Cross Correlation Residual Phase Noise Measurements using Two HP3048A Systems and a PC Based Dual Channel FFT Spectrum Analyser

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ABSTRACT

This paper describes a cross correlation residual phase noise measurement system based on two HP3048A systems and a dual channel FFT spectrum analyser consisting of a PC card containing two 16bit (125MS/Sec) A to D converters. A measurement system noise floor of -200 dBc/Hz is achieved for 100,000 correlations. Residual phase noise measurements are also performed on low noise L-Band microwave amplifiers, developed at York. The key features of the cross correlation technique and the different window functions required during measurement are discussed.

INTRODUCTION

Residual noise can be described as the noise added to a signal when it's processed by a two port network. It consists of both AM and PM components but in this paper we will primarily be dealing with the measurement of the PM component. A conventional residual phase noise measurement system is illustrated in Fig. 1. A low noise sinusoidal signal source is used to provide a reference signal at the input to the measurement system. This signal is then split into two separate paths that provide the drive signals to a phase detector which is typically a double balanced mixer. It is assumed that the source phase noise is correlated at the inputs to the phase detector and therefore cancels. A phase shifter is used to adjust the delay in one of the paths such that the signals mix in quadrature. This is the point at which a mixer is most sensitive to any phase fluctuations between its input signals. Finally, the output of the mixer is low pass filtered to remove any unwanted RF components before being amplified and fed to an FFT analyzer. The FFT analyser then plots the spectral density of the phase noise at the output of the mixer.

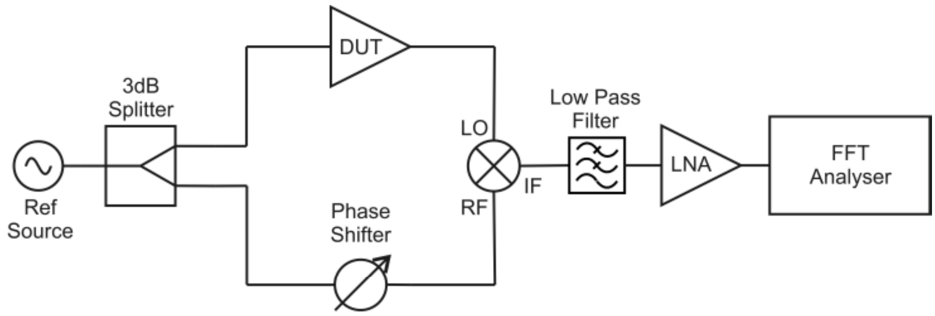

Fig. 1. Simplified single channel residual phase noise measurement system.

The noise floor of a system utilising this single channel measurement technique is highly dependent on and limited by the noise floors of the mixer, filters and low noise amplifier (LNA)[1]. This type of system can have a residual phase noise floor in the region of -180 dBc/Hz at high offset frequencies[2].

THE CROSS CORRELATION TECHNIQUE

The cross correlation technique can be used to suppress the mixer, filter and baseband LNA noise from the measurement results [1]. Using this method the uncorrelated noise in each channel can be suppressed by a factor of $\sqrt{N}$, where N is the number of cross correlations. In this type of system two independent channels are used and their outputs are fed to a cross spectrum analyser. Fig. 2 shows the simplified diagram of a cross correlation residual phase noise measurement system.

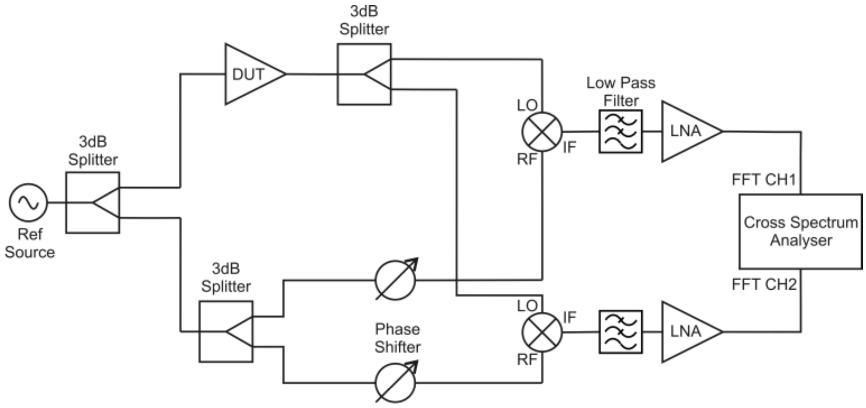

Fig. 2. Simplified cross correlation residual phase noise measurement system.

