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REDUCED COMPLEXITY SOFT DETECTION FOR DOUBLY ITERATIVE RECEIVER USING 64 – QAM

BY

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Abstract. The spectral efficiency for a communication system can be increased using a flexible modulation and coding diagram such as bit interleaved coded modulation combined with a high order modulation diagram like Quadrature Amplitude Modulation (QAM). In this paper, we propose two methods to reduce the complexity of the soft detection for a doubly iterative decoder using space-time turbo codes and a large number of transmit and receive antennas for 64–QAM modulation. The reduced complexity calculation methods for the log-likelihood ratio (LLR) for the current demodulated symbol consist in considering a certain number of symbols belonging to the first three outlines around the respective symbol, instead of all 64. In the first method, some additional points are required when no information exists for a bit from the six ones corresponding to a certain symbol. In the second method, the additional points are avoided, but instead a fixed distance is used in the numerator of the denominator of LLR value. The average detection time is reduced by about 48% compared to the original diagram, while the total time required for the receiver diagram decreases with about 7% up to 18%, depending on the SNR values and on the number of outer iterations.

Keywords: space-time turbo codes; doubly iterative decoder; soft estimates.

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1. Introduction

Space – time turbo codes with iterative decoding (Foschini & Gans, 1998) have become an area of great interest in the last two decades. It has been demonstrated that the use of an iterative demodulation – decoding approach can lead to a performance of 2 to 3 dB from the channel outage probability for quasistatic fading channels and perfect channel state information available at the receiver. In (Biglieri *et al.*, 2005), Biglieri presented a block diagram for a doubly iterative receiver based on the minimum mean square error (MMSE) criterion (Biglieri *et al*., 2003). A new diagram has been presented with a significantly reduced complexity for a large number of transmit and receive antennas, compared to the previous diagram proposed by Stefanov and Duman (Stefanov & Duman, 2001). Since the large number of antennas increases the receiver complexity, a spatial interference canceling diagram is used, ensuring a good compromise between complexity and performance.

The high order modulation diagrams have the advantage of larger data rates and better spectral efficiencies for radio communications systems. The disadvantage is that the performance of the iterative receivers depends critically on the size of the signal constellation and a high order modulation diagram is less robust to noise and interference, leading to performance degradation.

In this paper addresses the detection block from the doubly iterative receiver and we propose two reduced complexity methods for the calculation of the log-likelihood ratio (LLR) values for coded bits.

The paper is structured as follows: in Section **2** we recall the system model consisting of the transmitter and the receiver block diagram. In Section **3** we present the proposed reduced complexity detection methods and in Section **4** the simulation results are presented. Section **5** concludes the paper.

2. System Model

We consider the same mobile communication system as in Rotopănescu *et al*., (2012), or Rotopănescu *et al*., (2016), with N_T transmit antennas and N_R receive antennas. The information bits are turbo-coded with coding rate R_c and block size of *NTN* modulated symbols, where *N* is the number of successive transmissions from the transmit antennas, corresponding to a codeword.

The signal at the modulator output is denotes by the symbol $x_{i,t}$ transmitted by antenna *i*, $1 \leq i \leq N_T$, at each time instant *t*, $1 \leq i \leq N$. The modulation function is $\boldsymbol{x}_t = \boldsymbol{f}\left(\boldsymbol{c}_t\right) = \left(f_1\left(\boldsymbol{c}_{1,t}\right),\ ...,\ f_t\left(\boldsymbol{c}_{N_T,t}\right)\right)^T$ and it transforms the MN_T components of the column vector of the coded symbols from the turbo decoder output, at time *t*, $\boldsymbol{c}_t = (c_{1,t},\ ...,\ c_{M \cdot N_T,t})^T$ into the column vector $x_t = (x_{1,t},\ ...,\ x_{N_T,t})^T$, where *M* is the number of bits transmitted per modulated symbol. $f_i(\cdot)$, $1 \le i \le N_T$, is the mapping function for the modulator corresponding to the i-th transmit antenna, and $\boldsymbol{c}_{i,t} = (c_{(i-1)\cdot M+1,t}, \ ..., \ c_{i\cdot M,t})^T$, $1 \le i \le N_T$, $1 \le t \le N$, is the vector with the coded bits that will be mapped in the modulated symbol and this symbol will be transmitted by antenna i at time t. In this paper we consider that all transmit antennas have the same modulation function, namely 64-QAM modulation.

The spectral efficiency refers to the information rate that can be transmitted over a given bandwidth in a specific communication system. The spectral efficiency is denoted by η and is equal to R_cMN_T where R_c is the coding rate of the turbo code. We denote by F the number of independent fading states of equal lengths, given by a single space time codeword matrix of length N_TN (it is assumed that F divides N).

One codeword of the turbo-code contains $F = N_TN/L$ distinct blocks with constant fading, where L is the number of symbols for which the fading remains the same and can take any value that may divide N_TN. In our simulations F is equal to one. The path gains $\alpha_{i,j}$ are modeled by complex independent Gaussian random variables with zero mean and variance 0.5, for each dimension.

The path gains are constant for L symbols corresponding to ηL information bits and they are independent for each L-symbols block. $\alpha_{i,j}$ is the path gain from the transmit antenna i to the receive antenna j, $1 \le j \le N_R$. $z_{i,j}$ is the noise sample for receive antenna j at time t and it is modeled as a complex Gaussian random variable with zero mean and independent real and imaginary parts with variance $N_0/2$. N_0 is the noise power spectral density. $y_{t,j}$is the signal received by antenna j at time t and it consists in the transmitted signals corrupted by Rayleigh fading and Gaussian noise, as follows:

$$y_{t,j} = \sum_{i=1}^{N_T} \alpha_{i,j} x_{t,i} + z_{t,j} . \tag{1}$$

Using a high order QAM constellation more bits per symbol can be transmitted. If the average energy of the constellation remains the same, the symbols must be closer to each other and thus they are more susceptible to be corrupted by noise.

