An Efficient Soft Output Computation for coded STBC Systems

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***Abstract*— Space time block codes (STBC) provide transmit diversity by using multiple transmit antennas. We obtain a lattice representation of the STBC by converting the complex channel model into a real matrix. In this paper, we propose a computationally efficient demapper in the coded STBC systems. Based on independently divided maximum likelihood (ML) decoding, we derive a simple equation for the soft bit metrics. With the efficient soft bit metrics, the computational complexity generally reduces more than 80%. By employing the proposed soft value generation, the computational complexity can be significantly decreased without any performance loss.**

I. Introduction

In future wireless digital communication systems, multiple antennas are expected to be employed in order to improve the system performance. Multiple antennas can be used at the transmitter and receiver, as a multi-input multi-output (MIMO) system. A MIMO system takes advantage of the spatial diversity and the multiplexing gain by utilizing spatially separated antennas in a dispersive fading environment [1] [2]. There are two approaches for the MIMO systems; space time coding (STC) and spatial division multiplexing (SDM). STC techniques are designed to provide transmit diversity [3] [4] [5]. Among them, space time block codes (STBC) exhibit an advantage of a simple decoding compared to trellis based codes.

When large signal constellations and/or many transmit antennas are involved, the complexity of maximum likelihood (ML) decoding in STBC systems becomes prohibitive. Therefore, there have been many efforts to reduce the complexity of the ML decoding such as sphere decoding with the linear lattice structure [6]. Most recently a new ML algorithm for uncoded STBC systems using a lattice representation is introduced in [7]. In [7], it has been shown that certain structures such as orthogonal designs allow a separate ML decoding of each in-phase/quadrature component by utilizing the lattice structure of the STBC. Also the equivalence between the symbol-wise ML decoding and the component-wise ML decoding is illustrated with reduced computational complexity in [7].

The STBC gurantees the higher transmit diversity but there is a loss in performance [5]. To overcome this disadvantage, the STBC is concatenated with other channel coding techniques such as convolutional codes and turbo codes showing good performance [8]. Such coded STBC systems with bit-level interleaver between the encoder and the modulator can obtain better performance and an additional time diversity.

The demapper is responsible for generating the log likelihood ratio (LLR) values for soft input Viterbi decoder in the single antenna case and the MIMO system [9] [10]. In this paper, we extend the results in [7] to the coded STBC systems which combine the STBC and convolutional code and adopt a new concept for the demapper implementation. The MIMO demapper in coded systems requires high computational complexity [11]. By utilizing the component-wise ML decoding in [7], we show that the complexity $\mathcal{O}(M_c)$ for the orthogonal design and $\mathcal{O}({M_c}^2)$ for the quasi-orthogonal design reduces to $\mathcal{O}({M_c}^{\frac{1}{2}})$ and $\mathcal{O}(M_c)$, respectively, where M_c indicates the constellation size. This proposed scheme can even be employed to the rotated quasi-orthogonal design [12]. Also we verify that the proposed method incurs no performance loss under various simulation circumstances.

This paper is organized as follows: In section II, we briefly present the system model and the lattice representation of STBC. In section III, we introduce an efficient LLR generation in various STBC designs based on the lattice representation. In section IV, we compare the computational complexity for the original LLR scheme and the proposed method. The simulation results of the proposed LLR scheme are presented in section V. Finally the paper is terminated with a conclusion in section VI.

II. System Model

We consider the following space time coding scheme as

$$\bar{\mathbf{y}} = \mathbf{C}\bar{\mathbf{h}} + \bar{\mathbf{n}} \tag{1}$$

where $\mathbf{C}$ represents the code matrix, $\bar{\mathbf{h}}$ indicates the complex channel vector and $\bar{\mathbf{n}}$ represents the independent and identically-distributed (i.i.d) additive complex Gaussian vector

