P4R.3 PHASE NOISE, COHERENCY, AND CLUTTER SUPPRESSION

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1. INTRODUCTION

Clutter is an unwanted radar return from nonmeteorological targets. Typically, these clutter targets are stationary, hard targets such as building, mountains, towers, etc. Other clutter returns, sea clutter, aircraft, birds, ships, and anomalous propagation are not from stationary products and thus are more difficult to characterize.

Clutter suppression is a feature of modern weather radar. The clutter filter consists of a high pass filter, eliminating the returns with low radial velocity (ideally zero velocity) characteristic of stationary clutter. The clutter filter is typically specified as the type of filter (Finite Impulse Response, FIR, or Infinite Impulse Response, IIR), the delay inherent in the filter (number of poles or points), the rejection depth of the clutter (50 dB), and their width ($0.05f_{PRF}$). Table I summarizes the filter specifications from a number of weather radar signal processors.

Table I The filter specifications from a number of weather radar signal processors.

Signal Processor	Filter Type	Poles	Pub. Depth (dB)	Meas. Depth (dB)
RVP-5	FIR	11	37	37
RVP-6	IIR	4	40	40
ESP-7	IIR	3	50	45
EDRP-8	IIR	4	50	45
EDRP-9	IIR	3	50	65
WSR-88D	IIR	5	50	65

Though not obvious from manufacturer descriptions, the maximum rejection depth of the clutter filter does ***not*** specify the maximum suppression supported by the system as a whole, rather it denotes a maximum guaranteed rejection if the system will support it. This is emphasized by the last column of Table I which shows measured clutter suppressions for a number of systems.

Another typical performance specification for a weather radar is the phase noise specification. The phase noise describes the variance of the phase angle measurement. Therefore, the phase noise is a specification of the maximum error allowed in the phase measurement.

Since a Doppler radar measures the Doppler shift through the phase shift induced by the different frequency, the phase noise of the system is limit the systems ability to differentiate velocity. As the clutter filters operation depends upon the ability to differentiate velocity (rejects those signals with low velocity), the system phase noise provides a limit to the amount of clutter rejection. This limit is specified by taking the variance of the phase and converting to decibel radians. Mathematically, this is written,

$$CS = -20\log\left(\Delta\phi\right). \qquad (1)$$

Table II lists the phase noise specification for a number of weather radar systems and the associated maximum clutter suppression.

Table II The phase noise specification for a number of weather radar systems and the associated maximum clutter suppression

Radar System	Max Phase Noise (deg)	Maximum Clutter Rejection (dB)
DWSR2501C	0.5	41
TVDR2500C	0.6	40
DWSR2501C/K	0.1	55
DWSR8501S	0.5	41
DWSR8501S/K	0.1	55
WSR-88D	0.18	50

The phase noise specification as noted in Table II provides a single value describing the phase noise. This is in fact a bit simplistic. The phase noise actually varies with respect to the frequency offset. Table III is a standard table representation of the phase noise spectrum, and Figure 1 shows the measured phase noise spectral power density in blue. Some manufacturers list the power levels not at the specific decade frequency, but rather at the center of the decade. This curve is plotted in green in Figure 1. Finally, the red line in Figure 1 shows the noise floor of the test equipment.

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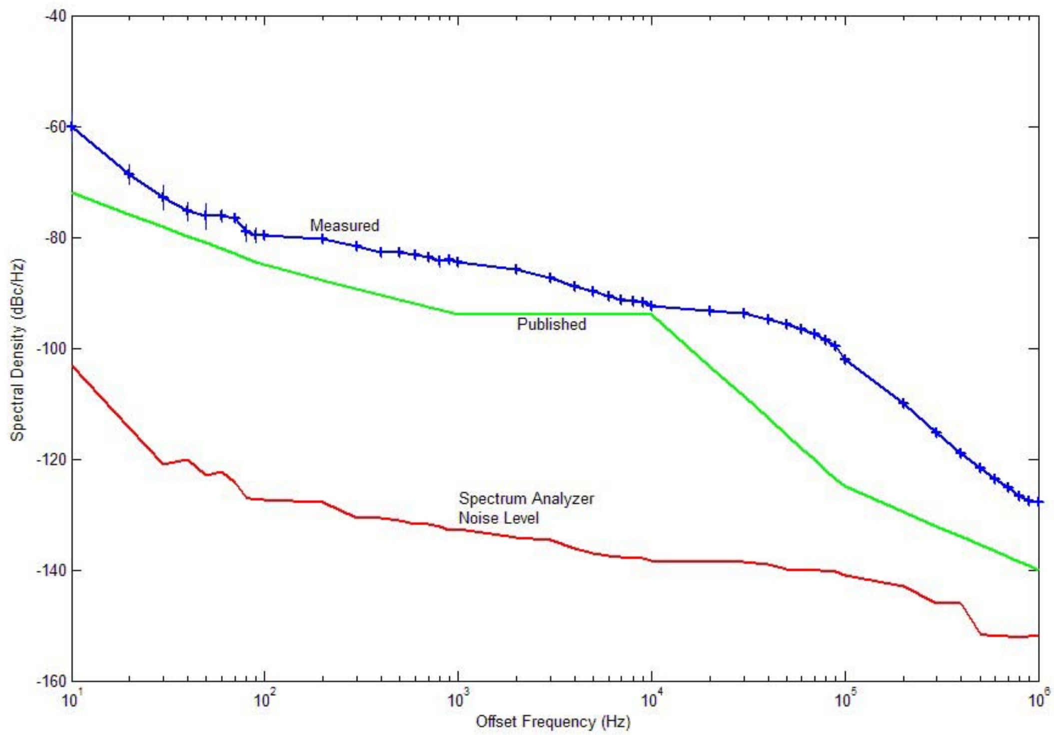


Figure 1. The spectral datain blue plotted along with the manufacturer reported data in green. The red curve is the noise floor of the measurement system. The manufacturer provided data in green is shows less noise than the measured results..

Table III. Phase noise performance table.

Frequency Offset (Hz)	Spectral Power (dBc/Hz)
10	-60
100	-80
1000	-84
10000	-92
100000	-102
1000000	-128

Another performance specification for radar systems is the *Integrated Phase Noise* (IPN). Integrated phase noise is just as it sounds, it is the integration of the phase noise spectrum. That is, by definition, to find the area under the spectral curve of the phase noise measurement. The spectral plot is given in logarithmic units, i.e. dB/Hz. To find the area, this needs to be converted to linear units, i.e. W/Hz. Figure 2 shows the linear curve for the blue spectrum of Figure 1.

The maximum clutter suppression of a system is related to the phase noise of the system. Empirically it is known that the maximum clutter suppression is directly related to the system phase noise variance provided the clutter filters have enough depth and a signal to noise ratio greater than the filter depth.

