Complexity Considerations for Unambiguous Acquisition of Galileo Signals

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Abstract - In order to obtain a higher spectral separation from the BPSK-like signals, i.e. GPS C/A code, the signals proposed for future Galileo and GPS M-code signals are processed using split-spectrum type modulations, such as Binary Offset Carrier (BOC) modulation. These BOC modulations create deep fades (ambiguities) in the envelope of the Autocorrelation Function (ACF) of signal, and therefore the acquisition and tracking of these signals pose new challenges. To overcome these problems, two approaches have been recently proposed in literature, referred either as "sideband techniques" (Betz, Fishman & al.) or "BPSK-like" techniques (Martin, Heiries & al.). These methods allow the use of a higher search step in time domain, but employ an modified reference PRN code at receiver, which lead to an increase in implementation complexity. Moreover, the BPSK-like method does not work for odd BOC modulation orders. In this paper we present an extension of BPSK-like method, which provides a significantly lower complexity in the correlation part, and it works for both even and odd, sine and cosine BOC modulation orders. This technique is compared with the existing sideband methods in terms of performance and implementation complexity. As a benchmark, we also keep the ambiguous BOC processing. For a further decrease in implementation complexity, we investigate the effect of different IIR and FIR filtering structures used for the side-band selection in the receiver. We use here an interpolated FIR filter structure which provides a lower computational complexity than a direct form FIR filter and has similar performance with the others filters. The analysis is done in the presence of realistic multipath fading channels and the signals are modeled according to the current proposals for Galileo system Open Service (OS).

1 Background and motivation

Future space-borne navigation receivers will operate with both GPS and Galileo navigation signals [1]. In order to provide a better spectral separation with existing GPS C/A code signals, the sine and cosine Binary Offset Carrier (BOC) modulations have been selected, for both Galileo and modernized GPS [2],[3].

A BOC(m,n) modulated signal is created by a square carrier modulation, where the signal is multiplied by a rectangular sub-carrier at sub-carrier frequency. The modulation parameters satisfy the relationships $m = \frac{f_{sc}}{f_{ref}}$ and $n = \frac{f_c}{f_{ref}}$ where f_{ref}=1.023 MHz is the reference frequency, f_{sc} is the sub-carrier frequency and f_c is the code rate. Thus the BOC modulation splits the signal spectrum into two symmetrical components, around the carrier frequency $f_{carrier}$. A generic characterization of the signal at baseband is given by the BOC modulation order $N_{BOC} = \frac{2f_{sc}}{f_c}$. While allowing a better usage of available bandwidth for different GNSS signals, the BOC modulation brings new challenges into the acquisition process, due to the deep fades (ambiguities) that appear into the ±1 chips interval around the maximum peak of ACF envelope. These ambiguities lead to an increased number of timing hypotheses in order to detect correctly the signal, thus the necessary step to search a given time-uncertainty window Δt_{bin} should be small enough in order to find the main lobe of ACF. Therefore the computational complexity of the acquisition process is increased, the computational load being inversely proportional with the step time bin Δt_{bin} [6]. Additionally, the tracking of BOC signals has to cope with more false lock points than traditional BPSK modulated signals. On the other hand, tracking accuracy increases if BOC modulation is used instead of BPSK modulation, because of the decrease in the width of the main lobe of the envelope of the ACF.