It can be seen that the reference signal is split into two paths, the first passes through the device under test (DUT). The signal at the output of the DUT is then split again where it provides the LO inputs to a pair of mixers. The second path from the reference source also passes through a 3 dB splitter whose outputs are used to provide the additional quadrature signals required to drive the mixers. This arrangement ensures that the noise added by the DUT is correlated in both channels while the noise generated by the mixers, filters and LNA remains un-correlated.

As described by Rubiola in[3], the noise present at the output of each mixer can be modelled using two noisy signals as shown in (1) and (2).

$$x(t) = a(t) + c(t) \xleftrightarrow[FFT]{} X(f) = A(f) + C(f) \tag{1}$$

$$y(t) = b(t) + c(t) \xleftrightarrow[FFT]{} Y(f) = B(f) + C(f) \tag{2}$$

Where *a(t)* and *b(t)* represent the uncorrelated instrument noise present in each channel and *c(t)* represents the correlated DUT noise. Taking the cross spectrum of these two signals and average over N samples we can write:

$$\overline{S_{XY}} = \frac{1}{N}\sum_{n=1}^{n=N}\left[X_n \times Y_n *\right] \tag{3}$$

Where n represents the sample index and * implies the conjugate. If we then substitute (1) and (2) into (3) we can write:

$$\overline{S_{XY}} = \frac{1}{N}\sum_{n=1}^{n=N}\left[\left(A_n + C_n\right)\times\left(B_n + C_n\right)*\right] \tag{4}$$

After multiplying out the brackets we can write (5):

$$\overline{S_{XY}} = \frac{1}{N}\sum_{n=1}^{n=N}\left[\left(A_n B_n *\right) + \left(A_n C_n *\right) + \left(C_n B_n *\right) + \left(C_n C_n *\right)\right] \tag{5}$$

If we assume that there is no correlation between the noisy signals *a(t)*, *b(t)* or *c(t)* then as the number of averages increases the uncorrelated terms in the cross spectrum - *AB*, *AC* and *CB* will all approach zero. The only remaining term, CC, represents the power spectral density of the correlated DUT noise.

MEASUREMENT SYSTEM

The main components of our measurement system are two HP 11848A phase noise test sets. These form part of the HP 3048A phase noise measurement system and they integrate a phase detector, LNA, PLL and several stages of filtering into a single unit. They exhibit a low single channel noise floor of below -180 dBc/Hz at carrier offsets greater than 10 kHz[2]. They are also fully computer controllable using Visual Basic software developed at the UK National Physical Laboratory (NPL) by David Adamson - (Co-Author).

A dual channel high speed data capture card with a 16 bit resolution and maximum sampling rate of 125 MSamples/Sec is used to capture the noise output from the 11848A units. A software based cross spectrum analyser has also been developed using the Java programming language. This fully multi-threaded software allows spectrum auto-correlation and cross-correlation as well as the application of multiple time domain windows to the sampled data. A block diagram of the complete measurement, including calibration components, is shown in Fig. 3. The components integrated into the HP 11848A units are shown enclosed inside the red dashed rectangles.

