Exploiting Orthogonality in DVB-S2X Through Timing Pre-Compensation

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***Abstract*—The superframing option of the recent DVB-S2X standard specifies, for the so-called SF-Pilot fields, the use of the orthogonal set of Walsh-Hadamard (WH) sequences. In order to exploit this orthogonality, waveforms coming from different beams to the k-th User Terminal (UT) should be quasi perfectly aligned in time. While in the downlink part of a terrestrial system this is quite straightforward, in satellite system, having a satellite as a relay, this is not the case, especially when large baudrates are considered in the transmission. A procedure to compensate for timing misalignment amongst waveforms is here presented and the advantages are quantified through numerical simulations.In particular, Channel State Information (CSI) estimation errors, which are fundamental for precoding techniques, are evaluated. While the focus of the work is on systems which enable precoding techniques, the procedure can be applied in each scenario which uses the superframing structure of DVB-S2X in an interference limited scenario.**

I. Introduction

In recent years, the growing interest in high data rates broadband services through high throughput satellite systems is fostering the investigation of enabler techniques which can guarantee a boost of the overall spectral efficiency with a reasonable complexity. In this context, multi-beam system architectures supported by an aggressive frequency re-use of the total bandwidth in many beams is essential when a bandwidth limited scenario is considered. When the focus is on the forward link of a satellite system as in the current work, signal processing techniques, namely precoding, are implemented to account for the mitigation of the interference at the UTs.

In order to construct the pre-coding matrix, it is foreseen that each terminal provides estimates of the channel parameters for all of the detectable received signals, i.e., the terminal specific waveform plus all of the detectable interferers [1], [10]. This requires that a terminal synchronizes not only onto its own signal, but also onto the detectable interferers. DVB-S2X superframe structure, defined in Annex E of [6], is, by design, the framing structure which enables the possibility of estimating the CSI in a satellite system [7].

While DVB-S2X relies on the use of Walsh-Hadamard [8] sequences in the SF-Pilots defnition, the advantages provided by orthogonal cross correlated codes [9] requires the transmitting waveforms to be perfectly synchronized in time, expecially when a scrambling sequence is used as it happens in satellite links. What is more important, the satellite payload introduces a timing misalignment amongst different transmitting antennas due to the different group delays (and paths) of the transponder filters. In general, the problem of the waveforms alignment is important not only for CSI estimation but also for precoding techniques which assume the transmitting waveforms to be quasi-perfectly synchronized [11].

A procedure to pre-compensate the timing misalignments introduced by the payload, which is presented in the current work, is foundamental to both increase the quality of the CSI estimations and to avoid performance degradation in the precoding process.

The paper is strctured as follow. In Section II the reference system is modelled, in Section III the impact of the timing misalignment wrt the Superframing option of the DVB-S2X is studied, then, in Section IV, the timing pre-compensation stategy is described and some results on the timing estimation procedure are shown, and finally in Section V a comparison in terms of CSI estimation between the pre-compensated and the non pre-compensated cases is evaluated with numerical results.

II. System model

We consider the forward link of a multi-beam satellite system when an aggressive frequency reuse is applied. In Figure 1, a pictorial representation of the system model is shown.

According to this model, the received analog signal at the input of the UT antenna, $y(t)$, at the generic k-th UT can be modeled as

$$y_k(t) = \sum_{n=1}^{N} r_{kn}(t) + n_k(t) \tag{1}$$

where $r_{kn}(t) = \hat{h}_{kn}(t)x_n(t)$ is the analog waveform received by the k-th terminal from the n-th beam with $\hat{h}_{kn}(t)$ the channel coefficient containing also the satellite impairments, $x_n(t)$ the transmitted information signal, and $n_k(t)$ the AWGN noise at terminal k. The number of beams is Nt while the number

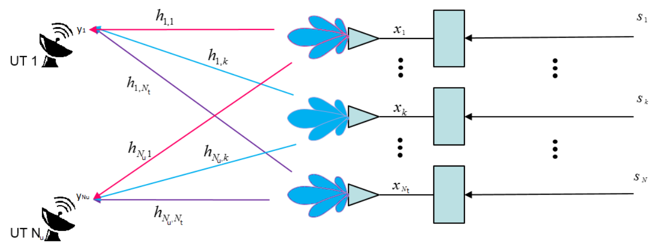

Figure 1: Block Diagram of the System Model for a multi-beam satellite system

of UTs is Nu. Each UT reference waveform is hence affected by the Signal-to-Noise-plus-Interference (SNIR) expressed in the following:

$$\mathrm{SNIR} = \frac{C}{N+I} = \frac{1}{\left(\frac{C}{N}\right)^{-1} + \left(\frac{C}{I}\right)^{-1}} \tag{2}$$

where C, N and I are respectively the reference signal, the noise and the interference power at the UT without any interference management techniques. Based on (1) and (2), the previous can be re-written as