The signal to noise ratio (SNR) is as defined, the ratio of the signal power to the noise power. This ratio is typically described in logarithmic units (dB). The SNR presents another fundamental limit to the maximum clutter suppression, for the signal cannot be suppressed (or accurately measured if it was) below the noise level. In fact, another definition of phase noise is the reciprocal of the SNR in linear units (Ivo, 2001),

$$\Delta\phi_{rms}^2 = \frac{1}{SNR_{lin}}. \tag{2}$$

This translates into a maximum clutter suppression of,

$$CS = SNR, \tag{3}$$

where in this case the signal to noise ratio is in logarithmic (dB) units. Thus, if the system is very quiet and the clutter filters are sufficiently deep, the

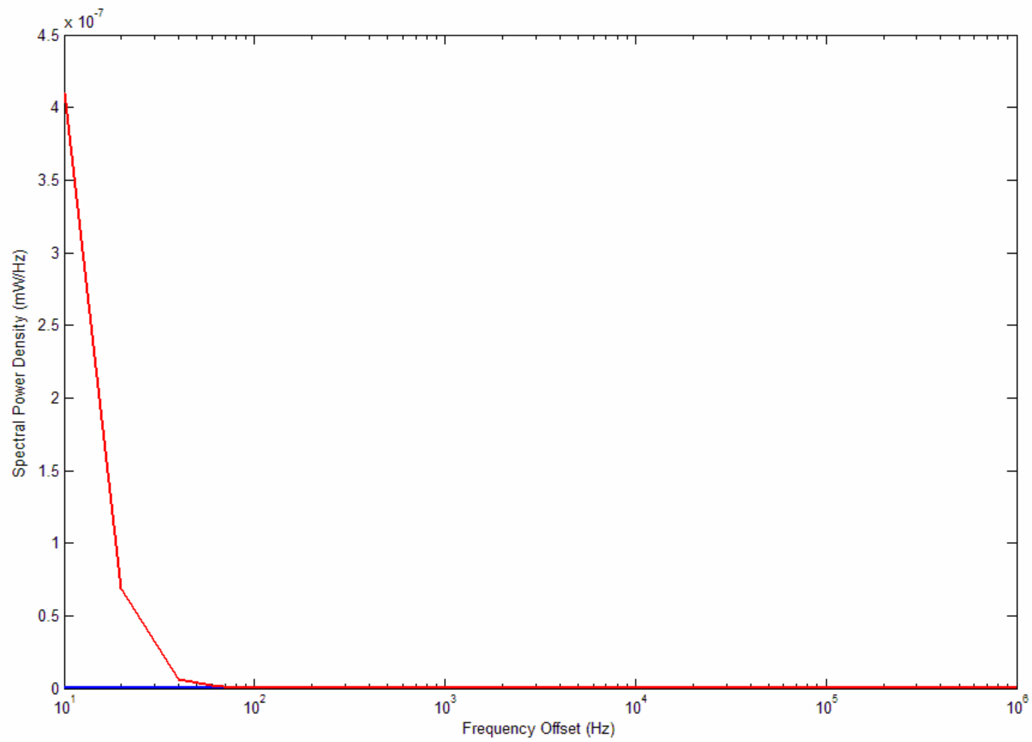


Figure 2. The phase noise spectrum of the EEC Oscillator in linear units, mW/Hz.

clutter rejection will be limited by the SNR. For example, suppose a system with a phase noise of only 0.06 deg (60 dB), has an SNR at the receiver of only 54 dB. The clutter rejection will be limited to 54 dB, though the rest of the system can support further rejection. This is an important result when evaluating clutter rejection at low signal levels, i.e. the signal being evaluated must be well above the noise level. This is also important for receiver design, with overlapping channels to extend dynamic range. Depending upon the channel selected by logic, the clutter rejection may be limited in the overlap range.

A common belief is that the maximum clutter rejection is directly related to the integrated phase noise (IPN). Basically, the idea is that the IPN will give a value 3 dB below that of the phase variance. Based upon this assumption, measuring the clutter suppression with a delay line assesses the IPN. In this paper, we demonstrate that this assumption is only true for clutter targets infinitely far from the transmitter.

The remainder of this paper will discuss how phase noise is measured, demonstrate the inconsistency of the IPN with the maximum clutter suppression, develop the theoretical approximations to the phase noise from spectral measurements of the system and apply these measurements to estimate the maximum clutter suppression of the system. An experimental apparatus and techniques are described to evaluate the theoretical estimates. Finally, results from the DWSR-8501S/K system delivered to Evansville, Indiana for the United States National Weather Service is provided.

2. PHASE NOISE MEASUREMENT

Phase noise measurement is as much art as it is science. The standard methodology for measuring the phase noise spectrum is to mix the signal from the device or system under test with that of a reference oscillator at the same frequency and view the spectrum of the results. Figure 3 schematically shows this method. To ensure an accurate measure of the device under evaluation, the phase noise of the reference oscillator should be significantly less than that of the device under test. Phase noise test equipment on the market implements this technique and outputs a spectrum of the device under test. A sample spectrum is shown in Figure 4.

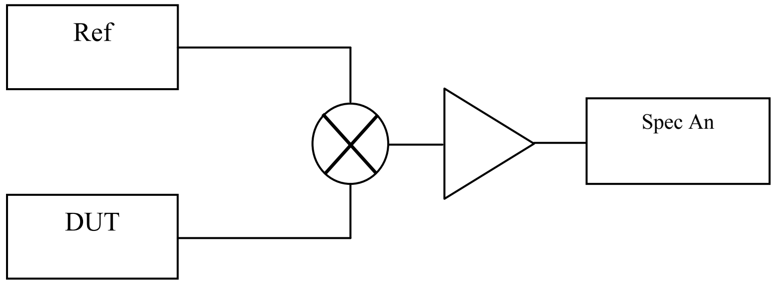


Figure 3. Schematic of a phase noise test configuration. The components identified as Ref is the reference oscillator, the DUT is the Device Under Test, and the Spec An is the spectrum Analyzer.

An alternative test arrangement is to split the signal from the device under test, and mix the result together, and obtain the spectrum. In this case, the resulting phase noise measurements will be twice as high (+3 dB higher) than the true phase noise of the system. This is due to the additive properties of signal mixing.

Throughout our studies, an Agilent E4445A Spectrum Analyzer with Phase Noise Personality function was used. This particular piece of equipment contains internal low noise sources with which to evaluate the phase noise. Figure 5 is a measured phase noise spectrum of a EEC Oscillator with the noise floor of the E4445A superimposed in blue. Note that the noise floor remains 10 to 20 dB below the oscillator spectrum. For each device for which data was collected, it was collected at least 10 times to allow statistics. We see that the phase noise of the measurement apparatus has minimal impact on the measurement of the oscillator phase noise.

The integration process involves dividing the frequency axis into small segments. The spectral power density values for each side of a segment is used to estimate the area under the curve for that segment. This is shown in Figure 6.

The resultant area (represents the power between these two offset frequencies) is given by the formula,

$$P_n = \frac{1}{2}\left(S(f_{n-1}) + S(f_n)\right)\left(f_n - f_{n-1}\right). \qquad (4)$$

All of the resultant areas are then summed to give the integrated phase noise, and the sum is converted to decibels.

