

Pulsed Phase Noise Measurements

Application Note

Products:

- R&S®FSWP

With advances in digital signal processing techniques modern wireless communications systems and radar systems have become increasingly digital and more constraints have been placed on packing more information into less and less bandwidth. On the wireless side, modulation formats have become more complicated with less margin for symbol error. On the radar side, modern radar systems are trying to extract more information about the targets they track and provide the ability to track slow moving targets such as automobiles and personnel in the presence of clutter.

For both communications and radar systems one of the key RF parameters that limits system performance is phase noise. As systems engineers develop new and more advanced RF systems, phase noise of oscillators and transmitters can no longer be overlooked.

This application note will focus on phase noise measurements for pulsed RF carriers such as those used for radar systems. We will address some of the physical limitations relating to pulsed phase noise measurement and introduce the capabilities of the new Rohde & Schwarz FSWP phase noise analyzer.

Note:

Please find the most up-to-date document on our homepage <http://www.rohde-schwarz.com/appnote/1EF94>.

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1 Introduction

The word radar is an acronym standing for radio detection and ranging. From this we know that the primary purposes for radar systems is to detect targets and determine their range or distance. Other key characteristics of the target may also be of interest such as: location, bearing from the radar, velocity of the target, and size of the target or possibly even some form of target identification.

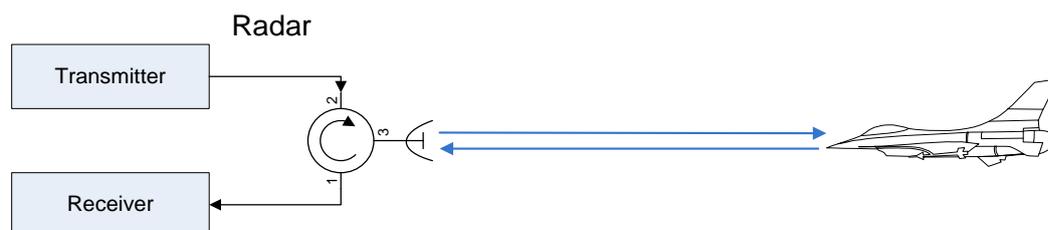


Figure 1. Simplified Radar Concept

Most radar systems employ pulse modulation, transmitting a short burst or pulse of RF energy then wait for the signal to travel to the target, reflect off of the target, and return to the radar system. By measuring the total elapsed time for the signal to travel to and return from the target the distance to the target can be determined by knowing the velocity of propagation of the RF signal. The bearing to the target can be determined by using a narrow-beam antenna and measuring angular displacement of the antenna from North to the target.

1.1.1 Radar Velocity Measurements

Before getting started with a discussion of pulsed phase noise, it would be good to provide some background relating to why pulse phase noise is important to radar systems engineers. The following discussion will provide some examples of typical radar applications illustrating what frequency offsets may be of importance.

The velocity of radar targets can be determined by measuring the Doppler shift of the radar's received signal relative to that of the transmitter. Essentially, if the target is moving toward the radar the effective wavelength of the transmitted signal is reduced or compressed by the relative motion of the target during each RF cycle. The Doppler frequency for a moving target can be expressed as:

$$f_d = \frac{2f_0}{c} (\vec{v} \cdot \hat{R}) = \frac{2f_0 v}{c} \cos \theta$$

Where:

- f_d = Doppler frequency
- f_0 = radar transmit frequency
- c = speed of light

v = target velocity vector

R = the radial unit vector relative to the radar antenna

Θ = the angle between the velocity vector of the target and the radar

Target velocity is not only important for predicting target track, but in many cases it is more important to differentiate moving targets from those that are stationary. Radars with the capability of tracking moving targets in the presence of large stationary targets such as buildings, mountains, and other forms of ground clutter are known as moving target indicator (MTI) radars. This becomes very important when we consider that the radar returns from clutter can be many times larger than that of a typical target of interest, sometimes as much as 80 to 100 dB larger.

A couple of simple examples might illustrate the Doppler frequency ranges of interest to a radar systems engineer. Let's consider two examples:

- First, let us consider a small low-velocity general aviation aircraft as seen by an airport surveillance radar.
- Second, we will consider the case of two fighter aircraft in a nose to nose engagement

Case 1:

For this example consider a Cesena 150 aircraft that is making an approach at a full-service airport equipped with an airport surveillance radar operating at a frequency of 2.7 GHz. Assume that the aircraft is flying at a speed of 80 km/Hr. on a direct radial to the radar (no cosine term in the Doppler equation.) Therefore:

$$f_d = \frac{2f_0V}{c} = \frac{2(2.7 \times 10^9)(8 \times 10^4 \frac{m}{Hr})}{(3 \times 10^8 \frac{m}{s})(3600 \frac{s}{Hr})} = 400 \text{ Hz}$$

Case 2:

For this example consider a fire control radar used on a fighter aircraft. Assume a radar with a transmitter frequency of 10 GHz that is designed to handle a maximum closing velocity of 500 m/s (~Mach 1.5). The maximum Doppler frequency for this radar would be:

$$f_D = \frac{2V}{\lambda} = \frac{2(500 \text{ m/s})(10 \times 10^9 \text{ Hz})}{3 \times 10^8 \text{ m/s}} = 33 \text{ kHz}$$

From these examples we can see that common Doppler frequencies for many radars are primarily in the audio range and that they are often very close to the carrier. We can also see that if the oscillators used in the radar systems have high phase noise in this frequency range it will limit the radar's ability to determine the velocity of the target. This is very important when we consider that the radar target returns from large clutter blocks also have the phase noise of the radar transmitter impressed on them and will generally overpower the small return from targets of interest.

So, as we can see, phase noise is very important to the radar systems engineer, but since most radar systems utilize pulse modulated carriers the ability to measure the phase noise of pulse carriers is fundamentally necessary.

Next we will review some of the basic characteristics of pulse modulated RF signals.

1.2 Characteristics of Pulse Waveforms

A good place to begin thinking about the characteristics of pulsed RF is to start with a basic pulse or even better, a square wave. Recall that a square wave can be created by plotting a sine wave and its odd harmonics. If you add up enough harmonics at the correct amplitudes you can start to see the square wave take shape.

In the following figure we have plotted a sine wave, its third, fifth, and seventh harmonics with the amplitude of each harmonic reduced by $1/n$, where n is the harmonic number.

