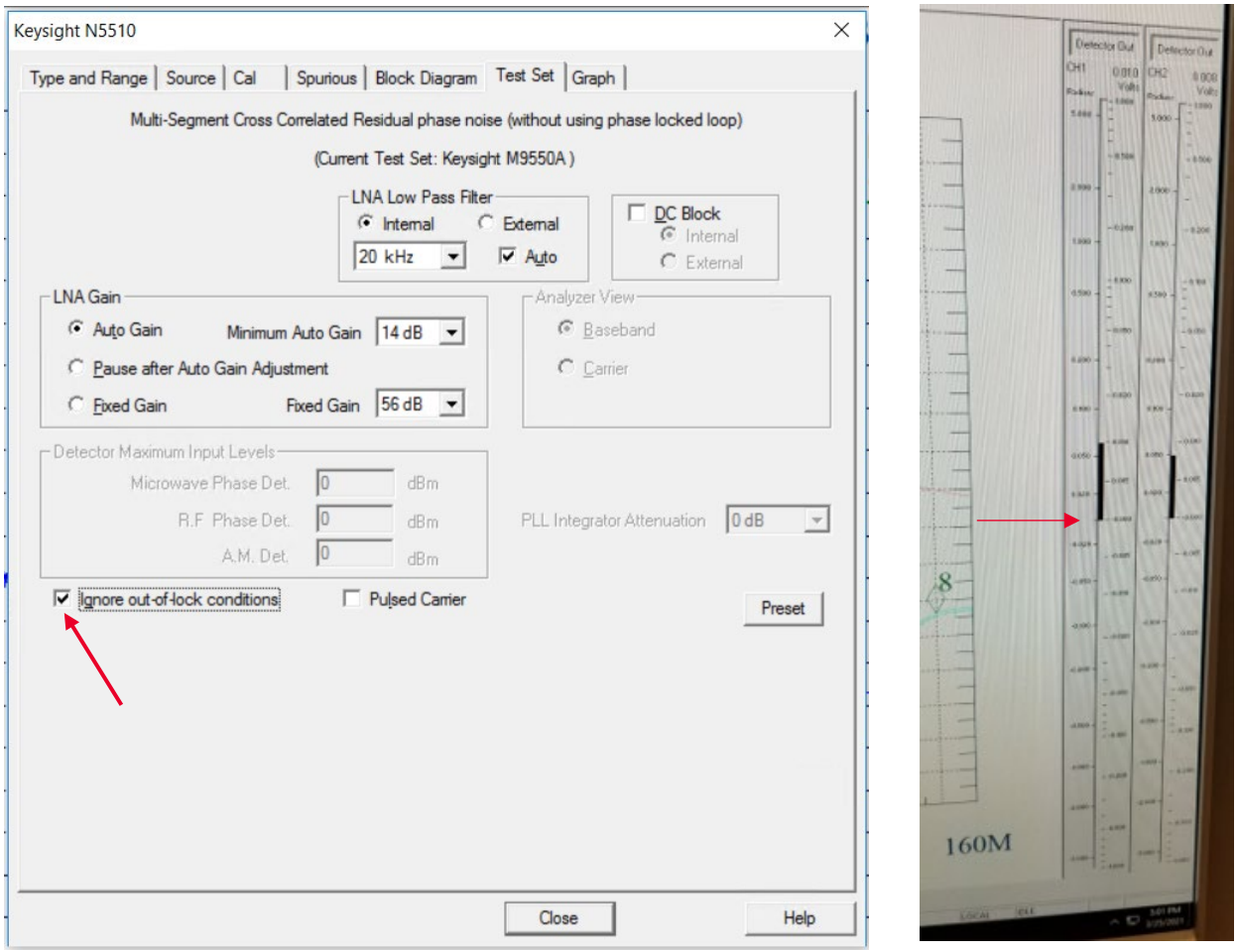


Figure 24. UI setting to allow DC Error (out-of-lock condition) as seen in the Quadrature Monitoring meters on the right. The red arrow shows 0 Volt point. The dark bars in the meters show small DC errors at the optimum AM suppression setting. This DC error is small compared to the full-scale Peak DC error when the RF and LO relative phase is 0°.

Note: The AM spur minimization at the N5511A baseband Monitor output may reach a floor due to small amount of AM/PM conversion in the DUT. It should typically not have a major impact on Stimulus AM suppression optimization. Referring to Figure. 3, the polarity of the AM conversion gain for relative phase between the RF and LO inputs, shown by "DC Volts" trace, is negative for relative phase greater than 90 degrees and positive for relative phase less than 90 degrees. During cross-correlation using the dual-channel absolute phase noise measurement system, it is

desirable to have the same polarity for each of the two channels to avoid a phenomenon known as "cross-spectral collapse," which may occur when the AM noise partially cancels the PM noise during cross-correlation and gives a wrong phase noise measurement at frequency offsets where the AM conversion gain polarities are opposite in the two channels. During adjustment of the AM null at the output of mixer 210 of each channel, the adjustment is done in the same direction for each channel, and the adjustment is stopped immediately after the minimum tone level from the AM modulation starts to increase again from the minimum value. This ensures that the relative phase between RF and LO inputs for each channel are either both slightly greater than 90 degrees or slightly less than 90 degrees.

Appendix

Impact of LO and RF Self-Detection on the Stimulus and DUT AM Noise Suppression

Leakage between the LO and RF ports creates a DC error due to self-detection in the PD. DC error also arises from the 2nd (even) order nonlinearity in the PD. These lead to an error in the Quadrature phase adjust that is based on DC-nulling at the PD output done by adjusting the delay of a phase shifter line stretcher. Due to this phase error, only a small amount of additional Stimulus phase noise leaks at the PD output as shown by the equation in the figure below, and as derived in the Appendix. However, the major impact of this phase error is reduced suppression of the Stimulus AM noise that appears at the PD output, as depicted in the figure below.

