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Performance Comparison of Terrestrial DVB Detection using LDPC and Turbo Codes

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Abstract - Last-generation and future wireless communication standards, such as DVB-T2 or DVB-NGH, are including multi-antenna transmission and reception in order to increase bandwidth efficiency and receiver robustness. The main goal is to combine diversity and spatial multiplexing in order to fully exploit the multiple-input multiple output (MIMO) channel capacity. Full-rate full-diversity (FRFD) space-time codes (STC) such as the Golden code are studied for that purpose. However, despite their larger achievable capacity, most of them present high complexity for soft detection, which hinders their combination with soft-input decoders in bit-interleaved coded modulation (BICM) schemes. This article presents a low complexity soft detection algorithm for the reception of FRFD space-frequency block codes in BICM orthogonal frequency division multiplexing (OFDM) systems and gives the performance comparision using Ldpc and Turbo codes. The proposed detector maintains a reduced and fixed complexity, avoiding the variable nature of the list sphere decoder (LSD) due to its dependence on the noise and channel conditions. Complexity and simulation based performance results are provided which show that the proposed detector performs close to the optimal log-maximum a posteriori (MAP) detection in a variety of DVB-T2 broadcasting scenarios.

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Strictly as per the compliance and regulations of:

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Eugin E [α] & Jeno Paul P [σ]

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I. Introduction

SPACE-TIME coding is one of the main methods in order to exploit the capacity of multiple-input multiple-output (MIMO) channels [1]. Since STC techniques use both time and spatial domains for coding data symbols, diversity and spatial multiplexing can be combined achieving robustness at the receiver with a higher data rate transmission. As a result, STC techniques have been incorporated in many of the last-generation wireless communications systems, including the new generation of terrestrial and mobile digital video broadcasting (DVB) standards. If STC is joined to multi-carrier modulation, such as orthogonal frequency-division multilexing (OFDM), space frequency block coding (SFBC) can be performed. This way, codeword's are fed into adjacent of the two consecutive OFDM symbols, translated to the time domain and transmitted through several transmit antennas. This transmission scheme is usually combined with bit-interleaved coded modulation (BICM) giving good diversity results in a wireless communication link [2].

In order to achieve the full MIMO diversity-multiplexing frontier [3], the proposals for the future generations of terrestrial, portable and mobile digital video broadcasting standards, such as DVB-NGH, focus on the combination of both diversity and spatial multiplexing [4], [5] through full-rate full-diversity (FRFD) codes [6]. The main drawback of full-rate codes arises from their very high decoding complexity, which grows exponentially with the number of transmitted symbols per codeword. In order to reduce the complexity of the detection process, hard detection techniques such as sphere decoding (SD) or low complexity STC designs [7], [8] can be used. Nevertheless, when iterative decoders, such as turbo or LDPC codes, are included in the reception chain, soft information on the conditional probabilities for all possible transmitted symbols is required in the form of log-likelihood ratios (LLR). Moreover, the computation of the LLRs for the whole set of transmitted symbols is unfeasible, specially for large constellation sizes. Hence, the soft MIMO detector has to select a group of candidates to be fed to the decoder in order to compute the required LLRs. Several algorithms that serve this purpose can be found in the literature, such as list sphere detection (LSD) [9], near-optimal soft SD [10], list fixed-complexity sphere detection (LFSD) [11], QR decomposition associated with the M-algorithm (QRM-MLD) [12] or bounded soft sphere detection (BSSD) [13].

The contribution of this paper is comparing the performance of soft detection algorithm for FRFD SFBC and its assessment in an Ldpc/Turbo-based BICM scenario. The generation of the candidate list is carried out by means of a fixed tree search, whose complexity does not depend on the channel conditions or the noise level. Since the complexity of the proposed soft detector is closely linked to the architecture of the tree search, we analyze different tree configurations in order to find the best balance between complexity and performance. Simulation results are provided for the Golden [14] and the FRFD Sezginer-Sari (SS) codes [8] in a DVB-T2 framework [15]. Although our research has been carried out on a terrestrial TV system, the results can be generalized for any MIMO bit interleaved coded modulation scheme. The remainder of the paper is organized as follows: Section II details the system model, focusing on soft detection and LLR calculation.

