Galileo Dual-Channel CBOC Receiver Processing under Limited Hardware Assumption*

Elena Simona Lohan and Heikki Hurskainen

Tampere University of Technology,
Korkeakoulunkatu 1, 33101 Tampere, Finland
www.cs.tut.fi/tlt/pos
{elena-simona.lohan,heikki.hurskainen}@tut.fi

Abstract. Composite Binary Offset Carrier (CBOC) modulation is currently proposed for future Open Service (OS) Galileo signals in E1 frequency band. CBOC consists of a weighted sum or difference of two sine Binary Offset Carrier (BOC) waveforms: a sine BOC(1,1) and a sine BOC(6,1) component. The transmitted OS signal has both data and pilot channels. Data and pilot channels use slightly different modulation, namely CBOC(+) (i.e., weighted sum of BOC(1,1) and BOC(6,1)) and CBOC(-) (i.e., weighted difference of those). At the receiver side, depending on the number of channels available, several approaches are possible: processing either data or pilot, or processing both channels with any of the BOC(1,1) and BOC(6,1) components, or with a weighted or time-multiplexed combination of those. Therefore, a significant number of receiver processing variants is possible. The focus here is on the architectures having a limited hardware available, when we assume that only two channels per satellite and per E1 Open Service signal are used at the receiver and when we have one-bit processing only. This allows us to either process both data and pilot channels with a single sine-BOC(1,1) reference, or to process only the data channel with both BOC(1,1) and BOC(6,1) components, and then combine them with appropriate weights. The question we address here is which of these two variants is better in terms of performance. The novelty of our solution comes from an analytical approach of this problematic and from the comparison of the two architectures in terms of tracking performance at various bandwidths. Our analysis focuses both on narrowband receiver cases (i.e., low front-end receiver bandwidts, of interest in mass-market applications), and on wideband receiver cases (more suitable for professonal receivers). The tracking results are analyzed in terms of tracking error variances and multipath error envelopes.

Keywords: Binary Offset Carrier (BOC), Composite Binary Offset Carrier (CBOC), Galileo, Global Navigation Satellite Systems (GNSS), Non-coherent Early Late Power discriminator (NELP) discriminator, narrowband GNSS receiver.

* The research leading to these results has received funding from the European Union's Seventh Framework Programme (FP7/2007-2013) under grant agreement number 227890 (GRAMMAR project). This research work has also been supported by the Academy of Finland.

G. Giambene and C. Sacchi (Eds.): PSATS 2011, LNICST 71, pp. 391–401, 2011.

1 Background and Motivation

For receiver manufacturers and vendors the upcoming changes in signal domain opens new market potential since new products are needed to get benefit of the increased accuracy and availability of GNSS signals. On the other hand, additional challenges are created by increased complexity needs of receiver. It may become impossible to implement optimal structures to receivers due short time-to-market requirements or dependency of 3rd party receiver Intellectual Property. Composite Binary Offset Carrier (CBOC) modulation has now been selected for the Galileo Open Service signal in E1 band [1]. CBOC modulation is a weighted superposition of two sine BOC-modulated signals: a BOC(1,1) and a BOC(6,1) component. The higher modulation BOC(6,1) requires more bandwidth than BOC(1,1), but has the ability to enhance the tracking performance. CBOC signal is a four-level signal, while BOC(1,1) and BOC(6,1) components are two-level signals, able to be implemented via 1-bit receiver processing.

Traditionally, CBOC signals have been processed either with a CBOC receiver (if a large bandwidth, e.g., 24.552 MHz, is available) [2,3,4], or with a sine BOC(1,1) receiver (for narrowband GNSS receivers) [5,6]. In the first case (CBOC-based processing), implementation is based on at least 2 bits. If we assume some hardware restrictions, such as 1-bit receiver and limited number of channels, processing the incoming CBOC signal with BOC(1,1) and/or BOC(6,1) componets separately makes more sense because it reduces the receiver complexity. Therefore, this is the problem we address in this paper: how to choose the dual-channel processing of Galileo CBOC signal from the limited hardware point-of-view. One-bit processing architectures for CBOC signal have been previously proposed in [4,7].

