CBOC – AN IMPLEMENTATION OF MBOC

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BIOGRAPHY

José-Ángel Ávila-Rodríguez is research associate at the Institute of Geodesy and Navigation at the University of the Federal Armed Forces Munich. He is responsible for research activities on GNSS signals, including BOC, BCS, and MBCS modulations. He is involved in the GALILEO program, in which he supports the European Space Agency, the European Commission, and the GALILEO Joint Undertaking, through the GALILEO Signal Task Force. He studied at the Technical Universities of Madrid, Spain, and Vienna, Austria, and has an M.S. in electrical engineering. His major areas of interest include the GALILEO signal structure, GNSS receiver design and performance, and GALILEO codes.

Stefan Wallner studied at the Technical University of Munich and graduated in 2003 with a Diploma in Techno-Mathematics. He is now research associate at the Institute of Geodesy and Navigation at the University of the Federal Armed Forces Germany in Munich. His main topics of interests can be denoted as the Spreading Codes and the Signal Structure of GALILEO.

Guenter W. Hein is Full Professor and Director of the Institute of Geodesy and Navigation at the University FAF Munich. He is responsible for research and teaching in the fields of high-precision GNSS positioning and navigation, physical geodesy and satellite methods. He has been working in the field of GPS since 1984 and is author of numerous papers on kinematic positioning and navigation as well as sensor integration. In 2002 he received the prestigious "Johannes Kepler Award" from the US Institute of Navigation (ION) for "sustained and significant contributions to satellite navigation". Presently he is heavily involved in the GALILEO program.

Emilie Rebeyrol graduated as a telecommunications engineer from the INT (Institut National des Télécommunications) in 2003. She is now a Ph.D student at the satellite navigation lab of the ENAC. Currently she carries out research on Galileo signals and their generation in the satellite payload in collaboration with the CNES (Centre National d'Etudes Spatiales), in Toulouse.

Olivier Julien is an assistant professor at the signal processing laboratory of ENAC (Ecole Nationale de l'Aviation Civile), Toulouse, France. His research interests are GNSS receiver design, GNSS multipath and interference mitigation and GNSS interoperability. He received his B.Eng in 2001 in digital communications from ENAC and his PhD in 2005 from the Department of Geomatics Engineering of the University of Calgary, Canada.

Christophe Macabiau graduated as an electronics engineer in 1992 from the ENAC in Toulouse, France. Since 1994, he has been working on the application of satellite navigation techniques to civil aviation. He received his Ph.D. in 1997 and has been in charge of the signal processing lab of the ENAC since 2000.

Lionel Ries is a navigation engineer in the Transmission Techniques and signal processing department, at CNES since June 2000. He is responsible of research activities on GNSS2 signal, including BOC modulations and GPS IIF L5. He is involved in the GALILEO program, in which he supports ESA, EC and GJU, through the GALILEO Signal Task Force. He graduated in 1997 from the Ecole Polytechnique de Bruxelles, at Brussels Free University (Belgium), in 1997, and received a M.S. degree from the Ecole Nationale Supérieure de l'Aéronautique et de l'Espace (SUPAERO) in Toulouse (France) in 1998.

Antoine DeLatour from CNES (Centre National d'Etudes Spatiales) is involved in the design of the Galileo signals, in the GPS/Galileo radiofrequency compatibility assessment and in the use of GNSS for space applications.

Laurent Lestarquit graduated from the Ecole Polytechnique of Paris in 1994, and then specialized in space telecommunication systems at the Ecole Nationale de l'Aéronautique et de l'Espace (SUPAERO) in Toulouse, France. Since 1996, he has been collaborating to several projects related to GPS space receivers (HETE2 and STENTOR). He is now involved in the GALILEO program and supports EC and ESA through the GALILEO Signal Task Force. He invented the ALTBOC 8PSK signal proposed for GALILEO in E5. From now on he will be involved in GALILEO orbit determination and time synchronisation.

Jean-Luc Issler is head of the Transmission Techniques and signal processing department of CNES, whose main tasks are signal processing, air interfaces and equipments in Radionavigation, Telecommunication, TT&C, High Data Rate TeleMetry, propagation and spectrum survey. He is involved in the development of several spaceborne receivers in Europe, as well as in studies on the European RadioNavigation projects, like GALILEO and the Pseudolite Network. With DRAST, he represents France in the GALILEO Signal Task Force of the European Commission.

ABSTRACT

Under the 2004 *Agreement on the Promotion, Provision, and Use of GALILEO and GPS Satellite-Based Navigation Systems and Related Applications*, the member states of the European Union and the United States agreed on working together, intensifying thus the cooperation on interoperability and compatibility issues between GALILEO and GPS. Among other topics, one important focus was the E1/L1 frequency band, centred at 1575.42 MHz, where the GALILEO Open Service (OS) signal and the modernized L1 civil (L1C) signal are going to be transmitted along with many other RNSS signals. The opportunity to design new signals in this preeminent radionavigation frequency band has significant importance to future users worldwide.

Recent joint efforts by United States and European experts have identified MBOC (multiplexed BOC) [1][2][3][4][5] as a promising joint solution for E1 OS and L1C, along with multiple sets of spreading waveforms that yield this optimized spectrum. The resulting optimized E1 OS and L1C spreading modulations enable receivers to obtain significantly better performance in multipath than with previously considered spreading modulations, along with other potential benefits. The optimized spreading modulation provides considerable flexibility for receiver designers, and simpler receivers that employ only BOC(1,1)-based processing, experience very modest performance degradation, compared to the baseline BOC(1,1) spreading modulation.

