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Analysis of Phase Noise and Jitter in Ring Oscillators

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Abstract
Voltage controlled oscillators (VCOs) have gain paramount importance in frequency modulation (FM) and pulse modulation (PM) circuits, phase locked loops (PLLs), function generators, frequency synthesizers etc. which are vital for communication circuits. CMOS based ring oscillators have tuning range, tuning gain and phase noise as the important characteristics. The most difficult task is that variation of phase due to stochastic perturbations. Phase noise has been the designer's primary concerned. The effect of oscillator's noise is one of the most insightful issues in the designing of modern RF telecommunication systems. A low phase noise with minimum power dissipation is rapidly preferred criteria for the design of voltage controlled ring oscillators (VCROs). A very simple and precise analysis of different phase noise models of ring VCOs and their causes is analyzed in this paper. For each case, the flicker noise and the white noise component of phase noise and jitter are considered which limits the signal. The important elements that determine the phase noise in VCOs are the transistor's flicker noise ($1/f$ noise), the output power level, and the quality factor (Q). A synchronized relationship among the effective noise components in the oscillatory circuits leads to good agreement for new design insights and also improves the performance.
Keywords: Ring Oscillators, Voltage Controlled Oscillator, Voltage Controlled Ring Oscillator, Phase noise, jitter.

1. Introduction

Voltage controlled ring oscillators (VCOs) are categorized among the class of oscillators where the frequency of the output oscillation can be varied by the biasing of a controlled voltage signal. They are fundamental electronic modules in the wide sectors of communication. They have variety of applications such as RF transceivers, modulator and demodulator, radio frequency identification devices (RFID) transponders [Rezvan Dastanian, Ebrahim Abiri, Mohammad Reza Salehi, Saeed Ghorbani (2015)] medical domains [Ahmet Tekin, Mehmet R. Yuce, and Wentai Liu (2008)] and clock data recovery circuit [Shruti Suman, K. G. Sharma, P. K. Ghosh (2015)]. Extensive research has been done for modeling of VCOs which is concerned with its instantaneous phase along with the signal amplitude and waveform shape. Consequently, modeling is often done in the phase domain.

Phase noise is the frequency domain representation of rapid short term random fluctuations in the phase of the waveform caused by time domain instabilities (jitter). Phase noise and jitter both indicate the stability of signal and are interrelated. Basically phase noise is the instability of a frequency expressed in frequency domain; while jitter is fluctuation of the signal waveform in the time domain.

The phase noise is a critical parameter which can compare the performance of VCO. The output of a practical oscillator can be written as

$$V_{out}(t) = A\cos(\omega_o t + \phi) \tag{1}$$

Here A is the amplitude of oscillator ω_o is the oscillation frequency and ϕ is the inherent phase. The output of ideal oscillator is shown in Figure1 (a). However, the oscillator is affected by internal and external noise. This noise makes an amplitude and phase fluctuation. The output of actual oscillator can be given by Equation (2) as

$$V_{out}(t) = A(t) f(\omega_o t + \phi) \tag{2}$$

The function f is periodic in 2π then, $A(t)$ and $\phi(t)$ are modeled as the fluctuation of amplitude and phase, respectively. The amplitude can be originally restored by the amplitude mechanism. However, the phase noise can't be brought back to the former state. The phase variation is very important and significantly considered. In order to mathematically understand the phase noises the output of the actual oscillator as shown in Figure 1(b). In the Figure 1, ω_o is the oscillation frequency, $\Delta\omega$ is the offset frequency. To quantify the phase noise, a unit bandwidth is considered at an offset $\Delta\omega$ with respect to ω_o .Thus the equation for describing to the phase noise is given by Equation (3) as

$$L_{total}\{\Delta\omega\} = 10\log\left[\frac{p_{sideband}(\omega_o + \Delta\omega, 1Hz)}{p_{carrier}}\right] \quad (3)$$

Equation (3) describes the phase noise which is noise power in a unit bandwidth divided by carrier power. The value of phase noise can be different according to the offset frequency. For the low phase noise the noise power in the unit bandwidth should be small and the power of the carrier should be large.

