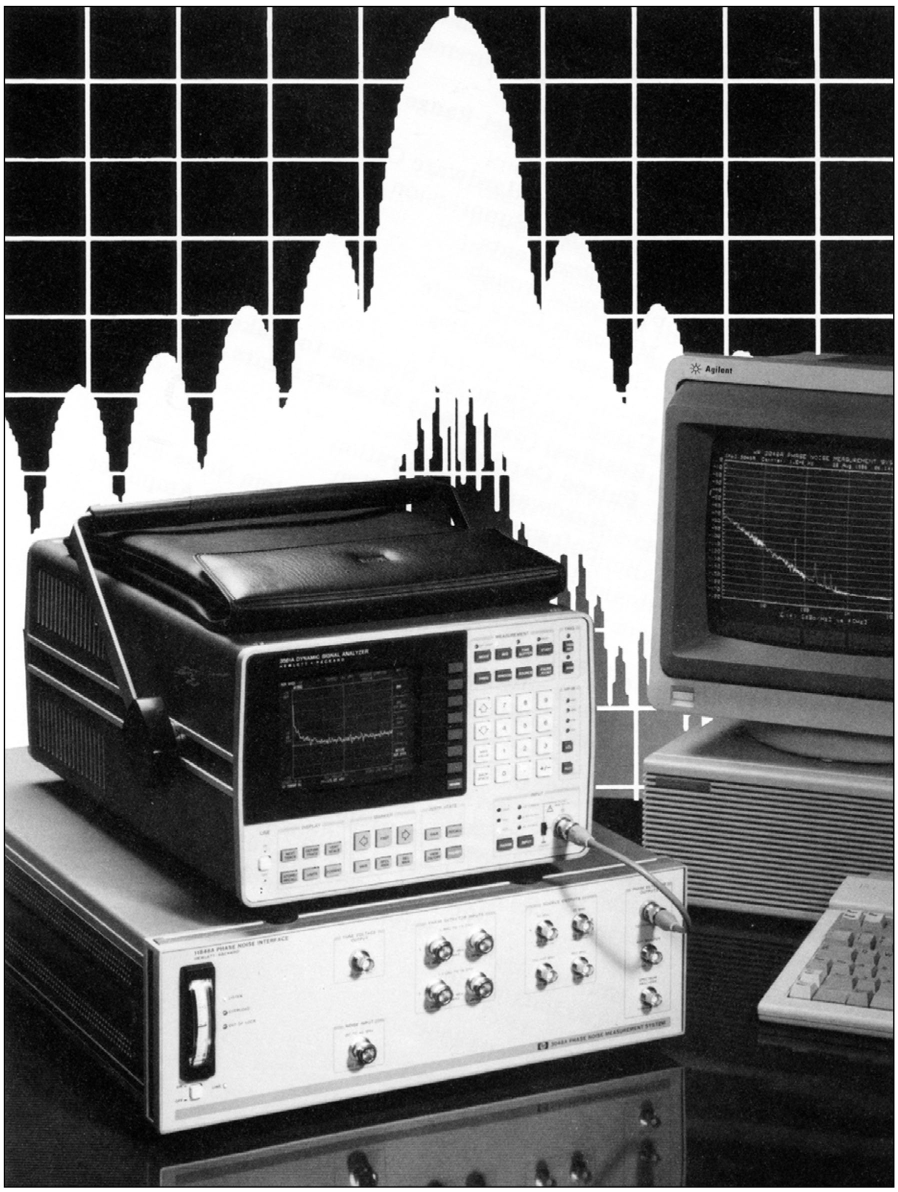
Agilent AN 386

Pulsed Carrier Phase Noise Measurements Using the Agilent 3048A Phase Noise Measurement System

Application Note



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Chapter 1. Introduction

Advances in RF and microwave communication technology have extended system performance to levels previously unattainable. Design emphasis on sensitivity and selectivity have resulted in dramatic improvements in those areas. However, as factors previously limiting system performance have been overcome, new limitations arise and certain parameters take on increased importance. One of these parameters is the phase noise of signal sources used in pulsed RF and microwave systems.

In pulsed radar systems, for example, the phase noise of the receiver local oscillator sets the minimum signal level that must be returned from a target in order to be detectable. In this case, phase noise affects the selectivity of the radar receiver, which in turn determines the effective range of the overall system.

Since the overall dynamic range of the radar system is influenced by the noise of the transmitted signal, it is not only important to know the absolute noise of the individual oscillators but to know the residual or additive noise of the signal processing devices like power amplifiers and pulse modulators. Because the final signal in most radar systems is pulsed, making absolute phase noise measurements on the pulsed carrier is essential to determining the overall performance of the system.

Pulsed carrier phase noise measurements have traditionally been made using custom test sets dedicated to that particular measurement need. With the introduction of the Agilent Technologies 3048A Phase Noise Measurement System, a general-purpose measurement tool now exists which has the capability to make phase noise measurements, with some limitations, on pulsed carriers and devices. This application note discusses the use and limitations of the 3048A system for making pulsed carrier phase noise measurements.

The assumption is made that the reader is familiar with the basic concepts of phase noise and CW phase noise measurement techniques. For a complete discussion of these topics, refer to Application Notes 283-3, 207, and 150-4; RF & Microwave Measurement Symposium paper "Choosing a Phase Noise Measurement Technique," literature number 1000-1118; and the RF and Microwave Phase Noise Measurement Seminar, literature number 1000-1132. Copies of these pieces of literature may be obtained at no cost by writing to:

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Chapter 2 reviews the fundamentals of pulsed carriers in the frequency and time domains. The majority of terms used in succeeding chapters are defined throughout Chapter 2. Chapter 3 discusses how the single sideband phase noise of a CW carrier is affected by the pulse modulation process. Chapter 4 discusses the effects a pulsed RF carrier has on the performance of a phase detector based measurement system such as the 3048A. Techniques are presented for dealing with these effects. The pulsed carrier phase noise measurement capabilities of the 3048A and 3048A/11729C are summarized at the conclusion of Chapter 4. Chapter 5 presents the recommended hardware configurations and the step-by-step measurement procedures for making residual (two port) measurements on pulsed RF carriers using the Agilent 3048A Phase Noise Measurement System. Chapter 6 goes on to present the recommended hardware configurations and step-by-step measurement procedures for making absolute measurements on pulsed carriers using the 3048A system and the 3048A/11729C combination. Chapter 6 discusses using the Agilent 11729C Carrier Noise Test Set as a low noise downconverter with the 3048A for making phase noise measurements on pulsed microwave carriers. In conclusion, Chapter 7 presents a brief discussion on making pulsed AM noise measurements using the 3048A system.

Chapter 2. Fundamentals of Pulsed Carriers

The formation of a square wave from a fundamental sine wave and its odd harmonics is a good way to begin a discussion of pulsed carriers and their representation in the time and frequency domains.

You might recall having plotted a sine wave and its odd harmonics on a sheet of graph paper, then adding up all the instantaneous values. If there were enough harmonics plotted at their correct amplitudes and phases, the resultant waveform would begin to approach a square wave. The fundamental frequency determined the square wave rate, and the amplitudes of the harmonics varied inversely to their number.

A rectangular wave is merely an extension of this principle. In fact, to produce a rectangular wave, the phases must be such that all the harmonics go through a positive or negative maximum at the same time as the fundamental. Theoretically, to produce a perfectly rectangular wave an infinite number of harmonics would be required. Actually, the amplitudes of the higher order harmonics are relatively small, so reasonably shaped rectangular waves can be produced with a limited number of harmonics. By changing the relative amplitudes and phases of the harmonies, both odd and even, an infinite number of waveshapes can be plotted.

To create a train of pulses (i.e., a waveform whose amplitude alternates between zero and one) with a series of sine waves, a dc component must be added. Its value equals the amplitude of the negative loops of the rectangular wave with sign reversed.

Consider a perfect rectangular pulse train as shown in Figure 1a, perfect in the sense that the rise time is zero and there is no overshoot or other aberrations. This pulse is shown in the time domain and if we wish to examine it in the frequency domain it must be broken down into its individual frequency components. Figure 1b superimposes the fundamental and its second harmonic plus a constant voltage to show how the pulse begins to take shape as more harmonics are plotted.

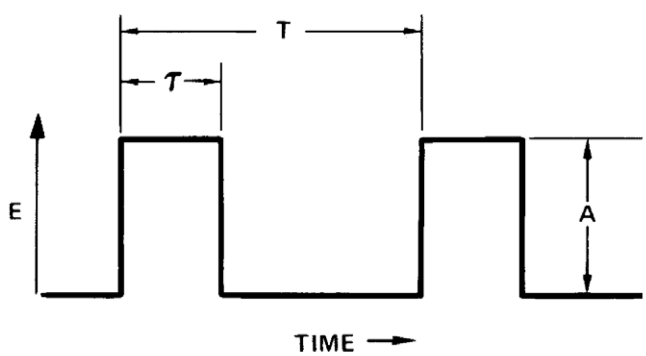


Figure 1a. Periodic rectangular pulse train

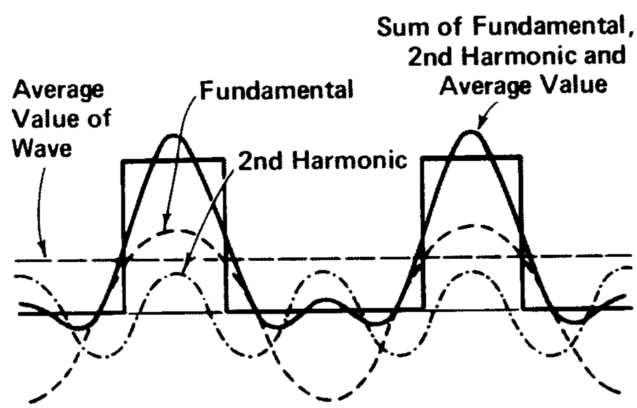


Figure 1b. Addition of a fundamental cosine wave and its harmonies to form rectangular pulses

A spectrum analyzer would in effect “unplot” these waveforms and present the fundamental and each harmonic in the frequency domain.

A frequency domain plot of this waveform would be as shown in Figure 2. This is an amplitude versus frequency plot of the individual waves which would have to be added together to produce the waveform. Since all the waves are integer multiples of the fundamental (PRF), the spacing between lines is equal to the PRF. The envelope of this plot follows a sinX/X function with the spectral line frequencies at $f_{LINE} = n \times 1/T$, for $n = 1,2,3 \ldots \infty$. Note that the nulls occur at integer multiples of the reciprocal of the pulse width.

Before proceeding to a discussion of modulating a CW RF carrier with a pulsed waveform, let’s define the terms used to represent the characteristics of a pulsed waveform.

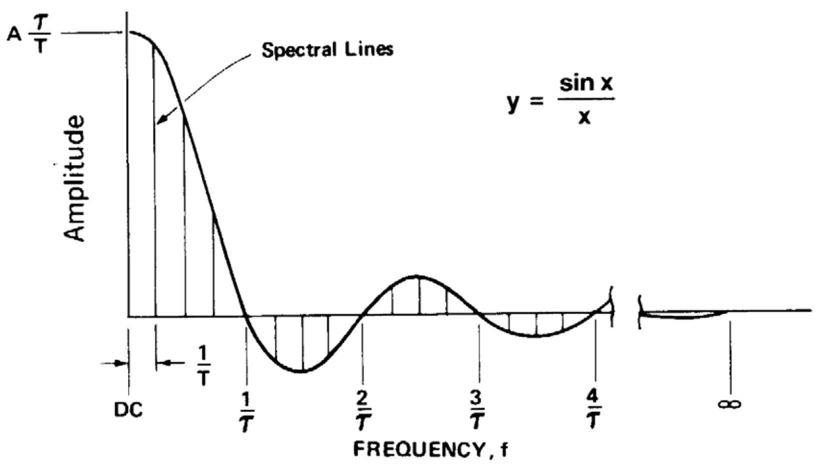


Figure 2. Spectrum of a perfectly rectangular pulse. Amplitudes and phases of an infinite number of harmonics are plotted, resulting in a smooth envelope as shown.

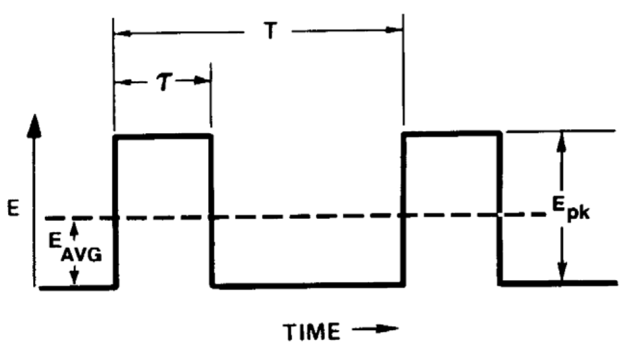


Figure 3. Basic characteristics of a pulsed waveform

Referring to Figure 3:

τ = Pulse width: Refers to the period of time during which the pulse is on, usually represented by the greek letter tau (τ).

T = Period: Refers to the time elapsed between the beginning of one pulse and the start of the next pulse (i.e., the time required to complete one cycle).

E_{PK} = Peak amplitude: Refers to the peak voltage level of each individual pulse.

E_{AVG} = Average amplitude: Refers to the average, or dc, value of the pulsed waveform.

$$E_{AVG} = E_{PK} \times \frac{\tau}{T}$$

PRF = Pulse repetition frequency: Refers to the frequency at which the pulses are repeated, the number of pulses per second.

$$PRF = \frac{1}{T}$$

Duty Cycle = Refers to the ratio of the pulse width to the pulse period. It represents the fraction of the time the pulse is on during one complete cycle.

$$Duty\ Cycle = \frac{\tau}{T}$$

With this background we can now apply the pulsed waveform as amplitude modulation to a continuous wave RF carrier. A pulsed carrier is typically a continuous wave carrier whose amplitude is modulated by a rectangular pulse train having a relative amplitude of one during each pulse and zero during the period between pulses. Pulsed carriers can also be generated by pulsing a frequency generating device, such as an oscillator, on and off. One of the fundamental differences between these two methods is that an amplitude modulated CW carrier is phase continuous from pulse to pulse, whereas the phase of a frequency generating device, which is pulsed on and off, is random. The Agilent 3048A, using the phase detector technique, can only measure the phase noise of phase-continuous signals. The phase detector technique requires that the two input signals be at quadrature (i.e., 90 degrees out of phase). If quadrature is lost, the system will terminate the measurement. Quadrature cannot be maintained if the phase from pulse to pulse is random.

From single-tone AM modulation theory we know that sidebands will be produced above and below the carrier frequency. The concept is the same for a rectangular pulse train, except that the rectangular pulse train is made up of many tones, which produce multiple sidebands commonly referred to as spectral lines in the frequency domain. In fact, there will be twice as many sidebands or spectral lines as there are harmonics contained in the modulating pulse.

Figure 4 shows the spectral plot resulting from amplitude modulating a CW carrier with a rectangular pulse train. The individual lines represent the modulation products (upper and lower sidebands) of the CW carrier and the rectangular pulse train (fundamental and harmonics of the PRF). The spectral lines will be spaced in frequency by the fundamental frequency of the PRF.

The spectral line frequencies can be expressed as:

$$F_L = F_C \pm (n \times PRF)$$

where: F_C = carrier frequency
PRF = pulse repetition frequency
n = 0, 1, 2, 3, ...

The "mainlobe" in the center and the "sidelobes" are shown as groups of spectral lines extending above and below the baseline. For perfectly rectangular pulses and other functions whose derivatives are discontinuous at some point, the number of sidelobes is infinite.

The mainlobe contains the carrier frequency represented by the largest spectral line in the center. Amplitude of the spectral lines forming the lobes varies as a function of frequency according to the expression $\dfrac{\sin \omega \frac{\tau}{2}}{\omega \frac{\tau}{2}}$ for a perfectly rectangular pulse.

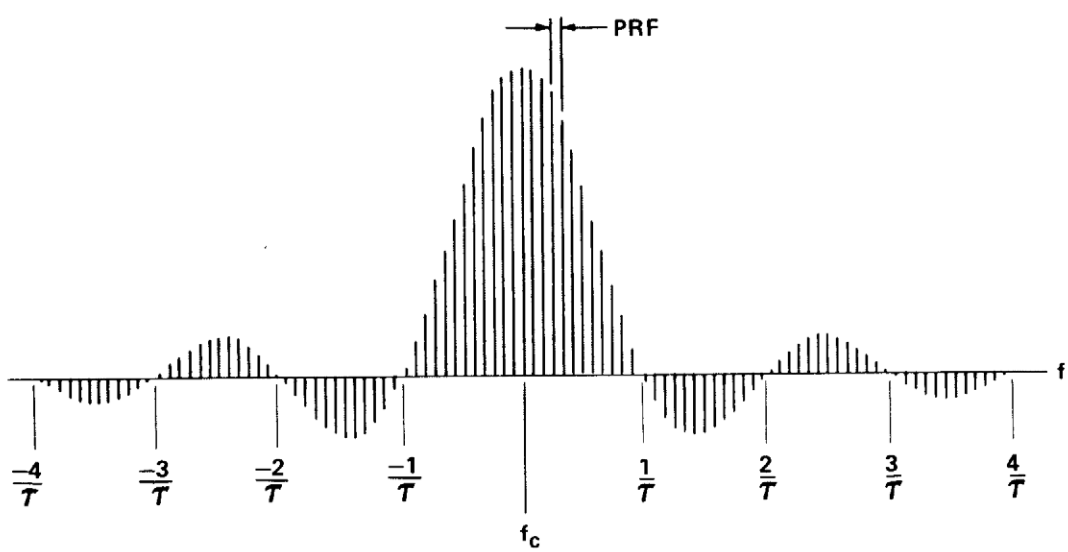


Figure 4. Resultant spectrum of a carrier amplitude modulated with a rectangular pulse

So for a given carrier frequency, the points where these lines go through zero amplitude are determined by the modulating pulse width only. As pulse width becomes shorter, minima of the envelope become further removed in frequency from the carrier, and the lobes become wider. The sidelobe widths in frequency are related to the modulating pulse width by the expression $f = 1/\tau$. Since the mainlobe contains the origin of the spectrum (the carrier frequency), the upper and lower sidebands extending from this point form a main lobe $2/\tau$ wide. Remember, however, that the total number of sidelobes remains constant so long as the pulse quality, or shape, is unchanged and only its repetition rate is varied. Figure 5 compares the spectral plots for two pulse widths, each at two repetition rates with carrier frequency held constant.

