A FAST UNAMBIGUOUS ACQUISITION ALGORITHM FOR BOC-MODULATED SIGNALS

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***Abstract*— This paper proposes a fast unambiguous acquisition technique for BOC-modulated signals. We remove the ambiguities (side-peaks) of the BOC autocorrelation function, exploiting a reduced-complexity (real and symmetric) filter composed of only 7 non-zero samples. The proposed scheme is applicable to both generic sine and cosine-phased BOC signals. Theoretical and simulation results show that the proposed method removes the ambiguities in the acquisition problem, without requiring any auxiliary signal in the receiver.**

***Index Terms*— Binary offset carrier (BOC) modulation, unambiguous acquisition, detection performance, global navigation satellite systems (GNSS).**

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I. INTRODUCTION

The binary offset carrier (BOC) modulation has been adopted in the new global navigation satellite systems (GNSSs) such as the (next European) Galileo and modernized-global positioning system (M-GPS) [1]-[2]. The BOC modulation is a square sub-carrier modulation, where the signal is multiplied by a rectangular sub-carrier (sine or cosine-phased). Both the sub-carrier frequency f_s and the pseudo-random noise (PRN) chipping rate f_c are typically an integer multiple of the reference frequency f_0=1.023 MHz. BOC signals are denoted as BOC (f_s, f_c) or BOC(m, n) where the parameter m stands for the ratio between f_s and the reference frequency f_0, while n stands for the ratio between the code rate and f_0. The main idea behind BOC modulations is to reduce the interference with the binary phase shift keying (BPSK)-modulated signal, which has a *sinc* function shaped spectrum, and it is used in existing GPS signals. BPSK-modulated signals, such as coarse acquisition (C/A) GPS codes, have most of their spectral energy concentrated around the carrier frequency [3]. On the other hand, BOC-modulated signals have low energy around the carrier frequency and two main spectral lobes further away from the carrier (thus, the name of *split-spectrum*, often used to indicate BOC modulation).

The main disadvantage of BOC-modulated signals is that their autocorrelation function (ACF) has a profile with more than one peak. In particular, if the BOC modulated sequence has $N_{BOC} = 2m / n$ elements, the number of positive and negative peaks in the ACF equals $(2 N_{BOC} - 1)$ and the peaks are separated in delay by the chip interval $T_c = 1/(2 \cdot f_s)$ [1]. This is a great drawback: in fact, multiple peaks allow lock on secondary peaks (not resolvable in the tracking stage as usually implemented) with a non - zero probability [4]. New unambiguous acquisition algorithms, with increased receiver complexity, have been proposed in the open literature (see for example [5]-[11]). Most of them try to mitigate the secondary peaks combining the correlation with the local BOC modulated signal replica and correlations with proper ancillary signal(s) or waveform(s). It has to be noted that the performance of the tracking stage can be degraded by destroying the sharpness of the main-peak of the BOC autocorrelation function during the acquisition phase. Some acquisition methods have been recently proposed to maintain the sharp main-peak of the BOC autocorrelation function [5], [12]. In particular, the authors in [12] propose a method to remove the side-peaks of the BOC autocorrelation completely, while keeping the sharp shape of the main-peak. Since the focus of this work is on code acquisition, a complete discussion about the tracking phase is out of the scope of this paper. However, in the case of our interest, efficient tracking methods, such as the one proposed in [13], can be used in cascade to our proposed technique to obtain a performance enhancement also in the tracking stage. Moreover, such a system can also take advantages of the low complexity required in the acquisition stage by our method. Hence, in the following of this paper we focus only on low-complexity acquisition methods. Recently, three low-complexity unambiguous techniques have been introduced by the authors in [11]. In particular, we are here referring to the *modified Betz and Fishman* (mB&F) method of [11], since it is, among these three new approaches, the one characterized by the best performances. In particular, the mB&F method of [11] is a reduced-complexity side-band (SB) method that, using SB correlation channels, removes the ambiguities of the ACF of the split-spectrum signals. This method will be referred throughout the paper as modified side-band (MSB) method.

In this work, we focus our attention on specific BOC-modulated signals that have been selected for the next generation of GNSSs [10]: • sinBOC(10, 5) used in GPS M-code; • sinBOC(1, 1) as the building block for the open service (OS) structure in the Galileo system. These signals are here used as a *proof of concept* of the proposed method, which can be applied to both generic sine- and cosine-phased BOC signals. We propose an effective unambiguous acquisition technique for BOC-modulated signals. In particular, we remove the ambiguities of the ACF of the split-spectrum signals, exploiting a reduced-complexity (real and symmetric) filter composed of only 7 non-zero samples. In this way, we perform the cancellation of the correlation side-peaks and, at the same time, we remove the unwanted replicas of the signal spectrum before proceeding towards the detection stage. The remainder of this work is organized as follows. Section II highlights both the signal model and the acquisition ambiguity problem that form the basis of the proposed innovative unambiguous technique, whose implementation and theoretical rationale are described in Section III. Numerical results (by theory and simulations) as well as comparisons with the modified SB technique are outlined in Section IV, and finally, our conclusions are depicted in Section V.

