Cross-Correlation-Function-Based Multipath Mitigation Method for Sine-BOC Signals

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Abstract. *Global Navigation Satellite Systems (GNSS) positioning accuracy indoor and urban canyons environments are greatly affected by multipath due to distortions in its autocorrelation function. In this paper, a cross-correlation function between the received sine phased Binary Offset Carrier (sine-BOC) modulation signal and the local signal is studied firstly, and a new multipath mitigation method based on cross-correlation function for sine-BOC signal is proposed. This method is implemented to create a cross-correlation function by designing the modulated symbols of the local signal. The theoretical analysis and simulation results indicate that the proposed method exhibits better multipath mitigation performance compared with the traditional Double Delta Correlator (DDC) techniques, especially the medium/long delay multipath signals, and it is also convenient and flexible to implement by using only two correlators, which is the case of low-cost mass-market receivers.*

Keywords

GNSS, multipath, BOC, DDC.

1. Introduction

The modernization of American GPS, Russia GLONASS, European Galileo and Chinese Compass are the new generation of Global Navigation Satellite Systems (GNSS), which will use the novel Binary Offset Carrier (BOC) modulation [1]. This is because that the BOC modulation provides GNSS signals with enhanced robustness against multipath and thermal noise, and increases the precision of range measurement compared with BPSK modulation [2].

Despite this potential of performance enhancement, multipath is still the dominant error source and the limiting factor in GNSS signal tracking. A multipath signal is a delayed version of the incoming signal that enters the receiver front-end and mixes with the direct line-of-sight (LOS) signal [3]. The positioning accuracy in GNSS is seriously degraded in the presence of multipath signals which cause tracking error in the delay lock loop (DLL), it is necessary to eliminate multipath errors in DLL discriminator for an efficient LOS signals tracking.

Several techniques have been proposed over the past decades to solve this problem. Among them, there are two representative ways which are respectively referred to as antenna techniques and signal processing techniques. Among the antenna techniques, the use of special multipath limiting antennas (i.e., choke ring or multi-beam antennas), the post-processing techniques is to reduce carrier multipath, the carrier smoothing to reduce code multipath [4]. Signal processing techniques are another way to solve the multipath errors, which can be identified into two subcategories. The first group relies on an estimation of parameters (delays, amplitudes and phases) of the LOS signal along with all other multipath components. The representative techniques of this group are Multipath Estimating DLL (MEDLL) [5], Modified Rake DLL (MRDLL) [6], and Fast Iterative Maximum Likelihood Algorithm (FIMLA) [7]. They achieve a significant performance improvement against multipath at the expense of a high complexity. Besides, most of these techniques suffer from the fact that they are partially ineffective against medium/short delays multipath. This is a strong limitation since most of the multipath signals tend to be close-in, medium/short delays type in practice [8].

Another group of signal processing techniques based on receiver internal correlation (RIC) technique is the most important approach [3]. The most known correlation-based code tracking algorithm is the traditional early-minus-late (EML), which is composed of one chip spacing between early and late correlator pair. The traditional EML has limited multipath mitigation capability, and therefore, several enhanced EML-based techniques have been introduced, especially to mitigate closely spaced multipath [8]. The Narrow Correlator (NC) is the earliest enhanced EML technique trying to mitigate the multipath error [9]. Another enhanced version of this type of structure is Double Delta Correlator (DDC), such as Strobe Correlator (SC) [10] and High Resolution Correlator (HRC) [11] which uses a higher number of correlators (i.e., 5 complex correlators for HRC) to achieve high accuracy at medium delay multipath in presence of multipath signals. However, a very narrow correlator spacing value and large finite bandwidth filters for these techniques are required to efficiently mitigate multipath errors. Besides, these techniques are sensitive to the short and long-delays multipath signals. Although a modified HRC scheme [12], which can reduce the error for short-delay multipath signals is proposed, it is still sensitive to the long-delay multipath [13].