Notice in the drawings how the spectral lines extend below the baseline as well as above. This corresponds to the harmonics in the modulating pulse, having a phase relationship of 180 degrees with respect to the fundamental of the modulating waveform. If these pulses were viewed using a spectrum analyzer, it would invert the negative-going lines and display all amplitudes above the baseline since a spectrum analyzer can only detect amplitudes and not phase.

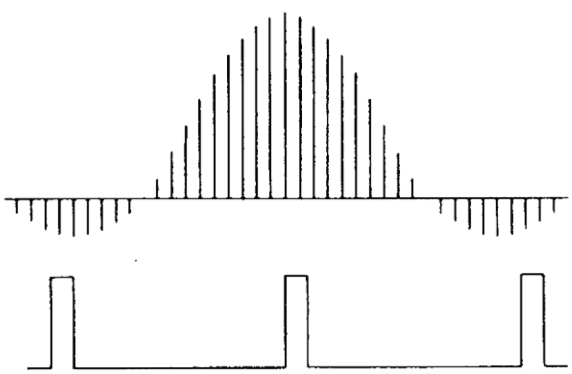


Figure 5a. Narrow pulse width causes wide spectrum lobes, high PRF results in low spectral line density.

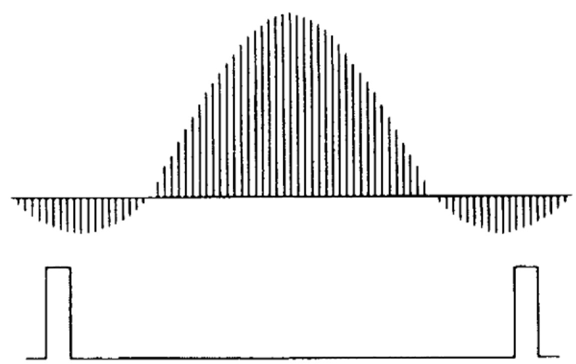


Figure 5c. PRF lower than 5a results in higher spectral density. Lobe width is same as 5a since pulse widths are identical.

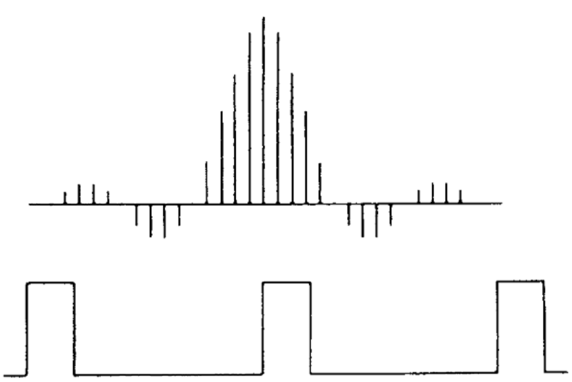


Figure 5b. Wider pulse than 5a causes narrower lobes, but line density remains constant since PRF is unchanged.

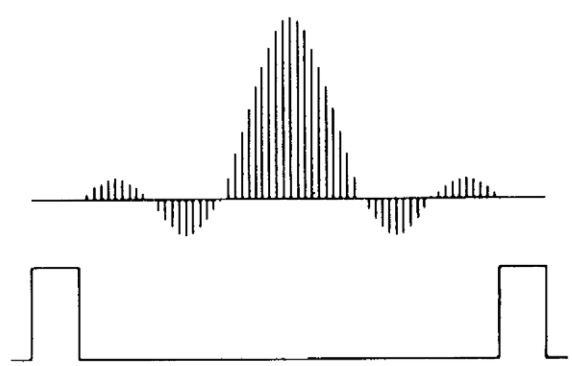


Figure 5d. Spectral density and PRF unchanged from 5c but lobe widths are reduced by wider pulse.

Before proceeding to a discussion of how the single sideband phase noise of a CW carrier is affected by the pulse modulation process, let's define the terms used to represent the characteristics of a pulsed carrier.

Referring to Figure 6:

f_c = Carrier frequency: Refers to the frequency of the unmodulated CW signal contained within the pulse envelope.

τ = Pulse width: Refers to the duration of the pulses. It is usually represented by the lower case greek letter tau (τ).

T = Pulse period: Refers to the time elapsed between the beginning of one pulse and the start of the next pulse.

PRF = Pulse repetition frequency: Refers to the frequency at which the pulses are transmitted–the number of pulses per second.

$$PRF = \frac{1}{T}$$

Duty cycle = Refers to the ratio of the pulse width (τ) to the pulse period (T). It represents the fraction of time the pulse is on during one complete pulse period.

$$Duty\ Cycle = \frac{\tau}{T}$$

P_{PK} = Peak power: Refers to the power of the individual pulses. If the pulses are rectangular–that is, if the power level is constant from the beginning to the end of each pulse–peak power is simply the peak power of the unmodulated CW signal.

P_{AVG} = Average power: Refers to the peak power of the pulse averaged over the pulse period, T. If the pulses are rectangular, the average power equals the peak power times the ratio of the pulse width, τ, to the pulse period, T.

$$P_{AVG} = P_{PK} \times \frac{\tau}{T}$$

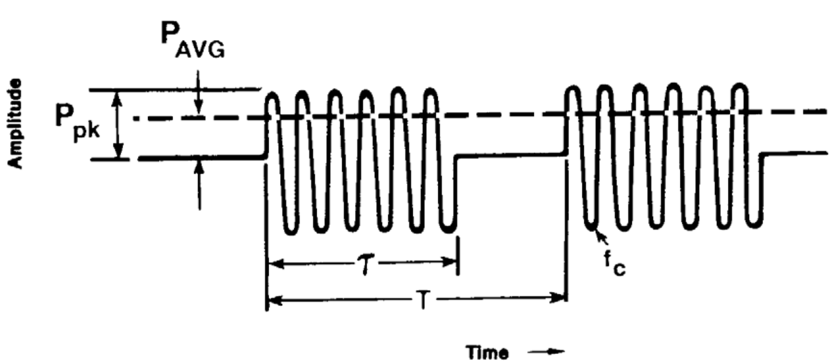


Figure 6. Basic characteristics of a pulsed carrier

Chapter 3. How Pulse Modulation Affects the SSB Phase Noise of a CW Carrier

Having defined the basic characteristics of pulsed waveforms and pulsed carriers in Chapter 2, we will now focus our attention on how pulse modulation affects the distribution of the SSB phase noise of a CW carrier.

As defined in Chapter 2, a pulsed carrier is actually a continuous wave carrier whose amplitude is modulated by a rectangular pulse train having a relative amplitude of one during each pulse and zero during the period between pulses. From modulation theory we know that any wave whose amplitude is modulated has two sidebands, an upper sideband and a lower sideband. Modulation theory also tells us that a portion of the waves total energy is contained in those sidebands. So one way of examining how the energy of a pulsed carrier is distributed is to look in the frequency domain at the sidebands produced when a CW carrier is pulse modulated.

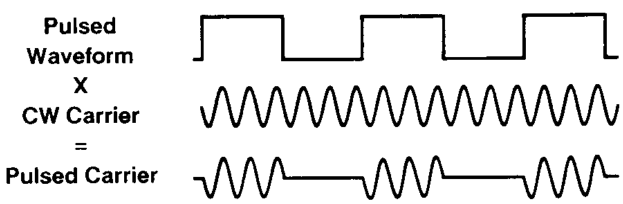


Figure 7. Multiplication of a CW carrier by a pulsed waveform results in a pulsed carrier.

Referring to Figure 7, amplitude modulation can be expressed in the time domain as the result of multiplying the CW carrier by a rectangular pulse train.

Multiplication in the time domain is analogous to the convolution of the pulsed waveform spectra and the CW carrier spectra in the frequency domain, as shown in Figure 8.

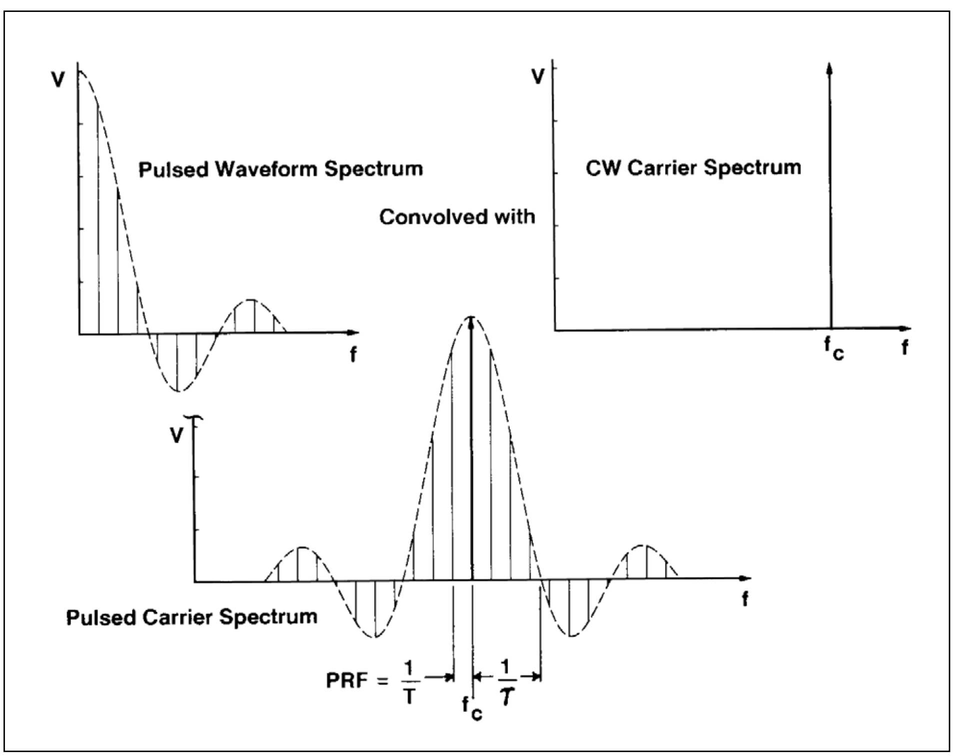


Figure 8. The spectrum of the product of two time functions is the convolution of their spectra.

When the CW carrier is amplitude modulated by the pulsed waveform, as illustrated in Figure 8, the sine wave represented by each spectral line in the pulsed waveform spectrum produces two sidebands, an upper sideband and a lower sideband. The fundamental produces sidebands at f_r (pulse repetition frequency) Hertz above and below the CW carrier. The second harmonic produces sidebands at $2f_r$ above and below the CW carrier, and so on. The zero frequency line produces an output at the carrier frequency. The convolution of the individual spectra produces all possible sums and differences of the CW carrier and all of the harmonic components contained in the modulating pulse.

If a single-sided spurious response is added to the CW carrier and pulse modulated, as illustrated in Figure 9, the spectrum of the pulsed waveform is convolved with both the CW carrier and the single-sided spur. The resultant spectrum is the sum of the individual spectra.

Figure 9 shows the spectrum which results from the convolution of the CW carrier with single-sided spur spectra and the pulsed waveform spectra. This example demonstrates two very important characteristics of the pulse modulation process:

1. The modulation process "aliased" a spur onto each of the PRF lines in the resultant pulsed carrier spectrum. The aliased spur is a sum or difference product of the spur and the pulsed waveform spectra weighted by the sinX/X function.

2. No matter how great the offset between the CW carrier and the spur, an alias of the spur will appear within an offset of ±PRF/2 of the central line.

If the single-sided spur were replaced with a double-sided spur and the pulse modulation process repeated, the double-sided spur would be aliased onto each of the PRF lines in the resultant pulsed carrier spectrum.

In the frequency domain a signal is no longer a discrete spectral line but spreads out over frequencies both above and below the nominal signal frequency in the form of modulation sidebands due to random phase fluctuations. This is the "phase noise" of the signal. At any given offset from the carrier the phase noise can be represented as a pair of discrete sidebands.

If the spur in Figure 9 is replaced with the phase noise of the CW carrier and the process repeated, as in Figure 10, the modulation process will alias the noise of the CW carrier onto each of the PRF lines in the pulsed carrier spectrum. As discussed in the case of the single-sided spur, the modulation process has two very significant effects on the composite signal:

1. The modulation process "aliased" the noise of the CW carrier onto each of the PRF lines in the pulsed carrier spectrum.

   The effect of this is that the composite noise at the central line (i.e., f_c) has been increased by the sum of the aliased noise on each of the PRF lines, weighted by the sinX/X function.

2. Phase noise information on the CW carrier at offsets above PRF/2 has been aliased to within ±PRF/2 of the central line in the resultant pulsed carrier spectrum.

   The effect of this, from a phase noise measurement perspective, is that after detection there is no new phase noise information above offsets of PRF/2 Hertz.

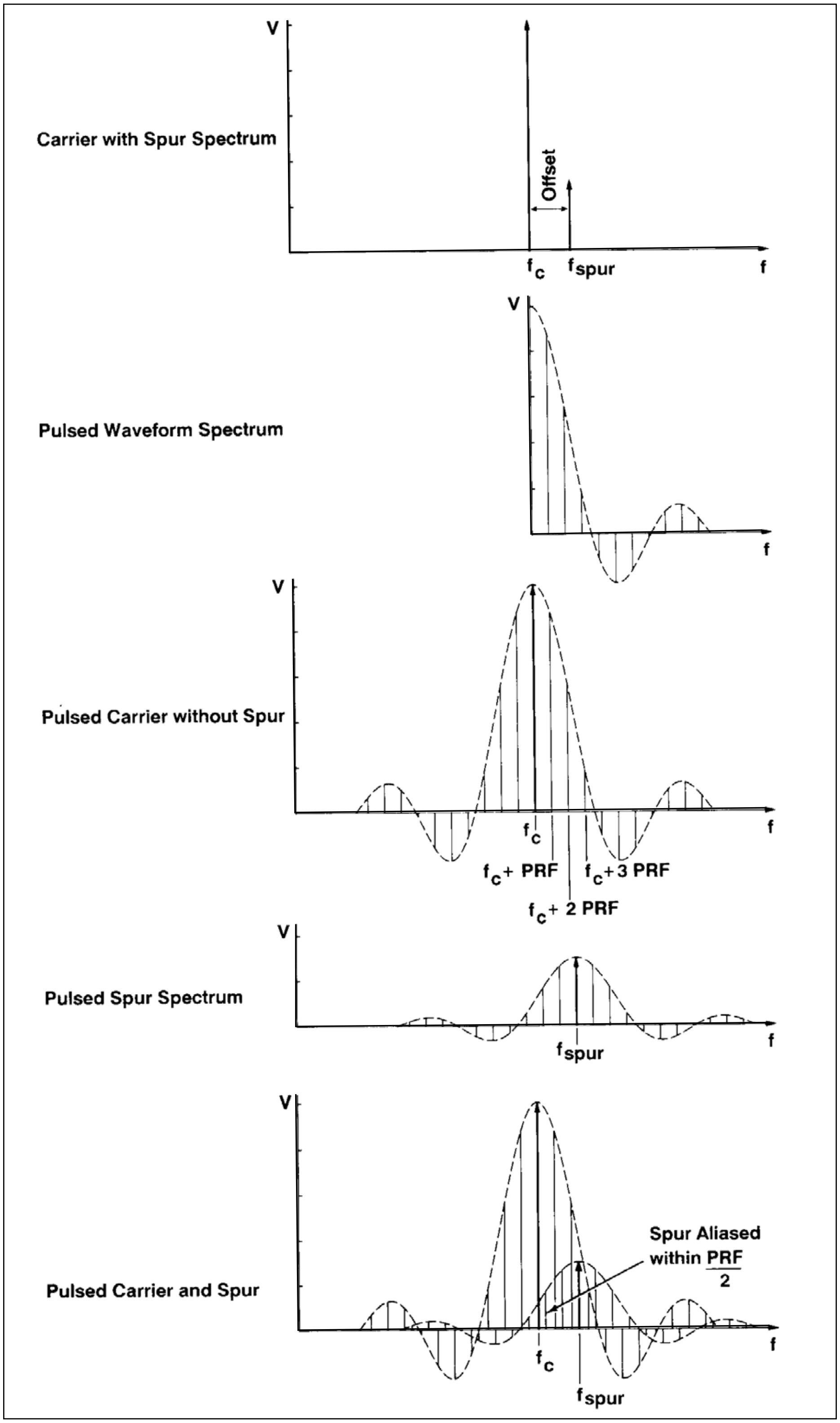
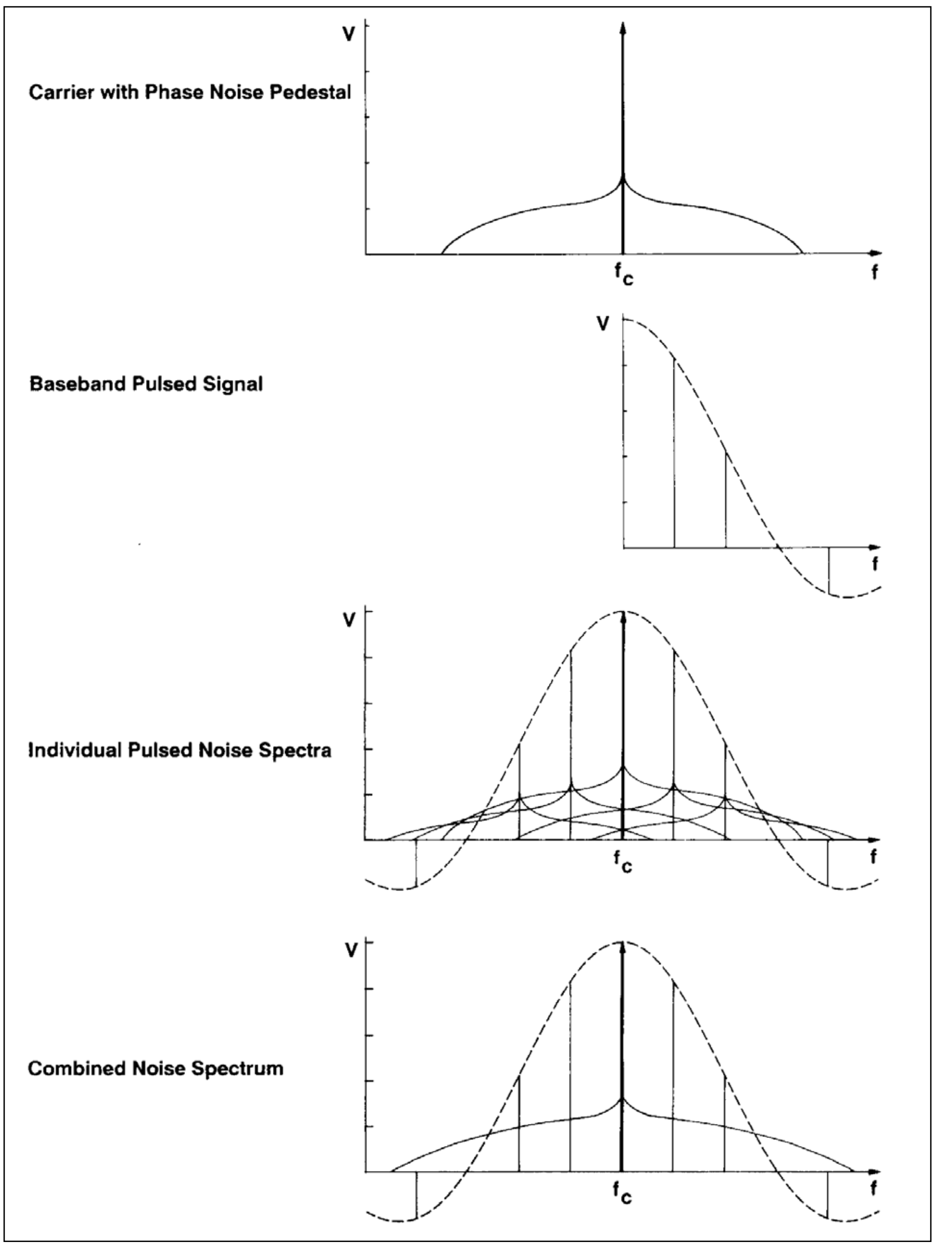