INTRODUCTION

The specific MBOC power spectral density (PSD) that has been proposed can be described as the convex sum of a BOC(1,1) PSD and a BOC(6,1) PSD[5]. Spreading waveforms corresponding to pilot and data components can be formed in a number of ways, including Composite BOC (CBOC), which adds or subtracts BOC(6,1) spreading symbols at the appropriate power level with BOC(1,1) spreading symbols, and time-multiplexed BOC (TMBOC), which produces a spreading waveform containing BOC(1,1) spreading symbols, interspersed with the appropriate fraction of time of BOC(6,1) spreading symbols. Depending upon the approach used, the pilot and data components can consist of either the same or different spreading time series, as long as they are selected to yield the same composite PSD resulting from the combination of both components.

This paper gives a detailed insight into the CBOC implementation of MBOC, presenting its definition and different potential ways of implementing it. Indeed, the high frequency BOC component of BOC(6,1) can be put on either data and pilot or alone on any of them. An extra degree of freedom is given by the possibility of having different power splits between data and pilot. This is indeed easily achieved with a CBOC implementation. Pros and cons will be discussed from a technical point of view. The impact on the definition in the time domain, phase plots and other relevant aspects will be equally analyzed. Additionally, since there exists the possibility that GPS will be implementing MBOC using the TMBOC approach and that Galileo will be using CBOC to materialize MBOC, the TMBOC approach will also be described. Moreover, the interesting possibility of acquiring and tracking the CBOC signal using not only CBOC-like but also prototype TMBOC-like implementations will be an important focus in this paper. This is of relevant interest since preliminary results speak for a high compatibility between GPS and Galileo even though each of them would implement MBOC in a different way.

OPTIMIZED WAVEFORMS FOR GALILEO E1 OS AND GPS L1C

GPS C/A code has been for a long time the dominant signal in the GNSS market, given its great simplicity and extended use. In spite of its very good performance, it has been recognized that much better performance can be obtained using spreading modulations that provide more power at high frequencies away from the centre frequency [6]. Binary offset carrier (BOC) spreading modulations [7] are one way to accomplish this, and indeed a BOC(1,1) spreading modulation was selected as the baseline for the future GALILEO E1 OS and GPS L1C signals. MBOC goes indeed in this direction and increments the amount of power on higher frequencies by adding a high frequency BOC component.

At a first sight, the number of potential MBOC solutions seems enormous but if additional constraints on the desired properties are considered, the number reduces considerably. In fact, MBOC, unlike Composite BCS (CBCS) [8], is formed using an antisymmetric sequence that is orthogonal to BOC(1,1) eliminating thus the tracking bias that CBCS presented. Additionally, it has been shown that among all the potential solutions that fulfilled the desired properties described in [9], BOC(6,1) was the best in terms of performance.

MBOC power spectral density

The multiplexed binary offset carrier (MBOC) PSD recommended in [2][3] is the PSD of the entire signal (pilot and data components together), denoted MBOC(6,1,1/11), where (6,1) refers to the BOC(6,1) part that is added with BOC(1,1) and 1/11 denotes the percentage of power of BOC(6,1) with respect to the total signal MBOC power [5]. The PSD of MBOC(6,1,1/11) is shown to be

$$G_{Signal}(f) = \frac{10}{11} G_{BOC(1,1)}(f) + \frac{1}{11} G_{BOC(6,1)}(f) \quad (1)$$

where $G_{BOC(m,n)}(f)$ is the unit-power PSD of a sine-phased BOC spreading modulation. It must be noted that the definition above corresponds to the ideal case of infinite bandwidth and that data and pilot are computed together.

Figure 1 next shows the increase in higher frequency power, compared to that of BOC(1,1). It must be noted that this PSD is the average of adding up data and pilot together. Therefore different power splits between data and pilot could result in the same spectrum although the performance of every channel considered individually would differ correspondingly.

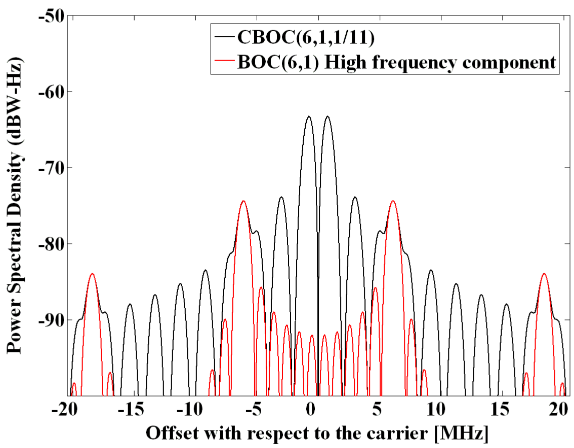

Figure 1. Power Spectral Density of MBOC(6,1,1/11), Showing MBOC(6,1,1/11)'s Additional Power at Higher Frequencies

The recommended MBOC(6,1,1/11) is a specific case of CBCS [8] with the difference that while MBOC does not specify the payload implementation of the signal, CBCS was based on a modified Interplex modulation. MBOC was selected to meet technical constraints in the Agreement [1], to retain a high degree of interoperability with receivers that might use BOC(1,1), and to facilitate implementation in satellites and receivers.

Since the definition of the recommended new signal for Galileo E1 OS and GPS L1C is in the frequency domain, different implementations in the time domain will fit into the definition above. CBOC and TMBOC are the two approaches that are being studied [5]. But before presenting the details of every particular implementation, it is important to mention the impact of changing the power split between data and pilot. The reason for this is that GPS is planning to have 75% of the power on the pilot channel and 25% on the data channel while Galileo's baseline is to have the same power for data and pilot. Therefore, Galileo E1 OS and GPS L1C are expected to have different power splits between the data and pilot channels.

As it is well known, Galileo will have a much higher data rate than GPS due to the presence of an integrity message and, as it can be shown, although it would be desirable that both GPS and Galileo would have the same implementation of MBOC and same power split, in the case of Galileo moving from a power split of 50/50 to another of 75/25 the BER would increase too much due to the high data rate which might be unacceptable for the ground segment receivers.