With a set-point phase error of θ_e there is an error τ_e in the phase shifter delay given by

$w_c(\tau + \tau_e) = \frac{\pi}{2} + \theta_e$ where $w_c\tau_e = \theta_e$

Replacing τ with $\tau' = (\tau + \tau_e)$ and substituting $w_c\tau_e$ with θ_e

$r(t) = \frac{1}{2}\cos(w_c\tau + \theta_e) + \frac{\emptyset_0}{4}[-\cos(w_c\tau + \theta_e + w_m t - w_m\tau') + \cos(w_c\tau + \theta_e - w_m t + w_m\tau') - \cos(w_c\tau + \theta_e - w_m t) + \cos(w_c\tau + \theta_e + w_m t)]$

Putting $w_c\tau = \frac{\pi}{2}$ and using $\cos\left(\frac{\pi}{2} - \theta\right) = \sin(\theta)$

$r(t) = \frac{\emptyset_0}{4}[-\sin(-w_m t + w_m\tau' - \theta_e) + \sin(w_m t - w_m\tau' - \theta_e) - \sin(w_m t - \theta_e) + \sin(-w_m t - \theta_e)]$

$r(t) = \frac{\emptyset_0}{4}[\sin(w_m t - w_m\tau' + \theta_e) + \sin(w_m t - w_m\tau' - \theta_e) - \sin(w_m t - \theta_e) - \sin(w_m t + \theta_e)]$

$r(t) = \frac{\emptyset_0}{2}[\sin(w_m t - w_m\tau')\cos(\theta_e) - \sin(w_m t)\cos(\theta_e)]$

$r(t) = \frac{\emptyset_0}{2}\cos(\theta_e)[\sin(w_m t - w_m\tau') - \sin(w_m t)]$

$r(t) = \emptyset_0\cos(\theta_e)\cos(w_m t - w_m\tau'/2)\sin(w_m\tau'/2)$

$r(t) = \emptyset_0\cos(\theta_e)\sin(w_m(\tau + \tau_e)/2)\cos(w_m t - w_m\tau'/2)$

Putting $\tau_e = \frac{\theta_e}{w_c}$ and $\tau = \frac{\pi}{2w_c}$

$r(t) = \emptyset_0\cos(\theta_e)\sin\left((\frac{\pi}{2} + \theta_e)\frac{w_m}{2w_c}\right)\cos(w_m t - w_m\tau'/2)$

The above equation shows that the PD gain reduces by the $\cos(\theta_e)$ term which is negligible for small θ_e.

Phase Detector Residual PN Suppression

The Suppression is the ratio of the power of $r(t)$ and $s(t)$:

Suppression = $[\frac{\emptyset_0^2}{2}\sin^2((\frac{\pi}{2}+\theta_e)\frac{w_m}{2w_c})]/[\frac{\emptyset_0^2}{8}]$

Suppression = $4\sin^2([\frac{\pi}{2}+\theta_e]\frac{w_m}{2w_c})$

For phase shifts of 90, 270, 450, … degrees i.e. for Odd n, it can be shown that

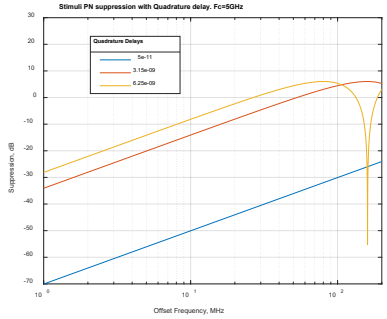
Suppression = $4\sin^2([n\frac{\pi}{2}+\theta_e]\frac{w_m}{2w_c})$

The $\cos(\theta_e)$ term does not appear in the above Suppression equation as it is common to both the wanted DUT phase noise gain and the Stimulus phase noise leakage gain of the PD, and cancels when taking the ratio when computing the Suppression.

Matlab Scripts

Stimulus Phase Noise Rejection Graph

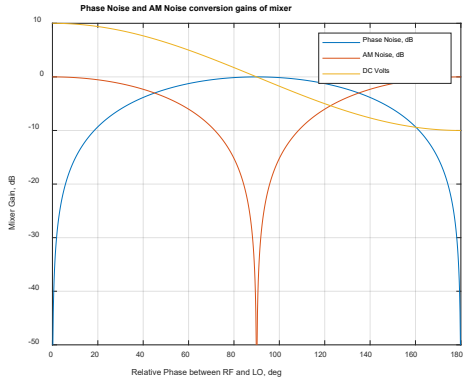
```
%% Stimuli PN Rejection as function of Quadrature Delay (N)
% Author: Rishi Mohindra, Keysight Technologies
clear all;close all;clc;
Fc=5000e6;
N=[1,63,125];
Td0=1/Fc/4;
Td=Td0*N;
fm=10.^(6:0.001:10)';
for n=1:length(Td)
    Lfm1(:,n)=4*(sin(pi*fm*Td(n))).^2;
end
figure(1);
semilogx(fm/1e6,db(Lfm1.^0.5));grid on;
lgd=legend(num2str(Td'));title(lgd,'Quadrature Delays');
xlabel('Offset Frequency, MHz');
ylabel('Suppression, dB');title(['Stimuli PN suppression with
Quadrature delay. Fc=',num2str(Fc/1e9),'GHz']);
axis([-inf 200 -70 30]);
```