II. BOC Signal Model and the Ambiguity Problem

A. Signal Model

BOC signals with sine- and cosine-phased offset carrier sent over single-path fading AWGN channels can be respectively expressed as [14]:

$$\begin{aligned} BOCs(t) &= A * u(t-\tau)c_{Ts}(t-\tau)a(t-\tau)\cos(w_{IF}t+\phi) \\ c_{Ts}(t) &= sign(\sin(2\pi f_s t)) \end{aligned} \tag{1}$$

$$\begin{aligned} BOCc(t) &= A * u(t-\tau)c_{Ts}(t-\tau)a(t-\tau)\cos(w_{IF}t+\phi) \\ c_{Ts}(t) &= sign(\cos(2\pi f_s t)) \end{aligned} \tag{2}$$

where *u(t)* is the PRN sequence, *a(t)* is the navigation message, $c_{Ts}(t)$ is the modulated subcarrier and *A*, w_{IF}, τ, ϕ, *n* are amplitude, intermediate frequency, time delay, phase, and noise components, respectively. In particular, sinBOC(1, 1) encodes, in the digital domain, a '+1' as a '+1 −1' sequence, while a '0' as a '−1 +1' sequence. Generalizing, for an arbitrary *NBOC* modulation order, in the sinBOC(*m*, *n*) case, a '+1' is encoded as an alternating sequence of '+1 −1 +1 −1 +1 ...', having *NBOC* elements, and a '0' (or '−1') is encoded as an alternating '−1 +1 ...' sequence, also having *NBOC* elements. Conversely, a cosBOC(1, 1) encodes, in the digital domain, a '+1' as a '+1 −1 −1 +1' sequence , and a '0' (or '−1') as a '−1 +1 +1 −1' sequence.

B.Acquisition and Ambiguity Problem

Conventional testing methods for the presence of a synchronization code signal (with a given spreading code offset) rely on power detection [15]. In the presence of additive white Gaussian noise (AWGN), it is well known that the optimum receiver performs a matched filtering operations followed by active PN correlator (or chip matched filter). Then, the correlator output is passed through the non-coherent detector and each sample output is then compared with the given threshold. Depending on the comparator output, the receiver either accepts corresponding code phase or continues to search. The received signal *r(t)* at the front-end acquisition receiver, after a multipath fading channel with AWGN, can be written as follows [16], [17]:

$$r(t) = \sqrt{P_r} e^{+j2\pi f_D t} \sum_{l=1}^{L} \alpha_l \sum_{n=-\infty}^{+\infty} b_n s_n(t - t_l) + n(t) \qquad (3)$$

where P_r accounts here for both the energy per bit and the channel gain, f_D is the Doppler shift, b_n are the data symbols while $s_n(t)$ is the (sine or cosine) BOC waveform expressed by (1) or (2) respectively, and $n(t)$ is the additive white noise. The multipath fading channel has *L* paths, each characterized by a complex path coefficient α_l and by a path time delay t_l.

The cross-correlation function $R(\tau)$ between *r(t)* and the local reference code *c(t)* is first estimated. Then, an *oversampling* factor N_s (i.e. the number of samples per BOC sub-chip) is used. In fact, practical acquisition and signal processing requires more than one sample per sub-chip interval; hence, using only one sample could result in a significant performance loss [14]. The sequential approach tests each detector's outputs and, once the maximum correlation result is larger than a threshold, true acquisition is declared. Since the side-peaks of the BOC ACF are characterized by a large amount of signal energy, under the influence of noise it is quite likely that one of side peak magnitudes exceeds the main peak, and false acquisition will happen [8]. As a consequence, the code tracking loop, performed after the acquisition stage, will initially lock on the side-peak (i.e. on the false acquisition) resulting in a dramatic degradation of the system performances. Therefore, avoiding the acquisition of side-peaks is more desirable than recover from false lock. In the next Section, we propose an innovative unambiguous acquisition technique for BOC-modulated signals.