The instantaneous frequency of a VCO is generally modeled as a linear relationship with its instantaneous control voltage [Shruti Suman, K. G. Sharma and P. K. Ghosh (2016)]. Figure 2 shows that the output phase of the oscillator is the integral of the instantaneous frequency as expressed in Equation (4) as

$$f(t) = f_O + K_O . v_{ctrl}(t) \quad (4)$$

$$\phi(t) = \int_{-\infty}^{t} f(t) dt \quad (5)$$

Where, $f(t)$ is the instantaneous frequency of the oscillator at time t, f_o is the quiescent frequency of the oscillator, K_o is called the oscillator sensitivity, or gain (Hertz per Volt). $\phi(t)$ is the VCO's output phase, V_{ctrl} is the time-domain control input or tuning voltage of the VCO. For analyzing a control system, the Laplace transforms of the above signals are useful; mathematically can be expressed as Equations (6) and (7) as

$$F(s) = K_O . v_{ctrl}(s) \quad (6)$$

$$\theta(s) = \frac{F(s)}{s} \quad (7)$$

In general CMOS circuits are sensitive to power supply, temperature variations as well as noise generated in IC's. Due to these effects, the propagation delay is variable with respect to its nominal value called as jitter. In an ideal oscillator, the spacing between transitions is constant but practically the transition spacing is variable. This uncertainty is known as clock jitter and increases with measurement interval ΔT (i.e., the time delay between the reference and the observed transitions). This variability accumulation ("jitter accumulations") occurs because any uncertainty in an earlier transition affects all the following transitions, and its effect persists indefinitely. For ring oscillators with identical stages, the variance will be given by $m\sigma_s^{\ 2}$ where m is the number of transitions during ΔT and $\sigma_s^{\ 2}$ is the variance of the uncertainty introduced by one stage during one transition. Noting that m is proportional to ΔT, the standard deviation of the jitter after ΔT seconds is

$$\sigma\Delta T = k\sqrt{\Delta T} \quad (8)$$

Where k is proportionality constant which can be determined by circuit parameters. Another instructive special case that is not usually considered is when the noise sources are totally correlated with one another. In this case, the standard deviations rather than the variances add. Therefore, the standard deviation of the jitter after ΔT seconds is proportional to ΔT as given by Equation (9) as

$$\sigma\Delta T = \varsigma\Delta T \quad (9)$$

where ς is also a proportionality constant. In most digital applications, it is desirable to decrease at the same rate as the period T. In practice, we wish to keep constant the ratio of the timing jitter to the period. Therefore, in many applications, phase jitter, which is is a more useful measure can be defined by Equation (10) as

$$\sigma\Delta\phi = 2\pi\frac{\sigma\Delta T}{T} = \omega_o\sigma\Delta T \quad (10)$$

Phase noise study of an oscillator considers its transfer function as linear with respect to infinitesimally small input noise perturbations to output phase. From oscillator theory, two conditions are required to make a feedback system to oscillate: the open loop gain must be greater than unity; and total phase shift must be 360° at the frequency of oscillation. Oscillator has positive feedback loop at selected frequency which is of keen interest [Vaishali, Shruti Suman, K. G. Sharma, P. K. Ghosh (2014)]. It is seen that frequency of such oscillation can be increased by decreasing the number of stages or by altering device dimensions, which often end up disastrously increasing the power consumption. The odd numbers of stages are well suited for single ended oscillators and will always oscillate whereas differential configurations can have both odd as well as even number stages such that one stage does not invert. Such differential even number of stages of is useful for generating quadrature or multiphase outputs [B. Razavi (2001)].

Thus, proper frequency stability is necessary for operation of the oscillator [Adel S.Sedra and Kenneth C.Smith (2013)]. Frequency Stability is a measure of the degree to which an oscillator maintains the same value of frequency over a given time. Phase noise is the short-term random frequency fluctuations of a signal which is measured in the frequency domain, and is expressed as a ratio of signal power to noise power measured in a 1 Hz bandwidth at a given offset from the desired signal. In other way, it is a measurement of uncertainty in phase of a signal. Both the systemic noise and the random noise contribute to the phase noise of the ring oscillator. The systemic noise such as common-mode supple noise can be avoided by using symmetric architectures, while the influence of the random noise such as thermal noise and flicker noise cannot be easily alleviated. There are various ways of improving phase noise such as increasing the width of the device and minimizing the channel length, which also add the power dissipation and many more.

This paper provides designers with an overview of phase noise and jitter. The section 2 gives relationship between phase noise and jitter. Section 3 focuses on describing different models of phase noise and jitter and their impact on system performance, and identifies techniques to minimize them. Section 4 gives conclusion and future scopes.