CBOC DEFINITION AND IMPLEMENTATION

To produce an MBOC(6,1,1/11) spectrum [5], the spreading symbols used are BOC(1,1) spreading symbols denoted $SC_1(t)$ and BOC(6,1) spreading symbols denoted $SC_2(t)$.

A possible CBOC implementation is based on the approach presented in [5] and [8], using four-level spreading symbols formed by the weighted sum of $SC_1(t)$ and $SC_2(t)$ symbols. A time domain representation of a CBOC implementation is shown in Figure 2:

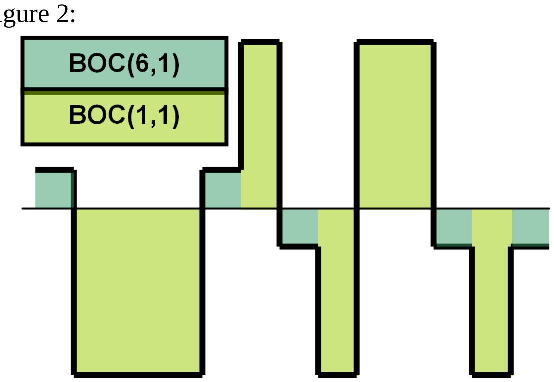

Figure 2. Pseudo-random time multiplexing of BOC(6,1) and BOC(1,1) in the CBOC solution

Two different implementations of CBOC could be considered for a 50%/50% power split between data and pilot components:

- CBOC symbols could be used on both data and pilot components, and formed from the sum of $\sqrt{10/11}\,SC_1(t)$ symbols – BOC(1,1) – and $\sqrt{1/11}\,SC_2(t)$ symbols – BOC(6,1) – , yielding the PSDs defined in (1).
- Alternatively, CBOC symbols could be used on only the pilot component, with the data component remaining all $g_{BOC(1,1)}(t)$. In this case the CBOC is formed from the sum of $\sqrt{9/11}\ SC_1(t)$symbols and $\sqrt{2/11}\ SC_2(t)$ symbols The resulting PSDs would be the same as (1) but the performance of data and pilot channel would be different favouring some applications to the detriment of others.

For the case of a CBOC waveform on both data and pilot components, the interplex modulation containing CBOC(6,1,1/11) and the rest of signals of E1 can be expressed mathematically as follows:

$$s(t) = A_1 \left[\begin{array}{l} \dfrac{C_A(t)}{2} \cdot \left(SC_1(t) \cdot \cos\theta_1 + SC_2(t) \cdot \cos\theta_2 \right) + \\ + \dfrac{C_B(t)}{2} \cdot \left(SC_1(t) \cdot \cos\theta_1 - SC_2(t) \cdot \cos\theta_2 \right) + \\ + j \cdot PRS(t) \cdot \dfrac{\sin\theta_1 + \sin\theta_2}{2} + IM(t) \end{array} \right]$$

$$IM(t) = -j \cdot C_A(t) \cdot C_B(t) \cdot PRS(t) \frac{\sin\theta_1 - \sin\theta_2}{2} \qquad (2)$$

where:

- A_1 is the amplitude of the modulation envelope, sum of the OS data and pilot signals, PRS and the Inter-Modulation product IM. The maximum possible value of A_1 that respects the agreement is a function of the percentage of power put on the BOC(6,1) component of the signal, and the relationship between the power put on the BOC(1,1) and BOC(6,1) component.
- θ_1 and θ_2 describe the angular distance of points of the 8-PSK modulation as described in Figure 3 which depends on the percentage of power that is put on the BOC(6,1) component.
- SC_1=sign[sin($2\pi R_{sc1}t$)] represents the BOC(1,1) modulation with R_{SC1}=1.023 MHz
- SC_2= sign[sin($2\pi R_{sc2}t$)] represents the BOC(6,1) modulation with R_{SC2}=6*1.023 MHz
- $PRS(t)$ is the PRS modulation BOC_{cos}(15,2.5)
- IM is the Inter-Modulation product
- $C_A(t)$ and $C_B(t)$ are the codes for the data and pilot channel respectively.

It can be observed that in this configuration, the data and pilot components use the BOC(6,1) sub-carrier with opposite signs.

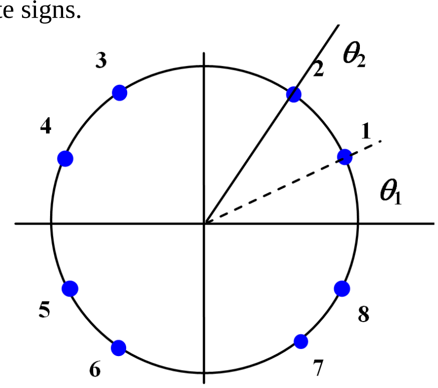

Figure 3. Modified Hexaphase modulation with constant envelope for the optimized signal

The modulation can be optimized so as to pseudo-randomly time-multiplex the BOC(1,1) and BOC(6,1) sub-carriers on the in-phase component. Indeed, rearranging the terms of the expression (2) above to make this pseudo-random time-multiplexing appear yields to the following expression:

$$s(t) = A_1 \left[\begin{array}{l} \dfrac{C_A(t) + C_B(t)}{2} \cdot SC_1(t) \cdot \cos\theta_1 + \\ + \dfrac{C_A(t) - C_B(t)}{2} \cdot SC_2(t) \cdot \cos\theta_2 + \\ + j \cdot PRS(t) \cdot \dfrac{\sin\theta_1 + \sin\theta_2}{2} + IM(t) \end{array} \right] \qquad (3)$$

The normalized autocorrelation function of the CBOC(6,1,1/11) spread spectrum time series (using the data and pilot components), is illustrated in Figure 4, along with the autocorrelation function for BOC(1,1). Observe that CBOC(6,1,1/11)'s correlation function main peak is narrower than that of BOC(1,1), while the widths at values of 0.5 and at the zero crossing are virtually the same.

