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A METHOD OF SIDE-PEAK MITIGATION APPLIED TO BINARY OFFSET CARRIER MODULATED GNSS SIGNALS TRACKING APPLIED IN GNSS RECEIVERS

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ABSTRACT

In this study, a new method of signal tracking technique in Global Navigation Satellite System (GNSS) is proposed. It is based on a combination of the autocorrelation function (ACF) with another cross correlation function in order to eliminate or reduce the power of the side peaks in ACF of Binary Offset Carrier (BOC) modulated signals. These types of modulated signals are adopted by both GNSSs like the modernized Global Positioning System (GPS) and Galileo. Moreover, this method still keep the sharp of main peak of ACF in order to maintain the advantage of BOC(n,n) signals in code tracking and multipath mitigation. In the proposed method, the output of the discriminator in delay tracking loop has no false lock point. The performance of multipath mitigation of the proposed method is better than Narrow Correlator method. The good performance of the proposed scheme in multipath mitigation has been tested using simulation results.

Keywords: binary offset carrier, multipath estimation, side peak cancellation, auto-correlation function, ambiguous tracking.

INTRODUCTION

Multipath is one of the most critical error sources for Global Navigation Satellite System (GNSS) receivers [1, 2]. Multipath is the propagation of the signals from satellites through different paths due to the reflection, diffraction of the direct signal on the surfaces nearby the receivers. Therefore, the signal reached the receiver superimposes the direct signal and some multipath signals. Multipath signals cause the distortion in correlation function between the received signal and the local signal generated by the receiver. The reception of multipath creates a bias and corruption into the time delay estimation of the Delay Lock Loop (DLL) of a GNSS receiver. This will lead to inaccurate position estimation or loss of lock on the signal. There are many solutions for mitigating the effects of multipath proposed in literature. Multipath mitigation techniques could be approached from several directions: pre-receiver techniques or antenna - based techniques are used before signal enters the receiver, the post-processing techniques are applied after the pseudo-range measurements have been produced and the code tracking techniques based on the correlation function [3]. Among them, the correlation - based multipath mitigation techniques are the most common and prominent approach due to their convenience in eliminating multipath errors in GNSS receivers.

The most widely known delay estimator is the standard wide correlator DLL or Early-Minus-Late (EML) loop [4]. However, this method fails to cope with multipath propagation. Therefore, several enhanced EML-based techniques have been proposed in order to mitigate the multipath immunity such as the Narrow Correlator (NC) [4], the Double Delta Correlator (DDC) [4]. Initially, these techniques were designed for legacy signal of Global Positioning System (GPS) called GPS L1 C/A (Coarse/Acquisition) [1]. Currently, both the modernized GPS and Galileo adopt new modulation for their new navigation signals. This new modulation is Binary Offset Carrier (BOC) which offers split-spectrum properties as well as better code tracking performance [5]. This modulation also provides the high degree of spectral separation between GPS L1 C/A signal and new navigation signals. Therefore, these signals could share the same carrier without interference [6]. Almost above - mentioned multipath mitigation techniques could be applied easily to BOC modulated signals. Moreover, based on new properties of BOC modulated signals, some new multipath mitigation techniques have been proposed as in [7-9]. These techniques use the correlation function between BOC signal and local replica of pseudorandom noise (PRN) code called BOC-PRN correlation function (CF) instead of the BOC-BOC one. The resulting performance of these techniques related to discriminator output of DLL in tracking stage in GNSS receivers like the performance of DDC for Binary Phase Shift Keying (BPSK) signal (e.g. GPS L1 C/A) [7].

However, the autocorrelation function (ACF) of BOC modulated signals have multi peaks (one main peak and some side peaks) that raise new challenges in code tracking [3]. Due to non-negligible side peaks, the risk of wrong peak selection in signal acquisition is higher. This problem is called ambiguity. Therefore, the correlation - based multipath mitigation techniques are not good enough for ambiguous cancellation, which is one of key motivations for present-day researchers. Several new innovative techniques have been proposed in order to avoid the ambiguous problem [10-15]. BPSK -like technique [11] considers the received BOC signals as the sum of two BPSK signals with carrier frequency symmetrically positioned on each side of the BOC carrier frequency. Therefore, each sideband is treated separately as a BPSK signal, which provides an unambiguous correlation function and a wider S-curve steady domain. However, this kind of method suffers from some disadvantages. The sharp peak of BOC ACF is destroyed, the root-mean-square (RMS) bandwidth of the received signal approaches to that of BPSK signal, and thus, it removes the advantages of BOC signal in multipath (MP) mitigation. In [15], the Sub-Carrier Phase Cancellation (SCPC) obtains a correlation function similar to BPSK one. It is based on generating an in-phase and a quadrature-phase local sub-carrier signals in additional to the local in-phase and quadrature-phase carrier signals. Thus, the received filtered signal is correlated with local BOC signal in both sub-carrier phase and sub carrier quadrature. Then, these two correlation channels are combined. However, the disadvantage of the SCPC method is that the shape of the main peak is destroyed. An innovative unambiguous tracking method named Autocorrelation Side-peak Cancellation Technique (ASPeCT) is described in [12]. This method subtracts the BOC-PRN CF from the BOC-BOC CF. Unfortunately, this technique is only applied to sine phased $BOC(n,n)$ modulated signals. Moreover, the performance of multipath mitigation of this technique is equivalent to the one of traditional correlation-based techniques.

To sum up, all above mentioned methods have different drawbacks. Some methods remove the advantage of BOC signal in multipath mitigation. The rest of methods provide a weak elimination of the side peaks. Thus, the side peaks always exist and cause the influence of the multipath on these methods.

