Two High-rate Space-Time Codes for Three and Four Transmit Antennas with Good Diversity-Multiplexing Tradeoffs

Xiaoming Dai, Shaohui Sun, Xiangming Li, and Yingmin Wang

***Abstract*—In this work, we present two high-rate quasi-orthogonal space-time coding (STC) schemes for three and four transmit antennas which are designed in conjunction with the serial interference cancellation (SIC) based receiver. The symbols of the proposed STCs are of judiciously designed diversity order which can be effectively exploited by the SIC based receiver to cancel interference as well as to obtain diversity gain. The proposed STC schemes' amicable quasi-orthogonal structures lend themselves to efficient low-complexity linear decoding. Simulation results show that the proposed STC schemes achieve better diversity-multiplexing tradeoffs than the existing conventional schemes.**

***Index Terms*—Quasi-orthogonal, space-time coding (STC), serial interference cancellation (SIC), diversity-multiplexing.**

I. Introduction

The multiple-input multiple-output (MIMO) system, which employs multiple antennas at both the transmitter and receiver, can provide diversity and/or multiplexing gain(s) [1]. Most mainstream MIMO transmission schemes optimize only either the diversity advantage [2] or the spatial division multiplexing (SDM) gain [3]. The orthogonal space-time block codes (OSTBCs) [2], [4] can offer full transmit diversity and low-complexity linear decoding. However, the upper-bound of the maximum possible code rate (the number of information symbols per channel use) of the OSTBC is 3/4 for cases with more than three antennas [5] (the rate-1 is achieved only for the case with two antennas proposed by Alamouti [4]). The vertical Bell layered space-time architecture (V-BLAST) [3] with a medium-complexity ordered successive interference cancellation (OSIC) based detector achieves full transmission rate at the cost of diversity gain. The space-time coding (STC) transmission schemes proposed in [6], [7] can achieve both full diversity and high transmission rate. However, the associated maximum-likelihood (ML) decoding complexity is much higher than that of the V-BLAST, especially when the number of transmitter antennas or the constellation size is large. While sphere decoding (SD) [8] can be utilized to simplify the implementation of the ML detector for many practical problems, there still exists the issue of the choice of the appropriate covering radius of the lattice, and the complexity can be still quite high if the number of transmitter antennas is large. It has been recently shown in [9] that expected complexity of the sphere decoding is exponential for fixed signal-to-noise ratio (SNR). Therefore, the complexity of the SD can be very high [9] in the low SNR regime.

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Therefore, STC designs that can achieve full/high transmit diversity and high rate, but requiring only moderate fixed decoding complexity are highly desirable for practical applications. The groupwise space-time coding (GSTC) [10] based schemes for three and four transmit antennas [11] which combine the advantages of the OSTBC and SDM provide both diversity as well as multiplexing gains and have been presented to the IEEE802. 11n draft for future potential applications.

In this work, we propose two high-rate STC structures that consist of partially overlapped 2×2 orthogonal sub-matrixes for three and four transmit antennas in conjunction with the SIC based receiver. The joint design of the STC and the SIC based receiver efficiently exploits transmit antenna diversity and the merits of the SIC based detector. Hence, the proposed STC schemes achieve better diversity-multiplexing tradeoffs than their counterparts of the GSTC based schemes [11].

By convention, Upper (lower) bold face letters denotes matrices (column vectors). $\|\cdot\|$ and $\cdot^H$ denote Euclidean norm and transpose and Hermitian of a vector and matrix respectively. $\lceil a \rceil$ represents the smallest integer larger than or equal to real number a. s_i stands for the i-th entry of vector $\mathbf{s}$. $h_{i,k}$ denotes the entry in the i-th row and k-th column of matrix $\mathbf{H}$.