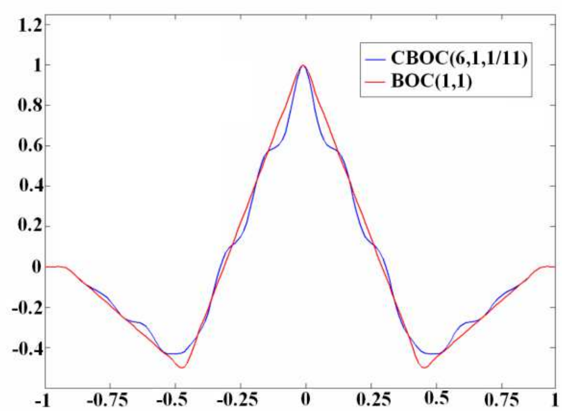

Figure 4. Normalized Autocorrelation Functions Computed over ±15 MHz Bandwidth

If the CBOC is considered only on the pilot component, the general expression for the composite signal may be defined slightly differently from (2), as follows:

$$\begin{aligned} s(t) &= k_1 D_D C_D SC_1(t) + \\ &+ C_P \{k_2 SC_1(t) - k_3 SC_2(t)\} \\ &+ jD_{PRS} C_{PRS} \{k_4 \, PRS(t) + IM(t)\} \end{aligned} \quad (4)$$

where the constants k_1, k_2, k_3 and k_4 are calculated from fixing the power split between data and pilot, the relationship of powers between OS and PRS and the percentage of power on BOC(6,1) with respect to the total OS power under the constraint of a constant envelope modulated signal.

The only inconvenient of this solution is that unless the BOC(6,1) is alternated in sign, depending on the power split between data and pilot, some solutions will present a cross-term in the spectrum that will make the resulting real implementation of the spectrum not to match 100% with the required PSD of MBOC, as it will be shown in the upcoming CBOC tracking section. Additionally, the number of phase plots increases as Figure 5 next shows.

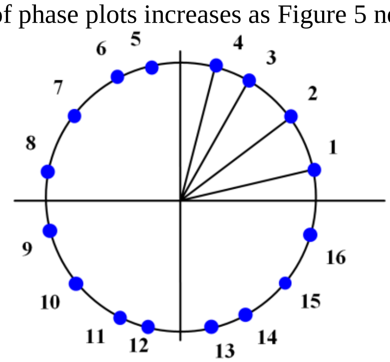

Figure 5. Modified phase modulation with constant envelope for the optimized signal and different power split between data and pilot.

TMBOC DEFINITION AND IMPLEMENTATION

TMBOC states for Time Multiplexed BOC modulation and is the other possible implementation of MBOC. It has a binary representation for both data and pilot components.

In a TMBOC spreading time series, different BOC spreading symbols are used in different slots in the time domain as Figure 6 shows, following a periodic or deterministic pattern.

As repeatedly explained, the total power can be divided between the pilot and data components in many different forms and still yield the same average agreed power spectral density [5]. One possible TMBOC implementation [4] proposed for GPS L1C is to place 75% of the total power on the pilot component while the other 25% of the power is on the data component. Additionally, this proposed solution uses all BOC(1,1) spreading symbols on the data component, while the pilot channel comprises 29/33 BOC(1,1) spreading symbols and 4/33 BOC(6,1) spreading symbols. Indeed, in spite of having all the BOC(6,1) on the pilot channel, the average power is shown to fulfil Equation (1).

If instead of having a power split of 75%/25% for pilot and data respectively we had 50%/50% power split between pilot and carrier component as it is the baseline for Galileo, a candidate TMBOC implementation would be to use all BOC(1,1) spreading symbols on the data component, and 2/11 BOC(6,1) spreading symbols on the pilot. Another imaginable option for a signal with 50%/50% power split between pilot and data components would be to place 1/11 BOC(6,1) spreading symbols on both the pilot and data, as it is also done in the typical CBOC implementation described above.

The exact locations of the BOC(6,1) spreading symbols depend on several considerations. For example, if BOC(6,1) symbols are placed in both the pilot and data components, the easiest would be to have the same placement for both components. But another possibility is to properly place the BOC(6,1) symbols to improve the spreading codes' auto and cross-correlation. Preliminary results have shown gains in the order of 1 dB for GPS L1C [5] and undergoing studies show similar results for Galileo E1 OS.

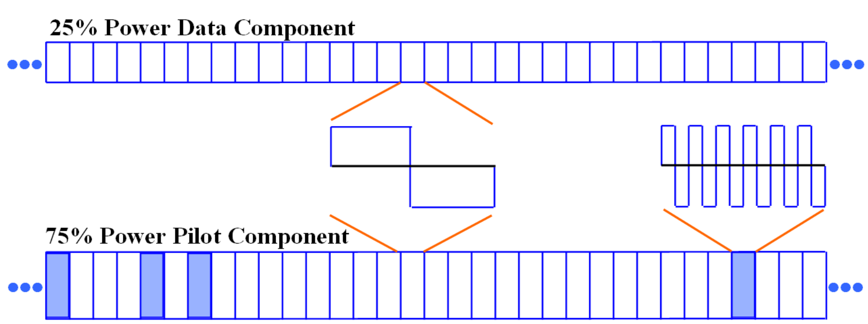

Figure 6. Example of TMBOC(6,1,4/33) Spreading Time Series, with All BOC(6,1) Spreading Symbols in the 75% Pilot Power Component

Now that we have defined the CBOC and TMBOC implementations of MBOC it is the moment to comment a little bit more on the flexibility of both approaches. Indeed, as we have commented above and regardless of the impact on the BER, the CBOC modulation can also have different power splits between data and pilot just as TMBOC. This brings an enormous flexibility since any power split between data and pilot can be achieved and additionally data and pilot channels could have different signals on it.

Nevertheless there is an important difference between both approaches in this regard, since while CBOC could adopt any imaginable percentage of power for the high frequency component, TMBOC can only adopt certain numbers given its time granularity.