Integrating the spectrum from the EEC oscillator shown in Figure 5, the integrated phase noise is found to be -54 dBc, yet using a 10 µs delay line, as discussed in the next section, this oscillator demonstrated a clutter rejection of 66 dB.

3. RESULTS

We evaluated the impact of the STALO phase noise on the clutter rejection of the radar system. This was done by characterizing the phase noise of the STALO using the Agilent E4445A Spectrum Analyzer Phase Noise Personality system. A Teledyne Brown 10 µs SAW delay line was utilized to characterize the clutter rejection capability of the system. A 10 µs delay line simulates clutter located a distance of 1.5 km ($cT_d/2$) from the radar. The clutter filters were applied to the received data and the resultant suppression noted. In addition, comparison with operational clutter rejection was obtained for some STALO's.

Figure 7 shows sample screen shots for the clutter rejection measurements from the DWSR-8501S/K delivered to Evansville, Indiana for the U.S. National Weather Service. Figure 7a shows that for the 10 µs delay, the clutter rejection was about 66 dB. Figure 7b was obtained during the operation of the radar system. A clutter target was located approximately 21 km from the radar and the clutter rejection was measured. As we can see, for a clutter target located 21 km away, the rejection was still over 60 dB.

Four oscillators, listed in Table IV, were characterized in terms of clutter suppression and integrated phase noise. Their phase noise spectra are plotted in Figure 8.

Table IV. Clutter suppression as measured with a 10 µs delay line and integrated phase noise for various synthesizers

STALO	Clutter Suppression (dB)	Integrated Phase Noise (dBc)
WSR-88D	65	-29
OSC - 1	52	-54
OSC - 2	38	-40
EEC Oscillator	66	-54

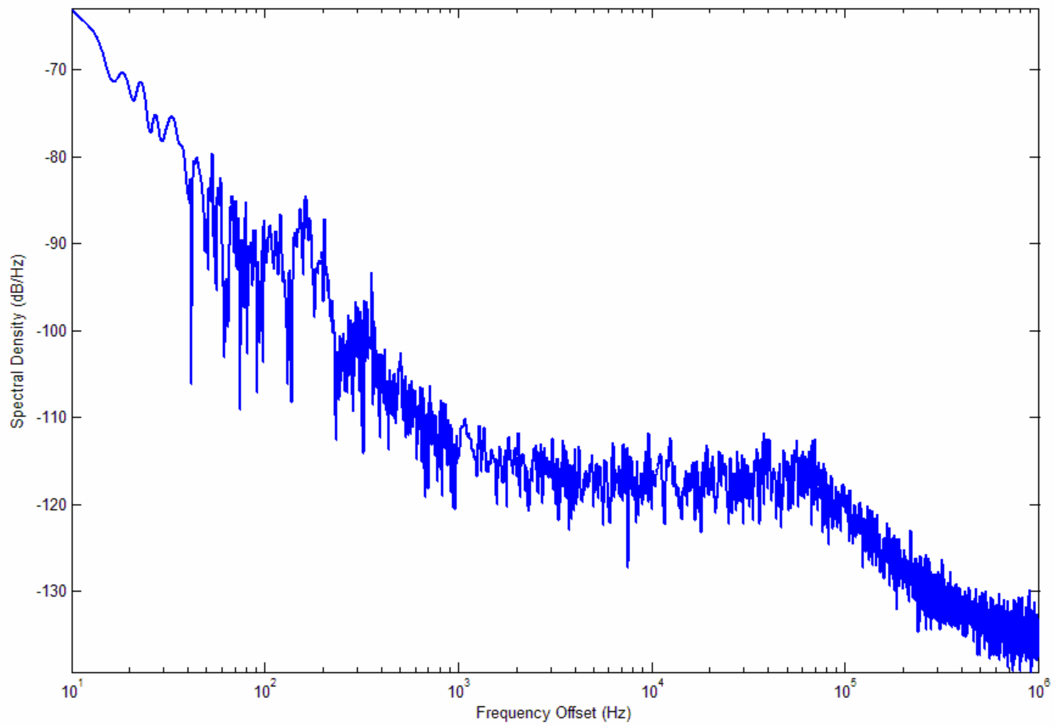


Figure 4. The spectrum of a device under test.

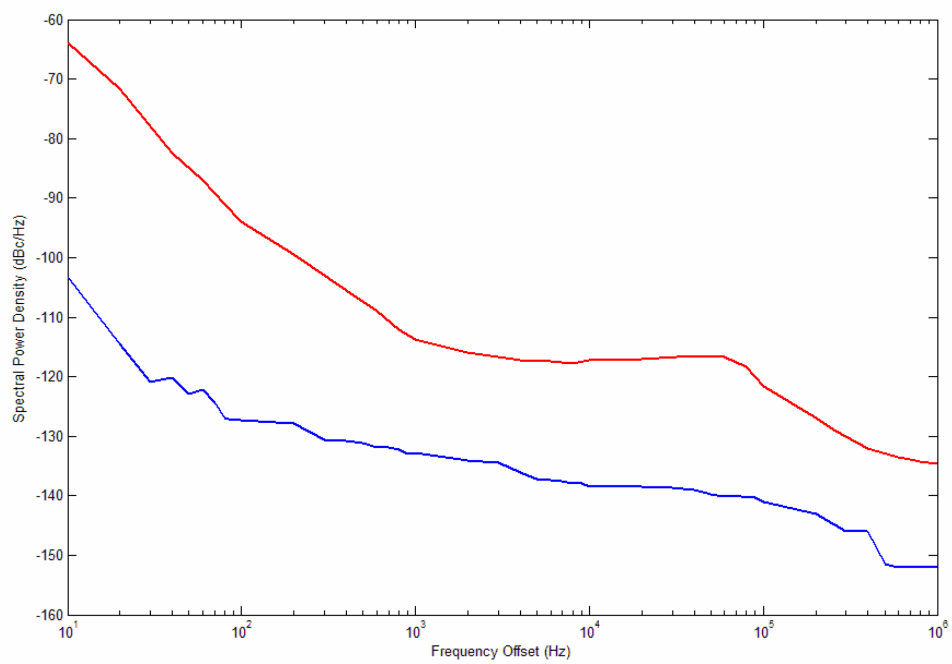


Figure 5. The phase noise spectrum of the EEC Oscillator in red with the noise floor of the E4445A superimposed in blue.