The novelty of these paper comes from analytical model of the two receiver architectures (data plus pilot processing versus data-only processing) and from the architecture we propose for processing the data-only channel in a dual-channel receiver (which is slightly different from the ones proposed in [4,7], as described in Section 2). To the best of the authors' knowledge such investigation of the CBOC receiver architectures with limited number of channels has not been made yet. We believe that the results reported here are important from the designer point of view, because it allows him or her to choose the best architecture according to the available receiver bandwidth.

This paper is organized as follows: the dual-channel architectures for GNSS signal tracking are discussed in Section 2, the analytical model used in our studies is presented in Section 3, the tracking error performance of discussed algorithms is studied in Section 4, and finally the multipath behavior of them is illustrated in Section 5.

2 Dual-Channel Architectures

The collaborative (or composite) tracking of both data and pilot components is discussed in detail in [8]; both non-coherent combining and coherent combining with relative sign recovery are discussed. In our work we assume a non-coherent combining channel architecture, where the code generators are capable of creating only binary outputs. The non-coherent architecture for the traditional architecture, namely the data-pilot tracking, is illustrated in Fig. 1. There, the upper channel tracks data signal and the lower is tracking pilot signal, both using only sine-BOC(1,1)-modulated replicas. The code generators have multiple, differently delayed outputs, each feeding its own correlator. The number of correlators per channel is typically three (e.g. narrow early-minus-late correlator (NELP) [9]) to five (e.g. high resolution correlator (HRC) [10]). The discriminator function outputs are combined, filtered and fed back to code generation. Since our paper focuses on the code tracking, the carrier tracking with its in-phase and quadrature-phase branches of correlators is omitted from figures to gain clarity.

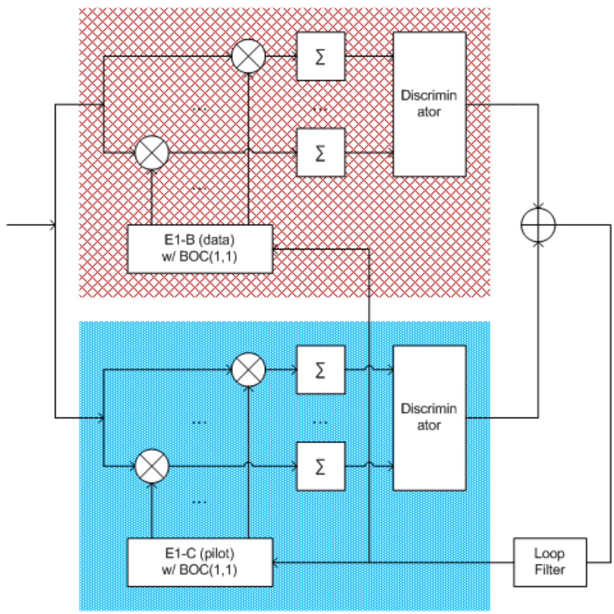

Fig. 1. Dual channel tracking for data and pilot signals (architecture 1)

Similarly, the tracking of four-level CBOC signal with a CBOC-modulated replica code becomes impossible due to binary generator outputs. In [4], a dual-channel approach for CBOC signal tracking has been presented as an alternative to the 4-level CBOC receiver tracking, where two channels are used: the first one for sine-BOC(1,1) and the second for sine-BOC(6,1) tracking. The discriminator results are weighted and combined to reach CBOC signal performance while still using binary subcarrier generation with some hardware overhead. The architecture of [4] has been the starting point of our proposed architecture from Fig. 2, with the main difference that, in our case, no time-multiplexing is used (we use a code-multiplexed approach, where the correlator outputs are weighted and combined as shown in Fig. 2).