2. Relationship between Phase Noise and Jitter

Phase noise and jitter are different ways of quantifying the same phenomenon but in different domains. Phase noise is the frequency domain representation of random fluctuations in the phase of a waveform. A high voltage swing with improved linearity of delay helps minimizing ring oscillator phase noise. Jitter is a random variation of the signal in time domain and essentially describes how far the signal period shows variation from its ideal value. These occur due to are identifiable interference signals; due to crosstalk between adjacent signal traces, EMI radiation sources in signal path, substrate noise and switching. For high speed circuits the prime requirement is of optimization of phase noise and jitter. This noise spreads the signal to adjacent frequencies, resulting in noise sidebands. The phase noise is typically expressed in dBc/Hz and represents the amount of signal power at a given sideband or offset frequency from the ideal carrier frequency [6]. Hence jitter is the time domain instability of the signal, usually expressed in picoseconds. There are three regions shown in Figure 3, the flat region $\left(1/f^{o}\right)$ at large offset frequency is the noise floor. The $\left(1/f^{2}\right)$ region is referred to as the “white frequency” variation region, since it is due to white, or uncorrelated, fluctuations in the period of the oscillator. The behavior in this region is dominated by the thermal noise in the devices of the oscillator circuit. At sufficiently low offset frequencies the flicker noise of devices usually comes into play and the spectrum in this region falls as $\left(1/f^{3}\right)$ [James Wilson (2010)].

The first step in calculating the equivalent RMS jitter is to obtain the integrated phase noise power over the desired range of frequency i.e., the area of the curve, A. The curve is divides into smaller of individual areas (A1, A2, A3, A4), as shown in Figure 4 and each one defined by two data points. Generally the upper frequency range for the integration should be twice the sampling frequency. The integration of each individual area yields individual power ratios. These ratios are then summed and converted back into dBc. Once the integrated phase noise power is known, the RMS phase jitter in radians is given by the equation as shown below in Equation (11) [Ulrich L. Rohde (1983)].

$$\text{RMS Phase Jitter (radians)} = \sqrt{2 10^{A/10}} \tag{11}$$

Division of Equation (11) by $2\pi f_o$ converts the jitter in radians to jitter in seconds given in Equation (12) as

$$\text{RMS Phase Jitter (Seconds)} = \frac{\sqrt{2 . 10^{A/10}}}{2\pi f_o} \tag{12}$$

Although the use of a deep submicron process allows the possibility of higher VCO frequency, it also introduces the problem of a higher 1/f noise. The high $1/f^3$ phase noise is due to the poor $1/f$ device noise in the deep submicron process. Even though the $1/f^3$ phase noise corner can be significantly, lowered by improving waveform symmetry [Adnan Gundel (2007)], the applicability is limited for ring oscillators since it is impossible to get symmetric rising and falling edges. Differential ring oscillators do not have symmetry advantage over single-ended peers since it is the symmetry of the half circuits that matters.

3. Analysis of Phase Noise using Different Models

The advantage of differential VCO compared to single-ended VCO is the superior common-mode noise (i.e. power supply and substrate noise) rejection ability. Therefore, the differential VCO is more suitable for modern mixed-signal IC on-chip environment, in which the digital circuitry will generate a substantial amount of power supply and substrate noise. Since the transition of each stage is triggered by the previous stage, at a single time only one stage in the ring is switching and thus contributing jitter [Boris Drakhlis (2001)]. Therefore the jitter analysis for the CMOS inverter is the best way to characterize the capacity of jitter optimization for different semiconductor processes.

The phase noise of the differential ring oscillator $L(\Delta\omega)$ has an inverse relationship with the control current and the voltage swing [A. Hajimiri and T. H. Lee (1998)] as

$$L(\Delta\omega) = \frac{KT}{I_{cntrl}\Delta V} \cdot \frac{f_o^2}{\Delta f^2} \tag{13}$$

The total average noise power P_n in a particular frequency band can be found by integrating the power spectral density (PSD) as given in Equation (14) as

$$P_n = \int_{f_1}^{f_2} PSD(f).df \tag{14}$$

The average noise power P_n for any random process can be expressed as

$$P_n = \lim_{T \to \infty} \frac{1}{T} \int_{-T/2}^{T/2} v_n^2(t)\,dt \tag{15}$$

Noise voltage V_n can be given as

$$v_n^2(t) = 4KTR \tag{16}$$

Also, switching time T for each cycle is defined in Equation (17)

$$T = \frac{C_L V_{DD}}{I} \tag{17}$$

Here, C_L is the load capacitance at the output of each node, V_{DD} is the power supply voltage and I is the current in the device during operation. Hence, fundamental noise analysis done for ring oscillator yields average noise power (P_n) as the function of time constant, delay at each stage and oscillation frequency which can be given by equation (18)

$$P_n = \frac{\Delta T}{T} \frac{4KTR}{1 + (2\pi F_0 RC)^2} \tag{18}$$

where T is the oscillation period, Δt is on-time of the transistor per delay cell, F_0 is the offset frequency of the carrier signal, RC is the time constant of the delay cell, and K is the Boltzmann's constant having value 1.38 x 10^{-23} J/K.