$$\mathrm{SNIR}_k = \frac{|h_{kk}|^2}{\sigma^2 + \sum_{\substack{n=1\\n\neq k}}^{Nt} \mid h_{kn} \mid^2} = \tag{3}$$
$$= \frac{1}{\left(\frac{|h_{kk}|^2}{\sigma^2}\right)^{-1} + \sum_{\substack{n=1\\n\neq k}}^{Nt} \left(\frac{|h_{kk}|^2}{|h_{kn}|^2}\right)^{-1}}$$

assuming that the average power of the transmitted symbols is normalized to one. This formula describes the SNIR of the reference waveform of the k-th UT. In the latter formulation it is clear that the SNIR is composed by two different quantities which are the SNR of the reference waveform and the summation of the C/I values for all the interferer beams. As a consequence, we can call as *"noise limited"* the region in which the impact of the SNR is much stronger than the impact of the interferers (which is for low SNR values), as *"interference limited"* the extreme opposite region where the behaviour of the SNIR depends mostly on the $\sum (C/I_i)^{-1}$ (which is for low C/I values) and an intermediate region where both components have an impact.

In the present work we focus the attention on the *"interference limited"* region (including the intermediate) since our goal is to report advantages in having aligned waveform which, in DVB-S2X, leads to exploiting orthogonal properties of SF-pilot sequences. A summary of the structure of the DVB-S2X pilots, which clarify the previous sentence, can be found in the next section.

While in terms of useful data to be detected and decoded the interest is only on the reference waveform, when the generic UT wants to calculate the CSI vector of all the possible detectable waveforms, both the reference and the interferer waveforms are useful data depending on the specific channel matrix coefficient to be estimated. As an example, when the k-UT is interested in computing the CSI for the channel h_{km}, the useful signal for the channel estimation algorithm is the m-th, hence the SNIR of the m-th waveform of the k-th UT is given by:

$$\mathrm{SNIR}_k^m = \frac{\mid h_{km} \mid^2}{\sigma^2 + \sum_{\substack{n=1\\n\neq m}}^{Nt} \mid h_{kn} \mid^2} = \tag{4}$$
$$= \frac{1}{\left(\frac{|h_{km}|^2}{\sigma^2}\right)^{-1} + \left(\frac{|h_{km}|^2}{\sum_{\substack{n=1\\n\neq m}}^{Nt} |h_{kn}|^2}\right)^{-1}}$$

The formulation is very similar to the previous one but it highlights another important information: the SNR that a weaker interferer (wrt the reference waveform) experiences is lower, hence, the impact of the *"noise limited"* region is higher.

In the following subsections, some consideration on the satellite channel model and on the received signal model at the UT due to the impairments are discussed.

A. Channel model

In this analysis, we consider a simplified but representative model of the satellite payload. The main sources for vanishing the original built DVB-S2X orthogonality (see Sec. III) are the carrier frequency and phase variations, as well as the time misalignments among the transmitted signals. To minimize the impact of frequency/phase offset, it is assumed a satellite payload architecture where a very stable master reference local oscillator is present for driving the frequency conversion stages of all transmitted signal chains. As far as the time misalignment is concerned, it is realistic to consider that the impact of different lengths of signal paths jointly to variations in the group delay function of the output filters can be limited into few nanoseconds (i.e., < 4 ns). [7]

B. Received signal model with impairments

Due to the impairments described in [7], the discrete time model (digital signal, not analog as in (1)) for the received composite signal at the k-th UT after the Analog-to-Digital conversion, which consists of the superposition of waveforms transmitted through N different interfering beams, is modeled as:

$$y_k[m] = \tag{5}$$
$$\mathrm{e}^{-j\left(2\pi\left(f_o + \frac{1}{2} f_d m\right) mT + \phi[m]\right)} \cdot$$
$$\sum_{n=1}^{N} h_{kn}[m] x_n \left[m - \tau_{kn} - \tau_d m\right] \mathrm{e}^{-j\left(2\pi \Delta f_{kn} mT + \theta_{kn} + \phi_{kn}^P[m]\right)} +$$
$$+ n_k[m]$$

where $x_n[m]$ is the waveform component received from the n-th antenna feed, and $h_{kn}[m]$ is the complex channel coefficient from the n-th feed to the k-th user, τ_{kn}, Δf_{kn} and θ_{kn} are respectively the time, frequency and phase offsets of the n-th received waveform at the k-th terminal. Moreover, due to the characteristics of the receiver, the composite signal is affected by the following common impairments which are time and frequency drift τ_d, f_d, frequency offset f_o and phase noise $\phi[m]$. In addition, $\phi_P[m]$ is the differential phase noise given by the payload. This won't be considered in this analysis because it can be considered negligible in one superframe

III. Orthogonality in DVB-S2X

In this section, the analysis of the SF-Pilots structure, which are composed by a WH sequence as well as a scrambling sequence that changes the properties of the sequences in case of time misalignment, is tackled. The impact of the timing misalignment amongst waveforms on the estimation algorithms is also studied through simulations. The mathematical formulation that relates the timing alignment to the orthogonality of the set of sequences is also reported.