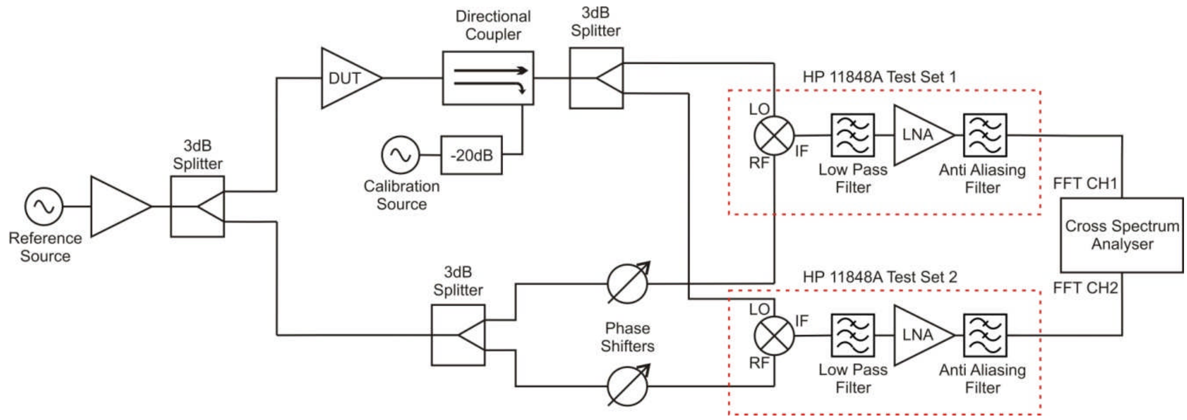

Fig. 3. Cross correlation residual phase noise measurement system including calibration components.

The reference source is provided by a battery powered ultra low phase noise dielectric resonator oscillator (DRO) developed at York. This oscillator exhibits PM and AM noise levels of less than -173 dBc/Hz at carrier offset greater than 10 kHz. Its low level of AM noise is particularly beneficial when making sensitive phase noise measurements. This is because a typical saturated double balance microwave usually only offers around 20dB to 30dB of AM noise suppression[4]. Once this has been exhausted the source AM noise will directly add to the residual phase noise and corrupt the measurement results. A power amplifier is connected to the output of the reference oscillator to increase the signal level to a value large enough to saturate the mixers. Finally, a selection high quality, double shielded, microwave cables were using to interconnect the various components.

Calibration

Calibration is achieved by injecting a tone into the measurements system via the directional coupler connected to the output of the DUT. With reference to Fig. 3, it can be seen that the calibration tone will appear as a spur at the output of each mixer. The spur frequency and carrier to spur amplitude ratio at the output of the directional coupler are noted and then the level of the calibration spur at the mixer outputs is measured using the cross spectrum analyser.

The accuracy of the calibration and resulting phase noise measurements is, in part, limited by the accuracy achieved during the measurement of the calibration spur. In order to minimise the high side lobe levels that will be present in a un-windowed spectrum we must apply a window function to the time domain data before transformation into the frequency domain.

For calibration purposes the amplitude accuracy of the calibration spur measurement is the most important factor. The Flat Top window, is an ideal candidate for this measurement. Time domain and frequency domain plots for this window function are shown in Fig. 4.