In the present study, a new architecture for BOC code tracking is proposed. The scheme involves five correlators named here very late, late, prompt, early and very early of both BOC-BOC CF and BOC-PRN CF to create new correlation function, which will be used for code tracking. The proposed correlation function could be applied to sine- phased $BOC(n,n)$ as well as cosine-phased $BOC(n,n)$ modulated signals related to ambiguity cancellation. For sine-phased $BOC(n,n)$ signal, the ambiguity problem will be removed. For cosine-phased $BOC(n,n)$, some side peaks will be removed and the rest side peaks will be minimized their occurrence. On other hand, the width of the main peak is still kept narrow. Therefore, the proposed method remains the advantage of $BOC(n,n)$ signals in code tracking in comparison to BPSK signal. Moreover, the proposed method gives a better performance in multipath mitigation.

The rest of this paper is organized as follows. In Section II, the characteristic of BOC signal and multipath signal model are described. After that, Section III shows the principle of our proposed method. The numerical results of comparison of our proposed method to traditional one are presented in Section IV. Finally, Section V concludes this paper.

BOC modulation and multipath signal model

The characteristics of BOC modulation

The baseband BOC modulated signals are the result of multiplying the non-return-to-zero (NRZ) PRN code and a synchronized square waveform subcarrier which is equal to the sign of a sine or cosine waveform [16]. Depending on the initial phase of subcarrier, the BOC modulated signal will be sine-phased BOC signal (or cosine-phased BOC one), denoted as $sBOC(m,n)$ (or $cBOC(m,n)$) if the initial phase of subcarrier is 0 radian (or $\pi/2$ radians). The parameter m stands for the ratio of the square wave frequency f_s to 1.023MHz, and n means the ratio of the chip rate f_c to 1.023MHz. Another important parameter of BOC modulated signal is modulation order N_B and defined as $N_B = 2\frac{m}{n}$. It is also remarked that N_B should be an integer number. For example, $N_B = 2$ for $BOC(n,n)$ and $N_B = 4$ for $BOC(2n,n)$. It is noted that $N_B = 1$ is the special case of BOC modulated signal, it is BPSK one. According to its definition in [16], BOC modulated signal could be expressed as

$$s_{BOC}(t) = s(t).sign(\sin(2\pi f_s t + \phi) \qquad (1)$$

where $s(t)$ is a baseband BPSK signal, $sign(.)$ is the sign operator.

The subcarrier will create one or more zero crossing on spreading chip duration. The number of zero crossing depends on subcarrier frequency and chip rate. The waveform of $sBOC(n,n)$ and $cBOC(n,n)$ modulated signals are shown in Figure-1.

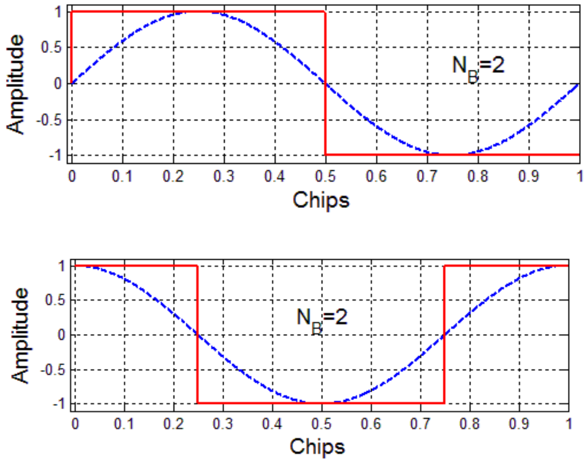

Figure-1. Waveforms of sBOC (n, n) (Top) and cBOC (n, n) (Bottom).

Let denote $T_c = 1/f_c$ as code duration in chips, the power spectral density (PSD) of the $sBOC(n,n)$ as well as $cBOC(n,n)$ signal are given by [5]

$$G_{sBOC}(f) = T_c \operatorname{sinc}^2(\pi f T_c) \tan^2\left(\frac{\pi f T_c}{2}\right)$$
$$G_{cBOC}(f) = T_c \operatorname{sinc}^2(\pi f T_c) \tan^2\left(\frac{\pi f T_c}{2}\right) \tan^2\left(\frac{\pi f T_c}{4}\right) \qquad (2)$$

Figure-2 illustrates the normalized PSDs of BPSK, $sBOC(n,n)$ and $cBOC(n,n)$ signals. As depicted, the PSD of BOC modulated signals is split symmetrically into two parts, and moves the main energy component away from carrier frequency. This will allow the co-existent of BPSK signal as well as BOC signal at the same carrier frequency.
The filtered ACF is related to the PSD by [17].

$$R(\tau) = \int_{-\infty}^{+\infty} G(f) H(f) e^{j2\pi f \tau} df \qquad (3)$$

where $H(f)$ is the transfer function of the receiver frontend filter. In case the frontend filter is ideal with bandwidth B , the filtered ACF becomes

$$R(\tau) = \int_{-B/2}^{+B/2} G(f) e^{j2\pi f \tau} df = \int_{-B/2}^{+B/2} G(f) \cos(2\pi f \tau) df \qquad (4)$$

The filtered ACF of $sBOC(n,n)$ and $cBOC(n,n)$ for some different bandwidths is shown in Figure-3. It is noted that when the bandwidth is reduced, the correlation peaks is attenuated. It can be seen that in addition to the main peak, the ACF of BOC modulated signals have some side peaks. The number of side peaks is $2(N_B - 1)$ and $2N_B$ for sine-phased BOC and cosine-phased BOC signals, respectively. The more the number of side peaks, the higher the risk of false lock in code tracking.