3.1 Lesson's Model

An early one-port model for the phase noise spectrum of oscillators was formulated by D.B Lesson's [J. McNeill (1997)]. Variations of this model is presently taken for the modeling of phase noise in oscillators reveals relationship between shot noise, power and quality factor. The phase noise can be closely related by equation (19) as

(19)

$$L(\Delta w) = 10\log\left[\frac{2KTF}{p}\left(1+\left(\frac{\omega_O}{2Q\Delta\omega}\right)^2\right)\left(1+\frac{\Delta\omega_{1/f3}}{|\Delta\omega|}\right)\right]$$

(20)

$$S(\Delta\omega) = \frac{\alpha}{\Delta\omega} + \frac{2KTF}{P}$$

where K is the Boltzmann's constant having value 1.38 x 10^{-23} J/K, P is the average signal power dissipated, T is the absolute temperature, F is fitting parameter for figure of noise, $\Delta\omega_{1/f3}$ is corner angular frequency between $1/f_3$ and $1/f_2$ components of flicker noise and Q is quality function of the oscillator, $\Delta\omega$ is frequency offset from the carrier (Hz) and ω_0 = carrier frequency (Hz) which is expressed as given by Equation (21) as

$$Q = \frac{f_o}{2\Delta f_{-3db}} \tag{21}$$

Here in equation (21), this input phase noise spectrum is expected to have two regions. One region is due to the additive white noise, 2FKT/Ps at frequencies around the oscillator frequency. The second region is due to $\alpha/\Delta\omega$ introduced by parameter variation at low frequencies. This includes both white noise and flicker noise which has a power spectral density inversely proportional to frequency. When Q increases due to decrease in -3db bandwidth results in sharpening of the peak in magnitude response. The analysis of noise in the system can be done by keeping in mind evaluation of noise added at the input, noise current generated due to the sources and thermal actions.

3.2 Razavi's Model

In Razavi's model for noise analysis an oscillator is considered as two port feedback system. These oscillators give output phase noise $L_T\{\Delta w\}$ of the ring oscillator in frequency domain with function of average power P_{avg} as given by following equations (22) and (23) as

(22)

$$\left|\frac{Y}{X}[j(\omega_o+\Delta\omega)]\right|^2 = \frac{1}{(\Delta\omega)^2\left[\left(\frac{dA}{d\omega}\right)^2+\left(\frac{d\Phi}{d\omega}\right)^2\right]}$$

In an oscillator with large Q, the required instantaneous change in frequency for given phase shift is smaller, thus giving better frequency stability.
This model adopts LTI approach to model differential CMOS ring oscillators as given in Equation (23) as

$$L(\Delta\omega) = 10.\log\left[\frac{16}{3}\frac{KTR}{V_{swing}^2}\left(\frac{\omega_O}{\Delta\omega}\right)^2\right] \tag{23}$$

It can be said that the input and the output nodes of any of the delay stages never reach the balanced state together. Therefore, inverter based ring oscillator never act as a linear amplifier during the transitions. Hence its phase noise cannot be analyzed with linearity assumption.

3.3 Hajimiri Model

A more accurate linear time invariant model is developed by Hajimiri and Lee [A. Hajimiri and T. H. Lee (1998), Donhee Ham, Ali Hajimiri (2003)]; which introduces Impulse sensitivity function to undertake effects of cyclostationarity virtue of noise. Phase change is proportional to change in voltage hence can be written as in Equation (24) as

$$\Delta\varphi = \Gamma(\omega_o \tau)\frac{\Delta V}{V_{\max}} = \Gamma(\omega_o \tau)\frac{\Delta q}{q_{\max}} \tag{24}$$

where $\Gamma(\omega_o \tau)$ is the time-varying "proportionality constant" called as impulse sensitivity function. Phase noise depends on the time when the noise current is injected. Oscillators have different noise sensitivity at different time instants over the period. It accounts for cyclostationarity through modulated ISF which can be given by Equation (25) as

$$\varphi(t) = \int_{-\infty}^{\infty} h_\phi(t,\tau) i(\tau) d\tau \tag{25}$$

A noise current component $i(t)$, whose frequency is near an integer multiple (n) of the oscillation frequency has the form of equation (26) which further yields simplified equation (27) as follows

$$i(t) = I_n \cos[(n\omega_o + \Delta\omega)t] \tag{26}$$

$$L(\Delta w) = 10 \log\left(\frac{\frac{\overline{i_n^2}}{\Delta f}\Gamma_{rms}^2}{2Q_{\max}(\Delta\omega)^2}\right) \tag{27}$$