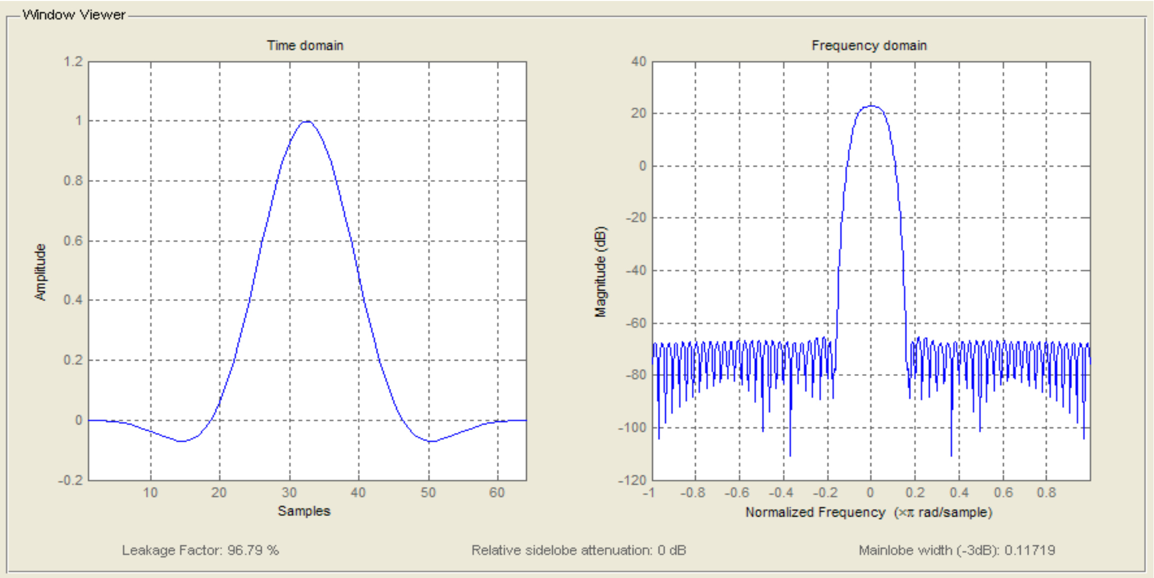

Fig. 4: Time domain (left) and frequency domain (right) plots of the flat top window function. Created using the Matlab Window Visualisation Tool.

When the calibration measurements have been completed the tone is removed and the coupled port of the directional coupler is terminated in 50 Ω. The measured noise can then be converted to single sideband phase noise, *L(f)*, using (6).

$$L(f) = S_{\varphi} + K_1 - K_2 - 6 \tag{6}$$

Where S_{φ} is the double side band phase noise spectral density, K_1 is the carrier to sideband spur ratio and K_2 is the amplitude of the calibration spur - all parameters are in decibels. The subtraction of an additional 6 dB is necessary in order to account for the fact that we are measuring double side band phase noise at the output of the mixers. The primary advantage of calibrating the instrument in this way is that the calibration is performed under the actual measurement conditions with the DUT in place. It should also be noted that ideally a calibration spur should be injected at every measurement frequency of interest; however, the frequency response of the HP11848A units is sufficiently flat this is not necessary.

In order to check the validity of the calibration a second independent calibration technique was used. This was developed by Dr. Bob Longstone at BAE Systems. The DUT was initially replaced with the series connection of a 44 dB attenuator and a 44 dB amplifier. This arrangement artificially increases the system noise floor to a value that can be calculated using (7) :

$$L(f) = -177 + N_A - P_i \tag{7}$$

Where N_A is the noise figure of the combined amplifier and attenuator and P_i is the power available at the input to the attenuator. If we substitute in the combined noise figure of 48.5 dB and an input power of 20 dBm then the theoretical phase noise floor can be calculated to be -148.5 dBc/Hz at offsets above the flicker noise corner. The measured noise floor was -148.4 dBc/Hz which is in good agreement with the theory.

SYSTEM NOISE FLOOR

The system noise floor is measured by removing the DUT from the equipment setup shown in Fig. 3 and replacing it with a straight through connection. The phase shifters were adjusted so that DC level at the output of each mixer was at a minimum. In order that the source phase noise did not de-correlate at large offset frequencies the group delay of each channel was measured to ensure that the phase shift between the mixer input ports was only 90° and not a multiple of 360° + 90°. The measurement was performed at a frequency of 1.25 GHz with mixer LO and RF port input powers of 16.1 dBm and 16.9 dBm respectively. The use of high power levels ensures that the mixers are truly saturated and minimises the signal channel noise floor of each HP 11848A unit. The signal processing was performed at a sampling rate of 2 MSamples/Sec with a data frame length of 262,144 points. This equated to a resolution bandwidth of 7.63 Hz. In order to minimise the FFT noise bandwidth no additional data windowing was performed. That is to say that the data was effectively multiplied by a uniform window resulting in a noise bandwidth identical to the resolution bandwidth.

After the calibration had been performed the residual noise floor was measured for increasing numbers of cross correlations. Fig. 5 details residual phase noise floor and acquisition time for each of these measurements. The final measurement of 100,000 cross correlations has an increased resolution bandwidth because the number of data points acquired in each frame was reduced in order to shorten the measurement time.

Fig. 5. Residual phase noise floor of the measurement system at 20 kHz offset.