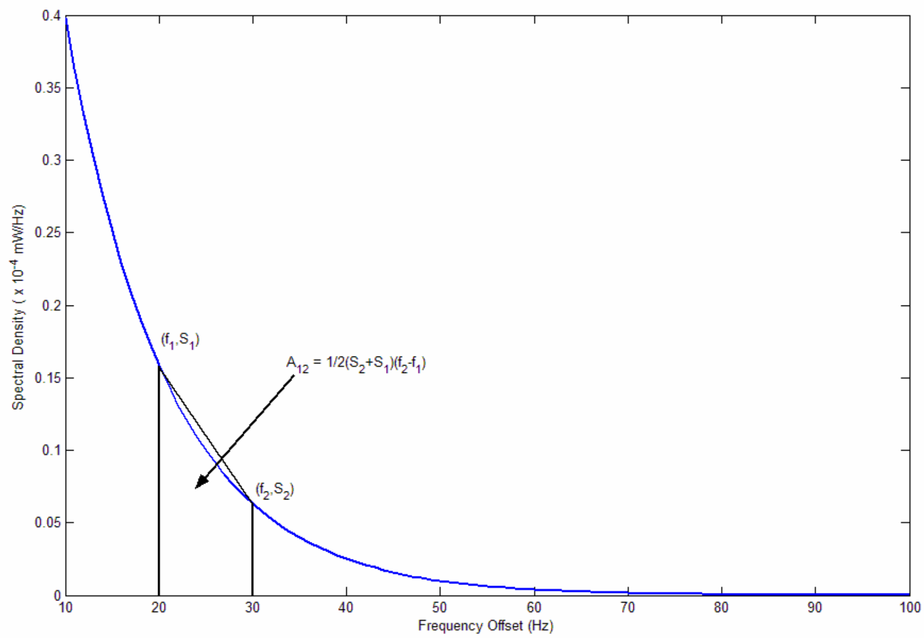


Figure 6 Finding the area under the curve between two points.

As we can see, the integrated phase noise does not correlate well at all with the clutter rejection. In fact the phase noise from small offset frequencies is clearly not as important as the noise from higher frequencies. This is a surprising result.

To explain the results summarized in Table IV, we need to look at the effect of a finite delay on the signal phase noise.

4. PHASE NOISE MODEL WITH DELAY

In developing the phase noise model with delay corrections, we need to understand the spectral function describing the signal before the addition of noise. Then, we can see what the addition of noise does to the signal and hence its behavior.

A very pure sinusoidal signal is described by,

$$s(t) = A_0 \sin(\omega_0 t + \phi_0) \tag{5}$$

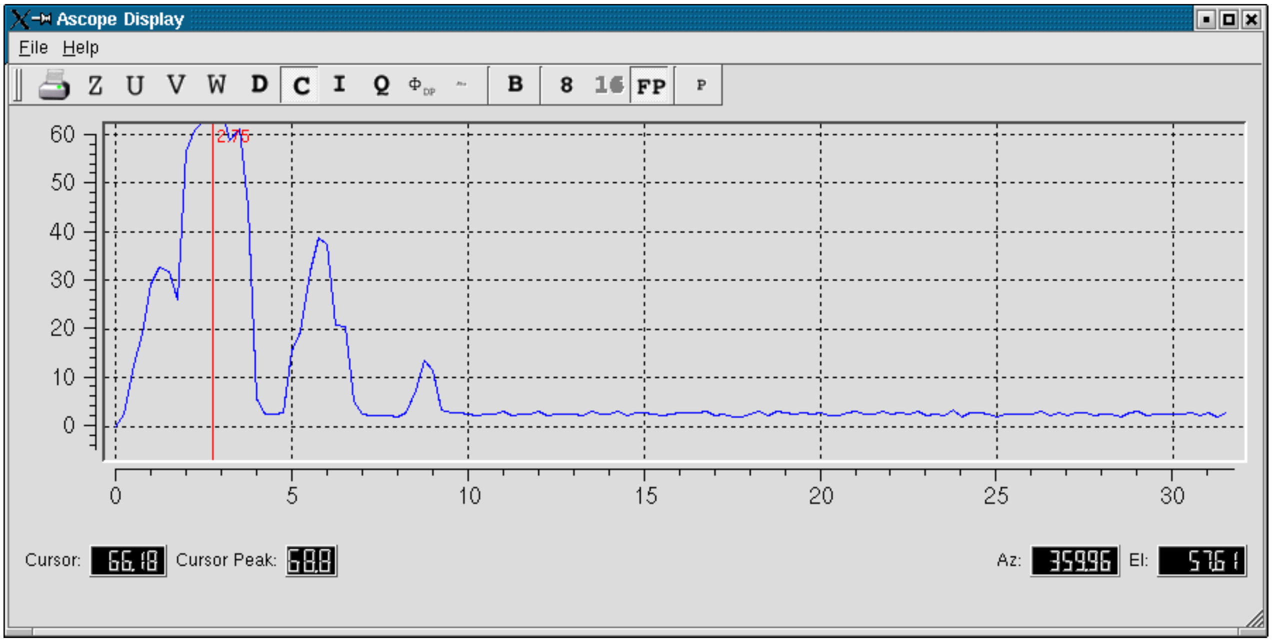
where A_0 is the amplitude, ω_0 is the central frequency, and ϕ_0 is the initial phase. Transforming the signal from temporal space (time based) to frequency space through a Fourier transform,

$$S(\omega) = \mathrm{F}(s(t)) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\infty} s(t) e^{j\omega t} dt \ , \tag{6}$$

