Spectral Analysis of Time-Domain Phase Jitter Measurements

Un-Ku Moon, *Senior Member, IEEE*, Karti Mayaram, *Senior Member, IEEE*, and John T. Stonick, *Member, IEEE*

***Abstract*—A simplified overview of time-domain jitter measurements is presented in this paper. The relationship between the time-domain jitter measurements and the power spectrum of the phase jitter is described using fundamental Fourier properties and basic random variables analysis. This leads to a unifying analysis and the results are in agreement with commonly accepted understanding of jitter accumulation in oscillators. The presented analysis also provides the basis for comparing different jitter measurements.**

***Index Terms*—Absolute jitter, jitter accumulation, jitter measurements, phase jitter, phase locked loop (PLL), phase noise, spectral analysis, time-domain jitter.**

I. INTRODUCTION

THE time-domain phase jitter measurement of an oscillator can be achieved using a number of different approaches. There exist a multitude of publications that refer to the various methods used for the measurement of phase jitter. Some of the common names/methods used to characterize jitter include "period jitter," "edge-to-edge jitter," "cycle-to-cycle jitter," "absolute jitter," "tracking jitter," etc. Even though each of these measurement types are important and well understood by groups of engineers within their applications, the authors find that the differences between the various measurements and how they are related to one another is somewhat confusing and has not been discussed in a simplified and unifying framework.

In this paper, our goal is to summarize the spectral analysis of various time-domain jitter measurements that are used by many oscillator and phase-locked loop (PLL) IC designers. We will establish a simplified way of looking at relationships between various time-domain measurements and their spectral relationship to well-known phase jitter power spectrums of both open-loop (oscillator only) and closed-loop PLLs.

In Section II, we start by reviewing the relationship between time and frequency domain (radians) phase jitter to voltage phase noise (dBc—referenced to carrier). In Section III we present a summary of various jitter measurement terminologies. In Section IV we briefly summarize the classic second-order signal and noise transfer functions of a PLL. We follow this in Section V with the spectral analysis of time-domain measurements obtained by a self-referenced open-loop PLL (oscillator only) measurement. In Section VI, we provide results for a self-referenced closed-loop PLL measurement. An example extending the established method for spectral analysis is described in Section VII. Final remarks and conclusions follow in Section VIII.

II. REVIEW OF PHASE JITTER AND VOLTAGE PHASE NOISE

In general, phase jitter/noise in the time-domain $\phi(t)$ as well as its power spectral density in the frequency domain are discussed in many papers both in the context of open-loop and closed-loop PLLs [1], [2]. Another commonly used measure of phase noise is *voltage* phase noise, commonly measured at the output of an oscillator by using a spectrum analyzer. This measurement is usually expressed in dBc (voltage power in decibel with respect to the power of the carrier/fundamental tone). In this section, we want to review the relationship between time and frequency domain jitter descriptions and voltage phase noise.

We start by considering a clean periodic ($T = 2\pi/\omega_o$) signal $s(t)$. A "jittery" signal (with phase noise) may be defined as

$$s_J(t) = s\left(t + j(t)\right) \tag{1}$$

where $j(t)$ represents the phase noise disturbance of the time variable in *seconds*. From the Fourier series expansion of $s(t)$ and ignoring the static dc phase offset, the signals $s(t)$ and $s_J(t)$ may be defined as

$$s(t) = c_1\cos(\omega_o t) + c_2\cos(2\omega_o t) + c_3\cos(3\omega_o t) + \cdots \tag{2}$$

$$\begin{aligned} s_J(t) =& c_1\cos(\omega_o t + \omega_o j(t)) + c_2\cos(2\omega_o t + 2\omega_o j(t)) + \cdots \\ =& c_1\cos(\omega_o t + \phi(t)) + c_2\cos(2\omega_o t + 2\phi(t)) + \cdots. \end{aligned} \tag{3}$$

In the above representation, $\phi(t) = \omega_o j(t)$ describes the phase noise in *radians*. As an aside we note that the scaling of the phase jitter between the fundamental tone and the harmonics is also clearly seen in $s_J(t)$: The phase jitter of the second harmonic $2\omega_o$ is twice as large ($2\phi(t)$) as the jitter of the fundamental $\omega_o(\phi(t))$. This makes sense since the time shift (noise) in seconds is the same for both the fundamental and the second harmonic and the period of the second harmonic is half that of the fundamental.