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The design of the new algorithm is presented for two FRFD SFBCs in Section III, while Section IV shows the simulation-based performance comparison of receiver in DVB-T2 scenarios. Finally, the main concluding remarks are drawn in Section V.

II. System Model

The basic structure of the Ldpc/Turbo-coded BICM-OFDM system is depicted in Fig. 1. As can be seen, the bit stream is coded, interleaved and mapped onto a complex constellation.

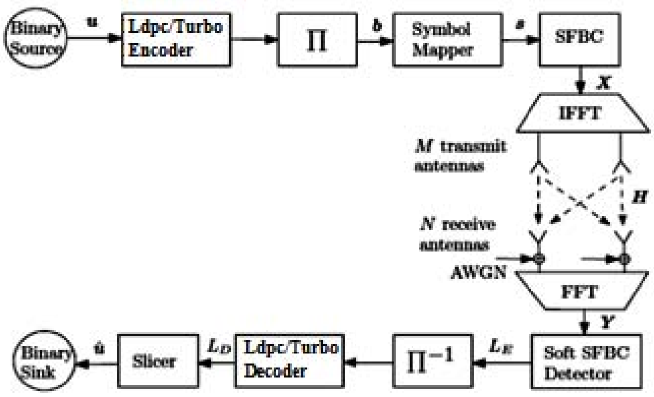

Figure 1 : Simplified diagram of a Ldpc/Turbo-based SFBC MIMO transmission and reception scheme based on DVB-T2

Next, a vector of Q symbols s is coded into space and frequency forming the codeword **X**, which is transformed into the time domain by an inverse fast fourier transform (IFFT) block and transmitted after the addition of fast Fourier transform (FFT) is carried out and the resulting signal **Y** of dimensions $N \times T$ can be represented mathematically as

$$\boldsymbol{Y} = \boldsymbol{HX} + \boldsymbol{Z}, \tag{1}$$

where H denotes the $N \times M$ complex channel matrix, **X** is any $M \times T$ codeword matrix composed by a linear combination of Q data symbols and **Z** represents the $N \times T$ zero-mean additive white Gaussian noise (AWGN) matrix whose complex coefficients fulfill $\mathcal{CN}$ (0, $2\sigma^2$) being σ^2 the noise variance per real component. The design of the codeword **X** follows the criteria defined in [16], [17] and will provide full rate when $Q = MT$, being T the frequency depth of the codeword. By taking the elements column-wise from matrices **X**, **Y** and **Z**, equation (1) can be vectorized as

$$\boldsymbol{y} = \breve{H}\boldsymbol{Gs} + \boldsymbol{z}, \tag{2}$$

where y, s and z are column vectors. The matrix $\breve{H}$ is the equivalent $NT \times MT$ MIMO channel written as

$$\breve{H} = \begin{bmatrix} H^1 & 0 & \dots & 0 \\ 0 & H^2 & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & H^T \end{bmatrix} \tag{3}$$

where we have a block diagonal of channel realizations H^c at the carriers $c = 1, \ldots, T$. The complex coefficient $h^c{}_{ij}$ corresponds to the channel from transmit antenna j to receive antenna i at the c-th carrier. The off-diagonal entries are zero matrices with dimensions $N \times M$. The matrix **G** is the generator matrix for the SFBC such that x = **G**s, where s corresponds to the symbol column vector $[s1, \ldots, sQ]^T$.

a) Soft Detection of SFBCs

The soft information required by the Ldpc/Turbo decoder is obtained though maximum a posteriori (MAP) detection, which consists of evaluating the LLR of the *a posteriori* probabilities of a bit bk taking its two possible values,

i.e.

$$L_D(b_k) = \ln \frac{\Pr(b_k = +|y)}{\Pr(b_k = -|y)} \text{ with } k = \{0, .., MT \log_2 P - 1\}.$$

