Jitter Measurements Using SpectreRF Application Note

Product Version MMSIM6.0
March 2006

Contents

Measuring Jitter

The procedures described in this application note are deliberately broad and generic. Your specific design might require procedures that are slightly different from those described here.

Purpose

This application note illustrates how to use the SpectreRF simulator within the Analog Design Environment (ADE) to measure jitter characteristics of the typical blocks which are used in analog and digital circuit design.

Audience

Users of SpectreRF in the Analog Design Environment.

Overview

This application note describes how to measure the jitter characteristics of blocks typically used in analog and digital circuits. It introduces a jitter measurement methodology in the framework of the SpectreRF time domain simulator. It also describes jitter metrics which sometimes create confusion when they are used inconsistently.

There are two major approaches to modelling jitter.

- Model jitter as the continuous analog output signal of an autonomous system such as a voltage controlled oscillator (VCO)
- Use an edge-sensitive thresholding circuit to observe and sample the output signal. This approach is primarily used for driven circuits.

In conclusion we present examples of both driven and autonomous circuits.

Timing Jitter and Its Metrics

Time (or phase) jitter is a key measurement in systems where the periodic behavior or exact event timing is crucial to system performance. The design process for high performance, low cost devices in today's communication systems requires comprehensive methods for modeling and analyzing both jitter and noise.

Jitter measures undesired fluctuations in the timing of events.

- In oscillators and frequency synthesizers period jitter quantifies the time domain uncertainty of the output signal.
- In thresholding and digital period jitter describes the misalignment of the significant edges of a digital signal from their ideal position in time.

In communication systems, "jitter is an abrupt and unwanted variation of one or more signal characteristics, such as the interval between successive pulses, the amplitude of successive cycles or the frequency or phase of successive cycles." See [1} in "References" on page 29 for more information.

Phase and Timing Jitter

Jitter can be considered as an undesired perturbation in time. You can represent a noisy signal as a noise-free signal $v_{ideal}(t)$ displaced in time with a stochastic process $j(t)$. The noisy signal is modeled as shown in Equation 1-1.

(1-1) $$v_n(t) = v_{ideal}(t + j(t))$$

The noisy signal can also be modeled in terms of the phase noise $\phi(t)$ as shown in Equation 1-2.

(1-2) $$v_n(t) = v_{ideal}\left(t + \frac{\phi(t)}{2\pi}T\right)$$

As you can see, phase noise leads to jitter.

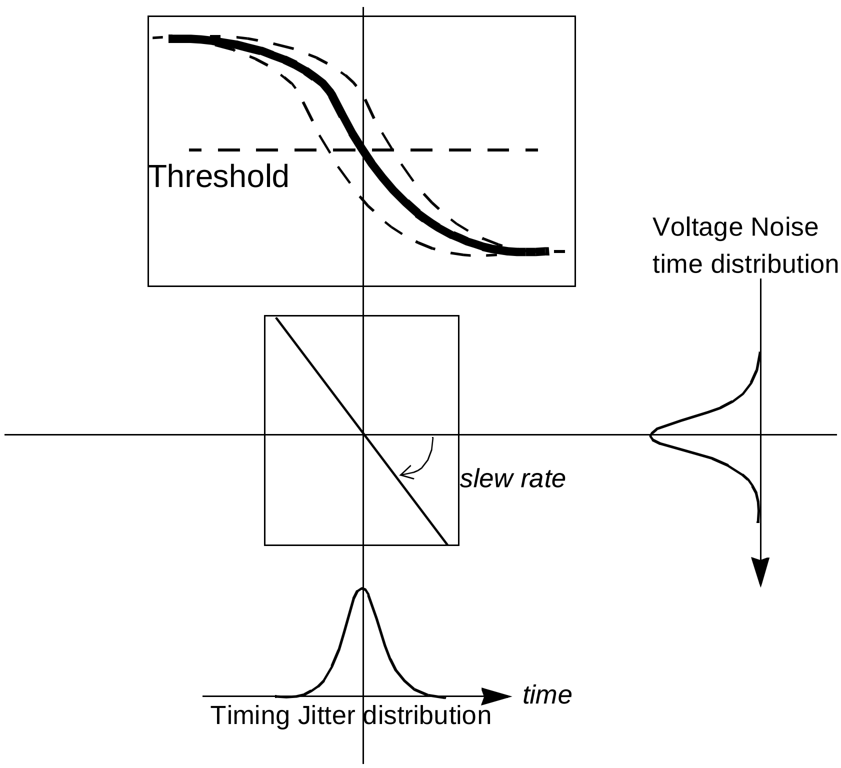
One definition of phase noise demonstrates its connection to jitter - *phase noise is rapid, short-term, random fluctuations in the phase of a wave, caused by time domain instabilities*, See [1} in "References" on page 29 for more information.

Important

In the remaining discussion, you can assume that the original signal and *j(t)* are both *T*-periodic functions. They both vary periodically with period *T*.

There are also situations when the noisy output signal is sampled while it is crossing the threshold. Since the noise $n(t)$ is cyclostationary (see [4] in "References" on page 29), its power varies periodically over time.

Figure 1-1 Jitter at the Threshold Crossing



The voltage noise at the point where the threshold is crossed causes the time shift of the edge by way of the slew rate of the signal as shown in Equation 1-4 and in Figure 1-1.

(1-3) $$v_n(t) = v_{ideal}(t) + n(t)$$

(1-4) $$var(j(t_{cross})) \cong \frac{var(v_n(t) - v_{ideal}(t))}{SlewRate^2} = \frac{var(n(t_{cross}))}{\left(\frac{d}{dt} v_{ideal}(t_{cross})\right)^2}$$

The variance $var(n)$ is the expected value of the square of the deviations of a random signal from its mean value and t_{cross} is the time of the threshold crossing.

Because the noise is assumed to be small enough that it does not affect the solution in any way that will significantly change the signal, you can approximate the actual signal slew rate with the slew rate you get from the ideal, jitter free, large signal model. Small signal approximation requirements and limitations also exist when you use the Pnoise analysis and its variations.

Jitter Metrics

Various jitter metrics characterize the statistics of the observed sequence of events.

When a periodic system has a reference time point, you can consider all other events in relation to that time point. For a driven system, such an event can occur when the periodic input signal crosses through a particular threshold value. The periodic output signal generates an ideal sequence of events in response to the input signal, separated by the period T of the system. Ideally the delay from the input edge to the output edge is a constant t_{delay}. This sequence of same direction threshold crossings by the output signal expects periodic threshold crossings at times $\{ iT + t_{delay} \}$ where the events are indexed with i.

(1-5) $$v_{ideal}((i+1) \cdot T + t_{delay}) = v_{ideal}(iT + t_{delay}) = v_{threshold}$$

The actual transition times $\{ t_i \}$ of the noisy signal will vary from the expected time as

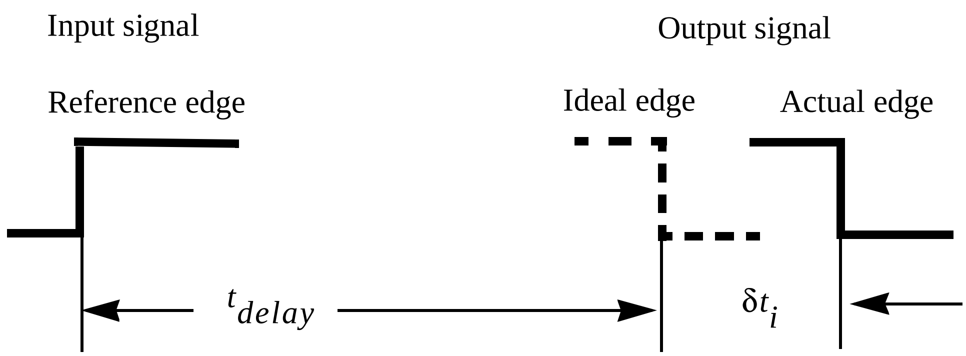
(1-6) $$v_n(t_i) = v_{ideal}(iT + t_{delay} + j(t_i)) = v_{threshold}$$