In order to see the relationship from $\phi(t)$ (and its power spectrum) to the voltage phase noise (dBc), let us consider the following phase modulation example. Assume a sinusoidal signal $a(t)$ at frequency of ω_o with a single frequency phase jitter $\phi(t) = b\sin(\omega_m t)$

$$\begin{aligned} a(t) =& A\cos(\omega_o t + \phi(t)) = A\cos(\omega_o t + b\sin(\omega_m t)) \\ =& A\cos(\omega_o t)\cos(b\sin(\omega_m t)) \\ & - A\sin(\omega_o t)\sin(b\ \sin(\omega_m t)). \end{aligned} \tag{4}$$

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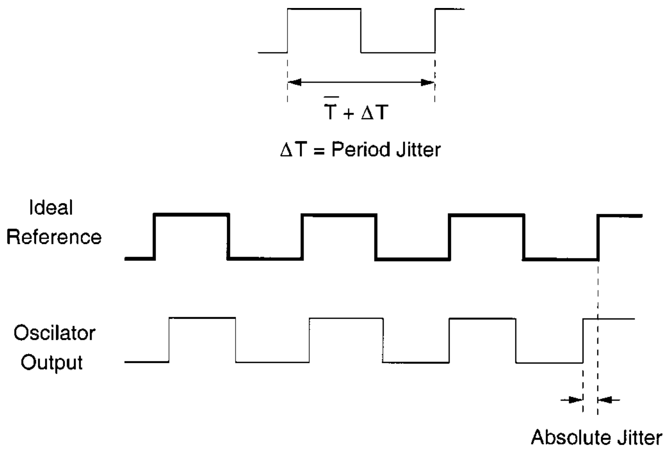


Fig. 1. Definitions of period jitter and absolute jitter.

As b represents the amplitude of phase noise in radians, if $b \ll 2\pi$ (the "narrow-band" FM approximation)

$$\begin{aligned} a(t) \approx& A\cos(\omega_o t) - A\sin(\omega_o t)\, b\sin(\omega_m t) \\ =& A\cos(\omega_o t) - \frac{1}{2}A\,b\cos([\omega_o - \omega_m]\,t) \\ &+ \frac{1}{2}A\,b\cos([\omega_o + \omega_m]\,t). \end{aligned} \tag{5}$$

Given the narrow-band approximation, the power spectrum of the oscillator with a single tone phase noise consists of the carrier and two side tones each offset from the carrier by the phase jitter frequency ω_m. The dBc value of each of the two side tones is $10\log(b^2/4)$. From this, we see the direct relationship between the phase jitter $\phi(t)$ (and the frequency domain spectrum of phase jitter $\Phi(\omega)$ which is an impulse function of weight $b/2$ at $\pm\omega_m$) and the representative voltage phase noise in dBc. This is valid only given the narrow-band approximation. With out the narrow-band assumption, Bessel functions of the first kind describe the spectrum [3]. Exact analysis involving Bessel functions can spread the phase jitter frequency (ω_m) out further from the carrier (see *modulation index* of FM). In the analysis of oscillator/PLL systems and applications, the narrow-band assumption is typically accurate.

In this section, we have shown the relationship between the phase jitter variable $\phi(t)$ and the *voltage* phase noise power spectrum measured in dBc. This was done to show the translational equivalence of the two measures and so that the subsequent discussions dealing with the phase variable $\phi(t)$ can be seen to apply to voltage phase noise in dBc as well.