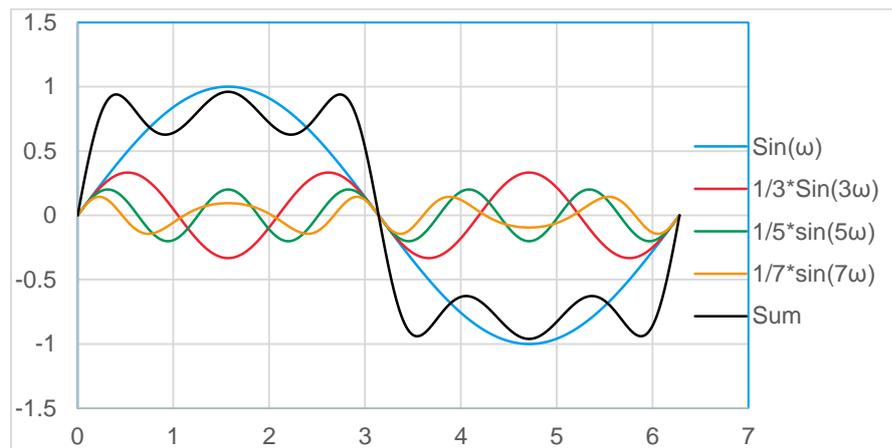


Figure 2. A sine wave and its odd harmonics

As you can see by the black trace, the sum of fundamental and the first three odd harmonics starts to pretty closely resemble a square wave. If we were to continue this exercise, by adding more odd harmonics, we could generate a nearly perfect square wave.

Creating a rectangular pulse is simply a continuation of this process where we force all of the harmonics to go through a positive or negative maximum at the same time as the fundamental. To create a perfect rectangular pulse we would need an infinite number of harmonics, but as you can see from our example above the amplitude of the higher order harmonics taper off pretty fast so we can build a reasonable pulse with a limited number of harmonics.

To create a pulse train whose amplitude varies from 0 Volts to some positive value we will need to add a DC component to compensate for the negative going portion of the sine waves we used to create the pulse train. This DC component represents the average value of the waveform.

As we think about pulsed waveforms it is also useful to think about their frequency content, basically the reverse process we have just gone through.

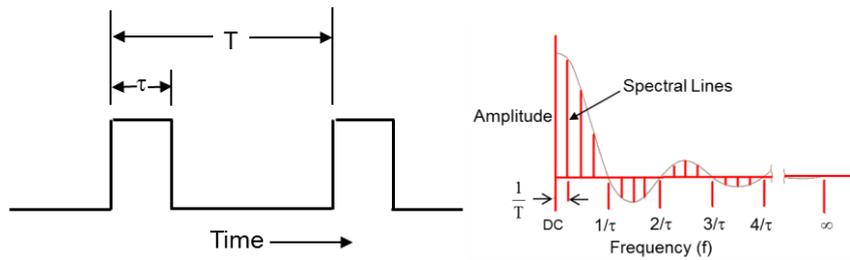


Figure 3. A pulse waveform and its Fourier transform

For the pulse train shown above two key parameters are the pulse width designated by the Greek character τ and the pulse period or pulse repetition interval (PRI) represented by the letter T.

Other key parameters describing the pulse train, shown in Figure 3 are:

Amplitude	The peak amplitude of the pulse is usually designated by the character A and generally is expressed in units of Voltage
PRF	The pulse repetition frequency (PRF) is the number of pulses per second and is equal to 1 divided by the pulse repetition interval ($1/T$)
duty cycle	The duty cycle or duty factor is the ratio of pulse width divided by the period (τ/T)
A avg.	The average amplitude of the pulse waveform. $A_{avg.} = A \times \text{duty cycle}$

In Figure 2, above, we plotted the various harmonics of a sine wave to obtain a time-domain view of a square wave. In Figure 3 we have effectively done the reverse process to obtain the frequency spectrum of a pulse train. Since all of the frequencies are integer multiples of the fundamental PRF the spacing between the frequency lines is equal to the PRF. The shape takes on the familiar $\sin(x)/x$ form, where the nulls in the spectrum occur at a spacing of $1/\tau$.

Since pulsed waveforms are periodic they can be expanded into a Fourier series to determine their frequency content, as follows:

$$V(f) = \frac{\tau}{T} + \frac{2\tau}{T} \sum_{n=0}^{\infty} \frac{\sin(n\pi\tau/T)}{n\pi\tau/T}$$

Where τ/T is the duty cycle or the average value of the waveform.

1.3 Pulsed RF Waveforms

Pulse modulated RF is produced by amplitude modulating the RF carrier with the desired pulsed waveform. Pulsed carriers can also be produced by gating an oscillator on and off; however, gated oscillators, such as magnetrons, are not phase continuous causing the starting phase of each pulse to be random. This random nature limits gated oscillators use in radars that process Doppler. In addition, the random phase

relationship from pulse to pulse generally prevents phase noise measurements on these sources when techniques such as phase detectors or synchronous detectors are used. Since the phase detector requires a stable reference to compare the measured signal to, it is impossible to measure phase noise with the phase of each pulse changing by large amounts.

Because of the limitations of gated oscillators we will focus our discussion of pulsed phase noise measurements to pulsed sources that use amplitude modulation to impress the pulse modulation on the RF carrier. Amplitude modulation results in the desired effect of producing an RF pulse train with continuous phase.

When thinking of the modulation process it is beneficial to remember that multiplication of the pulse waveform and the RF, carrier in the time domain, is the same as convolution of the pulse spectrum and the spectrum of the RF carrier in the frequency domain.

From AM modulation theory we know that when a carrier is modulated with a single tone a modulation side tone is produced on each side of the carrier. This can be verified by analyzing the trigonometry identity that describes the multiplication of two sine waves. Since the spectrum of the pulse train is a series of harmonics of the PRF we can then deduce that the AM modulation of a carrier with a pulsed waveform will result in a carrier with modulation sidebands on each side of the carrier. As mentioned above, the modulation process will convolve the carrier spectrum with the spectrum of the pulsed waveform, essentially sliding the pulse spectrum up to the carrier frequency. This will result in a spectrum that is symmetrical on either side of the carrier and is composed of spectral lines spaced $n \cdot \text{PRF}$. The amplitude of the spectral lines will have the expected $\sin(x)/x$ shape with the spectral nulls occurring at $1/\tau$.

The following diagram illustrates what happens as we hold the PRF constant and change the pulse width. This change is shown as a change in the lobe width of the spectrum and is seen by moving left and right on the diagram. Additionally, the change in the spectrum due to changing the PRF is shown by moving up and down in the diagram. Notice as the PRF is decreased the density of the spectral lines is increased.