In order to eliminate these ambiguities in the ACF envelope and thus to enable the use of a higher step in the acquisition process, two different methods have been proposed in literature so far. These methods are referred either as "BPSK-like techniques", introduced by Martin, Heiries & al. [4],[5] (denoted here by **M**&**H**) or as "sideband techniques" (referred here as **B**&**F**), which have been proposed by Betz, Fishman & al. [6],[7],[8] and analyzed also in [9]. We remark that these methods can be also used in tracking, but the main focus here is the acquisition part. The main idea behind these techniques is that the BOC modulated signal can be approximated as a superposition of two shifted BPSK-modulated. The effect of sub-carrier modulation can be eliminated by using a pair of single sideband correlators. We may have a single sideband (SSB) or a dual sideband (DSB) receiver, depending if only one of sideband correlators is used (either positive or negative), or both sideband correlators are used and combined non-coherently. In the referred literature both these methods have been tested only with even BOC modulation orders and both use a modified reference PRN code at the receiver, which leads to a higher complexity. We have shown in [10] that the M&H method fails to work for odd BOC modulation orders and we have also introduced a new modified approach, which provides a lower implementation complexity and works for both even and odd BOC orders.

The complexity of the acquisition depends on both the correlation and the filtering (sideband selection) parts. In the existing literature which treats these unambiguous methods, the sideband selection is usually performed assuming ideal filtering and the effects of real filtering are ignored [4][6][7]. However, since the distortion introduced to the useful signal by sideband filtering in a true receiver may affect the performance, this point is worth being investigated further. In [5] the filtering effects on the M&H method were studied using Butterworth filters, which induce a translation on correlation function due to the time group delay. In [11] we have considered the design of digital bandlimiting receiver filters for BOC-modulated and oversampled Galileo signals, for different filtering structures.

In this paper, we improve further the architecture of the proposed modified approach introduced in [10], by reducing the number of filters used for band selection. The implementation complexity of the unambiguous methods is investigated, considering both the correlation and the sidebands filtering parts. We study the effect of several IIR and FIR filtering structures and we study a interpolated FIR filter, which provides a lower complexity, at the similar performance with the other filtering methods.

In Section II the unambiguous acquisition methods are presented. The signal model and the used filtering structures are described in Section III. The next section compares the implementation complexities and the performances of the studied unambiguous techniques. The conclusion are drawn in Section V.

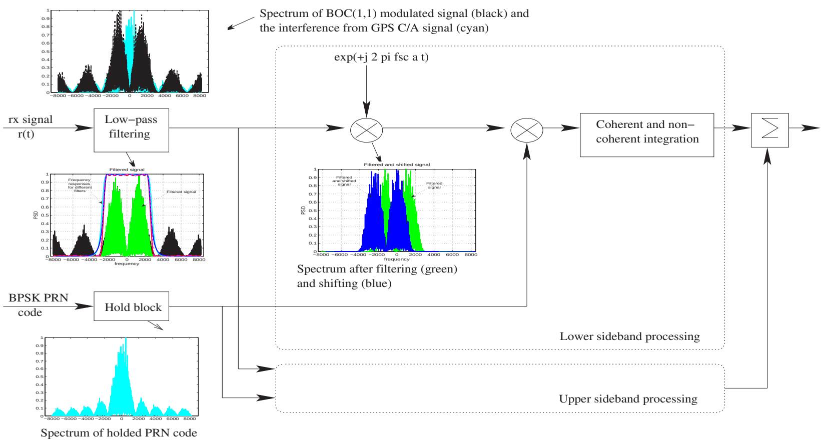

Figure 1: Block diagram of unambiguous proposed method (modified Martin & Heiries), dual sideband processing, even BOC modulation order N_{BOC}=2.

2 Unambiguous acquisition methods

In the **B&F** method the receiver filters out, individually, the main lobes of the received signal and those of the reference code. The correlation function between the received signal and reference code, on each sideband, will be unambiguous and will resemble the ACF of a BPSK-modulated signal. However, the shape of resulting ACF is not exactly the one of a BPSK-modulated signal, since there are information losses due to selection of main lobes alone. In single sideband (SSB) method, only one of the bands (upper or lower) is used when forming the decision statistic and 3 sideband selection complex filters are needed, 2 for the in-phase and quadrature parts of the received signal, and one for the real code. However, the SSB approach suffers of high non-coherent integration losses [6]. This loss can be compensated if dual sideband (DSB) processing is used, when both sidebands are kept and combined non-coherently, but this leads to rather high complexity since 6 complex filters are needed for DSB processing [8].