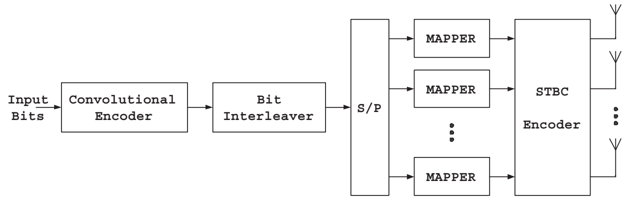

Fig. 1. Coded STBC transmitter structure

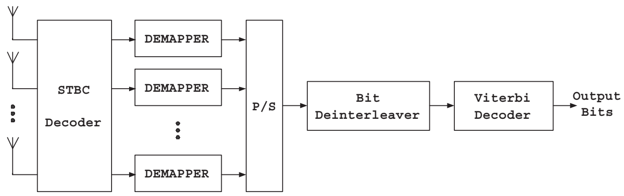

Fig. 2. Coded STBC receiver structure

with variance σ_n^2. The ith column $\mathbf{c}_i$ of the matrix $\mathbf{C}$ indicates the transmit data stream through the ith transmit antenna.

First, we define the transmit vector $\bar{\mathbf{x}} = [x_1\ x_2 \cdots\ x_{N_t}]^t$. Here $[\ ]^t$ denotes the transpose of a matrix and subscript N_t indicates the number of transmit antennas. According to the linear lattice representation illustrated in [6], the received signal vector (1) can be decomposed as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \tag{2}$$

where $\mathbf{y} = [\ \Re[\bar{\mathbf{y}}^t]\ \Im[\bar{\mathbf{y}}^t]\ ]^t$, $\mathbf{x} = [\ \Re[\bar{\mathbf{x}}^t]\ \Im[\bar{\mathbf{x}}^t]\ ]^t$ and $\mathbf{n} = [\ \Re[\bar{\mathbf{n}}^t]\ \Im[\bar{\mathbf{n}}^t]\ ]^t$. Here $\mathbf{H} \in R^{2T\times 2N_t}$ defines the effective channel matrix and T is the block size of the STBC. Figures 1 and 2 show the transmitter structure and the receiver structure of the coded STBC system in the bit-interleaved coded modulation (BICM) structure, respectively. We use a convolutional code with code rate R_c in this structure. In Figure 2, the demapper generates the LLR values, where the LLR value $LLR(u)$ is defined as

$$LLR(u) = \log\frac{P[u=1]}{P[u=0]}.$$

Here $P[u]$ denotes the probability that the random variable takes on the value u.

III. Efficient LLR Computation for STBC

In this section, we present an orthogonal STBC scheme with $N_t = 2$ and 4. First we derive the simplified LLR equation for the orthogonal code designs [3] [4]. Then we extend to the quasi-orthogonal case with $N_t = 4$ [13].

A. The Alamouti scheme ($N_t = 2$)

The Alamouti code provides full rate and full diversity in the complex orthogonal signal space with two transmit antennas. The code construction $\mathbf{C}$ is given by

$$\mathbf{C} = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix}.$$

We transform the received vector into (2) as

$$\begin{bmatrix} y_{1,I} \\ y_{2,I} \\ y_{1,Q} \\ y_{2,Q} \end{bmatrix} = \begin{bmatrix} h_{1,I} & h_{2,I} & -h_{1,Q} & -h_{2,Q} \\ -h_{2,I} & h_{1,I} & -h_{2,Q} & h_{1,Q} \\ h_{1,Q} & h_{2,Q} & h_{1,I} & h_{2,I} \\ -h_{2,Q} & h_{1,Q} & h_{2,I} & -h_{1,I} \end{bmatrix} \begin{bmatrix} x_{1,I} \\ x_{2,I} \\ x_{1,Q} \\ x_{2,Q} \end{bmatrix} + \begin{bmatrix} n_{1,I} \\ n_{2,I} \\ n_{1,Q} \\ n_{2,Q} \end{bmatrix}.$$