In order to overcome those limitations of the aforementioned methods, a new multipath mitigation method based on cross-correlation function (CCF) is proposed in this paper. This method is convenient and flexible to implement by designing the modulated symbols of the local signal. Theoretical approximate analysis and simulation results obtained with sine-BOC(1,1) and sine-BOC(10,5) signals show that it has better multipath mitigation performances compared with the aforementioned methods. The remainder of this paper is organized as follows. In Section 2, the derivations of CCF are given out. Section 3 proposes a multipath mitigation method for sine-BOC signals, some implementation issues are also given in this section. Section 4 presents the simulation results and discussions for the proposed method. And Section 5 analyzes the code tracking noise evaluation for the proposed method, followed by the conclusions in Section 6.

2. Derivations of Cross-correlation Function

All of the RIC multipath mitigation techniques are based on the shapes of internal correlation between the received signal and the local signal, which is different from the usual local replica. The main difficulty of the RIC technique design is how to select the spreading code chip waveform of local signal. It is desired to define a parameterized local signal model the chip waveform of which has a high degree of freedom and is easy to generate in receivers to provide more opportunities for waveform optimization. In this section, the CCF between the received sine-BOC signal and the local signal based on the concept of step-shape code symbol (SCS) signal is given out, which is the basis of the investigation of a newly developed multipath mitigation technique in Section 3.

2.1 Sine-BOC Modulation Model

A sine-BOC modulated signal is the product of a Non-Return-to-Zero (NRZ) spreading code with a synchronized sine-phased square wave sub-carrier. In the navigation community, a sine phased BOC modulated signal is normally expressed as sine-BOC(m,n), where m is the ratio of the square wave frequency f_s to 1.023 MHz, and n denotes the ratio of the spreading code rate f_c to 1.023 MHz. The ratio $M = 2m/n$ is referred to as modulation order, which is constrained to positive integer [2].

Using the original definition from [14], a sine-BOC modulated signal can be considered a special case of the SCS signal, which uses the kind of SCS waveform. A detailed description of the SCS signal can refer to [14], which is defined as follows

$$s_{\mathrm{SCS}}(t) = \sum_{i=-\infty}^{\infty} c_i p_{\mathrm{SCS}}(t - iT_c) \tag{1}$$

where $\{c_i\}$ represents pseudorandom code symbols (which may be periodic), $p_{SCS}(t)$ is the SCS waveform, and T_c refers to the period of the modulated symbol. The SCS waveform is divided into M segments, each with equal length $T_s = T_c/M$, M is referred to as the order of SCS signal, and in each segment the level remains constant. Then the SCS waveform is given by

$$p_{\mathrm{SCS}}(t) = \sum_{k=0}^{M-1} d_k \varphi_k(t) \tag{2}$$

with

$$\varphi_k(t) = \begin{cases} 1 & t \in [kT_s, (k+1)T_s] \\ 0 & \text{others} \end{cases} \tag{3}$$

where d_k is the projection of $p_{SCS}(t)$ onto the $\varphi_k(t)$ subspace, and could adopt any real value. To meet the energy normalization condition of the SCS chip waveform, the coded symbol $\{d_k\}$ must satisfy

$$\frac{1}{M}\sum_{k=0}^{M-1} {d_k}^2 = 1 . \tag{4}$$

Every SCS waveform $p_{SCS}(t)$ can be determined by the given shape vector $\boldsymbol{d} = [d_0, d_1,\ldots,d_{M-1}]^{\mathrm{T}}$ and spreading sequence rate $f_c = 1/T_c$. A sine-BOC(kn,n) signal is equivalent to a SCS signal whose shape vector is $\boldsymbol{d}_{\mathrm{BOC}} = [1, -1_1,\ldots,1, -1]^{\mathrm{T}}$.