Figure 9. Convolution of a CW carrier with spur spectra and a pulsed waveform spectra

From our discussion of pulse basics (refer to Chapter 2, Figure 5), we know that, for a fixed pulse width, the density or spacing of the PRF spectral lines is inversely related to the PRF (high PRF results in low spectral line density, low PRF results in high spectral line density). Using this relationship one can see that, for a fixed pulse width, the increase in noise at f_c will be inversely proportional to PRF (i.e., low PRF = high spectral line density large increase in noise at f_c; high PRF = low spectral line density = small increase in noise at f_c).

Relating PRF back to duty cycle we can see that, for a fixed PRF, increasing the duty cycle (relative to some nominal value) will result in a decrease in noise at f_c due to a narrowing of the lobes.

Figure 10. Noise aliasing with pulse modulation



As an upper limit, the noise at f_c will increase by:

$\approx 10\log_{10}$ (*number of PRF lines to the first null*)

This is a worse case approximation and assumes that the noise contribution at each PRF line out to the first null is equal.

It should be noted that the shape of the CW carrier's phase noise energy distribution curve will affect the degree to which the noise is degraded at different offsets from the carrier. Typically, the slope of the close-in phase noise (0 to 100 Hz) will be very steep, on the order of 20 to 40 dB per decade. Due to the sinX/X envelope, aliased noise at these offsets will be well below the CW noise and little degradation will be seen. At higher offsets, up to ½ PRF, the degradation will be more apparent, especially if the CW noise curve has a pedestal, as illustrated in Figure 10. Since a pedestal 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 constantly decreasing slope. As previously discussed, above ½ PRF the noise is simply repeated about each PRF line.

Before proceeding to a review of the measurement technique used in the Agilent 3048A to make pulsed carrier phase noise measurements, one important question remains to be answered. All of the spectral plots shown so far have been of amplitude versus frequency. But nothing has been said about how this amplitude relates to the amplitude of the unmodulated carrier in the time domain.

We can clear up this discrepancy by developing a mathematical equation for a pulsed RF carrier and then examining the coefficients of the amplitude terms. Recall from Figure 7 that, in the time domain, amplitude modulation can be expressed as the result of multiplying the CW carrier by the pulsed waveform. It follows that the mathematical equation for a pulsed RF carrier can be derived by multiplying the equations for the CW carrier and the pulsed waveform as follows:

Unmodulated Carrier: $f_1(t) = A \cos \omega_c t$

Pulsed Waveform:

$$f_2(t) = \frac{\tau}{T}\left[1 + 2\sum_{n=1}^{\infty} \frac{\sin n\,\omega_0 \frac{\tau}{2}}{n\,\omega_0 \frac{\tau}{2}} \cos n\,\omega_0 t\right]$$

Pulse Modulated Carrier: $f_3(t) = f_1(t) \times f_2(t)$

$$f_3(t) = A\frac{\tau}{T}\cos\omega_c t + A\frac{\tau}{T}\sum_{n=1}^{\infty}\frac{\sin n\,\omega_0 \frac{\tau}{2}}{n\,\omega_0 \frac{\tau}{2}}\left[\cos(\omega_c + n\,\omega_0)t + \cos(\omega_c - n\,\omega_0)t\right]$$

Examination of the amplitude term shows that after being pulsed modulated, the amplitude of the unmodulated carrier (A) is reduced by the ratio of τ to T (i.e., the duty cycle).

In equation $f_3(t)$ the peak amplitude of the unmodulated carrier, A, is expressed as a voltage. The decrease in power of the modulated carrier (which appears as the central line in the pulsed carrier spectrum) relative to the power of the unmlated CW carrier cae expressed in dB as follows (recall that power is proportional to voltage squared):

Decrease in carrier power in dB

$$= 10\, Log_{10} \left(\frac{E_2}{E_1}\right)^2$$

$$= 10\, Log_{10} \left(\frac{A\,\frac{\tau}{T}}{A}\right)^2$$

$$= 20\, Log_{10} \left(\frac{\tau}{T}\right)$$

E_2 = P_{PK} amplitude of central line in pulsed carrier spectrum
E_1 = P_{PK} amplitude of unmodulated carrier

The average power of the pulsed carrier will also be affected by the duty cycle, as was discussed in Figure 6. The reduction in the average power of the pulsed carrier (i.e., the total spectral power of the pulsed waveform) relative to the peak power of the unmodulated carrier, can be expressed in dB as follows:

Decrease in pulsed carrier spectral power

$$= 10\, Log_{10} \left(\frac{P_{AVG}}{P_{PK}}\right)$$

$$= 10\, Log_{10} \left(\frac{P_{PK}\,\frac{\tau}{T}}{P_{PK}}\right)$$

$$= 10\, Log_{10} \left(\frac{\tau}{T}\right)$$

P_{AVG} = Average power of the pulsed carrier

P_{PK} = Peak power of unmodulated carrier

This apparent contradiction—i.e., carrier power drops by 20 $\log_{10}(\tau/T)$ but spectral power drops by 10 $\log_{10}(\tau/T)$—is most easily explained as follows: pulsing a CW carrier results in its power being distributed over a number of spectral components (carrier and sidebands). Each of these spectral components then contains some fraction of the total power.

Chapter 4. How Pulsing the Carrier Affects the Phase Detector Measurement Technique

Having reviewed the basics of pulsed carriers and having discussed how pulsing a CW carrier affects its SSB phase noise, we will now turn our attention to the effects a pulsed RF carrier has on a phase detector based phase noise measurement system. The phase detector method is the recommended measurement mode for making both residual and absolute noise measurements on pulsed carriers using the Agilent 3048A.

It is assumed that the reader understands the principles and operation of the phase detector method of measuring SSB phase noise as well as the definition of residual and absolute measurements. A complete discussion of these topics can be found in the references listed in Chapter 1. However, let's quickly review this method to establish a basis for understanding the problems which are encountered when applying this method to the measurement of pulsed carriers.

Figure 11 shows the basic block diagram of the phase detector method of measuring SSB phase noise. The phase detector (typically a double balanced mixer) is used to convert phase fluctuations to voltage fluctuations which are then displayed on a spectrum analyzer.

The use of mixers as phase detectors is based upon the fact that when two identical frequency, constant amplitude signals are input to a mixer, a dc output which is proportional to the phase difference between the two signals is generated. The output at the IF port contains the sum and difference of the frequencies input to the LO and RF ports. If the RF and LO signals have identical frequencies, their difference is 0 Hz, or dc, and will vary as the cosine of the phase difference between the LO and RF signals. Their sum, which is twice the input frequency, can be selectively filtered off (if not already beyond the frequency response of the IF port) by an LPF following the mixer. By maintaining a 90-degree phase difference between the RF and LO ports the mixer is operated in its most linear region (i.e., $\Delta V_{IF} \propto \Delta\Phi_{LO\text{-}RF}$ with the constant of proportionality being the phase detector constant, K_Φ, in volts/radian). The low noise amplifier (LNA) following the LPF amplifies the baseband signal above the noise of the spectrum analyzer. How the DUT is connected to the basic configuration determines whether the measurement will be an absolute or a residual measurement.

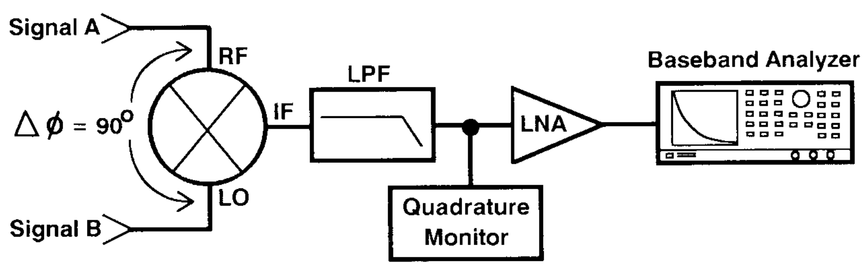


Figure 11. Block diagram of phase detector method of measuring SSB phase noise

The discussions in this chapter will focus on the specific areas where the system's response to a pulsed signal is different than its response to a CW signal. Since the effects of pulsing the carrier are the same for either measurement (i.e., absolute or residual), the discussions will focus on these effects as they relate to the basic block diagram and not to a particular measurement configuration. The basic block diagram in Figure 11 will be referenced throughout this chapter when explaining the differences in response and their effect on system performance. Where possible, methods for dealing with these effects will be presented. This chapter, then, will serve as a reference for subsequent chapters to explain the differences in measurement technique which must be applied to pulsed signals as compared to CW signals. The recommended hardware configurations for making absolute or residual measurements on pulsed carriers will be presented in Chapter 5.

System Noise Floor

A primary consideration when making any phase noise measurement is the system noise floor. The system noise floor represents the lowest level of noise which the system is capable of measuring for a given measurement configuration (i.e., the maximum system sensitivity for a given configuration). When using the phase detector technique as shown in Figure 11, the system noise floor (i.e., system sensitivity) is set by the noise floor of the mixer, the noise floor of the LNA, and the phase detector constant. The LNA following the LPF amplifies the baseband signal above the noise floor of the spectrum analyzer, thereby removing the spectrum analyzer as a limiting factor. The noise floor of the LNA and the noise floor of the phase detector are fixed by design but the phase detector constant can vary from measurement configuration to measurement configuration. The phase detector constant specifies the sensitivity of the phase detector in converting phase fluctuations to voltage fluctuations.

The magnitude of the phase detector constant is a function of the maximum output voltage level of the phase detector. The greater the maximum output voltage level is the more sensitive the phase detector becomes. The maximum output voltage level of the phase detector is a function of input drive level. It follows then that anything which affects the input drive level to the phase detector also affects the phase detector constant (i.e., the sensitivity of the phase detector). If the assumption is made that the signal at the LO input is strong enough to completely turn on the diodes, the maximum output voltage of the phase detector then becomes directly proportional to the RF input drive level. The RF drive level to the phase detector is set by the magnitude of the unmodulated carrier. Under pulsed conditions the amplitude (voltage) of the unmodulated carrier is reduced by 20 $\log_{10}$ (duty cycle). This in turn decreases the maximum output voltage which decreases the sensitivity of the phase detector. As the sensitivity of the phase detector decreases, the smallest increment of phase change which it can detect (i.e., convert to a voltage at its output) gets progressively larger. Viewed from a system perspective the minimum level of phase noise detectable by the system has increased. In other words, the system noise floor has degraded under pulsed conditions. The magnitude of this degradation can be determined by examination of the phase detector output under CW and pulsed conditions.

For a CW signal the output of an ideal phase detector, as a function of time, would appear as shown in Figure 12. The voltage fluctuations, V_{IF}, represent the sum of the noise of the reference and DUT (absolute measurement), or the sum of the internal noise floor and the noise added by the two port device (residual measurement). It should be pointed out that the vertical scale is greatly exaggerated for purposes of illustration.

Pulsing the carrier has several adverse effects when using the phase detector technique to measure phase noise. Phase noise measurements of pulsed carriers can be made, with some limitations, using the phase detector technique if these effects are accounted for.

For a pulsed carrier the phase detector output becomes a sampled version of the CW phase noise, with the sampling rate being the PRF. For an ideal phase detector the only time there would be an output is when the pulse is on and there is a phase difference between the RF and LO ports. If there was no phase difference between the RF and LO ports, the output would be 0 Vdc with no evidence of pulse envelope or PRF feedthrough. Ideally, the LPF following the phase detector only has to filter the sum product of the RF and LO ports. The output of an ideal phase detector for a pulsed carrier would be as shown in Figure 13. Again note the greatly exaggerated vertical scale used for purposes of illustration.

$$V_{IF} = (\Delta\varphi_{LO-RF}\ (t) \times K\varphi)$$

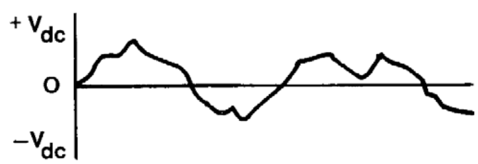


Figure 12. Phase detector output for a CW signal

$$V_{IF} = \left(\Delta\varphi_{LO-RF}\,(t)\right)\left(K\varphi_{CW} \times \frac{\tau}{T}\right)$$

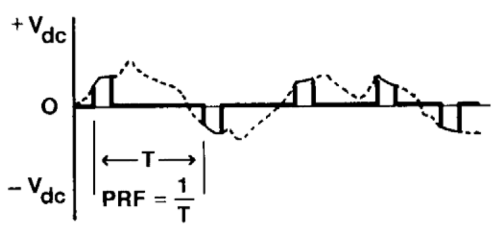


Figure 13. Phase detector output for a pulsed carrier

The effective sensitivity of the phase detector (i.e., the phase detector constant) is now scaled by the duty cycle of the pulsed carrier, just as in the case of the average or dc voltage of a rectangular waveform. For a pulsed measurement then:

$$K\varphi_{\text{PULSED}} = K\varphi_{\text{CW}} \times \frac{\tau}{T}$$

As previously discussed, this directly affects system sensitivity (i.e., system noise floor). Logarithmically, the degradation of the system noise floor is given by:

$$20\ Log_{10}\ (duty\ cycle)$$

Measurement Offset Range

When using the phase detector technique to make CW phase noise measurements with the Agilent 3048A system, the measurement offset range is set by the bandwidths of various circuit components—such as filters and amplifiers—in the Agilent 11848A Interface, and the bandwidth of the RF spectrum analyzer. For a standard 3048A system the measurement offset range is limited to 100 kHz, which is the bandwidth of the Agilent 3561A Dynamic Signal Analyzer. An RF spectrum analyzer can be added to increase the offset range to 40 MHz.

However, when measuring pulsed carriers, the valid measurement offset range is limited, based on sampling theory, to one-half the PRF. A pulsed RF carrier is essentially a sampled version of the unmodulated carrier. Sampling theory states that if a band-limited signal is amplitude modulated with a periodic pulse train—corresponding to extracting equally spaced time segments—it can be recovered exactly by low-pass filtering, if the fundamental frequency of the modulating pulse train is greater than twice the highest frequency present in the band-limited signal. In Chapter 3 we saw that the waveform which results from amplitude modulating with a periodic pulse train is a periodic function of frequency consisting of a sum of shifted replicas of the unmodulated carrier scaled by the PRF (see Figure 10, page 14). Sampling theory states that if the sampling frequency is twice the highest frequency in the band-limited signal there will be no overlap between the shifted replicas of the band-limited signal and that the band-limited signal will be faithfully reproduced at integer multiples of the sampling frequency, as shown in Figure 14(c). However, if the band-limited signal is under-sampled (i.e., the sampling frequency is less than twice the highest frequency in the band-limited signal) there will be overlap of the shifted replicas, as shown in Figure 14(d). This effect, in which the shifted replicas overlap, is referred to as *aliasing*. When aliasing occurs, the original frequency takes on the identity or alias of a lower frequency. Consequently, portions of the original signal are folded back onto itself when under-sampling occurs.