The uncertainty is represented by the sequence

(1-7) $$\{\delta t_i\} = \{t_i - (iT + t_{delay})\}$$

This variation in the delay between a triggering event and a response event is the *edge-to-edge timing jitter as shown in Figure* 1-2, It is also called *absolute* or *aperture jitter.*

Figure 1-2 Edge-to-Edge Jitter



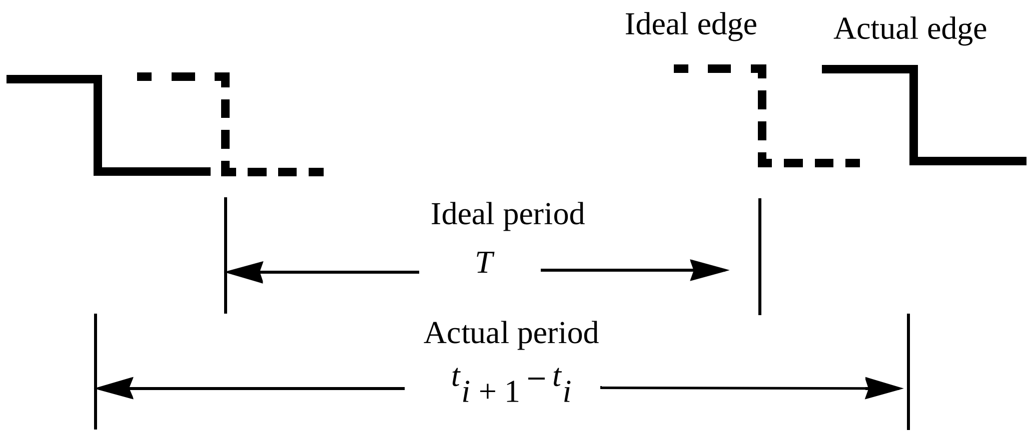
(1-8) $$J_{ee}^{i} = \sqrt{var(t_i)}$$

The presence of a noise free input signal implies that the edge-to-edge jitter metric can only apply for driven systems.

The cycle or period between two similar output signal transitions is affected by the noise on both edges. The variation in the length of the period from the ideal value (as expressed in Equation 1-9), is characterized by the standard deviation also known as the *period* jitter or *cycle* jitter, see Figure 1-3 and Equation 1-10.

(1-9) $$J_{c}^{i} = \{t_{i+1} - t_i - T\}$$

Figure 1-3 Period Jitter

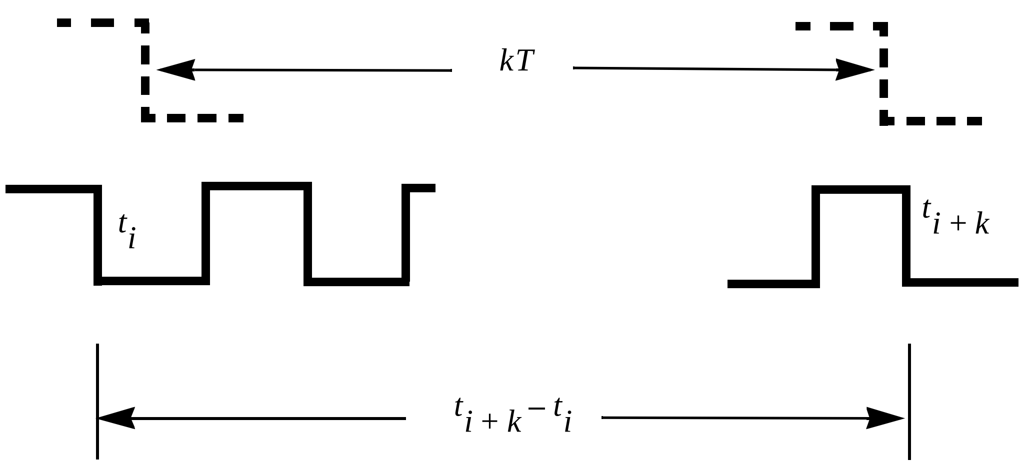


(1-10) $$J_c^i = \sqrt{var(t_{i+1} - t_i)}$$

If two transitions are separated in time by *k* cycles or periods, as shown in Figure 1-4, the standard deviation of the length of *k* cycles from nominal is computed using the following sequence.

(1-11) $$\left\{j_c^i(kT)\right\} = \{t_{i+k} - t_i - kT\}$$

Figure 1-4 K-Period or K-Cycle Jitter



It is designated as long term *k-cycle jitter* or *k-period jitter* computed.

(1-12) $$J_c^i(kT) = \sqrt{var(t_{i+k} - t_i)}$$

Another metric, *cycle-to-cycle jitter*, is produced by the second difference of *absolute jitter*.

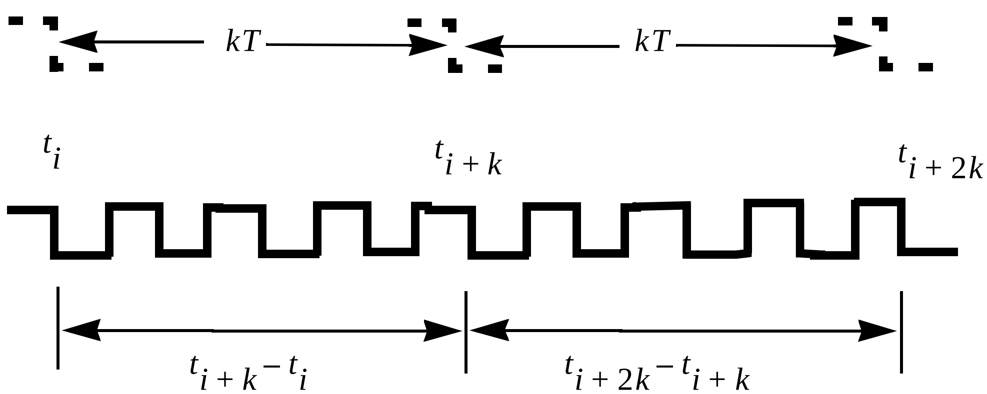
(1-13) $$\left\{ j_{cc}^i \right\} = \{((t_{i+2} - t_{i+1}) - (t_{i+1} - t_i)\}$$

Also called *adjacent period jitter,* it characterizes the local change in the period as

(1-14) $$j_{cc}^i = \sqrt{var(t_{i+2} - t_{i+1}) - (t_{i+1} - t_i)}$$

Figure 1-5 illustrates *cycle-to-cycle jitter* or *adjacent period jitter.*

Figure 1-5 Cycle-to-Cycle Jitter or Adjacent Period Jitter for General k



In the more general case of k cycles adjacent to another k cycles the metric is the sequence

(1-15) $$\left\{j_{cc}^{i}(kT)\right\} = \{(t_{i+2k} - t_{i+k}) - (t_{i+1} - t_i)\}$$

The standard deviation of the length of *k* adjacent periods is a long term jitter called either *k-long cycle-to-cycle jitter* or *k-long adjacent period jitter* as in Equation 1-16.

(1-16) $$j_{cc}^{i}(kT) = \sqrt{var(t_{i+2k} - t_{i+k}) - (t_{i+k} - t_i)}$$

For stationary or T-cyclostationary noise, the metrics do not depend on the location of the events, which is defined by "i".

Note: Simplified notations, as shown in Table 1-1, are used in the remainder of this document.

Table 1-1 Simplified Jitter Notations

Notation	Definition
J_{ee}	Edge-to-Edge Jitter
J_c	Cycle Jitter
J_{cc}	Cycle-to-Cycle Jitter
$J_c(k)$	K-Cycle Jitter
$J_{cc}(k)$	K-Long Cycle-to-Cycle Jitter

Synchronous Jitter Versus Accumulating Jitter

Circuits that are driven by externally generated periodic signals or clocks, exhibit *synchronous jitter*. In this case you can measure the *absolute jitter*, since the reference input edge exists in the system. The name *edge-to-edge jitter* is reserved for the input referred jitter in such systems.