This brings a higher bit error rate and thus higher-order QAM can deliver more data less reliably than lower-order QAM, for constant mean constellation energy. Using higher-order QAM without increasing the bit error rate requires a higher signal-to-noise ratio (SNR) either by increasing signal energy, or by reducing the noise, or both.

The QAM constellation symbols are normally arranged in a square grid with equal vertical and horizontal spacing, so that the most common used QAM constellation have a number of symbols equal to a power of 4, such as 4 – QAM, 16 – QAM, 64 – QAM, 256 – QAM, and so on. In general, the number of symbols of the QAM modulation is $4^m = 2^{2m}$, where the number of the bits in each constellation symbol is $M = 2m$ (m is an integer). The symbols are represented in a complex plane having the in-phase component on the real axis and the quadrature component on the imaginary axis.

The transmitter block diagram is presented in Fig. 1. It performs a coded modulation with bit interleaving and antenna diversity, as described in (Caire *et al*., 1998).

The receiver block diagram in (Biglieri *et al*., 2005; Trifina *et al*., 2011) uses MMSE iterative algorithm and a linear MMSE interface (Biglieri *et al*., 2003) and it is shown in Fig. 2.

The turbo decoding algorithm is the Max-Log-APP (Benedetto *et al*., 1997; Vogt & Finger, 2000). We note that turbo decoding algorithm is an iterative algorithm and the receiver diagram is also an iterative one. Therefore, we mean by inner iteration the iteration in the turbo decoder and by outer iteration the iteration in the global receiver diagram shown in Fig. 2.

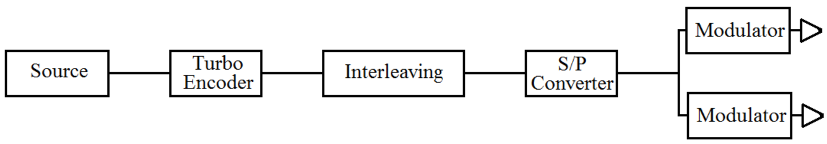

Fig. 1 – Transmitter block diagram.

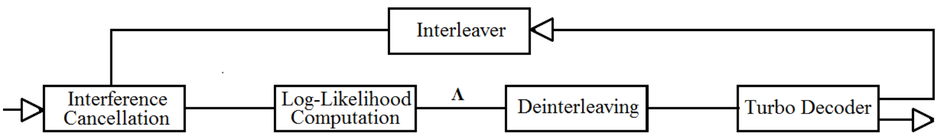

Fig. 2 – Receiver block diagram.

3. Reduced Complexity Detection

The linear MMSE interface used by the receiver, consists in a linear filter modeled by a matrix that minimizes the mean square error as described in (Biglieri *et al*., 2003). The filtered signal is transmitted to the interference canceling block.

The output of this block is generated according to the algorithm described in (Biglieri *et al*., 2005). The suboptimal simplified log-likelihood ratios of the coded bits $c_{q,t}$, $1 \le q \le MN_T$, $1 \le t \le N$, are given by :

$$\widehat{\Lambda}\left(c_{q,t}\right)=\ln\frac{\sum\limits_{x=f_i\left(\boldsymbol{c}_{i,t}\right):c_{q,t}=1}\exp\left\{-\frac{\left|\tilde{y}_{i,t}^{(k)}-x\right|^2}{\left(\boldsymbol{K}^{(k)}\right)_{ii}}\right\}}{\sum\limits_{x=f_i\left(\boldsymbol{c}_{i,t}\right):c_{q,t}=0}\exp\left\{-\frac{\left|\tilde{y}_{i,t}^{(k)}-x\right|^2}{\left(\boldsymbol{K}^{(k)}\right)_{ii}}\right\}} \tag{2}$$

where: $i=1+\left\lfloor\frac{q-1}{M}\right\rfloor$, $\tilde{y}_{i,t}^{(k)}$, and matrix $\boldsymbol{K}^{(k)}$ are described in (Biglieri *et al*., 2005) and k is the number of outer iterations. Fig. 3 presents the signal constellation for 64-QAM modulation using the Gray coded bit-mapping.