When sampling band-limited signals with a sampling frequency that is equal to exactly twice the highest frequency in the original signal, the shifted replicas will begin to overlap at the original signals center frequency plus one-half the sampling frequency. If under-sampling occurs, signal information which is at offsets above one-half the sampling frequency is folded or aliased to offsets below one-half the sampling frequency.

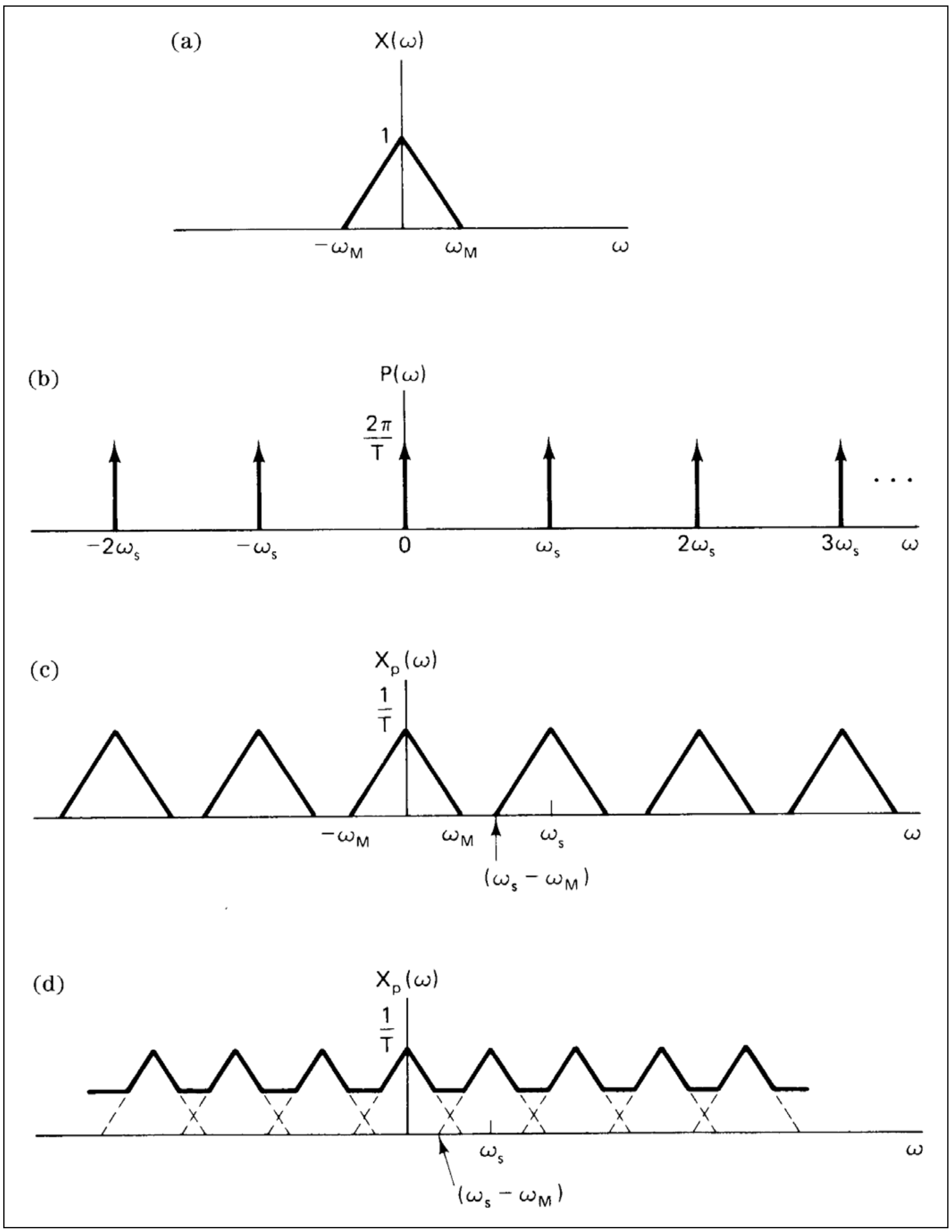


Figure 14. Effect in the frequency domain of sampling in the time domain

Since the phase noise of a CW carrier is not band-limited (i.e., there is information at offsets from 0 to infinity Hertz) when it is amplitude modulated with a periodic pulse train, it is constantly being under-sampled and aliasing will always occur, as was illustrated in Figures 9 and 10 in Chapter 3. This means that signal information at offsets above one-half the PRF will be aliased to offsets below one-half the PRF. Since all of the phase noise information of the CW carrier above offsets of one-half the PRF will be aliased below PRF/2 of the central line in the pulsed waveform spectrum, one only needs to measure out to the point where the overlap occurs (i.e., PRF/2). Beyond that point there is no new information since we would begin to measure the same phase noise information which was just measured around the central line, only now we would begin to measure it around each PRF line. If the phase noise was directly measured around each PRF line in the pulsed spectrum, the same noise spectrum which is seen around the central line would be seen to repeat exactly. While it is not necessary to limit the measurement offset range to one-half the PRF, it is necessary to recognize that the measured data is only useful to one half the PRF.

Up to this point the discussions have been based on the characteristics of an "ideal" mixer used as a phase detector. Unfortunately real mixers do not exhibit such ideal characteristics. Attention will now be focused on these non-ideal characteristics and their effects on the measurement process.

Mixer dc Offset

Theoretically, when signals having identical frequencies are applied to the RF and LO ports of a mixer, the dc voltage at the IF port should be 0 Vdc when the phase difference between the RF and LO ports is 90 degrees (i.e., in quadrature). In practice, real mixers exhibit a dc offset at the quadrature point. The dc offset is the deviation from 0 Vdc that is seen at the mixer output when the RF and LO inputs are in quadrature. This dc offset is of concern because the measurement path through the 3048A system is dc coupled and a large amount of dc offset can overload the LNA, forcing it to be removed from the measurement path. Removing the LNA degrades the system noise floor by 20 to 30 dB. When making CW measurements the phase difference between the signals at the LO and RF ports, $\Delta\Phi_{LO\text{-}RF}$, can be adjusted off quadrature to cancel the dc offset of the mixer, as shown in Figure 15. When making CW measurements the 3048A system can automatically compensate for dc offsets of up to one half the peak voltage of the beat-note and apply appropriate correction factors to the phase detector constant to maintain linearity and system accuracy.

However, setting quadrature is not as straightforward when measuring pulsed carriers. When setting up a pulsed measurement one would normally connect the pulsed DUT to the RF input of the phase detector and the reference signal to the LO input. Under these conditions the RF path is pulsed and the LO path is CW. During the pulse off, interval power is only applied to the LO port, which produces a dc offset at the IF port. Adjusting the phase a few degrees off quadrature does not directly cancel this dc offset. The phase must be adjusted off quadrature by an amount necessary to produce an average dc value of 0 Vdc, as shown in Figure 16.

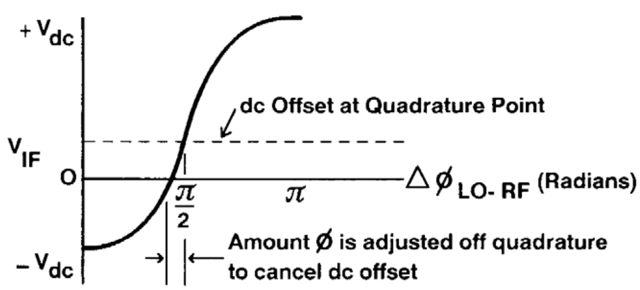


Figure 15. Cancelling dc offset by adjusting $\Delta\Phi_{LO-RF}$ off quadrature

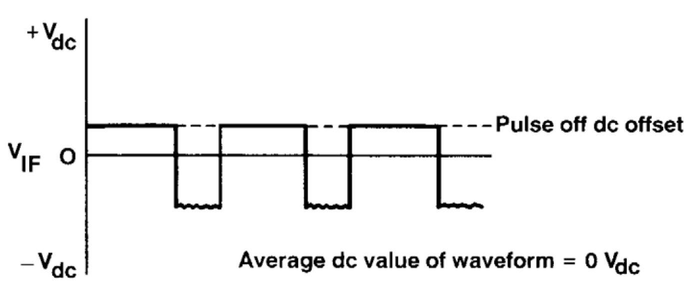


Figure 16. Compensating for dc offset by adjusting $\Delta\Phi_{LO-RF}$ to produce an average dc value of 0 Vdc

Note that there are some negative consequences associated with doing this. As can be seen in Figure 16, this method produces high PRF feedthrough. If the magnitude of the feedthrough is sufficient to overload the LNA, an LPF would be required. The amount of phase shift required to produce a waveform with an average value of 0 Vdc may move the point of measurement far from quadrature, thereby degrading the accuracy of the measurement. This effect becomes more pronounced as the duty cycle is reduced. As the pulse on time becomes a smaller and smaller percentage of the total pulse period, the amplitude of the pulse on time must be increased more and more to produce an average value of 0 Vdc. Since the amplitude of the pulse on time out of the phase detector is a function of the phase difference at its input ports ($\Delta\Phi_{LO-RF}$), increasing the amplitude moves the measurement point farther and farther from the true quadrature point of 90 degrees. Additionally, moving the measurement point off quadrature increases the AM sensitivity of the phase detector, which would further degrade the accuracy of the measurement.

An alternative approach, which minimizes the negative consequences previously mentioned, is to adjust the phase for minimum deviation from dc offset during the pulse on period, as shown in Figure 17.

The advantages of this method are:

- PRF feedthrough is minimized.
- LPF for PRF may not be necessary.
- Point of measurement can be as close to quadrature as the CW case.

However, since the dc offset has not been cancelled by averaging the waveforms dc value to 0 Vdc, the average dc value must pass through the LNA—possibly overloading it and forcing it to be removed from the measurement path. The negative consequence of removing the LNA, as previously discussed, is severe degradation of the system noise floor.

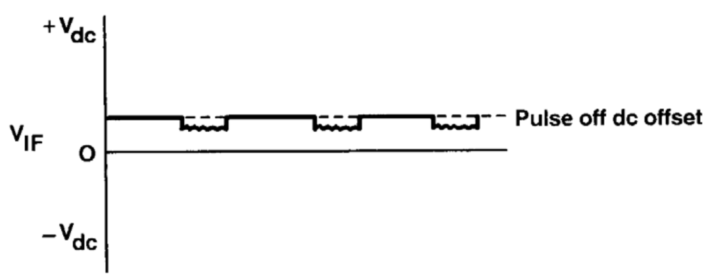


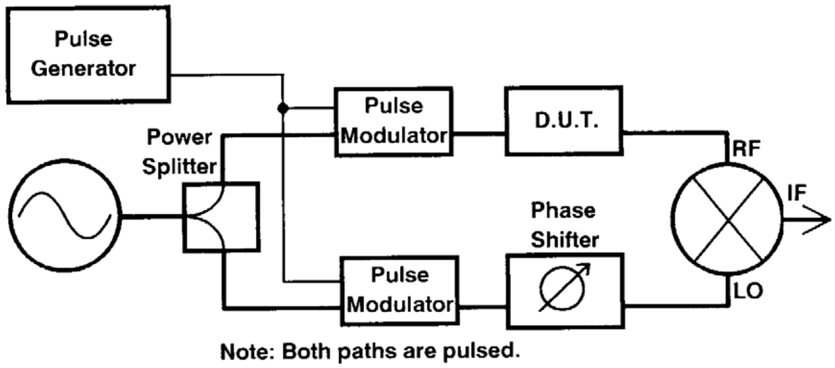
Figure 17. Adjusting phase for minimum deviation from dc offset during pulse on period

The recommended method for measuring the phase noise of pulsed carriers in a way that eliminates the dc offset problem is to pulse both the LO and RF paths to the phase detector. This technique also minimizes AM noise problems (AM noise problems will be discussed in the following section). By pulsing both paths to the phase detector, an output is produced only during the pulse on interval. This effectively eliminates the dc offset caused by the LO signal being present during the pulse off interval.

The advantages of this method are:

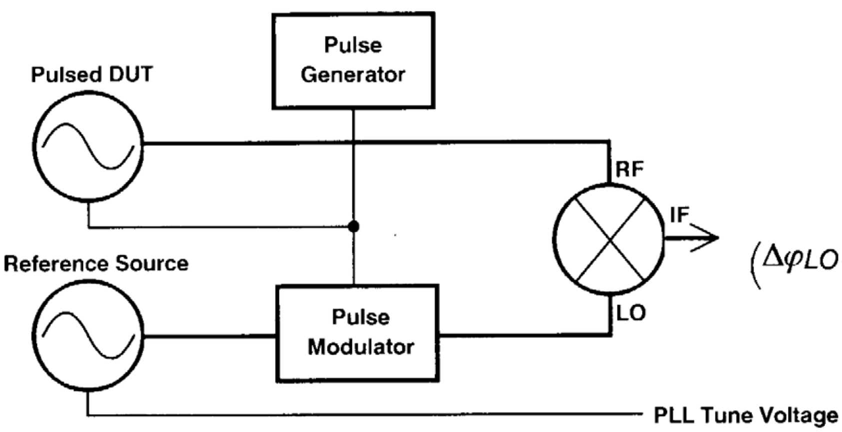
- PRF feedthrough is minimized.
- PRF filter may not be necessary.
- Point of measurement can be as close to quadrature as the CW case.
- AM noise contribution is minimized.

The basic block diagram for making a pulsed residual measurement is shown in Figure 18. Figure 19 shows the basic block diagram for making a pulsed absolute measurement using this technique. Figure 20 shows the output of the phase detector, at the quadrature point, when pulsing both the RF and LO paths to the phase detector.



$$V_{IF} = \left(\Delta\varphi_{LO-RF}(t)\right)\left(K_{\varphi CW} \times \frac{\tau}{T}\right)$$

Figure 18. Recommended configuration for making pulsed residual noise measurement



$$V_{IF} = \left(\Delta\varphi_{LO-RF}(t)\right)\left(K_{\varphi CW} \times \frac{\tau}{T}\right)$$

Figure 19. Recommended configuration for making pulsed absolute noise measurement

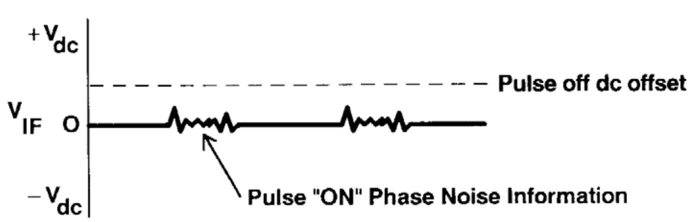


Figure 20. Phase detector output, at quadrature, when pulsing both inputs

LO AM Noise Suppression

Analogous to the single sideband phase noise definition, AM noise is defined as the noise power in one AM modulation sideband, divided by the total signal power, in units of dBc/Hz. AM noise plays an important role in phase noise measurements because it can affect the accuracy of the phase noise measurement.

One benefit of a double balanced mixer configuration is the suppression of AM noise associated with the local oscillator. AM noise components near the LO signal frequency will produce noise within the bandwidth of the IF port. Compared to single ended mixers, converted noise will appear suppressed in a double balanced mixer by 30 to 40 dB at lower UHF frequencies, while at microwave frequencies values of 10 to 20 dB are common. LO AM noise suppression is also evidenced when a double balanced mixer is operated as a phase detector.

Typically, when measuring the noise of CW signals, the level of AM noise is well below the level of phase noise and does not degrade the accuracy of the phase noise measurement. However, this situation changes somewhat when attempting to measure the phase noise of a pulsed carrier.

As previously discussed, under pulsed conditions the phase detector constant is scaled by the duty cycle which reduces the level of phase noise seen on the output of the phase detector. When only the DUT path to the phase detector is pulsed, LO AM noise is present at the output of the phase detector 100% of the time while the DUT phase noise is only present during the pulse on time. Consequently the LO AM noise becomes a larger component of the total measured noise. If both paths to the phase detector are pulsed the LO AM noise returns to the same relative level as in the CW case.

Additionally, if only the DUT path is pulsed, the LO AM noise component will begin to dominate at very low duty cycles. If both paths are pulsed the LO AM noise component will always be present in the same relative level as in the CW case, at all duty cycles. Therefore, pulsing both paths to the phase detector allows measurements to be made at lower duty cycles.

Phase Transients

As a consequence of the pulse modulation process, the instantaneous phase of the DUT output signal may undergo rapid fluctuations during turn-on and turn-off. These rapid fluctuations will appear as transients at the output of the phase detector, as shown in Figure 21.

These phase transients can only be observed on a wideband oscilloscope connected to the AUX monitor port on the Agilent 11848A Interface. They cannot be evaluated with a spectrum analyzer on either the pulsed RF carrier or the AUX monitor port since a spectrum analyzer provides no information about the shape of the waveform in the time domain. These transients, if present, can have several adverse effects on the measurement process.