The architecture for dual channel tracking for CBOC signal is illustrated in Fig. 2. Here, the differences comppared with the architecture 1 of Fig. 1 are: the code generator on the lower channel produces the spreading code for data signal with BOC(6,1) component and amplitude weighting factors ($\sqrt{1-\alpha^2}$ and α) are applied to discriminator outputs from BOC(1,1) and BOC(6,1) channels, respectively. The α^2 factor refers to the power percentage of BOC(6,1) component.

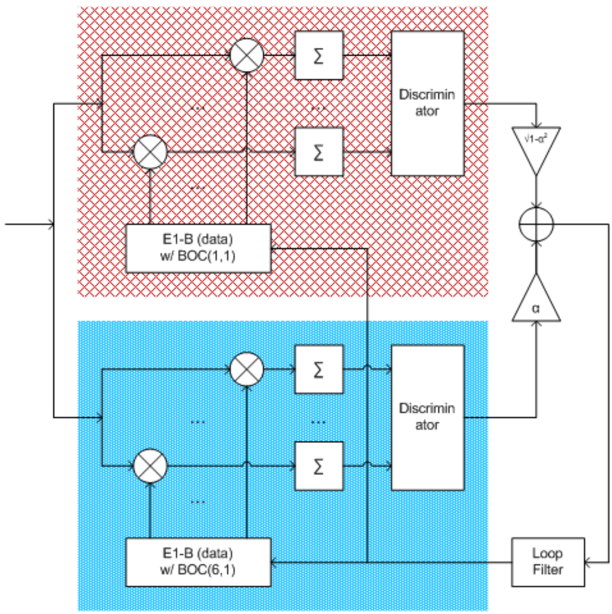

Fig. 2. Dual channel tracking of CBOC signal (architecture 2)

3 Analytical Model

We will adopt here the model introduced in [11] and detailed in [5]. The block diagram of the transmitter-receiver chain for Galileo data or pilot channel is illustrated in Fig. 3. The modulation at the receiver side is dependent on the channel type (CBOC(+) if data, and CBOC(-) if pilot channel) and it is characterized by the $H_{tx}(f)$ transfer function. The modulation at the receiver side, characterized by $H_{rx}(f)$ transfer function can be either a sine BOC(1,1)-modulation (in case we process both data and pilot channels), or a weighted combination of sine BOC(1,1) and sine BOC(6,1) components. The weighting factor is to be decided separately, according to the receiver tracking variance, as explained further on. The correlation block contains a multiplier and an Integrate ad Dump (I&D) block.

The overall effects of the channel $H_c(f)$ and the bandwidth-limiting filter $H_f(f)$ can be lumped in a single term $H(f) = H_c(f)H_f(f)$. This paper focuses on single-path static channel case, therefore assuming that $|H_c(f)| = 1$, in order to find out the maximum achievable performance. This analysis can be straightforwardly extended to multipath fading channels, under a variety of scenarios.

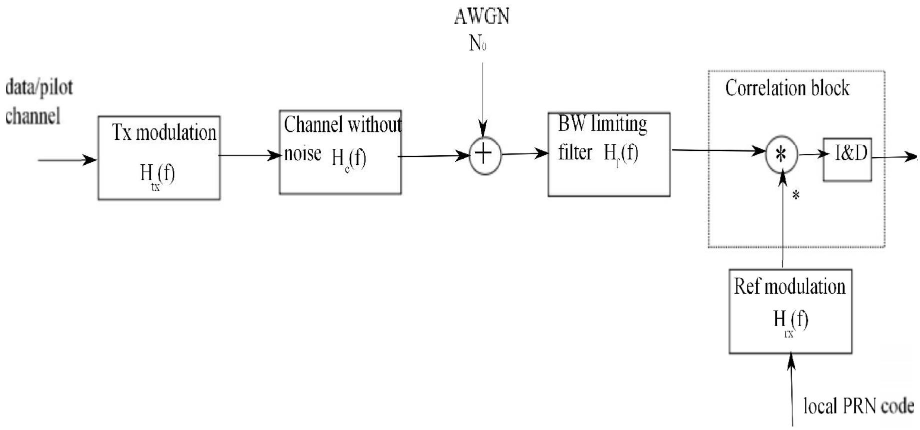

Fig. 3. Transmitter-receiver block diagram in terms of transfer functions

The normalized Power Spectral Density (PSD) of the received signal $\overline{G}_s(f)$ and, respectively, of the noise $\overline{G}_n(f)$ after the correlation can be therefore written as:

$$\begin{cases} \overline{G}_s(f) = \dfrac{|H_{tx}(f)H_{rx}(f)||H(f)|^2}{\displaystyle\int_{-\infty}^{\infty} |H_{tx}(f)H_{rx}(f)||H(f)|^2 df} \\ \overline{G}_n(f) = \dfrac{|H_{rx}(f)|^2|H(f)|^2}{\displaystyle\int_{-\infty}^{\infty} |H_{rx}(f)|^2|H(f)|^2 df} \end{cases} \tag{1}$$

The impulse response of a sine-BOC(m,n) modulation is given by [11]:

$$h_{tx}(t) = p_{T_B}(t) \circledast \sum_{i=0}^{N_B-1} (-1)^i \delta\big(t - i\frac{T_c}{N_B}\big) \tag{2}$$

where $\circledast$ is the convolution operator, $c(t)$ is the pseudorandom code sequence, $N_B = \frac{2m}{n}$ is the BOC modulation order, $T_c = 1/f_c$ is the chip interval, f_c is the chip rate, and $p_{T_B}(t)$ is the convolution between a rectangular pulse of support $T_B = \frac{T_c}{N_B}$ and the front-end filter used to limit the signal bandwidth.

The transfer functions, $H_{tx}(f)$ and $H_{rx}(f)$ can be computed according to the modulation type, using eq. (2) and following the derivations similar to [5] (the full derivations are not included here due to lack of space):

1. Data and pilot channel processing with sine BOC(1,1):

$$\begin{cases} H_{tx,data} = e^{-j\pi f T_c}\frac{sin(\pi f T_c)}{\pi f}\left(\sqrt{\frac{10}{11}}e^{-j\pi f\frac{T_c}{2}}tan(\frac{\pi f T_c}{2}) + \sqrt{\frac{1}{11}}tan(\frac{\pi f T_c}{12})\right) \\ H_{tx,pilot} = e^{-j\pi f T_c}\frac{sin(\pi f T_c)}{\pi f}\left(\sqrt{\frac{10}{11}}e^{-j\pi f\frac{T_c}{2}}tan(\frac{\pi f T_c}{2}) - \sqrt{\frac{1}{11}}tan(\frac{\pi f T_c}{12})\right) \\ H_{rx} \quad = e^{-\frac{3j\pi f T_c}{2}}\frac{sin(\pi f T_c)}{\pi f}tan(\pi f\frac{T_c}{2}) \end{cases} \quad (3)$$

2. Data-only channel processing with a weighted combination of sine BOC(1,1) and sine BOC(6,1):

$$\begin{cases} H_{tx,data} = e^{-j\pi f T_c}\frac{sin(\pi f T_c)}{\pi f}\left(\sqrt{\frac{10}{11}}e^{-j\pi f\frac{T_c}{2}}tan(\frac{\pi f T_c}{2}) + \sqrt{\frac{1}{11}}tan(\frac{\pi f T_c}{12})\right) \\ H_{rx} \quad = e^{-j\pi f T_c}\frac{sin(\pi f T_c)}{\pi f}\left(\sqrt{1-\alpha^2}e^{-j\pi f\frac{T_c}{2}}tan(\frac{\pi f T_c}{2}) + \alpha tan(\frac{\pi f T_c}{12})\right) \end{cases} \quad (4)$$

Above, α^2 is the power percentage of sine BOC(6,1)-component in the reference signal (and $1-\alpha^2$ is the power percentage of sine BOC(1,1)-component).