III. Jitter Measurement Terminology

Let us first briefly define the terminology that we will use in describing various jitter measurements. The two most important and commonly used definitions, period jitter and absolute jitter, are shown in Fig. 1. This illustration is similar to that used in [4]. The term *period* jitter, is the variation measured between one rising edge to the next. The duration measured is the period of the oscillation. Some have appropriately called this measure *edge-to-edge* jitter, *cycle* jitter and sometimes the term *cycle-to-cycle* jitter is used.[1] This period jitter measurement is most commonly quoted in digital applications (e.g., microprocessors) where the high-speed data propagation is critically dependent on proper data latching from one clock edge to the next.

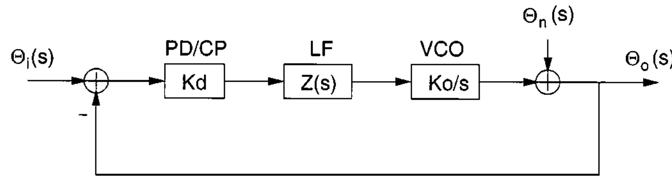


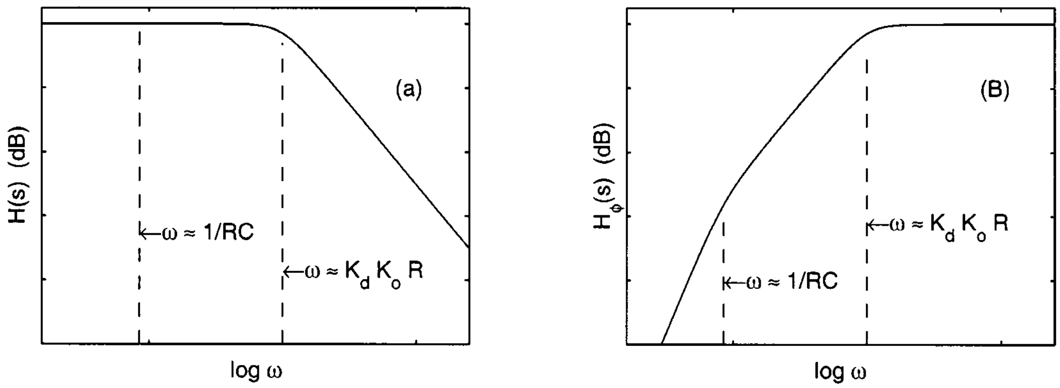
Fig. 2. PLL system. Θ_i is the input phase, Θ_o is the output phase, and Θ_n is the phase noise at the output of the VCO.

Another commonly quoted jitter measurement is the *absolute* jitter. Sometimes the term *tracking* jitter is used instead. However, in the closed-loop PLL, both the absolute jitter and the tracking jitter measurements are referring to the same jitter. In a typical lab environment, the absolute jitter measurement of a closed-loop PLL is obtained by triggering off the input reference frequency (normally an accurate "jitter-free"[2] crystal source) and measuring the PLL output. Given the resources, one can also estimate the absolute jitter by obtaining the PLL output clock edges for a very long period of time and using the average clock period as the "reference." The term tracking jitter is intended to be specific about how this measurement is made, specifically how well the output of the closed-loop PLL tracks the nearly ideal input reference. While the absolute jitter means the same in a closed-loop PLL, the same term can also be used for the open-loop PLL (oscillator only). The open-loop absolute jitter is almost never measured (but can be estimated by measurement) for practical purposes, but the concept is found useful as we elaborate on this in Section V. Because the term tracking jitter is less commonly used and the term absolute jitter is more generally useful, we prefer the term absolute jitter. The absolute jitter measurement (in closed-loop PLL) is quoted for many (if not most) applications including data converters and telecommunication systems. This measure is especially important in digital communication systems where the bit-error-rate of a transmit to receive channel is directly affected by the amount of the transmitter clock jitter and by the accuracy (e.g., jitter) of the receiver's clock and data recovery (CDR) system.