Note that the columns of the above equivalent channel matrix are orthogonal with each other. From the orthogonal property in [7], the ML solution can be simply divided into a set of in-phase and quadrature components as

$$\hat{x}_{1,I} = \arg\min_{x_{1,I}\in Z} ||\mathbf{y} - \mathbf{h}_1 x_{1,I}||^2 = \arg\min_{x_{1,I}\in Z} \left|\left|\bar{\mathbf{y}} - \begin{bmatrix} \bar{h}_1 \\ \bar{h}_2 \end{bmatrix} x_{1,I}\right|\right|^2 \tag{3}$$

$$\hat{x}_{2,I} = \arg\min_{x_{2,I}\in Z} ||\mathbf{y} - \mathbf{h}_2 x_{2,I}||^2 = \arg\min_{x_{2,I}\in Z} \left|\left|\bar{\mathbf{y}} - \begin{bmatrix} \bar{h}_2 \\ -\bar{h}_1 \end{bmatrix} x_{2,I}\right|\right|^2 \tag{4}$$

$$\hat{x}_{1,Q} = \arg\min_{x_{1,Q}\in Z} ||\mathbf{y} - \mathbf{h}_3 x_{1,Q}||^2 = \arg\min_{x_{1,Q}\in Z} \left|\left|\bar{\mathbf{y}} - j\begin{bmatrix} \bar{h}_1 \\ -\bar{h}_2 \end{bmatrix} x_{1,Q}\right|\right|^2 \tag{5}$$

$$\hat{x}_{2,Q} = \arg\min_{x_{2,Q}\in Z} ||\mathbf{y} - \mathbf{h}_4 x_{2,Q}||^2 = \arg\min_{x_{2,Q}\in Z} \left|\left|\bar{\mathbf{y}} - j\begin{bmatrix} \bar{h}_2 \\ \bar{h}_1 \end{bmatrix} x_{2,Q}\right|\right|^2 \tag{6}$$

where $\mathbf{h}_i$ indicates the ith column of the equivalent channel matrix $\mathbf{H}$.

Next, based on the above new ML metrics (3)-(6), we derive a computationally efficient method for generating soft LLR values for Viterbi decoder input. Let $d_{i,m}$ be the mth bit ($m = 1, 2, \cdots, \log_2 M_c$) of the constellation symbol at the ith transmit antenna ($i = 1, 2$). Then the conventional LLR equation for the Alamouti system is denoted as

$$\begin{aligned} LLR(d_{i,m}) &\triangleq \log\frac{P[d_{i,m}=1|\mathbf{y}]}{P[d_{i,m}=0|\mathbf{y}]} \\ &= \log\frac{\sum_{\mathbf{x}\in S^1_{i,m}} P[\mathbf{x}|\mathbf{y},\mathbf{H}]}{\sum_{\mathbf{x}\in S^0_{i,m}} P[\mathbf{x}|\mathbf{y},\mathbf{H}]} \end{aligned} \tag{7}$$

where $S^b_{i,m}$ denotes a set consisting of symbol vectors whose mth bit $d_{i,m}$ is equal to $b = 0$ or 1.

We consider the new LLR value based on the lattice representation. Denoting $d^I_{i,n}$ and $d^Q_{i,n}$ as the nth ($n = 1, 2, \cdots, \frac{\log_2 M_c}{2}$) bit of the symbol at the ith transmit antenna, for the in-phase and quadrature component, respectively, the LLR derivation becomes

$$\begin{aligned} LLR(d^I_{i,n}) &\triangleq \log\frac{P[d^I_{i,n}=1|\bar{\mathbf{y}},\bar{\mathbf{h}}]}{P[d^I_{i,n}=0|\bar{\mathbf{y}},\bar{\mathbf{h}}]} \\ &= \log\frac{\sum_{\mathbf{x}\in X^1_{i,n,I}} P[\mathbf{x}|\mathbf{y},\mathbf{H}]}{\sum_{\mathbf{x}\in X^0_{i,n,I}} P[\mathbf{x}|\mathbf{y},\mathbf{H}]} \end{aligned} \tag{8}$$