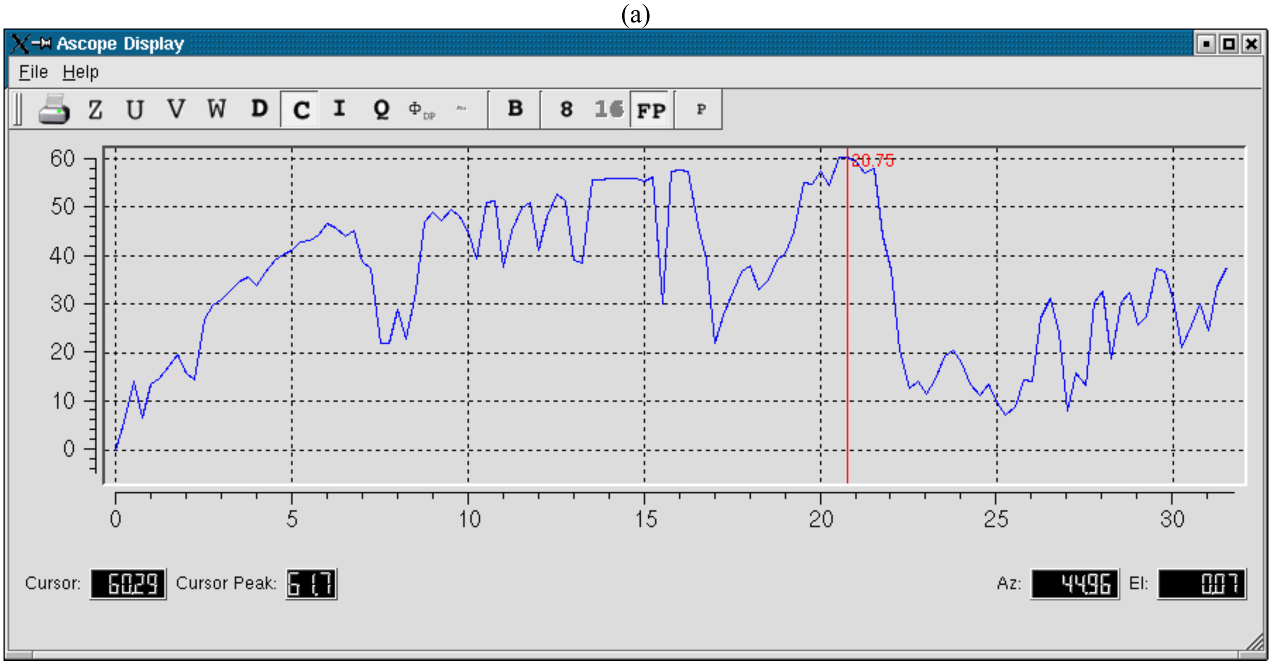
the result is a delta function,

$$S(\omega) = \frac{A_0}{\sqrt{2\pi}} \delta(\omega - \omega_0) . \tag{7}$$

The delta function has the properties of having value 1 at the center frequency and zero elsewhere, i.e.



(a)



(b)

***Figure 7.** A-scope screen shots of the clutter rejection tests of the DWSR-8501S/K delivered to Evansville, Indiana for the United States National Weather Service, with the fixed frequency STALO installed. (a) A 10 µs RF delay line test, and (b) real clutter target approximately 21 km from the radar.*

$$\delta(\omega-\omega_0)=\begin{cases}1 & \omega=\omega_0\\ 0 & \omega\neq\omega_0\end{cases}. \qquad (8)$$

Thus, the spectral function for the pure sinusoidal signal described in Eqn (6) is,

$$S(\omega)=\begin{cases}\dfrac{A_0}{\sqrt{2\pi}} & \omega=\omega_0\\ 0 & \omega\neq\omega_0\end{cases}. \qquad (9)$$

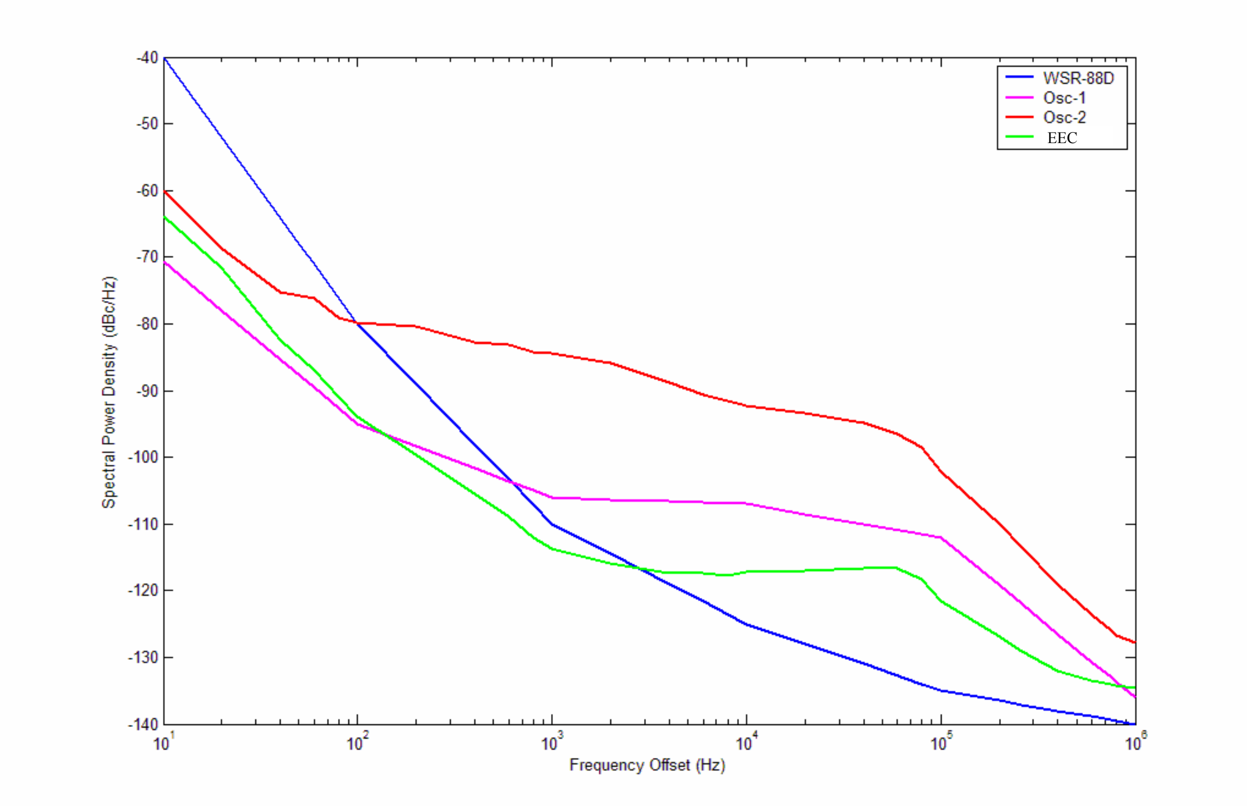


Figure 8. The phase noise spectra for the oscillators listed in Table IV. The WSR-88D legacy oscillator is in blue, Oscillators 1 and 2 are magenta and red respectively, and the EEC oscillator delivered with the Evansville DWSR-8501S/K is in green.

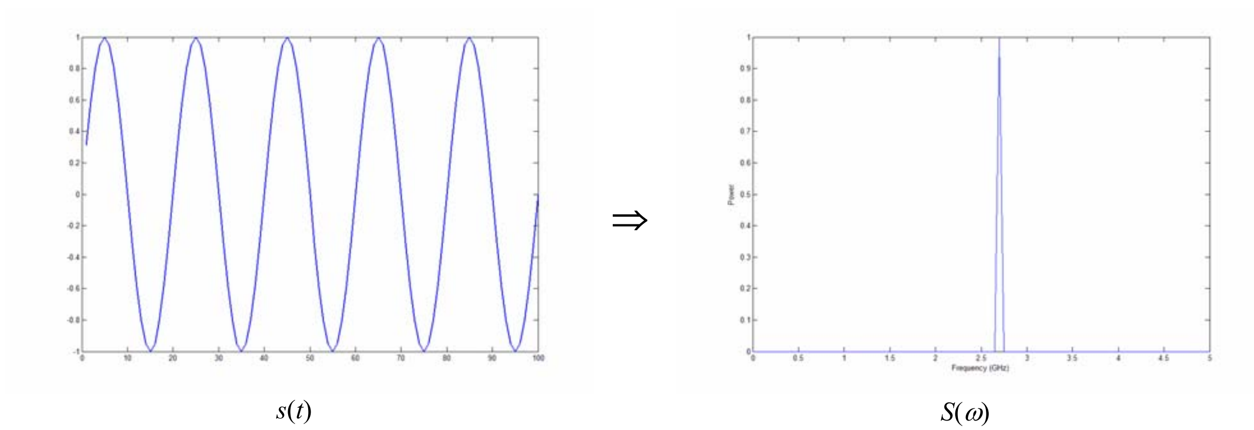


Figure 9. The spectral density (on the right) obtained via Fourier Transform of the signal on the left.

