THE EFFECT OF MASKED SIGNAL ON THE PERFORMANCE OF GNSS CODE TRACKING SYSTEM

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Abstract

The main purpose of this paper is to describe the code tracking performance of a non-coherent digital delay lock loop (DLL) or coherent DLL while tracking GNSS signal in the presence of signal masking. The masking effect is usually caused by buildings that obscure the signal in either a periodic or random manner. In some cases, ideal masking is used to remove random or periodic interference. Three types of the masked signal are considered - no masking, periodic masking, and random masking of the signal input to the receiver. The mean time to lose lock (MTLL) of the code tracking loop are evaluated, and some numerical result and simulation results are reported. Finally, the steady-state tracking errors on the performance of the tracking loop in interference environment are also presented.

Keywords: delay lock loop, mean time to lose lock, masked signal, GNSS

1. Introduction

The Global Navigation Satellite System (GNSS) provides three dimensional absolute position, velocity, and time information. An important class of predicting the performance of a GNSS system is the ability of the code tracking loop to maintain “lock”. When the signal was masked or blanked, most the GPS positioning method suffer degradation in accuracy and increase the processing time. The masking effect is usually caused by buildings, wall, trees, or random manners that obscure the signal periodically. Certainly make use of ideal masking to overcome and remove interference too. The goal of this paper is to analyze these effects on GPS code tracking system. The code tracking system considered here is a second order delay lock loop (DLL) with an early-late gating of 1/2 chip. The conventional DLL (coherent or non-coherent) has been reported in the literature; see, e.g., [1][2][3]. In this paper, we first discussed the code tracking performance in the presence of noise and signal masking. Three masking function are considered: no-masking, periodic masking, and random masking. Finally, the mean time to lose lock (MTLL) of the code tracking loop for GNSS signal blanking are also presented. The most commonly used performance criteria for the DLL are the tracking jitter and MTLL. The MTLL denotes the mean time that a tracking loop remains synchronized. The calculation of the MTLL is based on Fokker-Planck techniques [2][3][4]. For the coherent second-order DLL, the MTLL that calculation based on the results in [5]. For non-coherent or higher-order loops can be described for example in [6][7].