2.2 CCF of SCS Signals

According to [14], [15], consider the CCF between received sine-BOC(kn,n) signal and a SCS signal which has the same chip rate f_c, pseudorandom code symbols $\{c_i\}$, and the order M, while the SCS chip waveform may be different from that of sine-BOC signal. Then, these two SCS signals can be expressed as follows, respectively

$$\begin{cases} s_{\mathrm{BOC}}(t) = \sum\limits_{i=-\infty}^{\infty}\sum\limits_{k=0}^{M-1} c_i (-1)^k \varphi_k(t - iMT_s) \\ s_{\mathrm{SCS}}(t) = \sum\limits_{j=-\infty}^{\infty}\sum\limits_{l=0}^{M-1} c_j d_l \varphi_l(t - jMT_s) \end{cases} . \tag{5}$$

Therefore, since the case of ideal spreading code symbols with $\mathrm{E}[c_i c_j] = \delta_{ij}$, the CCF of these two signals is

$$R_{\mathrm{B/S}}^{\mathrm{CCF}}(\tau) = \begin{cases} \left(\dfrac{\tau}{T_s} - k\right)\times(r_{k+1} - r_k) + r_k \\ , \qquad \tau \in [kT_s, (k+1)T_s] \\ \left(\dfrac{\tau}{T_s} - k + M\right)\times(r_{k-M+1} - r_{k-M}) + r_{k-M} \\ , \qquad \tau \in [(k-M)T_s, (k-M+1)T_s] \\ 0, \qquad \text{others} \end{cases} \tag{6}$$

where

$$r_k = \begin{cases} \frac{1}{M}\sum_{i=0}^{M-1-k}(-1)^i d_{i+k}, & 0 \le k \le M-1 \\ \frac{1}{M}\sum_{i=0}^{M-1+k}(-1)^{i-k} d_i, & 1-M \le k < 0 \\ 0, & |k| \ge M \end{cases}. \quad (7)$$

A schematic diagram of $R_{\mathrm{B/S}}^{\mathrm{CCF}}(\tau)$ is shown in Fig. 1. It is obvious that the CCF of two SCS signals is piecewise linear between kT_s and $(k+1)T_s$ within [$-T_c$, T_c], and $R_{\mathrm{B/S}}^{\mathrm{CCF}}(kT_s) = r_k$, for $k \in [1-M, M-1]$ and $k \in \mathrm{Z}$.

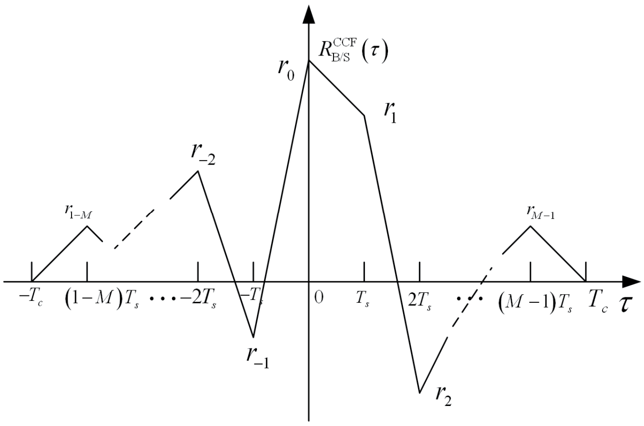

Fig. 1. Schematic diagram of the CCF of $R_{\mathrm{B/S}}^{\mathrm{CCF}}(\tau)$.

From Fig. 1, it can be noted that changing the modulated symbol shape vector $\boldsymbol{d}$ can change the shape of $R_{\mathrm{B/S}}^{\mathrm{CCF}}(\tau)$. This is the theoretical basis of RIC techniques. When considering designing the SCS chip waveform of local signal, since local signal does not relate to amplifying and transmitting in the receiver, it does not need to satisfy the request of constant modulus, and its SCS chip waveform should be easy to generate. Taking the realizing complexity into consideration, it is necessary to construct a more practicable multipath mitigation by using as few correlators as possible. However, to the authors' knowledge, there is no available theoretical formula of the CCF of two signals for multipath mitigation by designing the modulated symbols of local signal. In next section, an analytical expression of that CCF for multipath mitigation is presented.