In the **M&H** technique, instead of using two filters (one for every main lobe), only one filter is used, centered at middle of carrier frequency band $f_{carrier}$ (respectively, at zero frequency if the filtering is done at the baseband as here). This filter has a bandwidth which includes the two main lobes and the secondary lobes between them (if any). Only two real filters (for the in-phase and for the quadrature components) are needed for the received signal, for both the SSB and DSB methods. The reference code is the BPSK-modulated code, held at the sub-sample rate (with the hold factor $N_s N_{BOC}$, where N_s is the number of samples per BOC interval) and shifted by $\pm f_{sc}$ (i.e. $\pm \frac{N_{BOC}}{2} f_{carrier}$) to the sub-carrier frequency. Simulations have shown that this method fails to work for odd BOC modulation orders, if shifting is done with this factor [10].

A less complex implementation method, proposed by the authors [10], is shown in Fig. 1. This method is a modified approach of M&H technique presented in [4], [5]. After filtering out both the main lobes and everything between them (if any), the received signal is shifted towards the middle of frequency band, by multiplying it with the exponential $exp(\pm j2\pi f_{sc} a t)$. The shifting parameter a depends on the order of sine BOC modulation used, as explained in [10] and is given by eq. (1):

$$a = \begin{cases} 1 & \text{if } N_{BOC} \text{ even} \\ \frac{N_{BOC}-1}{N_{BOC}} & \text{if } N_{BOC} \text{ odd} \end{cases} \quad (1)$$

If the shifting is done with $a = 1$ for odd BOC modulation orders, we loose the peak after the correlation, and therefore, the M&H method becomes invalid. In contrast to M&H method, this shifting factor allows a proper acquisition for both even and odd BOC modulation orders.

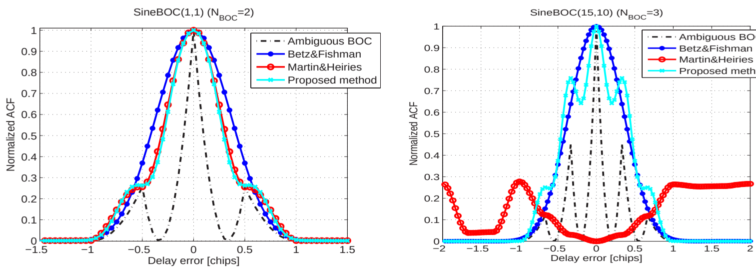

Figure 2: Illustration of the envelope of the correlation functions: **Left plot**, even BOC order, N_{BOC}=2, DSB processing, **Right plot**, odd BOC order, N_{BOC}=3, SSB processing. The ambiguous BOC shape is also shown as a reference.

The differences to the original M&H method is that the shift is applied to the received signal and not to the code, thus the reference PRN code is a BPSK modulated sequence, held at sub-sample level. Here, unlike in [10], the shifting is applied to the received signal after filtering. Therefore only two real filters are sufficient, for both SSB and DSB processing. The correlation between received signal and reference code can be performed either in frequency domain or in time domain. If the correlation part is done in time-domain, the implementation complexity can be tremendously decreased, since the reference code is a sequence of $\pm$1s, and thus, the complex multiplication between the received signal and the reference code can be done via simple additions and sign inversions [12].

The shapes of the envelope of the ACF after the unambiguous processing are shown in Fig. 2, for both even and odd BOC modulation orders. In the left plot N_{BOC}=2 and DSB processing is assumed, while in the right plot N_{BOC}=3 and SSB processing is used. In case of an even BOC order, for the M&H and proposed methods, some small residual peaks can be observed, due to the fact that there is some interference from the neighbor bands into the main lobe of the signal spectrum (since they are not filtered out completely, as it is the case for the B&F method). In case of proposed technique, these residual peaks are slightly higher for odd N_{BOC} than for even N_{BOC}. Also, it can be observed that for an odd BOC modulation order, the original M&H method fails to work completely.