II. The Proposed STC Schemes for Three and Four Transmit Antennas

A. System and Channel Model

A discrete-time baseband channel model and quasi-static flat Rayleigh fading is assumed in this work. The quasi-stationarity means the channel characteristic remain constant for the period of transmission of an entire frame, with a duration of T symbols periods. For a multiple-antenna communication system with N_T transmit and N_R receive antennas, the transmitted and received signals are given by

$$\mathbf{Y} = \sqrt{\frac{\rho}{N_T}}\mathbf{HS} + \mathbf{V} \quad (1)$$

where $\mathbf{Y} = [\mathbf{y}_1, \mathbf{y}_2, ..., \mathbf{y}_t, ..., \mathbf{y}_T]$ denotes the matrix of complex received signals with the size of $N_R \times T$. $\mathbf{S} = [\mathbf{s}_1, \mathbf{s}_2, ..., \mathbf{s}_t, ..., \mathbf{s}_T] \in C^{N_T \times T}$ represents the matrix of complex transmitted signals, $\mathbf{H} \in C^{N_R \times N_T}$ denotes the channel matrix, and the additive noise $\mathbf{V} \in C^{N_T \times T}$ is $\mathcal{CN}(0, 1)$ (zero-mean, unit-variance, complex-Gaussian) distributed that is spatially and temporally white. ρ

is the SNR at each receive antenna. The channel coefficients $h_{i,k}, i = 1, ..., N_R, k = 1, ..., N_T$ are modeled as independent identically distributed (i.i.d.) complex circular Gaussian random variables, each with a $\mathcal{CN}(0, 1)$ distribution.

B. Review of the SIC Based Receiver

The SIC detector was first proposed in [12] and the key idea of SIC is that signals are decoded one after another, with the receiver canceling interference after each symbol. The decoded data for the first symbol is re-encoded and by using accurate channel knowledge, can be made to very closely resemble its received signal. Hence, it can be subtracted out of the composite received signal, and the second symbol to be decoded experiences less interference than it otherwise would have. In addition to its simplicity and amenability to implementation, SIC is also well-justified from a theoretical point of view. Simple successive interference cancellation implementation with optimal coding was shown to nearly achieve the Shannon capacity of multiuser additive white gaussian noise (AWGN) channels, assuming accurate channel estimation and a large spreading factor [13]. Other more recent work has proven that SIC with single user decoding in fact achieves the Shannon capacity region boundaries for both the broadcast (downlink) [14] and multiple access (uplink) multiuser channel scenarios [15]. The OSIC based detection algorithm of the V-BLAST system [3] is summarized as follows:

We summarize the OSIC based detection algorithm of the V-BLAST system [3] as follows:

1) At the first detection stage ($i = 1$), Let $\mathbf{H}^1 = \mathbf{H}$ and $\mathbf{y}_t^1 = \mathbf{y}_t$.
2) At the i-th detection stage, a nulling matrix $\mathbf{W}^i$ is calculated as

$$\mathbf{W}^i = (\mathbf{H}^i)^+ = \left((\mathbf{H}^i)^H(\mathbf{H}^i)\right)^{-1}(\mathbf{H}^i)^H \tag{2}$$

Choose the layer with largest signal-to-interference-plus-noise ratio (SINR) to detect. This corresponds to choosing the row of $\mathbf{W}^i$ with the minimum squared Euclidean norm and the corresponding row as the zero-forcing nulling vector

$$k_i = \arg\min_{j \notin \{1,\cdots,N_T-i+1\}} \left\|(\mathbf{W}^i)_j\right\|^2$$

$$\mathbf{w}_{k_i} = \left(\mathbf{W}^i\right)_{k_i}$$

where $(\mathbf{W}^i)_{k_i}$ denotes the k_ith row of matrix $\mathbf{W}^i$. Then, project the received signal $\mathbf{y}$ onto the nulling direction and perform the slicing operation to detect the symbol as follows

$$\hat{s}_{k_i} = Q\left(\mathbf{w}_{k_i}^H \mathbf{y}_t^i\right)$$

where $Q(\cdot)$ function describes the slicing operation that chooses the constellation point closest to the argument.
3) cancellation. Subtract the detected signal form the received sinal to get

$$\mathbf{y}_t^{i+1} = \mathbf{y}_t^i - \mathbf{h}_{k_i}\hat{s}_{k_i}$$

where $\mathbf{h}_{k_i}$ denotes the k_ith column of $\mathbf{H}$. Eliminating k_ith column of $\mathbf{H}^i$, the channel matrix is changed to

$$\mathbf{H}^{i+1} = \mathbf{H}^i_{\overline{k_i}}$$

where $\mathbf{H}^i_{\overline{k_i}}$ denotes the matrix obtained by eliminating the k_ith column from $\mathbf{H}^i$.
4) Recursion. Repeat the steps 2-3 for $i = i+1$ until all layers are detected ($i = N_T$).