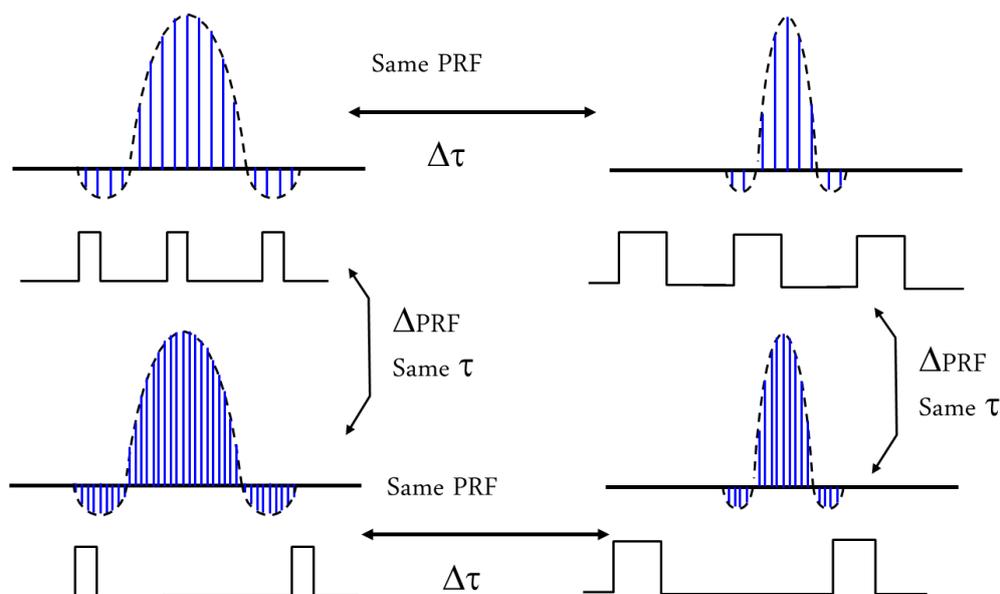


Figure 5, Spectrum of a pulse modulated carrier

From the above figure, the phase noise side bands of the carrier can be clearly identified and it can be noted that the carrier's phase noise has been impressed on each of the PRF lines of the pulsed RF spectrum. Notice also that a characteristic bath-tub curve is formed between each line of the spectrum. From this we also notice that there is no new phase noise information in the spectrum for frequency offsets greater than PRF/2. Because of this, phase noise measurements of pulsed carriers are usually made between the carrier and a maximum offset frequency less than or equal to PRF/2.

The general practice for pulsed phase noise measurements is to insert a low-pass filter following the phase detector. The filter has a corner frequency less than PRF/2 and ensures the phase noise measurement is made over a valid offset frequency range.

1.4.1 Differences Between Pulsed and CW Phase Noise

When comparing phase noise of a pulsed RF source to that of the same RF source without pulse modulation the first question that is generally asked is: Why is the phase noise of the pulsed carrier higher than for CW, particularly at higher offset frequencies?

Below is a graphic showing the phase noise of a signal generator with and without pulse modulation.

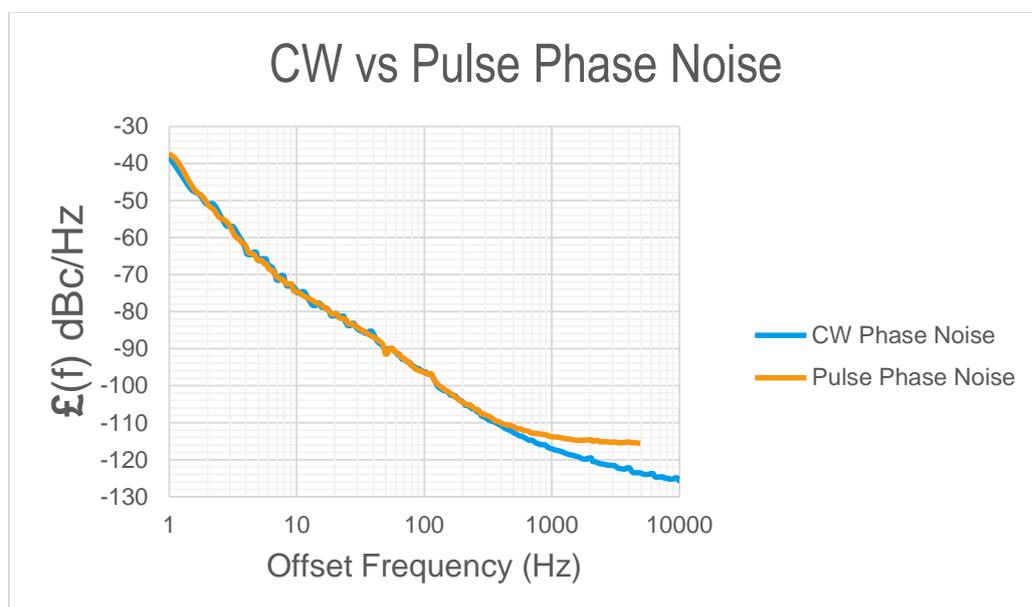


Figure 6, CW vs pulsed phase noise

The above figure shows the CW and pulse phase noise of a 1 GHz carrier, as measured on a Rohde & Schwarz FSWP phase noise analyzer. The pulse modulation had a pulse width of 10 μ s with a PRF of 10 kHz. Note that the maximum offset frequency shown for the pulsed measurement is 5 kHz (PRF/2). From the figure, you

can see that the pulsed phase noise starts to deviate from the CW phase noise at an offset of approximately 400 Hz. and is about 8 dB higher at PRF/2 (5 kHz). Comparing this graph to the spectrum we measured in Figure 5 you can see a similar shape starting in the curve.

Recalling our explanation of the AM modulation process of a CW carrier with a pulse waveform, the sine wave represented by each spectral line in the pulse spectrum (see Figure 3) produces two side bands, an upper sideband and a lower sideband. The convolution of the individual spectra produces all possible sums and differences of the CW carrier and all of the harmonic components of the modulating pulse. If phase noise sidebands are present on the carrier and the carrier is pulse modulated the spectrum of the pulsed waveform is convolved with the carrier and its noise sidebands.

Essentially, the modulation process aliased the noise sidebands of the carrier onto each of the PRF lines of the pulse spectrum weighted by a $\sin(x)/x$ function, see Figure 8.

Note: No matter how great the offset between the CW carrier and the phase noise sidebands of the carrier, an alias of the noise sidebands will appear within an offset of range of $\pm PRF/2$ from the carrier and will not be filtered off by the PRF/2 low-pass filter.