$$\begin{aligned} LLR(d^Q_{i,n}) &\triangleq \log\frac{P[d^Q_{i,n}=1|\bar{\mathbf{y}},\bar{\mathbf{h}}]}{P[d^Q_{i,n}=0|\bar{\mathbf{y}},\bar{\mathbf{h}}]} \\ &= \log\frac{\sum_{\mathbf{x}\in X^1_{i,n,Q}} P[\mathbf{x}|\mathbf{y},\mathbf{H}]}{\sum_{\mathbf{x}\in X^0_{i,n,Q}} P[\mathbf{x}|\mathbf{y},\mathbf{H}]} \end{aligned} \tag{9}$$

where a symbol subset $X^b_{i,n,I}$ consists of symbol vectors whose nth in-phase bit $d^I_{i,n}$ at the ith transmit antenna is equal to $b = 0$ or 1. $X^b_{i,n,Q}$ is similarly defined for quadrature bits. The new LLR metrics (8) and (9) give rise to the calculation savings because the candidate subset changes from $S^b_{i,m}$ to $X^b_{i,n,I}$ and $X^b_{i,n,Q}$. The subset $S^b_{i,m}$ in (7) consists of $M_c/2$ complex symbols, while the newly defined subset $X^b_{i,n,I}$ and $X^b_{i,n,Q}$ are composed of $\sqrt{M_c}/2$ real number components. Therefore when computing the LLR values, the conventional method in (7) requires the complex operations with M_c candidates. In contrast, the proposed method as in (8) and (9) requires the real operations with $\sqrt{M_c}$ candidates. Hence the proposed LLR derivation gives a considerable computational reduction. A detailed explanation of the complexity reduction amounts will be given in section IV.

B. Orthogonal STBC design ($N_t = 4$)

The maximal rate full diversity STBC with four transmit antennas is considered in [4]. It was shown in [5] that the complex orthogonal design for $N_t > 2$ is unable to achieve full rate and full diversity simultaneously. The rate 3/4 orthogonal design for $N_t = T = 4$ introduced in [4] has the structure as

$$\mathbf{C} = \begin{bmatrix} x_1 & x_2 & x_3 & 0 \\ -x_2^* & x_1^* & 0 & -x_3 \\ -x_3^* & 0 & x_1^* & x_2 \\ 0 & x_3^* & -x_2^* & x_1 \end{bmatrix}.$$

Then the effective channel model can be rewritten as

$$\mathbf{H} = \begin{bmatrix} h_{1,I} & h_{2,I} & h_{3,I} & -h_{1,Q} & -h_{2,Q} & -h_{3,Q} \\ h_{2,I} & -h_{1,I} & -h_{4,I} & h_{2,Q} & -h_{1,Q} & h_{4,Q} \\ h_{3,I} & h_{4,I} & -h_{1,I} & h_{3,Q} & -h_{4,Q} & -h_{1,Q} \\ h_{4,I} & -h_{3,I} & h_{2,I} & -h_{4,Q} & -h_{3,Q} & h_{2,Q} \\ h_{1,Q} & h_{2,Q} & h_{3,Q} & h_{1,I} & h_{2,I} & h_{3,I} \\ h_{2,Q} & -h_{1,Q} & -h_{4,Q} & -h_{2,I} & h_{1,I} & -h_{4,I} \\ h_{3,Q} & h_{4,Q} & -h_{1,Q} & -h_{3,I} & h_{4,I} & h_{1,I} \\ h_{4,Q} & -h_{3,Q} & h_{2,Q} & h_{4,I} & h_{3,I} & -h_{2,I} \end{bmatrix}.$$

Following the observations in [7] for the above effective channel $\mathbf{H}$, we note that the ML estimations of $x_{i,I}$ and $x_{i,Q}$ for $i = 1, 2, 3$ are decoupled for each in-phase/quadrature component. For $x_1 = x_{1,I} + jx_{1,Q}$, as an example, the ML estimations of $\hat{x}_{1,I}$ and $\hat{x}_{1,Q}$ are obtained by

$$\begin{aligned} \hat{x}_{1,I} &= \arg\min_{x_{1,I} \in Z} ||\mathbf{y} - \mathbf{h}_1 x_{1,I}||^2 \\ &= \arg\min_{x_{1,I} \in Z} ||\bar{\mathbf{y}} - \bar{\mathbf{h}}_{1,I} x_{1,I}||^2 \end{aligned}$$