MBOC PERFORMANCE ASSESSMENT

Several different parameters were considered during the optimization process of the E1 OS signal waveforms. Among all, the most important characteristic in this process was the improvement of the tracking and multipath performance maintaining the compatibility and interoperability with the existing systems.

MBOC presents a clear improvement in all these metrics with respect to BOC(1,1) showing figures that are close to those of BOC(2,2) and even superior for certain bandwidths [5].

OPTIMAL TRACKING OF CBOC

The objective of this section is to describe the tracking performances using the CBOC modulation when only one channel of the Galileo E1 OS (data or pilot channel) is used. As it has been explained in the previous sections, the waveforms used on the data or pilot channels depend upon the choice of the whole Galileo E1 modulation. When looking at (2) and (4), it can be seen that the expression of each channel waveform can differ. Two cases were exposed herein:

- The use of a CBOC(6,1,1/11) where the data and pilot channels have a BOC(6,1) component with opposite signs (see(2)). In this case, the power of the BOC(6,1) part is 1/11 of the channel total power.
- The use of a CBOC(6,1,2/11) where the pilot (or data) channel has a BOC(6,1) component with alternating sign, while the other channel is a pure BOC(1,1) (see (4)). In this case, the power of the BOC(6,1) part is 2/11 of the pilot channel total power.

For the sake of completeness, another scenario is considered where a CBOC(6,1,1/11) is used both on the data and pilot channels, both with a BOC(6,1) component with alternating sign. In this case, the data and pilot channels have a BOC(6,1) power of 1/11 of the channel total power.

The 3 following signal models can then be used:

$$CBOC(6{,}1,p,'-')(t)=C(t)\{V\,SC_1(t)-W\,SC_2(t)\}$$
$$CBOC(6{,}1,p,'+')(t)=C(t)\{V\,SC_1(t)+W\,SC_2(t)\}$$
$$CBOC(6{,}1,p,'+/-')(t)=\begin{cases}C(t)\{V\,SC_1(t)+W\,SC_2(t)\} & \text{even chips}\\ C(t)\{V\,SC_1(t)-W\,SC_2(t)\} & \text{odd chips}\end{cases}$$

where $p=\dfrac{W^2}{V^2+W^2}$ (5)

The autocorrelation function of these three CBOC waveforms is given by:

$$R_{CBOC('-')}(\tau)=\begin{pmatrix}V^2R_{BOC(1,1)}(\tau)+W^2R_{BOC(6,1)}(\tau)\\ -2VWR_{BOC(1,1)/BOC(6,1)}(\tau)\end{pmatrix}$$
$$R_{CBOC('+')}(\tau)=\begin{pmatrix}V^2R_{BOC(1,1)}(\tau)+W^2R_{BOC(6,1)}(\tau)\\ +2VWR_{BOC(1,1)/BOC(6,1)}(\tau)\end{pmatrix}$$
$$R_{CBOC('+/-')}(\tau)=\left(V^2R_{BOC(1,1)}(\tau)+W^2R_{BOC(6,1)}(\tau)\right) \quad (6)$$

The autocorrelation functions shown in (6) clearly exhibit the existence of the cross-term between the BOC(1,1) and BOC(6,1) parts for the CBOC('+') and CBOC('-') cases, which is the reason why they would not meet the MBOC definition if used alone, as commented above. Figure 7 shows the autocorrelation functions of each of the CBOC type. It is compared to the corresponding TMBOC autocorrelation functions. It can be seen that:

- the percentage of BOC(6,1) power in the signal channel (data or pilot) total power will shape the correlation function. The higher the value of p, the more the autocorrelation function will have the undulations of the pure BOC(6,1) autocorrelation function.
- The sign of the BOC(6,1) component will also shape the correlation function: with a '-', the main peak of the autocorrelation function is narrower.
- The TMBOC and the CBOC('+/-) have very close (not to say the same) autocorrelation functions (thus both respecting, if used alone, the MBOC constraint).

Assuming traditional processing of the incoming signal, it is well-known that the tracking performances of that signal will be influenced by the shape of its autocorrelation function. Thus it can be expected that according to the CBOC type, the performances will be different. Three main criterions are studied herein: (1) the risk of biased tracking, that will influence the existence of potential false lock points, (2) the code tracking noise induced by thermal noise, and (3) the multipath-induced code tracking error.

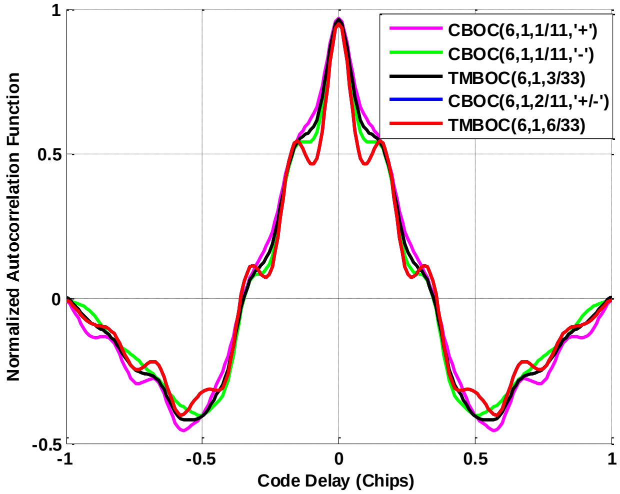

Figure 7 –CBOC and TMBOC Autocorrelation Function

False Tracking Points

Looking at the autocorrelation function, it is easy to understand that the existence of secondary peaks can lead to stable false lock points. In that respect, it can be seen that the CBOC(1/11) is not likely, in any of its types, to lead to stable false lock points close to the expected lock point. Indeed, the first false lock point would be approximately for a code delay of 0.6 chips, resulting in a bias of around 175 metres, thus easily detectable. For the CBOC(6,1,2/11,'+/-'), on the other hand, the existence of false lock points seems unavoidable. The closest one should be around 0.15 chips, equivalent to a measurement bias of 43 metres. Thus, it might be more difficult to detect.