Assuming statistically independent information bits and using the Bayes' rule, the a posteriori information $L_D(b_k/\boldsymbol{y})$ can be expressed as $L_D(b_k/\boldsymbol{y}) = L_A(b_k)+L_E(b_k/\boldsymbol{y})$, where $L_A(b_k)$ and $L_E(b_k/\boldsymbol{y})$ denote the a priori and extrinsic information, respectively. Considering our non-iterative model depicted in Fig. 1 and the vectorized model (2), the extrinsic information can be written using the Max-log approximation as

$$\begin{aligned} L_E(b_k/y) \approx\ & \frac{1}{2}\max_{b\in B_{k,+1}}\left\{-\frac{1}{\sigma^2}\|y-\breve{H}G_s\|^2\right\} \\ & -\frac{1}{2}\max_{b\in B_{k,-1}}\left\{-\frac{1}{\sigma^2}\|y-\breve{H}G_s\|^2\right\}, \end{aligned} \tag{4}$$

Where $\mathbb{B}_{k,\pm1}$ represents the sets of $2^{MT \log 2\, P-1}$ bit vectors $\boldsymbol{b}$ having $b_k = \pm1$ and the symbol column vector s = $map(\boldsymbol{b})$ is the mapping of the vector $\boldsymbol{b}$ into the symbols o column vector $\boldsymbol{s}$. The main difficulty in the calculation of (4) arises from the computation of the metrics $\|\boldsymbol{y}- \breve{\boldsymbol{H}}\boldsymbol{Gs}\|^2$ since a calculation of P^Q metrics is necessary for a FRFD SFBC, being P the modulation order. This becomes unfeasible for high modulation orders unless the calculation of (4) can be reduced. As a result, a good approximation based on a candidate list $\mathcal{L}$ is proposed in [9] in order to reduce the calculation of L_E in (4). The list includes $1 \le N_{cand} < P^Q$ vectors s with the smallest metrics and the number of candidates N_{cand} must be defined sufficiently large in such a way that it contains the maximizer of (4) with high probability. Hence, (4) can be approximated as

$$\begin{aligned} L_E(b_k/y) \approx\ & \frac{1}{2}\max_{b\in\mathcal{L}\cap B_{k,+1}}\left\{-\frac{1}{\sigma^2}\|y-\breve{H}G_s\|^2\right\} \\ & -\frac{1}{2}\max_{b\in\mathcal{L}\cap B_{k,-1}}\left\{-\frac{1}{\sigma^2}\|y-\breve{H}G_s\|^2\right\}, \end{aligned} \tag{5}$$

III. Fixed-Complexity Detection

The design of efficient detection algorithms is one of the greatest challenges when implementing full-rate SFBC. Given the high complexity of performing an exhaustive search, special focus has been drawn into developing lower complexity detection algorithms that yield a close-to-ML performance. The LSD is one of the most remarkable approach but its complexity order is bounded by *(P^Q)* in the same way as the SD [9]. Even though the list of candidates corresponds to the set $\mathcal{L}$ of the smallest metrics, the complexity of performing such a selection may be considerably high for low signal-tonoise ratio (SNR) scenarios. Furthermore, an unsuitable choice of the initial radius may lead to a shortage in candidate points, which forces the algorithm to restart with a looser sphere onstraint. In order to limit the complexity and to facilitate the computation of soft detected symbols, a fixed-complexity tree-searchstyle algorithm was proposed in [18] for spatial multiplexing schemes, coined list fixed-complexity sphere decoder (LFSD).

The main feature of the LFSD is that, instead of constraining the search to those nodes whose accumulated Euclidean distances are within a certain radius from the received signal, the search is performed in an unconstrained fashion. The tree search is defined instead by a tree configuration vector $n = [n_1, . . . , n_{MT}]$, which determines the number of child nodes (n_i) to be considered at each level. Therefore, the tree is traversed level by level regardless of the sphere constraints. Once the bottom of the tree is reached, the detector retrieves a list of N_{cand} candidate symbol vectors.

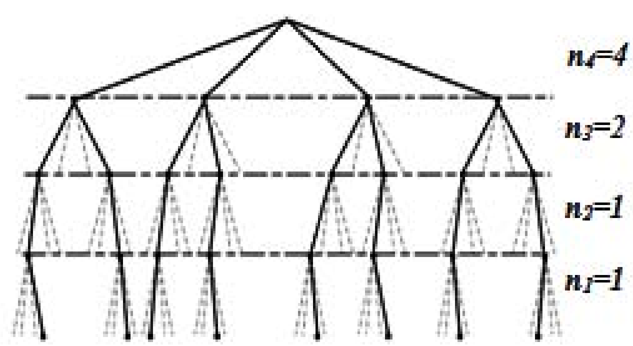

Figure 2 : Fixed-complexity tree search of a QPSK-modulated signal using a tree configuration vector of n = [1, 1, 2, 4]