4 Tracking Error Variances and Optimal Weighting Factor

The code tracking error variance, given in s^2 (squared seconds) for a signal with normalized PSD $\overline{G}_s$ used in a non-coherent early-minus late power delay tracker or correlator (NELP), in the presence of a colored Gaussian noise with normalized PSD $\overline{G}_n$ and operating at a carrier-to-noise density ratio C/N_0 is, according to [12]:

$$\sigma^2_{NELP} = \frac{B_L(1-0.5B_L T)}{(2\pi)^2 C/N_0 I_2^2}\left(I_1 + \frac{I_3 - I_4}{4C/N_0 T(I_5)^2}\right) \quad (5)$$

where $I_i, i = 1, \ldots, 5$ are the following integrals:

$$\begin{aligned} I_1 &= \int_{-\frac{B_W}{2}}^{\frac{B_W}{2}} \overline{G}_{y_s}(f)\overline{G}_{y_n}(f)sin^2(\pi f \Delta_{EL} T_c)df \\ I_2 &= \int_{-\frac{B_W}{2}}^{\frac{B_W}{2}} f\overline{G}_{y_s}(f)sin(\pi f \Delta_{EL} T_c)df \\ I_3 &= \int_{-\frac{B_W}{2}}^{\frac{B_W}{2}} \overline{G}_{y_s}(f)\overline{G}_{y_n}(f)df \\ I_4 &= \int_{-\frac{B_W}{2}}^{\frac{B_W}{2}} \overline{G}_{y_s}(f)\overline{G}_{y_n}(f)e^{j2\pi f \Delta_{EL} T_c}df \\ I_5 &= \int_{-\frac{B_W}{2}}^{\frac{B_W}{2}} \overline{G}_{y_s}(f)cos(\pi f \Delta_{EL} T_c)df \end{aligned} \quad (6)$$

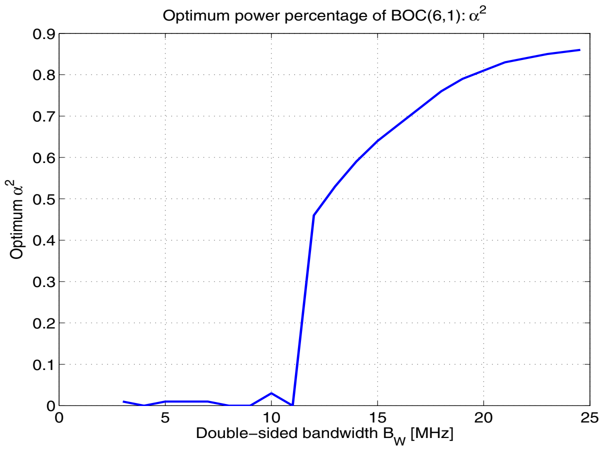

Fig. 4. Optimum power percentage of component sine BOC(6,1) in data channel processing with combined BOC(1,1)/BOC(6,1)

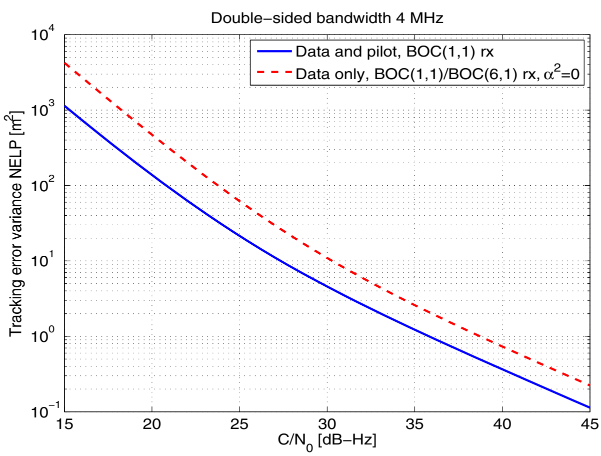

Fig. 5. Tracking error variance of the two dual-channel architectures at $B_W = 4$ MHz

Above, B_L is the NELP loop bandwidth (in Hz), Δ_{EL} is the early-late spacing (in chips), and $T = Nc{*}10^{-3}$ is the coherent integration in seconds, N_c being the number of codewords used in the coherent integration. For example, under the ideal rectangular filter assumption, the margins of the integrals in 5 can simply be replaced by $-B_W/2$ and $+B_W/2$, where B_W is the double-sided bandwidth at the receiver.