First, if the amplitude of the transient exceeds ±0.2 V_{pk} at the output of the LNA (equivalent to 4 mV at the output of the phase detector since LNA gain is equal to 50), the LNA will be removed from the measurement path, thereby degrading the system noise floor by 20 to 30 dB. In the 11848A Interface, LNA overload is monitored by an AC coupled peak detector on the output of the LNA. The overload trip point, which has been established for CW operation, is set at ±0.2 V_{pk} at the LNA output. As shown in Figure 21, the amplitude of the leading or trailing phase transient may exceed the CW trip point, but only for a fraction of the pulse width. Even though the LNA is not experiencing an overload condition across the entire pulse width, the system—having sensed an overload condition—will remove the LNA from the measurement path. Under pulsed conditions it is acceptable to leave the LNA IN if no saturation occurs on that portion of the signal where the phase noise information is. In order to overcome this situation, the system—when in PULSE MODE—gives the operator the capability to override the hardware overload detector. The operator can toggle the LNA IN or OUT while observing the phase detector output on the wideband oscilloscope connected to the AUX monitor port of the 11848A Interface. If that portion of the waveform which contains the phase noise information (typically the central portion) shows any signs of saturation, the LNA must be left OUT. However, if that portion of the waveform which contains the phase noise information shows no signs of saturation with the LNA IN even though the phase transients themselves may show signs of saturation—as shown in Figure 22—the LNA can be left in providing the remaining LNA IN criteria are satisfied.

Secondly, since the phase transients are coincident with the pulse they will appear as PRF feedthrough, which may necessitate the use of a PRF filter. If the magnitude of the transients cause the LNA to be overloaded and to be removed from the measurement path by the system, a PRF filter can be inserted before the LNA. A properly designed PRF filter will sufficiently reduce the magnitude of the phase transients so that the LNA can be left in the measurement path. If a PRF filter is required an external phase detector must be used, as discussed in the next section.

Finally, if the phase transients exceed 0.2 radians for more than 10% of the pulse width, as shown in Figure 23, the measurement accuracy is degraded. Under these conditions measurement accuracy can be determined manually by inserting a phase modulator into one path to the phase detector, introduce a phase modulation sideband and calibrate on sideband level. This technique is discussed in the RF & Microwave symposium paper "Residual Phase Noise and AM Noise Measurements and Techniques," p/n 1000-1126. This paper is available from your local Agilent sales and service office.

Figure 21. Phase transients at output of phase detector as observed on oscilloscope connected to the AUX monitor port

Figure 22. Phase transients with saturation as observed on oscilloscope connected to AUX monitor port

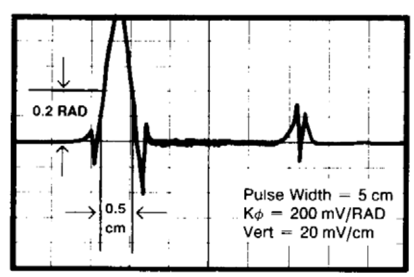


Figure 23. Phase transients exceeding 0.2 radians for more than 10% of pulse width as observed on oscilloscope connected to AUX monitor port

PRF Feedthrough

Under pulsed conditions an ideal phase detector, at the quadrature point, would only produce an output voltage when the pulse is on and there is a phase difference between the LO and RF inputs. In the ideal case if there was no phase difference between the input ports the output would always be 0 Vdc. No pulse envelope or PRF feedthrough would be seen. In the ideal case the LPF following the phase detector would only have to filter off the sum product of the RF and LO ports. However, in the practical case, PRF feedthrough will exist. Throughout this chapter we have discussed two of the mechanisms which contribute to PRF feedthrough: phase transients, and moving away from the quadrature point to compensate for mixer dc offset. Additionally, since the port to port isolation of the mixer is not infinite some carrier feedthrough will exist as a result of the mixing process itself.

It is the magnitude of the PRF feedthrough which is of concern in a pulsed carrier phase noise measurement. Too much PRF feedthrough can overload the LNA and/or the baseband spectrum analyzer, resulting in measurement accuracy and system noise floor degradation. Under these conditions (i.e., excessive PRF feedthrough) an additional PRF filter can be added to the measurement path to reduce the magnitude of the feedthrough, as shown in Figure 24.

The PRF filter should be designed to have the response characteristics shown in Figure 25.

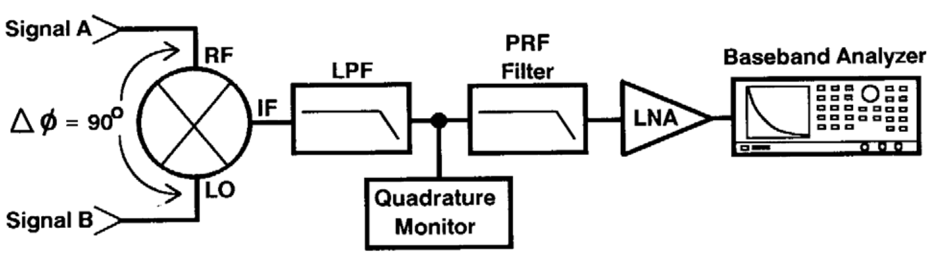


Figure 24. Proper location of PRF filter in measurement path

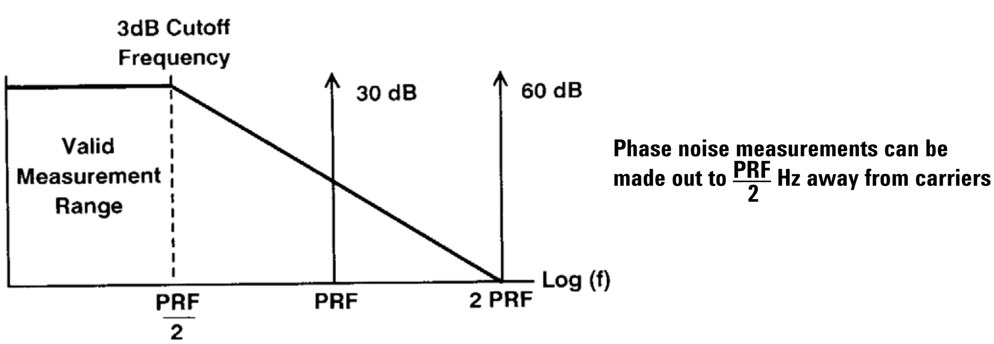


Figure 25. PRF filter response characteristics

The absolute magnitude of the PRF feedthrough for a given measurement configuration cannot be empirically determined before the measurement is attempted. Consequently, the need for a PRF filter typically manifests itself during the measurement process. However, experience has shown that for many measurement configurations the measurement can be made successfully without a PRF filter. When a PRF filter is required an external phase detector must also be used since there is no direct access to the output of the internal phase detectors in the 11848A Interface. Figures 26 and 27 show the recommended hardware configurations for making pulsed phase noise measurements when a PRF filter is required.

When PRF feedthrough is excessive and a PRF filter is necessary, the Watkins Johnson WJM9H can be used as an RF phase detector from 5 MHz to 1.6 GHz, and the NORSAL DBM 1-26 can be used as a microwave phase detector from 1.2 GHz to 18 GHz.

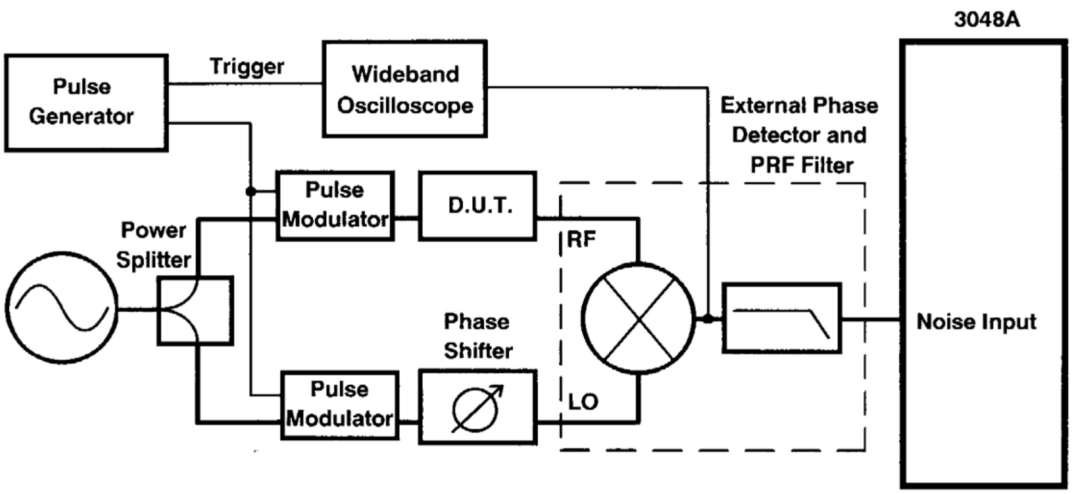
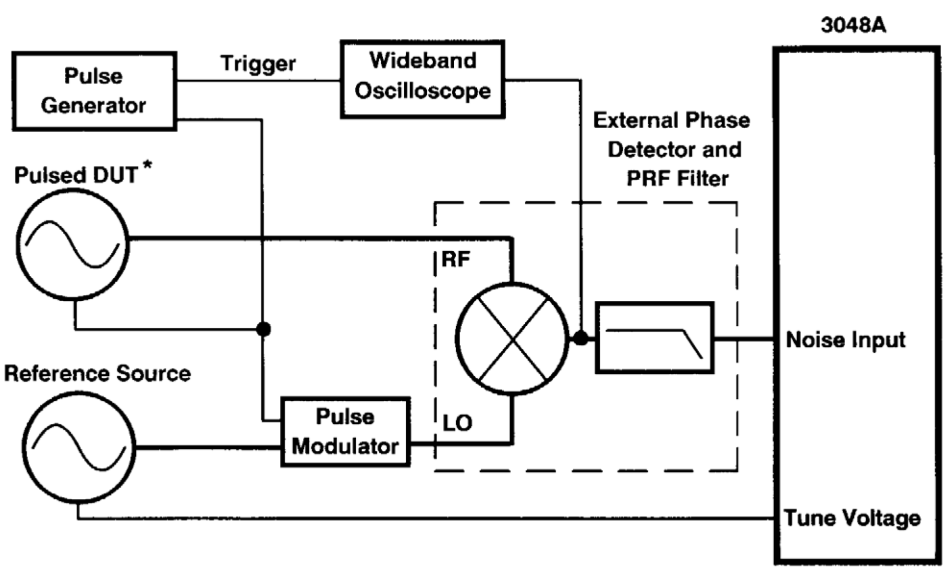


Figure 26. Recommended hardware configuration for making pulsed residual measurements when a PRF filter is required



* If DUT does not have internal pulse modulator then put pulse modulator in "R" path and connect to output of pulse generator.

Figure 27. Recommended hardware configuration for making pulsed absolute measurements when a PRF filter is required

Minimum Duty Cycle

One of the limitations of the Agilent 3048A Phase Noise Measurement System, when used to make pulsed carrier phase noise measurements, is the minimum duty cycle which the system will measure. Many applications call for measurements at very low duty cycles (i.e., << 1%), which may be below the capabilities of the 3048A system. Let's briefly discuss what the minimums are and why they were selected.

The recommended minimum duty cycles for the supported hardware configurations are:

Agilent 3048A—5% for absolute measurements and 1% for residual measurements
Agilent 3048A/11729C—5% for absolute

For the residual case, 1% was selected to provide an acceptable noise floor for the measurement. Using the microwave detector and assuming that the level of the input signal is at the minimum rating for the detector (i.e., 0 dBm, which gives the worst noise floor for the detector), one would get a phase detector constant of 0.11 volt/radian which would yield a system noise floor of approximately –165 dBc. A 1% duty cycle would degrade the noise floor by 40 dB (i.e., $20 \log_{10} 0.01$) to approximately –125 dBc. Note that this is a worst case analysis. If the duty cycle decreases further then the system noise floor continues to rise. It is possible to measure lower duty cycles when making residual measurements providing the system noise floor remains low enough to be useful.

For the absolute case, which uses a PLL to maintain quadrature, the criteria is driven by the signal level necessary to keep the PLL functioning correctly. The PLL was designed to maintain lock with a minimum phase detector constant of 20 mVolts/radian. At duty cycles less than 5% the amplifiers in the PLL do not have enough gain to provide the necessary signal level to keep the loop locked. It may be possible to make absolute measurements at duty cycles less than 5% if the PLL can maintain lock.

The Agilent 3048A system is guaranteed to operate correctly at the recommended minimums. Below the recommended minimums the system may or may not function correctly. If a measurement is made below the recommended minimums and the system operates correctly then the accuracy of the measurement is not in question.

For pulse widths less than 5 microseconds a special filter needs to be added to the 11848A Interface. This filter keeps the amplifiers that drive the 3561A from saturating on harmonics of the PRF. The following special ordering options add this filter to a new or used 11848A Interface:

3048A-E50:	adds filter to new system, order with system
11848A-H05:	adds filter to new 11848A, order with 11848A
11848A-G05:	factory retrofit, adds filter to used 11848A, 11848A must be returned to factory for filter retrofit
11848A-K05:	Field Retrofit Kit. Adds filter to 11848A. Customer installable. Requires 3585A/B to recalibrate 11848A.

Summary

The phase detector method is the recommended measurement mode for making residual and absolute single sideband phase noise measurements on pulsed carriers when using the Agilent 3048A System. This chapter has focused on the effects a pulsed carrier has on the phase detector method of phase noise measurement. Taking these combined effects into consideration, the pulsed carrier phase noise measurement capabilities of the 3048A system and the 3048A/11729C combination are summarized in Table 4.1

Table 4.1 Agilent 3048A and 3048A/11729C Pulsed Carrier Measurement Capabilities

	Capabilities	
	Absolute	**Residual**
3048A		
Frequency range	Determined by phase detector (Internal or External)	
Minimum duty cycle	5%	1%
Minimum PRF	2 x PTR	No Minimum
Noise Floor	Phase detector noise scaled by duty cycle	
Accuracy	±2 dB .01 Hz to 1 MHz offsets ±4 dB 1 MHz to 40 MHz offsets	
3048A/11729		
Frequency range	10 MHz to 18 GHz, to mm with specials	N/A
Minimum duty cycle	5%	N/A
Minimum PRF	2 x PTR	N/A
Noise Floor	8662/11729 CW noise folded by pulse spectrum	N/A
Accuracy	±2 dB .01 Hz to 1 MHz offsets ±4 dB 1 MHz to 40 MHz offsets	N/A

Frequency range: The frequency range of the internal phase detectors is: 5 MHz to 1.6 GHz (low frequency phase detector) and 1.2 GHz to 18 GHz (high-frequency phase detector). When using an external phase detector, the carrier frequency range becomes a function of the external phase detector used. Since the output of the phase detector/filter combination is at baseband, the carrier frequency range is determined directly by the external phase detector used.

Minimum Duty Cycle Range: The minimum duty cycle for absolute measurements is 5% and for residual is 1%. Recall that in an absolute measurement a control signal for the phase locked loop must be generated which requires a slightly higher signal level out of the phase detector.

Minimum PRF range: For a residual measurement, in which the PLL is not used, there are no minimum PRF restrictions. For an absolute measurement the minimum PRF is determined by the PTR (peak tuning range). The critical requirement is that the PRF be well outside the PLL bandwidth. If the PRF fell within the loop bandwidth, the PLL would try and respond to it, resulting in an unstable loop. In the ideal case, the PRF should be an order of magnitude higher than the bandwidth of the PLL. In the 3048A system the PLL bandwidth is derived from the PTR so it is convenient to express the PRF limitations as a function of the PTR (recall that the PTR is derived from the tuning characteristics of the VCO source used for the measurement). For an absolute measurement the PRF is limited to 2 x PTR (PTR ≤ PRF/2).

Noise Floor: For the 3048A system the measurement noise floor (i.e., the noise that would be measured if the DUT were removed) would be the phase detector noise scaled by the duty cycle (i.e., 20 $\log_{10}$ [duty cycle]). For the 3048A/11729C combination the measurement noise floor would be influenced by: 1) the phase detector noise scaled by the duty cycle, and 2) the 8662/11729 or 8663/11729 CW noise folded by the pulse process. Again, the phase detector noise is scaled by 20 $\log_{10}$ (duty cycle). As an approximation the CW noise of the 866X/11729 is scaled by 10 $\log_{10}$ (# of PRF lines to first null). This is only a rough approximation. The aliased noise from each PRF line, scaled by the sinX/X function, would have to be individually summed at each offset to determine the actual level of the aliased noise.

Accuracy: The Agilent 3048A System maintains its specified accuracy of ±2 dB for offsets of .01 Hz to 1 MHz, and ±4 dB for offsets of 1 MHz to 40 MHz. The 3048A/11729C combination has the same accuracy as the 3048A system.