A. Analysis of the DVB-S2X SF-Pilots

The Annex E of the DVB-S2X standard [6] specifies the optional framing structure which enables the use of precoding techniques for the forward link of a satellite system. Basically, the superframe structure fixes the use of:

1) A constant superframe length of 612540 symbols
2) A preamble of 720 symbols which includes the Start of Super-Frame (SOSF) and the Super Format Frame Indicator (SFFI)
3) 2 different scrambling sequence definitions (Gold codes) for the so-called reference and payload data scramblers

The super-framing structure is introduced for multiple purposes, in particular to increase resilience to co-channel interference by other beams thanks to the use of different scrambling sequences and to support synchronization algorithms thanks to the fixed framing structure.
In the following, the main and general features of the superframe structure related to the synchronization procedures are listed:

- in order to take advantage from orthogonality properties in case of perfect alignment amongst waveforms (signals from different beams), Walsh-Hadamard sequences are used for SoSF and Pilot fields;
- Pilot field is composed by the concatenation of 2 Walsh-Hadamard sequences, respectively of 32 and 4 symbols for the long PLFrame, and 32 and 16 symbols for the short PLFrame;

The SFFI field specifies the format of the superframe. For precoding purposes, the format specifications to be considered are the Bundled PLFRAMES (64800 payload size) with SF-Pilots and the Bundled PLFRAMES (16200 payload size) with SF-Pilots. For the two specified formats, different pilot structures are defined. In the following, a brief description of the structures is reported.

1) Bundled PL-Frames - Format 2: In Figure 2, the structure of the Bundled Long PL-Frame is shown. The main characteristics to keep in mind for the preset work are:

- pilot field length is 36 symbols
- the number of pilot fields in 1 superframe is 639
- the distance between two consecutive pilot fields is 956 symbols (including 36 pilot symbols)
- the superframe consists of 9 Bundled PLFrames of length 64800 symbols

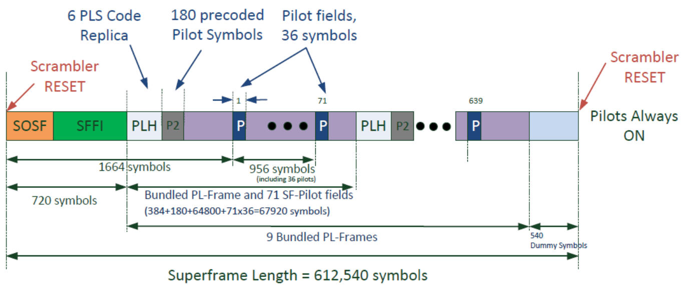

Figure 2: Structure of the Superframe Bundled PLFrame with Precoded and not Precoded Pilots - Format 2 [6]

2) Bundled PL-Frames - Format 3: The main characteristics to keep in mind for synchronization purposes are:

- pilot field length is 48 symbols
- the number of pilot fields in 1 superframe is 324
- the distance between two consecutive pilot fields is 1887 symbols (including 48 pilot symbols)
- the superframe consists of 36 Bundled PLFrames of length 16200 symbols

For the sake of the synchronization and estimation performance, it is fundamental to note also that, when precoding is enabled:

- SOSF and SF-Pilots are not-precoded;
- SOSF and SF-Pilots consist of beam-specific orthogonal Walsh-Hadamard sequences;
- two different scrambling sequences are applied to the superframe: the first sequence, the so called *reference data scrambler*, is the same for all beams, it overlays only the SoSF and SF-Pilots, and is restarted at each Start of Super Frame. The second scrambling sequence, the so called *payload data scrambler*, is beam dependent, overlays the data payload, and provides resilience to co-channel interference.
- the superposition of the beam-specific Walsh-Hadamard sequences in the SOSF and SF-Pilots and the *common Reference Data Scrambler* yields a unique beam-specific *signature* that can be used for waveform/beam identification

There are two main reason why we have orthogonality only in the alignment case:

- the scrambling sequence ontop essentially modifies the caracteristic of the sequences in case of delay different from 0
- when the oversampled correlation is considered, even fully orthogonal sequences show a correlation values equal to 0 only for kT_s when k is integer and T_s is the symbol period.

B. Impact of time misalignment of Bundled frames on Channel Estimation

As previously described, an important impairment to be taken into account for channel estimation purposes is the time misalignment between waveforms at the UT. Considering an interference limited scenario due to an aggressive frequency reuse amongst beams, known-symbols assisted algorithms provide advantages in distinguishing beam-specific waveforms from the superimposed signal at the UT. Pilot-Aided algorithms are strictly influenced by the correlation properties of the set of sequences used. In case of time alignment, as in a downlink scenario of a wireless system, the use of a set of orthogonal sequences, as the Walsh-Hadamard set, is a preferable choice for it provides very good cross- correlation properties between sequences under some constraints. On the other hand, the autocorrelation properties are not suitable for detection purposes due to the presence of several sidelobes in the autocorrelation function [8]. The use of a scrambling sequence on top which operates symbol by symbol, shared between all signals (as in DVB-S2x Annex E framing structure described in [6]) is very helpful in terms of autocorrelation function and in terms of frequency spectrum to avoid spectral lines, but it limits orthogonality to the case of perfect alignment only. In the following we want to verify this last statemet by comparing the correlation functions amogst 2 sets of sequences: the Walsh-Hadamard only and the resulting sequence given by the product of the Walsh-Hadamard and the scrambling sequence.