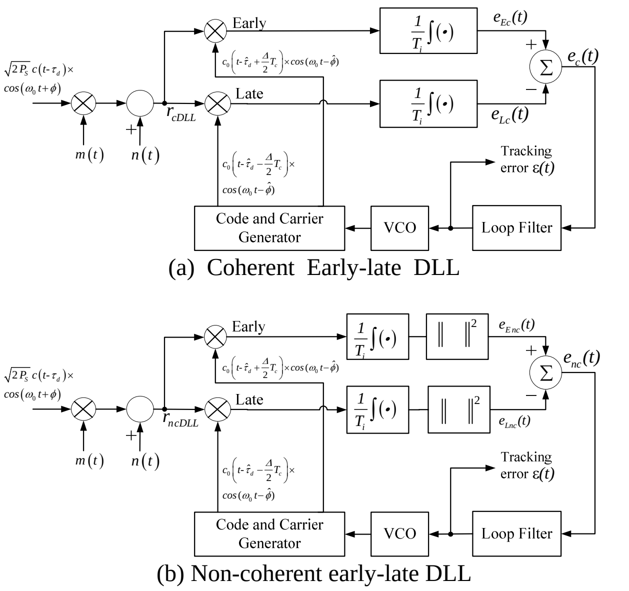
2. Code Tracking Loop Description

2.1 Delay Lock Loop Models

The goal of this paper is to investigate the code tracking performance in the signal masked environment. The purpose of code tracking is to perform and maintain fine synchronization. A common fine synchronization strategy is to depict a code tracking loop which can track the code phase in the presence of a small frequency error. After the correct code phase is acquired by the code tracking loop, a phase lock loop (PLL) can be used the track the carrier frequency and phase. In this paper, we describe two alternate methods, coherent delay lock loop and non-coherent delay lock loop, for GNSS code tracking system. Figure 1(a) demonstrates the model for both the data modulated signal and code phase error measurement. The C/A signal is generated and is controlled by a masking switch that can randomly or periodically “masks” the signal. The generated signal combines the masked signal and white Gaussian noise then fed into the delay lock loop. It was assumed that perfect carrier demodulation was utilized in the simulation. The delay lock loop considered here are first order coherent DLL (c-DLL) and noncoherent DLL (nc-DLL) and the discriminator is referred to as the normalized early minus late with an early-late gating of 1/2 chip. From Figure 1(a), the early-late discriminator output error signal is then passed through the loop filter to the VCO that steering the clock of the local PRN code generator and that is designed not only to reduce the tracking error but also to produce an accurate estimate of the original signal at its output. The model for the nc-DLL is illustrated in Figure 1(b), and is the same as Figure 1(a), except that the each low pass filter of Figure 1(a) with an lowpass filter followed by a square-law envelope detector. The difference between the situations to choose an nc-DLL and a c-DLL depends on if the carrier frequency and phase are known or not. If both of them are known already, the c-DLL can be utilized. When neither of them are known, the nc-DLL must be in use to track the received code. Thus, we assume the incoming signal is RF filtered and down-converter to IF and remove the data modulation in c-DLL and carrier phase recover. Besides, the Doppler perturbation is neglected in this simulation. The received mathematical model in the baseband is given by

$$r_{cDLL}(t)=\sqrt{2P_s}\,c\left(t-\tau_d\right)m^{(\ell)}\left(t\right)cos\left(\omega_0 t+\phi\right)+n(t) \qquad (1)$$

where P_s is the IF signal power, $c(\cdot)$ is the pseudonoise PN signal with chip rate $f_c = 1/T_c$, τ_d is the time delay and varies with time, the signal frequency, $f_0 = \omega_0 / 2\pi$, is assumed to be known, $m^{(\ell)}(t)$ is the masking function that there are three masked model are considered: no masked, periodic masked, and random masked signal of the input to the receiver. The notation (ℓ), $\ell = 1,2,3$ denote different form of masked model.



(a) Coherent Early-late DLL

(b) Non-coherent early-late DLL

Figure 1. Both the baseband c-DLL and nc-DLL, respectively. (a) C-DLL modeling for masked signal (b) Nc-DLL modeling for masked signal

$n(t)$ is the total receiver noise which include the thermal noise, possible co-channel interference, and the lower multipath signal and usually modeled as white Gaussian distribution that is expressed as

$$n(t) = \sqrt{2}\, n_I(t) cos(\omega_0 t) + \sqrt{2}\, n_Q(t) sin(\omega_0 t) \qquad (2)$$

where $n_I(t)$ and $n_Q(t)$ are in-phase and quadrature-in-phase low-pass noise components that are independent, zero mean with power spectral density $N_0/2$. The nc-DLL mathematical model is similar to the c-DLL except that the navigation data $b(t)$ is considered in this model, thus the model is given by

$$r_{ncDLL}(t) = \sqrt{2P_s}\, c(t-\tau_d) b(t) m^{(\ell)}(t) cos(\omega_0 t + \phi) + n(t) \qquad (3)$$

where $b(t)$ is a binary valued unit function (± 1).

2.2 Masking Model

In the following, we introduced the masking model in this simulation. Three kinds of masking models are considered, it is no-masking, periodic masking, and random masking. The defined process $m(t)$ is a collection or ensemble of time waveforms that change from 1 to 0, change from 0 to 1, or remain the same at possible transition time point that are equally spaced a distance T apart along the time axis. The mathematical no-masking function is given by

$$m^{(1)}(t) = 1 \quad \text{for all} \quad t \in [0, T], \qquad (4)$$

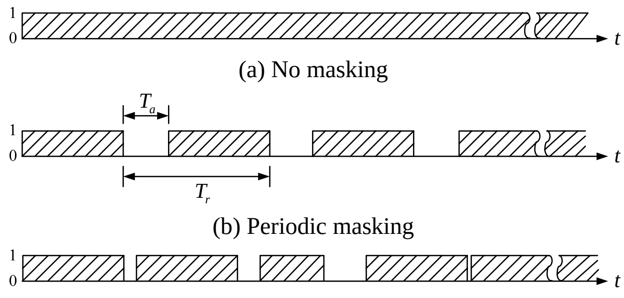
where T is the closed time interval. The periodic masking function are modeled to have a T_a ms duration in a period of T_r ms for a duty cycle of $T_p = (T_a/T_r) \times 100\%$, and is described on each subinterval for i th period of the time axis as

$$m^{(2)}(t) = \begin{cases} 1 & \text{for} \quad iT_r + T_a \le t \le iT_r \\ 0 & \text{for} \quad (i-1)T_r < t < (i-1)T_r + T_a \end{cases}$$
$$\text{for} \quad i = 0,1,2,\cdots,K \qquad (5)$$

where K is the total period number. Finally, the random masking function model, $m(t)$, using 1 and 0 as values for random.