It is worth noting that the set *G* composed the vectors of the set $\mathcal{L}$ with the smallest metrics given by the LSD, but provides sufficiently small metrics and diversity of bit values to obtain accurate soft information. A representation of an LFSD tree search is depicted in Figure 2 for a QPSK modulation and a tree configuration vector of $n = [1, 1, 2, 4]$.

a) Soft-output LFSD algorithm

For the sake of simplicity, the equations for the aforementioned FRFD codes will be rearranged so that ML metrics can be given by $\|\bar{y} - \boldsymbol{H}_{eq}\bar{s}\|^2$, where $\bar{y}$ and $\bar{s}$ are the received and transmitted signals respectively, reorganized according to the Corresponding SFBC code, and H_{eq} is the effective equivalent channel, which will be defined later for the SS and Golden codes.

A level-by-level computation of the metrics requires the conversion to the following equivalent system

$$\|U(\bar{s} - \hat{\bar{s}})\|^2, \tag{6}$$

where *U* is obtained through the Cholesky decom-position of $H^H_{eq}H_{eq}$ and $\hat{\bar{s}} = H^{\dagger}_{eq}\,\bar{y}$. Given the triangular structure of *U*, it is now possible to compute the accumulated Euclidean distances (AED) up to level *i* recursively by traversing the tree backwards from level $i = MT$ down to $i = 1$. The Euclidean distances that must be minimized in the cost function in (6) can be equivalently represented in a tree search fashion as

$$D_i = u_{ii}^2|\bar{s}_i - Z_i|^2 + \sum_{j=i+1}^{MT} u_{jj}^2|\bar{s}_j - Z_j|^2 = d_i + D_{i+1}, \tag{7}$$

and

$$Z_i = \hat{\bar{s}}_i - \sum_{j=i+1}^{MT} \frac{u_{ij}}{u_{ii}}\left(\bar{s}_j - \hat{\bar{s}}_j\right). \tag{8}$$

Therefore, the n_i symbols to be evaluated at each level *i* are chosen in accordance with the Schnorr-Euchner enumeration [19], being their corresponding partial Euclidean distances d_i computed and accumulated to the previous level's AED, that is, D_{i+1}. Once the bottom of the tree has been reached, a sorting operation is performed on the $n_T = \prod_{i=1}^{MT} n_i$ Euclidean distances in order to select the N_{cand} symbol vectors with the smallest metrics. This latter sorting procedure can be avoided if the tree configuration vector n is chosen so as to yield $n_T = N_{cand}$. In such a case, the complexity of the algorithm is reduced at the expense of a degradation in the quality of soft information as the selected metrics are higher in value.

b) Ordering Algorithm

The performance of the LFSD soft-detector in uncoded scenarios is strongly dependent on the ordering algorithm of the channel matrix and the choice of the tree configuration vector [18]. The ordering algorithm proposed in [18] to enhance the performance of the LFSD and FSD detectors was based on the fact that it was possible to mitigate the error propagation derived from ruling out several tree branches by ordering the several columns of the channel matrix according to their *quality*.

More precisely, the FSD ordering scheme dictates that the subchannel with the worst norm needs to be processed first, since all the constelation symbols are evaluated at the first level (see Fig. 2), and therefore, there is no error propagation to the remainder levels. However, in the specific case of spacefrequency- coded systems, the effect of the ordering algorithm on the overall performance relies on the type of code utilized. In order to verify this assumption, two 2 × 2 FRFD SFBC codes have been assessed.

i. *Golden Code [14]*

For the Golden code, the data symbol vector s is transformed into the transmitted codeword as follows:

$$\mathbf{X_g} = \frac{1}{\sqrt{5}}\begin{bmatrix} \alpha(s_1+\theta s_3) & \alpha(s_2+\theta s_4) \\ i\bar{\alpha}(s_2+\bar{\theta} s_4) & \bar{\alpha}(s_1+\bar{\theta} s_3) \end{bmatrix}, \quad (9)$$