3. Proposed Method for Multipath Mitigation

3.1 Method Description

From (6) and (7), it can be seen that since the chip waveform shape of the received sine-BOC signal is known, the CCF of two SCS signals entirely depends on the shape of local signal spreading chip waveform. The key of RIC technique is the design of the local signal's modulated symbols, which need to be analyzed by CCF of the signals with various step-shape modulated symbols. Note that the SCS chip waveform of local signal corresponds to an unique point in M-dimensional whose coordinate vector is $\boldsymbol{d} = [d_0, d_1, \ldots, d_{M-1}]^{\mathrm{T}}$, thus, by changing the value of d_k, one can adjust the shape of CCF.

After building the relationship between the shape of CCF and the value of a vector $\boldsymbol{d}$, the good chip waveform should be searched for mitigating multipath. As we know, the sharper of the main peak, the better of multipath mitigation performance (e.g., the multipath mitigation performance of sine-BOC(n,n) signal is better than that of BPSK(n), since the main peak of sine-BOC is shaper than that of BPSK). To ensure a much sharper correlation main peak along with fewer side peaks at larger delays and provide better resolution than the received sine-BOC signal, the CCF must satisfy the three requests:

$$\begin{cases} (1)\text{The main peak should be an ideal triangle} \\ (2)\text{The main peak should be sharper than} \\ \quad \text{that of autocorrelation function (ACF)} \\ (3)\text{The side peaks are as few as possible} \end{cases}. \quad (8)$$

In order to shape the CCF into an ideal triangle, one have to make the modulated of vector $\boldsymbol{d}_{\mathrm{SCS}}$ should be the same with $\boldsymbol{d}_{\mathrm{BOC}}$ at the polarity inversion every two bits, as shown in Fig. 2 (both the solid and dot ellipse line). Since the CCF of two SCS signals is piecewise linear, the modulated symbol $p_{SCS}(t)$ of SCS signal should be divided into $X \times M$ segments to obtain a more shaper main peak compared with the autocorrelation function (ACF), each with equal length $T_s = T_c/(X \times M)$, where the parameter $X \in (0, +\infty]$, and is constrained to positive integer. Thus, the CCF of two SCS signals can be changed by

$$R_{\mathrm{B/S}}^{\mathrm{CCF}}(\tau) = \begin{cases} \left(\frac{XM\tau - kT_c}{T_c}\right) \times (r_{k+1} - r_k) + r_k \\ , \qquad \tau \in [kT_s, (k+1)T_s] \\ \left(\frac{XM\tau - kT_c + XMT_c}{T_c}\right) \times (r_{k-XM+1} - r_{k-M}) + r_{k-XM} \\ , \quad \tau \in [(k-XM)T_s, (k-XM+1)T_s] \\ 0, \qquad \text{others} \end{cases} \quad (9)$$

with

$$r_k = \begin{cases} \frac{1}{XM}\sum_{i=0}^{XM-1-k}(-1)^{\left\lfloor \frac{i}{XM} \right\rfloor} d_{i+k}, & 0 \le k \le XM-1 \\ \frac{1}{XM}\sum_{i=0}^{XM-1+k}(-1)^{\left\lfloor \frac{i-k}{XM} \right\rfloor} d_i, & 1-XM \le k < 0 \\ 0, & |k| \ge XM \end{cases} \quad (10)$$

where the operator $\lfloor \cdot \rfloor$ represents the fixing operation.

The shape vectors of sine-BOC(kn,n) signal can be expressed as follows

$$\boldsymbol{d}_{\mathrm{BOC}} = \left(\underbrace{1,\cdots,1}_{X} \underbrace{-1,\cdots,-1}_{X}, \underbrace{1,\cdots,-1}_{X(M-2)} \right)^{\mathrm{T}} . \tag{11}$$

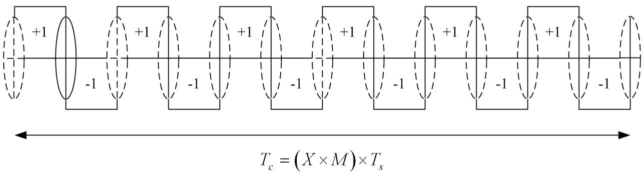

Fig. 2. Sine-phased sub-carrier for the sine-BOC modulation.