The ordering of detection of the substreams is chosen to maximize the SINR at each stage and is equivalent to the global maximization of the minimum substream SINR [3]. The SIC based receiver has been widely acclaimed for providing a reasonable tradeoff between performance and complexity for MIMO system [3]. However, one main disadvantage of any system using nulling and canceling based approach has the drawback that errors occurred in detecting the transmitted symbols are propagated further into subsequent symbols due to interference subtraction. If the previous decisions are correct, the diversity order of the i-th interference cancellation (IC) stage of the V-BLAST system [16] is $N_R - N_T + i$, As a result, the detection of the first stage suffers the most severe interferences. The error propagation resulted from low diversity of early detection stages will severely degrade the overall performance. It has been shown in [16] that the total error rate is dominated by the *first-step* bit error rate (BER). Furthermore, errors in channel estimates further exacerbate the detrimental effects of the error propagation.

Increasing the number of transmit antennas N_R can alleviate the low diversity order of the detection of early stages. However, it may not be an appealing solution for the size- and power-stringent handheld mobile device. The ordering operation [3] can mitigate the error propagation, however, it can not fully compensate the low diversity orders of early detection stages, especially for high modulation alphabets or in the low SNR regime [17].

As mentioned beforehand, the overall system performance is usually limited by the symbols with worst performance, this gives us the hint that improving the performance of the *first* layer will be the most effective in enhancing the overall system performance.

C. Proposed Codewords

For N_T transmit antennas, a complex orthogonal space-time block code is described by a $T \times N_T$ transmission matrix $\mathbb{G}_{N_T}$, where each entry in $\mathbb{G}_{N_T}$ is a linear combination of the M variables $s_1, s_2, ..., s_M$ and their conjugates [19]. $\mathbb{G}_{N_T}$ can send M symbols from a signal constellation in a block of T channel uses. Since T time slots are used to transmit M symbols, the code rate of $\mathbb{G}_{N_T}$ is defined as $r_{\text{code rate}} = M/T$ [20].

Based on the observations of Section II-B, we propose the codewords for three and four transmit antennas as shown, respectively, in Fig. 1(a) and Fig. 1(b). The proposed STC schemes' code rates $r_{\text{code rate}}$ of the three and four transmit antennas are 4/3 and 7/4, respectively.

With $N_R = 3$ receive antennas, the system model of the

$$\mathbb{G}_3 = \begin{bmatrix} s_1 & s_2^* & s_4 \\ s_2 & -s_1^* & s_3 \\ s_4 & s_3^* & s_1 \end{bmatrix}$$

(a) $N_T = 3$.

$$\mathbb{G}_4 = \begin{bmatrix} s_1 & s_2^* & s_5 & s_7^* \\ s_2 & -s_1^* & s_3 & s_6^* \\ s_5 & s_3^* & s_1 & s_4^* \\ s_7 & s_6^* & s_4 & -s_1^* \end{bmatrix}$$

(b) $N_T = 4$.

Fig. 1. The proposed codeword matrixes for $N_T = 3$ and $N_T = 4$ transmit antennas.

proposed STC for the $\mathbb{G}_3$ can be expressed as

$$\begin{bmatrix} y_{11} & y_{12} & y_{13} \\ y_{21} & y_{22} & y_{23} \\ y_{31} & y_{32} & y_{33} \end{bmatrix} = \sqrt{\frac{\rho}{3}} \begin{bmatrix} h_{11} & h_{12} & h_{13} \\ h_{21} & h_{22} & h_{23} \\ h_{31} & h_{32} & h_{33} \end{bmatrix} \begin{bmatrix} s_1 & s_2^* & s_4 \\ s_2 & -s_1^* & s_3 \\ s_4 & s_3^* & s_1 \end{bmatrix} + \begin{bmatrix} v_{11} & v_{12} & v_{13} \\ v_{21} & v_{22} & v_{23} \\ v_{31} & v_{32} & v_{33} \end{bmatrix} \tag{3}$$