We note here that one of the most commonly quoted papers by Maneatis [5] uses the term cycle-to-cycle jitter and tracking jitter. However, this does not mean that these terms are the norm. For example, Herzel and Razavi [4] use the term cycle jitter and long term jitter, and cycle-to-cycle jitter is used to describe the jitter that represents the difference between two adjacent periods (see period-to-period jitter in Section VII). When we use a time-domain jitter measurement terminology in this paper (Sections V–VII), we make sure to define each measurement by a simple mathematical expression to avoid any potential confusion. If all jitter measurement definitions are accompanied by mathematical expressions, we can now generalize the description of jitter measurements whether we are looking at a

[1]While the term cycle-to-cycle jitter is used to describe period jitter in [5], it is also used to describe the "difference between two consecutive periods" in [4].

[2]The source has jitter that is low enough so that it is effectively nonexistent compared to that which is being measured.



(a) (b)

Fig. 3. PLL s-domain transfer functions. (a) $H(s) = (\Theta_o(s))/\Theta_i(s)$. (b) $H_\theta(s) = (\Theta_o(s))/\Theta_n(s)$.

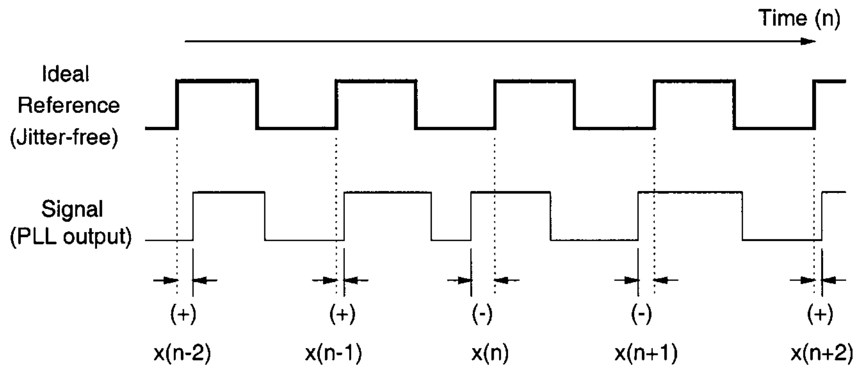


Fig. 4. Absolute jitter in the time domain.

single period variation (edge-to-edge, cycle-to-cycle), variation of multiple periods/edges/cycles, variation of period measurements adjacent to one another (period-to-period jitter in Section VII), or any other possible combination.

IV. Second-Order PLL

In the following, we briefly summarize the s-domain transfer functions of a classic second-order PLL. The presented block diagram and transfer functions are a result of rudimentary math that is familiar to the reader, but we present it here since the discussions in the following sections will refer to these fundamentals.

When the PLL loop filter is an ordinary first-order filter (second-order PLL loop), $Z(s) = 1/sC + R$, the input to output closed-loop transfer function of Fig. 2 becomes

$$H(s) = \frac{\Theta_o(s)}{\Theta_i(s)} = \frac{\left(\frac{K_d K_o}{C}\right)(1 + sRC)}{s^2 + s\left(\frac{K_d K_o}{C}\right)RC + \frac{K_d K_o}{C}} \tag{6}$$

where the effective phase-detector gain is $K_d = I_p d/2\pi$. This example is consistent with a series resistance–capacitance *RC* loop filter used in a typical charge-pump PLL [6] with a charge-pump current I_p (d is a constant for a given system). The effect of a commonly used "ripple bypass capacitor" is omitted in this discussion for simplicity. The resulting input to output transfer function can be compared to the standard two-pole system transfer function

$$H(s) = \frac{2\zeta\omega_n s + \omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \tag{7}$$

where $\omega_n = \sqrt{(K_d K_o)/C}$ and $\zeta = R/2\sqrt{K_d K_o C}$. This transfer function is shown in Fig. 3(a) for a mildly overdamped system ($\zeta > 1$ with two closely separated real poles). This mildly overdamped example is once again chosen for simplicity. In the same PLL system, the transfer function from the output-referred voltage-controlled oscillator (VCO) phase noise (from open-loop) to the output of the PLL (closed-loop), as shown in Fig. 2, yields

$$H_\theta(s) = \frac{\Theta_o(s)}{\Theta_n(s)} = \frac{s^2}{s^2 + 2\zeta\omega_n s + \omega_n^2}. \tag{8}$$

The resulting transfer function is again plotted for closely separated poles (mildly overdamped) in Fig. 3(b). The discussions in the following sections will refer to this classic two-pole PLL system.