Number of Cross Correlations	System Noise Floor at 20 kHz Offset (dBc/Hz)	Measurement time (Seconds)	Resolution Bandwidth (Hz)
100	< -185 dBc/Hz	73	7.63 Hz
1000	< -190 dBc/Hz	730	7.63 Hz
10,000	< -195 dBc/Hz	7300	7.63 Hz
100,000	< -200 dBc/Hz	3620	122.07 Hz

Fig. 6, Fig. 7 and Fig. 8 show plots of residual phase noise floor for 1000, 10,000 and 100,000 cross correlations respectively. The blue and red (upper) traces are the noise floors for each independent channel and the green (lower) trace is the dual channel cross correlated noise floor. It should be noted that due to the large variance of the measured noise the top of noise line was used to provide a conservative estimate for the noise floor. If we compare these three figures, the reduction in noise floor is clearly visible as the number of correlations increase. The phase noise floor is reducing by approximately 5 dB for ever factor of 10 increases in the number of cross correlations and this is in good agreement with the theory. Unfortunately the close to carrier noise, at frequency offsets below 1 kHz, does not show the same level of suppression. This is thought to be a result of the high levels of spurs that are present in the 10 Hz to 1 KHz region.

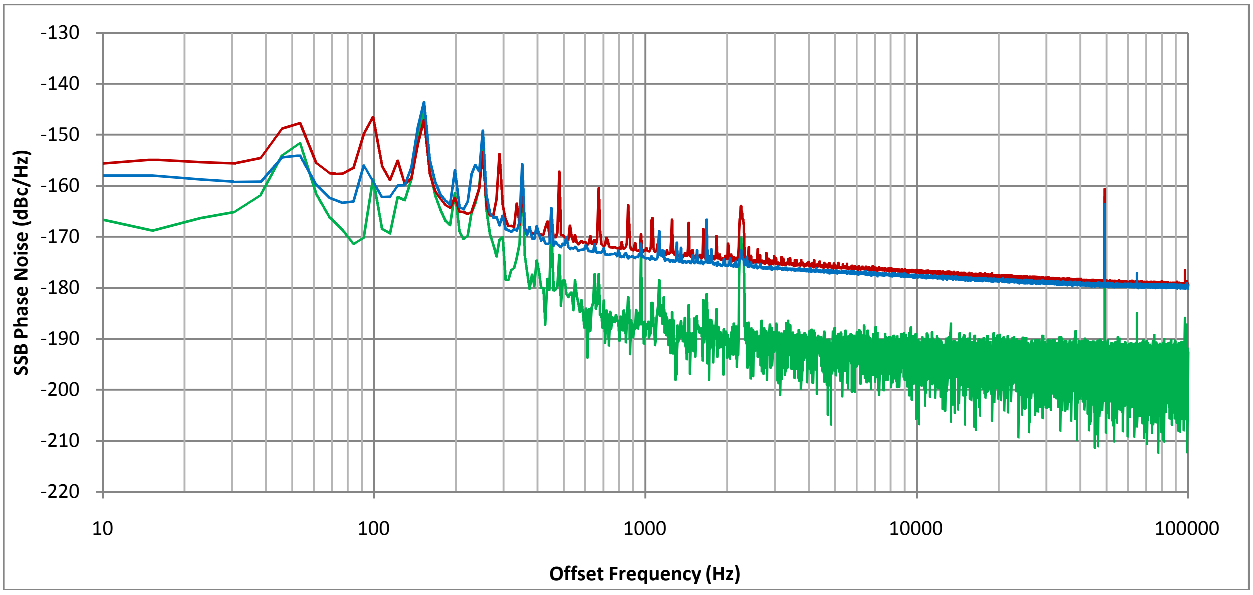

Fig. 6. Residual phase noise floor of the measurement system after 1000 cross correlations.