which can be rewritten as the following equivalent form

$$\underbrace{\begin{bmatrix} y_{11} \\ y_{21} \\ y_{31} \\ y_{12}^* \\ y_{22}^* \\ y_{32}^* \\ y_{13} \\ y_{23} \\ y_{33} \end{bmatrix}}_{\triangleq \mathbf{y}} = \sqrt{\frac{\rho}{3}} \underbrace{\begin{bmatrix} h_{11} & h_{12} & 0 & h_{13} \\ h_{21} & h_{22} & 0 & h_{23} \\ h_{31} & h_{32} & 0 & h_{33} \\ -h_{12}^* & h_{11}^* & h_{13}^* & 0 \\ -h_{22}^* & h_{21}^* & h_{23}^* & 0 \\ -h_{32}^* & h_{31}^* & h_{33}^* & 0 \\ h_{13} & 0 & h_{12} & h_{11} \\ h_{23} & 0 & h_{22} & h_{21} \\ h_{33} & 0 & h_{32} & h_{31} \end{bmatrix}}_{\triangleq \mathcal{H}_{\text{equivalent}}} \underbrace{\begin{bmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{bmatrix}}_{\triangleq \mathbf{s}} + \underbrace{\begin{bmatrix} v_{11} \\ v_{12} \\ v_{13} \\ v_{12}^* \\ v_{22}^* \\ v_{23}^* \\ v_{13} \\ v_{23} \\ v_{33} \end{bmatrix}}_{\triangleq \mathbf{v}}. \tag{4}$$

The $\mathcal{H}_{\text{equivalent}}$ for the $\mathbb{G}_4$ with $N_R = 3$ is given as:

$$\mathcal{H}_{\text{equivalent}} = \begin{bmatrix} h_{11} & h_{12} & 0 & 0 & h_{13} & 0 & h_{14} \\ h_{21} & h_{22} & 0 & 0 & h_{23} & 0 & h_{24} \\ h_{31} & h_{32} & 0 & 0 & h_{33} & 0 & h_{34} \\ -h_{12}^* & h_{11}^* & h_{13}^* & 0 & 0 & h_{14}^* & 0 \\ -h_{22}^* & h_{21}^* & h_{23}^* & 0 & 0 & h_{24}^* & 0 \\ -h_{32}^* & h_{31}^* & h_{33}^* & 0 & 0 & h_{34}^* & 0 \\ h_{13} & 0 & h_{12} & h_{14} & h_{11} & 0 & 0 \\ h_{23} & 0 & h_{22} & h_{24} & h_{21} & 0 & 0 \\ h_{33} & 0 & h_{32} & h_{34} & h_{31} & 0 & 0 \\ -h_{14}^* & 0 & 0 & h_{13}^* & 0 & h_{12}^* & h_{11}^* \\ -h_{24}^* & 0 & 0 & h_{23}^* & 0 & h_{22}^* & h_{21}^* \\ -h_{34}^* & 0 & 0 & h_{33}^* & 0 & h_{32}^* & h_{31}^* \end{bmatrix} \tag{5}$$

It is straightforward to derive $\mathcal{H}_{\text{equivalent}}$ for the proposed codewords $\mathbb{G}_3$ and $\mathbb{G}_4$ with arbitrary number N_R ($>= \lceil r_{\text{code rate}} \rceil$, otherwise the uncoded system is under-determined) of receiver antennas.

The design of unequal diversity order of the symbols is to some extent inspired by the fact that [18] an unequal distribution of received powers is amicable for the SIC based detector to perform well in code-division multiple-access (CDMA) systems, especially when the interference cancellation is *imperfect* which is inherent in the SIC detection process.

With a OSIC detector, the symbol s_1 has the highest diversity order (assumed to be detected first in general) and the first layer BER is substantially reduced. As a result, the error propagation is, thus, significantly mitigated in the detection of subsequent layers. The symbols of lower diversity orders of the proposed codeword can benefit from accurate detection of the previously detected symbols with higher diversity orders. The reliable estimates from the previously detected symbols are then utilized in detection of subsequent symbols, thus, result in a better overall BER performance. Therefore, the judiciously designed diversity orders of the symbols of different layers are efficiently exploited by the SIC based receiver to suppress interfering signals as well as to obtain diversity gain. The diversity gain obtained in the SIC detection process can be leveraged to increase the transmission rate, thus, achieving a better multiplexing-diversity tradeoff (demonstrated by numerical results in Section III).

The unequal diversity order based STC scheme lends itself to the SIC based detector, even for the fixed-order SIC (FSIC) one, where the symbols at the SIC receiver side are processed in accordance with the data index, s_1 is detected first, and the s_2 is detected second, and so on, i.e., s_1, s_2,...,$s_{N_T r_{\text{code rate}}}$ etc.