$$m^{(3)}(t) = random(0,1), \quad \text{for all} \quad t \in [0, T] \qquad (6)$$

The defined process $m(t)$ is a normal distribution random process. Figure 2 shows the no-masking, periodic masking, and random masking model.



(a) No masking

(b) Periodic masking

(c) Random masking

Figure 2. Three masking model for (a) no-masking, (b) periodic masking, and (c) random masking

2.3 Masking Signal Terms in the Delay Lock Loop

The code tracking loops under consideration are illustrated in Figure 1. Figure 1(a) is a coherent code tracking loop that forms a early correlator output minus the late correlator output to drive the loop filter. The digital baseband signal pass through the bandpass filter and correlated with early and late correlators can be approximated as follows:

$$\begin{aligned} e_{Ec}(t) &\approx \frac{1}{T_i}\int_0^{T_i} r_{cDLL}(t) c_0\left(t - \hat{\tau}_d + \frac{\Delta}{2}T_c\right) cos(\omega_0 t - \hat{\phi}) dt + \\ &\sqrt{2}\left\{ \frac{1}{T_i}\int_0^{T_i} n_I(t) cos(\omega_0 t) c_0\left(t - \hat{\tau}_d + \frac{\Delta}{2}T_c\right) cos(\omega_0 t - \hat{\phi}) dt + \right. \\ &\left. \frac{1}{T_i}\int_0^{T_i} n_Q(t) sin(\omega_0 t) c_0\left(t - \hat{\tau}_d + \frac{\Delta}{2}T_c\right) cos(\omega_0 t - \hat{\phi}) dt \right\} \end{aligned} \qquad (7)$$

$$
\begin{aligned}
& e_{Lc}(t) \\
& \approx \frac{1}{T_i}\int_0^{T_i} r_{cDLL}(t)c_0\left(t - \hat{\tau}_d - \frac{\Delta}{2}T_c\right)\cos(\omega_0 t-\hat{\phi})dt + \\
& \sqrt{2}\left\{\frac{1}{T_i}\int_0^{T_i} n_I(t)cos(\omega_0 t)c_0\left(t - \hat{\tau}_d - \frac{\Delta}{2}T_c\right)cos(\omega_0 t-\hat{\phi})dt + \right. \\
& \left. \frac{1}{T_i}\int_0^{T_i} n_Q(t)sin(\omega_0 t)c_0\left(t - \hat{\tau}_d - \frac{\Delta}{2}T_c\right)cos(\omega_0 t-\hat{\phi})dt\right\}
\end{aligned} \tag{8}
$$

The output of early correlator, $e_E(t)$, is given by correlation between the incoming baseband signal $r_{cDLL}(t)$ and the locally generated replica $c_0\left(t - \hat{\tau}_d \pm (\Delta/2)T_c\right)cos(\omega_0 t-\hat{\phi})$ in which $c_0(\bullet)$ is the local generated spreading code, $j=\sqrt{-1}$, $\hat{\phi}$ is the estimation phase error of phase ϕ , T_i is the integration time, $\hat{\tau}_d$ is its DLL output estimated values of code delay τ_d. Substituting both (1) and (2) into (7) and (8), the early and late correlator output can be expressed as

$$
e_{Ec}(t) \approx \sqrt{P_s}m^{(\ell)}(t)R_c\left[\left(\frac{\tau_d - \hat{\tau}_d}{T_c}+\frac{\Delta}{2}\right)T_c\right]e^{j(\phi-\hat{\phi})}+N_E(t) \tag{9}
$$

$$
e_{Lc}(t) \approx \sqrt{P_s}m^{(\ell)}(t)R_c\left[\left(\frac{\tau_d - \hat{\tau}_d}{T_c}-\frac{\Delta}{2}\right)T_c\right]e^{j(\phi-\hat{\phi})}+N_L(t) \tag{10}
$$