Here, Γ_{rms}^2 is the root mean square (RMS) value of the ISF, where $\frac{\overline{i_n^2}}{\Delta f}$ is the total square noise spectral density per hertz for long channel CMOS transistors, which can be expressed mathematically as in Equation (28) [J. McNeill (1997)] .

$$\frac{\overline{i_n^2}}{\Delta f} = \frac{4KT}{L^2}\mu_{eff} Q_{inv} \tag{28}$$

$$Q_{inv} = C_{ox} W L V_{eff} \gamma \tag{29}$$

As the linearized bulk charge, surface potential at source and surface potential at the drain plays vital significant therefore, drain source conductance comes into design given by Equation (30) as

$$g_d = \mu_{eff} C_{ox} \frac{W}{L} V_{eff} \tag{30}$$

The equation (28) can be modified as

$$\frac{\overline{i_n^2}}{\Delta f} = 4KT\gamma g_d \frac{A^2}{H_z} \tag{31}$$

Here, with γ = 2/3 for MOSFETs in the saturation region and γ varies from 2/3 to 1 as the drain-to source voltage V_{DS} varies from zero to the onset of the saturation. For long-channel MOSFETs in the saturation region, the value can be expressed as given in equation (32) as

$$\frac{\overline{i_n^2}}{\Delta f} = \frac{8}{3} KT \mu_{eff} C_{ox} \frac{W}{L} V_{eff} \tag{32}$$

With the advancement, waveform symmetry is of great importance for minimization of phase noise strongly dependent on the oscillation amplitude. For short channel noise, the linearity criteria for resistance and channel length does not satisfy due to degraded charge carrier mobility; also the γ (excess noise factor due to hot electron effect) observed to be two or three times larger than that of long-channel MOS transistors in saturation.

These models for non-linearity, is prescribed to obtain time dependent sensitivity in the oscillators.
Therefore, the dependency of the oscillation frequency on the channel length L is given by following equation (33) as

$$\frac{i_n^2}{\Delta f}\alpha\begin{cases}\frac{1}{L^3} & L \succ\succ L_c \\ \frac{1}{L^2} \text{ when} & L \prec\prec L_c\end{cases} \tag{33}$$

3.4 Demir's Model

This model is complex, pure mathematical based and involves no circuit intuition. This model is non linear, mathematically involved and CAD oriented. It can be described by one dimensional differential equation. Q-factor definition is not utilized throughout analysis. It develops solid foundation of phase noise regardless of operating mechanism. It requires solving of complex differential equations but more accurate. It's able to predict injection locking behavior of oscillator hence its universal model. Demir's model represents oscillator by a group of equations in the form as expressed by Equations (34) [Ali Hajimiri and Thomas H. Lee (2003)] as

$$\frac{\partial x(t)}{\partial \tau} = f(x(t)) \tag{34}$$

When the oscillator is perturbed by a small perturbation $b(t)$, the output voltage where $v(t)$ is the orbital deviation. Finally the phase noise resulting from the voltage perturbation $b(t)$ can be obtained by solving the following one dimensional differential equation (35) as

$$\frac{\partial \theta}{\partial \tau} = v(t+\theta(t))B(t+\theta(t))(b(t)) \tag{35}$$

3.5 Ham's Model

This model is recent and contributions in the scenario of phase noise and jitter domains which act as a bridge between conventional and existing noise theories. It is simple and intuitive model which is valid for both LTI and LTV modeling of phase noise. In this model, concept of virtual damping rate is taken as a fundamental measure of phase noise [Y. Tsividis (1988), Asad A. Abidi (2006), A. Demir (1998)]. This model is capable to explain up conversion, down conversion of noise in the vicinity of integral multiple of oscillation frequency. Analysis using this not requires long simulation time and does not have computational overhead.
Considering that all oscillators start oscillating at the same time instant, it's found that the ensemble average over time exhibits exponential damping. If the phase diffusion is due to white noise, the variance of which signifies the width of the probability distribution is given by equation (36) as

$$\varphi^2(t) = 2DT \tag{36}$$

Here phase diffusion constant 'D' is the reciprocal of the exponential time constant in damping of ensemble/time average and is also called as virtual damping rate. For LTV analysis time varying effects are taken in to account for evaluating diffusion constant.