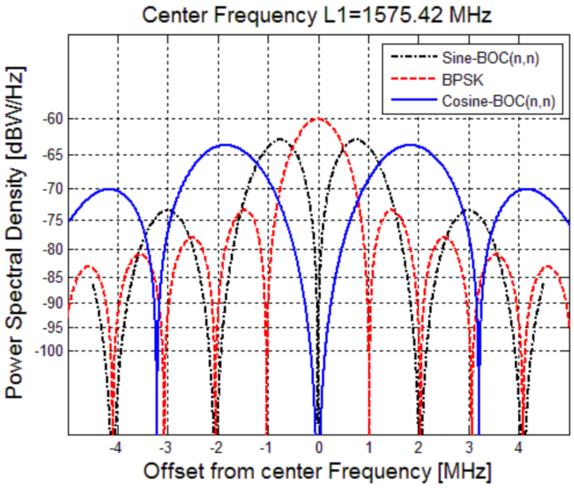

Figure-2. PSDs of BPSK, sBOC (n, n) and cBOC (n, n) signals.

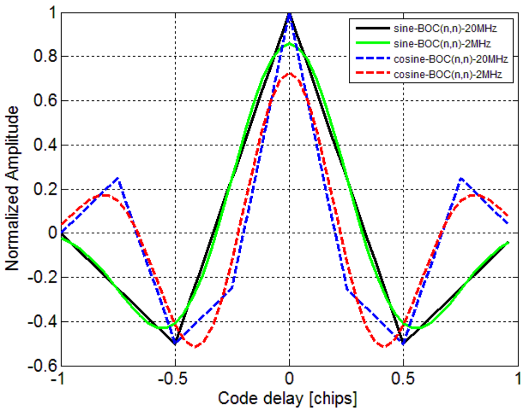

Figure-3. The effects of the frontend filter bandwidth to ACF.

The multipath signal model

The navigation signal $s(t)$ is typically transmitted over multipath static or fading channel. It means that before reaching the receiver, the GNSS signal is reflected, scattered and diffracted in the surrounding environment. As consequence, beside a Line-of-Sight (LOS) component, additional replicas with different delays, phases and frequencies also arrive at the receiver. Therefore, the incoming signal can be expressed as

$$r(t) = \sum_{i=0}^{L-1} \alpha_i s(t - \tau_i) e^{j(2\pi f_D t + \theta_i)} + n(t) \qquad (5)$$

where $r(t)$ is the received signal, L is the number of channel paths, α_i and τ_i are respectively the amplitude and delay of each replica (typically assumed to be slowly varying or constant within the observation interval), f_D is the Doppler shift introduced by the channel and $n(t)$ is Additive White Gaussian Noise (AWGN) of the channel.
The multipath effect cause the error in both code delay and carrier phase estimation. However, the error in carrier phase estimation depends on the carrier frequency (e.g. band $L1 = 1575,42 MHz$), thus, it is much small. Whereas, the code delay is used to determine the pseudo-ranges to each satellite and therefore the error in its estimation is the main contributor to the error in the final solution. Consequently, most multipath mitigation techniques proposed in literature by researchers try to reduce the code delay error [4].

Ambiguous problem

While ACF of BPSK signals is a triangle, the BOC signals have a sawtooth-like, piecewise linear ACF which has multiple peaks and multiple zero crossing points. In a traditional GNSS receiver, the code tracking loop is a DLL, where the input signal is correlated with three replicas of the code. A traditional DLL structure adapted to BOC signals is shown in Figure-4. When using a traditional Early-Minus-Late Power (EMLP) tracking loop, the discriminator output is expressed as [1]

$$\begin{aligned} D_{EMLP} &= \left(I_E^2+Q_E^2\right)-\left(I_L^2+Q_L^2\right) \\ &= \left[\left(R_I\left(\tau-\delta/2\right)\right)^2+\left(R_Q\left(\tau-\delta/2\right)\right)^2\right]- \\ &\left[\left(R_I\left(\tau+\delta/2\right)\right)^2+\left(R_Q\left(\tau+\delta/2\right)\right)^2\right] \end{aligned} \qquad (6)$$

where I_E, Q_E, I_L, Q_L denote respectively the in-phase component and quadrature component output of Early (E) and Late (L) correlator arm and δ is early-late spacing in chips.

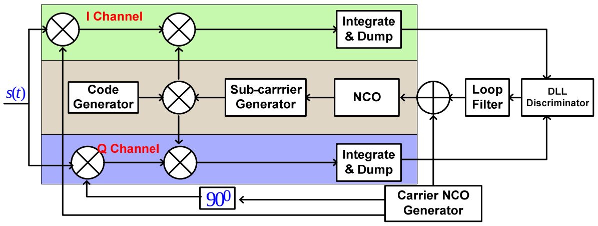

Figure-4. Traditional tracking loop structure.

The ACFs and discriminator characteristic curves (S-curves) for $sBOC(n,n)$ and $cBOC(n,n)$ signal with early-late spacing $\delta = 0.1chips$ are shown in Figure-5. It can be found that two (02) stable false lock points for $sBOC(n,n)$ signal (four (04) ones for $cBOC(n,n)$ signal) are identified due to side-peaks of ACF. The tracking loop can be locked on one of the side-peaks, which could result in intolerable biased measurements. This problem is called as ambiguity for BOC signal tracking. In order to avoid the ambiguous problem, all side peaks should be removed or at least minimized occurrence. Therefore, the probability of wrong peak selection could be reduced.