With $\theta = \frac{1+\sqrt{5}}{2}$ (the golden number), $\bar{\theta} = \frac{1-\sqrt{5}}{2}$,

$$\alpha = 1+i-i\theta \text{ and } \bar{\alpha} = 1+i-i\bar{\theta}.$$

The vectorization of (1) with Xg implies the following equivalent channel

$$\boldsymbol{H_{eq}} = \begin{bmatrix} h_{11}^1(1+i\bar{\theta}) & h_{12}^1(-\theta+i) & h_{11}^1(\theta+i) & h_{12}^1(-1+i\bar{\theta}) \\ h_{21}^1(1+i\bar{\theta}) & h_{22}^1(-\theta+i) & h_{21}^1(\theta+i) & h_{22}^1(-1+i\bar{\theta}) \\ h_{12}^2(1+i\theta) & h_{11}^2(1+i\bar{\theta}) & h_{12}^2(\bar{\theta}+i) & h_{11}^2(\theta+i) \\ h_{22}^2(1+i\theta) & h_{21}^2(1+i\bar{\theta}) & h_{22}^2(\bar{\theta}+i) & h_{21}^2(\theta+i) \end{bmatrix} \quad (10)$$

The fact that Golden code does not equally disperse the symbol energy in all spatial and temporal directions, generates an unbalanced structure of the transmitted symbols since the norms of the code weights in (10) are not equal, i.e. $1+\vartheta^2 \neq 1+\bar{\theta}^2$. Thus, given this difference in the absolute value of the weights, one of the symbols in each transmitted pair (s_1, s_3) and (s_2, s_4) always has a higher power than the other. Hence, the norms of the equivalent subchannels are $\|h_1\|^2 \neq \|h_3\|^2$ and $\|h_2\|^2 \neq \|h_4\|^2$ in any case, being h_j the j-th column vector of H_{eq}. This unbalanced structure allows for the implementation of a new ordering procedure in order to improve the overall system's performance.

On the other hand, if the channel is assumed quasi-static over adjacent carriers, are $\|h_1\|^2 \approx \|h_3\|^2$ and $\|h_2\|^2 \approx \|h_4\|^2$. An important feature when considering the optimum ordering approach is the tree configuration vector that will shape the search tree. As opposed to the LFSD detector presented in [18] for spatial multiplexing MIMO transmission, the tree configuration vector for the detection of the Golden Code has been set to $n = [k, k, P, P]$, where $k < P$. With such a tree structure, an exact ML search is performed in the first two levels of the tree, and therefore, there is no error propagation down to the next levels.

Consequently, by ordering the equivalent channel matrix in such a way that the *worst* sub-channel is processed in the first two levels of the tree, the probability of finding vectors with smaller metrics is increased. Moreover, it has to be taken into account that the symbols belonging to the same pair need to be detected together in the non-ML part of the tree search for a better performance of the algorithm, since there exists a correlation between their corresponding sub-channels due to the code structure.

The equivalent ordered channel matrix $\boldsymbol{H_{eq}^{ord}}$, which will be used in the detection of the Golden Code, can then be described as

$$\boldsymbol{H_{eq}^{ord}} = [h_{\tilde{b}}\ h_b\ h_{\tilde{w}}\ h_w], \quad (11)$$

Where

$$w = argmin_{j\in S}\ \|h_j\|^2, \quad (12)$$

and

$$b = argmax_{j\in S, j\neq w, \tilde{w}}\ \|h_j\|^2. \quad (13)$$

The two symbols that compose a symbol pair are represented as $(a, \bar{a})$ and the set of symbol indices is $S = \{1,\ldots,MT\}$. Given the chosen tree configuration vector, one can notice that the order of the first two selected symbols can be switched without having any impact on the final performance of the system.

ii. *The SS Code*

This 2×2 SFBC scheme was designed to enable optimum detection with lower complexity than the Golden code. The low complexity detection property of multistrata codes such as the SS code, was analyzed in [7], [8],where it was proven that optimum output was obtained with two symbol-by-symbol detection stages of complexity P and a ML detection of complexity $\mathcal{O}(P^2)$. The main difference between the SS [8] and the silver codes [7] is the larger coding gain of the latter. The SS code is the combination of two Alamouti schemes whose codeword can be written as

$$\boldsymbol{X_{ss}} = \begin{bmatrix} as_1+bs_3 & as_2+bs_4 \\ -cs^*_2-ds^*_4 & cs^*_1-ds^*_3 \end{bmatrix}, \quad (14)$$

With $a = c = 1/\sqrt{2}$, $b = \frac{1-\sqrt{7}+i(1+\sqrt{7})}{4\sqrt{2}}$ and $d = -ib$.