The diagonal 2×2 sub-matrixes in the square box of Fig. 1(a) and Fig. 1(b) are orthogonal, thus, greatly alleviating the self interference. The amicable quasi-orthogonal structure makes it amenable for efficient low-complexity linear decoding, and it also *alleviates* the error propagation in the SIC detection process.

III. Simulation Results

In this Section, we provide simulation results of the proposed STC schemes, the GSTC based schemes employing the OSIC based detector and the zero-forcing linear detector (ZF-LD) [11]. The codewords of the GSTC based schemes [11] of $N_T = 3$ and $N_T = 4$ antennas are, respectively, expressed as

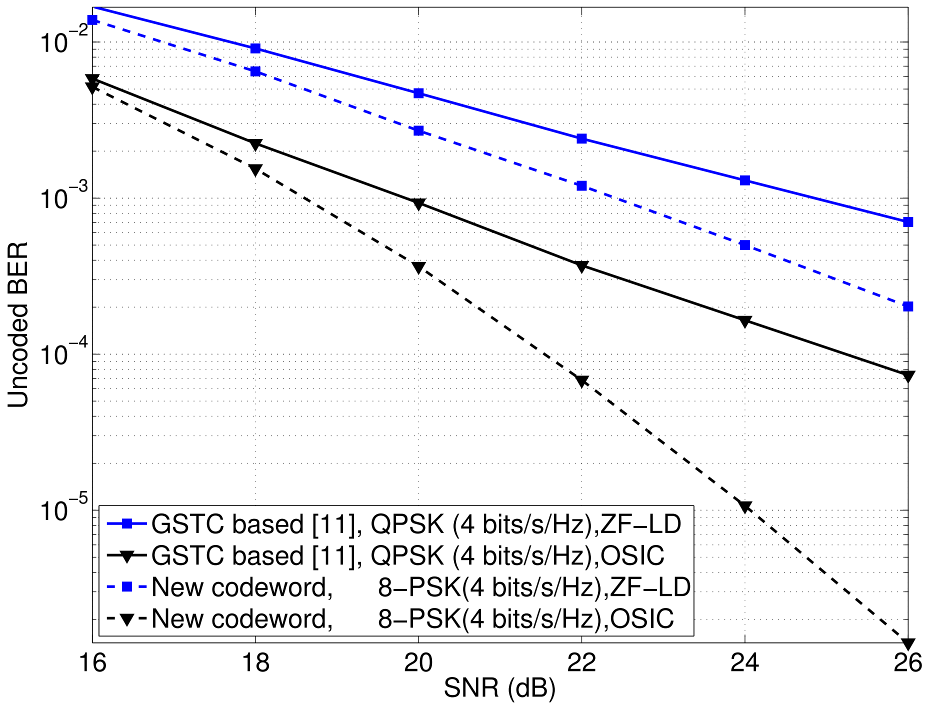

Fig. 2. BER curves of the proposed STC scheme and the GSTC based scheme [11] for three transmit and two receive antennas.