However, the request (3) makes the modulated symbol shape value of $\boldsymbol{d}_{\mathrm{SCS}}$ at the polarity inversion every two bits equal to zeros except for the second polarity inversion (see the solid ellipse line), so as to obtain fewer side peaks compared with ACF. Therefore, the shape vectors of local SCS signal can be obtained as

$$\boldsymbol{d}_{\mathrm{SCS}} = \left(\underbrace{0,\cdots,0}_{X-1}, \sqrt{XM/2}, -\sqrt{XM/2}, \underbrace{0,\cdots 0}_{X(M-1)-1} \right)^{\mathrm{T}} . \tag{12}$$

The CCFs between the received sine-BOC signals and local signals using the proposed method for sine-BOC(1,1) and sine-BOC(10,5) signals are shown in Fig. 3 and Fig. 4, respectively, without front-end filter.

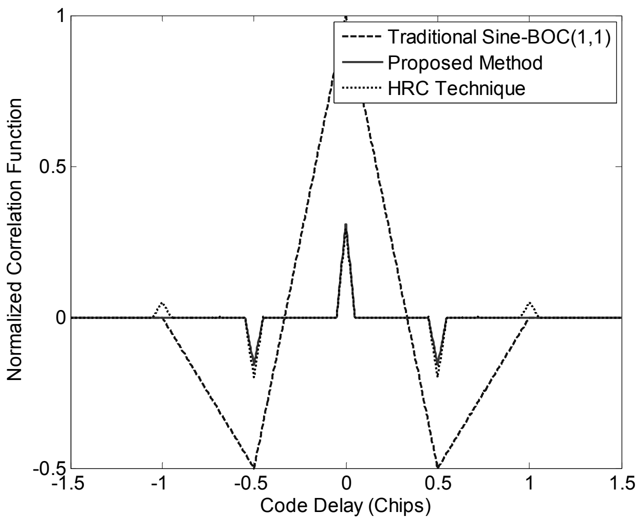

Fig. 3. Normalized correlation function of sine-BOC(1,1) signal, HRC and the proposed method, without front-end filter.

For comparison, the CCFs between the received sine-BOC signals and the local signals, which using the HRC technique and the traditional local code replicas are also shown in the figures. In the simulation, we assume that the parameter $X = 10$, the early-minus-late spacing for HRC technique are set to 0.05 chips and 0.025 chips for sine-BOC(1,1) and sine-BOC(10,5) signals, respectively. From Figs. 3, 4, it can be seen that the CCFs of the main peaks using the proposed method are triangular and the baseline widths of the triangles are much narrower than the ones of the traditional sine-BOC signals. Although the shape of the CCFs using the proposed method is similar with that of the HRC technique, the number of side peaks for the proposed method are fewer than that of HRC technique, and the amplitudes of side peaks for the proposed method are degraded relative to HRC technique. Therefore, it can provide better performance to resist the effect of multipath at medium/long multipath delays than HRC technique.

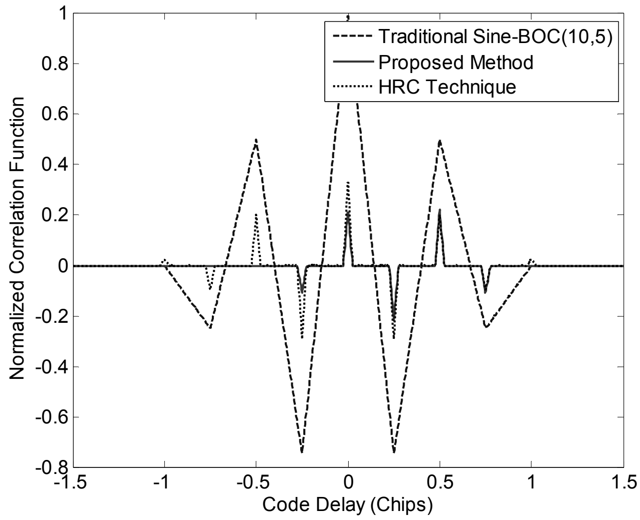

Fig. 4. Normalized correlation function of sine-BOC(10,5) signal, HRC and the proposed method, without front-end filter.