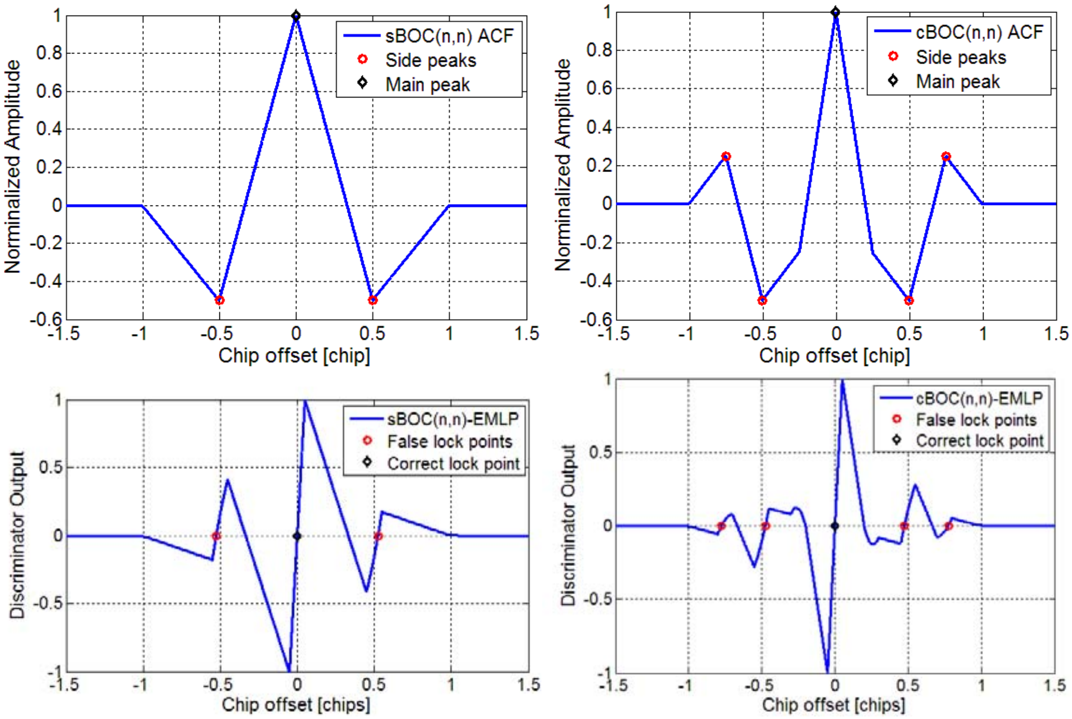

Figure-5. The ACFs (Top) and S-curves (Bottom) of EMLP DLL for sBOC (n, n) (Left) and cBOC (n, n) (Right).

The proposed method of multipath estimation

In order to overcome the limitation of $BOC(n,n)$ modulated signal, in term of side-peak cancellation, we propose a new method based on the combination of BOC-BOC CF with BOC-PRN CF. For the BOC-PRN CF, instead of dispreading the satellite signal by multiplying by a locally generated replica of $BOC(n,n)$ in code tracking loop, the received signal is multiplied only by the PRN code [7]. With unlimited bandwidth of pre-correlation filter, BOC-PRN CF could be expressed as:

For $sBOC(n,n)$

$$R_{BOC/PRN}(\tau)=\frac{1}{2}\left[tri\left(\frac{\tau-1/2}{1/2}\right)-tri\left(\frac{\tau+1/2}{1/2}\right)\right] \qquad (7)$$

where $tri(x/y)$ is triangular function of width $2y$, centered in $x=0$ where it has a unity value.

For $cBOC(n,n)$

$$\begin{aligned}R_{BOC/PRN}(\tau)=\frac{1}{4}\Bigg[&tri\left(\frac{\tau-1/4}{1/4}\right)+tri\left(\frac{\tau+1/4}{1/4}\right)\\&-tri\left(\frac{\tau-3/4}{1/4}\right)-tri\left(\frac{\tau+3/4}{1/4}\right)\Bigg]\end{aligned} \qquad (8)$$

Figure-6 illustrates shapes of two BOC-PRN correlation functions for $sBOC(n,n)$ and $cBOC(n,n)$ signal.

In case of BOC-BOC CF, this CF could be written as

For $sBOC(n,n)$

$$R_B(\tau)=tri\left(\frac{\tau}{1/2}\right)-\frac{1}{2}\left[tri\left(\frac{\tau-1/2}{1/2}\right)+tri\left(\frac{\tau+1/2}{1/2}\right)\right] \qquad (9)$$

For $cBOC(n,n)$

$$\begin{aligned}R_B(\tau)=tri\left(\frac{\tau}{1/4}\right)-\frac{1}{4}\Bigg[&tri\left(\frac{\tau-1/4}{1/4}\right)+tri\left(\frac{\tau+1/4}{1/4}\right)\\&-tri\left(\frac{\tau-3/4}{1/4}\right)-tri\left(\frac{\tau+3/4}{1/4}\right)\Bigg]\\&-\frac{1}{2}\left[tri\left(\frac{\tau-1/2}{1/4}\right)+tri\left(\frac{\tau+1/2}{1/4}\right)\right]\end{aligned} \qquad (10)$$

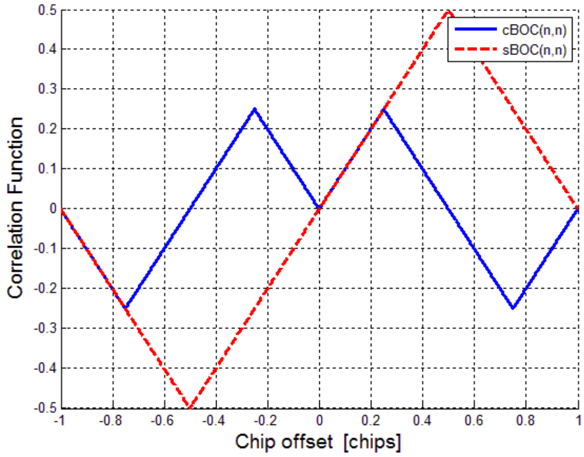

Figure-6. BOC-PRN CF for sBOC (n, n) and cBOC (n, n) signals.