AM and PM Gain of Phase Detector

```
%% AM and PM gain of Phase detector
% Author: Rishi Mohindra, Keysight Technologies
ph=[0:pi/1000:pi]';
phDeg=ph*180/pi;
PN_gain=db(sin(ph));
AM_gain=db(cos(ph));
plot(phDeg,PN_gain,phDeg,AM_gain);
title('Phase Noise and AM Noise conversion gains of mixer');
ylabel('Mixer Gain, dB');xlabel('Relative Phase between RF and LO,
deg');
legend('Phase Noise','AM Noise');
axis([-inf inf -50 0]);grid on;

figure(2);
plot(phDeg,PN_gain,phDeg,AM_gain,phDeg,10*cos(ph));
title('Phase Noise and AM Noise conversion gains of mixer');
ylabel('Mixer Gain, dB');xlabel('Relative Phase between RF and LO,
deg');
legend('Phase Noise, dB','AM Noise, dB','DC Volts');
axis([-inf inf -50 10]);grid on;
```



Residual Phase Noise Floor for E8257D at 5.5 GHz

```
%% Residual Phase Noise Floor with Quadrature delay
% Author: Rishi Mohindra, Keysight Technologies
Fc=5.5e9;
fm=10.^(0:0.01:9)';
N=[1,3,11];
F=Fc+fm;
Td0=1/Fc/4;
Td=Td0*N;
% Ph=2*pi*F*Td-pi/2*N;
% PdRej=db(sin(Ph));

Ph=pi*fm*Td;
PdRej=db(2*sin(Ph));

figure(3);semilogx(fm,PdRej);grid on;
lgd=legend(num2str(Td'));title(lgd,'Quadrature Delays');
xlabel('Offset Frequency, Hz');
ylabel('Suppression, dB');title(['Stimulus PN suppression vs Offset
Freq. Fc=',num2str(Fc/1e9),'GHz']);
axis([-inf inf -200 0]);

PN=[ % 5.5 GHz measured PSG Phase Noise
    1       -55;
    3.3     -75;
    10      -85;
    33      -91;
    1e2     -96 ;
    330     -111;
    1e3     -122;
    2e3     -130;
    5e3     -131;
    10e3    -132.5;
    33e3    -132;
    100e3   -130.5;
    200e3   -128;
    330e3   -130;
    1e6     -145;
    3.3e6   -157;
    10e6    -162.5;
    33.3e6  -164;
    100e6   -165;
    160e6   -165;
    1e9     -165];
AN=[ % 5.5 GHz measured PSG AM Noise
    1       -122;
    3.3     -119;
    10      -119;
    33      -121;
    1e2     -126 ;
```

```
    330     -130;
    1e3     -135;
    2e3     -136;
    5e3     -134;
    10e3    -132;
    33e3    -135;
    100e3   -138;
    330e3   -141
    1e6     -148;
    3.3e6   -155;
    10e6    -162;
    33.3e6  -167;
    100e6   -170;
    160e6   -171;
    1e9     -171];
PD_AM_Rej=-30; %dB

PNintp=interp1(PN(:,1),PN(:,2),fm,'cubic');
PNnet=PNintp+PdRej;

ANintp=interp1(AN(:,1),AN(:,2),fm,'cubic');
ANnet=ANintp+PD_AM_Rej;

PowerDUT=0; % dBm
ThermalNoiseFloor=-177-PowerDUT; % dBc/Hz
ThermalNoiseFloorLine=ones(length(fm),1)*ThermalNoiseFloor;
figure(4);semilogx(fm,PNintp,fm,ANintp,fm,ANnet,fm,PNnet,fm,ThermalNoi
seFloorLine);grid on; xlabel('Hz, offset');ylabel('dBc/Hz');
title(['Residual Phase Noise Floor with Quadrature delay.
Fc=',num2str(Fc/1e9),'GHz']);
axis([-inf 200e6 -270 -50]);yticks((-270:10:-50)');
lgd=legend(['PSG PN    ';'PSG AM    ';'AM Floor
';num2str(Td','%10.4e\n');'Thermal   ']);title(lgd,'Quadrature
Delays');
```