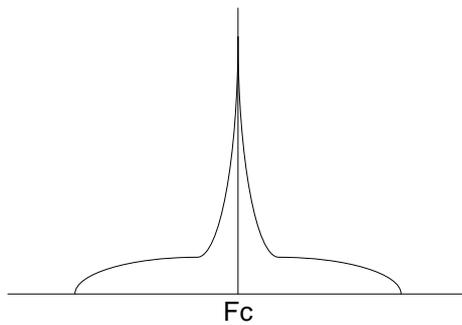


Figure 7, Spectrum of a carrier and its noise sidebands

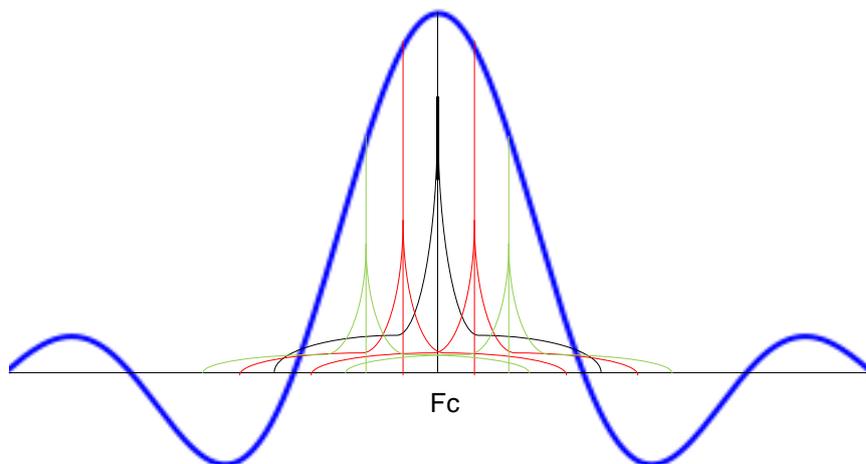


Figure 8, Spectra of noise sidebands aliased onto each PRF line

Note: For a fixed pulse width the increase in noise at F_c will be inversely proportional to the PRF (low PRF corresponds to high spectral line density and more noise; whereas, increasing the PRF will decrease the noise at F_c). Also, for a constant PRF increasing the duty cycle results in decreased noise at F_c —due to narrowing of the spectral lobes. From a worst-case standpoint the noise will increase by:

$$\text{Noise increase} \leq 10 \log_{10} (\text{Number of PRF lines to the first } \sin(x)/x \text{ null})$$

As mentioned above, this is a worst-case approximation and assumes that the noise contribution at each PRF line out to the first $\sin(x)/x$ null is equal. The reason this is not an exact relationship is because of the spectral shape of the phase noise sidebands. Generally, near the carrier the noise level is falling very fast, probably 20 to 40 dB per decade. Because of the $\sin(x)/x$ envelope of the pulse modulated spectrum aliased noise at these offsets will be well below the CW noise; however, at higher offset frequencies, up to $\text{PRF}/2$, the degradation will be more apparent, especially if the CW noise curve has a pedestal. Since a pedestal in the phase noise curve represents a relatively constant energy level over a range of offset frequencies the combined energy of the aliased noise will be greater than for a region with constantly decreasing slope.

1.4.2 Trying to Visualize How Noise Changes with Pulse Parameters

From the discussion above, you can see how the phase noise of a pulsed carrier can change as a function of PRF and pulse width. This can best be visualized by looking at an actual phase noise measurement at different pulse widths.



Figure 9, Differences in phase noise caused by changing pulse width

Figure 9, above, shows three phase noise measurements of a pulsed carrier superimposed over one another. All of these measurements were made at a constant PRF of 10 kHz.

The first measurement was made using a 10 μ s pulse width and is shown on the blue (middle) trace of Figure 9. For the second measurement the pulse width was increased to 50 μ s and is shown on the green (or lower) trace. From the trace you can see that the phase noise decreased about 6 dB. For this measurement, the pulse width was increased by a factor of 5 which decreased the width of the pulsed spectrum main lobe by a factor of five also decreasing the number of PRF lines under the main lobe by a factor of five. Using the equation shown above we can see that we could expect a maximum reduction in phase noise for the above scenario of:

$$10 \cdot \text{Log}_{10} (1/5) = -6.9\text{dB}$$

Again, this is simply a worst case approximation of the change in phase noise.

For the last measurement shown in Figure 9 the pulse width was reduced to 1 μ s. Therefore, the pulse width was reduced by a factor of 10 from the first measurement (blue trace) which increased the width of the main lobe of the $\sin(x)/x$ spectrum by a factor of 10 also increasing the number of PRF lines under the main lobe by a factor of ten, which should increase the phase noise of the carrier, see yellow (top) trace of Figure 9.

From this it is apparent that one should not expect the pulsed carrier phase noise of an oscillator to be the same as its CW phase noise. In addition, it is also apparent that engineers and technicians making phase noise measurements of pulsed carriers should have a good understanding of their signal generation chain and the pulse modulation used to fully understand if their measurement results reasonable.

2 Rohde & Schwarz FSWP

The Rohde & Schwarz FSWP is a self-contained one-box cross correlation phase noise analyzer. The FSWP can perform phase noise and AM noise measurements on CW as well as pulsed carriers for carrier frequencies up to 50 GHz, without additional external hardware. In addition to phase noise measurements the FSWP, with Option B1, provides a full featured signal and spectrum analyzer with up to 80 MHz of digital analysis bandwidth.



Figure 10, FSWP Phase Noise Analyzer

The following FSWP options can be added to extend the analyzer's performance and capabilities:

Option	Description
B1	Signal and Spectrum Analyzer
B4	High stability OCXO
B60	Cross correlation
B64	Residual (two-port) phase noise measurements
B80	80 MHz analysis bandwidth
K4	Pulsed phase noise measurements Note: This option is required for pulse phase noise measurements
K6	Pulse measurements--detailed pulse analysis and trends--not necessary for phase noise measurement
K7	Analog modulation analysis for AM, FM, and ϕ_m
K30	Noise figure measurements
K70	Vector Signal Analysis

The R&S FSWP brings a new level of performance, speed, and ease of use to phase noise measurements. Most measurements can be made by simply connecting the signal to the analyzer and pressing a button. Once the measurement has completed both AM and phase noise traces can be displayed, this is also the case for pulsed carrier noise measurements. In the case of pulsed carriers, the analyzer automatically measures the RF carrier frequency, pulse width, and PRF of the signal and then configures the analyzer to make the measurement; without user intervention.

2.1 Phase Noise Measurement with the FSWP

Before getting into how the FSWP performs a phase noise measurement it would be good to review the classic phase detector method of phase noise measurement.