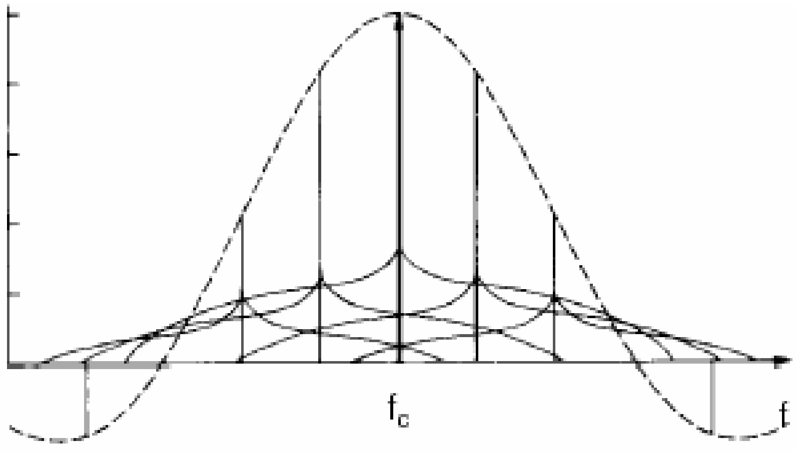


Figure 10. The effect of the phase noise across several PRT intervals. The vertical lines represent the pulses at the PRT interval. The dashed line is the power envelope in frequency space. The solid lines are the noise envelope.

The spectrum and related signal is summarized Figure 9.

Real signals do not present themselves as delta functions in frequency space. This is due to the stochastic processes associated with random noise. We can model the noise as normally stochastic function.

A mathematical model for the sinusoidal signal including the functions modeling the noise is (Scheer, 1993 and Goldman, 1989),

$$s(t) = \left[A_0 + \Delta A(t)\right]\sin\left[(\omega_0 t + \phi_0) + \Delta\phi(t)\right], \quad (10)$$

where the perturbation $\Delta A(t)$ is called the amplitude noise and $\Delta\phi(t)$ is the phase noise. The amplitude noise term is typically 20 dB or more less than the phase noise. Thus, when measuring and estimating the noise contribution to signal degradation, the noise considered is phase noise. However, this assumption is not always true, particularly when relativistic effects are considered in high power systems.

Most weather radars transmit pulses at a constant rate or interval called the pulse repetition frequency PRF and pulse repetition time PRT respectively. In this case, we will have a number of overlapping phase noise curves in each PRT The phase noise spectrum from one pulse enters the subsequent PRT's to increase its noise level. This effect is demonstrated in Figure 10.

Attempting to estimate all of the overlapping contributions within a PRT appears to be quite daunting, but appearances can be deceiving. If we look carefully at one of the PRT sections and consider only the phase noise contributions from it and subsequent PRT intervals, it is clear that the integration of all the noise is in fact the same as integrating the signal in frequency space over the bandwidth,

$$I_{PN} = \int_{-f_{BW}}^{f_{BW}} S(f)\,df \ , \quad (11)$$

Functionally, we can integrate just one-half of the bandwidth and multiply the integral by two,

$$I_{PN} = 2\int_{0}^{f_{BW}} S(f)\,df \ , \quad (12)$$

This integral is the integrated phase noise.

As an example, consider the following spectral function,

$$S(f) = \begin{cases} \dfrac{10^{-4}}{f^3} & 0 \le f < f_{BW}/4 \\ \dfrac{10^{-8}}{f^2} & f_{BW}/4 \le f < f_{BW}/2 \\ \dfrac{10^{-12}}{f^1} & f_{BW}/2 \le f < 3f_{BW}/4 \\ \dfrac{10^{-13}}{f^0} & 3f_{BW}/4 \le f < f_{BW} \end{cases} , \quad (13)$$

The integrated phase noise is,

$$I_{PN} = 2\left[\int_{0}^{f_{BW}/4} \frac{10^{-4}}{f^3}df + \int_{f_{BW}/4}^{f_{BW}/2} \frac{10^{-8}}{f^2}df + \int_{f_{BW}/2}^{3f_{BW}/4} \frac{10^{-12}}{f^1}df + \int_{3f_{BW}/4}^{f_{BW}} 10^{-13}df\right]. \quad (14)$$

Realizing the first integral returns an indefinite result, we start the initial frequency not from zero, but slightly offset by an amount ε. Thus, Eqn. (14) becomes,

$$I_{PN} = 2\left[\int_{\varepsilon}^{f_{BW}/4} \frac{10^{-4}}{f^3}df + \int_{f_{BW}/4}^{f_{BW}/2} \frac{10^{-8}}{f^2}df + \int_{f_{BW}/2}^{3f_{BW}/4} \frac{10^{-12}}{f^1}df + \int_{3f_{BW}/4}^{f_{BW}} 10^{-13}df\right]. \quad (15)$$

Integrating this function we have,

$$I_{PN} = 10^{-4}\left[\frac{1}{\varepsilon^2} - \frac{16}{f_{BW}^2}\right] + 10^{-8}\left[\frac{4}{f_{BW}}\right] + 10^{-12}\left[2\ln\left(\frac{3}{2}\right)\right] + 10^{-13}\left(\frac{f_{BW}}{2}\right). \quad (16)$$

Assuming that ε = 10 Hz and the bandwidth is, f = 1 MHz (10^6 Hz), the integrated phase noise is,

$$I_{PN} = 1.0 \times 10^{-6} \text{ W or } -60 \text{ dBc.} \quad (17)$$

This system would achieve 60 dB of clutter suppression provided the signal to noise ratio is large enough.

This procedure seems simple enough provided we know the functional form of the phase noise. That is precisely the problem. When we take a phase noise measurement, we don't obtain an arithmetic form of the phase noise; rather a graphical form of the phase noise function is presented. If we obtain a set of the data values either from the plot or directly from the measuring device we can perform numerical integration.

Recall that the integration process finds the area between the curve and the axis. This process can be approximated by finding the area under the curve between two data points. As knowing the functional form is required to get an exact area, we can approximate this area by drawing a line between two nearest data points and finding the area of the resulting trapezoid. The area of a trapezoid is given by,

$$A_{trap} = \frac{1}{2}(y_1 + y_2)(x_2 - x_1). \quad (18)$$

If the data samples are close enough together, this technique will provide a good estimate of the curve. The phase noise plots are somewhat complicated because the Spectral function is described in logarithmic units (dBc/Hz). These values must be converted to linear units (W/Hz) before integration.