3 Signal model and filtering structures

The performance of the unambiguous techniques has been verified through extensive simulations, assuming a fading multipath channel model. At the transmitter, after spreading and BOC modulation, the data sequence is oversampled with an oversampling factor N_s, which represents the number of sub-samples per BOC sub-chip interval. We have considered here that the signal is modulated using sine BOC(1,1) modulation. In case of a BOC modulated signal, the zeros in the ACF may induce a missed detection due to a zero sampling. If a higher search step is used (i.e. Δt_{bin}=0.5 chip), this oversampling becomes necessary, in order to have sufficient resolution in the delay estimation process. The signal received via an L_{path} fading channel can be written as in eq. (2):

$$r(t) = \sqrt{E_b} e^{+j2\pi f_D t} \sum_{n=-\infty}^{n=+\infty} b_n \sum_{l=1}^{L} \alpha_{n,l}(t) s_n(t-\tau_l) + \eta(t) \quad (2)$$

where E_b is the bit energy of signal, b_n is the data bit corresponding to the n-th code epoch, $\alpha_{n,l}(t)$ is the time-varying complex fading coefficient of the l-th path during the n-th code epoch, τ_l is the corresponding path delay (rounded to integer multiples of the sampling interval T_s), f_D is the Doppler shift, and $\eta(\cdot)$ is the additive white noise added by the channel.

$$s_n(t) = \sum_{k=1}^{S_F} (-1)^{kN_{BOC}} c_{k,n} \sum_{m=0}^{N_{BOC}-1} (-1)^m p(t - nS_F N_{BOC} \times N_s T_s - kN_{BOC} N_s T_s - mN_s T_s) \quad (3)$$

The signal is correlated with a reference signal $s_{ref}(t, \widehat{\tau}, \widehat{f}_D, n_1)$, which can include either the PRN code and BOC modulation or only the PRN code. For the ambiguous BOC case, the reference signal is the BOC-modulated PRN code given in eq.(3), where $c_{k,n}$ is the k-th chip value corresponding to the n-th data symbol, S_F is the spreading factor or the code epoch length (e.g., for GPS C/A signal, $S_F = 1023$), $p(\cdot)$ is a train of rectangular pulses, and T_s is the sampling interval ($T_s = 1/(N_s f_c)$). The $(-1)^{kN_{BOC}}$ factor was added in order to take into account the situation of odd modulation orders.

In the low-complexity method, the reference signal is the BPSK-modulated PRN sequence, held at the same rate $f_c N_{BOC} N_s$ of the received signal. The main lobes of the received filtered signal are shifted down or up, around the center frequency (eq. 4):

$$r_{sh}(t) = r_{filt}(t) e^{\pm j2\pi f_{sc} a t} \quad (4)$$

Depending on the method, at the receiver part either only the main lobes of the incoming BOC signal are extracted or the whole bandwidth of signal containing the main lobes and the secondary lobes between them is selected. After correlation, the signal is coherently averaged over N_c ms (Fig.1) and next, non-coherently averaged over N_{nc} blocks. The maximum coherence integration length N_c is dictated by the coherence time of the channel and by the stability of the clock oscillator. The decision statistic is formed in a hybrid manner, by splitting the code-Doppler search space into several code-Doppler windows.