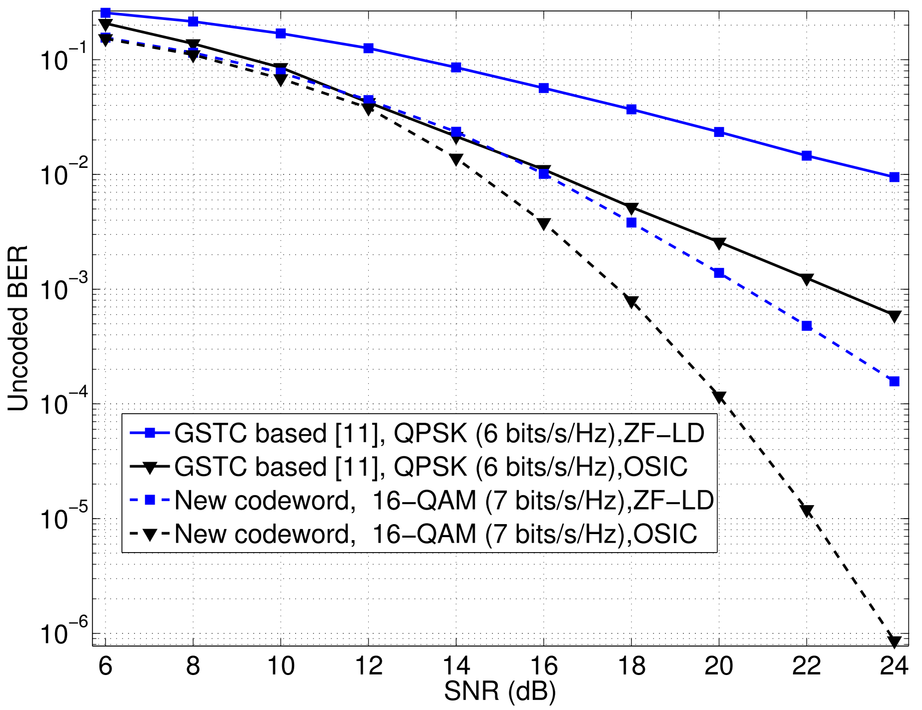

Fig. 3. BER curves of the proposed STC scheme and the GSTC based scheme [11] for four transmit and three receive antennas.

$\mathbb{G}_3 = \begin{bmatrix} s_1 & s_2^* \\ s_2 & -s_1^* \\ s_3 & s_4^* \end{bmatrix}$, and $\mathbb{G}_4 = \begin{bmatrix} s_1 & s_2^* \\ s_2 & -s_1^* \\ s_3 & s_4^* \\ s_5 & s_6^* \end{bmatrix}$. Therefore, the code rates are 2 and 3 for the GSTC based schemes of three and four transmit antennas, respectively.

We assume that the channels between the transmit antennas and the receive antennas are independent quasi-static flat Rayleigh-fading channels. The elements of the MIMO channel are obtained from an independent and identically distributed (i.i.d.) complex Gaussian distribution with zero mean and unit variance. The channel state information (CSI) was assumed to be perfectly known at the receiver.

A. $\mathbb{G}_3$

Fig. 2 provides the uncoded average BER versus SNR performance of the proposed codeword of eight phase shift keying (8-PSK) and the GSTC based scheme [11] of quadrature phase shift keying (QPSK). Therefore, the transmission rates of both schemes are the same (i.e., 4/3*3=2*2=4 bits/s/Hz). Fig. 2 shows that the proposed codeword using the OSIC based receiver obtains about 4 dB gain over the GSTC based scheme [11] at the BER of 1×10^{-3}. The proposed codeword with ZF-LD detection achieves about 3.8 dB gain over that of the GSTC at the BER of 1×10^{-2}.

B. $\mathbb{G}_4$

For four transmit antennas, we applied 16-ary quadrature amplitude modulation (16-QAM) and QPSK for the proposed codeword and the GSTC based scheme [11], respectively. Fig. 3 illustrates that the proposed STC scheme with OSIC receiver gives about 5 dB gain over the GSTC based scheme at the BER of 1×10^{-3}. About 7.5 dB gain is obtained by the proposed STC scheme with the ZF-LD at the BER of 1×10^{-2}. The proposed codeword using the ZF-LD *even* achieves better performance than the GSTC based scheme with the more complex SIC receiver in the medium-to-high SNR regime as shown in Fig. 3. The proposed codeword achieves a higher transmission rate of 7 bits/s/Hz (=7/4*4) than the GSTC based scheme [11] of 6 bits/s/Hz (=3*2) with the OSIC and ZF-LD receivers in this setup.

The uncoded BER curve of proposed STC scheme has a larger slope than that of the hybrid STC [11] in the high-SNR regime, which implies that proposed STC scheme has a better diversity gain. Therefore, proposed STC achieves a better diversity-multiplexing tradeoff, which is consistent with the analysis in Section II-C.

IV. Conclusions

A partially overlapped 2×2 orthogonal structure is exploited to design two high-rate STC schemes for three and four transmit antennas with the SIC based receiver in this work. The proposed unequal diversity based STC schemes can be effectively exploited by the SIC based receiver to cancel interference as well as to obtain diversity gain. The proposed SIC-amenable STCs' quasi-orthogonal structure alleviates the error propagation which is inherent in the SIC based system, it is also amicable for the low-complexity linear decoding.

There is a consensus among the industry representatives [11], [21] to mainly propose and deploy in the initial stage of next-generation systems MIMO schemes with up to four transmit antennas at the base-station and up to *two* antennas at the mobile device due to cost and space constraints. Therefore, the proposed STC schemes are of appeals for applications in both symmetric and asymmetric deployments. The efficient LD- and the SIC-amenable characteristic of the proposed STC schemes are also appealing for power-stringent mobile devices.

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