Eq. (7) and Eq. (9) show that there are two triangles of BOC-PRN CF locating on the side peaks of the $sBOC(n,n)$ ACF with the same magnitude. For $cBOC(n,n)$, Eq. (8) and Eq. (10) also show that two triangles of BOC-PRN CF locate on two of four side peaks of the $cBOC(n,n)$ ACF with the same magnitude. Therefore, by subtracting $R^2_{BOC/PRN}(\tau)$ from $R^2_B(\tau)$, two undesired side peaks could be removed. This is the first stage of the proposed method and it is written as

$$R_{s1}(\tau)=R_B^2(\tau)-R^2_{BOC/PRN}(\tau) \qquad (1)$$

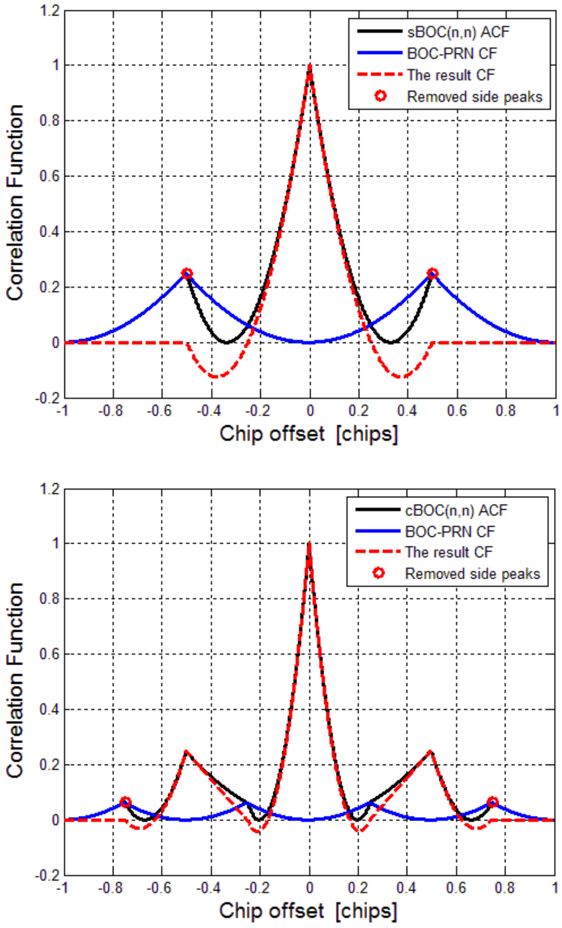

Figure-7. The first stage of removing side peaks for sBOC (n, n) signal (Top) and cBOC (n, n) signal (Bottom).

Figure-7 shows the correlation functions after the first stage of combination. Indeed, as shown in Figure-7, only two outer side peaks of the $cBOC(n,n)$ ACF are removed by this combination. For two inner side peaks, the proposed method will try to minimize their magnitudes. To do so, three (03) replicas (early, late and prompt) of CF in the first step $R_{s1}(\tau)$ are needed to compute the proposed correlation function as:

$$
\begin{aligned}
R_{proposed}(\tau) &= kR_{s1}^2(\tau) - R_{s1E}(\tau)R_{s1L}(\tau) \\
&= kR_{s1}^2(\tau) - R_{s1}(\tau-\delta/2)R_{s1}(\tau+\delta/2)
\end{aligned}
\qquad (2)
$$

where $R_{s1E}(\tau), R_{s1L}(\tau)$ are early, late versions respectively of $R_{s1}(\tau)$, δ is early-late spacing (in chips) and k is a modified factor in order to optimum the performance of the proposed method. The proposed correlation function with $k=1$ is shown in Figure-8 along with BOC-BOC ACF and BOC-PRN CF for $sBOC(n,n)$ and $cBOC(n,n)$ signals

As shown in Figure-8, the proposed method still removes completely all side peaks for $sBOC(n,n)$ signals. For $cBOC(n,n)$ signals, although this method could not remove all side peaks, the magnitudes of remain side peaks are reduced significantly. In the proposed correlation function, the magnitudes of the side peaks are equivalent to the ones of outer side peaks in the BOC-BOC ACF. It can be said that two the biggest side peaks in the BOC-BOC ACF have been removed. Therefore, the risk of wrong peak selection could be reduced. Consequently, the risk of false lock in code tracking loop is reduced. Finally, as illustrated in Figure-8, the width of the main peak in the proposed correlation function is much narrower than the one in the BOC-BOC ACF in both case of $sBOC(n,n)$ and $cBOC(n,n)$ signals. It means that all advantages of BOC modulated signal in code tracking are still kept. Moreover, the performance of the proposed method in multipath mitigation could be better than traditional method (Narrow Correlator [18]) as well as other techniques mentioned in the first section. This is a motivation of the proposed method.

In order to implement the proposed method in GNSS receivers, a new DLL architecture based on this method can be proposed. This new architecture is depicted in Figure-9. As shown in Figure-9, the proposed DLL architecture still uses EMLP discriminator as the NC method.