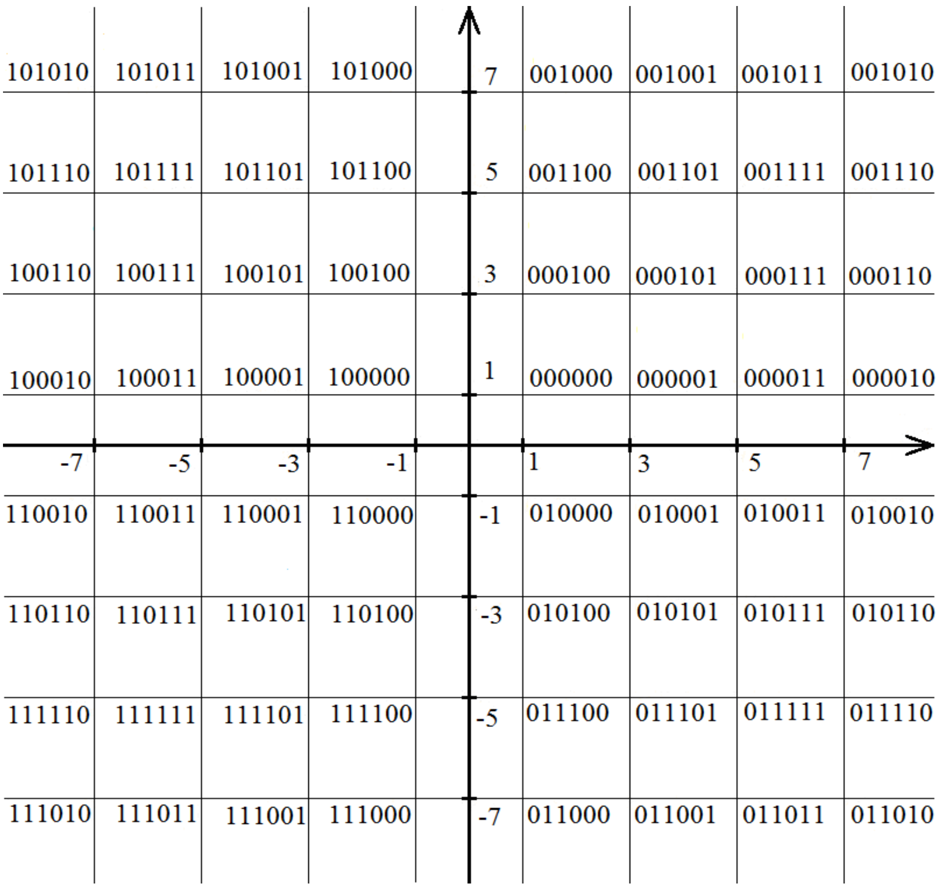

Fig. 3 – Gray coded bit mapping for 64-QAM signal constellation.

From formula (8) in (Biglieri *et al*., 2005) we can write:

$$\tilde{y}_{i,t}^{(k)} = x_{i,t} + q_{i,t}^{(k)},\ 1 \le i \le N_T,\ 1 \le t \le N\ , \tag{3}$$

where $x_{i,t}$ is the symbol transmitted by antenna *i* at time *t* and $q_{i,t}^{(k)}$ is the noise and the residual spatial interference at outer iteration *k*. Variable $q_{i,t}^{(k)}$ is assumed to be modeled as a complex Gaussian random variable and thus can be noticed that when a symbol is transmitted, the symbols around this symbol are more likely to be received.

As described in (Rotopanescu *et al*., 2016), the average energy for this constellation is

$$E_{64-\mathrm{QAM}} = 42. \tag{4}$$

The global turbo coding rate is 1/2 and the spectral efficiency for 64-QAM modulation, in the case of 16 transmit antennas is 48 bits/s/Hz.

In formula (2) above, the sums from the numerator and denominator are performed for all the 64 QAM symbols, each of them consisting of 6 bits. In order to reduce the detection complexity and to reduce the detection time of the received symbols, in this paper we propose a reduced complexity calculation method for the log-likelihood ratios of the coded bits. Specifically, in (2), only a reduced number of QAM symbols is used, instead of all 64.

Further, by *outline* we define all the symbols belonging to a square that has in its center a certain symbol, or a part of a square, if that certain symbol is close to the edge of the constellation. In order to simplify the detection method we first demodulate the current received symbol $\tilde{y}_{i,t}^{(k)}$, and then in the sums from (2) we consider only the symbols belonging to the first 3 outlines around the demodulated symbol. The choose of the first 3 outlines was determined by simulation as being the minimum number of outlines for which there is no performance degradation in decoding. We note that there are demodulated symbols for which no symbol among its outlines contains the bit value 0 or 1 at a certain position from the six possible ones, so that there is no information for the bit on that position. Thus, there is no value corresponding to the sums from the numerator or denominator in (2). In order to remove this possibility, two methods are proposed.

In the Method 1, a number of additional symbols is required in the cases mentioned above.

For example, for symbol 5 (000101), marked with a circle in Fig. 4 *a*, all the symbols of the 3 outlines are 0, 4, 12, 1, 13, 3, 7, 15 marked with a square, 48, 32, 36, 44, 40, 16, 8, 17, 9, 19, 11, 18, 2, 6, 14, 10 marked with a triangle, and 53, 49, 33, 37, 45, 41, 52, 20, 21, 23 marked with star, as seen in Fig. 4 *a*. For symbol 10 (001010), marked with a circle in Figure 4.b, the symbols from the three outlines are 15, 11, 14 marked with a square, 5, 13, 9, 7, 6 marked with a triangle, 0, 4, 12, 8, 1, 3, 2 marked with a star, as presented in Fig. 4 *b*. Because all the 16 symbols from the three outlines for symbol 10, including 10, have the first two bits equal to 0, some additional points are required. The additional point 40, marked with hexagon in Fig. 4 *b*, has the first bit equal to 1 and the additional point 18, marked with pentagon in Fig. 4 *b*, has the second bit equal to 1.

As another example, for the constellation symbol 42, marked with a circle in Figure 4.c, the total number of selected symbols from the first three outlines is 16 (including symbol 42). These symbols are 46, 47, 43 marked with a square, 38, 39, 37, 45, 41 marked with a triangle, 34, 35, 33, 32, 36, 44, 40 marked with a star. The additional symbol 8, marked with a hexagon, is required because this symbol has the first bit equal to 0, while all the previous 16 symbols have the first bit equal to 1. The additional symbol 50, marked with pentagon, is required because this symbol has the second bit equal to 1, while all the previous 16 symbols have the second symbol equal to 0. The total number of symbols for which the bit values 0 and 1 for the six bits appear in all the 18 symbols for symbol 42 are given in Table 1.