V. Self-Referenced Measurement of an Open-Loop PLL

The phase noise properties (jitter characteristics) of an open-loop PLL (VCO) are well understood and the explanation

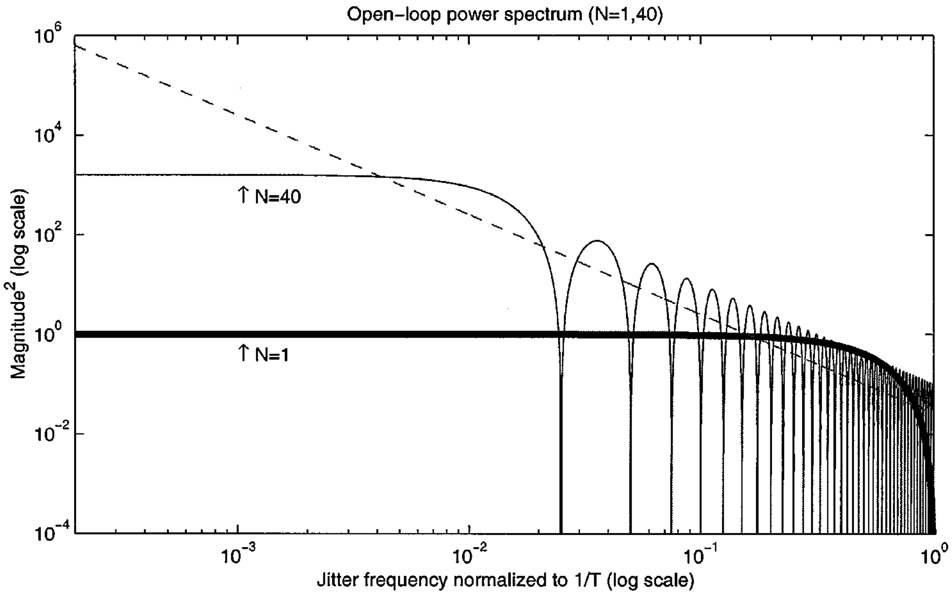


Fig. 5. Open-loop self-referenced jitter measurement.

of the source of phase noise in integrated oscillators has received much attention [1], [7]–[11]. In the following analysis, we will start from an assumed phase noise power spectrum.

Given that an open-loop PLL does not have a fixed phase/clock reference, let us first assume that measurements can be collected over a very large (infinite) time window. Then we can develop a mathematical model such that the absolute, sampled phase jitter is defined by the discrete time-domain function $x(n)$. As graphically shown in Fig. 4, $x(n)$ is the *rising-edge deviation* from the ideal phase/clock reference. This ideal phase reference is the infinite time averaged frequency clock, which in reality can not be measured. The power spectrum of such an open-loop phase noise $|X(\omega)|^2$ is discussed in a number of papers [1], [7], [9], [12], where $|X(\omega)|^2 \propto 1/\omega^2$ for white noise.

The *period* deviation may be defined as $p_1(n) = x(n) - x(n-1)$. This is the self-referenced measurement of one period delay and as mentioned above is commonly referred to as the *period jitter*. The expression for $p_1(n)$ implies that if the phase jitter (or phase deviation) is the same in two consecutive edges, the period deviation for that cycle is zero. The period jitter is here described in terms of the sampled absolute phase jitter $x(n)$.

To obtain the power spectrum of the period jitter $p_1(n)$, or more generally, the jitter of a self-referenced measurement with N-periods delay, define

$$p_N(n) = x(n) - x(n-N). \tag{9}$$

The discrete-time Fourier transform of $p_N(n)$ is given by

$$\begin{aligned} P_N(\Omega) =& X(\Omega) - X(\Omega)e^{-j\Omega N} \\ =& X(\Omega)\left(1 - e^{-j\Omega N}\right) \\ =& X(\Omega)e^{-(j\Omega N)/2}(2j)\sin\left(\frac{\Omega N}{2}\right). \end{aligned} \tag{10}$$