By replacing eqns. 1, 3, and 4 into eqn. 5, we obtain a traking error variance as a function of the receiver weighting factor α. A numerical optimization has been performed, as shown in Fig. 4, and the optimum α factor (for the case when data-only channel is processed with a weighted combination of BOC(1,1)/BOC(6,1))

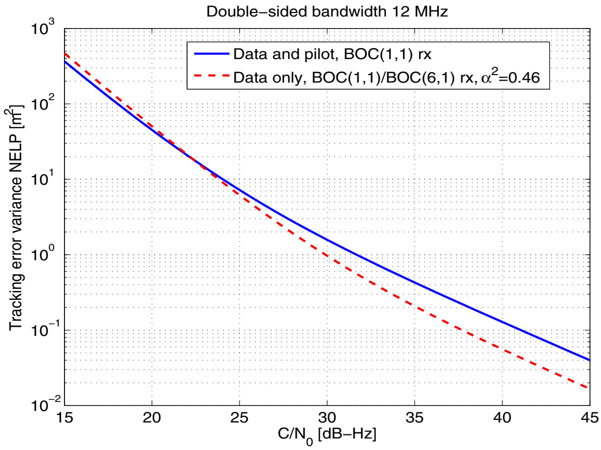

Fig. 6. Tracking error variance of the two dual-channel architectures at $B_W = 12$ MHz

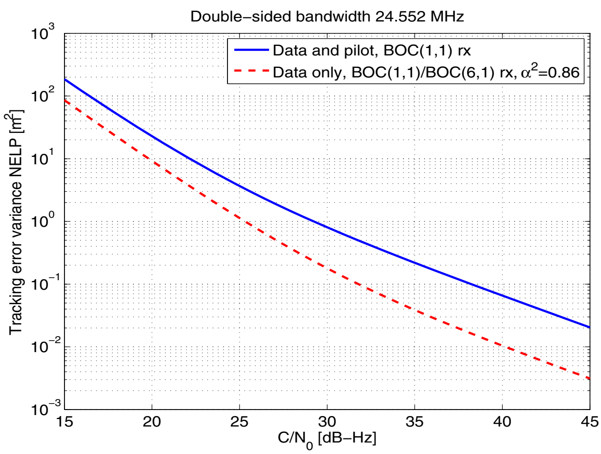

Fig. 7. Tracking error variance of the two dual-channel architectures at $B_W = 24.552$ MHz

was found to be bandwidth dependent. For double-sided bandwidths up to 12 MHz, the optimum processing is with a very weak or no sine BOC(6,1) component ($\alpha^2 = 0 - 0.02$). For sufficiently high bandwidths (i.e., larger than 12 MHz, a strong sine BOC(6,1) component was found beneficial (α^2 ranging from 0.46 at 12 MHz till 0.86 at 24.552 MHz).

The next question addressed here is whether processing the data and pilot together, both with a sine BOC(1,1) signal is better (in terms of tracking error variance) than processing the data channel alone with a weighted combination of sine BOC(1,1)/BOC(6,1). The optimal power weight α^2 found above will be considered in this comparison. Illustrative plots at two extreme double-sided bandwidths and a middle-case bandwidth are shown in Figs. 5, 6, and 7.

Clearly, from the tracking error variance point of view, architecture 1 is better that architecture 2 at low bandwidths (up to about 12 MHz double-sided bandwidth), while architecture 2 (which makes use also of the BOC(6,1) component) is better at higher bandwidths.