We count the total number of considered symbols in Method 1 for which the bit with number l, $l = 1, 2, ..., 6$, is equal to 0 or to 1, in the case of all 64 symbols of the 64-QAM constellation. The sums of these numbers are given in columns 2 and 6 in Table 2. In the method from (Biglieri *et al*., 2005) all the 64 symbols are considered, resulting in 32 symbols for each of the six bits 0 or 1. Thus for all 64 symbols the total number of considered symbols for each of the six bits is $32 \cdot 64 = 2048$. In the columns 4 and 8 in Table 2 we give the percentages of the number of symbols considered in the proposed Method 1 compared to that in (Biglieri *et al*., 2005). Globally, for all the six bits in each of the 64 symbols the results are given in the last row in Table 2, in the corresponding columns.

Considering all symbols to be used equally likely, it results that bits 0 and 1 are also equally likely for each of the six positions. Thus, the global average complexity for computation of LLR in Method 1 results to be equal to 48.05% from that of the method in (Biglieri *et al*., 2005).

In Method 2, as an alternative for the additional points, instead of the value $\left|\tilde{y}_{i,t}^{(k)} - x\right|^2$ in the denominator or the numerator of LLR expression from (2), we use the squared distance from the point corresponding to the demodulated symbol to the respective additional point. Because in all cases the additional points are in the fourth outline of the demodulated symbol, this distance is equal to the distance between two successive points on one axis in the M-QAM constellation, multiplied by four. If the distance between two successive points on an axis in the M-QAM constellation is weighted so that for N_T transmit antennas the average energy is equal to one, then value $\left|\tilde{y}_{i,t}^{(k)} - x\right|^2$ is approximated by the quantity $\left(4\cdot\frac{2}{\sqrt{E_{M\text{-QAM}}\cdot N_T}}\right)^2=\frac{64}{E_{M\text{-QAM}}\cdot N_T}$. E_{M-QAM} is the average energy of the M-QAM constellation when the distance between two successive points on an axis is equal to two. Thus, for 64-QAM and 16 transmit antennas, as considered in this paper, value $\left|\tilde{y}_{i,t}^{(k)}-x\right|^2$ is approximated by the quantity $4/42\cong 0.09524$.

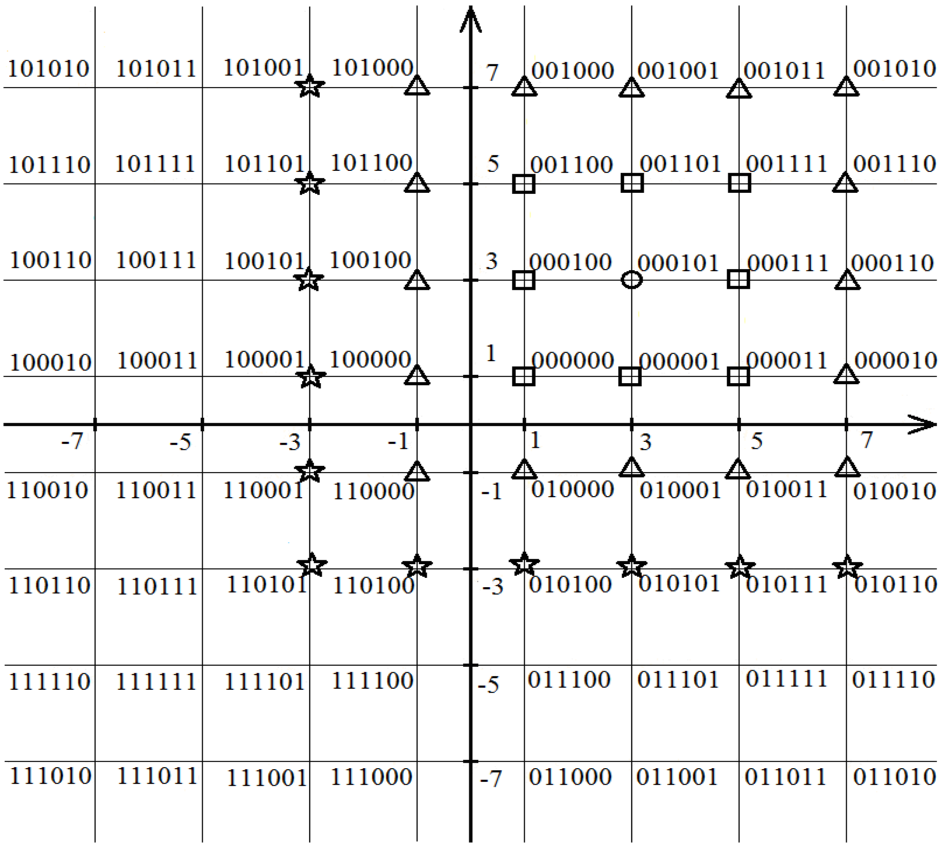

Fig. 4 *a* – Outlines for symbol 5 (000101).

We note that in Method 2, we have to test if for a demodulated symbol we have no information for the bit 0 or 1 in one of the six positions. For example, if we initialize the value of denominator or numerator from (2) with zero, when we compute LLR for a bit, we have to test if this value is zero or not. This test is not required in Method 1.

The sums of the numbers of symbols from outlines without additional points, for which the bit with number l, l = 1, 2, ..., 6, is equal to 0 or to 1 are given in columns 3 and 7 in Table 2. In the columns 5 and 9 in Table 2 we give the percentages of the number of symbols considered in Method 1 compared to that in (Biglieri *et al.*, 2005). Globally, for all the six bits in each of the 64 symbols the results are given in the last row in Table 2, in the corresponding columns.

The global average complexity for computation of LLR in the proposed Method 2 results to be equal to 47.27% from that of the method in (Biglieri *et al.*, 2005). However, we note that for this method an additional test is required and, thus, the complexities for the two methods are comparable.