III. UNAMBIGUOUS BOC ACQUISITION TECHNIQUE

In all the following analysis, we consider the base-band signal model meaning that *the carrier frequency has been removed beforehand* [11]. Hence, after a block performing conventional BOC demodulation, such as the one used in [19]-[21], it is well known that in presence of additive Gaussian noise optimum receiver performs matched filtering operations followed by active PN code correlator (or chip matched filter) [22]. The block scheme of our method is sketched in Fig. 1. Task of the proposed procedure is to completely remove the ambiguities of the ACF of the split-spectrum signals. To solve the ambiguity problem, we need to remove both the correlation side-peaks and the unwanted replicas of the signal spectrum before proceeding towards the detection stage. More in details, the rationale behind the proposed algorithm is as follows. Referring to Fig. 1, the output of the matched filter, i.e. the complex envelope of the BOC ACF, is now sampled with a sampling interval T_s/N_s with an over-sampling factor N_s (where T_s is the sample period [25]). Then, the squared magnitude of the estimated cross-correlation is evaluated. It has to be noted that the complex envelope of the BOC (f_s, f_c) ACF has a power spectral density (PSD) presenting the two principal spectrum lobes centered around the sub-carrier frequency $\pm f_s$. In practice, we can think of this signal as the output of a conventional double side-band amplitude modulator (DSB-AM). Hence, before removing the ambiguities, we need to first shift the main lobe of the BOC PSD to zero frequency, i.e. we need to demodulate. The squaring operation aims at performing this goal, acting as a typical DSB-AM non-coherent demodulator. In particular, it acts as a digital product detector [23]. In fact, squaring the complex envelope of the BOC ACF results in obtaining both the main spectrum lobe centered at zero frequency (signal demodulation) and two more replicas of the main lobe, centered at $\pm 2f_s$. Now the BOC ACF side-peaks can be removed, and then high-frequency components in the PSD can be filtered out. Finally, it has to be underlined that the output of the squaring operation is the vector signal with real elements in the time domain. This means that the following filtering operations can be performed (in the time domain) at a reduced complexity, avoiding complex sums and multiplications.

A.Side-Peaks cancellation

In order to eliminate the multiple peaks of the ACF of BOC-modulated signals, we exploit a reduced-complexity finite impulse response (FIR) filter. As illustrated in Fig. 2, the impulse response of this filter is composed by only 3 non-zero samples, to reduce its computational complexity as well as not to rely on the ACF tails that are too far from the maximum of the function. In particular, the impulse response of the proposed filter is as follows.

$$h_1(t) = \frac{1}{2}\delta(t+\tau^*) + \delta(t) + \frac{1}{2}\delta(t-\tau^*) \qquad (4)$$

where $\delta(t)$ is the continuous Dirac pulse. The value of the parameter τ^* must be chosen according to the particular BOC function under investigation and, hence, to the current analyzed BOC order, as it will be shown in the following Section. This finite impulse response corresponds, in the frequency domain, to the following transfer function:

$$H_1(\omega) = 1 + \cos\omega\cdot\tau^* \qquad (5)$$

that is a raised cosine function with roll-off = 1. The result of the convolution with the (square modulus of the) BOC ACF is clearly visible in Fig. 2 and can be written as follows:

$$\left|R_{un}\left(\tau\right)\right|^2 = \left|R\left(\tau\right)\right|^2 + \frac{1}{2}\cdot\left|R\left(\tau-\tau^*\right)\right|^2 + \frac{1}{2}\cdot\left|R\left(\tau+\tau^*\right)\right|^2 \quad (6)$$

where $R_{un}(\tau)$ is the unambiguous ACF and $R(\tau)$ is the ambiguous ACF . We have 3 ACFs, modulated by the filter's coefficients that now must be summed together to obtain the final output. The result of the convolution process is shown in Fig. 2 where the cancellation of the side-peaks is evident. The filter has been designed in order to have the 3 non-zero samples centered in $-\tau^*$, 0, and τ^*. In particular, this reduced-complexity raised cosine filter must properly filter the main lobe of the BOC PSD. This means that the value of the parameter τ^* must be equal to $T_s/2$.

B.Unwanted replica filtering

Then, as said before, the squaring operation produces two unwanted replicas of the BOC spectrum. In order to eliminate these spectrum replicas, centered in $\pm 2f_s$, we have to implement another reduced-complexity raised cosine filter with now two zeros in $\pm 2f_s$. This means that in the time domain we can use again a raised cosine filter (and again with roll-off $\gamma = 1$) composed by 3 non-zero samples, as before, whose impulse response is as follows:

$$h_2(t) = \frac{1}{2}\delta(t+2\cdot\tau^*) + \delta(t) + \frac{1}{2}\delta(t-2\cdot\tau^*) \quad (7)$$

The only difference with the first filter is that now the three taps are centered in 0, and $\pm T_s$ (i.e. $2\cdot\tau^*$), respectively The transfer function of this second filter can be now written as:

$$H_2(\omega) = 1 + \cos 2\cdot\omega\cdot\tau^* \quad (8)$$

that is, again, a raised cosine function with roll-off = 1. Finally, we can design a unique filter, whose impulse response is the convolution between the impulse responses of previously defined raised cosine filter, obtaining a reduced-complexity triangular filter with 7 non-zero samples, as shown in Fig. 3. In particular, the impulse response of the triangular filter can be expressed as follows:

$$\begin{aligned} h_{filt}(t) &= h_1(t)\otimes h_2(t) = \\ &= \frac{1}{4}\left[\delta(t+3\cdot\tau^*)+\delta(t-3\cdot\tau^*)\right] + \frac{1}{2}\left[\delta(t+2\cdot\tau^*)+\delta(t-2\cdot\tau^*)\right] + \frac{3}{4}\left[\delta(t+\tau^*)+\delta(t-\tau^*)\right] + \delta(t) \end{aligned} \quad (9)$$

and its transfer function is expressed as:

$$H_{filt}(\omega)=H_1(\omega)\cdot H_2(\omega)=\left(1+\cos\omega\cdot\tau^*\right)\cdot\left(1+\cos 2\cdot\omega\cdot\tau^*\right)= \\ =\frac{1}{2}\cos 3\cdot\omega\cdot\tau^*+\cos 2\cdot\omega\cdot\tau^*+\frac{3}{2}\cos\omega\cdot\tau^*+1 \tag{10}$$

It has to be noted that the convolution with the finite impulse response of the unique proposed filter, eq. (9), do not imply complex signals. Hence, the proposed technique can be defined a fast method, meaning that the computational complexity of the overall process can be kept at minimum (see Section IV) and its implementation can be straightforwardly optimized by exploiting parallel processors.

C. Detection stage

The power detector simply accumulates W blocks of the square magnitude of the estimated cross-correlation, and then compares the k-th currently examined decision variable $Z_k(\tau)$ to a pre-selected threshold:

$$Z_k(\tau)=\frac{1}{W}\sum_{w=1}^{W}\left|R_{filt}^{w}(\tau)\right|^2 \tag{11}$$

where $R_{filt}^{w}(\tau)$ is the BOC ACF (filtered by eq (9), i.e. the unambiguous function without spectrum replicas), of the w-th block. The constant false alarm rate (CFAR) procedure is often employed to perform effective tests. The CFAR test is accomplished in two successive parts: first, a threshold is determined to limit the false-alarm probability (P_{FA}), at a given chosen value under the H_0 hypothesis (i.e. absence of the tested code with the considered code-offset); second, the detection probability (P_D) is evaluated under the hypothesis H_1 (i.e. presence of the tested code with the correct code-offset) for the previously determined threshold. The test threshold can be asymptotically tuned from a straightforward evaluation of the Gaussian integral for a fixed P_{FA}, under the H_0 hypothesis, as follows [18]:

$$\nu=E\left[Z_k(\tau)\big|_{H_0}\right]+\left(2\cdot\mathrm{var}\left[Z_k(\tau)\big|_{H_0}\right]\right)^{1/2}\cdot erf^{-1}\left(1-2\cdot P_{FA}\right) \tag{12}$$

Furthermore, the theoretical (i.e. asymptotic) probability of detection (for the above threshold) can be similarly determined in the H_1 hypothesis by means of the expression [16]:

$$P_D = \frac{1}{2} + \frac{1}{2} \cdot erf\left[\left(-\nu + E\left[Z_k(\tau)\big|_{H_1} \right] \right) \cdot \left(2 \cdot \mathrm{var}\left[Z_k(\tau)\big|_{H_1} \right] \right)^{-1/2} \right] \quad (13)$$

IV. NUMERICAL RESULTS

Several simulation trials were performed to validate the innovative acquisition procedure, proposed in Section III versus the MSB approach of [11]. Since we are interested in verifying the applicability of the devised method to remove the ambiguities threats of the BOC-modulated signal, in all the following results we have considered a simple (single-path) channel model. Hence, eq. (3) can now be re-written as follows [14]:

$$r(t) = \sqrt{P_r} e^{+j2\pi f_D t} \sum_{n=-\infty}^{+\infty} b_n s_n (t - t_1) + n(t) \quad (14)$$

where t_1 is the time delay of the single-path channel, and $s_n(t)$ is the (sine or cosine) BOC waveform as in (3). The P_D of the new unambiguous method is evaluated under the CFAR procedure versus Carrier-to-Noise ratio (CNR) of practical interest. In the reported results, four samples per BOC chip (i.e. N_s = 4), a $P_{FA}=10^{-3}$ and white Gaussian noise have been used. We have assumed that the Doppler-frequency information exists at the receiver's side, i.e. there is GPS assistance (see [14]). Several Monte–Carlo simulation trials (10^6 independent runs) have been implemented to numerically evaluate the detection probabilities of the two methods, while the theoretical results have been obtained using (12)-(13). Fig. 4 shows the performance of the two methods for a cosBOC(1,1) and a cosBOC(10,5) modulated signal, while Fig. 5 illustrates the cases of sinBOC(1,1) and sinBOC(10,5). As it can be easily seen, for both the cases of a cosBOC(1,1) and a sinBOC (1,1) modulated signal, the proposed unambiguous technique outperforms the existing one and the simulation results perfectly match the theoretical ones, confirming the validity of the theoretical assumptions of Section III. Referring to the case of cosine and sine-phased BOC(10,5) modulated signals, it is evident that, again, the new method is capable of removing the ambiguity threat in the acquisition process. In particular, the curves referring to the two methods almost overlaps, meaning that the performance gain of the proposed method now reduces (due to the numerous side-peaks of this kind of BOC-modulated signal).