The vectorization of (1) for $\boldsymbol{X}_{ss}$ implies a rearrangement of the received and transmitted signals such that $\bar{\mathbf{y}} = [y_{11} y_{21} y^*_{12} y^*_{22}]^T$ and $\bar{\mathbf{s}} = [s_1, s_2, s^*_3, s^*_4]^T$ respectively.

The equivalent channel can be expressed as:

$$\boldsymbol{H}_{eq} = \begin{bmatrix} ah^1_{11} & -ch^1_{12} & bh^1_{11} & -dh^1_{12} \\ ah^1_{21} & -ch^1_{22} & bh^1_{21} & -dh^1_{22} \\ (ch^2_{12})^* & (ah^2_{11})^* & (dh^2_{12})^* & (bh^2_{11})^* \\ (ch^2_{12})^* & (ah^2_{21})^* & (dh^2_{22})^* & (bh^2_{21})^* \end{bmatrix}, \quad (15)$$

When considering the equivalent channel in (15), it is worth noting that the equivalent subchannels for the symbol pairs ($s1$, $s3$) and ($s2$, $s4$) have very similar norms. This is due to two main factors. On one hand, both symbol pairs undergo almost the same channel conditions as they are assigned to adjacent carriers ($H^1 \approx H^2$) in quasi-static fading channels). On the other hand, the code weights a, b, c and d imposed by the SS code fulfill a power constraint for linear dispersion codes [20], which forces the symbols to be dispersed with equal energy in all spatial and temporal directions, i.e. $|a| = |b| = |c| = |d| = 1/\sqrt{2}$.

The consequence of employing such a code is that the difference of the norms of the equivalent subchannels is negligible and, therefore, performing a matrix ordering stage does not provide any remarkable performance enhancement. As it will be shown in the next section, the proposed ordering approach yields close-to-optimum performance when combined with the suggested tree configuration vector. Moreover, the matrix ordering process only requires the computation of MT vector norms as opposed to other ordering algorithms such as FSD [18] or V-BLAST [21], which need to perform $MT - 1$ matrix inversion operations.

c) Bit LLR generation for the proposed list fixed-complexity detector

The expression for the LLRs in (5) can be rewritten to comply with the equivalent system in (6) as

$$L_E(b_{k|\mathbf{y}}) \approx \frac{1}{2}\min_{b\in\mathfrak{G}\cap B_{k,+1}} \frac{1}{\sigma^2}\|\bar{y} - \boldsymbol{H}_{eq}\bar{s}\|^2 - \frac{1}{2}\min_{b\in\mathfrak{G}\cap B_{k,-1}} \frac{1}{\sigma^2}\|\bar{y} - \boldsymbol{H}_{eq}\bar{s}\|^2, \quad (16)$$

Note that the list of candidates $\mathcal{L}$ of the ML/LSD detector has been substituted by the set G of the LFSD.

IV. Simulation Results

The performance of the overall system has been assessed by means of the bit error rate (BER) after the Ldpc/Turbo decoder. The DVB-T2 parameters used in the simulations are: 64800 bits of length of the Ldpc/Turbo block, $R = 2/3$ of Ldpc/Turbo code rate, 16-QAM modulation, 2048 carriers as FFT size and 1/4 of guard interval. The simulations have been carried out over a Rayleigh channel (Typical Urban of six path, TU6), commonly used as the simulation environment for terrestrial digital television systems [22]. Perfect CSI has been considered at the receiver.