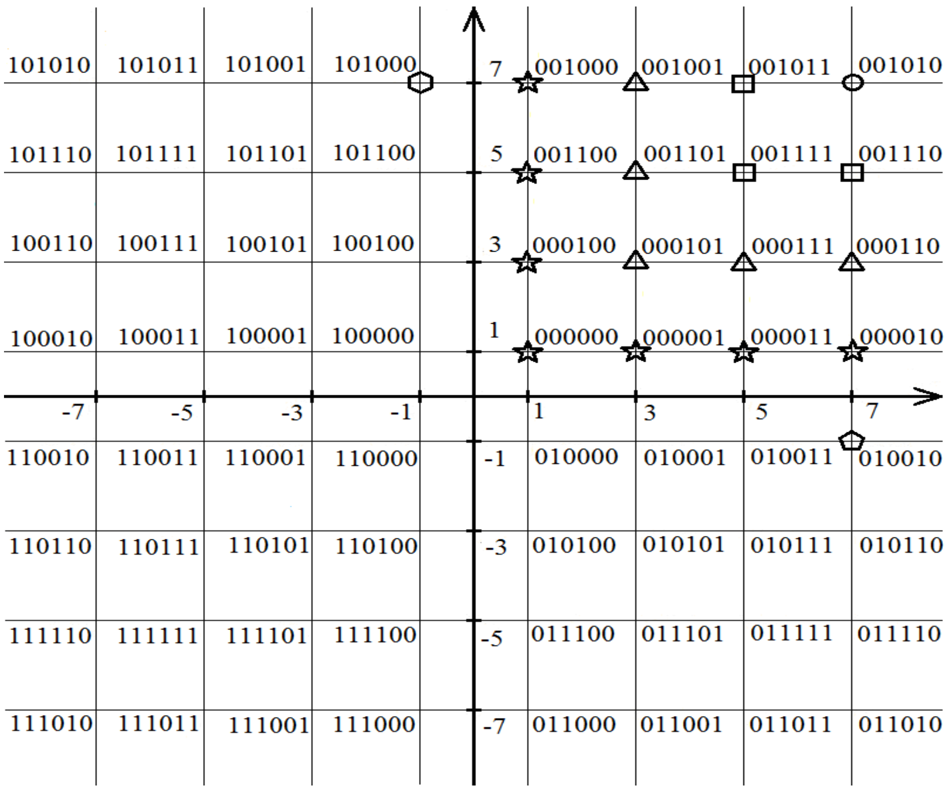

Fig. 4 *b* – Outlines for symbol 10 (001010).

In the following, for a number of received sequences corresponding to the same number of information blocks at transmitter, by ***detection time*** we mean the time in which Log-Likelihood Computation block in Fig. 2 computes LLR values for the coded bits and by ***total time*** we mean the time in which the entire receiver shown in Fig. 2 processes all the received sequences.

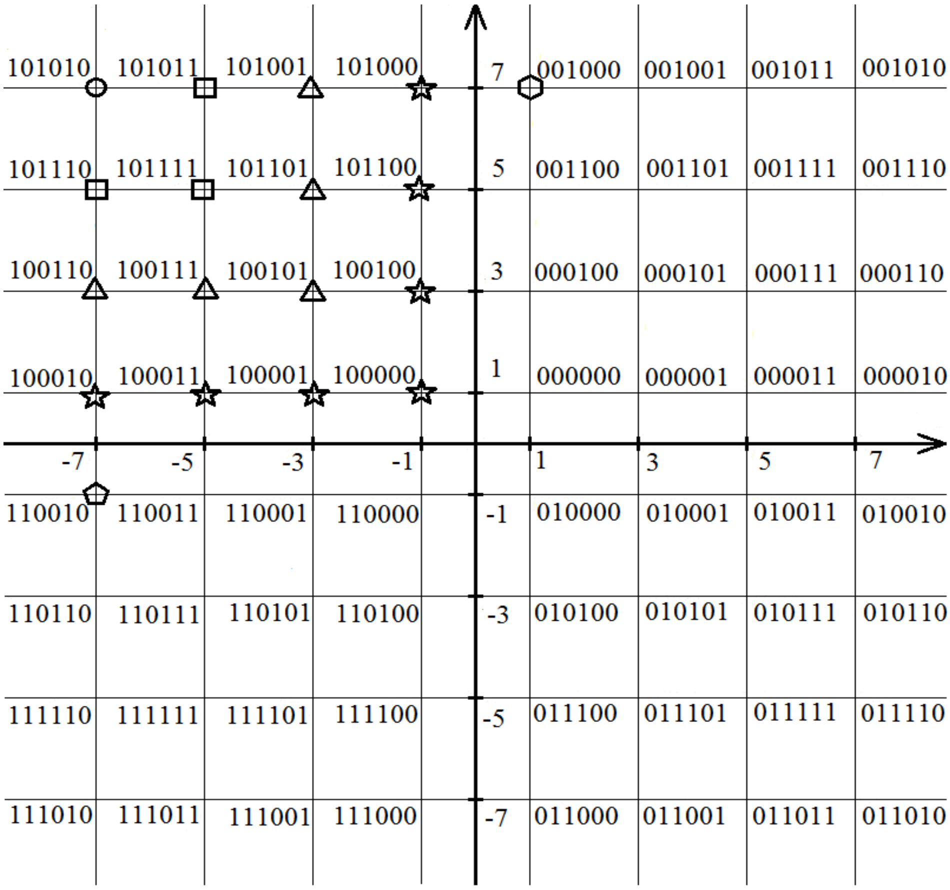

Fig. 4 *c* – Outlines for symbol 42 (101010).