5 Multipath Behaviour

The cross-correlation shape at the receiver side can be computed via the inverse Fourier transform of the PSD $\overline{G}_s(f)$ given in eq. (1). The Multipath Error Envelopes (MEE) for a second path 3 dB weaker than the first one are shown

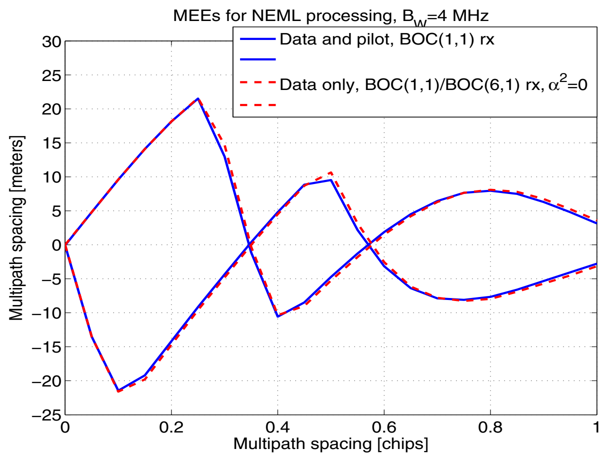

Fig. 8. Multipath Error Envelope (MEE) at $B_W = 4$ MHz

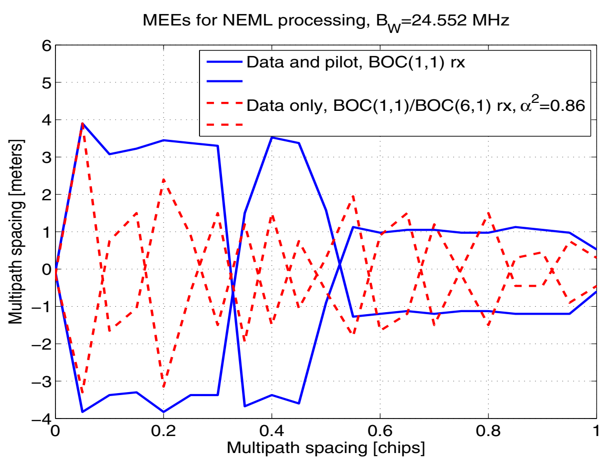

Fig. 9. Multipath Error Envelope (MEE) at $B_W = 24.552$ MHz

in Figs. 8 and 9, for double-sided bandwidths of 4 and 24.552 MHz. A MEE shows the robustness in the presence of multipaths [13], when the channel has two paths, spaced at the distance showed on x-axis of Figs. 8 and 9, and with the second path with 3 dB smaller than the first one. The double lines in the egend of the plots in Figs. 8 and 9 are due to the upper and lower multipath error envelopes. As seen in Fig. 8, at low bandwidths (i.e., narrowband receiver), there is no difference in the multipath performance between the two architectures. When the double-sided bandwidth increases (as seen in Fig. 9), the second architecture when data-only channel is processed via both BOC(1,1) and BOC(6,1) receivers, has a clear advantage also in terms of MEE, due to the higher resolution capability of BOC(6,1) component. Indeed, we can get few meters better multipath resistance with the joint BOC(1,1)/BOC(6,1) architecture of Fig. 2.

6 Conclusions and Design Issues

The processing of CBOC signal with 1-bit receiver and limited number of channels (e.g., two channels per satellite for E1 signals) can be done in two main architectures: either by processing both the data and and the pilot channels with a sine BOC(1,1) receiver (and ignoring completely the BOC(6,1) component) as shown in Fig. 1, or by processing only the data channel (needed to extract navigation data) and using both BOC(1,1) and BOC(6,1) correlations, as shown in Fig. 2. In the second approach, the power division between BOC(1,1) and BOC(6,1) components at the receiver should be optimized separately. Our studies show that, when the receiver double-sided bandwidth is less than about 12 MHz, the optimum processing is given by the first architecture: data and pilot are processed via BOC(1,1) component, and the BOC(6,1) component is completely ignored. The gain in terms of tracking error variance is about 3 dB (over the second architecture), as seen in Fig. 5. This is a completely intuitive result, since data and pilot powers are split in half at the transmitter, and we expect to lose 3 dB when processing data-only or pilot-only channels.