3.2 Implementation Issues

The study of the multipath mitigation performance of the proposed method has shown that the multipath mitigation performance of the proposed method is relevant to the value of $\boldsymbol{d}_{\mathrm{SCS}}$. Consequently, the new architecture of the non-coherent narrow early-minus-late (EML) tracking loop based on the proposed method is depicted in Fig. 5. Note that the only two correlators are employed by the proposed method, which is small compared with those which require dozens of correlators to mitigate multipath, such as the methods proposed in [10]-[13], as illustrated in Fig. 5. The received sine-BOC signal is first multiplied with the local carrier, and then down converted to baseband in-phase (I) and quadrature-phase (Q) signals. The local sequence generator generates early and late spreading sequence with a spacing Δ between them. Each sequence is modulated by the symbol $p_{\mathrm{SCS}}(t)$, respectively, and then does multiplication with the baseband I and Q signals in correlator. The results of this multiplier are resampled by the integrate and dump accumulators with the duration time T, and then the CCF is given by

$$R_i^{\mathrm{CCF}}(\tau) = \sqrt{I_i^2 + Q_i^2} \tag{13}$$

where i = E, L indicates early (E) or late (L). The final result of the discriminator is

$$D(\tau) = \left(R_E^{\mathrm{CCF}}(\tau)\right)^2 - \left(R_L^{\mathrm{CCF}}(\tau)\right)^2. \tag{14}$$

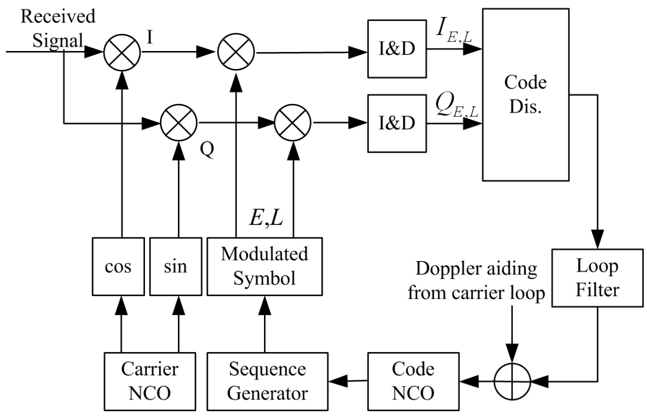

Fig. 5. New delay lock loop architecture for the proposed method.

4. Simulated Performance of Proposed Method

To simulate the effect of multipath on the code tracking, consider a simple model of multipath as a one-path specular reflection having some amplitude relative to the direct path, arriving at some phase and delay, and with all values time-invariant over the time period of interest.

Fig. 6 shows the multipath-induced error envelopes for traditional sine-BOC(1,1) signal tracking, as well as for HRC technique and the proposed method with $X = 10$ for a signal-to-multipath-amplitude ratio of 10 dB and a 24 MHz (double sided) front-end filter and early-late spacing $\Delta = 0.2$ chips. And the case of sine-BOC(10,5) signal tacking with a 30 MHz front-end filter, $\Delta = 0.1$ chips is presented in Fig. 7.