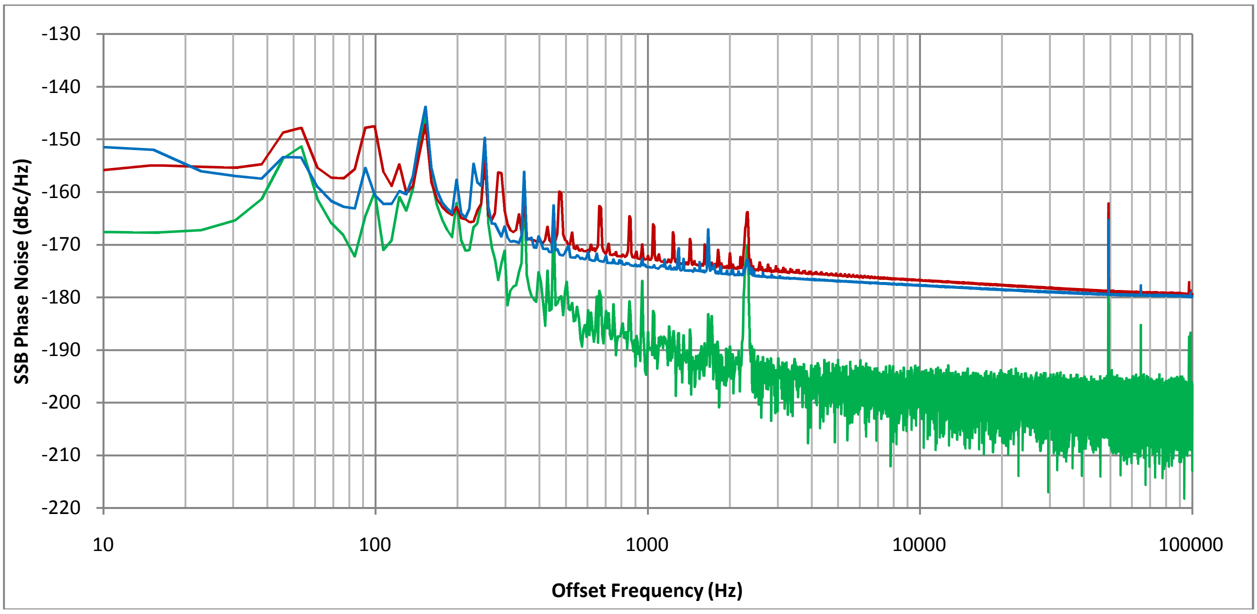

Fig. 7. Residual phase noise floor of the measurement system after 10,000 cross correlations.

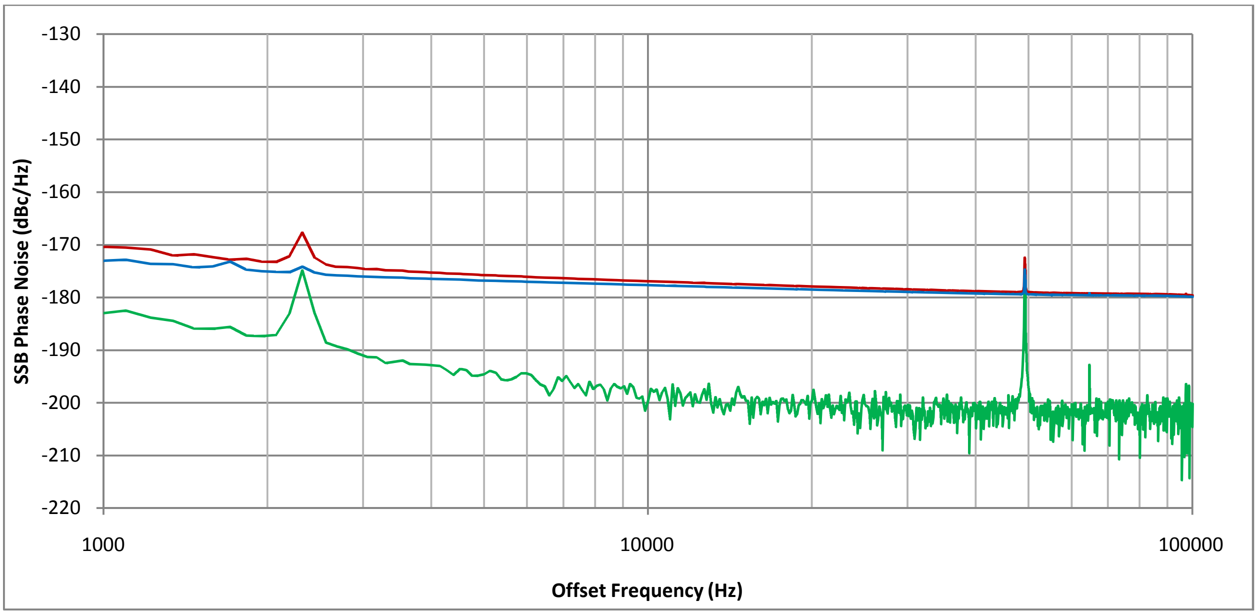

Fig. 8. Residual phase noise floor of the measurement system after 100,000 cross correlations. This measurement was made with an increased resolution bandwidth of 122.07 Hz in order to reduce the measurement time.