2.1.1 Phase Detector Method of Phase Noise Measurement

Generally, high performance phase noise measurements are made using a calibrated phase detector, where phase deviations around the carrier are converted to a voltage. Following the phase detector the detector's output is filtered with a low-pass filter and amplified before being fed to a high-performance analog to digital converter or Fourier analyzer.

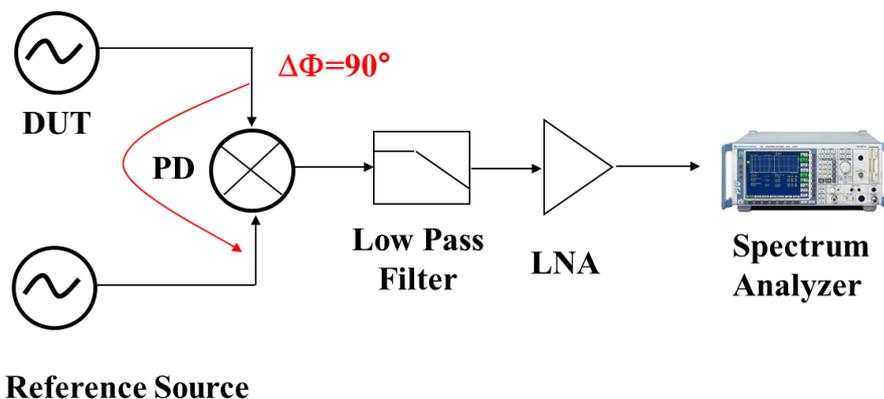


Figure 11, Phase detector method for phase noise measurement

In most cases a mixer is used as the phase detector with the device under test (DUT) connected to the input port of the mixer and a reference oscillator connected to the LO port. The reference oscillator is set up to be at the same frequency as the DUT and in phase quadrature. With the two input signals in quadrature the DC output voltage of the mixer is zero and short-term phase fluctuations are converted to an AC voltage. The Fourier analyzer then displays the spectral density of phase fluctuations which is converted to phase noise.

The above technique is often referred to as the reference oscillator phase lock loop method because it is generally necessary to provide a feedback path from the phase

detector's output to the reference oscillator to maintain a quadrature relationship between the DUT and the reference oscillator.

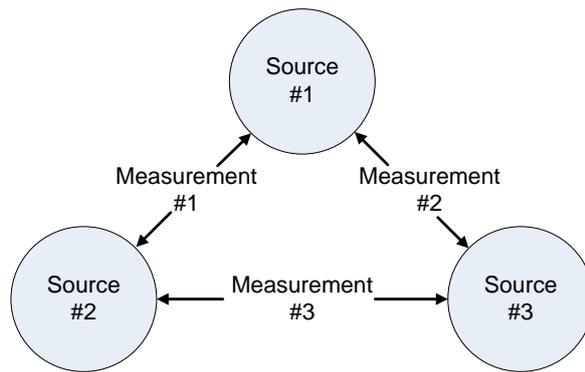
For pulsed carrier phase noise measurements the above block diagram provides some limitations. When the pulsed carrier is compared to a CW reference oscillator the output voltage of the phase detector also becomes pulsed and will contain an average DC component (due to the duty cycle of the pulsed waveform) that often saturates the low-noise amplifier. To prevent this condition it is generally necessary to also pulse modulate the reference source with the same pulse waveform that is used to modulate the DUT, essentially time-gating the measurement.

Adding pulse modulation to the reference source adds the complication that the user is often required to view the output of the phase detector on an oscilloscope to ensure that the DUT and reference source pulse modulation are synchronized. If these modulation signals are not perfectly synchronized the output of the phase detector will exhibit so called "rabbit ears" and video feed through which can also overdrive the low-noise amplifier.

In addition, the user must also select an appropriate low-pass filter to filter off the higher offset frequencies that are greater than $PRF/2$. For most phase noise test systems the filters are generally analog filters and in many systems the filter needed may not be available by a simple selector settings. For these cases where the correct $PRF/2$ filter isn't available the user is forced to use an external phase detector and an external low-pass filter, which greatly increase the chances of external noise corrupting the phase noise measurement.

2.1.2 Cross Correlation Phase Noise Analyzers

One of the limitations of the reference oscillator phase lock loop method of phase noise measurement is that phase noise performance of the system is limited by the phase noise of the reference oscillator. The common rule of thumb is that the phase noise of the reference oscillator must be about ten times better than the device being measured. This can often be a serious performance limitation. One way around this limitation has been to use two identical oscillators and assume that the actual phase noise is 3 dB lower than measured. In many cases the identical source concept is not valid and a better solution is desired. For many years the better way was to make three phase noise measurements on three similar oscillators moving each of the oscillators from the reference position to the DUT position. Following these measurements the phase noise of each oscillator could be determined by solving three equations in three unknowns.



The Three-source comparison determines the noise of each source, provided each source is comparable (3 to 6 dB difference)

Figure 12, Three-source comparison method

Continuing the quest for a simpler phase noise measurement technique with an improved noise floor, engineers recognized that if the DUT signal were split and fed to two independent reference oscillator phase lock loop systems the internally generated noise of the measurement system would be uncorrelated in both channels while the noise from the DUT would be correlated in both channels. With this concept cross correlation could be used to reduce the noise generated by the instrumentation.

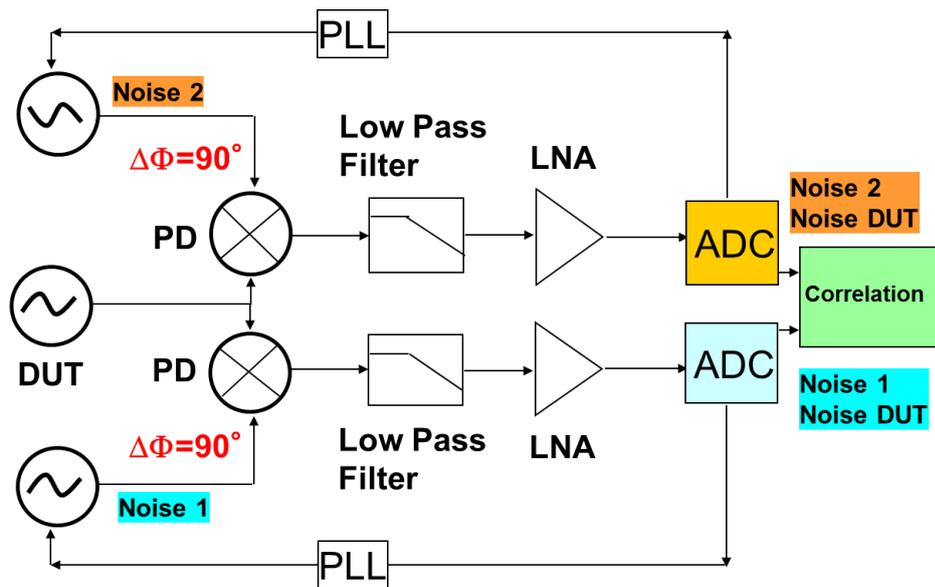


Figure 13, Cross correlation phase noise measurement

In a cross correlation phase noise measurement system, as depicted in Figure 13, all of the instrumentation noise from Channel 1 is uncorrelated with that generated in Channel 2; therefore, when cross correlation is performed on the outputs of the two independent channels only the DUT signal correlates and measurement noise is reduced by $5 \text{ Log}_{10} (M)$, where M is the number of correlations. Using this technique we can reduce the system noise by 10 dB if we perform 100 cross correlations.