3.1 Traditional filtering structures

Since the complexity of implementation depends on both the correlation and the filtering parts, selecting a suitable filtering structure can lower even further the receiver complexity. We have considered here different digital filter realizations, such as FIR and IIR structures. So far, only the effect of IIR filtering for sideband selection has been studied in the context of unambiguous methods [5],[8]. The choice of most suitable filtering structure (IIR or FIR) depends on the specific performance criteria. If filter design and linearity of phase response are not so critical, an IIR structure is sufficient. However, there is a trade-off between IIR and FIR filters; for instance FIR filters can be designed to have exactly linear-phase response, are always stable, easy to implement in fixed point version and they require less datapath precision. On the other hand, the IIR filters have fewer computations, but they are conditionally stable, and the exact linear phase response is not possible. Also, they can be quite complex to be implemented in fixed-point version, typically require greater precision and may produce limit cycles. FIR filtering implementations typically require more multiplications and summations than IIR filters with similar performance, but since some computer architectures are better suited for FIR filtering, the computation speed of an IIR filter is not necessary faster than that of a FIR filter. For a FIR structure, due to symmetry of coefficients, some practical implementations allow for a linear-phase response using only half of the number of coefficients.

There are many attempts to develop improved filtering structures, for different kind of applications, in order to lower the computational complexity. These are the result of extensive research to find structures which are computationally efficient and insensitive to quantization error. For example, other IIR structures can be considered, such as wave digital filter structures, lattice-ladder, etc., which offer a lower number of multiplications and consists a topic of future research [13].

As parameters of the filter design process we have the filter's bandwidth B_T (in MHz), the passband and stopband edge frequencies, taken as $f_p{=}\frac{B_T}{2}$ MHz, respectively $f_s{=}f_p{+}tr_{width}$ MHz. The transition band tr_{width} is the boundary between the passband and stopband, which occurs around the filter's cutoff frequency. The width of this transition band is the filter's roll-off and, in general, a faster roll-off gives a sharper frequency response. However, for any FIR design algorithm, the filter order required to meet a given specification is inversely proportional to the transition width allowed, if the peak ripple remains the same [11]. In order to have a good trade-off between the filter's order and the selectivity of filter, we have selected, for the presented simulations, a transition width of $0.4 f_p$ MHz.

3.2 Interpolated (FIR) filters

Another class of digital filters considered here, which can implement lowpass FIR filters with significantly reduced computational workload, are the interpolated FIR (**IFIR**) filters [14]. The IFIR filters provide an efficient means of narrowband filtering, yield linear-phase response and can meet the given specification, with a reduced number of multipliers. The filter consists of a cascade of subfilters. Each stage has a periodic frequency response, with a different periodicity. The transfer function of the interpolated filter is (eq. 5):

$$H(z) = \prod_s H_s(z_s^L), \tag{5}$$

hence the periodicity of response of the s^{th} stage is $\frac{F_s}{L_s}$, where F_s is the sample rate. Different periodicities are obtained by replacing each unit delay in prototype filter by a delay of L_s samples, resulting in an "upsampled" impulse response. The stage with the shortest period (greatest L_s) shapes the transition band, while the images of its passband are attenuated by the other stages. The attribute "interpolated" refers to the time domain interpretation: adjacent impulse responses of a narrowband filter are strongly correlated. Therefore, the impulse response can be constructed by interpolating a sparse (i.e. upsampled) prototype impulse response with a lowpass filter. Notice that interpolated filters can be derived from multistage decimators (interpolators) by moving the filter stages to the input (output) sample rate using the well-known multirate identities.

The principle sets no constraints to the types (e.g. FIR or IIR) or implementation of the subfilters. Due to relaxed subfilter requirements and reducing the time-domain redundancies, IFIR filters possess significantly lower multiplication rates than corresponding direct-form single rate filters, and even lower

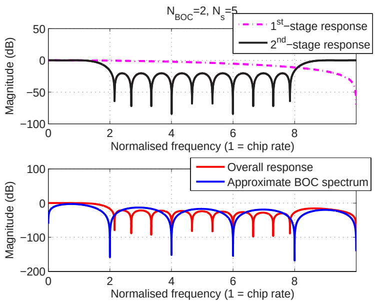

Figure 3: Frequency response of the two-stage interpolated FIR filter.