The measurement is between the ideal input signal and the noisy output signal. The cycle (or period) jitter and the cycle-to-cycle (or adjacent period) jitter are also measured for driven systems, but they are different in the sense that the measurement starts and ends with a noisy output signal. But even in this situation, both transitions are the direct result of a transition by the ideal periodic input signal.

For autonomous circuits where there are no ideal reference transitions, you are limited to using self-referred jitter metrics. This jitter is *accumulating jitter*. The next cycle transition is the result of the previous cycle output, so the jitter variance accumulates from cycle-to-cycle.

The *synchronous* jitter for driven systems is measured using strobed periodic noise simulation in SpectreRF. The slew rate is also automatically computed at the requested threshold crossing events. The final results are presented using the Direct Plot form. The entire simulation is executed in one run as long as you know the values of the threshold and outputs of the system.

The accumulating jitter is computed using phase noise measured by SpectreRF.

The conversion between phase noise, or strobed noise, and jitter will be described in the sections to follow.

Jitter Computation

Both strobed noise and phase noise measurements in SpectreRF produce the power spectral density (PSD) of the random noise which can be converted to the RMS value of random jitter.

In the case of synchronous jitter, the slew rate is used to convert the strobed noise PSD into absolute time jitter PSD. By definition, the scalar value of the absolute jitter is the total area of its power spectrum.

(1-17)
$$J_e^2 = \int_{-f_c/2}^{f_c/2} S_j(f)df = \frac{2}{\left(\frac{d}{dt}v_{ideal}(t_c)\right)^2}\int_0^{f_c/2} S_n(f)df$$

where $S_n(f)$ is the instant noise power spectrum from sampled noise analyses. The lower limit of integration is selected based on several considerations, such as system bandwidth and the validity of the flicker noise model, which can grow unbound as the frequency decreases.

Other metrics are computed from the spectral density of the absolute jitter using Equations 1-18 and 1-19.

(1-18) $$J_c^2(kT_c) = 4\int_{f_{lower}}^{f_{upper}} S_j(f)\sin^2(\pi k f T_c)df$$

and

(1-19) $$J_{c-c}^2(kT_c) = 16\int_{f_{lower}}^{f_{upper}} S_j(f)\sin^4(\pi k f T_c)df$$

The same Equations 1-18 and 1-19 are used for autonomous systems. These Equations use phase noise and its relationship with the time jitter which follows from Equations 1-1 and 1-2.

(1-20) $$S_j(f) = \frac{S_\phi(f)}{(2\pi f_c)^2}$$

Converting RMS to Peak-to-Peak Jitter

The random jitter addressed in this document is typically characterized in terms of root mean squared average (or RMS). It is a natural way to represent the unbounded probability distribution function of a random process. In order to combine or analyze the random jitter with deterministic jitter, which is bounded, there is a way to convert RMS into the peak-to-peak bound under certain conditions.

Assuming that only a finite amount of error is allowed, you can find the value of jitter, beyond which, the error will be less probable than the established limit. A bit error rate (BER) is such a limit. For a Gaussian distribution of the jitter, you can find the scaling factor to convert RMS to the peak-to-peak range. SpectreRF uses a table which has the scaling factor for the most typical values of BER.

Long Term Jitter

Long term jitter is of limited use because of the limitation of small signal analyses we use. The variations that SpectreRF models cannot grow large enough to break important assumptions about the validity of small signal approximation. Large low frequency noise components that lead to large long term jitter break the assumption that the noisy curve is smooth on transition. If operating point is affected by the noise, the slew rate of the jitter free signal can not be used to approximate long term jitter.

Jitter Units in SpectreRF

Measuring timing jitter involves several basic units. The default unit is seconds. Absolute units such as seconds are not always optimal. Using absolute units, you cannot see the effect of the jitter on the system period or on other timing properties.

Relative units such as parts per million (PPM) and the unit interval (UI), UI can be the clock period or another characteristic time cycle.

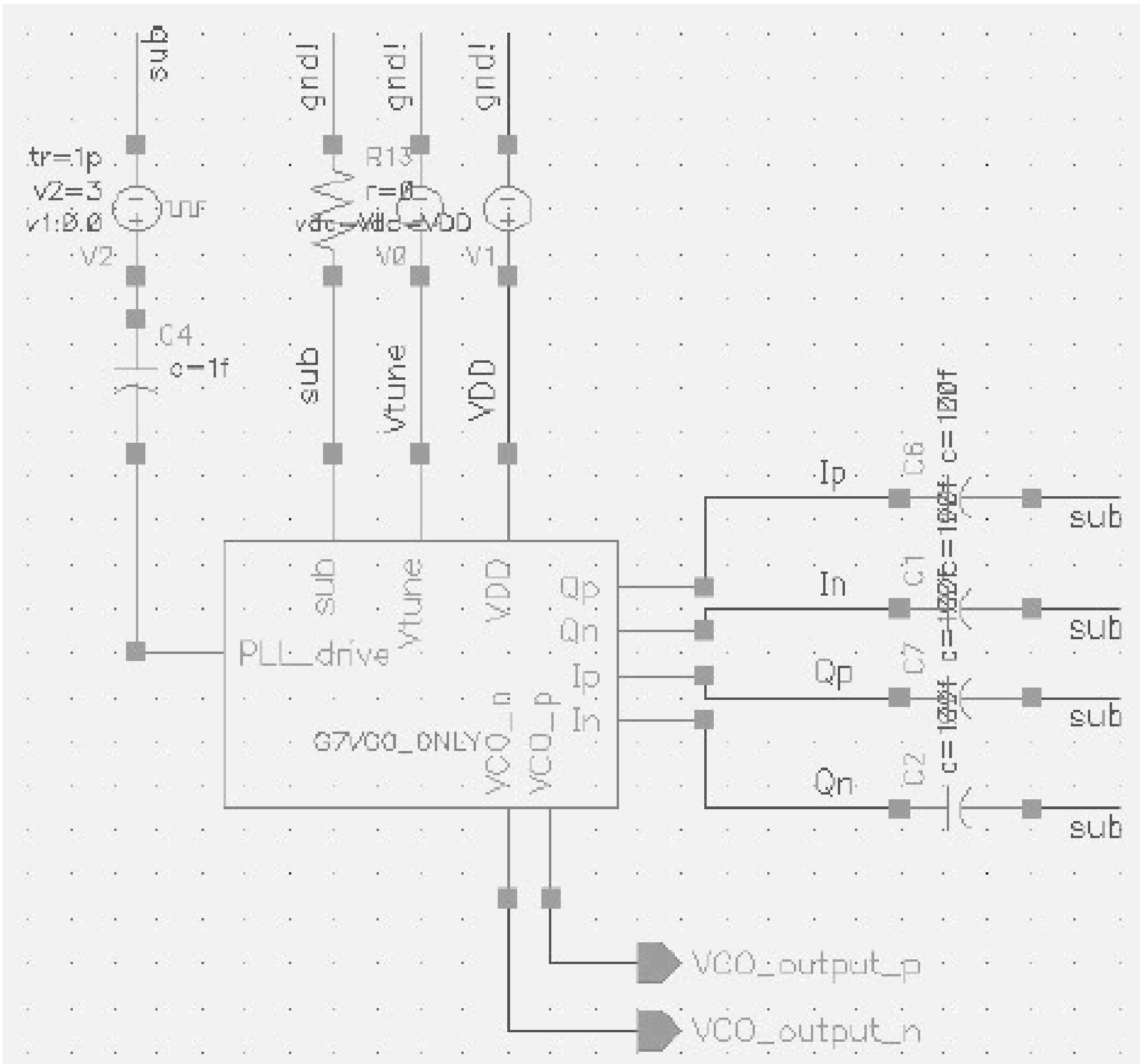
Jitter and Phase Noise Measurement Examples.