Then, we have also analyzed the case of multi-path propagation, considering the same six-path Rayleigh-faded channel model of [24] with exponential path decay. For the sake of compactness, we refer here only to the sine BOC modulated signals, since we have verified that the cosine BOC modulated signals suffer the same performance degradation. In particular, Fig. 6 shows the probability of detection of the sinBOC(1,1) and sinBOC (10,5) modulated signal in presence of multi-path for both the two analyzed methods. It can be easily seen from the graphs that in this operating scenario, the system performance are greatly affected by the multipath fading channel. Once again, the performance of the two techniques almost overlaps, thus confirming the validity of our approach for code acquisition of BOC signals. Moreover, multipath mitigation is expected to be taken care of in the tracking stage (see for example [16]) while in here we focus on low-complexity acquisition. Also, we remind that the proposed technique is a fast method: in fact, it performs the ambiguities removal process exploiting a real and symmetric filter of only 7 non-zero samples. It must be kept in mind that the proposed technique is a fast method: in fact, it performs the ambiguities removal process exploiting a real and symmetric filter of only 7 non-zero samples. The number of required sums and products is extremely reduced even compared with the low-complexity MSB method of [11], where complex multiplications and sums are used. In particular and as stated in [11], the complexity of the acquisition structure depends on the number of needed filters and required operations (sums and products) to be realized.

For the filtering part, it has to be underlined that the MSB method needs 8 real filters in the dual sideband approach and 4 real filters in the single sideband approach. Conversely, since in our method we apply the filtering directly on the non-coherent correlation function, only 1 real filter is needed. This means a drastically reduction in terms of require computational complexity. Then, let us now focus on the computational complexity of the proposed technique, by analyzing the total number of performed operations. According to the scheme of Fig. 1, the number of arithmetic operations to be realized by the receiver processor are: -first, evaluate the square modulus of the received signal (with N_s samples per

BOC sub-chip); -then, perform the convolution with the 7 non-zero taps filter. The first operation involves complex multiplications, whereas the filtering phase is only between real signals (the squared modulus of the signal and the impulse response of the real filter). The whole computation can be implemented either by sequential or parallel processing, depending on the available hardware. In agreement with the results of [11], we focus here only on the sequential implementation and we assume BOC signals of spreading factor S_F and N_s samples per BOC sub-chip. The approximate computing time needed by the entire procedure, operating on sequential processors in the receiver device, is reported below in terms of the time required by the number of arithmetic operations:

Performed operation	No. of Real Sums	No. of Real Products
• Squared modulus	$2 \bullet S_F \bullet N_s$	$4 \bullet S_F \bullet N_s$
• Filtering phase	$7 \bullet S_F \bullet N_s$	$7 \bullet S_F \bullet N_s$

For the sake of compactness, we have reported only the number of real products and sums assuming that, in sequential implementation, the time of one complex product is the same as four real products and two real sums, whereas the time of one complex sum time is equivalent to two real sums. Notwithstanding our proposed approach denotes the same performances of the MSB method of [11], it requires a small number of operations to be realized, as shown in Tab. 1. In fact, considering a sinBoc(1,1) signal with S_F = 4092 and N_s =2, the MSB method requires more than 10^8 total number of arithmetic operations, whereas it requires more than 10^{10} operations with S_F = 4092, N_s =16. Conversely, our fast method only needs a total number of arithmetic operations equal to $1.6368 \cdot 10^5$, with S_F = 4092 and N_s =2, while it requires $1.306 \cdot 10^6$ operations in the second case, that represents a remarkable reduction of the overall system computational complexity.

V. CONCLUSION

A novel unambiguous acquisition technique applicable to both generic sine and cosine-phased BOC signals has been here proposed. The proposed method removes the ambiguity threats (side-peaks) of the BOC autocorrelation, exploiting a reduced-complexity (real and symmetric) filter composed of only 7 non-zero samples. Theoretical and simulation results show the validity of our method in improving the system acquisition performances, with no increasing of the system computation complexity.