Note that in any case, due to the dominant BOC(1,1) component and its secondary peak located at 0.5 chips, a false lock detector is necessary in order to make sure that the receiver is tracking the signal based on the autocorrelation main peak.

Thermal Noise-Induced Code Tracking Error

Figure 8 shows the code tracking noise for a dot-product (DP) discriminator for the different CBOC cases considered, as well as the case when the incoming signal is a pure BOC(1,1) (with equal signal power) (the theoretical formula can be found using Appendix C in [10]). It can be seen that the best performer is the CBOC('+/-') with 2/11 of BOC(6,1) power, taking full advantage of its higher power at high frequencies. Within the CBOC(1/11) case, it can be seen that the CBOC(1/11,'+') has the lowest performance, while the CBOC(1/11,'-') has the best, as expected since its autocorrelation function's main peak is the narrowest. Still, all the CBOC modulations bring a significant improvement compared to the tracking of a pure BOC(1,1) modulation. That improvement, in terms of equivalent C/N_0, is between 1.9 and 3.1 dB if a CBOC(1/11) modulation is used and equals 4.2 dB if a CBOC(6,1,2/11,'+/-') is used.

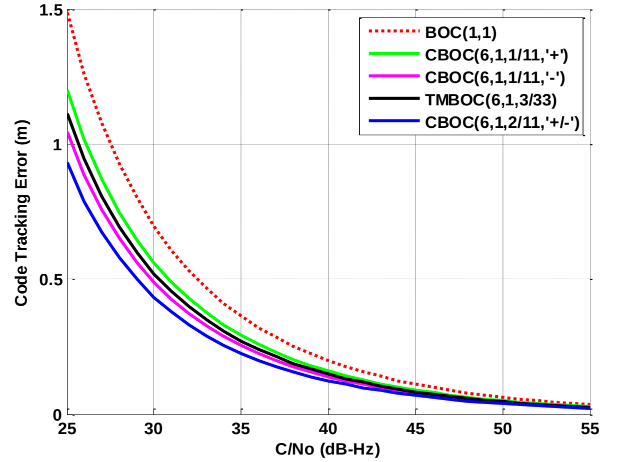

Figure 8 – BOC(1,1), CBOC and TMBOC DLL Tracking Performance Assuming a DP Discriminator, a 1 Hz Loop Bandwidth, a 1/12 Chip E-L Spacing, a 4 ms Integration Time, and a 12 MHz One-Sided Filter

Multipath-Induced Tracking Error

Multipath-induced tracking errors are also dependent upon the autocorrelation function shape. It is thus interesting to compare the performance of the different CBOC candidates against multipath. A common figure of merit is the multipath running average error as done in [5]. It is plotted in Figure 9 for an early-late spacing of 1/12 chips and a one-sided front-end filter of 12 MHz. Once again, it can be seen that the CBOC(6,1,2/11,'+/-') tracking has the best performance. This is mostly due to the narrow peaks constituting its autocorrelation function. Comparing the CBOC(1/11) cases, multipath rejection is more effective for the CBOC(6,1,1/11,'-') tracking case, followed by CBOC(6,1,1/11,'+/-') tracking and then CBOC(6,1,1/11,'+') tracking. Note that all these multipath results for the CBOC are much better than for pure BOC(1,1) tracking.

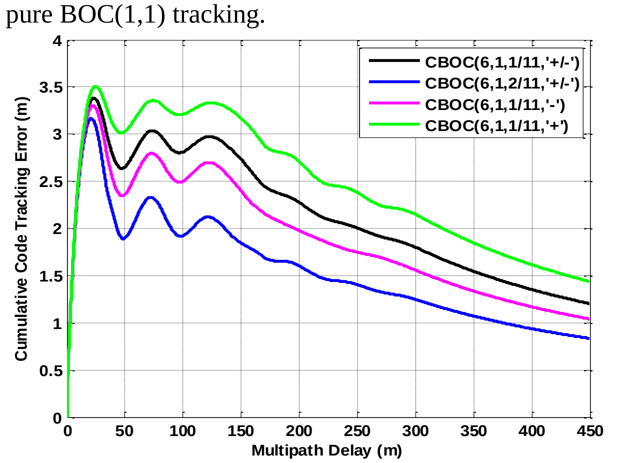

Figure 9 – CBOC Multipath Running Average Error with a 1/12 Chip E-L Spacing and a 12 MHz One-Sided Filter

Conclusions on Optimal CBOC Tracking

It has been seen that for the CBOC(1/11) cases, the best tracking performances were achieved for the CBOC(6,1,1/11,'-'). However, it has to be reminded that in this case, the counter part is the use of a CBOC(6,1,1/11,'+') on the other channel to meet the MBOC specification, and that exhibits the worst performance (within the CBOC(1/11) family).

Another option is the use of a CBOC(6,1,1/11,'+/-') on the data and pilot channels, in which case both channels would have the same tracking performances, and would also offer excellent interoperability. The drawback in this case would be the extra complexity required for the implementation of the alternation.

Finally, the option of using a CBOC(6,1,2/11,'+/-') on one channel and a pure BOC(1,1) on the other channel would allow having one channel with excellent tracking performance (4.2 dB higher equivalent C/N_0 than pure BOC(1,1) for tracking noise and best multipath performance), while the other channel would just use a pure BOC(1,1) modulation. Again, the main drawback would be the extra complexity needed to do the alternation.

However, the traditional processing of a CBOC signal, as shown in this section, implies that a replica of the CBOC signal has to be locally generated by the receiver. As the CBOC is a linear combination of two sub-carriers, it has more than two levels. This means that the local replica has to be encoded on at least 2 bits, which signifies the need for more challenging receiver architecture. This could be detrimental to the use of CBOC and it is then interesting to look at techniques that would only use local replicas encoded on 1-bit, while maintaining interesting tracking performances.