Table 1: Filters complexity (number of operations for 1 ms processing) per real filter for different filtering structures.

Operation	IIR direct form	Lattice IIR	Loss-less discrete integrators IIR	FIR single stage	Interpolated FIR S stages
N_{adds}	$2N_{IIR} \cdot N_{spc}$	$3N_{IIR} \cdot N_{spc}$	$2N_{IIR} \cdot N_{spc}$	$(N_{FIR} - 1) \cdot N_{spc}$	$\left(\sum_{s=1}^{S} N_{IFIR}(s) - 1\right) \cdot N_{spc}$
N_{muls}	$(2N_{IIR} + 1) \cdot N_{spc}$	$N_{IIR} \cdot N_{spc}$	$N_{IIR} \cdot N_{spc}$	$\lceil \frac{N_{FIR}}{2} \rceil \cdot N_{spc}$	$\sum_{s=1}^{S} \left\lceil \frac{N_{IFIR}(s)}{2} \right\rceil \cdot N_{spc}$

complexities can be reached by multirate filtering. We have implemented a two-stage (S=2) IFIR filter, with stage-wise decimation factors given by the BOC-modulation order N_{BOC}, respectively by the oversampling factor N_s. The filter type of the s^{th} stage is a direct form FIR filter, but it can be also a halfband FIR structure, in case of higher N_{BOC} and/or oversampling factors, when more than two stages can be used. The stage-wise and overall magnitude responses of the IFIR design, and also the approximate BOC spectrum, are shown in Fig. 3.

In order to compare the number of operations required by different methods, we have selected several typical FIR and IIR filtering structures, which are shown in Table 1. Here, we do not intend a exhaustive description of possible filtering structures, but rather some estimates about the expected complexity of the filtering part. The filters orders are denoted by N_{IIR}, N_{FIR} and N_{IFIR}. The term N_{spc}=$N_s N_{BOC} S_F$ represents the number of sub-samples per code epoch.

4 Overall complexity considerations and performance

At the complexity of acquisition part contribute both the filtering and the correlation parts. We consider here that correlation is performed in time-domain. If the reference code is a sequence of $\pm$1s, as in the modified M&H approach, a less complex correlation method can be employed [12]. This approach has no multiplications, just additions and sign inversions. The number of N_{adds} real additions per each frequency bin and for SSB processing can be computed as (eq. 6):

$$N_{adds} = 2\left(\left(N_\tau(S_F - 1) + N_s N_{BOC} - 1\right)\left(N_s N_{BOC} D_{max}/N_\tau - 1\right) + N_s N_{BOC} S_F - 1\right) \quad (6)$$

In the above equation, N_τ=$N_s N_{BOC} \Delta t_{bin}$ is the searching step in time, expressed in samples and D_{max} is the maximum delay search range in chips. For full search D_{max} is equal to the code length, but it can be much less if assisted acquisition is employed.

For B&F or M&H methods, since the correlation is done with a modified reference sequence (that requires a greater word-length), the direct, more computationally expensive approach must be used, in order to compute the correlations [12]. Therefore, we have the following number of real addition (eq.7) and multiplications (eq.8) [12]:

$$N_{adds,DirForm} = 2\left(3S_F N_s N_{BOC} - 1\right)\frac{N_s N_{BOC} D_{max}}{N_\tau} \tag{7}$$

$$N_{muls} = 4(S_F N_s N_{BOC})\frac{N_s N_{BOC} D_{max}}{N_\tau} \tag{8}$$

By comparing eqs.(6) and (7), for a step time bin of half of chip, it has been calculated that $N_{adds,DirForm}$ is approximately 6 times greater than N_{adds}. Thus, for the time-based correlation stage, in case of DSB processing, the B&F method needs about $12N_{adds}$ additions and $2N_{muls}$ multiplications. For SSB processing, the number of additions and multiplications are halved. For the M&H method, DSB processing, the necessary number of additions is $6N_{adds}+6N_{spc}$ and the number of multiplications is $2N_{muls}+4N_{spc}$. Here the term $N_{spc}=N_s N_{BOC} S_F$ is due to shifting of the received signal with the exponential $exp(\pm j2\pi f_{sc}at)$. This shifting is done at sample level, for both the in-phase and quadrature components of the received signal. In case of M&H SSB approach, there are $6N_{adds}+3N_{spc}$ additions and $N_{muls}+4N_{spc}$ multiplications.