Phase detector based cross correlation systems have been leading the industry for the past decade or so, but they still have the disadvantage that the phase noise measurement still is dependent on analog components and their limitations as signal processing elements. In an effort to further improve the state-of-the-art of phase noise measurements Rohde and Schwarz developed a new phase noise analyzer that moved the majority of the signal processing to the digital domain where performance would be more repeatable and system level calibration could be simplified.

2.1.3 FSWP Block Diagram

As mentioned previously, the R&S FSWP phase noise analyzer is a cross correlation analyzer. Like the cross correlation system mentioned above, the FSWP splits the input signal and feeds it to two independent measurement channels for processing, but that is where the similarity ends.

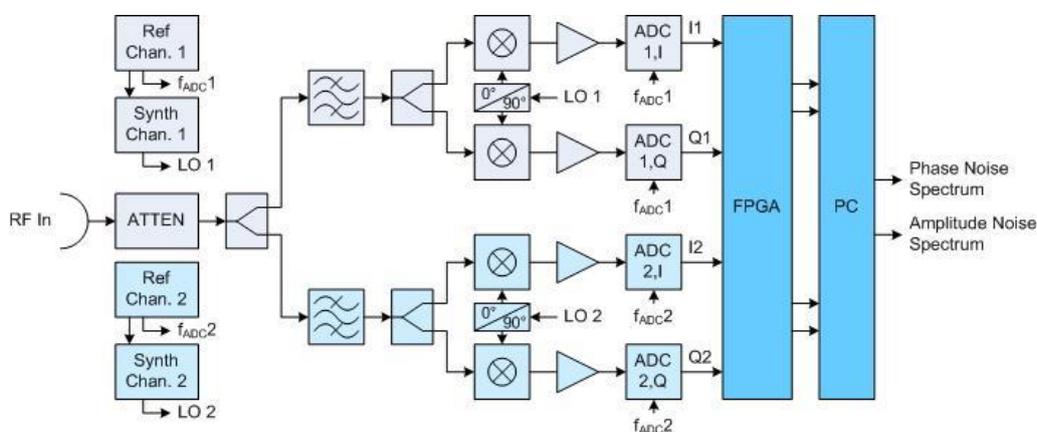


Figure 14, FSWP RF block diagram

Figure 14 shows a simplified block diagram of the FSWP. From the diagram the two cross correlation channels can clearly be identified, with dedicated local oscillators and reference oscillators shown on the left. As with other cross correlation systems the signal is split between channel 1 and 2. Following the splitter the signal for each channel is filtered by a band-pass filter and then fed to the input of an I-Q mixer. The analog I-Q mixer using an extremely low noise internal reference oscillator shifts the signal to a low frequency or zero IF, depending on the offset frequency to be measured. The outputs of the channel 1 and 2 I-Q mixers are fed to either a low-noise or limiting amplifier and then to a dedicated 100 Msa/s ADC. The outputs of all four ADCs are then fed to an FPGA and PC software for further signal processing.

The complex baseband signals from each channel, I and Q, are sampled and digital signal processing is performed in real time on an FPGA.

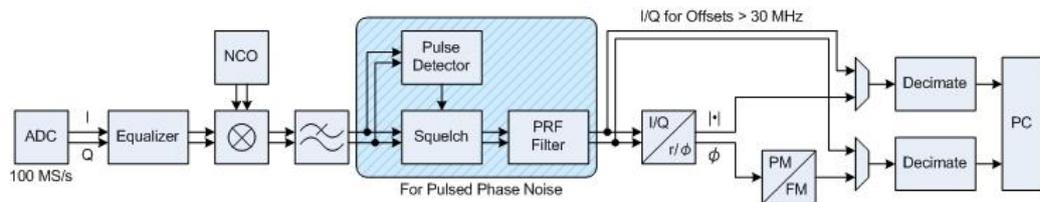


Figure 15, FSWP FPGA signal processing

Figure 15 shows the overall flow of the FPGA processing for one channel. The signals from the I and Q digitizers are next equalized and fed to a digital down converter that provides an I-Q stream for subsequent signal processing. The combination of an analog I-Q mixer and a digital equalizer keeps the AM rejection above 40 dB compared to 15 to 30 dB of a traditional analog PLL, thus reducing the chance of cross-spectrum collapse.¹ Cross-spectrum collapse can occur when AM noise contaminates both inputs to a cross-spectrum analyzer and becomes phase inverted due to the non-ideal behavior of mixers that are used as phase detectors.

The digital down converter provides precise I and Q signals that are not corrupted by common I-Q demodulator impairments such as I-Q imbalance and quadrature errors. Following the digital down converter the signal is low-pass filtered to remove higher order mixing products. From here the signal is routed to the pulse detector and squelch circuit that is used for pulsed carrier measurements, we will describe this processing in more detail in a few paragraphs. For CW carriers, the signal is fed to a digital FM demodulator, which would replace the traditional phase detector in the reference oscillator phase lock loop system shown in Figures 11 and 13.

The digital FM demodulator provides low frequency drift information to the digital down converter to correct for drift of the DUT signal and provides short-term frequency fluctuations to the instrument PC for FFT processing into the spectral density of frequency fluctuations $S_v(f)$ which is easily converted into the spectral density of phase fluctuations $S_\phi(f)$ and single sideband phase noise. In addition, a digital AM demodulator operates in parallel with the FM demodulator providing concurrent measurements of amplitude and phase noise.

2.1.4 Pulsed Carrier Phase Noise Measurements with FSWP

As mentioned previously, several issues complicate phase noise measurements of pulsed carriers, such as: lack of unique phase noise information for offset frequencies greater than $PRF/2$, the need for a low-pass filter at $PRF/2$, and problems with pulse modulating the reference source to minimize overdrive of the low-noise amplifier. One additional problem worth mentioning is that of pulse desensitization.