Differential VCO

This CMOS differential voltage controlled oscillator is a part of the design by Helic S.A. Intended for the wide range Wireless LAN. It is a part of the block that includes the VCO and Divide by 2 Prescaler. It uses a Cadence Generic CMOS PDK. Typical noise parameters were added to the original CMOS model for the demonstrative purposes only. The VCO runs at f(LO)=5GHz. The output noise of interest is at offset frequency of 1kHz to 2.5GHz from the frequency of the LO signal.

In the first example we are only interested in the loaded VCO behavior. In order to have an efficient simulation, we use only small input part of the Prescaler as a dummy load without frequency division. The top level is shown on Figure 1-6.

Figure 1-6 Differential VCO



If the VCO with actual frequency divider is to be simulated, the output of the oscillator will be at the M^{th} harmonic, where M is the divide ratio. The PNoise output offset frequency need to be relative to the M^{th} harmonic.

Measurements Testbench

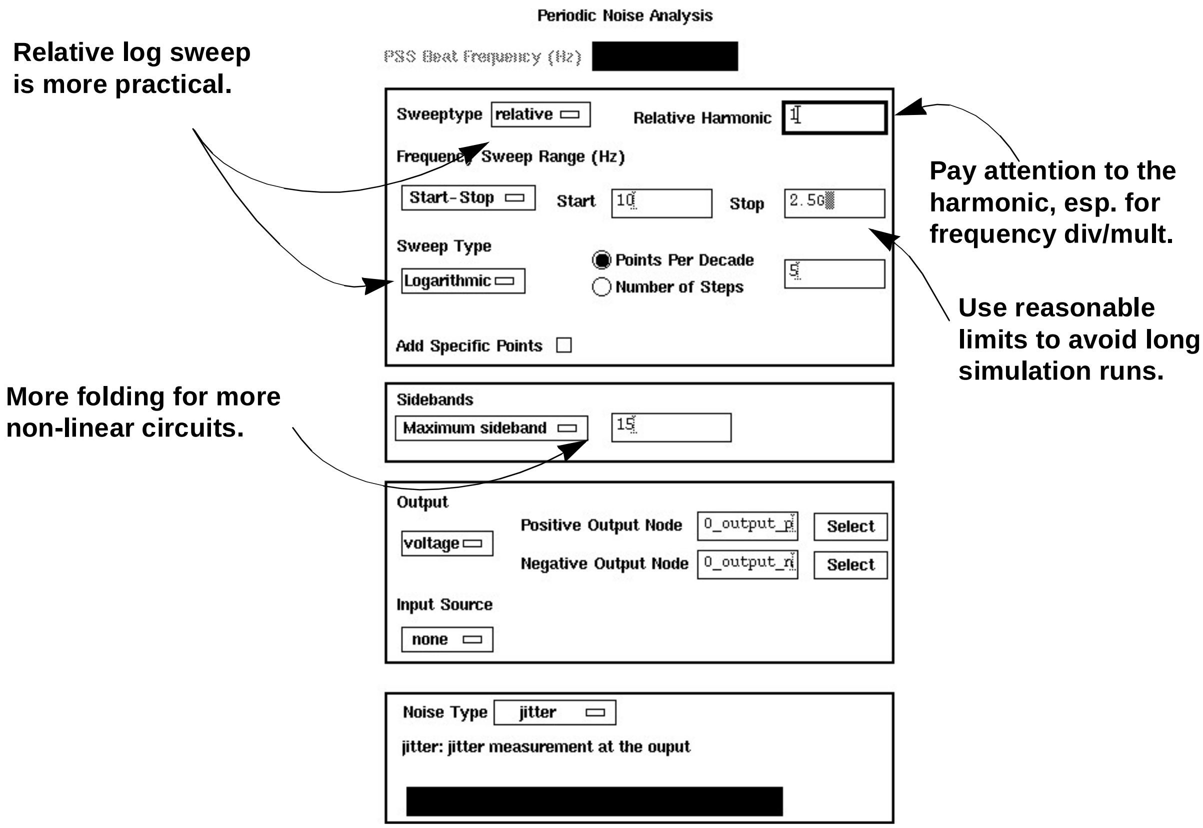
To measure the jitter for the autonomous circuit above we will setup PSS and PNoise analyses. The PSS analysis is a regular autonomous simulation. First, we use periodic steady state analysis to find the periodic trajectory or the large signal periodic solution of the system. Next, we use periodic noise analysis to find the phase noise and provide all the information we need to extract the jitter.

Only if the M divider is attached as the load, PSS will use the divider output to find the period of the large signal solution. The PSS fundamental will be M times smaller then VCO output frequency.

The PNoise analysis form is used to setup the Jitter measurements. The output is the output of the VCO.

A relative log sweep is recommended for optimal computation. The sweep limits should include the lowest offset frequency of the interest, but one has to realize that at the frequencies close to the LO the small signal approximation will break down and the information at those frequencies is not valid. That depends on the Q of the oscillator and on the presence of the flicker noise. Another factor could be the bandwidth of the circuit itself. The highest limit is usually on the order of f (LO)/2 to f (LO) since we are only interested in the phase noise around the output harmonic. This will be less accurate for low Q oscillators, when noise “tail” will extend far beyond f_(LO)/2 - part of the noise will not be accounted for in the computations. The bandwidth of the circuit will often be the factor here too. More frequency points are always helpful to increase an accuracy of the jitter computations, you will have to trade off the frequency range, number of point per decade and the maximum sidebands that will be used for folding in PNoise.

Figure 1-7 VCO Jitter Measurement Setup



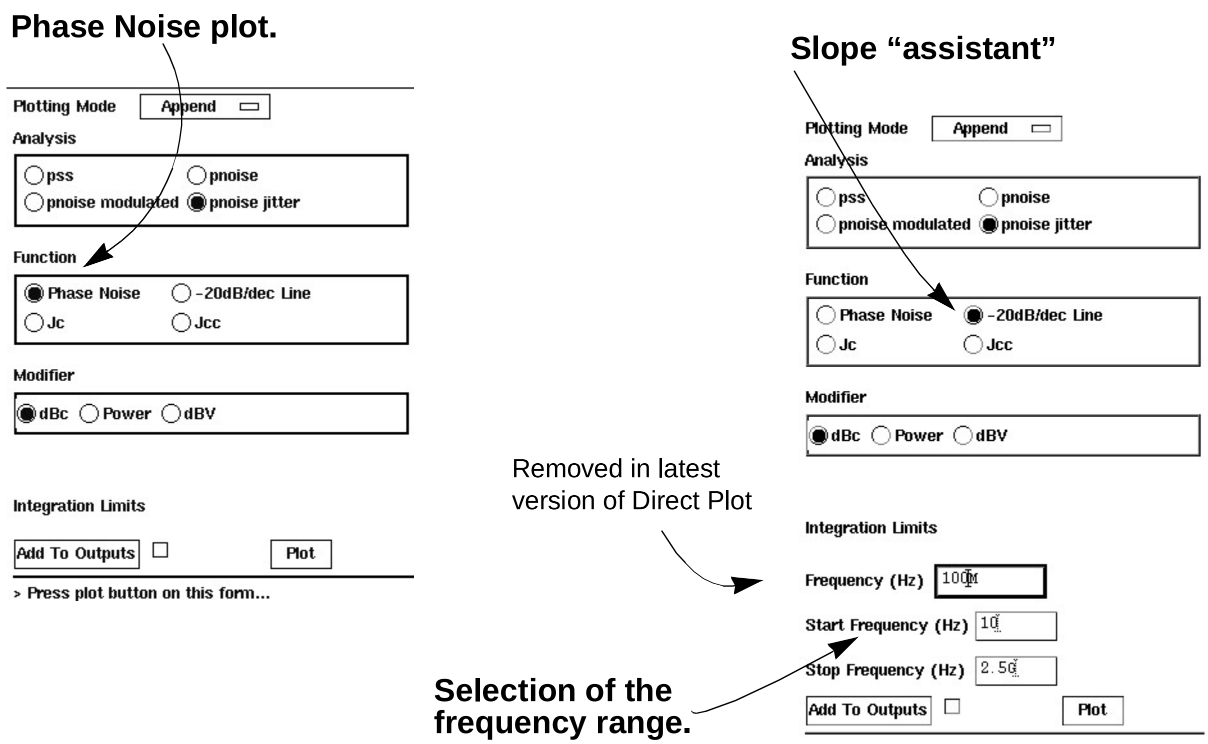
The Pnoise output will be used to compute the jitter. Since we are using the phase noise to calculate the jitter, the netlist will have two pnoise analyses in it. They are required to find upper and lower sidebands and their cross correlation in order to separate the total PNoise results into the PM and AM components, just like in modulated PNoise analysis type.