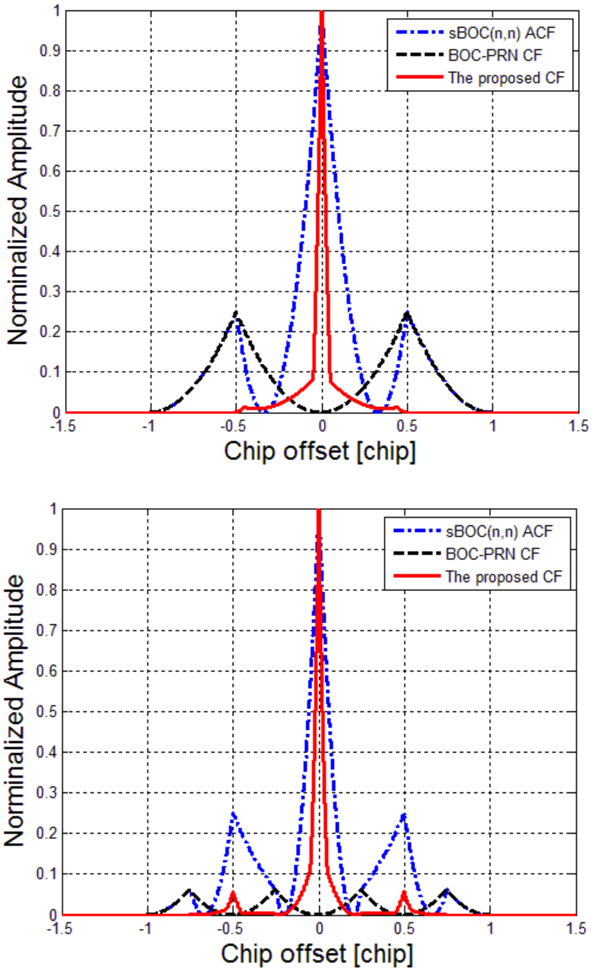

Figure-8. The proposed method for sBOC (n, n) signal (Left) and cBOC (n, n) signal (Right).

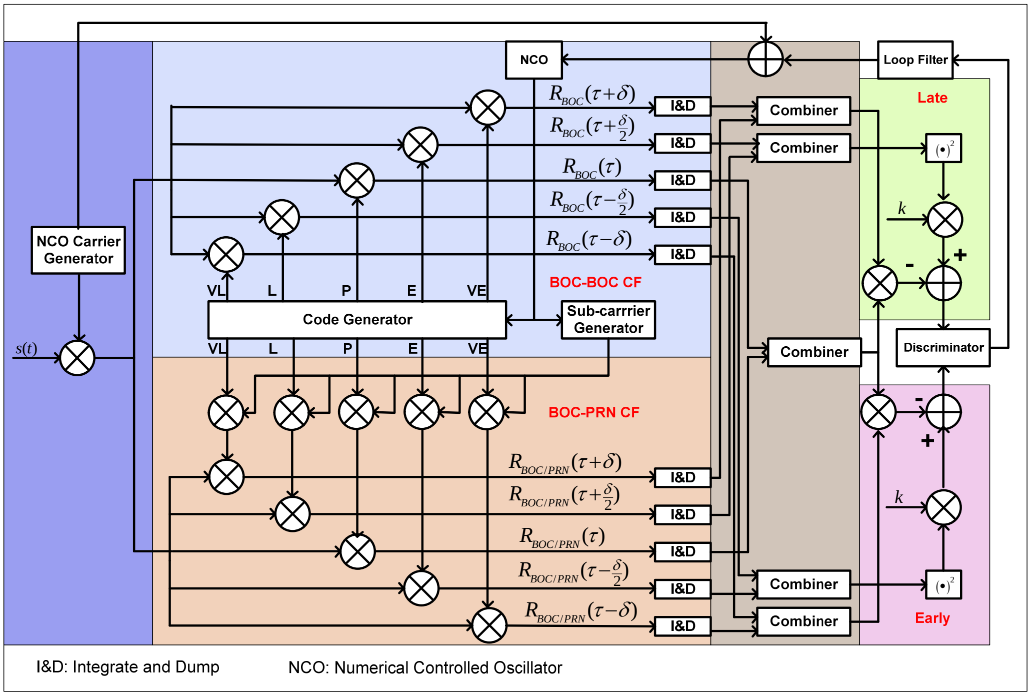

Figure-9. The DLL structure of the proposed method.

However, to create the early and late replicas of the proposed correlation function $R_{proposed}(\tau)$, five correlators named very late (VL), late (L), prompt (P), early (E) and very early (VE) of $R_{s1}(\tau)$ should be used. Therefore, the early and late replicas of the proposed CF can be modeled as

$$R_{proposedE}(\tau) = kR_{s1}^2(\tau-\delta/2) - R_{s1}(\tau-\delta)R_{s1}(\tau)$$
$$R_{proposedL}(\tau) = kR_{s1}^2(\tau+\delta/2) - R_{s1}(\tau)R_{s1}(\tau+\delta) \quad (3)$$

Now, the principle of the proposed scheme has been explained in detail. The test and verification for the proposed correlation function as well as its code tracking performance could be implemented via simulations. Moreover, it is important to assess the influence of multipath propagation on the code tracking performance to ensure that it does not imply significant drawbacks. These problems are studied in the next section.

SIMULATION RESULTS AND DISCUSSIONS

Simulations have been carried out in closely spaced multipath scenarios for $sBOC(n,n)$ and $cBOC(n,n)$ signals for finite frontend bandwidth. The channel model used in the simulation is Rayleigh fading, where the number of channel paths is one multipath component. The channel path is assumed to be a decay Power Decay Profile (PDP) which can be expressed as [19]

$$\alpha_i = \frac{p_o}{\sqrt{\tau_0}} e^{-\frac{\tau}{2\tau_0}} \quad (4)$$

where τ_0 is the typical multipath delay and p_0 is the total amount of multipath power, originating from all multipath sources. The received GNSS signal was sampled with oversampling factor $N_s = 24$ ($N_S = 48$ for BPSK signal) in order to have the same number of samples per code chip for BPSK, $sBOC(n,n)$ and $cBOC(n,n)$ signals. The early-late spacing $\delta = 0.1 chips$ .Therefore, the path delay of multipath signal varies from $\tau = 0$ to $\tau = 1.05 chips$. In theory, the longer path delays could not cause multipath errors with this value of early-late spacing.