In the low-complexity unambiguous DSB approach, the number of additions is decreased to $2N_{adds}+6N_{spc}$ and there are $4N_{spc}$ multiplications, due to the shifting stage. In the low-complexity method, SSB processing, there are $N_{adds}+3N_{spc}$ additions and $4N_{spc}$ multiplications. The term N_{spc} is much smaller than N_{adds} and N_{muls}, especially at high D_{max}.

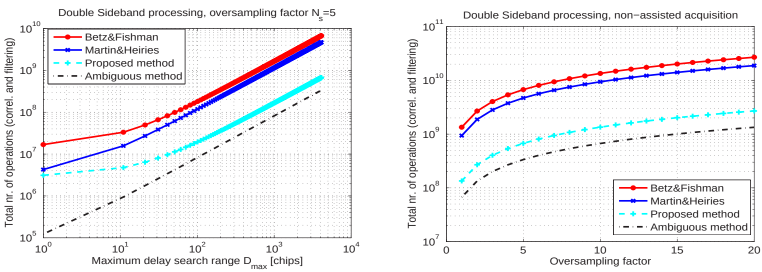

Figure 4: Comparison of total complexity (both the correlation and filtering processes), in terms of multiplications and additions, for unambiguous and ambiguous BOC processing, assuming Interpolated FIR filtering; **right plot** number of operations vs. delay search range, **left plot** number of operations vs. oversampling factor

Fig. 4 illustrates the total number of operations, for both correlation and filtering parts, assuming that an interpolated FIR filter was used for sidebands selection. The IFIR filter satisfies the following design requirement: the loss in passband is set to r_p=0.1 dB and the stopband attenuation to r_s=20 dB; the bandwidth is B_T=2.046 MHz, passband frequency is f_p=1.023 MHz and stopband frequency is f_s=1.4322 MHz. In order to satisfy these requirements, the following IFIR filter orders are necessary: $N_{IFIR}(1)$=5 at 1^{st} stage and $N_{IFIR}(2)$=15 at 2^{nd} stage. If for this particular design case, other filtering structures were employed, the required filter orders are N_{IIR}=7 (for IIR structure), respectively N_{FIR}=30 (for a FIR filter). Therefore, it can be concluded that the IFIR filter gives a similar number of operations as the IIR direct form and lattice structures, and a much lower number than the FIR filter.

As it can be observed from both plots of Fig. 4, the low-complexity proposed method brings a significant decrease in the computational complexity, when compared with the B&F and M&H methods and it is very close to the bound given by the ambiguous BOC processing. It is worth to mention that the main burden in the receiver complexity comes form the correlation part (i.e. in case of non-assisted acquisition and N_s=5, the number of operations for filtering part is about 4.9×10^6, while for the correlation part is about 670×10^6).

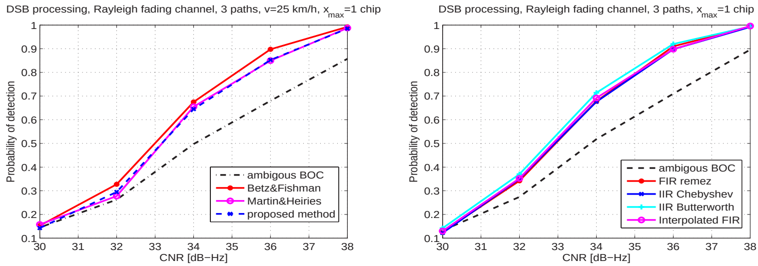

Figure 5: **Left plot**: Performance for different unambiguous methods, DSB processing, Rayleigh fading channel, 3 paths, sideband selection with IFIR filter, 2.046 MHz bandwidth, 0.0058 passband ripple and 0.1 stopband ripple; **Right plot**: Performance for different filtering structures used in sideband selection, DSB processing, SinBOC(1,1), Rayleigh channel, 3 paths, Δt_{bin}=0.5 chips, sideband selection filtering is realized with the proposed low-complexity method