AMPLIFIER MEASUREMENTS

A measurement of a 1.25 GHz medium power silicon amplifier has been performed, this amplifier had a noise figure of 8 dB. The amplifier was powered using batteries in order to reduce the potential for additional interference. A power level of 17.2 dBm was provided to its input and the theoretical far from carrier residual phase noise floor was calculated using (7) as -186.2 dBc/Hz. Fig. 9, Fig. 10 and Fig. 11 show plots of the amplifiers residual phase noise for 100, 1000 and 5000 cross correlations respectively.

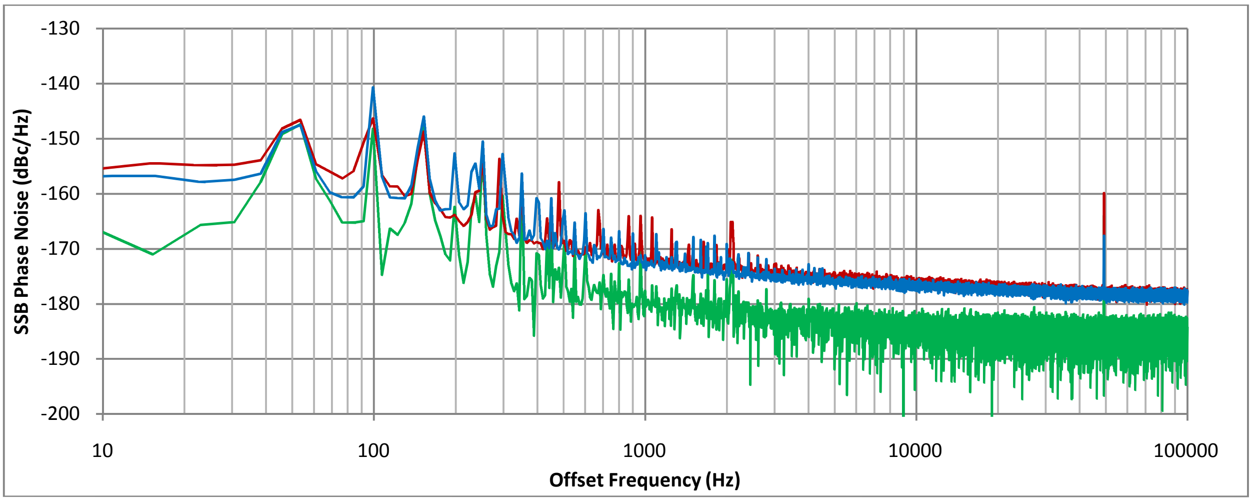

Fig. 9. Residual phase noise floor of a 1.25 GHz amplifier after 100 cross correlations.