The proposed correlation function

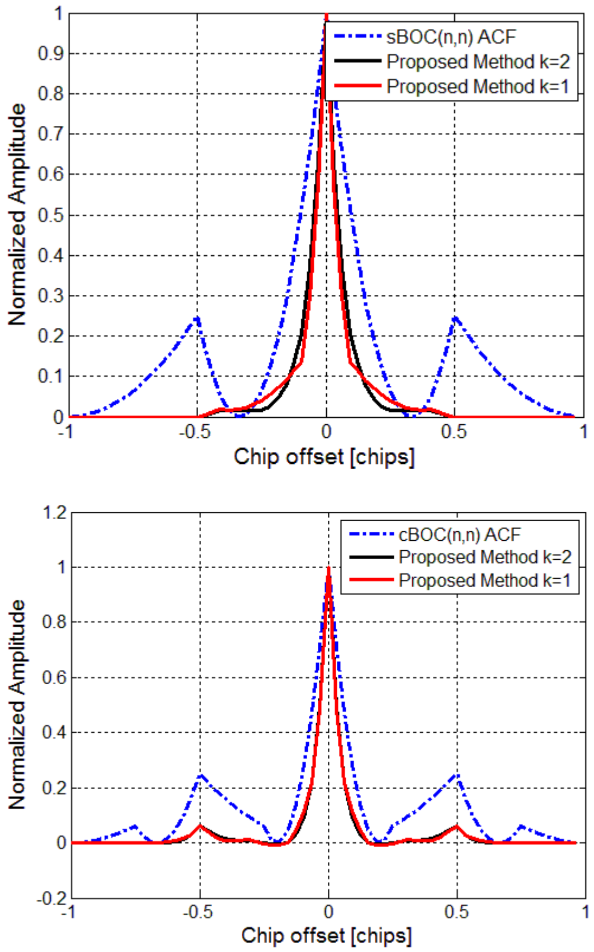

Figure-10. ACF of sBOC (n,n) and the proposed CF for sBOC (n, n) (Top) and cBOC (n, n) (Bottom).

For verifying the characteristic of the proposed correlation functions, the received GNSS signal only includes a single direct component and the frontend filter has infinitive bandwidth. The shape of correlation functions of the proposed method with $k = 2, k = 1$ along with the ACF for $sBOC(n,n)$ and $cBOC(n,n)$ signals are illustrated in Figure-10. From this Figure, it can be observed that the main peak of the proposed method is narrower than that of $sBOC(n,n)$ ACF. Moreover, there is no undesired side peaks in the proposed correlation function.

For $cBOC(n,n)$ signal, along with the main peak, there are 02 side peaks, and thus the ambiguity could not be completely removed. However, the number of side peaks of the proposed method is still fewer than that of $cBOC(n,n)$ ACF. So that, the risk of wrong peak selection is smaller. It is noted that if $k \geq 1$, there is a little effect on the proposed correlation function. If $k < 1$, it can destroy the shape of the proposed correlation function.

The S-curves of discriminators ouputs

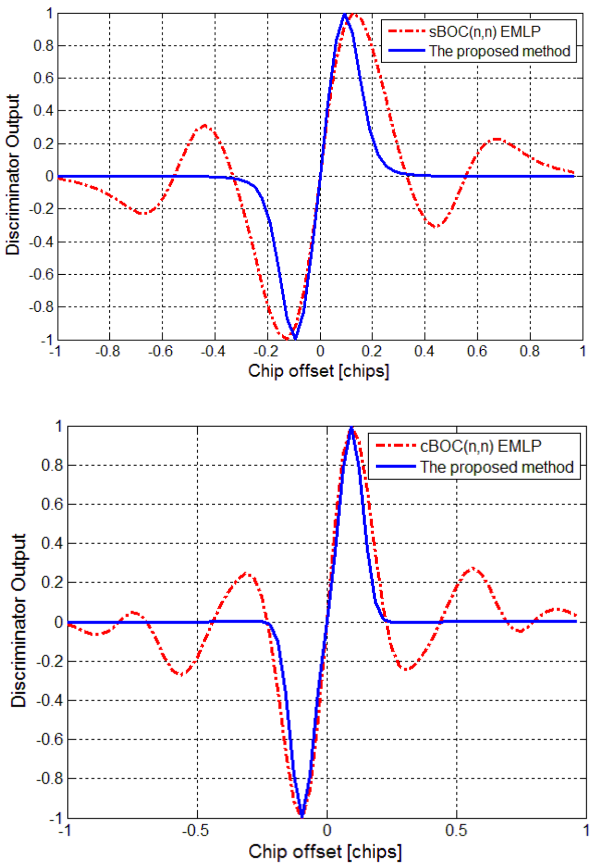

Figure-11. S-curves with the proposed and NC method for sBOC (n, n) (Top) and cBOC (n, n) (Bottom).

The performance of the tracking loops can be assessed by studying the discriminators curves (S-curves). These curves present the discriminators' outputs as a function of code delays. Figure-11 illustrates the characteristics of the S-curves with the proposed method and NC method for $sBOC(n,n)$ and $cBOC(n,n)$ signals. In both simulations, the correlator structure in DLL is Narrow correlator with early-late spacing $\delta = 0.1 chips$ and the frontend filter bandwidth of $6MHz$. Obviously, in comparison to the traditional method, the proposed method has no false lock point for $sBOC(n,n)$ signal. The probability of appearance of false lock point $cBOC(n,n)$ signal is significantly reduced. On the other hand, two tracking parameters of interest that can be derived from the S-curves are the linear tracking region, the region in which the discriminator reacts perfectly without any bias, and the stability region (pull-in region), defined as the region where the discriminator reacts in the correct direction. Both these region are located around the zero code delay error. As shown in Figure-11, the proposed method could provide the same of stability region and linear tracking region as the traditional method. It means that the performance of code tracking of the proposed method as robust as the one of traditional method.