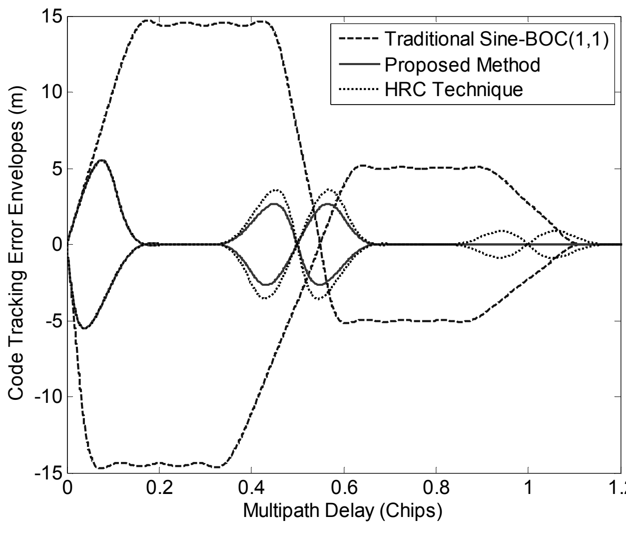

Fig. 6. Code tracking multipath envelopes for sine-BOC(1,1), HRC and the proposed method, with 24 MHz front-end filter.

From Fig. 6, it is obvious that the proposed method has a marked improvement multipath mitigation performance than the traditional method for a sine-BOC(1,1) signal. And notice that the difference between the proposed method and HRC technique is minimal at short multipath delays within [0;0.15] chips, while for multipath delays within [0.2;1.2] chips, the proposed method seems to mitigate multipath better than HRC technique. This phenomenon and the fact that the number of side peaks for the proposed method are fewer than that of HRC technique, and the amplitudes of the side peaks for the proposed method are degraded relative to those of HRC technique.

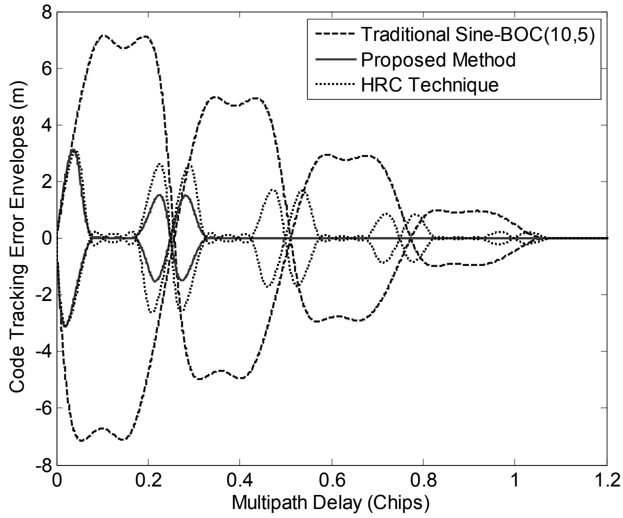

Fig. 7. Code tracking multipath envelopes for sine-BOC(10,5), HRC and the proposed method, with 30 MHz front-end filter

Fig. 7 shows the performance of the multipath mitigation of the proposed method for a sine-BOC(10,5) signal with 30 MHz filter bandwidth. Note that at all multipath delays, the proposed method seems to mitigate multipath much better than the traditional method. The difference between the proposed method and HRC technique is still minimal at short multipath delays within [0;0.35] chips. However, the proposed method provides better performance than HRC technique at medium/long multipath delays, especially the multipath delay effects are completely mitigated within [0.4;1.2] chips. Furthermore, with the increase of *M*, the multipath mitigation performance of the proposed method is improved more obviously.

5. Code Tracking Noise Evaluation

Although the proposed method provides substantial mitigation of medium and long delay multipath, this method results in CCF whose peak value is reduced. This means that there will be a penalty in thermal noise performance associated with this method, as well as HRC technique. The post-correlation I and Q prompt channel samples are modeled as

$$I_P = AR_{\mathrm{B/S}}^{\mathrm{CCF}}(\tau)\cos\theta + n_{IP}, \tag{15}$$

$$Q_P = AR_{\text{B/S}}^{\text{CCF}}(\tau)\sin\theta + n_{QP}, \tag{16}$$

$$A = \sqrt{2CN_0T_c} \tag{17}$$

where T_c represents the pre-detection integration interval, C/N_0 denotes the carrier-to-noise ratio at the detector, θ is the carrier phase angle, n_{IP} and n_{QP} refers the I and Q channel noise are correlated with the local signal, respectively.