Where $R_C(\bullet)$ denotes autocorrelation function of the PN sequence, and both the $N_E(t)$ and $N_L(t)$ are total noise of correlation signal. The autocorrelation of this function is defined as

$$
R_C(\tau_d)=\frac{1}{T_i}\int_0^{T_i} c(t)c(t-\tau_d)\,dt = \begin{cases} 1-\dfrac{|\tau_d|}{T_c}, & |\tau_d| \le T_c \\ 0, & |\tau_d| > T_c \end{cases} \tag{11}
$$

where τ is the phase shift of the replica function. When the phase of the replica is the same as the original function, $\tau_d = 0$, the maximum correlation is obtained. Next we assumed that the value of $\phi-\hat{\phi}$ in Eqs. (9) and (10) is $\phi-\hat{\phi}=0$ at baseband. Because neglecting the sum frequency term which will have no effect at baseband. Then the error signal can be described by

$$
\begin{aligned}
e_c &= e_{Ec} - e_{Lc} \\
&= \alpha\sqrt{P_s}\,S_\Delta^c(\varepsilon_c)+\nu_c
\end{aligned} \tag{12}
$$

in which the noise density of ν_c is N_ν (one-sided), α can be regarded the average power reduction factor after masking function effect. Under this assumption, that the inverse masking duration is much greater than the loop bandwidth. The residual power, also call the "modulation self-noise", could be neglected [5] in this loop, and the shape of discriminator characteristic function $S_\Delta^c(\varepsilon_c)$ is given by Eq. (13)

$$
S_\Delta^c(\varepsilon_c) = R_C\left[\left(\varepsilon_c+\frac{\Delta}{2}\right)T_c\right] - R_C\left[\left(\varepsilon_c-\frac{\Delta}{2}\right)T_c\right] \tag{13}
$$

The quantity $\varepsilon_c = (\tau_d - \hat{\tau}_d)/T_c$ is the phase delay error of the loop. The value of $S_\Delta^c(\varepsilon_c) \neq 0$ will lead to a static error in the DLL. Assumed the estimated phase error $|\varepsilon_c| \le \Delta/2$, and an wide front-end bandwidth, and we will find it convenient to define a normalized slope K_d by

$$
K_d^c = \left.\frac{d\,S_\Delta^c(\varepsilon_c)}{d\varepsilon_c}\right|_{\varepsilon_c=0} = 2(2-\Delta), \quad \text{for } |\varepsilon_c| \le \Delta/2 \tag{14}
$$

Then we may express the error signal as

$$
e_c = \alpha\,K_d^c\,\varepsilon_c\sqrt{P_s}+\nu_c \tag{15}
$$

Figure 3 depicts the block diagram of the linearized code tracking loop in which $1/s$ is the transfer function of the voltage control oscillator (VCO), $F(s)$ is the loop filter, and $\alpha K_d^c\sqrt{P_s}$ is the sensitivity of the discriminator.

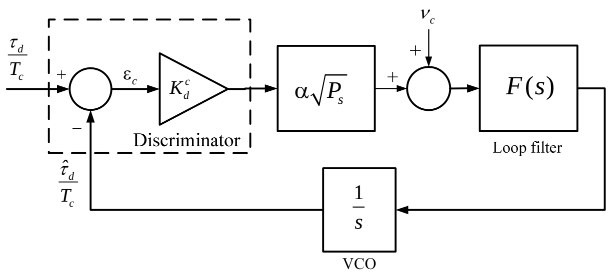


Figure 3. Linearized coherent DLL.

For the first-order DLL, $F(s)=1$, and the closed-loop transfer function $H_c(s)$ is

$$
H_c(s) = \frac{\dfrac{\alpha K_d^c\sqrt{P_s}}{T_c}F(s)\dfrac{1}{s}}{1+\dfrac{\alpha K_d^c\sqrt{P_s}}{T_c}F(s)\dfrac{1}{s}} \tag{16}
$$