The effects of multipath

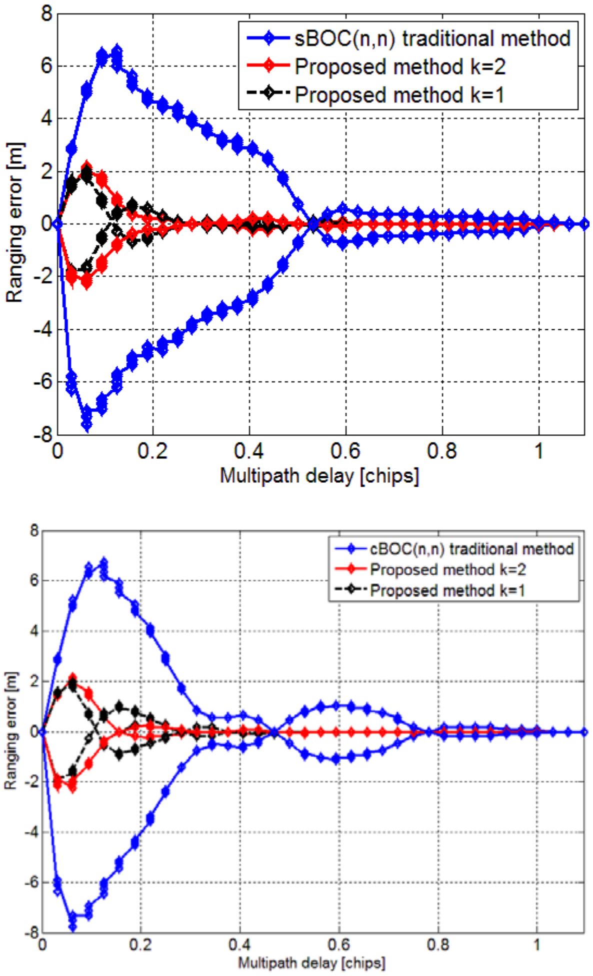

Figure-12. MEE for sBOC(n,n) (Top) and cBOC(n,n) (Bottom) with NC method and proposed method.

In order to assess the influence of multipath propagation, one of the main error sources in GNSS, to the performance of code tracking, the most common way is Multipath Error Envelope (MEE) [19]. A line-of-sight (LOS) signal and a single multipath signal are considered in this scenario. The error envelopes are calculated by determining the zero crossing point of the DLL discriminator. The multipath environment is rural environment which is mainly characterized by trees and forests located along the rural highways. For elevation angle of satellite of $E = 25^0$, the typical multipath characteristic can be [19]:

$$SMR = 13.5dB \qquad \tau_0 = 0.1901chips \tag{5}$$

where SMR is signal-to-multipath ratio. It is the ratio between the power of the LOS signal and multipath signals.

Based on the environmental multipath characteristics, the results MEE of the proposed method with $k = 2, k = 1$ along with the NC method for $sBOC(n,n)$ and $cBOC(n,n)$ signals are illustrated in Figure-12. As shown in this Figure, the error envelopes are completely reduced in the proposed method for all the code delays. The error envelopes of the proposed method decrease to reach nearly zero for multipath delays which are greater than 0.6chips with respect to the LOS signal. It means that the proposed method has a perfect performance for medium-to-long delayed multipath. For short-delayed multipath (delay is smaller than 0.2chips), it over-performs the traditional method. These results show the efficiency of the proposed method with respect to that of the traditional method in terms of multipath mitigation.

Optimization of the value of modified factor

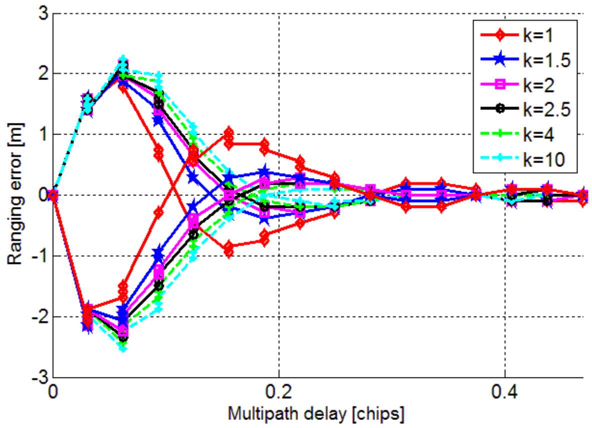

Figure-13. Optimization of modified factor for the proposed method.

As shown in Figure-12, the value of modified factor k could affect the performance of the proposed method in multipath mitigation. Therefore, the value of k should be optimized in order to reduce the multipath error. It means that the optimization is done according to the estimated multipath errors on average, in such a way to minimize them. The optimization criterion is multipath error envelope (MEE). A good delay tracking structure should provide small average errors, small worst errors and small multipath delay after which MEE becomes zero. For convenience, the search ranges for k were between $+1$ and $+10$, with a step of 0.5. Because, if the value of k is smaller than $+1$, the shape of the S-curve is destroyed and the delay tracking structure will be sensitive to multipath signal. If the value of k is larger than 10, there is a little change in the multipath mitigation. Figure-13 illustrates the MEE of the proposed method with some values of modified factor k. As shown in Figure-13, it can be seen that the value of k should be chosen between $+1.5$ and $+2$ for the better multipath mitigation. However, the variation of the value of k in this range could not impact significantly to the performance of the delay tracking structure.

CONCLUSIONS

In this paper, an efficiency method for multipath mitigation in GNSS receivers is proposed. It is based on the nonlinear combination of BOC-BOC ACF and BOC-PRN CF. The proposed method could be applied to $sBOC(n,n)$ and $cBOC(n,n)$ signals. It completely eliminates the ambiguity problem when applying to $sBOC(n,n)$ signal. For $cBOC(n,n)$ signal, the power of side peaks is very small with respect to that of main peak, thus, the risk of locking on side peaks could be reduced. In addition, the multipath mitigation performance of the proposed method is significantly better than NC method. Therefore, it works for medium-delayed and long-delayed multipath. Although the performance of the proposed method is sensitive to short-delayed multipath, it still outperforms traditional method.

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