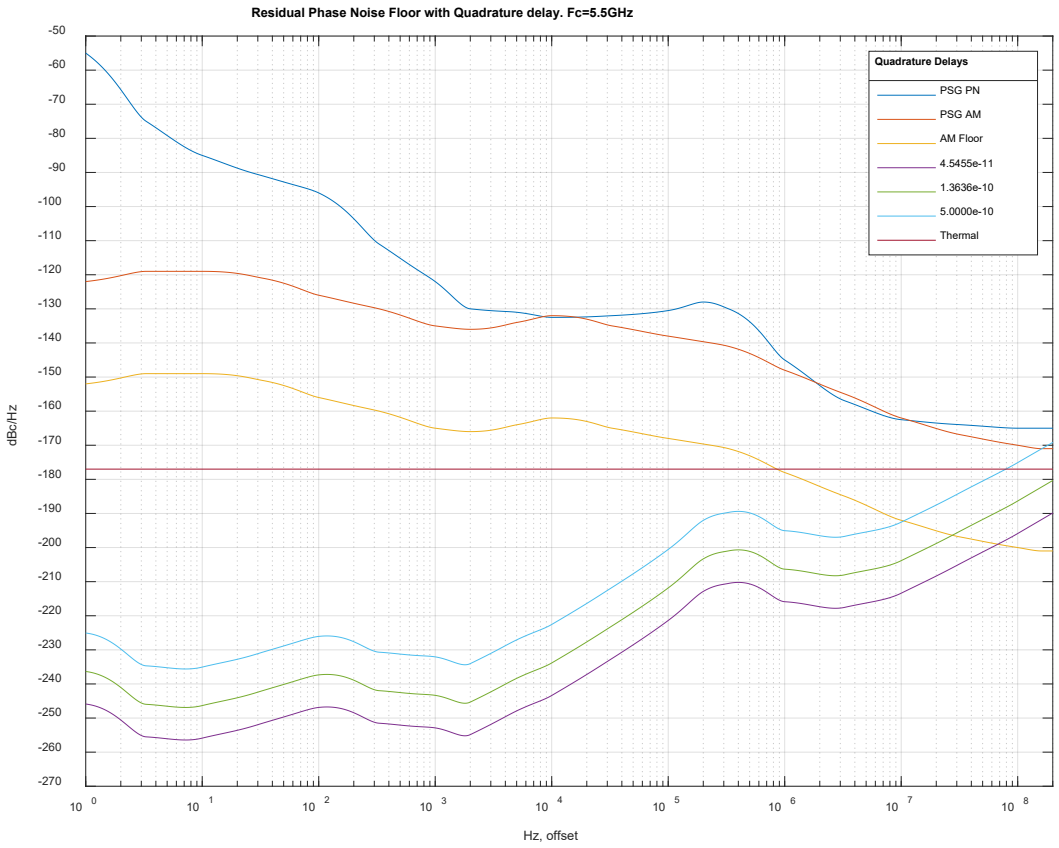


Figure 25. Residual Phase Noise Floors from AM and PM noise of E8257D at 5.5 GHz. The Matlab script plots the above figure for E8257D measured phase noise and AM noise at 5.5 GHz. The AM Floor is shown 30 dB below the measured AM noise, based on AM Suppression optimization. The lower 3 traces are the floors for n = 1, 3, 11 quadrature delay multipliers. The user can change the PN and AN noise data in the script along with other variables to model different sources.

Stimulus Phase Noise and AM Noise Suppression: Verification with ADS Simulation

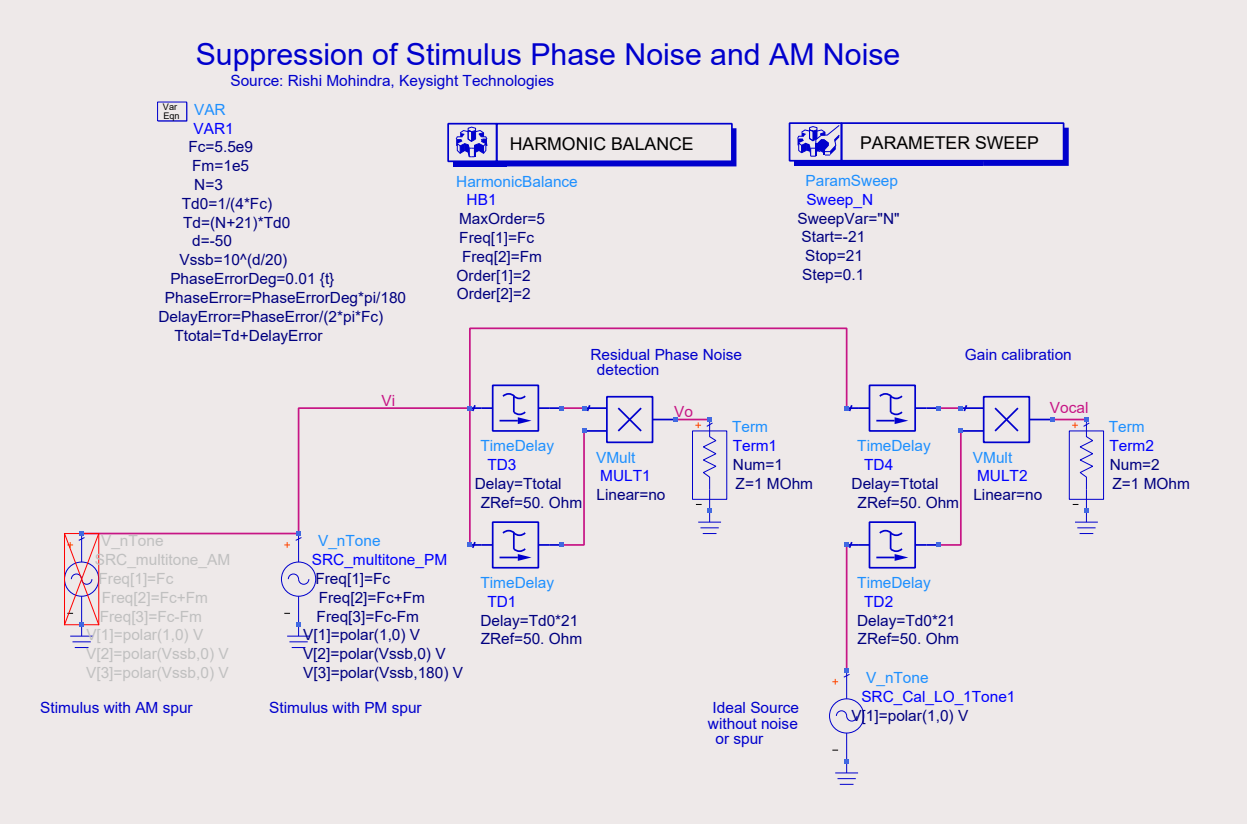
The AM and PM suppression of the Stimulus Noise by the PD is demonstrated below for a Residual Phase Noise measurement set up. In this Harmonic Balance simulation, two circuits are used:

- For the Residual Phase noise and AM noise suppression by the mixer, with output signal Vo and Stimulus input signal Vi. The RF and LO signals are from the Stimulus with a relative delay set by the variable N (multiplier for Quadrature Delay Td0 = 1/4Fc). As discussed earlier, N is an odd integer for Quadrature Phase.
- For calibrating the mixer conversion gains for AM and PM spurs. The output signal for this mixer is Vocal. The RF is the Stimulus with PM or AM Spur while the LO is an ideal signal. The RF-LO relative delay is identical to the first circuit.