In particular, let B_n be the equivalent noise bandwidth, the variance on phase error (jitter) estimation (in the unit of T_c^2) due to noise in the closed-loop operation is [9]

$$
\begin{aligned}
E[e_c] = \sigma_{\varepsilon\_c}^2 &= \frac{1}{\left(\alpha K_d^c\sqrt{P_s}\right)^2}\int_{-\infty}^{\infty}|H_c(f)|^2 S_{\nu_c}(f)df \\
&= \frac{N_\nu B_n}{\left(\alpha K_d^c\right)^2 P_s}
\end{aligned} \tag{17}
$$

where $S_{\nu_c}(f)$ represents the power spectral density of noise ν_c, $E[\bullet]$ denote the ensemble average, and the noise density N_ν of ν_c (after low-pass filtering to smooth the transients) has a value at low frequency of $(K_d^c N_0\Delta/T_c)$. Thus, the variance of the error (in the unit of T_c^2) can be expressed as

$$\sigma^2_{\varepsilon\_c} = \frac{B_n \Delta}{\alpha^2 K_d^c \left(P_s/N_0\right)} \quad \text{for } \sigma_{\varepsilon\_c} \ll 1 \qquad (18)$$
$$= \frac{B_n \Delta}{2\alpha^2 (2-\Delta)\left(P_s/N_0\right)},\ chip^2$$

In the previous discussion, we assumed a coherent signal to baseband and removal of the data demodulation prior to the DLL tracking. In the following, we consider an nc-DLL to code tracking system. Figure 1(b) illustrated the nc-DLL and it operates on a modulated carrier and simply add a square-law envelope detector after the lowpass filter in Figure1 (a). Assume that the quantization error can be neglected, and then the output of error signal driving the loop filter is given by

$$e_{nc} = e^2_{Enc} - e^2_{Lnc} \qquad (19)$$
$$= \alpha' P_s S_\Delta^{nc}\left(\varepsilon_{nc}\right) + \nu_{nc}$$

with $S_\Delta^{nc}\left(\varepsilon_{nc}\right) = R_C^2\left[\left(\varepsilon_{nc} + \frac{\Delta}{2}\right)T_c\right] - R_C^2\left[\left(\varepsilon_{nc} - \frac{\Delta}{2}\right)T_c\right]$ (20)

where e^2_{Enc} and e^2_{Lnc} are the early component and late component at the correlator output, respectively. The quantity $\varepsilon_{nc} = \left(\tau_d - \hat{\tau}_d\right)/T_c$ is the phase delay error of the loop and α' is the power reduction factor after masking function $m(t)$ effect. Owing to the low-pass filtering operation, $m^2(t)$ fluctuates around its mean value. The residual power, also call the “modulation self-noise”, could be neglected [5] in this loop. The discriminator noise ν_{nc} is the white Gaussian noise assumed to be normally distributed with zero mean and variance $\sigma^2_{\nu_{nc}}$. The power spectrum density of ν_{nc} is derived in [1] and [8], from this, the standard deviation can be described as follow:

$$\sigma^2_{\nu_{nc}} = \alpha' P_s N_0 B_n + 2 N_0^2 B_n^2 \qquad (21)$$

where B_n is loop noise bandwidth. P_s/N_0 is the carrier to noise ratio of the satellite. For the linearized discriminator gain K_d^{nc} is given by

$$K_d^{nc} = \left(\left.\frac{d S_\Delta^{nc}(\varepsilon_{nc})}{d\varepsilon_{nc}}\right|_{\varepsilon_{nc}=0}\right) = 2\left(2-\Delta\right), \quad \text{for } \left|\varepsilon_{nc}\right| \le \Delta/2 \qquad (22)$$

For the numerical simulations, the error signal e_{nc} given in (17) is normalized to the input noise power $P_n = N_0 B_n/2$. Thus, the normalized signal power and loop noise are described as follow:

$$P_s' = \frac{P_s}{P_n} = \frac{2\left(P_s/N_0\right)}{B_n} \qquad (23)$$

$$\nu'_{nc} = \frac{\sigma_{\nu_{nc}}}{P_n} = \sqrt{8 + \frac{4\alpha' P_s/N_0}{B_n}} \qquad (24)$$

Therefore, the normalized error signal can be expressed in the form of:

$$e'_{nc} = \alpha' P_s' S_\Delta^{nc}\left(\varepsilon_{nc}\right) + \nu'_{nc} \qquad (25)$$
$$= \alpha'\left(\frac{2\left(P_s/N_0\right)}{B_n}\right)K_d^{nc}\varepsilon_{nc} + \sqrt{8 + \frac{4 P_s/N_0}{B_n}}$$