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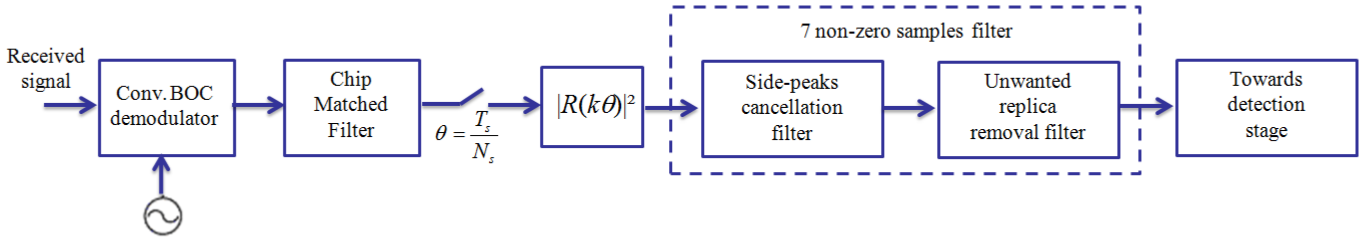

Fig. 1. Block scheme of the proposed unambiguous procedure

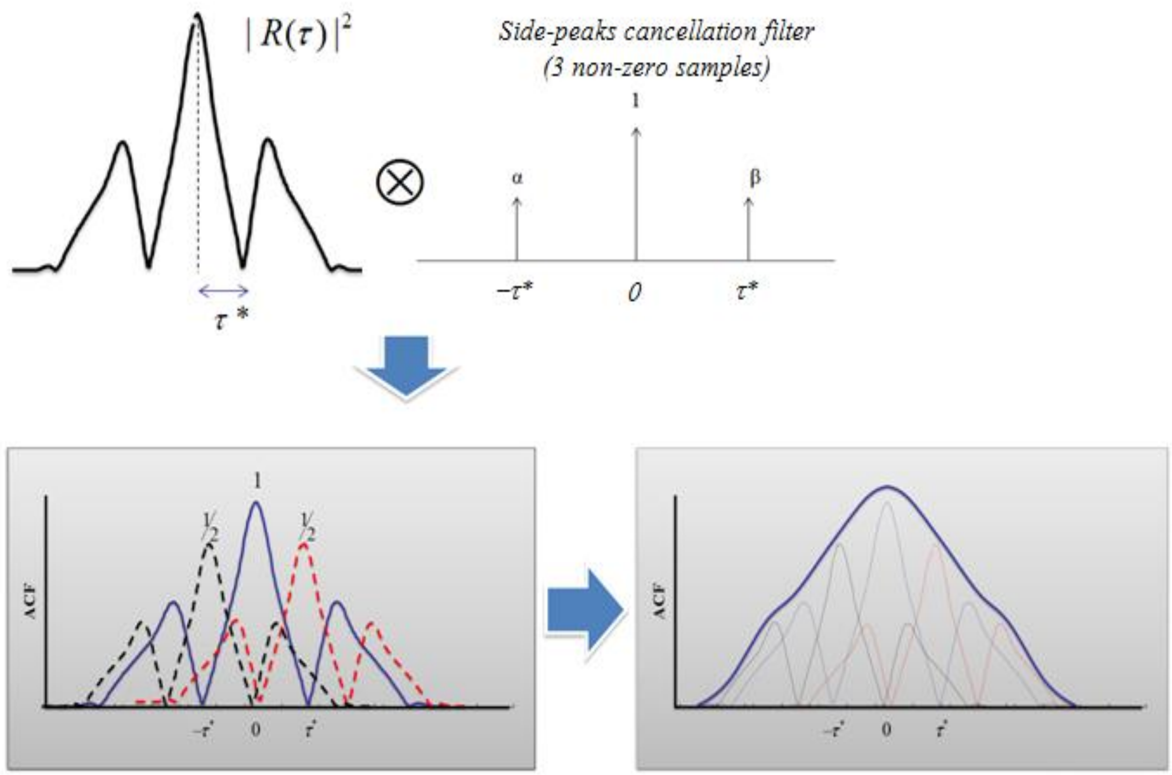

Fig. 2. Scheme of the side-peaks cancellation (ambiguities removal) procedure by means of the 3 non-zero samples filter.