a) Candidate Choice

When working with the ML metrics of LSD, i.e. the list L, the higher the number of candidates is, the more accurate the LE approximation is. Nevertheless, when considering the $\mathfrak{G}$ list of LFSD, the optimum value for N_{cand} will depend on the tree configuration vector n. Thus, the higher the value of n_T, the better the approximation is. In order to choose a suitable number of candidates for the detection algorithm, a battery of tests have been carried out. Fig. 3 depicts the bit error performance after the detection stage for a given SNR of 14.4 dB with different tree search configurations and N_{cand} values. The effect of the ordering preprocessing stage is also depicted in this figure. The analyzed tree search configuration vectors $\boldsymbol{n}$ have been obtained by setting $k = 1, 2, 3$, which is equivalent to calculating $n_T = P^2, 4P^2, 9P^2$ Euclidean distances, respectively. On one hand, one can observe that the list ML approximation (5) converges for $N_{cand} > 30$. This involves that computing a very large number of candidates is not necessary in order to obtain a good L_E approximation of (4) for the proposed non-iterative scheme in Fig.1. However, we should take into account that the exact computation of (4) provides a higher performance enhancement compared to applying the list ML approximation of (5) when iterative configurations are used [23].

On the other hand, a similar behavior for the fixed-complexity detector can be noticed, where the higher the value of k, the better the performance we obtain. Furthermore, it is noticeable that the ordering algorithm provides a performance enhancement such that the $k = 2$ LFSD approximates the BER values for the exhaustive MAP detector. Note that the BER degrades for a higher number of candidates with the tree search configuration $k = 1$.This is due to the fact that if we choose a large N_{cand} value from $n_T = P^2$ Euclidean distances, the probability of achieving the smallest or close to the smallest metrics is reduced. For $k > 1$, this effect is mitigated.

b) Performance comparison over DVB-T2 BICM

This section presents the performance assessment of the proposed list fixed-complexity soft detector over a SFBC DVB-T2 broadcasting scenario. The number of candidates considered for this study is $N_{cand} = 50$. Below graphs shows BER curves versus SNR for different configurations of the proposed algorithm in the detection of Golden and SS codes. For the Golden code, it is noteworthy that the ordering

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algorithm provides a gain of 0.4 and 0.25 dB compared to the non-ordering case for $n_T = P^2$ and $n_T = 4P^2$, respectively. However, as previously stated, the ordering algorithm does not provide any performance gain with the SS code. In this case, the subchannel norms of the symbol pair $(a, \tilde{a})$ are completely equal, i.e., $\|h_b\|^2 = \|h_{\tilde{b}}\|^2$ and $\|h_w\|^2 = \|h_{\tilde{w}}\|^2$, being negligible the enhancement provided by the ordering procedure.

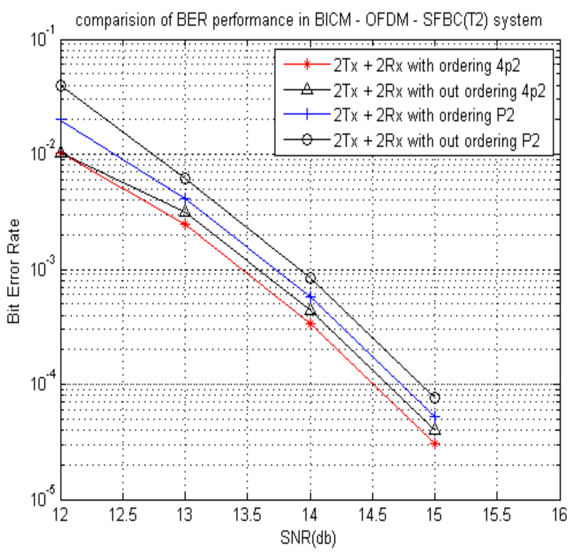

Figure 3 : Performance comparision with and with out ordering algorithm using LDPC codes

The Fig 3. shows that the performance Comparision based on different equlidian distances by using with and with out ordering algorithms.Here the first black colour plot represents that by with out using the ordering algorithm for the equlidian distance (p^2), when BER (bit error rate) is 10^{-4} the signal to noise ratio (SNR) will be 14.9db. And the second blue colour plot represent that by using the ordering algorithm for the equlidian distance (p^2), when BER (bit error rate) is 10^{-4} the signal to noise ratio (SNR) will be 14.7db. Next the third black colour plot on the graph represents that by with out using the ordering algorithm for the equlidian distance ($4p^2$), when BER (bit error rate) is 10^{-4} the signal to noise ratio (SNR) will be 14.6db. And finally the fourth red colour plot represents that by using the ordering algorithm for the equlidian distance ($4p^2$), when BER (bit error rate) is 10^{-4} the signal to noise ratio(SNR) will be 14.5db.