The interesting and novel result comes when higher bandwidths are available: in here, there is a tracking error variance performance gain when BOC(6,1) processing is also used (i.e., architecture of Fig. 2) and this performance gain can compensate for the 3 dB loss when only the data channel is processed. Indeed, according to the available receiver bandwidth, and by choosing properly the weighting factor between BOC(1,1) and BOC(6,1) components, as illustrated in Fig. 4, we can get up to 5 dB performance gain in the tracking error variance with the second architecture and also an enhacement in the multipath error bias (i.e., smaller multipath error envelopes).

Therefore, the design recommendations we adopt are bandwidth dependent: for a narrowband receiver as those employed in mass-market solutions, the classical architecture of processing data and pilot channels with a sine BOC(1,1) receiver is the best choice, while in a wideband receiver (here, meaning with the receiver double-sided bandwidths higher than 12 MHz), a processing of the CBOC signal with both BOC(1,1) and BOC(6,1) channels is the best option.

References

1. European GNSS (Galileo) Open Service, Signal in space interface control document (OS SIS ICD) (February 2010)
2. Rodriguez, J., Wallner, S., Hein, G., Rebeyrol, E., Julien, O., Macabiau, C., Ries, L., DeLatour, A., Lestarquit, L., Issler, J.: CBOC - An Implementation of MBOC. In: First CNES Workshop on Galileo Signals and Signal Processing, Toulouse, France (October 2006)
3. Hein, G., Avila-Rodriguez, J., Wallner, S., Betz, J.W., Hegarty, C., Rushanan, J., Kraay, A., Pratt, A., Lenahan, S., Owen, J., Issler, J., Stansell, T.: MBOC: The New Optimized Spreading Modulation Recommended for GALILEO L1 OS and GPS L1C. In: Inside GNSS - Working Papers, vol. 1(4), pp. 57–65 (2006)
4. Julien, O., Macabiau, C., Avila Rodriguez, J.-A., Wallner, S., Paonni, M., Hein, G.W., Issler, J.-L., Ries, L.: On Potential CBOC/TMBOC Common Receiver Architectures. In: Proc of. ION GNSS 20th International Technical Meeting of the Satellite Division, Fort Worth, TX, pp. 1530–1542 (September 2007)
5. Lohan, E.: Analytical performance of CBOC-modulated Galileo E1 signal using sine BOC(1,1) receiver for mass-market applications. In: Proc. of IEEE Position Location and Navigation Symposium, PLANS (May 2010)
6. Artaud, G., Ries, L., Dantepal, J., Issler, J., Grelier, T., Delatour, A.: CBOC performances using software receiver. In: 2nd Workshop on GNSS Signals & Signal Processing - GNSS SIGNALS 2007, Nordwijk, Netherlands (October 2007)
7. Macabiau, C., Issler, J.-L., Ries, L., Julien, O.: Two for one: tracking Galileo CBOC signal with TMBOC. Inside GNSS journal (Spring 2007)
8. Borio, D., Mongredien, C., Lachapelle, G.: Collaborative code tracking of composite gnss signals. IEEE Journal of Selected Topics in Signal Processing 3, 613–626 (2009)
9. Dierendonck, A.V., Fenton, P., Ford, T.: Theory and performance of narrow correlator spacing in a GPS receiver. Journal of the Institute of navigation 39, 265–283 (1992)
10. McGraw, G., Braasch, M.: GNSS multipath mitigation using high resolution correlator concepts. In: Proc. of ION National Technical Meeting, San Diego, CA, pp. 333–342 (January 1999)
11. Lohan, E.S., Lakhzouri, A., Renfors, M.: Binary-Offset-Carrier modulation techniques with applications in satellite navigation systems. Wiley Journal of Wireless Communications and Mobile Computing (July 2006), doi:10.1002/ wcm.407
12. Betz, J., Kolodziejski, K.: Extended theory of early-late code tracking for a bandlimited GPS receiver. Journal of Navigation 47(3), 211–226 (2000)
13. Avila-Rodriguez, J., Wallner, S., Hein, G.: How to Optimize GNSS Signals and Codes for Indoor Positioning. In: ION GNSS 19th International Technical Meeting of the Satellite Division (September 2006)