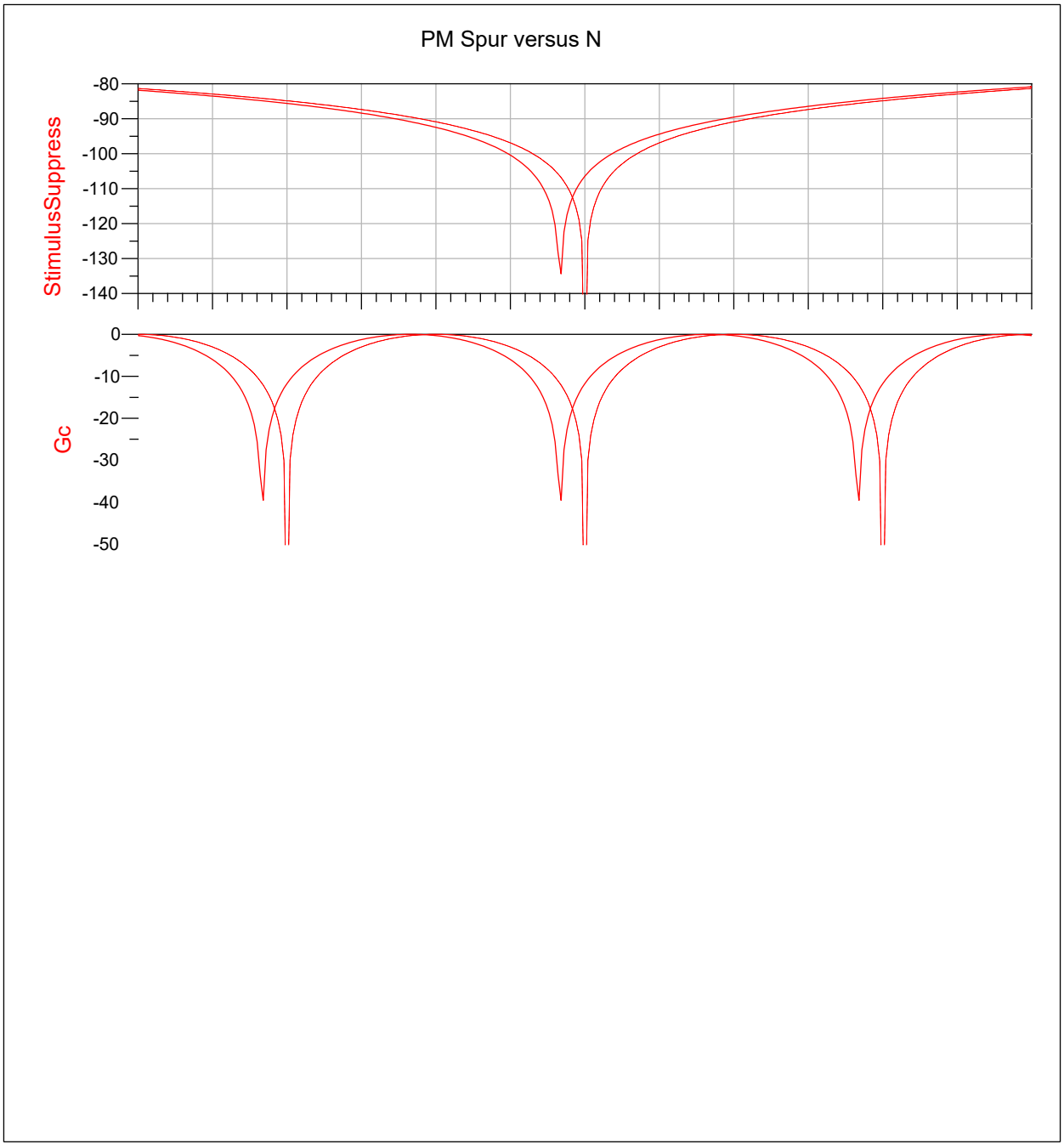
The Stimulus can be selected with either AM or PM spur by enabling one of the two signal sources as shown. Each of these two signal sources comprise of the Carrier along with two sideband spurs at relative level of 'd' dB each. The only difference between the two Stimulus sources is the 180° phase for the lower sideband PM spur.

The PD mixer is an ideal voltage multiplier.

The Line Stretcher phase shifters are ideal delay blocks. Delay in the LO path is kept constant while that in the RF path is swept by changing N from -21 to +21.



Below are simulation results for AM and PM spur suppression, as a function of N, and also quadrature phase error.