We denote by DT_{orig} and RT_{orig} the detection time and the total processing time at receiver, respectively, for the original diagram from (Biglieri *et al*., 2005). Similarly, we denote by DT_i and RT_i, $i \in \{1,2\}$, the detection time and the total time at receiver, respectively, for the reduced complexity diagram using method number i proposed in this paper. Let $\alpha_i = DT_i / DT_{\text{orig}}$ be the percentage of the required detection time in the proposed diagram with method i compared to that in the original diagram and let $\beta = DT_{\text{orig}}/RT_{\text{orig}}$ be the percentage of the required detection time compared to the total time at receiver in the original diagram. Then for the proposed reduced complexity detection diagram with method i the total time at receiver is equal to a percentage of

$$1-\frac{RT_i}{RT_{orig}}=1-\frac{DT_i+(1-\beta)\cdot RT_{orig}}{RT_{orig}}=1-\frac{\alpha_i\cdot\beta\cdot RT_{orig}+(1-\beta)\cdot RT_{orig}}{RT_{orig}}=(1-\alpha_i)\cdot\beta, \quad (5)$$

compared with the total time at receiver in the original diagram.

4. Simulation Results

In this paper the simulations were performed for 64-QAM (M = 6), considering the same scenario as in (Rotopanescu *et al*., 2012). The turbo encoder uses a random interleaver of length 2112 for 64-QAM modulation. The global rate of the turbo code is 1/2 (with alternative puncturing of the parity bits). The forward and feedback generator polynomials are (5, 7) in octal form, and the interleaver between the turbo encoder and the serial to parallel convertor is a random one.

The number of transmit and receive antennas is 16. The space-time codeword is a matrix with 16 rows (the number of transmit antennas) and 44 columns for 64-QAM modulation. The number of distinct blocks with constant fading, F, is equal to 1. The turbo decoder uses the Max-Log-APP algorithm, described in (Trifina *et al*., 2011), and performs maximum 10 iterations.

To cancel spatial interference, k = 0, k = 1 and k = 4 outer iterations were used. In (Trifina *et al*., 2011), an analysis was performed to evaluate the influence of the extrinsic information scaling coefficient, denoted by s, on the BER/FER performance of the system with QPSK modulation. As in (Rotopanescu *et al*., 2012) and (Rotopanescu *et al*., 2016), we consider the same extrinsic information scaling coefficient that performs the best FER and BER performance. Therefore, for k = 0, the scaling coefficient is s = 0.9, for k = 1, s = 0.8 and for k = 4, s = 0.75. The Monte Carlo simulation results are given in Figs. 5 and 6, for BER and FER performances, respectively, using k = 0, 1 and 4. These figures show the performance of the MMSE doubly iterative receiver through BER/FER, versus signal-to-noise ratio per bit (E_b/N_0). The simulations are made for the original diagram and for the proposed reduced complexity detection diagram using the methods described in Section **3**. Fig. 5 and Fig. 6 represent the BER/FER performances for the three diagrams. We can see that these performances are almost the same for all diagrams.

From Fig. 6 we note that the coding gain increases proportionally with the number of outer iterations k, as we will analyze in what follows. Increasing the outer iteration number of the MMSE iterative decoder up to 4 leads to improved performance. In (Biglieri *et al*., 2005) it was shown that further increasing the number of outer iterations does not lead to additional performance improvement.

From simulation results we also observe that for k = 4, the outer iterations introduce relatively more errors compared to k = 1, and the supplementary coding gain obtained for k = 4 is smaller than the one achieved when k = 1.

Tables 3, 5 and 7 show the detection times and the receiver times, in seconds, for the original diagram from (Biglieri *et al*., 2005) and for the proposed diagram with Method 1, when 1000 information blocks are simulated, for k and s specified above.

Table 1

The Total Number of Symbols Considered in Method 1for which the Bit Number l, l=1, 2, ..., 6, is Equal to 0 or 1, for Symbol 42

Bit position (l)	The total number of symbols for which the bit number l is equal to 0	The total number of symbols for which the bit number l is equal to 1
1	1	17
2	17	1
3	9	9
4	10	8
5	9	9
6	10	8

Table 2

The Sums of the Total Number of Considered Symbols in the Proposed Method for which the Bit Number l, l=1, 2, ..., 6, is Equal to 0 or 1, for all 64 Symbols of the 64-QAM Constellation

Bit position (l)	The total number of symbols for which the bit number l is equal to 0		Percentage of the number of symbols considered in the proposed methods		The total number of symbols for which the bit number l is equal to 1		Percentage of the number of symbols considered in the proposed methods	
	Method 1	Method 2	Method 1	Method 2	Method 1	Method 2	Method 1	Method 2
1	984	968	48.05%	47.27%	984	968	48.05%	47.27%
2	984	968	48.05%	47.27%	984	968	48.05%	47.27%
3	1168	1144	57.03%	55.86%	800	792	39.06%	38.67%
4	992	968	48.44%	47.27%	976	968	47.66%	47.27%
5	1168	1144	57.03%	55.86%	800	792	39.06%	38.67%
6	992	968	48.44%	47.27%	976	968	47.66%	47.27%
All six bits	6288	6160	51.17%	50.13%	5520	5456	44.92%	44.40%

Table 3

Detection times and receiver times in seconds for the original diagram from (Biglieri et al., 2005) and for the proposed diagram with Method 1, for k=0, s=0.9, when 1000 information blocks are simulated

SNR (dB)	DT_{orig}	RT_{orig}	β	DT_1	RT_1	α_1	RT_1/RT_{orig} from simulation	$(1-\alpha_1)\beta$
3	9.160	72.385	12.65%	4.154	67.542	45.35%	93.31%	93.08%
5	9.387	50.872	18.45%	5.068	45.989	53.99%	90.40%	91.51%
7	9.083	37.357	24.31%	4.871	32.518	53.63%	87.05%	88.73%
9	8.926	31.276	28.54%	4.802	26.802	53.80%	85.70%	86.81%