The normalized correlation function between two sequences is given by the well-known formula:

$$R_{ij}[n] = \frac{1}{N_{Pil}} \left| \sum_{m=0}^{Npil-1} o_i[m] o_j^*[m-n] \right| \tag{6}$$

where N_{pil} is the length of the sequences, o_i is an orthogonal Walsh-Hadamard sequence and $*$ represent the complex conjugate. This correlation function depends on the properties of the set of sequences which, for the case of Walsh-Hadamard set, is basically having a cross-correlation equal to 0 when $n = 0$. There exists also a subset of the WHs which has a cross-correlation value of the circular cross-correlation equal to 0 for each integer delay [8] in terms of symbols (hence only when not-shaped/not-oversampled sequences are considered) but these considerations are not valid in this case since SF-pilots are not repeated continuosly and the circular correlation is not representative of the correlation which occurs at the receiver.

When the scrambling sequence $g[m]$ is considered, a new set of sequences should be considered which is given by $c_i[m] = g[m]o_i[m]$,where, according to the Superframing description, **g** is the same for all the waveforms. Another property of the scrambling sequence, as it is defined in the [6] is that $g[m]g^*[m] = 1$. If we substitute the new sequences in the equation we obtain:

$$R_{ij}[n] = \frac{1}{N_{Pil}} \left| \sum_{m=0}^{Npil-1} c_i[m] c_j^*[m-n] \right| = \tag{7}$$

$$\frac{1}{N_{Pil}} \left| \sum_{m=0}^{Npil-1} g[m] o_i[m] g^*[m-n] o_j^*[m-n] \right| =$$

From this formulation it can be noticed that, when $g[m]g^*[m-n] = 1$ for each m in the range $m = 0, .., N_{pil}$, the correlation function is exactly the same as equation 6 and this happens when $n = 0$, hence, when the two sequences are time aligned. To demonstrate with some numerical results the formulation, in Figure 3 it is shown the superposition of the oversampled correlation functions between one selected sequence and all the other possible combinations. In should be specified that, to consider the effects of the payload data symbols on the correlation, which is a more realistic case, the corelation functions are calculated as follows:

$$R_{ij}[n] = \frac{1}{N_{Pil}} \left| \sum_{m=0}^{Npil-1} r_i'[m+n] c_j^*[m] \right| \tag{8}$$

where $n = [-920, -920 + 1/ns, 0, ..., 955 - 1/ns, 955] \in \mathbb{R}$, $m \in \mathbb{N}$ and ns is the oversampling factor. r_i' is the i-th received stream given by the successive concatenation of three vectors which are:

$$r_i'[m] = \begin{cases} x1_i[m] & \text{when } -920 \le m < 0 \\ c_i[m] & \text{when } 0 \le m < N_{pil} - 1 \\ x2_i[m] & \text{when } 955 \ge m \ge N_{pil} \end{cases} \tag{9}$$

In the latter definition, both $x1$ and $x2$ are random data symbols having QPSK modulation for the sake of simplicity. The figure clearly shows that the orthogonality between sequences happens in case of perfect alignment only since all the correlation values in delay 0 are equal to 0. The red curve which has a peak in delay 0 is of course the autocorrelation function R_{ii}. This justifies the importance of pre-compensating the timing misalignment due to the satellite payload.

When we consider a symbol rate of 500 MBaud, the time misalignment between signals has an impact on the correlation values, hence, it has an impact on the estimation performance. A maximum misalignment of ± 3 symbols is the worst case for such a symbol rate, which all users of the same group will experience. In fact, having a variation ($\pm$ standard deviation) in the group delay function as stated in section II-A limited to 4 nanoseconds, it is reasonable to consider, as a worst case, 3 times the standard deviation variation as the maximum timing misalignment, which corresponds to $3(\pm\sigma)Rs = \pm 3 symbols$.