An example of such a method is the separate correlation of the incoming CBOC signal with, on one side a pure BOC(1,1) replica, and on the other side, a pure BOC(6,1) replica. A simple linear combination of these two correlation values would result in the exact same correlation value than the CBOC autocorrelation value and thus the exact same tracking performances. However, this processing requires twice as many correlators as the traditional CBOC tracking. The following part introduces a new CBOC tracking technique that intends to remove that problem.

TRACKING CBOC WITH A 1-BIT RECEIVER

The idea here is to correlate the incoming CBOC signal with a locally generated signal obtained by time-multiplexing a BOC(1,1) sub-carrier and a BOC(6,1) sub-carrier. This local replica can be modelled as:

$$TM61(t) = \begin{cases} C_m(t)\cdot SC_1(t), & if \quad t \in S_2 \\ C_m(t)\cdot SC_2(t), & if \quad t \in S_1 \end{cases} \qquad (7)$$

where S_1 is the union of the segments of time when a BOC(1,1) sub-carrier is used, while S_2, the complement of S_1 in the time domain, is the union of the segments of time when a BOC(6,1) sub-carrier is used.

Using such a local sub-carrier for the correlation process allows the use of the BOC(1,1) and BOC(6,1) components while keeping a 1-bit encoded local replica and a limited number of complex correlators per channel. In the following, α will designate the percentage of the BOC(6,1) sub-carrier used in one code length, and $\beta = 1-\alpha$ represents the BOC(1,1) local sub-carrier part. Note that the sign of the BOC(6,1) local sub-carrier in the local replica is taken according to the sign of the BOC(6,1) used in the CBOC signal. The tracking technique will be referred to as TM61(α).

The resulting cross-correlation functions of the TM61 local replica with the CBOC incoming signal are, according to the type of CBOC received:

$$R_{CBOC('-')/TM61(\alpha)}(\tau) = \begin{pmatrix} \beta V R_{BOC(1,1)}(\tau) + \alpha W R_{BOC(6,1)}(\tau) \\ -(\beta W + \alpha V) R_{BOC(1,1)/BOC(6,1)}(\tau) \end{pmatrix}$$

$$R_{CBOC('+')/TM61(\alpha)}(\tau) = \begin{pmatrix} \beta V R_{BOC(1,1)}(\tau) + \alpha W R_{BOC(6,1)}(\tau) \\ +(\beta W + \alpha V) R_{BOC(1,1)/BOC(6,1)}(\tau) \end{pmatrix}$$

$$R_{CBOC('+/-')/TM61(\alpha)}(\tau) = \frac{1}{2}\begin{pmatrix} (\beta^- + \beta^+) V R_{BOC(1,1)}(\tau) \\ + (\alpha^- + \alpha^+) W R_{BOC(6,1)}(\tau) \\ + ((\alpha^- - \alpha^+)(W - V) R_{BOC(1,1)/BOC(6,1)}(\tau)) \end{pmatrix} \qquad (8)$$

where $\alpha^- = 1-\beta^-$ is the percentage of BOC(6,1) with a negative sign used in the TM61 replica (with respect to the negative part of the CBOC('+/-')), and $\alpha^+ = 1-\beta^+$ is the percentage of BOC(6,1) with a positive sign (with respect to the positive part of the CBOC('+/-')).

Two main effects can be observed from these TM61/CBOC cross-correlation functions:

- each correlation function is weighted, which is comparable to the CBOC autocorrelation case (see (6)), with the difference that the weights are controlled by the value of α, and
- the cross-correlation between TM61 and CBOC signals induces correlation losses that will impact the post-correlation SNR and thus the processing of the signal after the correlation operation. These correlation losses are represented in Figure 10 for different values of p. In any represented case, the

choice of a low value for α results in reduced correlation losses.

When considering code tracking noise assuming a DP discriminator, it is recommended to generate different local TM61 replicas on the prompt correlator and on the early and late correlators. Indeed, using a DP discriminator, the prompt correlator only influences the tracking noise squaring losses. Thus, it is important to choose a local prompt replica that minimizes the correlation loss with the incoming CBOC signal. Hence, according to Figure 10, the use of a pure BOC(1,1) (case α =0) replica for the prompt local replica appears recommended. It is also important to realize that the prompt correlator output is used for phase tracking and data demodulation. Thus, choosing a low correlation loss, inducing a high SNR, is also significantly profitable for these two processes. Consequently, in the remaining part of this article, the prompt local replica of the TM61 tracking technique is considered to use a pure BOC(1,1) sub-carrier.