Following [9], [10], the code tracking error standard deviation (in meters) caused by noise for a first order delay lock loop (DLL) is given by

$$\sigma_{DLL} = \lambda_c k_{DLL}\sqrt{r_{EML}} \tag{18}$$

with

$$k_{DLL} = \sqrt{\frac{2B_LT_c}{1-2B_LT_c}} \tag{19}$$

where $\lambda_c = 293$ m/chips, B_L is the noise equivalent bandwidth in Hz, and r_{EML} denotes the variance of the DLL discriminator output. The reduction in noise, i.e. (15) and (16) due to the DLL will be the same for all correlator methods [9], [10]. For conventional sine-BOC signal code tracking, HRC tracking, and the proposed method tracking, the variance of the DLL discriminator output r_{EML} can be expressed as

$$\begin{aligned} r_{\text{Traditional}} &= \frac{\Delta}{2(1+C/N_0T_c)} \\ r_{\text{HRC}} &= \frac{\Delta}{(1+C/N_0T_c)} \\ r_{\text{Proposed}} &= \frac{(X+8)\Delta}{18(1+C/N_0T_c)} \end{aligned} \tag{20}$$

where Δ is the early-late spacing in chips.

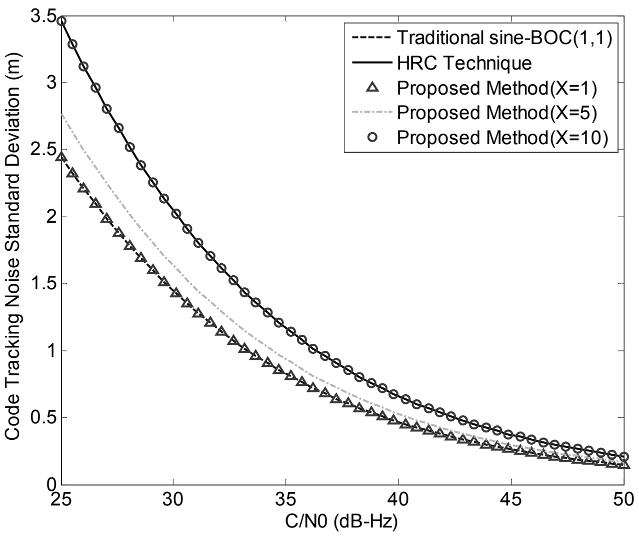

Fig. 8. Code tracking noise standard deviation for sine-BOC(1,1), HRC and the proposed method.

Fig. 8 shows the code tracking noise standard deviation for the proposed method with $B_L = 0.5$ Hz and $T_c = 1$ ms. For comparison, it also shows the code tracking noise standard deviation for the traditional BOC tracking method and HRC technique. The figure indicates that the code tracking error for the proposed method with $X = 10$ is worse than the traditional BOC tracking, as well as HRC technique (solid line, nearly completely below the dotted line). While $X = 1$, the proposed method is the same as the traditional sine-BOC. Due to the narrow spacing values and the degraded magnitude of the main peak, the proposed method suffers the effect of the noise. However, these effects on the proposed method can be offset by use of narrower loop bandwidths, although practical considerations impose limits on narrowing loop bandwidths [16].

6. Conclusions

In this paper, an efficient method for multipath mitigation in GNSS systems is presented. The proposed method uses a CCF with a tunable parameter X for multipath mitigation by designing the modulated symbols of the local SCS signal. The simulation results indicate that the proposed method provides better performance for multipath mitigation than HRC technique, especially at medium/long multipath delays, with minor or negligible degradation in noise performance. In the aspect of hardware, the proposed method is easy and flexible to implement, because only two correlators are needed for mitigating multipath signal. Besides, the proposed method provides a new concept based on the CCF to mitigate multipath. Future works will focus on finding more optimum local SCS waveform under the proposed framework to resist the effect of multipath, with less degradation than the existing methods.

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