Accordingly, the power spectrum of $P_N(\Omega)$ is

$$|P_N(\Omega)|^2 = |X(\Omega)|^2 \, 4\sin^2\left(\frac{\Omega N}{2}\right). \tag{11}$$

Equation (11) describes the power spectral density of the self-triggered measurement with N cycle/period delays. If $N = 1$ then (11) describes *period jitter*. To compute the time-domain variance, we can integrate the power spectrum [13]

$${\sigma_N}^2 = \int_{-\pi}^{\pi} |X(\Omega)|^2 \, 4\sin^2\left(\frac{\Omega N}{2}\right) d\Omega. \tag{12}$$

Recall that for the open-loop case $|X(\omega)|^2 \propto 1/\omega^2$ and furthermore, to avoid having to express the *sampled* absolute phase noise in the frequency-domain properly (including the aliasing effects), $X(\omega)|_{\text{sampled}} = X(\Omega)$, (12) may be rewritten as

$$\begin{aligned} {\sigma_N}^2 &= \int_{-\infty}^{\infty} |X(\omega)|^2 4\sin^2\left(\frac{\omega N}{2}\right) d\omega \\ &\propto \int_{0}^{\infty} \frac{1}{\omega^2} 4\sin^2\left(\frac{\omega N}{2}\right) d\omega. \end{aligned} \tag{13}$$

This integral is identical to (12) and fully describes the spectral density of the self-referenced jitter measurements that may now be analyzed numerically.

The expression inside the integral of (13) is plotted in Fig. 5 for one period delay ($N = 1$) and for 40 period delays ($N = 40$). In this figure, the dashed line represents the absolute open-loop jitter $|X(\omega)|^2 \propto 1/\omega^2$ and the solid lines represent the expression inside the integral of (13) for $N = 1$ (thick line) and $N = 40$ (thin line).

From (13), we can deduce that the variance (${\sigma_N}^2$) is proportional to the number of period delays N for self-referenced measurements. This can be observed by using the Fourier transform

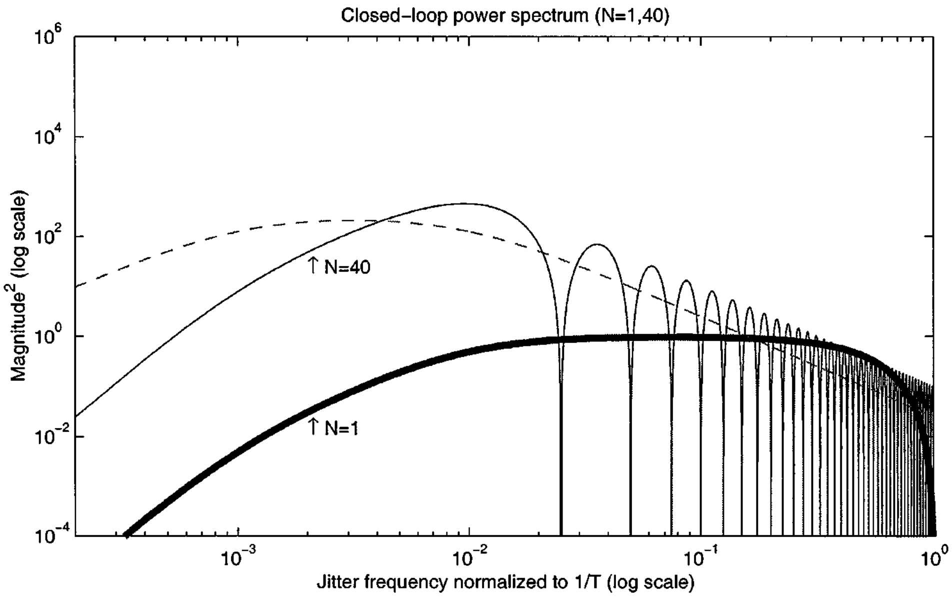


Fig. 6. Closed-loop self-referenced jitter measurement.