Fig. 5 presents the simulation results, in terms of detection probabilities. The detection probability is the probability to have the decision variable higher than a decision threshold, provided that we are in the correct bin. In order to have a fair comparison between different decision variables, the false alarm probability P_{fa} was set to a fixed value 10^{-1} and the corresponding threshold was computed in order to meet this P_{fa}. The detection probability was computed based on the estimated threshold. The simulations were performed for a fading Rayleigh channel model, with 3-paths decaying power profile. The BOC modulation order was set to N_{BOC}=2, for all the presented simulations. The maximum separation x_{max} between successive paths is set to 1 chip and the delay spacing is assumed to be uniformly distributed. The mobile speed is 45 km/h, N_c=20, N_{nc}=2, N_s=5 and Δt_{bin}=0.5 chip. The maximum frequency incertitude range in searching space is 9 kHz. The test statistic was built over time-frequency windows of size 12 chips x 1 kHz, using hybrid search.

The performances of the different unambiguous methods are presented in Fig. 5, left plot, for a DSB processing. Here, the sideband selection filtering is realized with an interpolated 2-stages FIR filter, with passband ripple δ_p=0.0058, respectively δ_s=0.1 stopband ripple. For both M&H and the low-complexity methods, the bandwidth width is set to B_T=2.046 MHz, in order to accommodate both main lobes of the received signal spectrum, while for B&F method B_T=1.023 MHz is enough to select one lobe. As it can be observed, the B&F method gives the best performance, with about 0.5 dB improvement comparing to the M&H and the low-complexity methods, which give similar results. As expected, the BOC processing gives the poorer results.

Fig. 5, right plot, compares the performances of different IIR and FIR filtering structures, in the context of the low-complexity unambiguous method. We have considered here a FIR filter design by Parks-McClellan optimal equiripple algorithm, IIR Chebyshev type I and Butterworth filters design (simulated by Matlab) and also the two-stage interpolated FIR structure. For the IIR filters (Butterworth and Chebyshev type I) the loss in passband is set to r_p=0.1 dB and the stopband attenuation is r_s=20 dB. For the FIR and the interpolated FIR filters, the passband ripple is set to δ_p=0.0058, respectively the stopband ripple to δ_s=0.1. The bandwidth width is B_T=2.046 MHz, the passband edge frequency f_p=1.023 MHz, respectively the stopband edge frequency is f_s=1.4322 MHz.

As illustrated in Fig. 5, right plot, all filters give roughly the same performance for the low-complexity unambiguous method. This behavior is expected, since the shapes of frequency responses for different filters are quite similar (as can be observed from Fig. 1, in the plot which illustrates the spectrum of the signal after filtering).

5 Conclusions

In this paper we have considered the complexity implementations of different unambiguous BOC processing methods. These complexity issues are important in the context of application for Galileo and modernized GPS systems. We have compared the existing unambiguous BOC acquisition methods with our proposed approach, which can significantly lower the complexity coming from the correlation part.

We have also taken into account the complexity coming from sideband selection part, considering different FIR and IIR filtering structures, which have similar performance in terms of detection probability. An interpolated FIR structure can provide lower computational complexity than FIR single stage and almost similar complexity with the direct-form and lattice IIR structures. If a FIR filtering structure is better suited for some practical implementation, this interpolated FIR filter is worth to be considered.

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