IV. PROPOSED STATEGY BASED ON TIMING ESTIMATIONS AT EACH UT

The proposed technique for the pre-compensation of the timing misalignments amongst waveforms is based on a feedback channel between UTs and GW. Since the time misalignment is due to the payload (hence we can assume fixed values

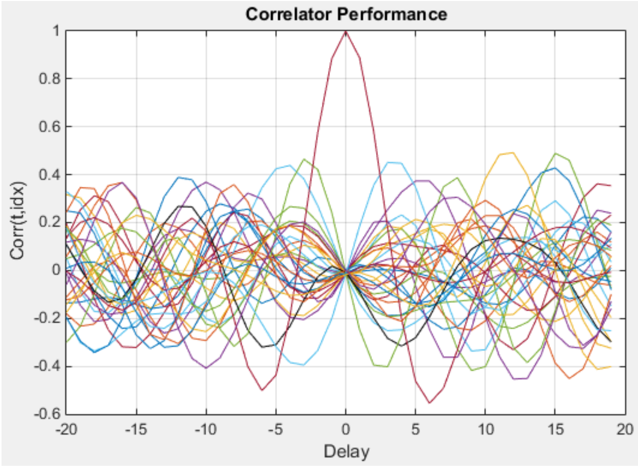

Figure 3: Superposition of all oversampled correlation functions amongst SF-Pilots. Delays are expressed in terms of kT_s/T_s where T_s is the symbol period and k the x-axis

for timing misalignments or at least manageble), the round trip delay of a satellite link is not an issue. Due to this, the pre-compensation procedure can be also seen (if needed) as a calibration procedure for the system. It is worth noting that, while in the following formulation we assume the use of precoding techniques at the trasmitted side, the proposed solution is very general and valid also in case this assumpion is alleviated. In the latter case, the precoded part of the following formulation should be removed while the non-precoded part is still valid.

Each UT provides an estimate of the time misalignments amongst waveforms based on the timing of the reference waveform. The i-th waveform to be transmitted by the GW is composed by two different parts, which are the non-precoded symbols and the precoded symbols. These two parts have clearly two different formulations. We define as P the space composed by the position of the SoSF and the SF-Pilots in the superframing structure while D is the complementary space for precoded symbols which are data, PLH and P2:

$$x_i[m] = \begin{cases} c_i[m] & \text{when } m \in P \\ \sum_j s_j[m]w_{ij} & \text{when } m \in D \end{cases} \tag{10}$$

where s_j is the data symbol from j-th stream and w_{ij} is the precoding weight. Due to the payload, as described in Section II, the transmitting waveforms are affected by a differential timing delay:

$$x_i[m+\tau_i] = \begin{cases} c_i[m+\tau_i] & \text{when } m \in P \\ \sum_j s_j[m+\tau_i]w_{ij} & \text{when } m \in D \end{cases} \tag{11}$$

The waveforms are then transmitted by the feeds on the i-th beam. After the channel matrix, the k-th superimposed stream at the k-th UT coming from the i-th beam is:

$$r_k[m] = \begin{cases} \sum_i c_i[m+\tau_i]h_{ki} & \text{when } m \in P \\ \sum_i \left(\sum_j s_j[m+\tau_i]w_{ij}\right) h_{ki} & \text{when } m \in D \end{cases} \tag{12}$$

which represents the received signal given by the superposition of misaligned waveforms, both in the non-precoded and precoded cases. We have already addressed the issue related to the loss of orthogonality of the non-precoded part which has an impact on the Pilot Aided (PA) algorithms at the receiver. What is more important is also the impact of the loss of orthogonality for precoded waveforms. In fact, if we highlight the part related to the useful symbol j from the precoded formulation, we obtain:

$$r_{kj}[m] = \sum_i s_j[m+\tau_i]w_{ij}h_{ki} \text{ when } m \in D \tag{13}$$

In case of timing misalignments, τ_i are different per various i. This introduces an amplitude variation of the same symbol which depends on the timing misalignment and the shaping filtering. It is clear that a pre-compensation technique is not only needed to exploit orthogonality in the CSI estimation but also to avoid performance degradations for precoding techniques, which assume perfect synchronization [11]. Assuming that a feedback channel from the UTs to the GW is available, a set of timing recovery algorithms, one for each detected waveform at the UT, can estimate a vector of $\hat{\tau}_i$ values for pre-compensation at the GW. It is mandatory that the timing pre-compensation is applied after precoding and after the shaping filtering as showed in Figure 6. When we applied the pre-compensation we obtain the following trasmitting waveforms at the GW:

$$x_i[m-\hat{\tau}_i] = \begin{cases} c_i[m-\hat{\tau}_i] & \text{when } m \in P \\ \sum_j s_j[m-\hat{\tau}_i]w_{ij} & \text{when } m \in D \end{cases} \tag{14}$$

Due to timing misalignments introduced by the payload, the waveforms transmitted at each satellite feed is:

$$x_i[m-\hat{\tau}_i+\tau_i] = \begin{cases} c_i[m-\hat{\tau}_i+\tau_i] \\ \sum_j s_j[m-\hat{\tau}_i+\tau_i]w_{ij} \end{cases} \tag{15}$$

We define $\hat{\tau}_i = \tau_i + \hat{\epsilon}_i$ where $\hat{\epsilon}_i$ the residual uncertainty that the timing recovery algorithm introduce in the estimate.

$$x_i[m-\hat{\epsilon}_i] = \begin{cases} c_i[m-\hat{\epsilon}_i] \\ \sum_j s_j[m-\hat{\epsilon}_i]w_{ij} \end{cases} \tag{16}$$

Clearly, depending on the value of the performance of the timing recovery algorithm and on the technique used to calculate the value from the estimate coming from all UTs in the same group and UTs in the foreign beam (which are able to detect the same waveforms), the obtained values of $\hat{\epsilon}_i$ can be very close to 0. While in the present work an averaging on the residuals from different UTs is applied, future works will be concentrated on finding an optimal technique to minimize the estimation error based on all the estimates provided by all UTs. When the estimated errors are random amongst users, the averaging operation on the various estimates can guarantee a quasi perfect alignment and hence orthogonality.