The Phase Noise Measurements

It is helpful to plot phase noise prior to beginning jitter measurements. It lets one determine the appropriate frequency range, and to decide which integration limits to use. Since at lower frequency offsets, the phase noise will grow boundless, it could be confusing at first how to decide on the lower limit of integration. But more careful examination of the meaning of the low frequency contribution to noise tells us that it can only contribute over the very long observation period and will not affect the jitter over reasonable period of time.

Another helpful feature is the -20dB/dec curve which could be placed on the phase noise plot at any frequency. The slope of -20dB/dec assistant let us to distinguish the regions of the 1/f^3, 1/f^2 and 1/f phase noise PSD. If you prefer, the quick manual calculations using simple white noise jitter approximation could be used after the proper region of 1/f^2 slope is determined [3]. The white noise approximation was used in the first release of the jitter measurements. It required the user to select the point for the approximation of the slope. In the later releases of Direct Plot, the integration is numerical, and the selection of the point is not needed anymore. So, the slope assistant is for informative purposes only now.

Figure 1-8 VCO Phase Noise on Direct Plot Form

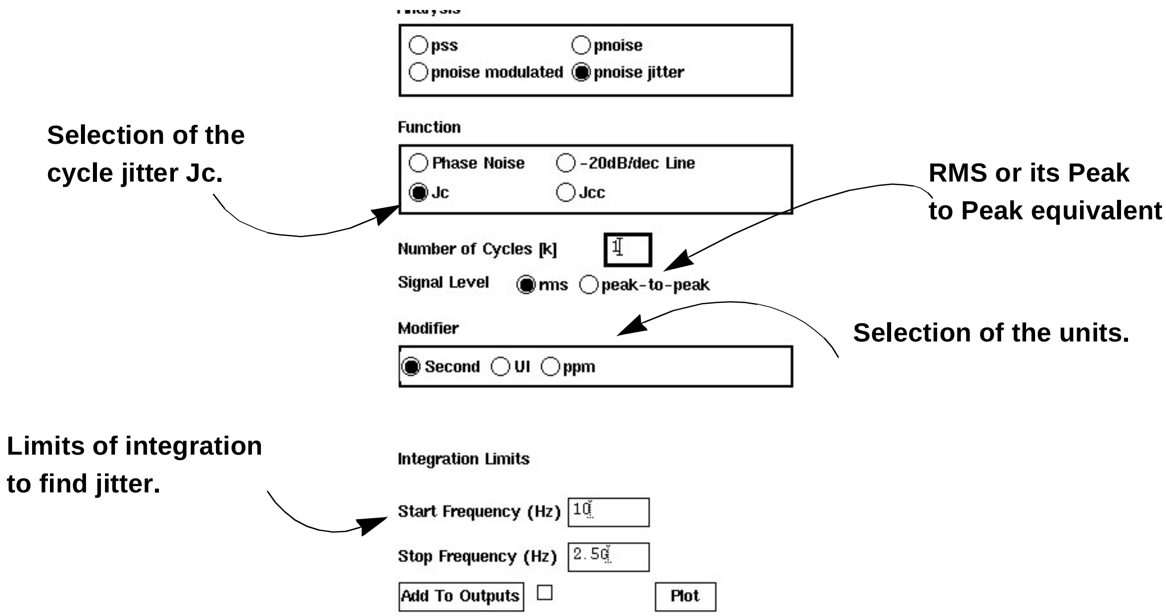


Since jitter is the scalar number, it will be placed together with the phase noise curve on the same plot, as shown in Figure 1-10 on page 20.

The Jitter Measurements

Based on the specs requirements, the phase noise curve, the actual values and the slope, one can decide on the limits of the integration to use. User will also select whether to compute period jitter or cycle to cycle jitter. The default units are in second, which could be changed. Number of cycles determines whether one period or K-periods jitter will be computed.

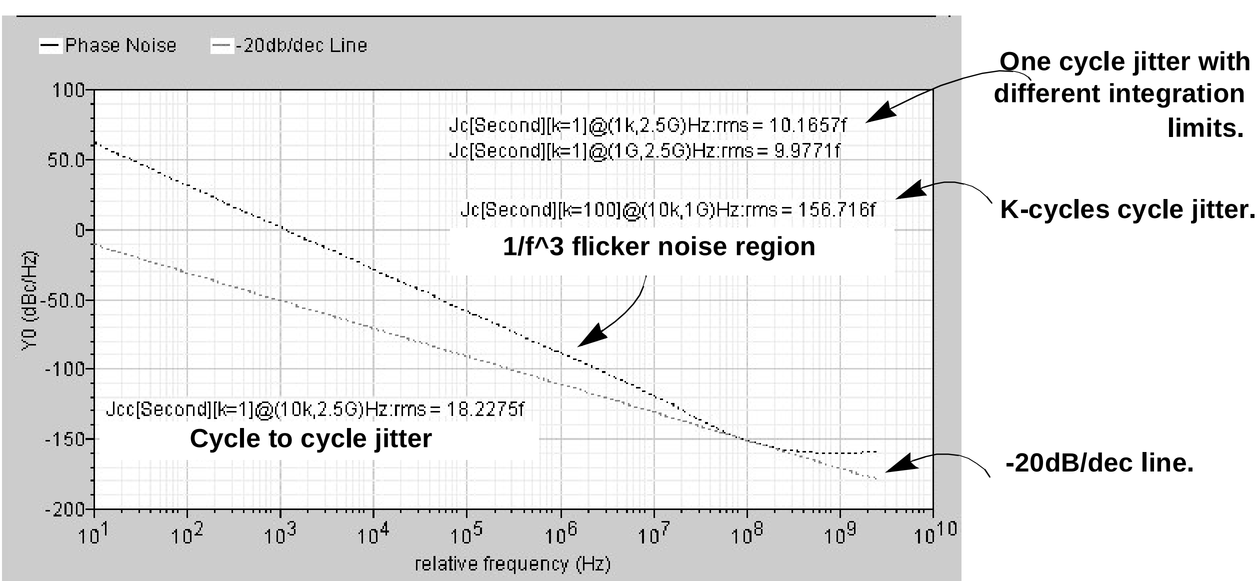
Figure 1-9 VCO Jitter Direct Plot Form



Jitter results confirm the fact the lower frequency have very small contribution to the short term jitter.