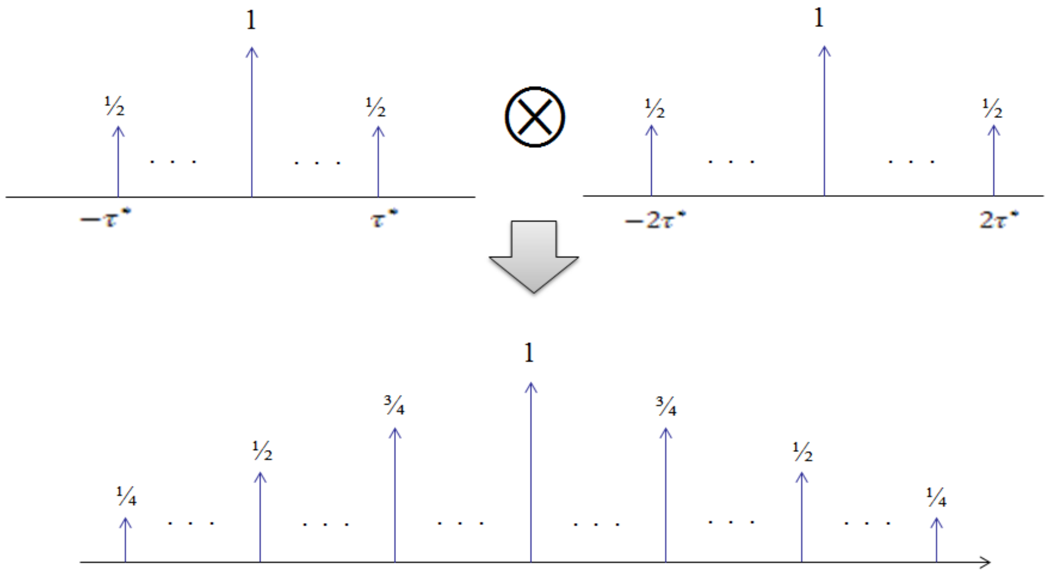

Fig. 3. Reduced-complexity 7 non-zero taps triangular filter

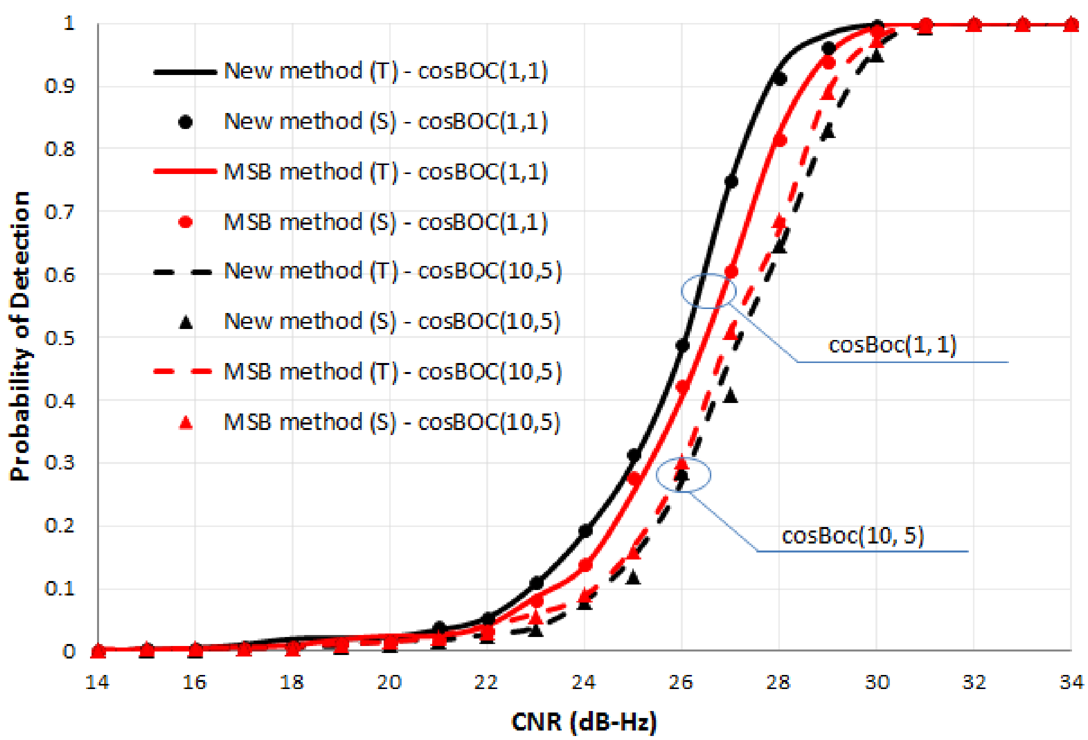

Fig. 4. Theoretical (T) and simulated (S) results for the PD the new and modified SB (MSB) method for cosBOC(1, 1) and cosBOC(10, 5) modulated signals.