Delay Correction

Clutter rejection involves transmitting a signal which is scattered off the clutter target located at a distance R (Scheer 1993, Goldman 1989). The scattered signal returns to the radar. The total travel time is given by,

$$T_{delay} = \frac{2R}{c}, \quad (19)$$

where c is the speed of light. Thus, the reference signal is compared to this time delayed signal. Part of the delayed signal remains correlated with the reference signal. The correlated sections will cancel an amount related to their correlation. If the delay is zero, there is no decorrelation and the signal will cancel completely either to the transmitter stability or the Signal to Noise ratio, whichever is less. If the delay is infinite (very long or a very low PRF), the two signals are completely decorrelated and the suppression will be equal to the integrated phase noise. The maximum suppression is between these two limits. The issue is to quantify the correction due to the delay time.

To estimate the form of the correction, we ignore the amplitude noise and look solely at the phase noise. With this approximation,

$$s(t) = A_0 \sin[(\omega_0 t + \phi_0) + \Delta\phi(t)]. \quad (20)$$

The instantaneous phase is given by the argument of the sine function,

$$\theta(t) = \omega_0 t + \phi_0 + \Delta\phi(t). \quad (21)$$

The instantaneous frequency is the derivative of the instantaneous phase,

$$\omega(t) = \dot{\theta}(t) = \omega_0 + \Delta\dot{\phi}(t). \quad (22)$$

The last term, $\Delta\dot{\phi}(t)$, is the frequency modulation of the signal.

Without knowing the exact form of $\Delta\phi(t)$, it is difficult to estimate the correction. To obtain an estimate in closed form, consider a sinusoidal modulation function,

$$\Delta\phi(t) = \gamma \sin(\omega_m t), \quad (23)$$

where γ is the modulation index given by,

$$\gamma = \frac{\omega_d}{\omega_m}, \quad (24)$$

ω_m is the frequency offset from the carrier, and ω_d is the maximum deviation frequency $\left(\frac{\omega_{BW}}{2}\right)$. The instantaneous phase and frequency are thus,

$$\theta(t) = \omega_0 t + \phi_0 + \gamma \sin(\omega_m t). \quad (25)$$

and

$$\omega(t) = \omega_0 + \gamma\omega_m \cos(\omega_m t). \quad (26)$$

Consider the difference between the instantaneous frequency at time t and that delayed by an amount t-τ. Mathematically, this is written,

$$\Delta\omega(\tau) = \omega(t) - \omega(t-\tau). \quad (27)$$

Expanding the difference, we get,

$$\Delta\omega(\tau) = [\omega_0 + \gamma\omega_m \cos(\omega_m t)] - [\omega_0 + \gamma\omega_m \cos(\omega_m (t-\tau))], (28)$$

and consolidating the terms, we have,

$$\Delta\omega(\tau) = \gamma\omega_m [\cos(\omega_m t) - \cos(\omega_m (t-\tau))]. \quad (29)$$

Using the trigonometric identity,

$$\cos(a-b) = \cos(a)\cos(b) + \sin(a)\sin(b), \quad (30)$$

we have,

$$\Delta\omega(\tau) = \gamma\omega_m [\cos(\omega_m t) - \cos(\omega_m t)\cos(\omega_m \tau) - \sin(\omega_m t)\sin(\omega_m \tau)].(31)$$

Simplifying we have,

$$\Delta\omega(\tau) = \gamma\omega_m [\cos(\omega_m t)[1 - \cos(\omega_m \tau)] - \sin(\omega_m t)\sin(\omega_m \tau)].(32)$$

To reduce Eqn. 33 , consider,

$$\Psi = A\cos(a+b). \quad (33)$$

Using the trigonometric identity,

$$\cos(a+b) = \cos(a)\cos(b) - \sin(a)\sin(b), \quad (34)$$

to expand Ψ , we have,

$$\Psi = A[\cos(a)\cos(b) - \sin(a)\sin(b)]. \quad (35)$$

Defining $a = \omega_m t$ and comparing Eqn. 32 with Eqn. 35, we have,

$$\Delta\omega(\tau) = \Psi = A\cos(\omega_m t + b), \quad (36)$$

where,

$$A\cos(b) = \omega_m \gamma [1 - \cos(\omega_m \tau)] \quad (37)$$

and

$$A\sin(b) = \omega_m \gamma \sin(\omega_m \tau). \quad (38)$$

Squaring both of these terms and adding, we have,

$$A^2 = \omega_m^2 \gamma^2 \{[1 - \cos(\omega_m \tau)]^2 + \sin^2(\omega_m \tau)\}. \quad (39)$$

Simplifying, we get,

$$A^2 = 2\omega_m^2 \gamma^2 [1 - \cos(\omega_m \tau)]. \quad (40)$$

Thus the instantaneous frequency deviation between the signal and a delayed version of the signal is,

$$\Delta\omega(t,\tau) = \omega_m \gamma \sqrt{2[1 - \cos(\omega_m \tau)]}\cos(\omega_m t + b) \ . \quad (41)$$

Integrating this give us the instantaneous phase deviation,

$$\Delta\phi(t,\tau) = \gamma \sqrt{2[1 - \cos(\omega_m \tau)]}\sin(\omega_m t + b). \quad (42)$$

Essentially, Eqn. 42 states that a signal mixed with a delayed version of itself has its modulation index, γ, modified by a function dependent on the delay. It is this delay function that will provide the delay correction to the spectral function. The delay correction function to the spectrum will be,

$$\Gamma(\omega_m, \tau) = 10\log[2\{1 - \cos(\omega_m \tau)\}]. \quad (43)$$

The correction will be valid up to an offset frequency, ω_m, that maximizes $1 - \cos(\omega_m \tau)$. This frequency, called the break frequency, will be when $\omega_m \tau = \pi$ or $\omega_m = \frac{\pi}{\tau}$. By definition, $\omega = 2\pi f$, so the break frequency in Hz is,