For the early-late non-coherent DLL, the discriminator can be linearized in its S-curve linear range and the error signal can be expressed as:

$$e'_{nc} = \alpha'\left(\frac{2\left(P_s/N_0\right)}{B_n}\right)K_d^{nc}\,\varepsilon_{nc} + \sqrt{8 + \frac{4\alpha' P_s/N_0}{B_n}} \qquad (26)$$
$$= \alpha'\left(\frac{2\left(P_s/N_0\right)}{B_n}\right)\left(2\left(2-\Delta\right)\varepsilon_{nc}\right) + \sqrt{8 + \frac{4\alpha' P_s/N_0}{B_n}}$$
$$= \frac{4\alpha' P_s/N_0\left(2-\Delta\right)\varepsilon_{nc}}{B_n} + \sqrt{8 + \frac{4\alpha' P_s/N_0}{B_n}}$$

where $\Delta \le 1$ is the spacing between the late and early correlators. The discriminator function, also called the noise-free S-curve, will depend upon the front-end bandwidth. Figure 3 show the S-curves for a wide front-end bandwidth (20MHz) and narrowband front-end bandwidth (2MHz) corresponding to three values of early minus late correlator spacing: $\Delta = 1$ chip, 0.5 chips, and 0.25 chips. For a large front-end bandwidth, the discriminator function is linear and when the error signal is within these bounds, the complete tracking loop is a linear feedback control loop and can be analyzed by basic linear control theory. Finally, for a first order DLL it is known that the closed-loop transfer function $H_{nc}(s)$ is given by

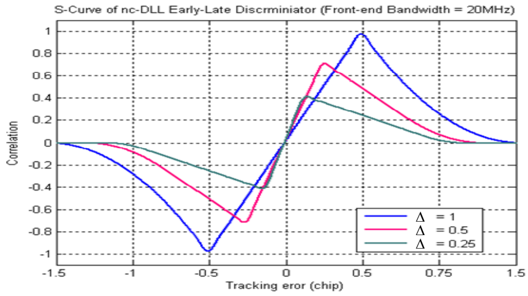
$$H_{nc}(s) = \frac{\frac{\alpha' K_d^{nc} P_s'}{T_c} F\left(s\right)\frac{1}{s}}{1 + \frac{\alpha' K_d^{nc} P_s'}{T_c} F\left(s\right)\frac{1}{s}} \qquad (27)$$

and the variance on phase error (jitter) estimation (in the unit of T_c^2) of an nc-DLL using the early-late power discriminator is as follow:

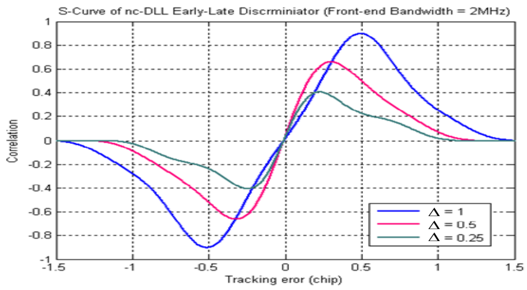
$$E\left[e'_{nc}\right] = \sigma^2_{\varepsilon\_nc} = \frac{1}{\left(\alpha' P_s' K_d^{nc} T_c\right)^2}\int_{-\infty}^{\infty}\left|H_{nc}(f)\right|^2 S_{\nu_{nc}}(f)df \qquad (28)$$

Now approximate $S_{\nu_{nc}}(f)$ with $S_{\nu_{nc}}(0)$, in other words assume that the noise spectral density is essentially flat across the loop bandwidth [9], thus

$$\sigma^2_{\varepsilon\_nc} \cong \frac{2B_n S_{\nu_{nc}}(0)}{\left(\alpha' P_s' K_d^{nc} T_c\right)^2} \qquad (29)$$
$$\cong \frac{B_n \Delta}{2\alpha'\left(P_s/N_0\right)}\left(1 + \frac{2}{\alpha'\left(2-\Delta\right)\left(P_s/N_0\right)T_c}\right),\ chip^2$$



(a)



(b)

Figure 3. S-curve of non-DLL early-late discriminator.