$$\begin{aligned} \hat{x}_{1,Q} &= \arg\min_{x_{1,Q} \in Z} ||\mathbf{y} - \mathbf{h}_4 x_{1,Q}||^2 \\ &= \arg\min_{x_{1,Q} \in Z} ||\bar{\mathbf{y}} - \bar{\mathbf{h}}_{1,Q} x_{1,Q}||^2 \end{aligned}$$

where $\bar{\mathbf{h}}_{1,I} = [\bar{h}_1 \ \bar{h}_2 \ \bar{h}_3 \ \bar{h}_4]^t$, $\bar{\mathbf{h}}_{1,Q} = [\bar{h}_1 \ -\bar{h}_2 \ -\bar{h}_3 \ \bar{h}_4]^t$ and $\mathbf{h}_i$ indicates the ith column of the channel matrix $\mathbf{H}$. In the same manner as for the Alamouti code case, we adopt the simplified ML estimation to calculate the LLR values in the orthogonal STBC demapper. Thereby, the LLR values can be calculated as

$$LLR(d^I_{i,n}) \triangleq \log \frac{\sum_{\mathbf{x} \in X^1_{i,n,I}} P[\mathbf{x}|\mathbf{y},\mathbf{H}]}{\sum_{\mathbf{x} \in X^0_{i,n,I}} P[\mathbf{x}|\mathbf{y},\mathbf{H}]}. \tag{10}$$

Again the LLR values for the quadrature components can be calculated similarly as in (10).

C. Quasi Orthogonal STBC design ($N_t = 4$)

In this section, we consider quasi-orthogonal designs for their ML estimation using the lattice representation. Compared to the orthogonal codes, the quasi-orthogonal codes are designed to achieve the full rate at the expense of the orthogonality of the code matrix $\mathbf{C}$ [12] [13]. Then, we extend the reduced complexity demapper derivation for this design. Note that the quasi-orthogonal code design is not unique. In this paper, we deal with one of quasi-orthogonal codes introduced with the code matrix in [12] [13] $\mathbf{C}$ [1]

$$\mathbf{C} = \begin{bmatrix} x_1 & x_3 & x_4 & x_2 \\ x_3^* & -x_1^* & x_2^* & -x_4^* \\ x_4^* & x_2^* & -x_1^* & -x_3^* \\ x_2 & -x_4 & -x_3 & x_1 \end{bmatrix}.$$

We can rewrite the corresponding effective channel $\mathbf{H}$ as

$$\begin{aligned} \mathbf{H} &= [\mathbf{h}_1 \ \mathbf{h}_2 \cdots \ \mathbf{h}_8] \\ &= \begin{bmatrix} h_{1,I} & h_{4,I} & h_{2,I} & h_{3,I} & -h_{1,Q} & -h_{4,Q} & -h_{2,Q} & -h_{3,Q} \\ -h_{2,I} & h_{3,I} & h_{1,I} & -h_{4,I} & -h_{2,Q} & h_{3,Q} & h_{1,Q} & -h_{4,Q} \\ -h_{3,I} & h_{2,I} & -h_{4,I} & h_{1,I} & -h_{3,Q} & h_{2,Q} & -h_{4,Q} & h_{1,Q} \\ h_{4,I} & h_{1,I} & -h_{3,I} & -h_{2,I} & -h_{4,Q} & -h_{1,Q} & h_{3,Q} & h_{2,Q} \\ h_{1,Q} & h_{4,Q} & h_{2,Q} & h_{3,Q} & h_{1,I} & h_{4,I} & h_{2,I} & h_{3,I} \\ -h_{2,Q} & h_{3,Q} & h_{1,Q} & -h_{4,Q} & h_{2,I} & -h_{3,I} & -h_{1,I} & h_{4,I} \\ -h_{3,Q} & h_{2,Q} & -h_{4,Q} & h_{1,Q} & h_{3,I} & -h_{2,I} & h_{4,I} & -h_{1,I} \\ h_{4,Q} & h_{1,Q} & -h_{3,Q} & -h_{2,Q} & h_{4,I} & h_{1,I} & -h_{3,I} & -h_{2,I} \end{bmatrix}. \end{aligned}$$