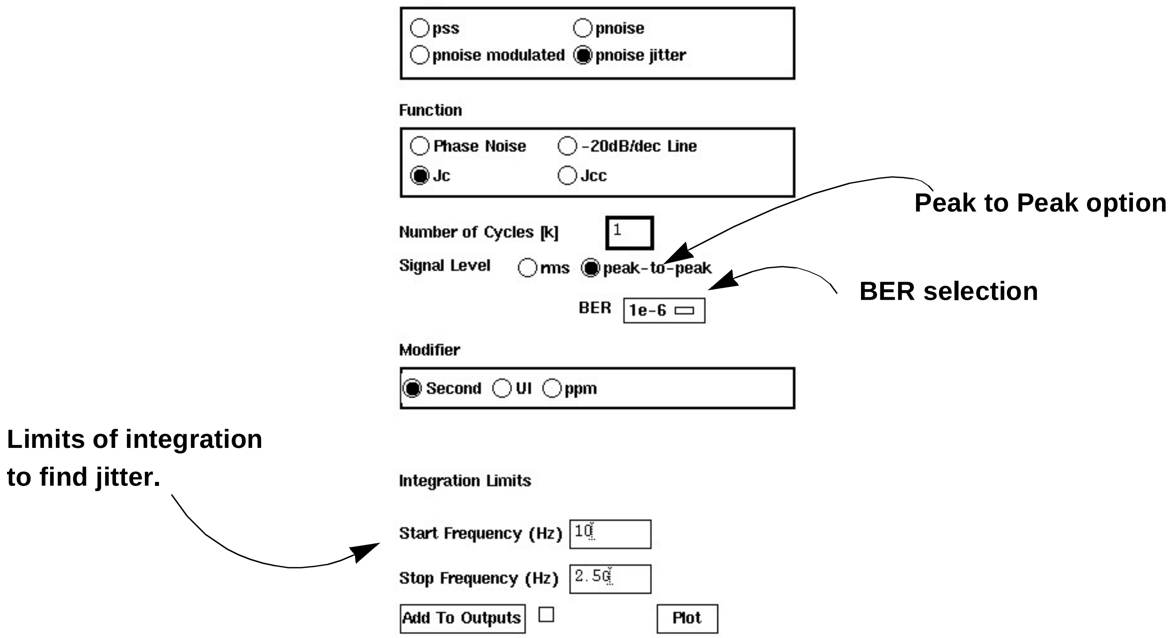
RMS jitter is typically used, but another option is the peak to peak measurement, which could be done for particular Beat Error Rate.

Figure 1-10 VCO Jitter and Phase Noise Results



You could also plot Peak to Peak jitter from the same form. Please note that it is only accurate under the assumptions of the Gaussian random noise distribution. The presence of the flicker noise will make it less accurate.

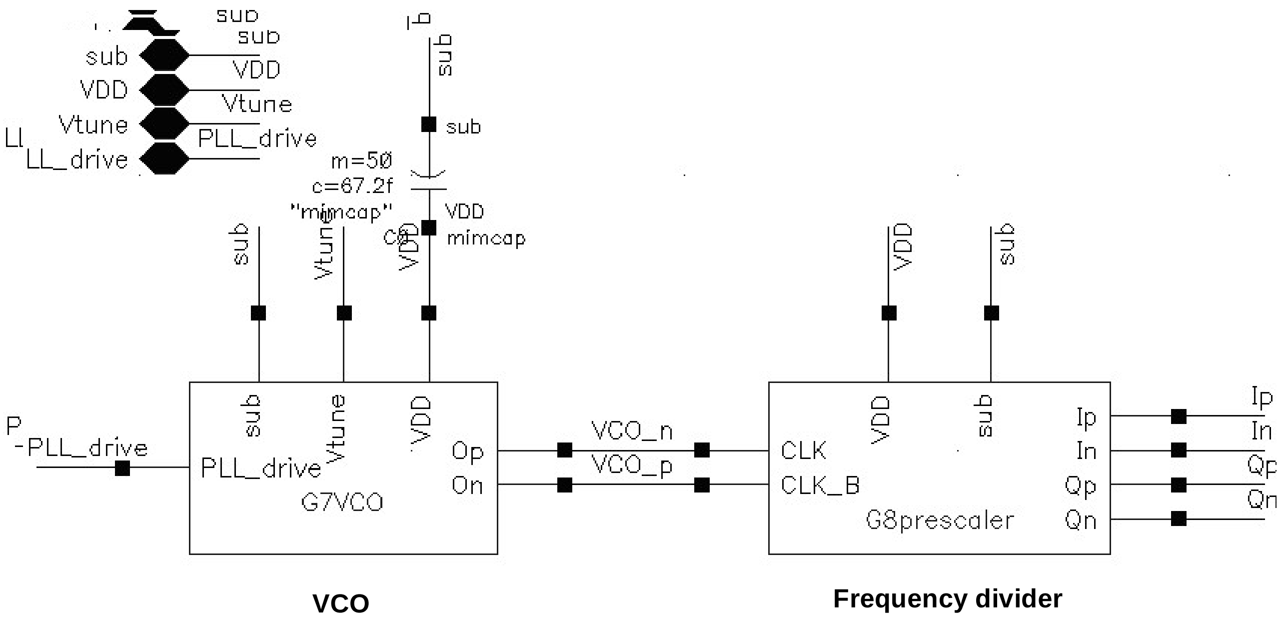
Figure 1-11 VCO Jitter, Peak to Peak Equivalent and the Direct Plot Form



VCO with Divide-By-2 Prescaler

As in the previous example, we have the VCO as a periodic signal generator. But this time it is driving the actual frequency divider, which is a second block in Figure 1-12. This time we are interested in seeing the resulting phase noise and jitter at the output (`Ip` and `In` are the output nodes for the I channel) of the divider.

Figure 1-12 Differential VCO with Frequency Divider/Prescaler



The output signal frequency is half the LO frequency and the offset frequency range of interest is reduced appropriately.

The Testbench for Measurements

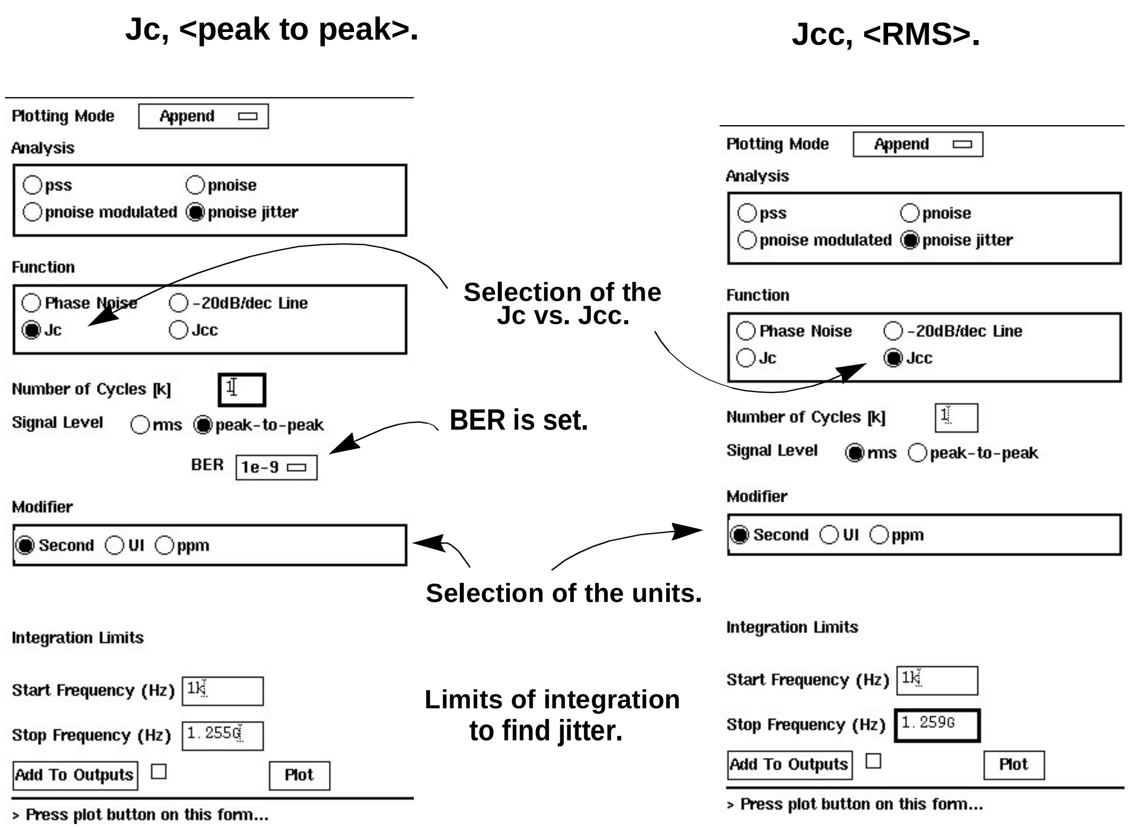
To measure the jitter for the autonomous circuit described in this section we setup PSS and PNoise analyses. The PSS simulation is setup to have fundamental frequency equal to the output frequency of the divider. The divider outputs are used in PNoise as the output nodes and the relative sweep is also referred to the first harmonic which represents the divider output in this analysis. The sweep range is up to half the output signal frequency.