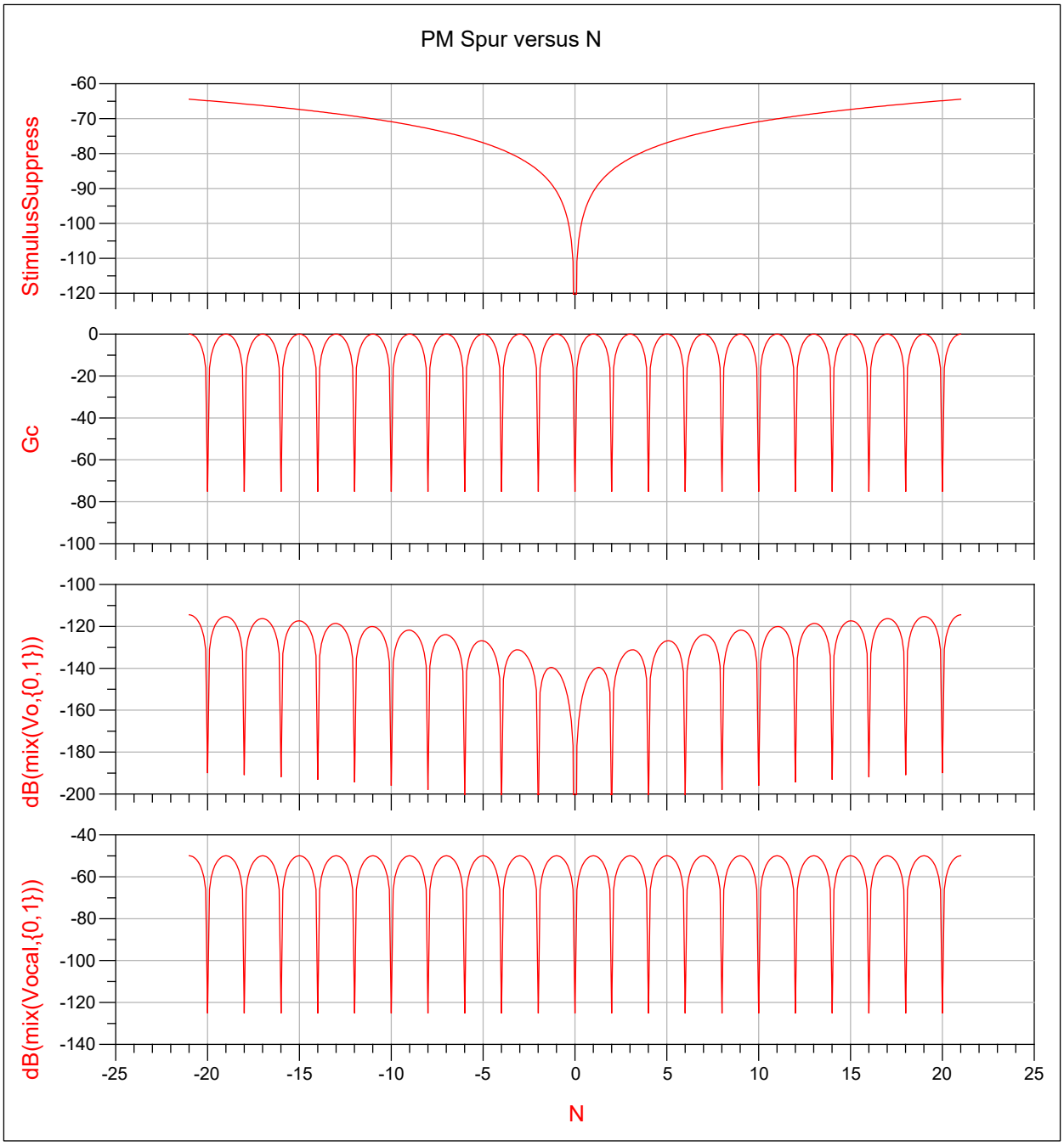


Impact of Quadrature Phase error of 15° at 5.5 GHz carrier. There is small impact on Stimulus PM suppression for Odd N at the at which the system operates.

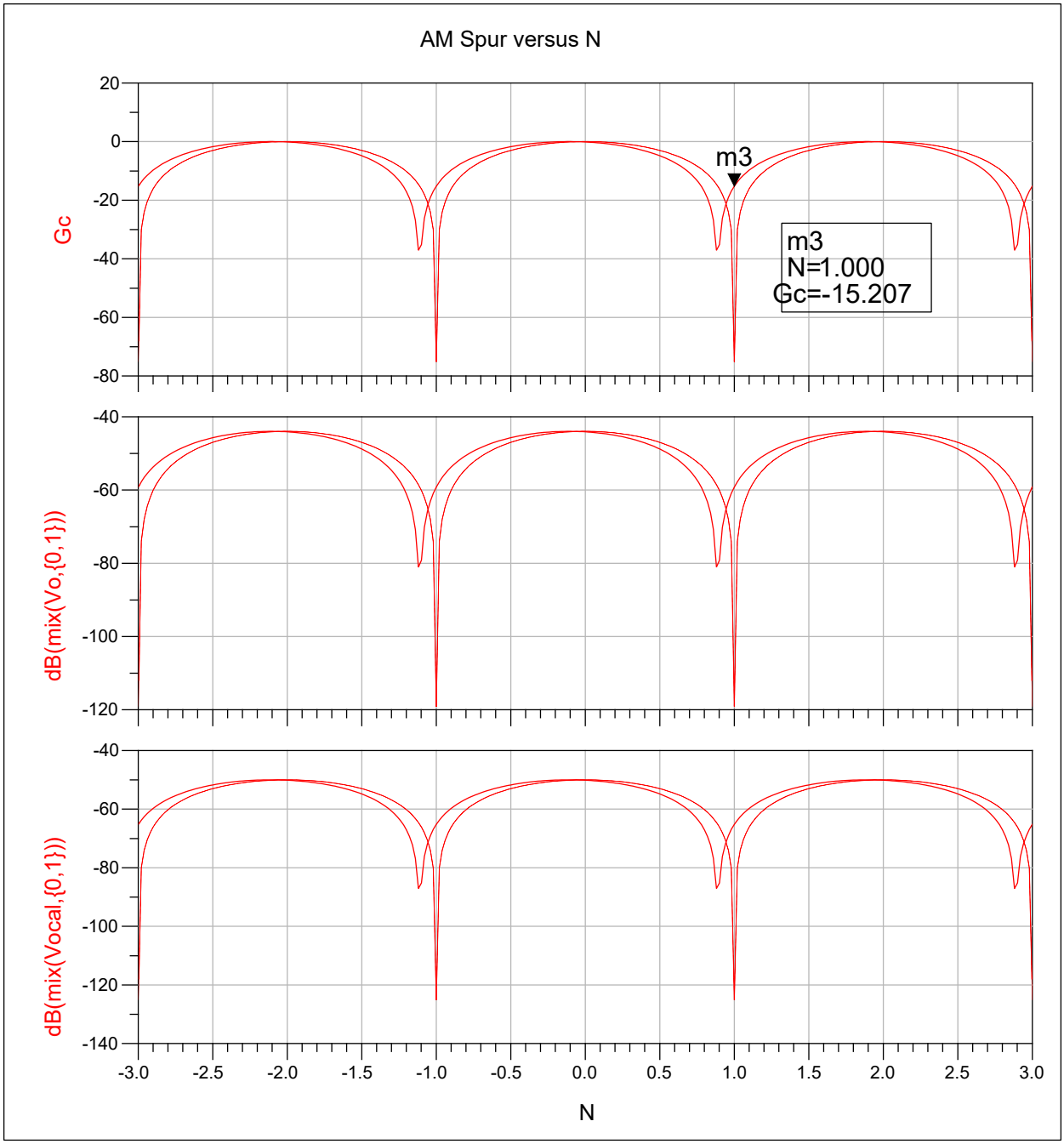
Measurement Equations:

Eqn dB(mix(Vocal,{0,1}))-d

Eqn usSuppress=dB(mix(Vo,{0,1}))-d-Gc



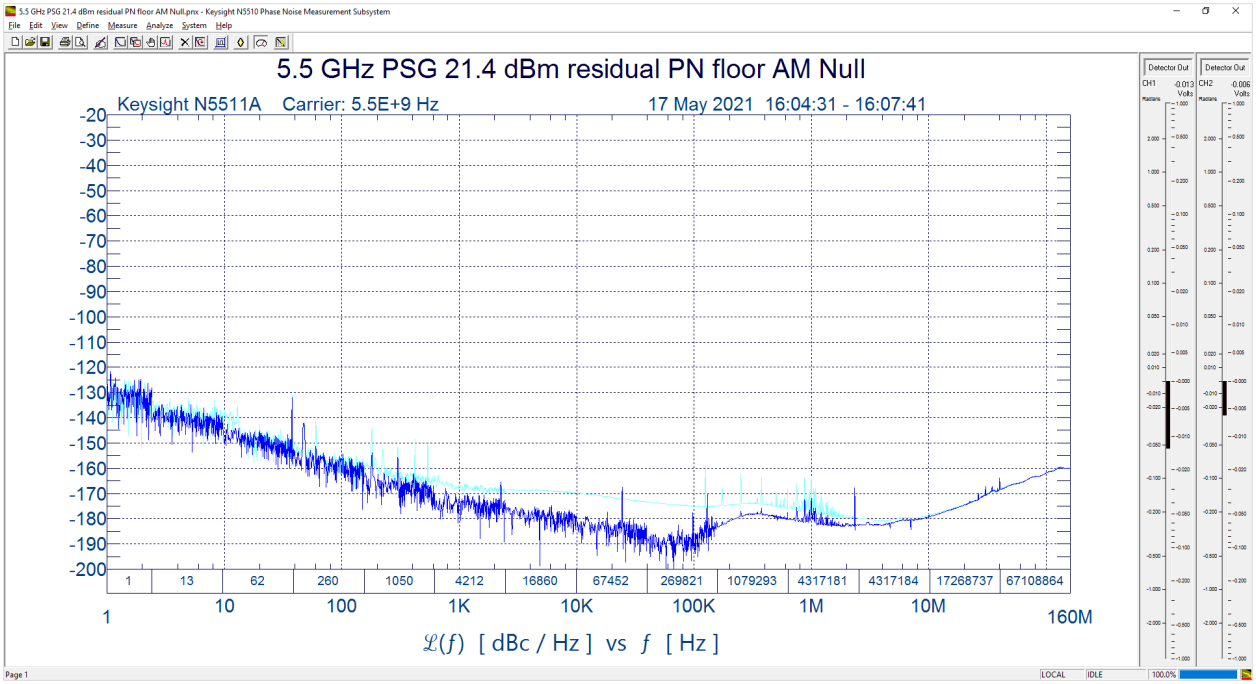
Top-most trace is Stimulus PM suppression. At 5.5 GHz carrier.