A. *Timing recovery algorithm at the receiver*

The synchronization/channel estimation procedure used in the present work is the one described in [1], [7]. After the frequency acquisition and the frame synchronization procedure, the time recovery for each user is required to compensate the timing offset of the superimposed signal.
Due to the received signal behavior, which is composed by several DVB-S2X signals, to estimate time misalignments between users, a PA algorithm is preferable. We choose to use a variant of the Early Late Gate (ELG) PA algorithm, one for each detectable user, which works on the autocorrelation function instead of the symbols. A block diagram of the algorithm is shown in Figure 4.

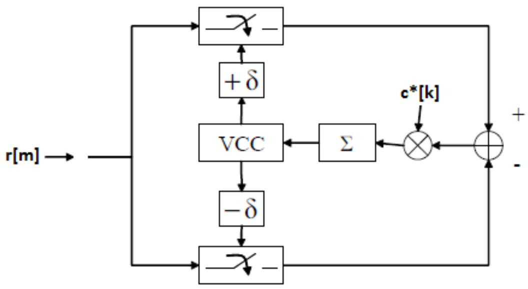

Figure 4: Block diagram of the ELG PA algorithm used in the synchronization chain

At first, the algorithm performs two re-sampler operations on the signal, one at δT_s and the other at $-\delta T_s$. These two signals are then subtracted and correlated with the known pilots at the receiver. The difference in the correlation value obtained between the two branches of the algorithm in Figure 4 is what it is basically driving the update via the δ value to be used. It is clear that this is an adaptive algorithm which depends on the δ value chosen (which represents an initial condition for the algorithm) and on the step size chosen to update the δ value at each iteration. The smaller the step size, the slower the convergence time, the better the accuracy. This step size, that we call γ, regulates the loop bandwidth of the loop filter used for the time tracking algorithm. It is worth noting that all these considerations are still valid in case the system is configured to feed-back information to the GW every Bundled frame instead of every superframe.
In the following, the performance of the timing recovery algorithm when the timing misalignment amongst waveforms is affecting the superimposed signal is presented. For the simulations, the receiver algorithms and impairments described in section C.5.3 of [7] are used. The reference scenario is given by the following interference distribution: $C/I_i = [0\ 4\ 8\ 12\ 16]dB$. Results show that, even for waveform I_4, which is the one corresponding to a C/I value of $12dB$, the mean and the standard deviation of the estimation error is very close to 0, hence, the timing estimation for misaligned waveforms can be accurately estimated through the PA-ELG algorithm.

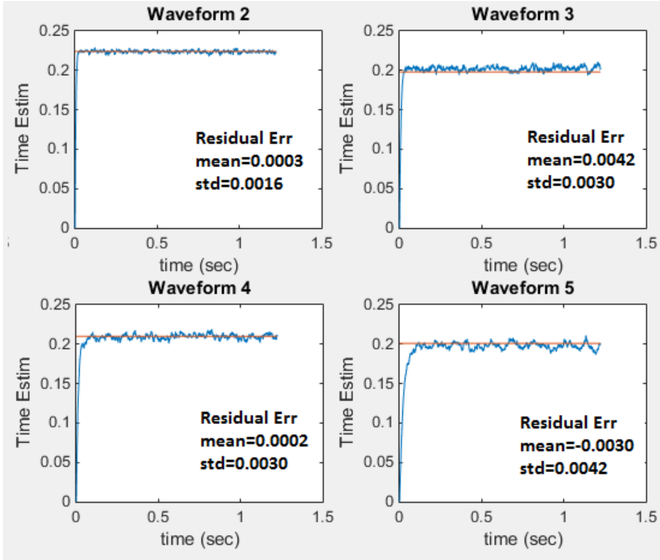

Figure 5: Timing Estimation results for waveforms I_1,I_2,I_3 andI_4 in case of misalignment (the error is normalized to the symbol time)

B. *Calculation of the compensation values*

In a satellite scenario, where the number of users is much more than the number of antennas, the computation of the statistics could have an impact on the computational complexity when each user has to feedback a vector of timing estimates related to different waveforms. Using various estimates by the same UT (successive Superframes) and several timing estimations from different UTs (the timing misalignments are fixed), leads to a very accurate time misalignment pre-compensation through an averaging operation.
After the calculation of the relative time misalignments amongst waveform, the final step is to pre-compensate the misalignments. It is quite straightforward that the compensation through polynomial interpolation or spline interpolation is done after the computation of the precoded-waveforms as shown in Figure 6. There is more than one reason for this:

- to compute the interpolation it is required that the transmitting streams are oversampled;
- as a general rule, whatever impairment that affects the waveforms between the precoding matrix W and the channel matrix H should be precompensated after the application of the precoding matrix to the waveforms.