relationship between $sinc$ and $rect$ functions and applying Parseval's relation for the equality

$$\sigma_N{}^2 \propto \int N^2 \frac{\sin^2\left(\frac{\omega N}{2}\right)}{\left(\frac{\omega N}{2}\right)^2} d\omega = \int \text{rect}^2\left(\frac{t}{N}\right) dt \propto N. \quad (14)$$

The numerical integration result of (13) is also shown in Fig. 7. The thick solid line is the variance of the self-referenced open-loop phase jitter, increasing linearly with the time delay (N). This is a well-known behavior that is discussed by McNeill in [9] and elsewhere. While an intuitive explanation for this behavior is provided by McNeill, our results confirm this behavior using straightforward mathematics. This result also demonstrates that our analysis provides a sound mathematical support of previously known observations.

VI. Self-Referenced Measurement of Closed-Loop PLL

In this section, we will use an approach similar to that of the previous section to derive the jitter characteristics for a closed loop PLL. The main distinction is that the absolute closed-loop jitter is not proportional to $1/\omega^2$, rather, due to the PLL feedback in operation, the absolute jitter has a finite magnitude over the loop bandwidth of the PLL. The power spectral density of the closed-loop absolute jitter is simply a filtered version of the open-loop absolute jitter ($1/\omega^2$). The filtering transfer function was discussed in Section IV and summarized in Fig. 3. A mildly overdamped PLL with a first-order loop filter ($Z(s) = (1 + sRC)/sC$) was used in this example [6], [14]. For changes in the order of the loop and the loop filter characteristics, one can calculate the closed-loop phase noise power spectrum by evaluating the transfer function to the output of the PLL. The conclusions that are made in the following remain true in either case.

The closed-loop equivalent of (13) for the closed-loop self-referenced measurement with N number of periods delay is

$$\sigma_{cl,N}{}^2 \propto \int_0^\infty |X_{cl}(\omega)|^2 4 \sin^2\left(\frac{\omega N}{2}\right) d\omega. \quad (15)$$

Shown in Fig. 6 is a plot of the power spectrum of the expression inside this integral. The dashed line approximates the absolute closed-loop jitter $|X_{cl}(\omega)|^2$ (a normalized bandwidth that is 1/100 of the oscillation frequency is assumed) of a mildly overdamped second-order PLL system summarized in Section IV and the solid lines represent the expression inside the integral of (15) for $N = 1$ (thick line) and $N = 40$ (thin line). The difference from the open-loop case stems from the fact, as mentioned above, that $|X_{cl}(\omega)|^2$ is not proportional to $1/\omega^2$ for the closed-loop case.

As mentioned in [9], we would expect the variance $\sigma_{cl,N}{}^2$ to approach twice the absolute variance $\sigma_x{}^2$ as N increases. McNeill [9] presents a simple and intuitive explanation for this behavior in the context of establishing open- and closed-loop relationships. Here the spectral analysis of the open-loop and closed-loop PLLs leads to the same conclusion. The added value of the discussions here is the familiar mathematical understanding of spectral reshaping and the integral of the phase jitter measurements. A normalized variance plot from the numerical calculations of (15) is shown in Fig. 7. The thin solid line in the figure shows how this closed-loop self-referenced jitter variance approaches *two times* the closed-loop absolute jitter (dashed line) as N increases.

VII. Extending the Analysis to Period-to-Period Jitter

In this section, we illustrate how to extend the analysis of the previous sections to other jitter measurements. We do this by analyzing what we call *period-to-period* jitter.

We define period-to-period jitter as the *change* in two consecutive single period jitter (self-referenced with $N = 1$) measurements. From this, we can deduce that the period-to-period jitter power spectrum is the period jitter spectrum multiplied by $|(1 - z^{-1})|^2$ (differentiation). We may also reach this result by following the approach we took in the previous sections. Star-