Chapter 5. Using the Agilent 3048A System to Make Residual (Two Port) Measurements On Pulsed Carriers

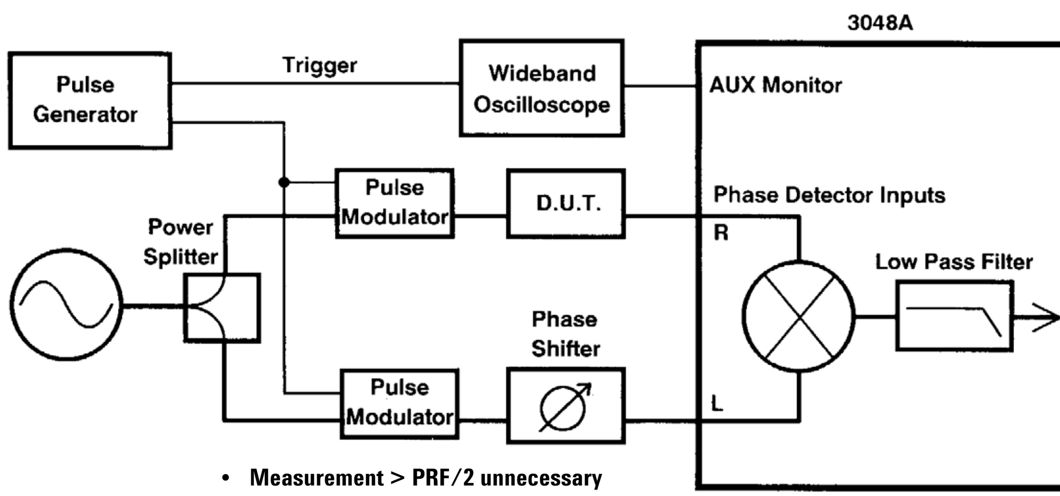
This chapter presents the recommended hardware configurations and step-by-step measurement procedures for making residual or two-port phase noise measurements on pulsed carriers. For a complete discussion of how pulsing the carrier affects phase noise measurements and how to address these effects when using the Agilent 3048A, refer to Chapter 4.

To successfully make a pulsed residual measurement the operator must ensure that the following items are correct:

1. Hardware configuration
2. Software configuration

Hardware Configuration

The recommended configuration for making pulsed residual measurements is to synchronously pulse both paths to the phase detector. Using this approach, the phase detector only produces an output during the pulse on interval. This eliminates many problems which occur when only the DUT path is pulsed. Figure 28 shows the recommended hardware configuration for making residual measurements. The hardware configuration shown in Figure 28 assumes that the source cannot be pulsed. However, if the source can be pulsed, the pulse modulators can be removed from the hardware configuration. Pulsing the source will also synchronously pulse both paths to the phase detector.



- **Measurement > PRF/2 unnecessary**
- **Duty cycles ≥ 1%**
- **Noise floor scales with duty cycle [degrades by 20 log (duty cycle)]**
- **PRF feedthrough may require PRF filter**

Figure 28. Recommended hardware configuration for making pulsed residual measurements using the Agilent 3048A

If PRF feedthrough is excessive and a PRF filter is necessary, an external phase detector/filter combination is required. The reason for this is that in the standard 11848A Interface, the phase detector output is not available at the front panel. When using an external detector, adding a PRF filter is just a matter of cabling. External phase detector/filter combinations (PRF and/or LPF) can be supplied as specials from Agilent. The Watkins-Johnson WJM9H can be used as an RF phase detector (5 MHz to 1.6 GHz), and the NORSAL DBM 1-26 can be used as a microwave phase detector (1.2 GHz to 18 GHz).

For reasons discussed in Chapter 4, both paths to the phase detector are pulsed. For f_c <2 GHz a double balanced mixer with a dc coupled IF port can be used as a pulse modulator. For f_c ≥2 GHz the Agilent 11720A Pulse Modulator can be used. The choice of phase shifter, power splitter, and mixer will depend on the frequency of the measurement. Some common devices are:

Mixer as pulse modulator:
Watkins-Johnson WJM9HC, 10 MHz to 1.5 GHz

Phase shifter:
Agilent 08741-60004 Mechanical Line Stretcher, 2.5 to 3.4 nsec variable delay

0 deg power splitter:

Mini-Circuits	MCL ZSCF 2-5	10 MHz to 1.5 GHz, Pwr max 1 watt
	MCL ZAPD-21	500 MHz to 2 GHz, Pwr max 10 watts
	MCL ZAPD-4	2 GHz to 4.2 GHz, Pwr max 10 watts

For pulse widths less than 5 usec a special filter needs to be added to the 11848A Interface to keep the amplifiers that drive the 3561A from saturating on harmonics of the PRF (see Chapter 4). A wideband oscilloscope is recommended as the quadrature monitor device in order to see phase transients which may be produced as a result of the modulation process. If the phase transient exceeds 0.2 radians for more than 10% of the pulse width, the accuracy of the measurement is degraded. Since the phase transients, if present, are coincident with the pulse they will appear as PRF feedthrough. For frequencies up to 4.2 GHz, the Agilent 8665A Signal Generator is the recommended source. For frequencies above 4.2 GHz, the Agilent 867X series of microwave signal generators offer good phase and AM noise performance.

Software Configuration

The 3048A software can measure two carrier types: CW and pulsed. When configured to measure pulsed carriers the software does not monitor quadrature for residual measurements. The reason for this is that the PRF lines can confuse the quadrature monitoring circuitry in the 11848A Interface. Therefore the user must manually set and monitor quadrature using the wideband oscilloscope connected to the AUX monitor port on the 11848A Interface.

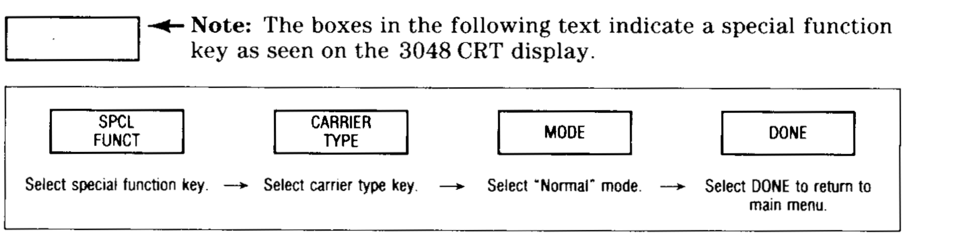
When making pulsed residual measurements the single-sided spur or double-sided spur calibration technique must be used if the system is to measure the phase detector constant automatically. Refer to the “Residual Phase Noise and AM Noise Measurements and Techniques” RF and Microwave Symposium paper for further information on spur calibration methods. The spur frequency should be kept below ½ PRF.

If the equipment required to calibrate using spurs is not available, the phase detector constant will have to be measured manually. The reason for this is that the 11848A circuitry cannot distinguish a PRF line from the desired beat note. Therefore, the user must manually measure and enter the phase detector constant. This is done by adjusting the phase shifter until the phase detector output voltage reaches its + and – peak voltage during the pulse-on time and then calculating the phase detector constant according to the formula:

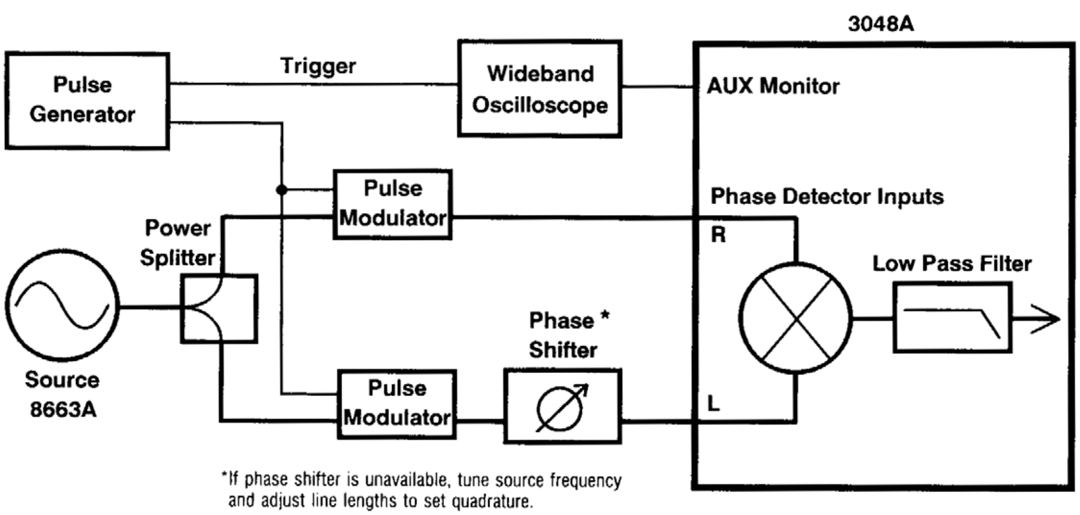
$$K\varphi_{PULSED} = \left(\frac{V_b+ \; + \; |\, V_b- \,|}{2}\right)\left(\frac{\tau}{T}\right)$$

Step-by-Step Measurement Procedure

The following step-by-step measurement procedure can be used when making residual measurements with the Agilent 3048A. Minor modifications will be required for different carrier frequencies, duty cycles, PRFs, source type, etc. Two measurement examples are given using the same fundamental procedure: 1) a noise floor measurement, and 2) a DUT measurement.



Example of how to use step-by-step measurement procedure



Example 1: Noise Floor using an Agilent 8663A as the source

PROCEDURE

1. Set up pulse generator

- Set correct output level for pulse modulator.
- Monitor pulse generator output on scope (dc coupled).
- Set pulse period and pulse width for: Duty cycle = 10% = τ/T PRF = 20 kHz = $1/T$
- Connect pulse generator output to pulse modulators.

2. Set up source

- Carrier frequency 1.5 GHz.
- Amplitude +16 dBm for 5 MHz to 1.6 GHz phase detector.
  +10 dBm for 1.2 to 18 GHz phase detector.

3. Set up the 3048A

SPCL FUNCT | CARRIER TYPE | Select Pulse mode | DONE

DEFINE MSRMNT

TYPE/ RANGE | Select:
- Measurement without PLL
- 10 Hz
- 100 kHz
- 4 averages

DONE

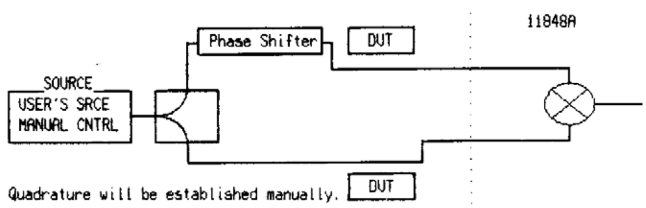
INST PARAMS | Select:
- 1.5 GHz carrier frequency.
- 1.5 GHz detector/input frequency.
- Internal phase detector 5 MHz to 1.6 GHz or internal 1.2 to 18 GHz phase detector, as appropriate.

DONE

CALIBR PROCESS | Select:
- Use current detector constant (input nominal constant, 30 mV for RF, 15 mV for microwave, will be measured later).

DONE

SOURCE CONTROL | Select:



DONE

DEFINE GRAPH | Select:
- Title
- 10 Hz to 100 kHz
- 0 to –170 dBc/Hz

DONE DONE

4. Calibrate detector constant

NEW MSRMNT | PROCEED

- At connect diagram, check hardware/connections.
- Connect scope to Aux Monitor port on 11848A Interface.
- Set scope to monitor waveform.
- Adjust phase shifter to positive peak (on scope) and record.
- Adjust phase shifter to negative peak and record.
- Detector constant = K (+ Vbpk + | –Vbpk |) (Duty Cycle)/2.

CALIBR PROCESS
- Input detector constant determined above.

DONE
- At connect diagram adjust phase shifter for quadrature, monitoring on scope.

5. PROCEED **with measurement**

– System will stop at the graph display LNA "IN" or "OUT."

– IF LNA "OUT" TOGGLE LNA "IN."

– Adjust phase shifter and monitor waveform on scope to determine if the LNA is saturated by the phase transients on the pulse.

– If the waveform is undistorted around quadrature, leave the LNA "IN" and PROCEED ; otherwise, TOGGLE LNA "OUT" and PROCEED

– System will measure phase noise.

Note: Do not move cables during measurement as this will change the results.

6. When the data is complete dump the graph and parameter summary.

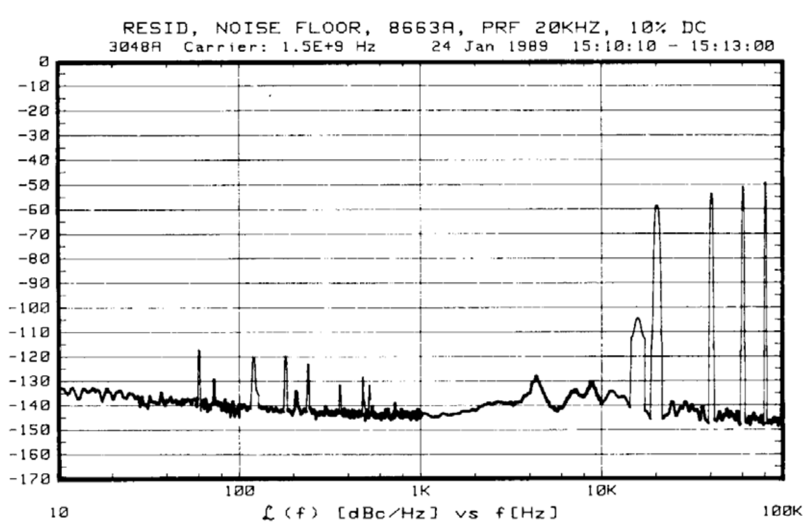
SHIFT HARD COPY will dump graphics and param summary.

Note the PRF feedthrough, aliasing effects, and any breaks in the graph.

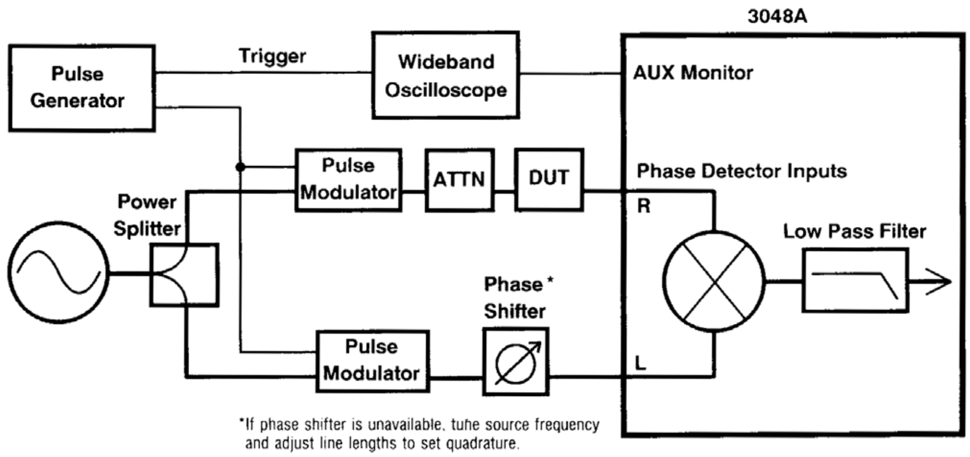
If PRF feedthrough is high and there are breaks in the graph, start a REPEAT MSRMNT and adjust the quadrature point at the connect diagram to reduce or change the PRF feedthrough.

Pertinent Measurement Parameters

Measurement Type:	Without Phase Lock	**K Detector Method:**	Entered
Start Offset Freq:	10 Hz	**Detector Constant:**	21 x 10^{-3}V/rad
Stop Offset Freq:	100 x 10^3 Hz		
Minimum Averages:	4	**Quadrature established by:**	Manual Phase Shifter
		Signal Source:	8662A Manual
Carrier Frequency:	1.5 x 10^9 Hz		
Detector Input Frequency:	1.5 x 10^9 Hz	**11848A LNA:**	In
Phase Detector:	1.2 to 18 GHz		



SETUP



Example 2: Residual noise of an Agilent 8447D amplifier

PROCEDURE

1. Set up pulse generator

- Set correct output level for pulse modulator.
- Monitor pulse generator output on scope (dc coupled).
- Set pulse period and pulse width for: Duty cycle = 10% = τ/T PRF = 20 kHz = 1/T
- Connect pulse generator output to pulse modulators.

2. Set up source

- Carrier frequency 1.25 GHz.
- Amplitude +16 dBm for 5 MHz to 1.6 GHz phase detector.
  +10 dBm for 1.2 to 18 GHz phase detector.

3. Set power into amplifier

- Measure the signal level out of the splitter on the RF spectrum analyzer.
- Attenuate the signal so that the amplifier is operating in its linear region (approximately 30 dB pad for the 8447D Amplifier).
- Connect the attenuated output of the splitter to the amplifier and measure the amplifier output to verify that the phase detector will not be damaged.

4. Set up the 3048A

[SPCL FUNCT] [CARRIER TYPE] Select Pulse mode [DONE]

[DEFINE MSRMNT]

[TYPE/ RANGE] Select:
- Measurement without PLL
- 10 Hz
- 100 kHz
- 4 averages

[DONE]

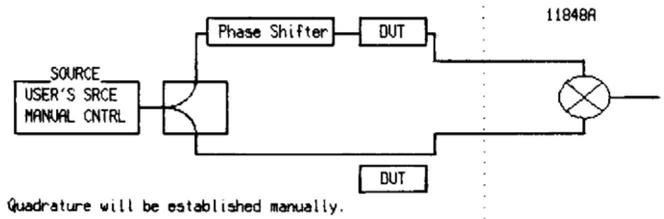
[INST PARAMS] Select:
- 1.25 GHz carrier frequency
- 1.25 GHz detector/input frequency.
- Internal phase detector 5 MHz to 1.6 GHz or internal 1.2 to 18 GHz phase detector, as appropriate.

[DONE]

[CALIBR PROCESS] Select:
- Use current detector constant (input nominal constant, 30 mV for RF, 15 mV for microwave, will be measured later).

[DONE]

SOURCE CONTROL — Select:



DONE

DEFINE GRAPH — Select:
- Title
- 10 Hz to 100 kHz
- 0 to –170 dBc/Hz

DONE DONE

5. Calibrate detector constant

NEW MSRMNT PROCEED

- At connect diagram, check hardware/connections.
- Connect scope to Aux Monitor port on 11848A Interface.
- Set scope to monitor waveform.
- Adjust phase shifter to positive peak (on scope) and record.
- Adjust phase shifter to negative peak and record.
- Detector constant = K (+Vbpk + | –Vbpk |) (Duty Cycle)/2.