¹ Nelson, C.W.; Hati, A.; Howe, D.A., "A collapse of the cross-spectral function in phase noise metrology", Rev. Sci. Instrum., vol. 85, 2014.

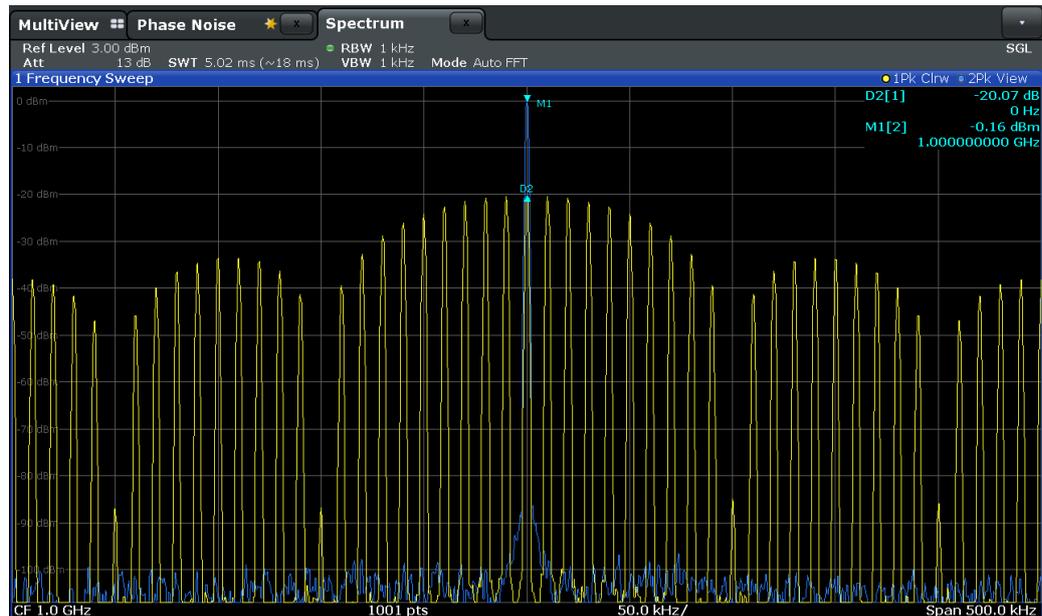


Figure 16, Spectrum showing pulse desensitization

Figure 16 shows the spectrum of a CW carrier (blue trace) and the spectrum of the same carrier with pulse modulation (yellow trace). Notice that the center line of the pulse spectrum has been reduced in amplitude by 20.07 dB from the CW carrier. Pulse desensitization occurs because the pulse modulation on the carrier spreads the spectral energy over a very large frequency span. The total signal power is still the same, it's just been spread over a much wider span (carrier and all sidebands). The difference between the carrier power when modulated and unmodulated is often referred to as pulse desensitization factor, where:

$$\text{Pulse desensitization (dB)} = 20 \text{ Log}_{10}(\tau/T) \text{ or } 20 \text{ Log}_{10}(\text{duty cycle})$$

For the pulsed spectrum shown in Figure 16, the pulse width was 10 μs and the PRI was 100 μs and as we can see the pulse desensitization is 20 dB.

How does this affect measurement of phase noise? For the signal we have been looking at the pulse modulation reduced the carrier power by 20 dB which moves the measurement 20 dB closer to the analyzer's noise floor. As may be recalled from phase noise theory, thermal noise can limit the extent to which you can measure phase noise. Thermal noise, as described by kTB at room temperature, is -174 dBm/Hz. Since phase noise and AM noise contribute equally to kTB , the phase noise portion of kTB is equal to -177 dBm/Hz (3 dB less than the total kTB power).

Theoretical KTB limits to phase noise measurements for low-level signals

P_s (dBm)	$\mathcal{L}(f)$ dBc/Hz
+10	-187
0	-177
-10	-167
-20	-157

If the power in the carrier signal becomes a small value, for example -20 dBm, the limit to which you can measure phase noise is the difference between the carrier signal power and the phase noise portion of kTB (-177 dBm/Hz - (-20 dBm) = -157 dBc/Hz). Higher signal powers allow phase noise to be measured to a lower dBc/Hz level.

The bottom line is that pulse desensitization decreases the measurement sensitivity of phase noise

measurements and systems using a phase detector require that a pulsed phase detector constant be used to compensate for the change in output level of the phase detector.

2.1.4.1 FSWP Pulse Carrier Signal Processing

The above discussion of pulse desensitization was included to clarify one important difference between the FSWP and classic phase noise measurement systems. From the FSWP DSP diagram shown in Figure 15 a sampling of the pulse waveform is routed to the PC software of the FSWP where an advanced pulse detection algorithm establish markers at the beginning and end of each pulse.

The pulse markers are used by the pulse detector block in Figure 15 to automatically determine the pulse width and PRI of the pulse modulated signal. The pulse width is used to establish a time gate around the pulse, such that the FSWP only processes phase noise information during the pulse on time. Since very high frequency resolution is required to measure phase noise close to the carrier the FSWP must use a long time records (many times longer than the PRI) for signal processing. Using gating type "Edge" the FSWP applies a precise time gate (typically 75% of the pulse width) inside the pulse and calculates a duty cycle based on the gate width and period of the waveform. Since the squelch block completely blanks the signal off during the time between pulses the noise level is reduced by a factor of $10 \cdot \text{Log}_{10}(\text{duty cycle})$, eliminating half of the pulse desensitization loss. In addition, knowing the duty cycle of the pulse waveform the FSWP DSP can apply gain to overcome most of the remaining pulse desensitization losses. For best pulsed carrier phase noise performance, use of the Edge trigger is required, see figure 17 and explanation, below.