In this code, pairs of the channel vectors $\{\mathbf{h}_1 \ \mathbf{h}_2\}$, $\{\mathbf{h}_3 \ \mathbf{h}_4\}$, $\{\mathbf{h}_5 \ \mathbf{h}_6\}$ and $\{\mathbf{h}_7 \ \mathbf{h}_8\}$ constitute quasi-orthogonal subsets. As shown in [7], the ML decoding can be carried out with pairs of the in-phase/quadrature components $\{x_{1,I}, x_{2,I}\}$, $\{x_{1,Q}, x_{2,Q}\}$, $\{x_{3,I}, x_{4,I}\}$ and $\{x_{3,Q}, x_{4,Q}\}$. Therefore, the joint ML estimation $\{\hat{x}_{1,I}, \hat{x}_{2,I}\}$, for example, can be obtained by

$$\begin{aligned} \{\hat{x}_{1,I}, \hat{x}_{2,I}\} &= \arg\min_{[x_{1,I}, x_{2,I}] \in Z^2} \left|\left| \mathbf{y} - [\mathbf{h}_1 \ \mathbf{h}_2] \begin{bmatrix} x_{1,I} \\ x_{2,I} \end{bmatrix} \right|\right|^2 \qquad (11) \\ &= \arg\min_{[x_{1,I}, x_{2,I}] \in Z^2} \left|\left| \bar{\mathbf{y}} - \begin{bmatrix} \bar{h}_1 & \bar{h}_4 \\ -\bar{h}_2 & \bar{h}_3 \\ -\bar{h}_3 & \bar{h}_2 \\ \bar{h}_4 & \bar{h}_1 \end{bmatrix} \begin{bmatrix} x_{1,I} \\ x_{2,I} \end{bmatrix} \right|\right|^2. \end{aligned}$$

Compared to the conventional symbol level joint ML detection, the component-wise joint ML detection proposed in [7] can greatly reduce the complexity for calculating the LLR

[1] In this paper, we consider a slightly different quasi-orthogonal design to simplify the analysis.

values. In the conventional demapper block, the soft bit values for quasi-orthogonal design are generated by

$$LLR(d_{i,m}) \triangleq \log \frac{\sum_{\mathbf{x}\in D^1_{i,m}} P[\mathbf{x}|\mathbf{y},\mathbf{H}]}{\sum_{\mathbf{x}\in D^0_{i,m}} P[\mathbf{x}|\mathbf{y},\mathbf{H}]}$$

where the subset $D^b_{i,m}$ consists of two symbol vector $\mathbf{x}$ whose mth bit in one of the two symbols $d_{i,m}$ is equal to $b = 0$ or 1. Using this simplified ML estimate (11), we can also represent the LLR values in the quasi-orthogonal system. The LLR value corresponding to the nth in-phase bit at the ith transmit antenna is given by

$$LLR(d^I_{i,n}) \triangleq \log \frac{\sum_{[x_{2i-1,I},x_{2i,I}]\in Y^1_{i,n,I}} P[x_{2i-1,I},x_{2i,I}|\mathbf{y},\mathbf{h}_{2i-1}\mathbf{h}_{2i}]}{\sum_{[x_{2i-1,I},x_{2i,I}]\in Y^0_{i,n,I}} P[x_{2i-1,I},x_{2i,I}|\mathbf{y},\mathbf{h}_{2i-1}\mathbf{h}_{2i}]}$$

where $Y^b_{i,n,I}$ defines a subset of two real number elements for the corresponding bit $d^I_{i,n}$. The LLR for the quadrature bit $d^Q_{i,n}$ can be calculated similarly as in the case of in-phase bit. We also confirm that the proposed scheme can be adopted to the rotated quasi-orthogonal design [12] without performance loss.

IV. Complexity Comparison

In the preceding section, we show that the proposed LLR computation leads to the reduction of operations for the demapping process. In the following, we will compare the complexity for the conventional LLR scheme and the proposed method. The overall complexity of the LLR calculation is measured by the number of real operations required to determine the soft value for one bit in the demapping process.