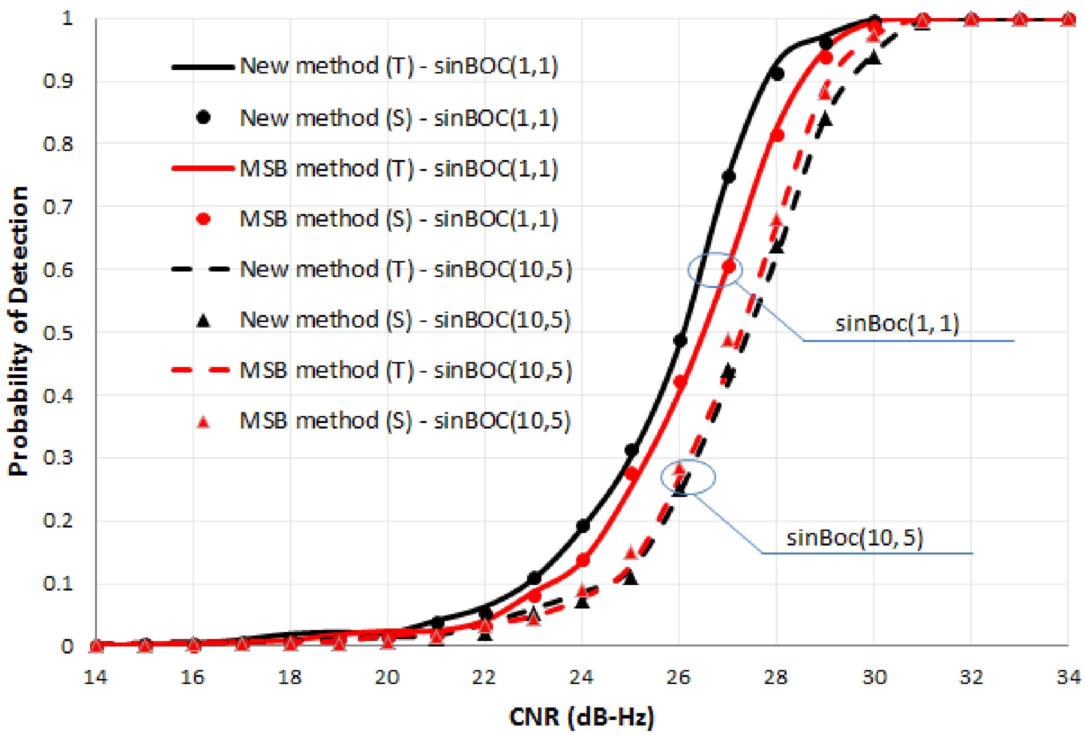

Fig. 5. Theoretical (T) and simulated (S) results for the PD the new and modified SB (MSB) method for sinBOC(1, 1) and sinBOC(10, 5) modulated signals.

Tab. 1. Total number of arithmetic operations required by both the conventional and new methods

Signal Parameters	S_F = 4092; N_S = 2	S_F = 4092; N_S = 16
# operations Conv. Method	$\approx 10^8$	$\approx 10^{10}$
# operations New Method	$\approx 1.6 \cdot 10^5$	$\approx 1.3 \cdot 10^6$
Gain (%)	≈99.84 %	≈99.98 %

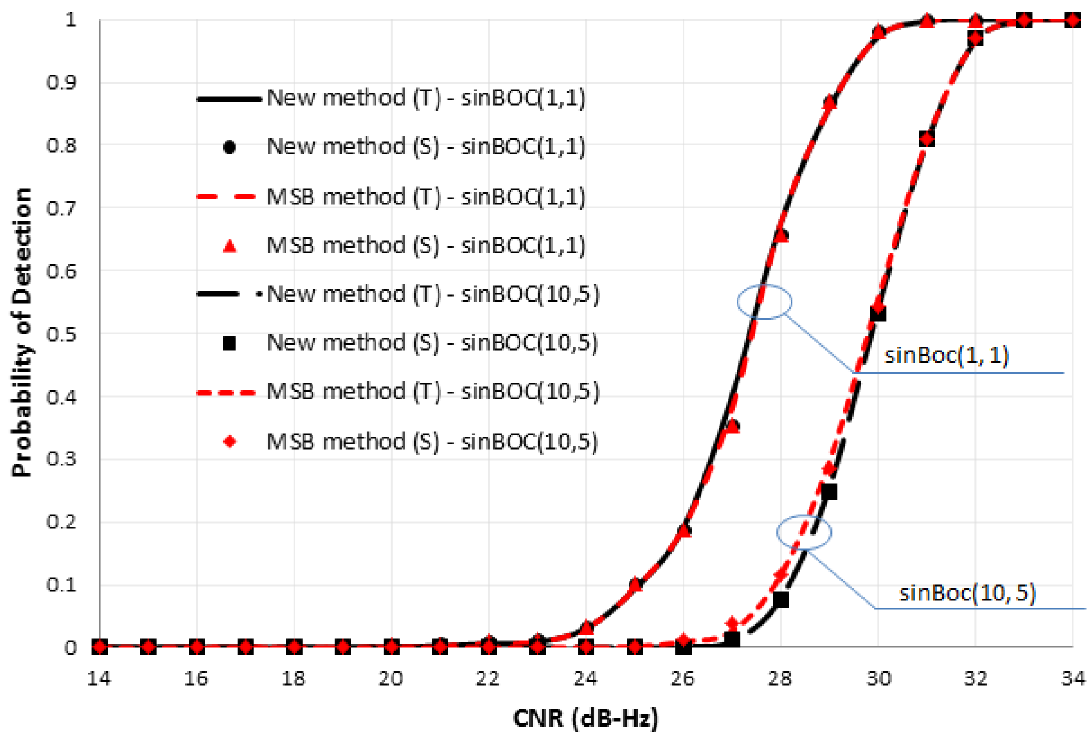

Fig. 6. Theoretical (T) and simulated (S) results for the PD the new and modified SB (MSB) method for sinBOC(1,1) and sinBOC(10,5) modulated signals in presence of a 6-path Rayleigh-faded channel.