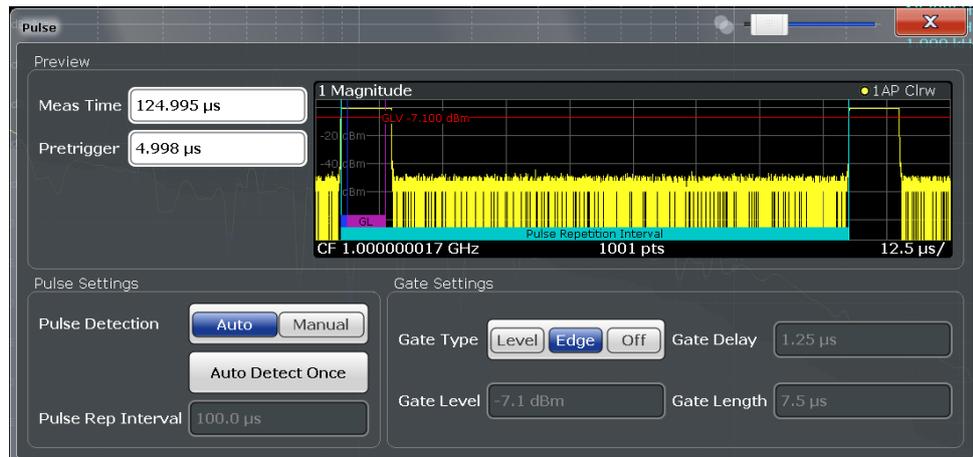


Figure 17, FSWP Pulse Configuration Menu.

Figure 17 shows the FSWP pulse configuration screen. From this menu the user can select whether he wants to use automatic or manual pulse detection. The default condition is to use automatic pulse detection. In this mode the analyzer will automatically find the pulse and center the analyzer's time gate over the central 75% of the pulse. Time gating can be turned off with the Gate Type radio buttons (resulting in lower system sensitivity) or can be set to a level mode where the user must specify a desired gate level. Finally, in manual pulse detection mode the user can specify the gate delay and gate length to focus the measurement on a specific region of the pulse. Looking at the graphic at the top of the menu, the gate delay is shown as a blue bar at the lower left of the pulse and is marked GD. The magenta bar shows the time that the gate is open and the FSWP is making its phase noise measurement. Generally, hundreds of pulses must be integrated together to make a phase noise measurement. This would suggest that waveforms with constant PRFs are most desirable. Advanced users might want to manually change the gate delay and length to measure phase noise in a particular region of the pulse. However, it should be noted that if the gate is moved outside the main area of the pulse additional noise will be introduced into the measurement, which will limit analyzer performance.

One additional point that is worthy of mention is that of required minimum pulse width to support a phase noise measurement. Recall from the block diagram that the four analog to digital converters following the I-Q mixers are 100 Msa/s digitizers. The FSWP requires at least ten samples in the pulse to make a phase noise measurement; therefore, establishing a minimum pulse width of 100 ns. Most phase lock loop based phase noise measurements systems experience fairly severe limitations for minimum duty cycle because of pulse desensitization. The R&S FSWP is not as limited pertaining to minimum duty cycle and can still make accurate measurements with duty cycles less than 1%.

Referring again to Figure 15, the blue highlighted block shows the signal processing used for measurement of pulsed carriers. Again, using the pulse start and stop markers from the K6 software the pulse detector generates a gate that is fed to the squelch block. The squelch block blanks the signal during pulse off time, which eliminates all broad band noise during the pulse off time, improving measurement dynamic range.

The next block is the IQLP. This is a DSP low-pass filter that is executed in FPGA to filter off the frequency components greater than PRF/2 which contain no unique phase noise information. This is a key advantage of the FSWP compared to most phase noise measurement systems. In most legacy phase noise measurement systems the user must determine the waveform PRF and manually select an appropriate analog low-pass filter. Generally, the correct filter is not implemented in the measurement system and the user is forced to use an external filter. To make matters worse, most of these systems have no provision for the user to connect the proper filter and he is forced to use an external phase detector to provide a connection point for the external filter.

The FSWP greatly simplifies this process by building the correct filter on-the-fly, freeing the user from having to do all of the work himself. In addition, the measurement is not corrupted by having to use external filters and phase detectors which provide more entry points for interfering signals and noise. Another key advantage of the FSWP's DSP based pulse detection and processing is that once we have gated the pulse and executed our measurements on a clean region from the center of the pulse we no longer have to deal with the switching transient portions of the pulse and do not suffer from the dramatic signal losses associated with pulse desensitization, as mentioned above.

Finally, it would be good to show an example of an FSWP noise measurement of a pulsed carrier. Again, we will use the 10 μ s pulse width we have previously discussed in this paper.



Figure 18, FSWP pulsed carrier noise measurement

Figure 18 shows AM noise and phase noise as simultaneously measured by the FSWP. From a few simple set up parameters to a completed pulsed phase noise measurement with just a few button presses and less than one minute of measurement time.

3 Conclusion

In summary, the Rohde & Schwarz FSWP brings a new level of performance and improved user experience to RF and microwave phase noise measurements. By moving the processing of phase noise from traditional analog phase detectors to a DSP based digital frequency discriminator the FSWP exhibits improved sensitivity with simplified system set up and calibration. By combining all required hardware in the box the user is freed from complicated intersystem cabling and experiences a significant reduction in system foot print. FSWP key contributions include:

- Simple graphical user interface with touch screen operation to improve user experience
- 1 MHz to 50 GHz frequency range, without external mixers or down converters
- Internal signal source (Option B64) for residual (two-port) phase noise measurements. In addition, no external delay lines are required for residual noise measurements.
- Pulse phase noise measurement with an automatic pulse detection system and PRF/2 low-pass filters implemented in DSP
- Two independent base band channels with DC to 40 MHz bandwidth
- Three built in low-noise power supplies to power external DUTs and provide tune voltages for VCOs. Power supply can source up to 2 Amperes of current.
- Simultaneous measurement of amplitude and phase noise without the need for an external AM detector
- 15 to 30 dB better AM suppression than traditional phase detector systems
- Optional built in signal/spectrum analyzer (Option B1) with up to 80 MHz of analysis bandwidth
- Optional built in Vector Signal Analyzer (Option K70)
- Optional pulse radar/EW pulse analysis software (K6)
- Optional analog AM, FM, and Φ M demodulation software (K7)
- Optional noise figure personality (Option K30)
- Instrument security (Option K33)

Lastly, in addition to the above FSWP contributions to the state-of-the-art in phase noise measurements the FSWP exhibits approximately a one-hundred fold speed improvement, as compared to the previous generation of Rohde and Schwarz phase noise measurement systems. The FSWP should be a welcome addition to any RF and microwave laboratory involved in pulsed carrier and CW phase noise measurements.

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Regional contact

Europe, Africa, Middle East
+49 89 4129 12345
customersupport@rohde-schwarz.com

North America
1 888 TEST RSA (1 888 837 87 72)
customer.support@rsa.rohde-schwarz.com

Latin America
+1 410 910 79 88
customersupport.la@rohde-schwarz.com

Asia Pacific
+65 65 13 04 88
customersupport.asia@rohde-schwarz.com

China
+86 800 810 82 28 | +86 400 650 58 96
customersupport.china@rohde-schwarz.com

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