The complexity comparison is determined under the following assumptions, [14]:

- The unit operand of additions and multiplications are defined as $1N_A$ and $1N_M$, respectively.
- A complex multiplication is equal to 4 real multiplications and 2 real additions $4N_M + 2N_A$.
- A complex addition is equal to 2 real additions $2N_A$.
- The norm and square operation can also be treated as one complex multiplication.

In this section, we consider not only multiplications but also additions to compare the complexity in detail. Table I compares the complexity for both the conventional scheme and the proposed LLR method in terms of the total number of operations. We first consider the orthogonal desgin (OD) with $N_t = 2$ (the Alamouti scheme). The first two terms are a function of M_c which account for the LLR computation in the denominator and numerator, while the last two terms take charge of the matched filter operation. For example, when the 16QAM modulation scheme is employed, the computation requirement reduces from $120N_M + 81N_A$ to $20N_M + 12N_A$, resulting in more than 80% reduction in the computational complexity.

The OD with $N_t = 4$ exhibits a similar tendency to the case of $N_t = 2$. In the case of 16QAM constellation, we can reduce the computational complexity by $1/4$ as the number of operations decreases from $128N_M + 89N_A$ to $32N_M + 22N_A$.

For the quasi-orthogonal (QO) designs, the computational savings are more significant, since we jointly decode two real components instead of two complex symbols as in the conventional case. Also shown in the third row of Table I, the number of operations increases exponentially with the number of antennas. For example, when we employ 16QAM with $N_t = 4$, the complexity reduction by a factor of 10 is achieved.

Figure 3 compares the number of operations for the LLR values per bit for the OD and the QO with $N_t \times N_r$ antenna configurations. Here the number of receive antennas N_r is fixed 1. In Figure 3, the dashed line indicates the conventional LLR scheme and the solid line represents the proposed LLR scheme. In this figure the quasi-orthogonal case shows the highest complexity savings for the proposed method. As the constellation size grows, the complexity for the proposed LLR scheme becomes much lower than the conventional scheme. Hence, for systems with high modulations, the proposed LLR scheme is a very promising technique in terms of the computational complexity.

TABLE I
Computation complexity comparison between the conventional scheme and the proposed LLR method.

	Conventional	Proposed
2×1 OD	$M_c(6N_M + 4N_A) + 24N_M + 17N_A$	$\sqrt{M_c}(2N_M + N_A) + 12N_M + 8N_A$
4×1 OD	$M_c(6N_M + 4N_A) + 32N_M + 25N_A$	$\sqrt{M_c}(2N_M + 1N_A) + 24N_M + 18N_A$
QO	$M_c^{\frac{N_t}{2}}\{(2N_t^2 + 4N_t)N_M + (2N_t^2 + 3N_t - 1)N_A\}$	$M_c^{\frac{N_t}{4}}\{(N_t^2 + 4N_t)N_M + (N_t^2 + 3N_t - 1)N_A\}$
4×1 QO	$M_c^2(48N_M + 43N_A)$	$M_c(32N_M + 27N_A)$