The aim of the *"Calculation Pre-Compensation values"* is to collect all the estimates from all the UTs and to compute an averaging operation on the correspondent timing estimations. Having all the estimates from all the UTs, the GW is able to compute the timing misalignments for all the waveforms to be transmitted. It is worth mentioning that, in case a specific UT is not able to estimate the timing misalignment of the weakest interferer waveform, another UT, which faces better SIR condition for that waveform, can estimate it.

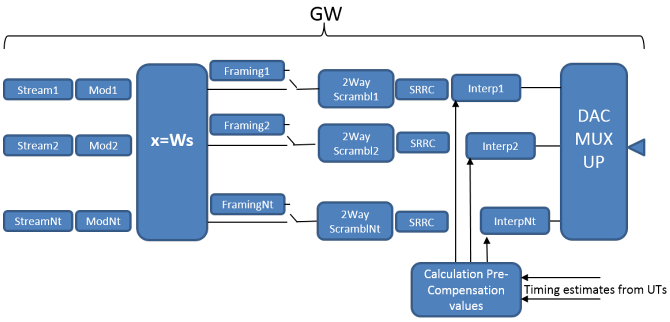

Figure 6: Timing Pre-compensation at the GW: block scheme

V. Numerical results

In the following, we report the numerical analysis and performance assessment considering the receiver algorithms (synchronization and channel estimation) described in [1], [7]. The results can be divided into two stages: in the first part, the timing estimation performance in presence of timing misalignment amongst waveforms (no orthogonality) are shown, then, results obtained in terms of CSI estimation errors for the cases of pre-compensated and not pre-compensated waveforms are reported. For this test, we first consider the frequency acquisition procedure in case of 500 Mbaud for the symbol rate to be considered, then we deal with the frame synchronization and verification procedure with particular attention on the timing drift compensation done using SoSF fields, after which the timing estimation can be assessed and tested, whose results are provided, and then the channel state information estimation results are presented. It is important to specify that the residuals from the previous estimation algorithms are included in the simulations.
For the numerical results to be provided, we consider, at the receiver side, a superposition of six waveforms given by the reference waveform, i.e. the only one to be decoded and five interferer waveforms. Assuming that C is the power of the reference waveform and I is the power of the considered interferer waveform, the six waveforms have the following *C/I* distribution (to uniformly test interferers performance):

$$\frac{C}{I_i} = [0\ 4\ 8\ 12\ 16]dB$$

The algorithm used for the channel estimation procedure is a Pilot Aided algorithm described by the following formula:

$$\hat{h}_i = A_i e^{j\varphi_i} \tag{17}$$

$$A_i = \frac{1}{N_{Pil} L_{Pil}} \sum_{k=1}^{N_{Pil}} \left| \sum_{j=1}^{L_{Pil}} y_k^p[j] c_{ki}^*[j] \right| \tag{18}$$

$$\varphi_i = \frac{1}{N_{Pil} L_{Pil}} \sum_{k=1}^{N_{Pil}} \angle \sum_{j=1}^{L_{Pil}} y_k^p[j] c_{ki}^*[j] \tag{19}$$

where $\angle$ is the angle function, $\hat{h}_i$ is the estimate for the $i-th$ beam, A_i and φ_i are respectively the amplitude and phase estimates for the $i-th$ waveform, N_{Pil} and L_{Pil} are the number of pilot fields (the number of consecutive pilot blocks over which the estimate is averaged) and the length of the pilot fields, $y_k^p[j]$ is the portion of the received signal corresponding to the $k-th$ block of the transmitted pilots within the Superframe and $c_{ki}^*[j]$ is the beam specific sequence (composed by a beam specific Walsh-Hadamard sequence and a scrambling sequence).

A comparison in terms of CSI erros for time aligned and time misaligned waveforms is shown and described.

The simulation parameters used in the channel estimation procedure are reported in the following:

- Rs = 500 MBaud;
- Roll-off = 0.05;
- Oversample = 4;
- C/I_n = [0 0 4 8 12 16] dB (where the first value is the reference waveform power);
- Time misalignments = [-3Ts ; +3Ts];
- Negligible frequency misalignments;
- Phase misalignments in the range $[-\pi/2; \pi/2[$;
- Residual from frequency estimation: Gaussian r.v. having standard deviation = 0.0003 Rs;
- Residual from timing estimation: Gaussian r.v. having standard deviation = 0.036 Ts;
- Residual from phase estimation for Phase Noise: Gaussian r.v. having standard deviation = Cramér Rao Bound (orthogonality provide performance very close to the CRB for the single user phase estimation);
- SNR (wrt the reference waveform) = 0-10dB;
- L_{Pil} = 32 symbols;
- N_{Pil} = 639 consecutive pilot fields (long Bundled Frame);