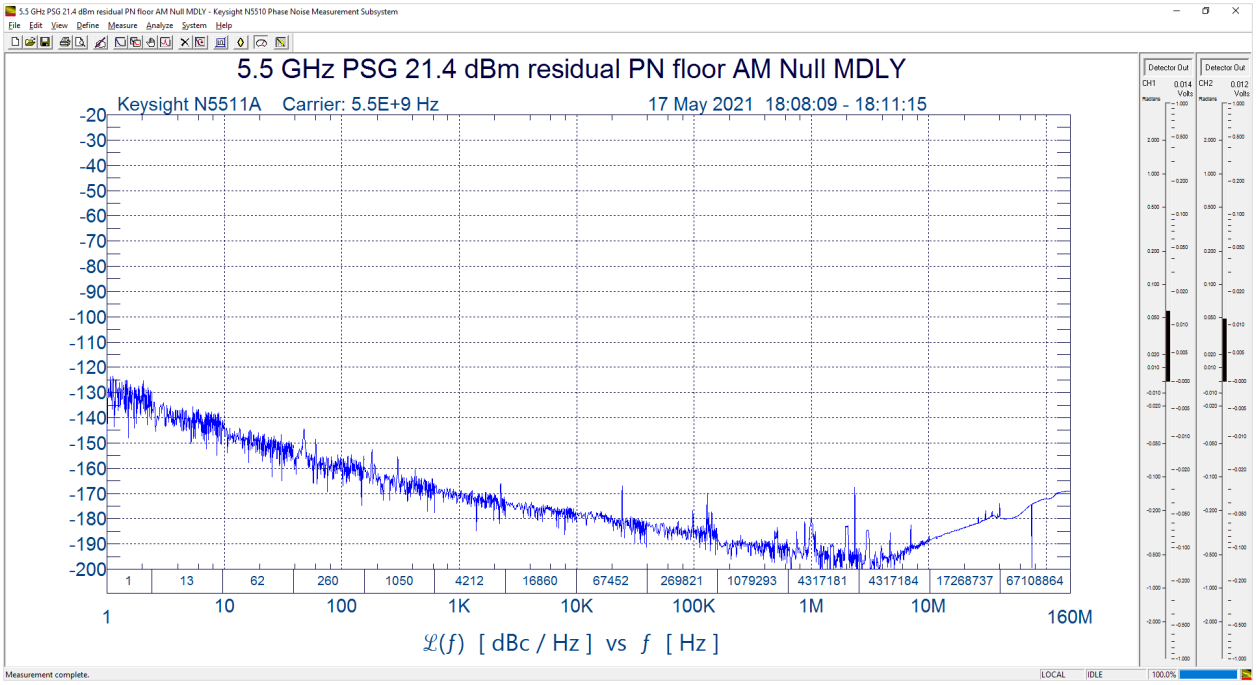


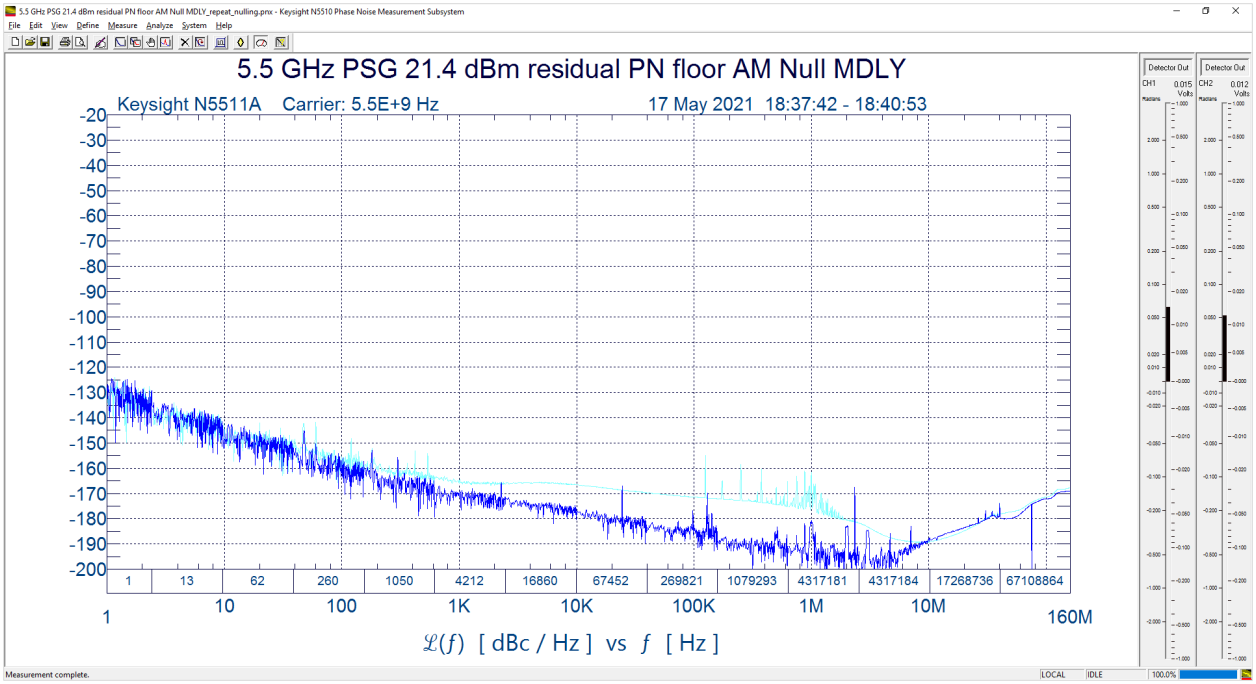
Impact of Quadrature Phase error of 10° at 5.5 GHz carrier. There is significant impact on AM Suppression for Odd N at which the system operates. The AM suppression is -15 dB for 10° phase error

Measurement Equations:

Eqn dB(mix(Vocal,{0,1}))-d


5.5 GHz PSG 21.4 dBm residual PN floor AM Null
Keysight N5511A Carrier: 5.5E+9 Hz
17 May 2021 16:04:31 - 16:07:41
-20
-30
-40
-50
-60
-70
-80
-90
-100
-110
-120
-130
-140
-150
-160
-170
-180
-190
-200
1
13
62
260
1050
4212
16860
67452
269821
1079293
4317181
4317184
17268737
67108864
1
10
100
1K
10K
100K
1M
10M
160M
ℒ(f) [dBc / Hz] vs f [Hz]
Detector Out
CH1
CH2
Page 1
LOCAL
IDLE
100.0%


5.5 GHz PSG 21.4 dBm residual PN floor AM Null MDLY
Keysight N5511A Carrier: 5.5E+9 Hz
17 May 2021 18:08:09 - 18:11:15
-20
-30
-40
-50
-60
-70
-80
-90
-100
-110
-120
-130
-140
-150
-160
-170
-180
-190
-200
1
13
62
260
1050
4212
16860
67452
269821
1079293
4317181
4317184
17268737
67108864
1
10
100
1K
10K
100K
1M
10M
160M
ℒ(f) [dBc / Hz] vs f [Hz]
Detector Out
CH1
CH2
Measurement complete.
LOCAL
IDLE
100.0%


5.5 GHz PSG 21.4 dBm residual PN floor AM Null MDLY
Keysight N5511A Carrier: 5.5E+9 Hz
17 May 2021 18:37:42 - 18:40:53
-20
-30
-40
-50
-60
-70
-80
-90
-100
-110
-120
-130
-140
-150
-160
-170
-180
-190
-200
1
13
62
260
1050
4212
16860
67452
269821
1079293
4317181
4317184
17268736
67108864
1
10
100
1K
10K
100K
1M
10M
160M
ℒ(f) [dBc / Hz] vs f [Hz]
Detector Out
CH1
CH2
Measurement complete.
LOCAL
IDLE
100.0%

References

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Product Information

- N5511A Phase Noise Test System
- PNMI: N5511A-K26
- E8257D PSG Signal Generators

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