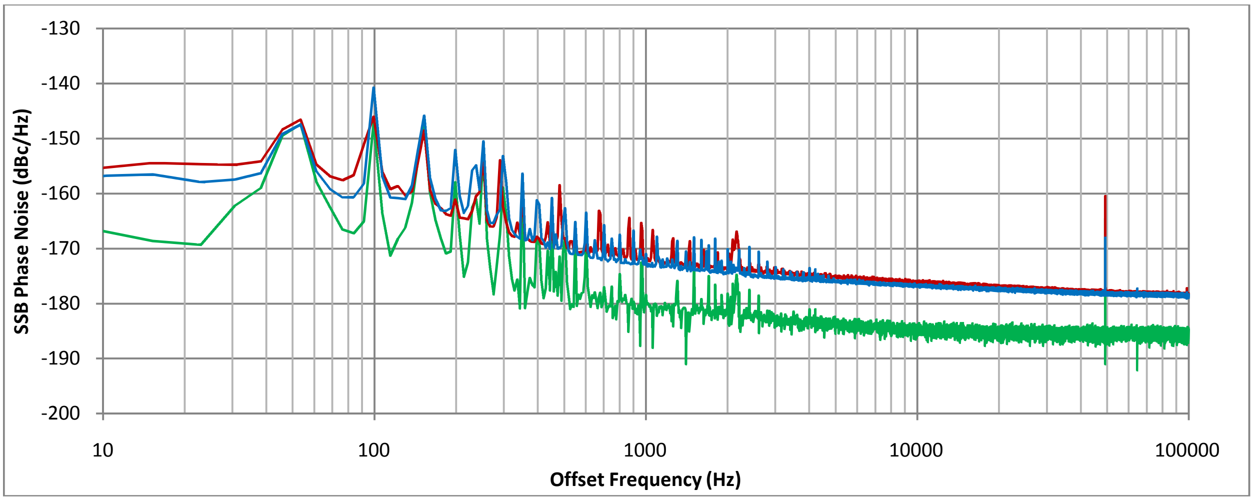

Fig. 10. Residual phase noise floor of a 1.25 GHz amplifier after 1000 cross correlations.

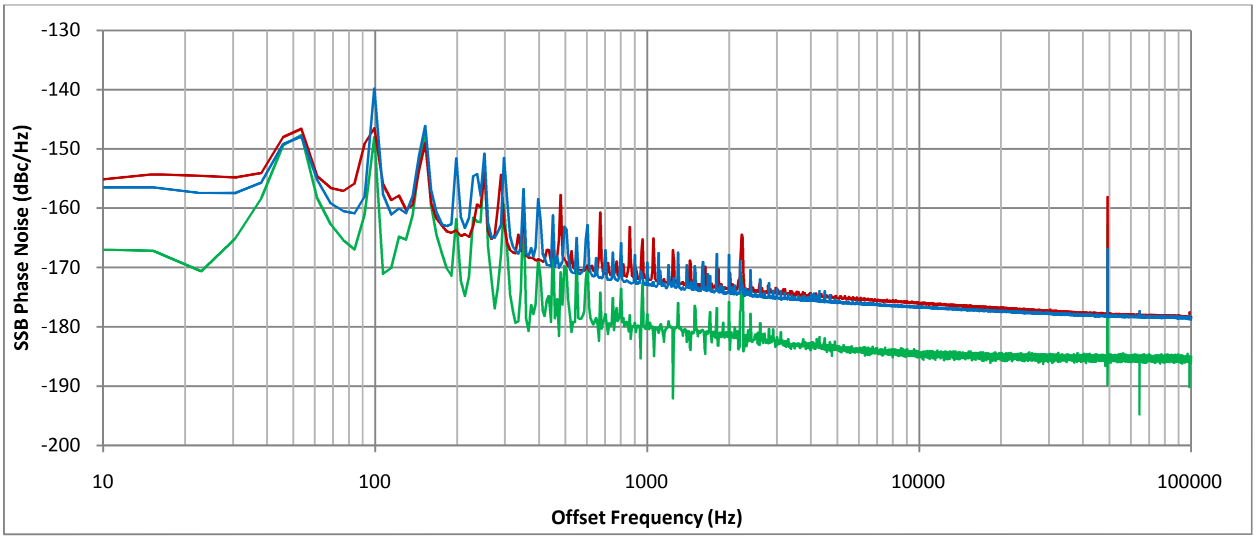

Fig. 11. Residual phase noise floor of a 1.25 GHz amplifier after 5000 cross correlations.

With reference to Fig. 9 it can be seen that after 100 correlations the measured noise still has a large variance. This is because the uncorrelated instrument noise is present in the measurement data. However, as the number of correlations increases it can clearly be seen that the measured noise is converging to an absolute value. At offsets above 10 KHz the residual phase noise is approximately -185 dBc/Hz, this is good agreement with the theoretical value of -186.2 dBc/Hz.

CONCLUSIONS

A cross correlation residual phase noise measurement system has been constructed with a noise floor of -200dBc/Hz for 100,000 correlations. This is an improvement of 20 dB when compared with the same system operating with a single channel. No additional AM suppression was required to achieve this result. However, the reduction in noise floor comes at the expense of increased measurement time and system complexity.

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