3. Performance Analysis and Simulation Results

In this section we investigate and compare the mean time to lose lock (MTLL) of different loop structures. The MTLL is a figure of a merit for the assessment of performance of GPS tracking loop by evaluating the ability of the tracking loop to maintain locked time. In other words, the MTLL characterizes the mean time that the DLL stays in its tracking range, while the DLL falls out of lock and a re-acquisition process is set going. One model for the MTLL is derived through solving the Fokker-Planck equation. The MTLL Equation solutions in the DLL and other code tracking loops are presented in [10], [11], and [12]. The final MTLL for the first order DLL with one chip correlator spacing is derived in [4] and shown to be:

$$MTLL = \frac{1}{4B_n\sigma_\tau^2}\int_0^{1.5}\int_\varepsilon^{1.5} exp\left[-\Lambda(x)/\sigma_\tau^2\right] exp\left[-\Lambda(y)/\sigma_\tau^2\right] dx\,dy \tag{30}$$

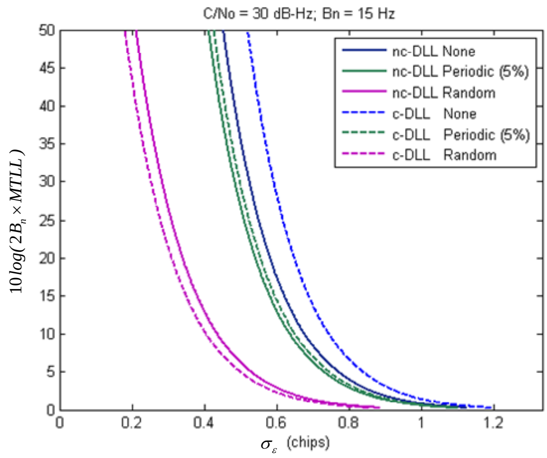
where $\Lambda(x)$ is the integral of the discriminator error function.

$$\Lambda(m) = \int^x S_\Delta(\varepsilon) \tag{31}$$

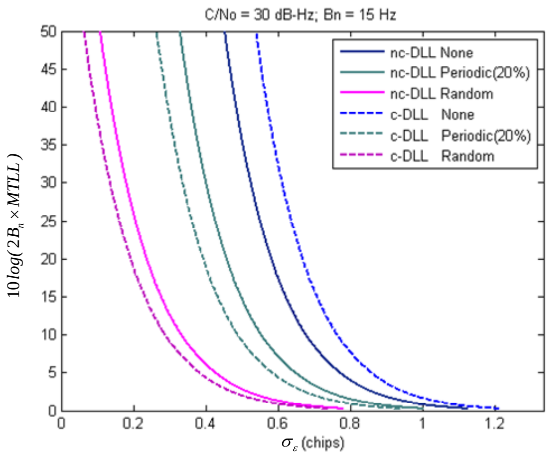
In [4], a general form of MTLL for different early-late correlator separation of 2*N* chip is given by

$$MTLL = \frac{1}{2\eta B_n\sigma_\tau^2}\int_0^{(N+1)/N} exp\left[-\Lambda(x)/\sigma_\tau^2\right]\int_\varepsilon^{(N+1)/N} exp\left[-\Lambda(y)/\sigma_\tau^2\right] dx\,dy \tag{32}$$

Actually, the result in (31) is valid for any closed loop operation, with an early-late discriminator. Figure 4 indicates the delay lock loop MTTL results for the case that C/No = 30 dB-Hz and $B_n = 15$ Hz. The masks are modeled to have 10 ms duration in a period of 200 ms. Thus the signal is masked once every 200 ms for a duty cycle T_p of 5%. As we can see, the random masking nc-DLL dropped lock slightly earlier than the periodic or the no-masking case. The simulation time was set to be 15 seconds simulation time. Thus the tracking results in the nc-DLL simulation model can be plotted in Figure 5. Direct evaluation for the simulated parameters of Eq. (29) with $\alpha' = 0.95$ with $\Delta = 1$ chip correlator separation yields the standard deviation, $\sigma_{\varepsilon\_nc} = 0.513\ chip$ and the simulation yields about $0.518\,chip$. The results show an increase in σ_{ε_nc} of about 2.4%, whereas the simulation shows an increase for periodic masking of about 3.6%, which is entirely close agreement. There are the same results in the c-DLL.