CALIBR PROCESS – Input detector constant determined above.

DONE – At connect diagram adjust phase shifter for quadrature, monitoring on scope.

6. PROCEED **with measurement**

- System will stop at the graph display LNA "IN" or "OUT."
- IF LNA "OUT" TOGGLE LNA "IN."
- Adjust phase shifter and monitor waveform on scope to determine if the LNA is saturated by the phase transients on the pulse.
- If the waveform is undistorted around quadrature, leave the LNA "IN" and PROCEED ; otherwise, TOGGLE LNA "OUT" and PROCEED
- System will measure phase noise.

Note: Do not move cables during measurement as this will change the results.

7. **When the data is complete dump the graph and parameter summary.**

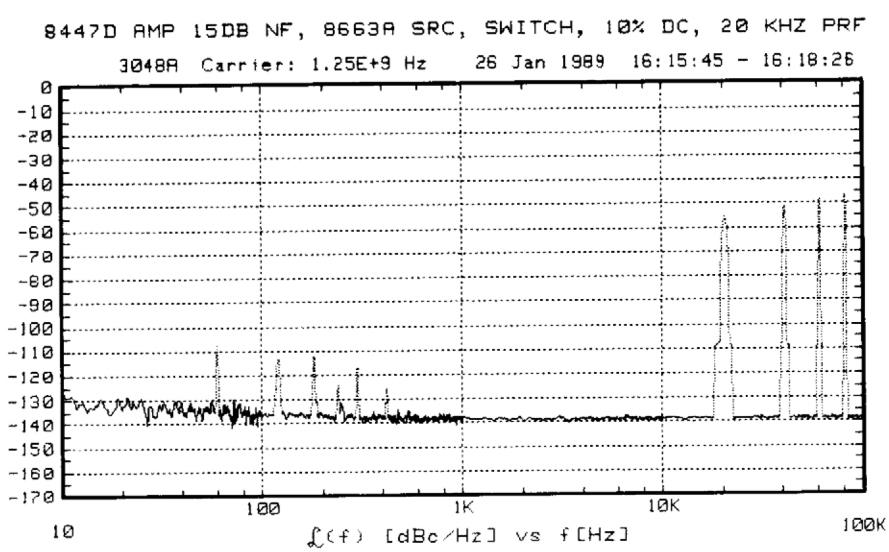
SHIFT [HARD COPY] will dump graphics and param summary.

Note the PRF feedthrough, aliasing effects, and any breaks in the graph.

If PRF feedthrough is high and there are breaks in the graph, start a [REPEAT MSRMNT] and adjust the quadrature point at the connect diagram to reduce or change the PRF feedthrough.

Pertinent Measurement Parameters

Measurement Type:	Without Phase Lock	**K Detector Method:**	Entered
Start Offset Frequency:	10 Hz	**K Detector Constant:**	12 x 10^{-3}V/rad
Stop Offset Freq:	100 x 10^{3} Hz		
Minimum Averages:	4	**Quadrature established by:**	Manual Phase Shifter
		Signal Source:	8662A Manual
Carrier Frequency:	1.25 x 10^{9} Hz		
Detector Input Frequency:	1.25 x 10^{9} Hz	**11848A LNA:**	In
Phase Detector:	5 to 1600 MHz		



Chapter 6. Using the Agilent 3048A System to Make Absolute Measurements on Pulsed Carriers

This chapter presents the recommended hardware configurations and step-by-step measurement procedures for making absolute phase noise measurements on pulsed carriers. For a complete discussion of how pulsing the carrier affects phase noise measurements and how to address these effects when using the Agilent 3048A, refer to Chapter 4. The recommended hardware configuration for using the Agilent 11729C Carrier Noise Test Set as a low noise downconverter for the 3048A will also be discussed.

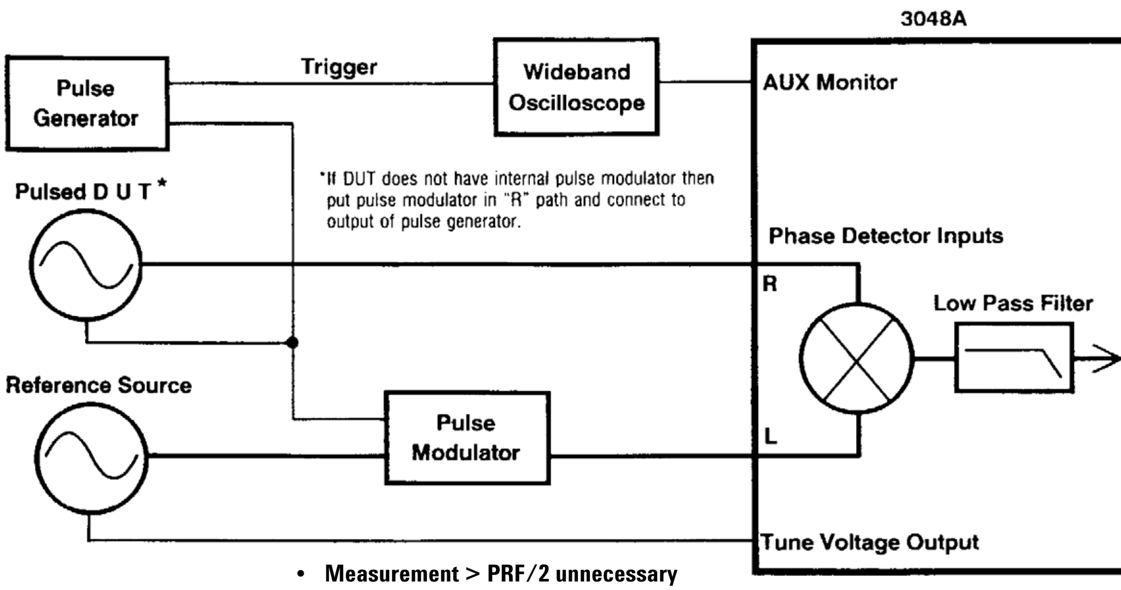
To successfully make a pulsed absolute measurement the operator must ensure that the following items are correct:

1. Hardware configuration
2. Software configuration

Hardware Configuration

The recommended hardware configuration for making pulsed absolute measurements using the 3048A System is shown in Figure 29.

The pulsed absolute measurement has the same special concerns as the pulsed residual measurement. Again, measurement offsets are useful to half the PRF, the noise floor scales with duty cycle and a PRF filter may be required to prevent overloading the LNA and baseband spectrum analyzer. Minimum duty cycle for a pulsed absolute measurement is 5% with the 3048A. This is higher than for a residual measurement due to the fact that a control signal for the phase locked loop must be generated which requires slightly higher signal levels out of the phase detector.



- **Measurement > PRF/2 unnecessary**
- **Duty cycles ≥ 5%**
- **Noise floor scales with duty cycle [degrades by 20_{log}(duty cycle)]**
- **PRF feedthrough may require PRF filter**

Figure 29. Recommended hardware configuration for making pulsed absolute measurements using the Agilent 3048A

Software Configuration

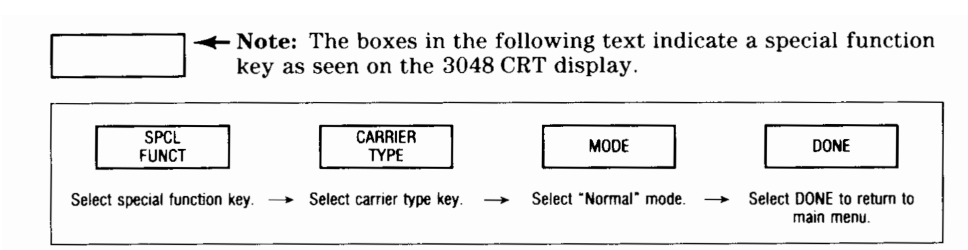
The Agilent 3048A software can measure two carrier types: CW and pulsed. When configured to measure pulsed carriers the software disables the zero beat routine for absolute measurements using the phase locked loop. This means that the operator must manually zero beat the sources to within 10% of the peak tune range. Additionally the user must select the “ignore out of lock” mode for the 3048A software. When “ignore out of lock” mode is selected all of the troubleshoot mode functions are enabled and the system will not check for an out of lock condition before or during a measurement. The “ignore out of lock” mode must be selected because of the PRF feedthrough. The out of lock circuitry will detect the PRF lines and interpret them as beatnotes. The presence of a beatnote indicates the loop is not phase locked. If the system was operating normally and an out of lock indication was detected the measurement would be stopped. When “ignore out of lock” mode is selected the operator is responsible for ensuring that phase lock is maintained during the measurement. This can be accomplished by connecting a wideband oscilloscope to the AUX monitor port on the 11848A Interface to verify the absence of a beatnote and to monitor the dc output level during the measurement.

The VCO tuning sensitivity cannot be measured under pulsed conditions, again due to the fact that the system cannot differentiate between a beat note and a PRF line. Consequently the VCO tuning constant must be entered manually. It can be determined from a CW measurement and entered for a pulsed measurement, or it can be known (as in the case of a signal generator) and entered into the system.

The peak tuning range (PTR) must be <PRF/2, otherwise when the phase detector constant is measured the beatnote will be confused with the PRF lines. This also guarantees that the PLL bandwidth will be less than the frequency of the PRF which is necessary for stability. If any of the PRF lines fell within the PLL bandwidth, the PLL would try to respond to their presence and the loop would become unstable.

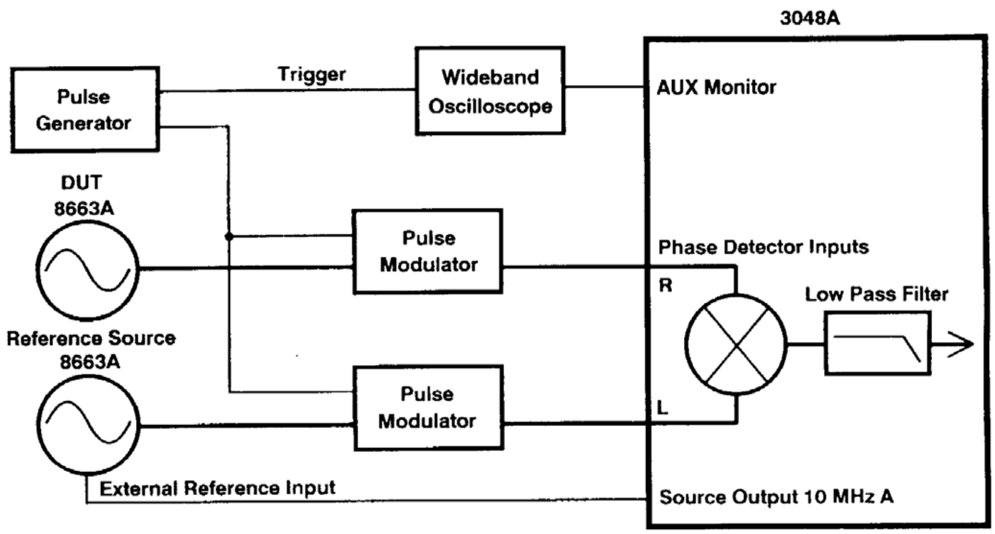
Step-by-Step Measurement Procedure

The following step-by-step measurement procedure can be used when making pulsed absolute measurements with the 3048A. Minor modifications will be required for different carrier frequencies, duty cycles, PRFs, source type, etc.



Example of how to use step-by-step measurement procedure

SETUP



Example 1: Agilent 8663A versus 8663A Pulsed Source Measurement

PROCEDURE

1. Set up pulse generator
- Set correct output level for pulse modulator.
- Monitor pulse generator output on scope (dc coupled).
- Set pulse period and pulse width for: Duty cycle = 10% = τ/T PRF = 20 kHz = 1/T
- Connect pulse generator output to pulse modulators.

2. Set up sources
- Carrier frequency 1 GHz.
- Amplitude +16 dBm for 5 MHz to 1.6 GHz phase detector.
  +10 dBm for 1.2 to 18 GHz phase detector.

3. Set up the 3048A

SPCL FUNCT	CARRIER TYPE	Select Pulse mode	
	MODE	Select: Ignore Out-of-Lock	DONE
DEFINE MSRMNT			
TYPE/ RANGE	Select:	– Measurement with PLL – 10 Hz – 40 MHz – 4 averages	DONE
INST PARAMS	Select:	– 1 GHz carrier frequency. – 1 GHz detector/input frequency. – Entered K_{VCO} (as appropriate).* – Center voltage: 0 – Tune range (as appropriate) – Phase Detector: 5 MHz to 1600 MHz	DONE

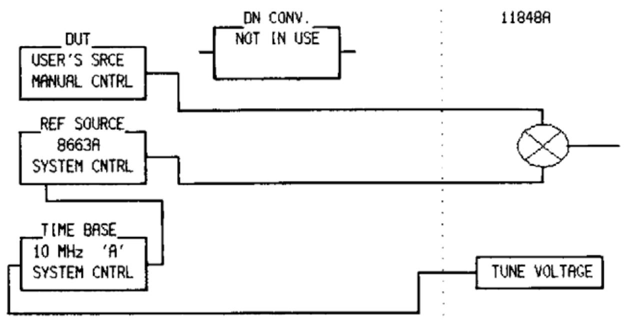
* The VCO Tuning Sensitivity cannot be measured under pulsed conditions. If the tuning sensitivity is not known accurately, measure the tuning sensitivity CW and enter the value.

Note: Verity that the PTR = K_{VCO} x Tune Range ≤PRF/2

CALIBIR PROCESS — Select:
- Measure the detector constant.
- Use current Tune Constant (enter) or Compute from entered K_{VCO}
- Phase Lock Loop WILL be verified.

DONE

SOURCE CONTROL — Select:



DONE

DEFINE GRAPH — Select:
- Title
- 10 Hz to 40 MHz
- 0 to –170 dBc/Hz

DONE DONE

4. Calibrate detector constant

NEW MSRMNT   PROCEED

- At connect diagram, check hardware/connections.
- Connect scope to Aux Monitor port on 11848A Interface.
- Set scope to monitor waveform.
- Verify that the beatnote is at "zerobeat." In pulse mode for source measurements the system will not verify zero beat and this must be done manually.
- Observe the beatnote on the 3561 while tuning the signal generator frequency to produce a beatnote that is <10% of the PTR.

5. PROCEED **with measurement**

- System will measure the phase detector constant and display the detector constant and approximate system noise floor.
- The system will determine whether or not to leave the LNA IN or OUT. Observe the pulse waveform on the oscilloscope, and toggle the LNA IN. If the waveform peak is less than 500 mV and it is not compressed or clipped severely, leave the LNA in and continue with the measurement.

PROCEED

- System will measure phase noise.

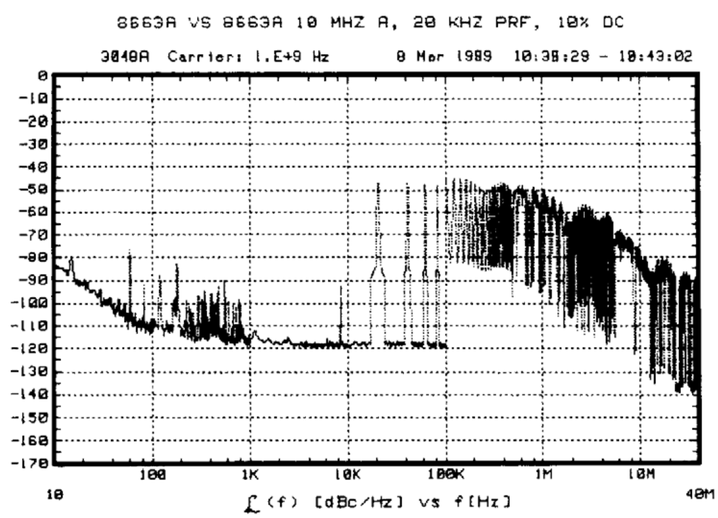
6. When the data is complete dump the graph and parameter summary.

SHIFT   HARD COPY   will dump graphics and parameter summary.

Note the PRF feedthrough, aliasing effects, and any breaks in the graph.

Pertinent Measurement Parameters

Measurement Type:	Phase Locked	**K_{VCO} Method:**	Entered
Start Frequency:	10 Hz	**K_{VCO}:**	1.7 x 10^3 Hz/Volt
Stop Frequency:	40 x 10^6 Hz		
Averages:	4	**Loop Suppr.:**	Verified
		Clsd Pll Bw.:	216.9 Hz
Carrier Frequency:	1 x 10^9 Hz	**Peak Tuning Range:**	2.58 x 10^3 Hz
Detector Input Frequency:	1 x 10^9 Hz	**Assumed Pole:**	738 Hz
Entered K_{VCO}:	1.7 x 10^3 Hz/volt	**D.U.T.:**	User's Source, Manual
Center Voltage:	10 x 10^{-3} Volts	**Reference Source:**	8663A, SYS
Tune Range:	±2 Volts	**Ext. Timebase:**	10 MHz 'A,' SYS, VCO
Phase Detector:	5 to 1600 MHz	**Down converter:**	Not in use
K Detector Method:	Measured	**11848A LNA:**	In
K Detector Method:	40.24 x 10^{-3} V/rad		



Using the Agilent 11729C with the 3048A

The recommended configuration for using the 11729C Carrier Noise Test Set to make absolute phase noise measurements on pulsed microwave carriers is to use it in conjunction with the 3048A system. In this configuration the 11729C is used as a low noise downconverter for the 3048A system. When either the 11729C/8662A or the 11729C/8663A is used as a low noise downconverter for the 3048A, only the microwave downconversion components in the 11729C are utilized. The pulsed microwave DUT is downconverted against the 640 MHz combline, and the IF output is phase detected in the 11848A Phase Noise Interface. None of the phase detector or phase lock loop circuitry in the 11729C is used in this configuration. In this configuration the 3048A controls the phase locked loop and drives the spectrum analyzers. The 3048A system software automatically compensates for loop suppression and spectrum analyzer effects. When the 11729C/866X is used as a low noise downconverter for the 3048A, the user has a calibrated, specified, automated measurement system with the lowest noise microwave reference available from Agilent.