Conclusion

With the scaling of technology in the recent years, high density of integration and fast speed are challenges to the design considerations such as power supply, stability and phase noise in circuits. The major focus has been laid on low phase noise and low jitter devices. Important VCO phase noise models such as Lesson's linear model, Razavi's Model and Hajimiri's time-variant model were analyzed. Abidi's model was studied for the phase noise and jitter process in CMOS inverter-based and differential ring oscillators. Demir's and Hem's Models are also discussed. A time-domain jitter calculation method is used to analyze the effects of white noise, while random VCO modulation is used for flicker noise. Analysis showed that in differential ring oscillators, white noise in the differential pairs dominates the jitter and phase noise, whereas the phase noise due to flicker noise arises mainly from the tail current. This analysis aids better understanding of phase and jitter phenomena. Apt minimization of noise distortion in the circuits leads to high performances.

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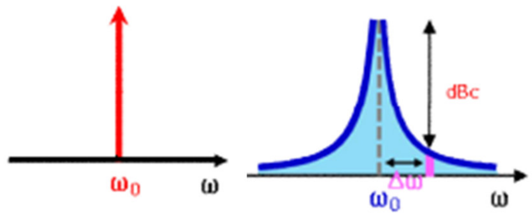

Figure 1 (a) Output spectrum of ideal oscillator (b) Output spectrum of actual oscillator

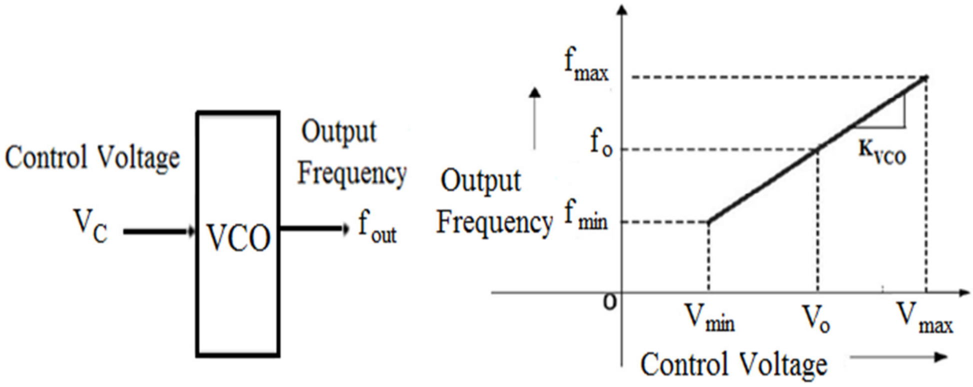

Figure 2 Fundamental model of ring VCO

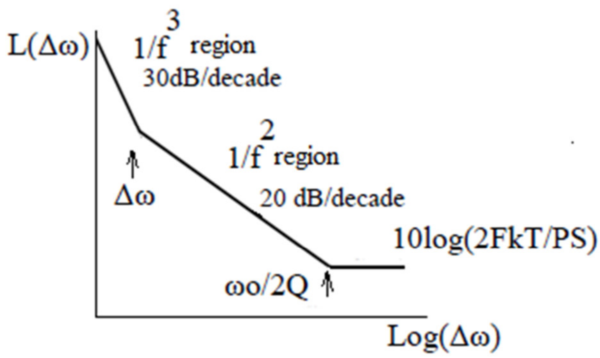

Figure 3 Phase noise representation with flicker noise component for ring VCO

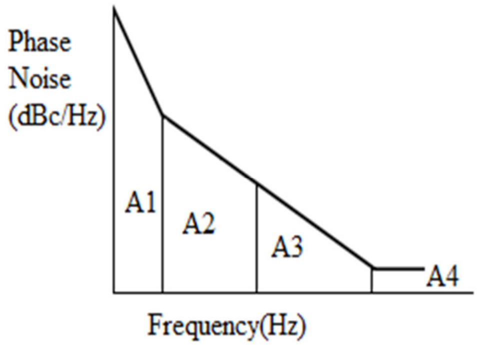

Figure 4 Calculation of jitter from phase noise