The Phase Noise Mmeasurements

Just as before, we plot the phase noise first. Note that the flicker corner here is sharply lower than for the VCO alone.

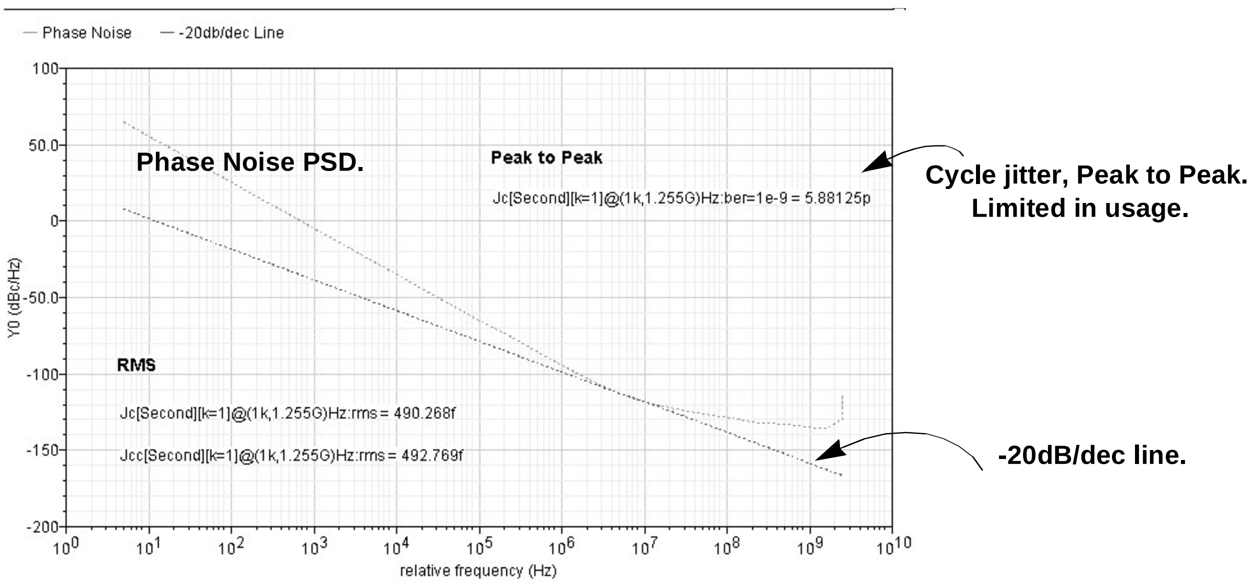
The Jitter Measurements

Figure 1-13 VCO with Divide-by-2 Jitter Direct Plot Form



We plotted J_c and J_{cc} metrics using both RMS and Peak-to-Peak options.

Figure 1-14 VCO with Divide by 2 and Jitter and Phase Noise Results



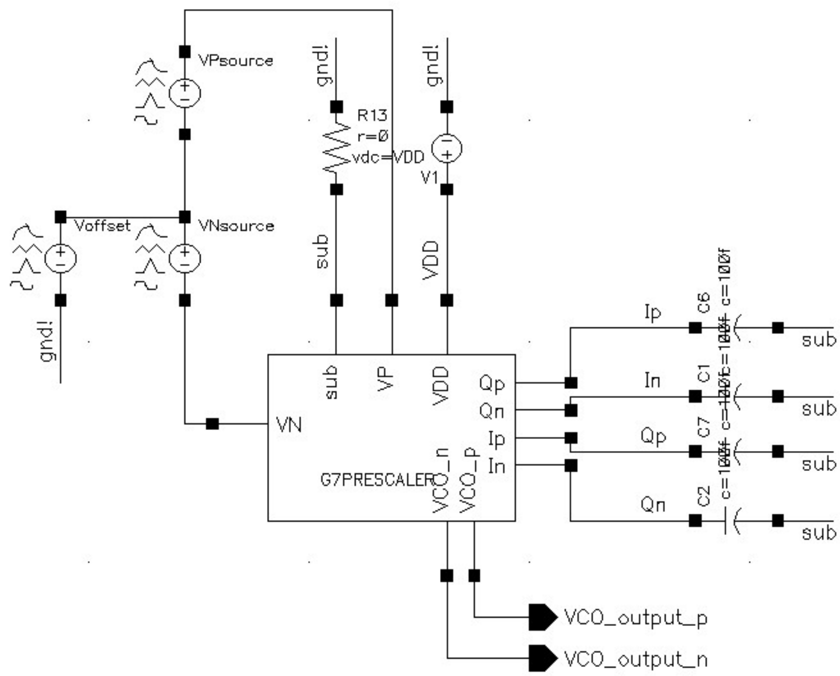
Jitter and Sampled Noise

Frequency Divider as a Driven Circuit

This time we remove the VCO and run the frequency divider alone. The inputs are representative of the VCO output. The output noise of interest is in the 1 kHz to 1.255 GHz range.

The top level Schematic is shown in Figure 1-15.