(a) Periodic duty factor = 5%



(b) Periodic duty factor = 20%

Figure 4. Numerical integration of MTLL for DLL. (a) Duty factor = 5%; (b) Duty factor = 20%.

Figure 6 shows the results with the same condition of Figure 5, except the periodic masking duty factor is increased to $T_p = 5,10,15,20,25,30,35,40,$ and 45%, respectively. The phase error for different DLL and masking model is given in Table 1.

Table 1. Phase error for different masking function.

DLL type		C/No = 30 dB-Hz; Bn = 15 Hz; $\Delta = 1$			
		Phase error (chip)			
		None	Periodic		Random
			5%	20%	
c-DLL	Theory	0.5	0.512	0.625	-
	Simu.	0.501	0.513	0.628	0.812
nc-DLL	Theory	0.5	0.513	0.567	-
	Simu.	0.506	0.518	0.578	0.751

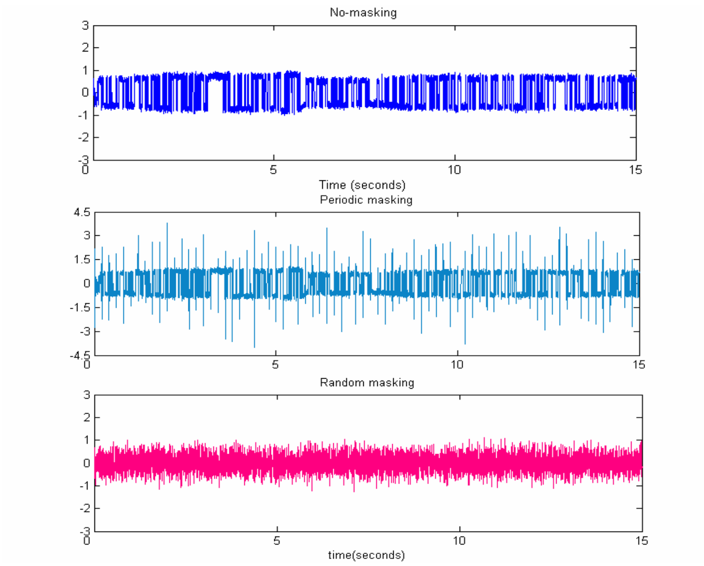


Figure 5. Nc-DLL tracking results with the three masking model at C/No = 30 dB-Hz and 5% duty factor.

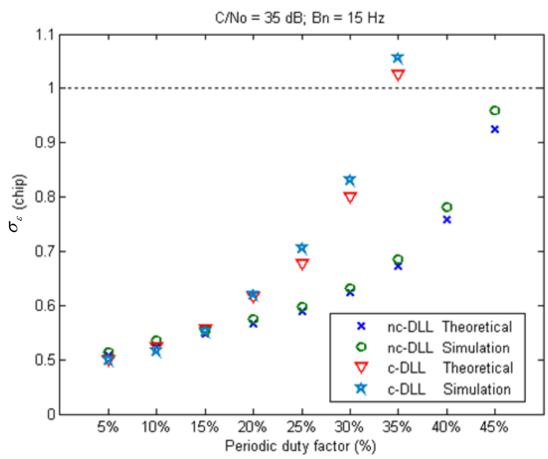


Figure 6. Periodic masking duty factor v.s phase error (chip)

4. Conclusion

In this paper, we have investigated the performance of the coherent and non coherent delay lock loop under various signal masked environment. The masked signal was modeled as periodic, random, and non-masked case. The MTLL and standard deviation of the phase error are used for assessing delay lock loop performance. Numerous masking values were simulated and it supposed that the masking duration is smaller than the inverse DLL loop bandwidth. The results show a little degradation to code tracking performance using the small pseudo random masking. However, the c-DLL had a higher phase error compared to the nc-DLL when the duty factor exceed to 30%. It is the effect of masking in c-DLL is to degrade the receiver power by $-10\,log\left(\alpha^2\right)$, where α is the power reduce factor loss relative the periodic masking duty factor. The tracking performance is also discussed under the effects of the signal masking. The results show that the random masking degrades the DLL performance more severe than the other masking models.

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