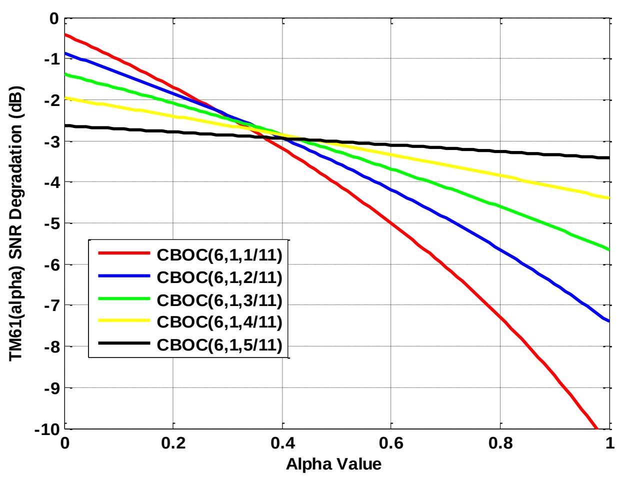

Figure 10 – Post-Correlation SNR Degradation due to the TM61/CBOC Correlation

The early and late correlator outputs will play a role in the code tracking noise in the sense that they will affect the gain of the discriminator as well as the noise correlation between the early and late correlators' output. Table 1 shows the tracking noise degradation resulting from use of the TM61 technique to track the different CBOC cases compared to optimal TMBOC tracking (with the same associated BOC(6,1) power) in terms of equivalent C/N_0. The comparison with the optimal TMBOC tracking is justified since the TMBOC modulation is the other candidate for the MBOC implementation. It can be observed that for the CBOC(1/11) cases, either a high (close to 1) or a low (close to 0) value of α is interesting. The equivalent C/N_0 degradations for the tracking noise in that case are 1.9 dB for the CBOC(6,1,1/11,'-'), 3 dB for the CBOC(6,1,1/11,'+/-') and 4 dB for the CBOC(6,1,1/11,'+'). Thus, the degradation is limited, although it is not satisfactory for the CBOC(6,1,1/11,'+') case. On the other hand, it seems that the use of high values of α (the extreme case being the use of a local pure BOC(6,1) replica) are well suited for the CBOC(6,1,2/11,'+/-'). In such a case, the TM61 tracking technique would only use pure local replicas (pure BOC(1,1) on the local prompt and pure BOC(6,1) on the local early and late correlators) and thus requiring a very simple receiver architecture with no multiplexing.

Table 1 - TM61(α) Tracking Noise Degradation wrt TMBOC in Terms of Equivalent C/N_0 for Different CBOC Configurations

Value of α for Early and Late TM61(α) Local Replicas	TM61(α) Tracking Noise Degradation wrt TMBOC in Terms of Equivalent C/N0 (dB)			
	CBOC(6,1,1/11,'x') vs TMBOC(6,1,1/11)			CBOC(6,1,2/11,'x') vs TMBOC(6,1,2/11)
	+	-	+/-	+/-
0	4	2	2.9	5
0.2	5.1	2.9	3.6	4.2
0.4	5.1	2.8	3.4	3.3
0.6	4.9	2.6	3.3	2.6
0.8	4.6	2.3	3.2	2.1
1	4.3	1.9	3	1.6

Concerning multipath rejection, it appears that for both the CBOC(1/11) and the CBOC(2/11) cases, a value of α around 0.5 would lead to the best results. However, values of α close to 1 also give close-to-optimal results. It also seems that using α =0 is not recommended for multipath rejection considerations. An example of multipath running average error using the TM61 tracking technique for an early-late spacing of 1/12 chips and a 12 MHz one-sided front-end filter is given for the CBOC(6,1,1/11,'+/-') for different values of α in Figure 11.

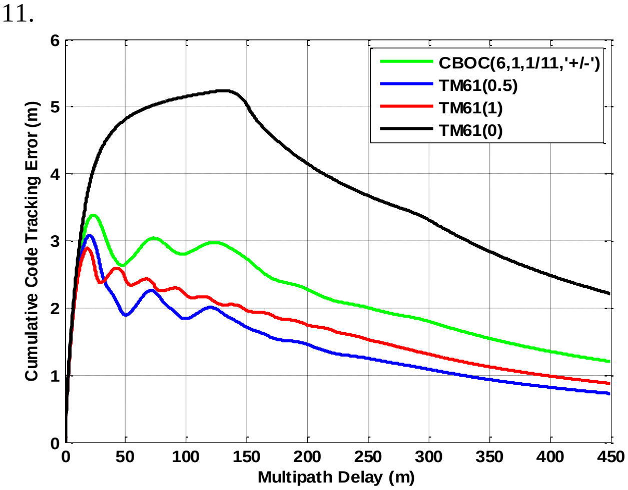

Figure 11 – TM61 Multipath Running Average Error Assuming a 1/12 Chip Early-Late Spacing, and a 12 MHz One-Sided Filter

Conclusions on CBOC Tracking by a 1-bit Receivers

It has been seen that a preferred version of the TM61 tracking technique would be to locally generate a pure local BOC(1,1) replica for the prompt correlator and a pure local BOC(6,1) replica for the early and late correlators. This provides with low-to-medium degradation for code tracking noise compared to optimal TMBOC tracking, depending upon the percentage of BOC(6,1) in the incoming CBOC signal, and close-to-optimal multipath mitigation. This also has the great advantage of offering very simple receiver architecture since no multiplexing is required and only pure sub-carriers are needed.

Further work shall be pursued to study the use of this TM61 technique with an incoming TMBOC signal in order, if implemented, to strengthen interoperability with GPS L1.

CONCLUSIONS

This paper has analysed the two possible implementations of MBOC, namely TMBOC and CBOC, emphasizing the goodness of the CBOC solution. Although both approaches generate in average the same spectrum when data and pilot are computed together, the time series are different in the time domain.
From the point of view of performance there are no differences of importance between the two implementations, and actually the performance depends on the final user configuration. Indeed putting all the power of the high frequency component on the pilot will be extremely positive if the receiver only uses the pilot channel for tracking but some receivers do use both channels and in this case there will be no difference.

The tracking performance of the optimal CBOC implementation was shown to be excellent with respect to tracking noise and multipath compared to the use of a pure BOC(1,1) modulation.

A new tracking technique, designed to use only a 1-bit local replica and a limited number of correlators, has been presented and its main results shown. It can take a very simple form, requiring only pure local BOC(1,1) and pure BOC(6,1) replicas without multiplexing. Its tracking performances in white noise show a low-to-medium degradation dependent upon the incoming CBOC (1.9 dB in terms of equivalent C/N_0 for the CBOC(6,1,1/11,'-'), 3 dB for the CBOC(6,1,1/11,'+/-'), 4 dB for the CBOC(6,1,1/11,'+') and 1.6 dB in the case of CBOC(6,1,2/11,'+/-')). Additionally, it exhibits a multipath resistance equivalent or even better than optimal CBOC tracking thus strongly outperforming pure BOC(1,1) tracking..

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