Figure 1-15 Differential Frequency Divide-by-2



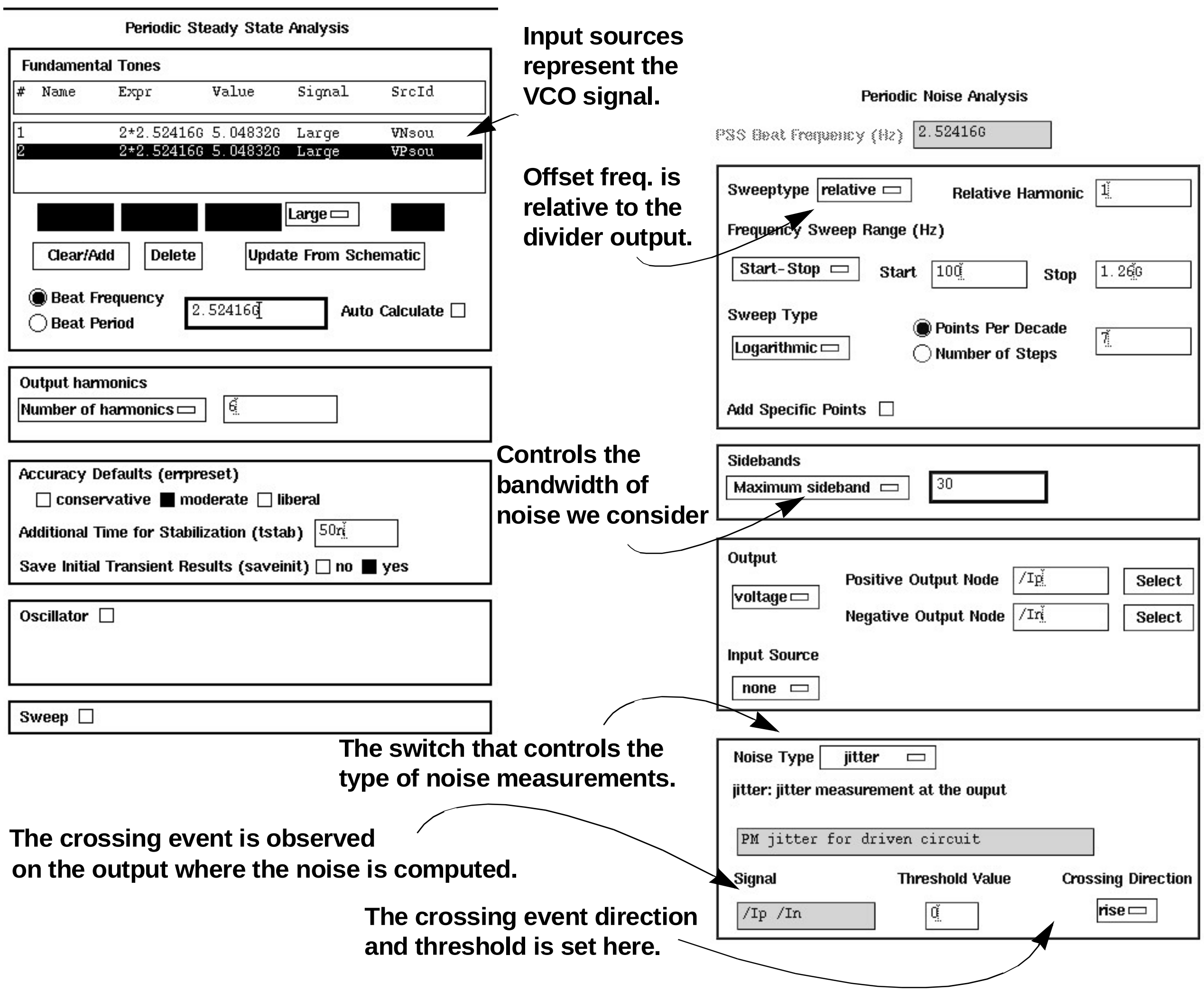
The Measurements Testbench

To measure the jitter for the driven circuit shown in Figure 1-15 we setup PSS and PNoise analyses. The input sources in our example represent the differential input signal coming from the VCO. We use ideal voltage sources which mimic the differential input. You can make the measurement more realistic by matching impedance and using a balun if the testbench is differential.

In the PNoise form, set up the threshold event. It will be internally processed. There is no need to run a PSS analysis and look for the time of the crossing. PNoise will internally pass the request to PSS for information about the crossing. PSS will find and return the timepoints that correspond to the crossing and sampled PNoise will run at those timepoints. PSS will also compute the slew rate at those locations. Periodic sampled noise analysis will find the instantaneous noise and provide the information required to extract jitter.

The output frequency after the division is the fundamental frequency of the PSS simulation for the divider, while the input source is the original, high frequency VCO signal.

Figure 1-16 Frequency Divider PSS and PNOISE Setup



In the PNoise setup, specify a large enough number of sidebands to include in the simulation. The ideal number of sidebands will include the entire bandwidth of the noise sources in the circuit. It can be very large. But a reasonable number is much smaller and can be found after some experience with the circuit. The sampled nature of the analysis will make it possible for even very high frequency noise to affect the final results, even if in a very insignificant way.

If you are not familiar with the noise properties of a circuit, it can take several experiments to find the optimal number of sidebands. Increasing the number of sidebands and observing the results will give you some idea. Sometimes, to speedup the simulation, you can reduce the number and see if the smaller number is sufficient. When the results are not changing significantly after a certain number of sidebands, there is no reason to further increase the number of sidebands.

The Sampled Noise Measurements

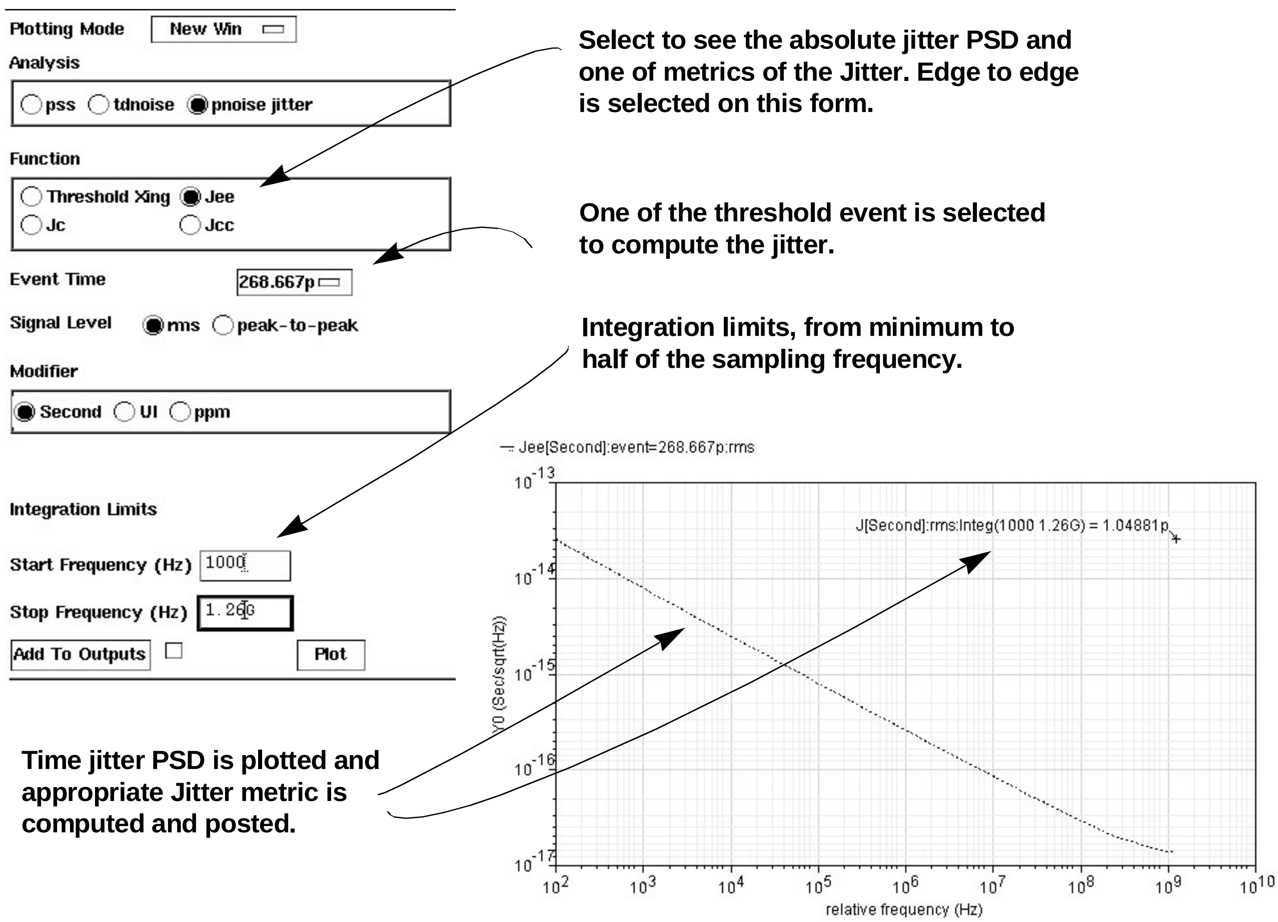
The sampled noise results always coexist with PM Jitter measurements. We will not address sampled noise in detail here, but all the time domain options are available from the appropriate Direct Plot form. Simply note that the time points at which the time domain PNoise analysis runs are the same time points at which Jitter is measured.

The Jitter Measurements

There is a separate Direct Plot form to measure jitter. It lets you select the crossing event that you need in the Event Time. Then you select the metric and the type (RMS vs. Peak to Peak). The integration limits are determined again by the specification used in the system.

The edge-to-edge jitter button will trigger both, the plotting of the absolute jitter PSD and the computation of the J_{ee}. Cycle (period) Jitter and Cycle to Cycle Jitter are also added to the same plot.

Figure 1-17 Frequency Divider for Input Referred Edge-To-Edge Jitter



References

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[2] Ken Kundert, *Introduction to RF Simulation and Its Applications*, IEEE Journal od Solid-State Circuits, vol. 34, pp 1298-2000, Sep. 1999

[3] Ken Kundert, *Predicting the Phase Noise and Jitter of PLL-Based Frequency Synthesizers*, The Designer s Guide, www.designers-guide.com, 2004

[4] W.A. Gardner, *Introduction to Random Processes with Applications to Signals and Systems*, 2nd ed. New York: McGraw Hill, 1989

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[6] Omid Oliaei, *Extraction of Timing Jitter from Phase Noise*, IEEE 2003

[7] IEEE Standards, *IEEE Standard Definitions of Physical Quantities for Fundamental Frequency and Time Metrology - Random Instabilities*, IEEE Std 1139-1999

Cadence application notes are available on SourceLink.