In LDPC coding one bit is generating one parity, so the error rate is 14.5db with ordering algorithm. This gives the better performance when compared with out ordering algorithm for detecting the signals at the different signal to noise ratio condition.

The Fig 4. shows that the performance Comparision based on different equlidian distances by using with and with out ordering algorithms. Here the red colour plot represents that by with out using the ordering algorithm for the equlidian distance (p^2), when BER (bit error rate) is 10^{-4} the signal to noise ratio (SNR) will be 14.45db. And the green colour plot represent that by using the ordering algorithm for the equlidian distance (p^2), when BER (bit error rate) is 10^{-4} the signal to noise ratio (SNR) will be 14.35db.

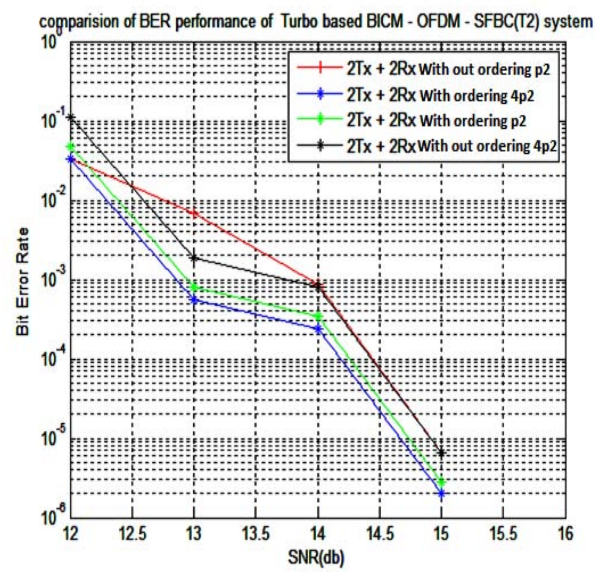

Figure 7 : Performance comparision with and with out ordering algorithm using Turbo codes

Above that the black colour plot on the graph represents that by with out using the ordering algorithm for the equlidian distance *($4p^2$)*, when BER (bit error rate) is 10^{-4} the signal to noise ratio (SNR) will be 14.4db. And finally the finally plot represents that by using the ordering algorithm for the equlidian distance ($4p^2$), when BER (bit error rate) is 10^{-4} the signal to noise ratio (SNR) will be 14.25db.

Thus in turbo codes one bit is generating two parity so the error rate is decreasing upto 14.25db with ordering algorithm. This gives the better performance when compared with out ordering algorithm for detecting the signals at the different signal to noise ratio condition. Also it will be giving the better performance with reduced complexity and achieve full rate for detecting the terrestrial digital TV signals in the different fading channel conditions like additive white gaussian noise and rayliegh channels.

V. Conclusion

Multi-antenna transmission using 2×2 FRFD codes, such as the Golden code or the SS code, increases the capacity allowing a higher data rate transmission with full diversity. This capacity increase involves a joint detection of a candidate list of four data symbol vectors in order to achieve soft information for the decoder. The complexity of this calculation can be reduced by means of different algorithms, being the most extended the LSD detector. However, the main drawback of this technique is its high and variable complexity, which can be upper-bounded by $\mathcal{O}(P^4)$. A list fixed-complexity detector with a novel ordering algorithm is proposed in this paper with the aim of approaching the performance of the LSD at a much lower complexity. Specifically, the complexity order can be reduced from $\mathcal{O}(P^4)$ to P^2 for the Golden code and from $\mathcal{O}(P^3)$ to P^2 for the SS code.

The analysis of the number of candidates shows that the list approximation does not need a high list size in order to converge to the exact soft information value. Provided simulation results show that a close-to-optimal detection can be achieved considering a reduced number of candidates (30 out of 65536 in 16-QAM). BER (bit error rate) simulation results show the close-to-optimal performance of the proposed low-complexity detector for both Golden and SS SFBC codes in a typical Ldpc/Turbo-based DVB-T2 broadcasting scenario. The performance is clearly improved when the proposed channel and candidate ordering algorithm is applied with Golden codes, though its effects are negligible for the SS code. In any case, the proposed detection algorithm can enable the realistic implementation and the inclusion of any FRFD SFBC code in any BICM-OFDM system such as the forth coming digital video broadcasting standards.

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