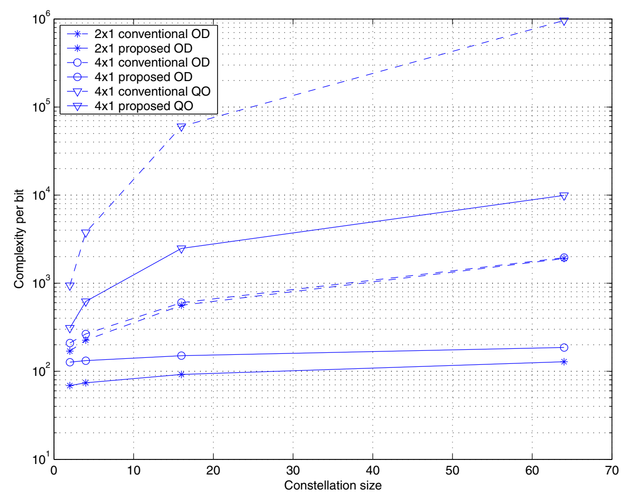

Fig. 3. LLR computation complexity comparison

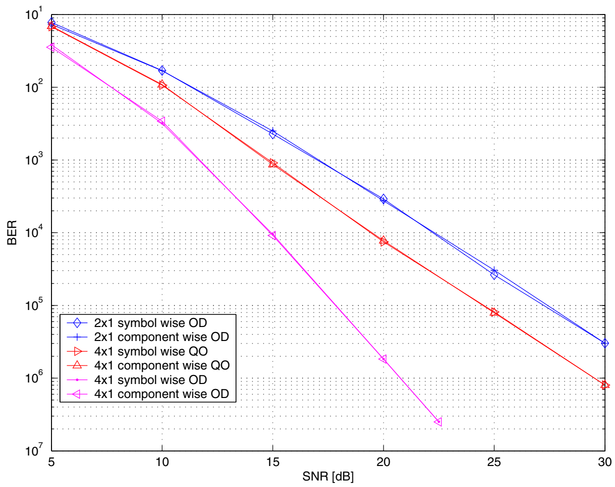

Fig. 4. Uncoded BER performance

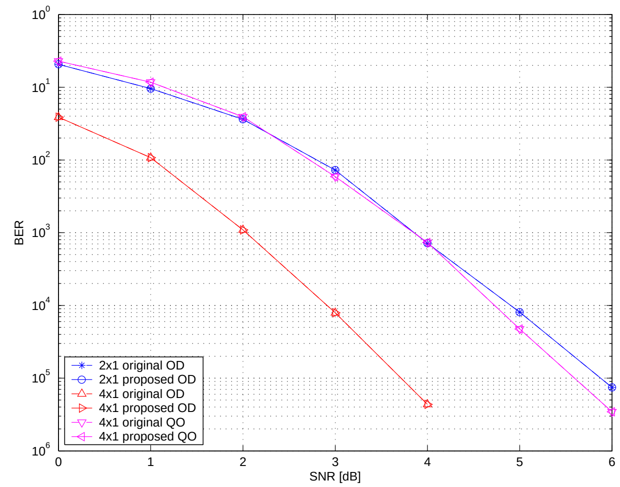

Fig. 5. BER performance of coded STBC systems

V. Simulation Results

In this section, we present simulation results to illustrate the validity of the proposed demapper scheme in the MIMO systems. In the simulation results, the x-axis represents the received signal to noise ratio (SNR) in dB and the y-axis indicates the bit error rate (BER).

First, Figure 4 depicts the performance comparison of the component-wise decoding algorithm and the conventional decoding in uncoded systems over flat fading channels. This simulation result confirms that the component-wise decoding algorithm provides the performance identical to the conventional symbol-wise decoding detection in uncoded systems.

Figure 5 demonstrates the performance of our proposed demapper in the coded STBC systems. In this simulation, the convolutional code memory is set to 6. We assume an independent Rayleigh fading channel and use the Gray labelled mapping. The code rate R_c is set to $1/2$ and 4QAM is employed. As a reult, 2×1 OD, 4×1 OD and 4×1 QO show the spectral efficiencies of 1, 0.75 and 1 bps/Hz, respectively.

In Figure 5, we confirm that the performance curves of two methods are identical in all systems, 2×1 OD, 4×1 OD and 4×1 QO. Therefore, the proposed demapper achieves the same performance as the original demapper with significantly lower complexity.

VI. Conclusion

In this paper, we have proposed a computationally efficient demapper calculation method utilizing the component-wise decoding scheme in [7] for the coded STBC systems. The STBC designs can be decoded separately exploiting the orthogonal property with the lattice representation. The complexity savings of the proposed method grows as the modulation size and the number of antennas increase. When applied to the quasi-orthogonal design, a complexity reduction by a factor of more than 10 is obtained. Regardless of this significant complexity reduction, the BER performance remains the same. The proposed demapper can be easily adopted to other systems, such as the orthogonal frequency division multiplexing (OFDM) technique.

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