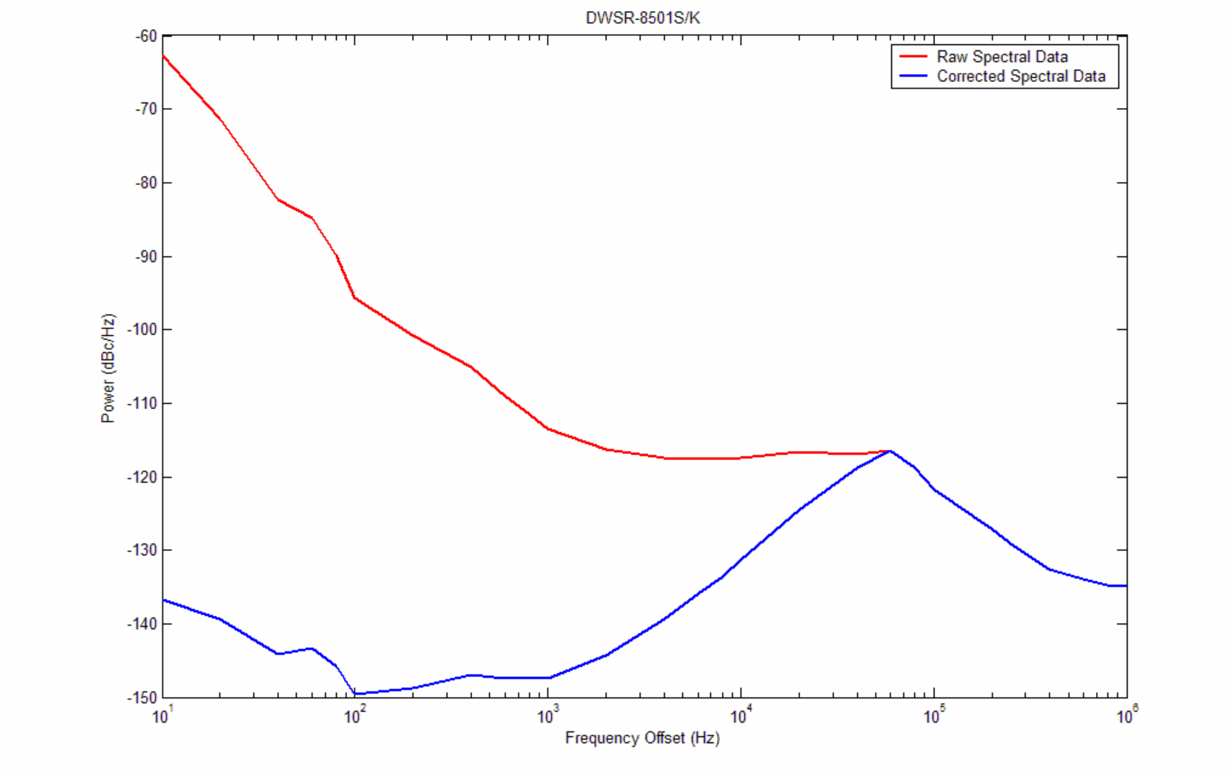


Figure 11. The impact of the delay correction on the phase noise spectrum. The red line is the measured phase noise. The blue line is the phase noise spectrum after the application of the delay correction.

$$f_b = \frac{1}{2\tau}. \quad (44)$$

Figure 11 shows the effect of applying the correction to the phase noise spectrum. The solid red line is the measured phase noise spectrum. The blue line is the phase noise spectrum with the delay correction applied. As we can clearly see, the impact of low frequency offsets is dramatically reduced.

Table V summarizes the results of the correction applied to the oscillators listed in Table IV. Figure 12 is the corresponding corrected spectrum plots.

Once the correction is applied, it is clear that the phase noise performance at high offset frequencies is of much greater interest for clutter rejection than those at low frequency. In addition, from the integrated phase noise, we find that the WSR-88D legacy oscillator as well as the EEC oscillator can easily support the measured performance. A result of this analysis is that as the clutter appears farther out in range (longer delay time), the ability of the system to cancel it is significantly degraded. This also is important for evaluating the specifications of a weather radar system.

Table V. Clutter suppression as measured with a 10 μs delay line and integrated phase noise for various synthesizers. Delay correction applied to the data.

STALO	Clutter Suppression (dB)	Integrated Phase Noise (dBc)	Integrated Phase Noise w/ Correction (dBc)
WSR-88D	65	-29	-75
OSC - 1	52	-54	-57
OSC - 2	38	-40	-45
EEC Oscillator	66	-54	-74

5. CONCLUSION

In this paper, we looked at the relationship between system phase noise and clutter rejection. We noted that the major contributors to a system's phase noise is the power supply, modulator, and tube

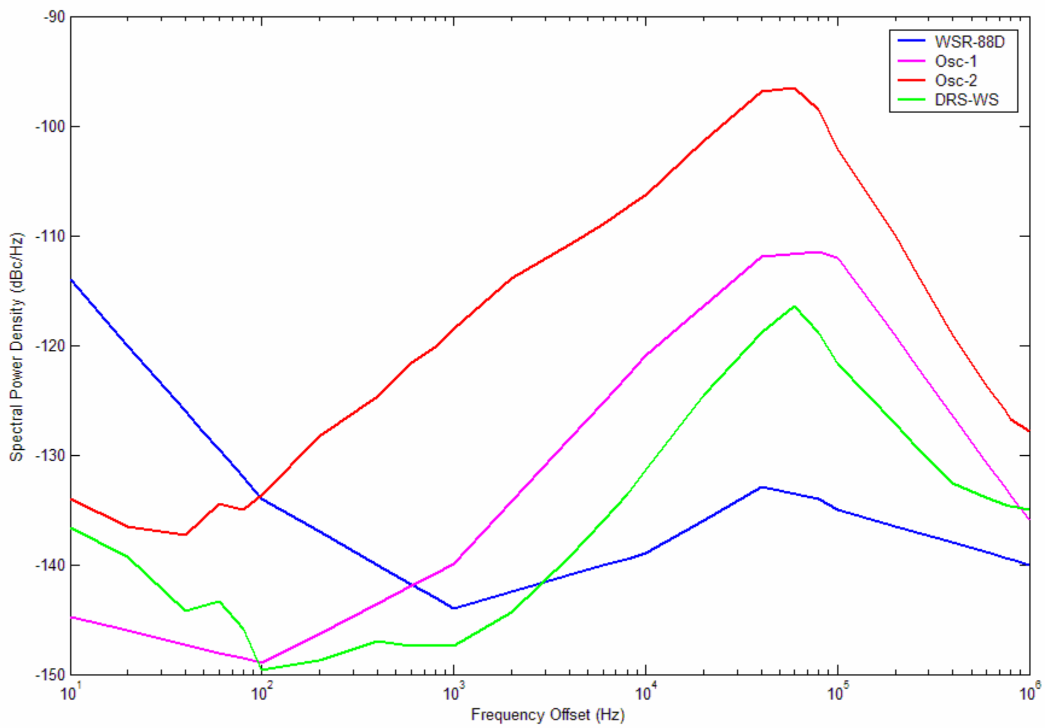


Figure 12. The corrected phase noise spectra for the oscillators listed in Table V. The WSR-88D legacy oscillator is in blue, Oscillators 1 and 2 are magenta and red respectively, and the EEC oscillator delivered with the Evansville DWSR-8501S/K is in green.

combination and the STALO. We used STALO's with different phase noise characteristics, transmitted pulses through a 10 μs delay line (corresponding to clutter located a mere 1.5 km from the radar, and determined the resulting clutter suppression.

From the data and analysis we are able to conclude that the relationship between system phase noise and clutter rejection is very complicated. First and foremost there needs to be sufficient signal with respect to noise for maximum clutter rejection. The signal to noise ratio (SNR) provides a limit to the amount of clutter rejection available.

Provided the SNR is great enough, the maximum clutter rejection is related to the integrated phase noise *corrected for the delay*. The net result, for STALO design is that the phase noise spectral power density levels near the carrier frequency are not as important for clutter rejection as those at greater frequency offsets from the carrier. Thus in designing a low noise STALO, more effort should be expensed at reducing high frequency noise.

6. REFERENCES

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