Table 4
Detection Times and Receiver Times in Seconds for the Original Diagram from (Biglieri et al., 2005) and for the Proposed Diagram with Method 2, for k = 0, s = 0.9, when 1,000 Information Blocks are Simulated

SNR (dB)	DT_{orig}	RT_{orig}	β	DT_2	RT_2	α_2	RT_2/RT_{orig} from simulation	$(1-\alpha_2)\beta$
3	9.160	72.385	12.65%	4.686	67.670	51.16%	93.49%	93.82%
5	9.387	50.872	18.45%	4.712	46.247	50.20%	90.91%	90.81%
7	9.083	37.357	24.31%	4.268	32.812	46.99%	87.83%	87.11%
9	8.926	31.276	28.54%	4.276	26.629	47.90%	85.14%	85.13%

Table 5
Detection Times and Receiver Times in Seconds for the Original Diagram from (Biglieri et al., 2005) and for the Proposed Diagram with Method 1, for k = 1, s = 0.8, when 1,000 Information Blocks are Simulated

SNR (dB)	DT_{orig}	RT_{orig}	β	DT_1	RT_1	α_1	RT_1/RT_{orig} from simulation	$(1-\alpha_1)\beta$
3	18.296	129.165	14.16%	9.418	121.058	51.48%	93.72%	93.13%
5	18.436	85.050	21.68%	9.428	77.514	51.14%	91.14%	89.41%
7	18.569	65.801	28.22%	9.097	57.907	48.99%	88.00%	85.61%
9	18.796	57.295	32.81%	9.904	49.132	52.69%	85.75%	84.48%

Table 6
Detection Times and Receiver Times in Seconds for the Original Diagram from (Biglieri et al., 2005) and for the Proposed Diagram with Method 2, for k = 1, s = 0.8, when 1,000 Information Blocks are Simulated

SNR (dB)	DT_{orig}	RT_{orig}	β	DT_2	RT_2	α_2	RT_2/RT_{orig} from simulation	$(1-\alpha_2)\beta$
3	18.296	129.165	14.16%	8.978	120.497	51.48%	93.72%	93.13%
5	18.436	85.050	21.68%	8.935	76.630	48.46%	90.10%	88.83%
7	18.569	65.801	28.22%	9.352	56.837	50.36%	86.38%	85.99%
9	18.796	57.295	32.81%	9.350	47.434	49.74%	82.79%	83.51%

Table 7
Detection Times and Receiver Times in Seconds for the Original Diagram from (Biglieri et al., 2005) and for the Proposed Diagram with Method 1, for k = 4, s = 0.75, when 1,000 Information Blocks are Simulated

SNR (dB)	DT_{orig}	RT_{orig}	β	DT_1	RT_1	α_1	RT_1/RT_{orig} from simulation	$(1-\alpha_1)\beta$
3	46.708	285.052	16.39%	22.255	259.451	47.65%	91.02%	91.42%
5	46.488	186.578	24.92%	22.974	162.407	49.42%	87.05%	87.40%
7	46.936	153.978	30.48%	22.471	129.733	47.88%	84.25%	84.11%
9	46.497	136.732	34.01%	22.596	112.921	48.60%	82.59%	82.52%

Table 8

Detection Times and Receiver Times in Seconds for the Original Diagram from (Biglieri et al., 2005) and for the Proposed Diagram with Method 2, for k = 4, s = 0.75, when 1,000 Information Blocks are Simulated

SNR (dB)	DT_{orig}	RT_{orig}	β	DT_2	RT_2	α_2	RT_2/RT_{orig} from simulation	$(1-\alpha_2)\beta$
3	46.708	285.052	16.39%	22.647	261.306	48.49%	91.67%	91.56%
5	46.488	186.578	24.92%	22.627	163.605	48.67%	87.69%	87.21%
7	46.936	153.978	30.48%	21.859	130.231	46.57%	84.58%	83.71%
9	46.497	136.732	34.01%	23.408	113.074	50.34%	82.70%	83.11%

Similarly, Tables 4, 6 and 8 show the detection times and the receiver times for the original diagram from (Biglieri *et al.*, 2005) and for the proposed diagram with Method 2. Percentages α_i, β, $(1-\alpha_i)\beta$, $i \in \{1,2\}$, defined at the end of Section **3**, and percentage RT_1/RT_{orig} resulted from simulation are also given in these tables.

We see that the receiver time for the proposed diagram with both methods, when outer iterations are performed, decreases with about 7% up to 18% compared to the original diagram from (Biglieri *et al.*, 2005). The complexity reducing is higher for high SNR and for a bigger number of outer iterations because in these cases the turbo decoder performs less iterations.

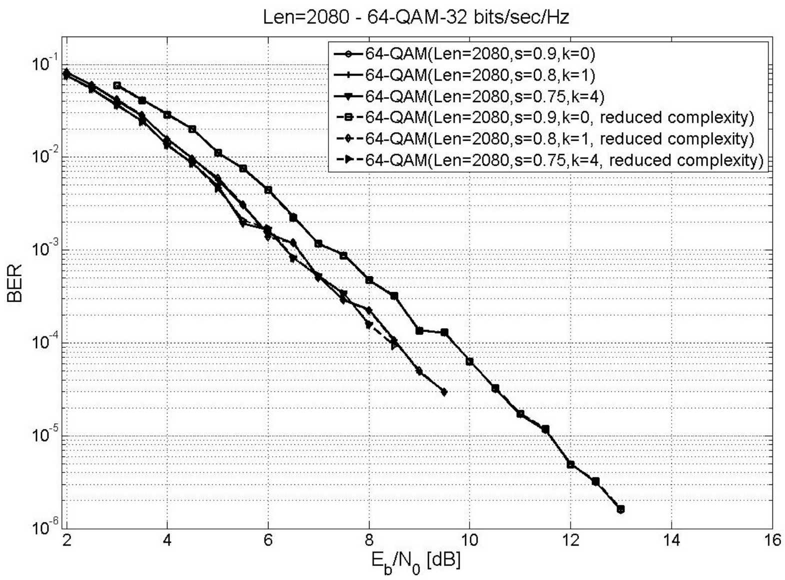

Fig. 5 – BER performances for the original diagram from (Biglieri *et al.*, 2005) and for the reduced complexity proposed diagram with both methods for 64-QAM modulation and for *k* = 0, *k* = 1, and *k* = 4 outer iterations.