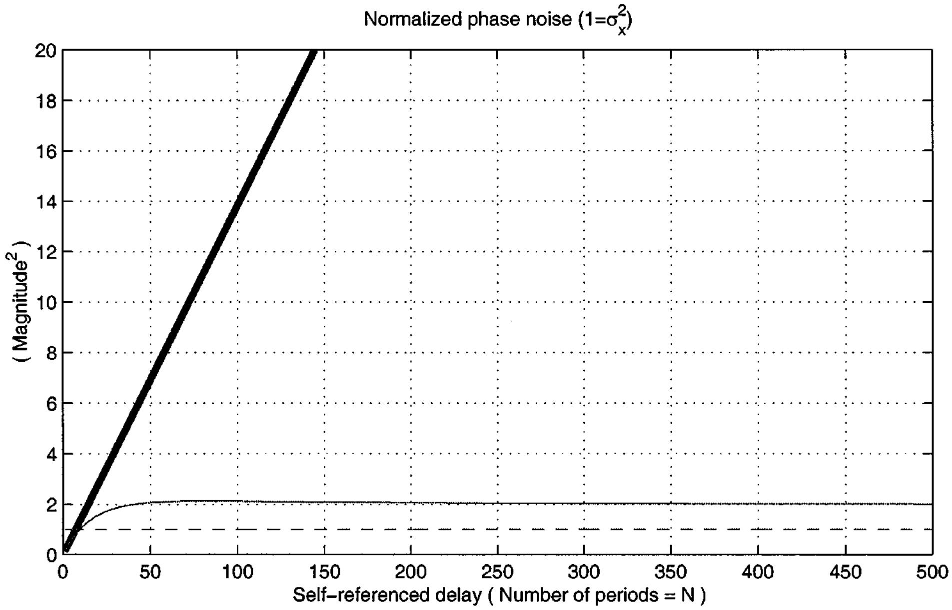


Fig. 7. Variance (integrated power spectrum) versus period delay.

ting with definition of period-to-period jitter

$$\begin{aligned} pp(n) =& [x(n) - x(n-1)] - [x(n-1) - x(n-2)] \\ =& x(n) - 2x(n-1) + x(n-2) \end{aligned} \quad (16)$$

and then following the steps taken in (9)–(13) for the open-loop PLL (oscillator only), it can be shown that

$$\begin{aligned} \sigma_{pp}{}^2 =& \int_{-\pi}^{\pi} |X(\Omega)|^2 16 \sin^4\left(\frac{\Omega}{2}\right) d\Omega \\ & \propto \int_0^{\infty} \frac{1}{\omega^2} 16 \sin^4\left(\frac{\omega}{2}\right) d\omega. \end{aligned} \quad (17)$$

The ratio of the numerical integration of the above expression to the numerical integration of (13) was found to be two, which is anticipated for the open-loop case (oscillator only) as each time (period) delay is uncorrelated to one another [9]. This doubling of the variance from the period jitter to period-to-period jitter can also be true for a closed-loop PLL when the loop bandwidth is low. However, as the loop bandwidth increases, the back-to-back period jitter becomes more correlated and the ratio varies from two. To ascertain by how much, we considered the condition where the loop bandwidth is 1/10 of the VCO output frequency (an extreme condition). In this case, the numerical computation of the closed-loop period-to-period jitter variance yielded approximately a 2.5 increase over the period jitter. Thus, the factor-of-two increase in variance ($\sqrt{2}\,\sigma$) from period jitter to period-to-period jitter seems to be a good approximation for both open-loop and close-loop cases with practical bandwidths.

VIII. Conclusion

We have shown that the various jitter measurements that are often taken for oscillators, operating either in open-loop or the closed-loop, can be explained using familiar Fourier and basic power spectrum properties. The spectral analysis of phase jitter measurements we have presented has provided an in-depth (and hopefully insightful) understanding of phase jitter relationships in the time-domain and frequency-domain. This spectral analysis also fills in the details for understanding the phase jitter for a specific number of clock cycle delays (N) in any self-referenced measurements. It was demonstrated, using the period-to-period jitter example, that the presented methods for understanding phase jitter can be extended to other jitter measurement definitions.

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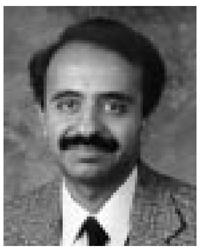

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