When the 11729C is used in conjunction with the 3048A, the accuracy of the measurement is specified at ±2 dB to offsets ≤1 MHz, and ±4 dB at offsets from >1 MHz to ≤40 MHz. The recommended hardware configuration for making pulsed absolute measurements using the 3048A/11729C is shown in Figure 30.

If a PRF filter is required, an external phase detector with PRF filter can be supplied as a special from Agilent.

The IF amplifier in the 11729C is a very high gain amplifier, so there is the possibility that during the pulse off interval there will be high level broadband noise on the IF output of the 11729C, which could potentially degrade the measured noise. If this is the case then one solution is to pulse (synchronous with the DUT) the other path into the phase detector. As discussed in Chapter 4, pulsing both paths into the phase detector also eliminates many problems associated with pulsing only the DUT path.

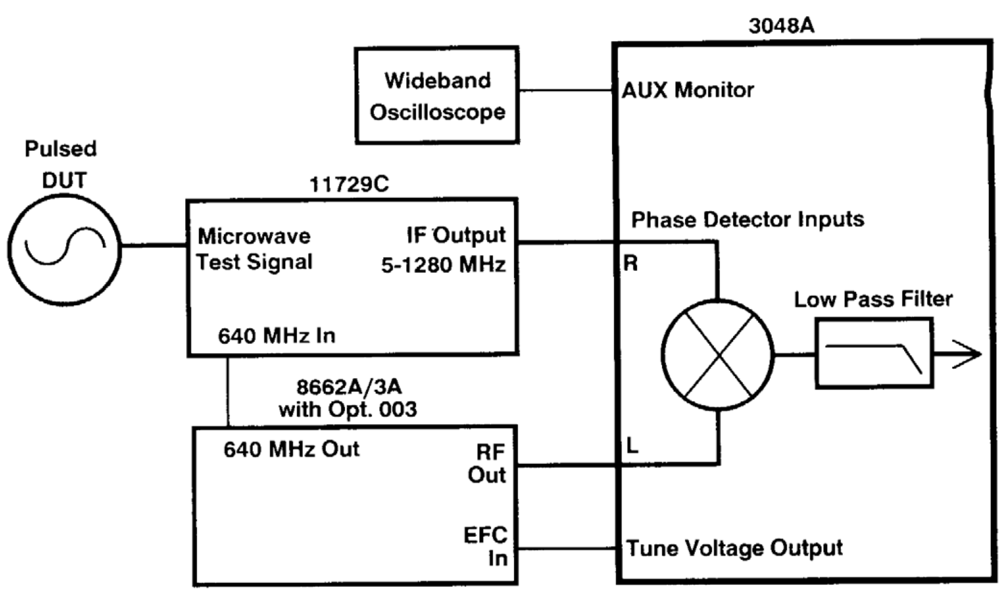


Figure 30. Recommended hardware configuration for making pulsed absolute measurements using the Agilent 3048A/11729C

Interpreting the Results

Figure 31 shows several plots of pulsed absolute phase noise measurements of an Agilent 8673D synthesizer. The graph on the top has a pulse width of 100 nanoseconds, a PRF of 1 MHz and a duty cycle of 10%. Note the 1 MHz and integer multiples of the PRF. If the phase noise of the pulsed measurement is compared to a CW measurement, one can see that the phase noise is identical out to 300 kHz and then the noise contributions from the PRF lines can be seen.

The phase noise plot on the bottom is the same source at a 10% duty cycle, but the PRF is now 10 kHz and the pulse width is 10 microseconds. The first PRF line appears at 10 kHz. One would expect the measured noise to be identical to that of the CW case out to an offset frequency of 5 kHz (i.e., ½ PRF), but on close inspection we see that the noise is the same out to only 1 kHz due to the sampling down or aliasing of noise at offsets above 5 kHz. Notice on the 1 MHz PRF plot that there is a significant spur at approximately 6 kHz. This spur is also present on the 10 kHz PRF plot but there is now a second spur at an offset of 4 kHz. This new spur is due to mixing of the 10 kHz PRF and the 6 kHz spur, causing a mixing induced spur at 4 kHz (the difference frequency). The break in the graph at 5 MHz is due to the fact that the PRF lines are too close together to be resolved by the spectrum analyzer. The RBW and the VBW of the spectrum analyzer could be adjusted to resolve these spurs by changing the segment table in the Agilent 3048A system software.

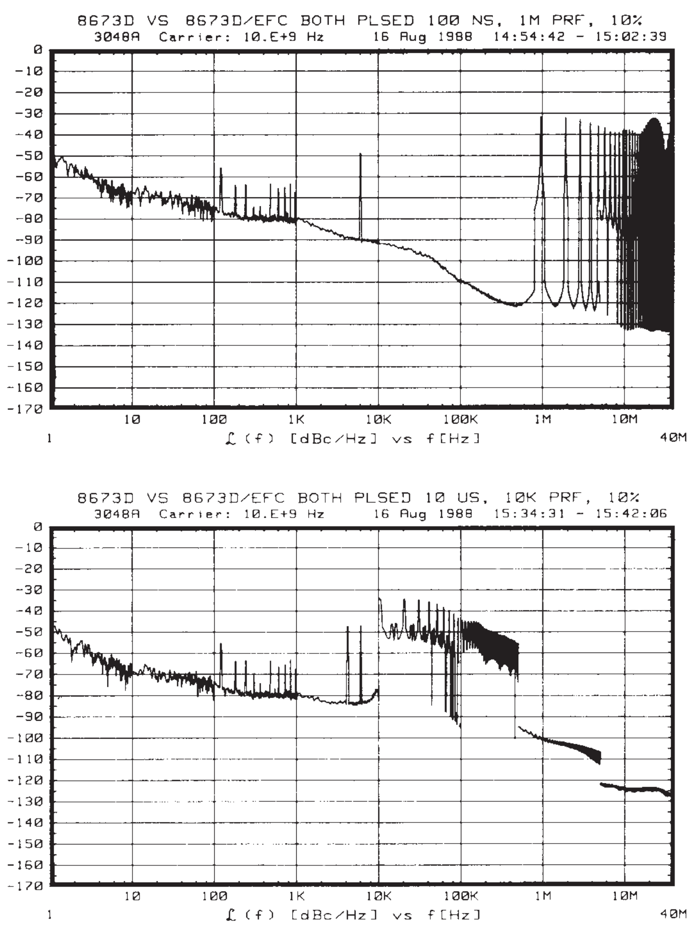


Figure 31. Pulsed synthesizer absolute phase noise plots

Chapter 7. Using the Agilent 3048A System to Make AM Noise Measurements on Pulsed Carriers

This chapter presents the recommended hardware configurations and step-by-step measurement procedures for making AM noise measurements on pulsed carriers. For a complete discussion of the problems and workarounds associated with making pulsed carrier phase noise measurements using the Agilent 3048A, refer to Chapter 4.

To successfully make a pulsed AM noise measurement the operator must ensure that the following items are correct:

1. Hardware configuration
2. Software configuration

Hardware Configuration

The recommended hardware configuration for making AM noise measurements using the 3048A System is shown in Figure 32.

The inside/outside dc block prevents ground loops. Ground loops can introduce additional noise into the measurement which would degrade the accuracy of the measurement. The AM detector consists of a low barrier Schottky diode detector and a filter network. The Schottky diode detectors will handle more power than the point contact detectors, and are equally sensitive and quiet. The AM detector filter network prevents the dc voltage component of the demodulated signal from saturating the system's low noise amplifier, and also sets the detector dc bias. Since this measurement is essentially doing envelope detection of the DUT signal, PRF feedthrough will be very high. This PRF feedthrough will overload the LNA, so a PRF filter is required to reduce the level of the PRF feedthrough.

As in the phase detector measurement, the AM detector must be calibrated. Most DUTs do not have calibrated AM modulation. It is usually necessary to substitute a second pulsed source with calibrated AM modulation in order to calibrate the AM detector. The basic calibration techniques are documented in the "Residual Phase Noise and AM Noise Measurements and Techniques" RF & Microwave Symposium Paper, p/n 1000-1126, available from your local Agilent sales office. The calibration methods are the same under pulsed conditions as under CW conditions with the caution that the calibration sidebands be <PRF/4.

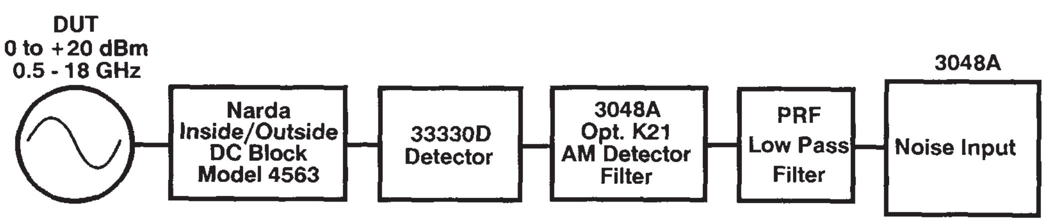


Figure 32. Recommended hardware configuration for making pulsed AM noise measurements using the Agilent 3048A

AM noise measurements are basically source measurements. The residual AM noise of a two-port device can be calculated by measuring the source's AM noise, then subtracting that from the measured output noise of the DUT, or by verifying that the source AM noise is less than the measured residual noise. The noise floor of this technique is the sum of the noise floor of the source and the measurement system.

Software Configuration

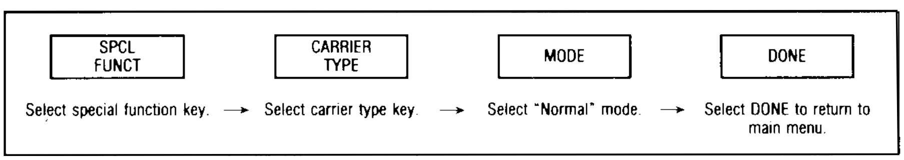
Pulsed AM noise measurements can be made using either CW carrier type or pulsed carrier type. The advantage of the pulsed carrier type is that it provides LNA control during calibration and measurement.

The system should be set to NORMAL mode.

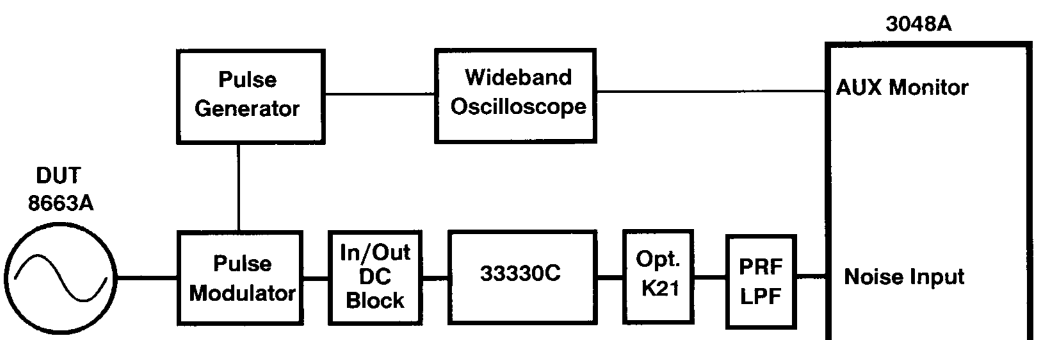
Step-by-Step Measurement Procedure

The following step-by-step measurement procedure can be used when making pulsed AM noise measurements with the Agilent 3048A. Minor modifications will be required for different carrier frequencies, duty cycles, PRFs, source type, etc.

Note: The boxes in the following text indicate a special function key as seen on the 3048 CRT display.



Example of how to use step-by-step measurement procedure



SETUP

- **Inside/outside dc block prevents ground loops.**
- **K21 provides detector bias and blocking cap.**
- **LPF filters PRF feedthrough.**

Example 1. Pulsed AM Noise Measurement of the Agilent 8663A

PROCEDURE

1. Set up pulse generator
- Set correct output level for pulse modulator.
- Monitor pulse generator output on scope (dc coupled).
- Set pulse period and pulse width for: Duty cycle = 10% = τ/T PRF = 20 kHz = 1/T
- Connect pulse generator output to pulse modulators.

2. Set up source (Calibrate and measure AM noise on same source)
- Carrier frequency 2 GHz.
- Amplitude +16 dBm for 5 MHz to 1.6 GHz phase detector.
  +10 dBm for 1.2 to 18 GHz phase detector.

3. Set up the 3048A

[SPCL FUNCT] [CARRIER TYPE] Select Pulse mode* [DONE]

[DEFINE MSRMNT]

* Pulsed AM noise measurements can be made using normal or pulsed mode. Pulse mode provides LNA control during calibration and measurement.

[TYPE/ RANGE] Select:
- AM Noise measurement
- 10 Hz
- 100 kHz
- 4 averages

[DONE]

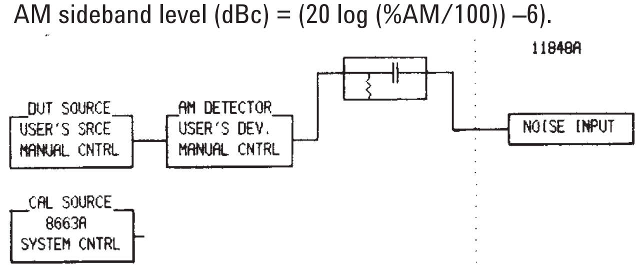
[INST PARAMS] Select:
- 2 GHz carrier frequency.
- 2 GHz detector/input frequency.
- Select external phase/AM detector.

[DONE]

[CALIBR PROCESS] Select:
- Double-sided Spur Method.
  (Input AM rate and approximate sideband level; system will calculate appropriate % AM for sideband level using:
  AM sideband level (dBc) = (20 log (%AM/100)) –6).

[DONE]

[SOURCE CONTROL] Select:



[DONE]

[DEFINE GRAPH] Select:
- Title
- 10 Hz to 100 kHz
- 0 to –170 dBc/Hz

[DONE] [DONE]

4. Calibrate detector constant

Most sources do not have calibrated AM modulation, it is usually necessary to substitute a second pulsed source with AM modulation in order to calibrate the AM detector. The basic calibration techniques are documented in the *"Residual Phase Noise and AM Noise Measurements and Techniques"* paper. The calibration methods are the same under pulsed conditions as under CW conditions with the caution that the calibration sidebands be <PRF/4.

The challenge for a pulsed AM measurement is to determine the correct power level to set on the pulsed calibration source. To set the calibration source power level:
- Connect the pulsed source under test output of the AM detector to the scope (50 Ω) at the CONNECT DIAGRAM.
- Measure peak voltage of pulsed source under test.
- Remove Source Under Test and substitute Calibration source.

– Set pulsed calibration source output to the same peak level and duty cycle measured for the source under test.
– Turn on AM and proceed with detector calibration.

NEW MSRMNT   PROCEED

– At connect diagram, check hardware/connections.
– Connect scope to Aux Monitor port on 11848A Interface.
– Set scope to monitor waveform.
– Verify that the calibration source is connected and AM is set.

5. PROCEED **with calibration**

– System will measure the calibration tone and display the detector sensitivity and system noise floor.

6. Connect diagram

– Disconnect the calibration source and replace with the source under test or simply turn off AM.

PROCEED with the AM noise measurement.

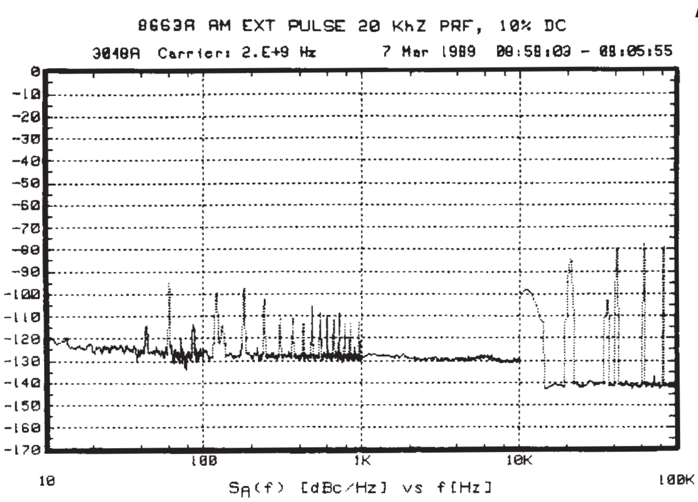
7. When the data is complete dump the graph and parameter summary.

SHIFT   HARD COPY   will dump graphics and parameter summary.

– Note the PRF filter cutoff.

Pertinent Measurement Parameters

Measurement Type:	AM Noise	**K Detector Method:**	Double Sided Spur
Start Frequency:	10 Hz	**Spur Amplitude:**	–14.515 dBc
Stop Frequency:	100 x 10^3 Hz	**Spur Frequency:**	5 x 10^3 Hz
Minimum Averages:	4	**K Detector Constant:**	6.231 x 10^{-3} V/rad
Carrier Frequency:	2 x 10^9 Hz	**OUT Source:**	User's Source, Manual
Detector Input Frequency:	2 x 10^9 Hz	**Cal. Source:**	8663A, SYS
Phase Detector:	External	**AM Detector:**	User's Device, Manual
		Agilent 11848A LNA:	In



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Printed in U.S.A. 11/00
5951-6743

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