Percentages α_i resulted from simulation are close to the average values of 48.05% or 47.27% theoretically determined in Section **3**. Percentage β increases along with the number of outer iterations and along with the value of SNR because in both cases the number of iterations performed by the turbo decoder decreases, while the detection time remains the same. Finally, we see that the percentages of RT_1/RT_{orig} values resulted from simulation are close to $(1-\alpha_i)\beta$.

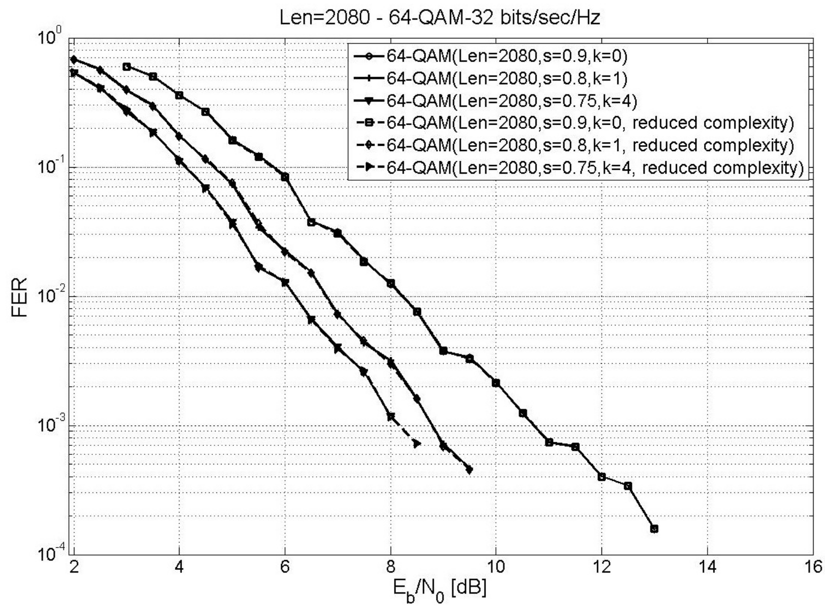

Fig. 6 – FER performances for the original diagram from (Biglieri *et al*., 2005) and for the reduced complexity proposed diagram with both methods for 64-QAM modulation and for k = 0, k = 1, and k = 4 outer iterations.

5. Conclusions

In this paper we have presented two methods to reduce the complexity detection in a doubly iterative receiver for MIMO transmissions with a large number of antennas. The proposed methods reduce the time for the computation of LLR for the current coded bit compared to the original detection diagram from (Biglieri *et al*., 2005). For 64-QAM modulation, the simplified detection methods consist in considering a certain number of symbols belonging to the first three outlines around the demodulated symbol, instead of all 64 symbols as in (Biglieri *et al*., 2005).

In Method 1 a number of additional points are added if all the outlines have no symbol containing the bit value 0 or 1 in one of the six possible bits for a symbol, so that there is no information for the respective bit, and thus there is no value corresponding to the sums from the numerator or denominator of LLR in formula (2).

In Method 2 the additional points are not required. Instead, the squared value of the distance from the received point to the additional point, in formula (2), is replaced by the corresponding distance from the demodulated symbol to the additional point. It was shown that for 64-QAM the detection time with the two proposed methods, averaged over all 64 symbols, is reduced by about 48% and 47.27%, respectively, compared to the original diagram from (Biglieri *et al.*, 2005).

The total time required for processing at receiver decreases with about 7% up to 18%. The complexity reducing is higher for high SNR values and for a bigger number of outer iterations because in both cases turbo decoder performs less iterations, and thus the percentage of detection time from the receiver time is bigger.

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DETECTOR SOFT DE COMPLEXITATE REDUSĂ PENTRU RECEPTOR DUBLU ITERATIV FOLOSIND 64-QAM

(Rezumat)

Eficiența spectrala pentru sistemele de comunicații poate creşte folosind modulații flexibile şi scheme de codare cu ordin mare, cum ar fi 64-QAM (Modulatie de amplitudinte in cuadratura). În această lucrare au fost propuse două metode care reduc complexitatea detecției soft pentru un decodor dublu iterativ ce foloseşte coduri turbo spațio-temporale şi un număr mare de antene de transmisie şi recepție. Metodele de calcul propuse pentru reducerea complexității a raportului de plauzibilitate constau în considerarea unui anumit număr de simboluri în loc de toate 64. În prima metoda sunt necesare câteva puncte adiționale atunci când nu există informație pentru un bit din cei 6 corespunzători unui simbol. În cea de-a doua metodă, punctele adiționale nu mai sunt folosite, dar se utilizează o distanță fixă în numărătorul şi numitorul raportului de plauzibilitate. Timpul de detecție este redus cu aproximativ 48% fată de schema originală, în timp ce timpul total necesar pentru schema receptorului scade cu 7% până la 18%, depinzând de valoarea SNR-ului şi de numărul de iterații.