In Figure 7, results obtained in terms of CSI amplitude mean and standard deviation errors in the case of misaligned and aligned (pre-compensated) scenario are shown versus the C/I value of the specific waveform. These values are calculated, starting from the amplitude errors obtained, by the following formula ($N_{iter} = 10000$):

$$A_{errdB} = 20 * log_{10}\left(\frac{1}{N_{iter}} \sum_{i=1}^{N_{iter}} \frac{A_{lin} + \epsilon_i}{A_{lin}}\right) \tag{20}$$

where A_{lin} is the value of the waveform amplitude in linear units and ϵ is the error from the estimation of the amplitude. While the solid lines are the mean values, the respective (same colour) dashed-dotted lines represent the window given by the standard deviation wrt the mean value, meaning that those curves are calculated as mean+standard deviation and mean-standard deviation. The orange and the red curves are the results obtained respectively in case of SNR (reference user) equal to 0dB and 10 dB for misaligned waveforms. On the other hand, dark blue and light blue are the results obtained respectively in case of SNR equal to 0dB and 10 dB for aligned waveforms. It is quite straightforward to notice that, for all curves, increasing the C/I the error is larger, which is quite obvious. By comparing the misaligned and the

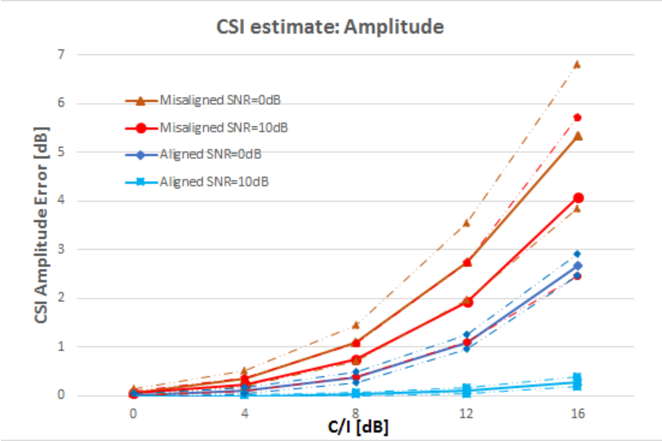

Figure 7: CSI amplitude errors (mean and standard deviation) in case of timing misalignment and alignment for different SNR values. The dashed lines of the same colous specify the window of the standard deviation wrt the mean value.

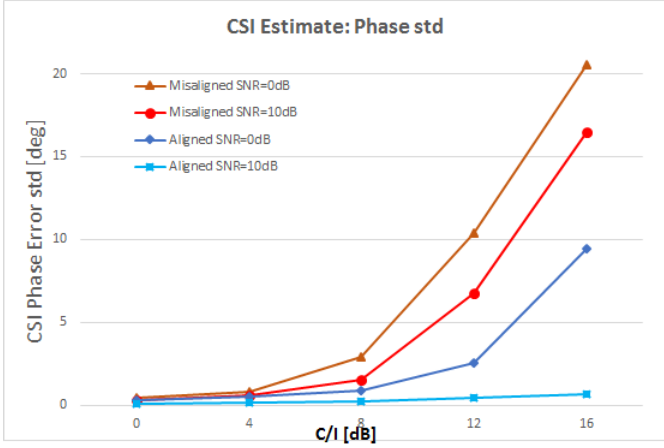

Figure 8: CSI phase errors (standard deviation) in case of timing misalignment and alignment for different SNR values.

aligned cases, it is also very clear that the pre-compensation at the trasmitter side provides huge gains in term of both CSI errors and accuracy of the estimation since both the mean and the standard deviation values are much larger in the misaligned case. In addition, it is worth noting that, considering a $C/I = 16dB$, there are about 2dB of difference between the mean values of the aligned and misaligned cases when the SNR of the reference waveform is 0dB while this difference increases when the considered SNR is 10dB. These results are in line with the considerations on the *"noise limited"* and *"interference limited"* regions since the weakest waveform is more affected by noise and the interference impact is smaller but still persistent.

In Figure 8, results obtained in terms of CSI phase standard deviation errors in the case of misaligned and aligned (pre-compensated) scenario are shown versus the C/I value of the specific waveform. The colour scheme is the same as in Figure 7 for the sake of clarity. The mean values are here not reported since they are very small compared to the standard deviation. As it happened for the amplitude case, the advantages of using the pre-compensation at the transmitter are quite huge, expecially considering that, for the weakest interferer, the standard deviation reaches values of 20 degrees, which is a not acceptable value.

VI. Conclusion

In this work, a timing pre-compensation procedure for the forward link of a multi-beam satellite system employing an aggressive frequency reuse is presented and numerical results in terms of CSI estimation errors are reported. The technique relies on UTs timing estimations and on the synchronization chain for DVB-S2X in [1]. By exploiting the orthogonality given by the WH sequences used in DVB-S2X, the CSI estimations obtained using the timing pre-compensation method are much more reliable even for limiting SNIR condition.

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